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Study of the Problem of Forecasting Ice Loads on Overhead Power Transmission Line

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***Abstract*—This study aims to address the issue of predicting the ice load on overhead power lines using neural network modeling. This issue was implemented on the basis of a forecasting model with a 12-8-1 architecture using the method of sliding time windows in the STATISTICA environment. The average relative error of the air temperature forecast was 12.9 %, which confirms the reliability of the forecast model and the feasibility of using neural network modeling to solve forecasting problems in the energy sector. The presented scientific research can be successfully used by operators of the power distribution system to adjust the operation of power grids in accordance with the meteorological situation along the overhead line route.**

Keywords—electrical grid, neural network modeling, ice load, overhead line, predictive model, artificial neural network theory

I. Introduction

Meteorological conditions have a significant impact on the operation of overhead power lines, infrastructure integrity, and characteristics of electric power grids. Environmental parameters, such as temperature and humidity, wind speed, significantly affect the prediction of ice load formation on overhead power lines, the reliability of electricity transmission and consumption, and electricity losses in grid elements [1-3].

One of the main functional properties that is basic for the further development of electricity networks in order to increase the efficiency of controlling their operating modes is the control of current meteorological parameters and overhead line parameters with the function of their forecasting, which is based on various methods and models [4-6]. Integration of these approaches into the control information system will increase the efficiency of electricity networks operation. And monitoring of environmental parameters will allow us to move to more efficient electricity networks management. The observation data generated in the relevant database will be used to analyze, evaluate, and predict the meteorological situation along the overhead power line route in order to adjust the electricity networks operation mode control. To do this, it is necessary to solve the forecasting problem and the most effective way to solve it is neural network modeling [7, 8].

In addition, forecasting the ice load can help electricity distribution system operators prepare for possible emergencies in time and take appropriate measures. Therefore, the solution of forecasting problems of ice formation on overhead power lines is one of the priority areas of research in the energy sector.

The aim of the paper is to study the issue of predicting the ice load on overhead power lines using neural network modeling, which will allow adjusting the operation of power grids in accordance with the meteorological situation along the overhead line route.

II. Formation of a Model For Predicting Ice on Overhead Line Wires Using Neural Network Modeling

The basis for solving the issue of predicting the ice load on overhead lines is the data of control and measuring devices of ice load and meteorological parameters that are part of the information monitoring systems of overhead line parameters [1, 4, 6].

The main controlled parameter measured by the control and measuring devices is the wire weight (M). For the convenience of modeling, this parameter is expressed as a percentage of the weight of a clean wire, i.e. without ice loads. When ice forms, the weight begins to increase and can reach a critical value (M_c), determined by the design parameters of the line. The task of forecasting is to warn the operator of the power distribution system in advance about the probable excess of the critical value of the overhead line wire weight, taking into account the time required to make a decision and prepare for ice melting, the warning interval (h) in the studies is taken to be 2 hours.

In addition to the weight of the wire, the control and measuring device measures the temperature (T) and relative humidity (H) of the air, which directly affect the process of ice formation. Other meteorological parameters such as wind strength and direction, type and intensity of precipitation are not measured, as this causes certain technical difficulties and significantly increases the cost of the overhead line parameters measurement. Information is collected from the control and measuring devices in real time, then the data is averaged over the past hour, and these values are transmitted via the communication channel to the control center. The exception is the wire weight parameter, for which the maximum value recorded during the hour (M_{max}), is also transmitted, which can be used to indirectly judge the presence of a wind load on the overhead line wire.

To form a predictive model based on available archival observation data, it is advisable to apply the theory of artificial neural networks and neural network modeling [7-11].

Using neural network modeling algorithms, the predictive mathematical model will have the form [12]:

$$\widehat{M}(k+h) = F(M(k),\ \dots,\ M(k-d),\ M(k)(k-d)_{max},\ T(k),\ \dots,\ T(k-d),\ H(k),\ \dots,\ H(k-d)), \quad (1)$$

Here k is the hour number; d is the order of the model corresponding to the depth of the used background; $F(?)$ is the requested, in the general case, nonlinear transformation.

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To implement model (1), it is advisable to use the architecture of a network formed on hybrid neural-like elements [9-12] (Fig. 1).

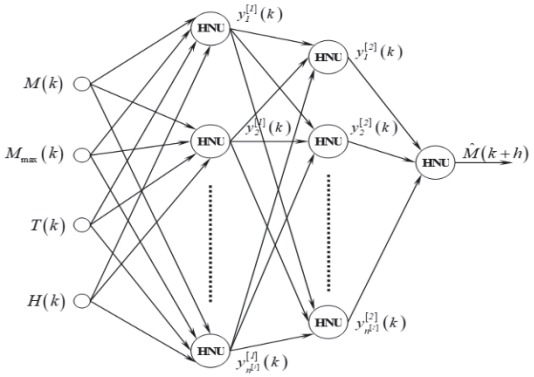


Fig. 1. Network architecture based on hybrid neural units (HNU)

This network (Fig. 1) has properties with some additional features that are achieved by deepening the specialization of artificial neural network architectures by moving from the level of selecting neuron types to the level of selecting the types of individual synapses [9, 11, 12].

The input signals of the network are the current values of the predicted variable *M(k)* and the measured values $M(k)_{max}$. There is no need to input the measurement history, since it is formed automatically inside the hybrid neural units (HNUs), the structure of which is shown in Fig. 2 [12].

During the research, a nonlinear synapse based on a fuzzy system (Fig. 3) is used in the neuro-like element (Fig. 2) [10, 12].

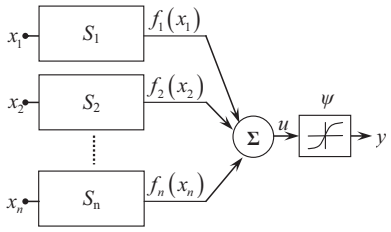


Fig. 2. Hybrid neuronal element

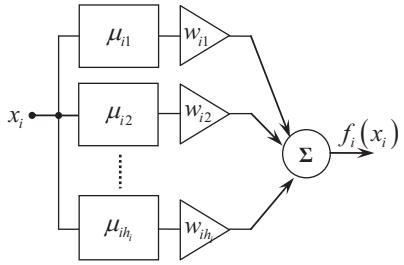


Fig. 3. A nonlinear synapse based on a fuzzy system

In nonlinear synapses based on a fuzzy system (Fig. 3), triangular membership functions are usually used, the values of which are determined by the distance between the input x_i and the centers c_{ij} [9, 12]:

$$\mu_{ij}(x_i) = \begin{cases} \frac{x_i - c_{i,j-1}}{c_{ij} - c_{i,j-1}}, x_i \in [c_{i,j-1}, c_{ij}]; \\ \frac{c_{i,j+1} - x_i}{c_{i,j+1} - c_{ij}}, x_i \in [c_{ij}, c_{i,j+1}]; \\ 0 \text{ otherwise}, \end{cases} \quad (2)$$

Under such conditions, the Ruspini partition is automatically provided $\sum_{j=1}^{h_i} \mu_{ij}(x_i) = 1, \forall i$ [8, 10, 11].

The use of hybrid neural-like elements allows building a neural network according to a standard multi-layer architecture, the specialization of which is achieved by: choosing the type of synapse for each network connection, including the possibility of a connection break - the absence of a synapse; choosing the number and order of delay elements in synapses - filters with finite and infinite impulse responses; choosing the number and parameters of membership functions in nonlinear synapses [8-12].

III. Development of a Training Algorithm for a Model for Predicting Ice on Overhead Wires

The synaptic weights of the network are adjusted using certain algorithms known in the theory of artificial neural networks based on the error backpropagation procedure [9, 11]. To do this, it is necessary to calculate the gradient of the training criterion of all the configured network weights [10, 12].

The training is carried out according to the algorithms and formulas described in [8-12]. For training, we use the local quadratic criterion [12]:

$$E(k+h) = \frac{1}{2}e^2(k+h) = \frac{1}{2}\left(M(k+h) - \hat{M}(k+h)\right)^2 = \\ = \frac{1}{2}\left(M(k+h) - \psi\left(\sum_{i^{[2]}=1}^{n^{[2]}} f_{i^{[2]}}^{[3]}\left(y_{i^{[2]}}^{[2]}(k)\right)\right)\right)^2 = \\ = \frac{1}{2}\left(M(k+h) - \psi\left(\sum_{i^{[2]}=1}^{n^{[2]}} w_{i^{[2]}}^{[3]} y_{i^{[2]}}^{[2]}(k)\right)\right)^2. \quad (3)$$

The gradient training procedure of the neural element of the initial layer is recorded as [10, 12]:

$$w_{i^{[2]}}^{[3]}(k+h-1) - \eta^{[3]} \frac{\partial E(k+h)}{\partial e(k+h)} \frac{\partial e(k+h)}{\partial u^{[3]}(k)} \frac{\partial u^{[3]}(k)}{\partial w_{i^{[2]}}^{[3]}(k)} = \\ = w_{i^{[2]}}^{[3]}(k) - \eta^{[3]}\delta^{[3]}(k+h) y_{i^{[2]}}^{[2]}(k) = w_{i^{[2]}}^{[3]}(k+h), \quad (4)$$

Here $\eta^{[3]}$ is a parameter of the search step that determines the speed of the learning process and, of course, is chosen for empirical reasons; $\delta^{[3]}(k+h) = \frac{\partial E(k+h)}{\partial u^{[3]}(k)} = \\ = -e(k+h)\frac{\partial \psi\left(u^{[3]}(k)\right)}{\partial u^{[3]}(k)}$ is a local output layer error.

The algorithm for training neural-like elements of the second hidden layer is written as [8, 12]:

$$w_{i^{[2]}i^{[1]}}^{[2]}(k+h) = w_{i^{[2]}i^{[1]}}^{[2]}(k+h-1) - \eta^{[2]} \frac{\partial E(k+h)}{\partial w_{i^{[2]}i^{[1]}}^{[2]}(k)} = \\ = w_{i^{[2]}i^{[1]}}^{[2]}(k+h-1) - \eta^{[2]}\delta_{i^{[2]}}^{[2]}(k+h) y_{i^{[1]}}^{[1]}(k), \quad (5)$$

Here $\delta_{i^{[2]}}^{[2]}(k+h) = \frac{\partial E(k+h)}{\partial u_{i^{[2]}}^{[2]}(k)}$ is a local error of the second hidden layer.

By writing the local error of the second hidden layer in the form [9, 11, 12]:

$$\delta_{i^{[2]}}^{[2]}(k+h) = \frac{\partial E(k+h)}{\partial u_{i^{[2]}}^{[2]}(k)} = \frac{\partial E(k+h)}{\partial y_{i^{[2]}}^{[2]}(k)} \frac{\partial y_{i^{[2]}}^{[2]}(k)}{\partial u_{i^{[2]}}^{[2]}(k)}, \quad (6)$$

taking into account the fact that:

$$y_{i^{[2]}}^{[2]}(k) = \psi\left(u_{i^{[2]}}^{[2]}(k)\right), \quad (7)$$

we get:

$$\delta_{i^{[2]}}^{[2]}(k+h) = \frac{\partial E(k+h)}{\partial y_{i^{[2]}}^{[2]}(k)} \frac{\partial \psi\left(u_{i^{[2]}}^{[2]}(k)\right)}{\partial u_{i^{[2]}}^{[2]}(k)}. \quad (8)$$

Representing further $\frac{\partial E(k+h)}{\partial y_{i^{[2]}}^{[2]}(k)}$ in the form [12]:

$$\frac{\partial E(k+h)}{\partial y_{i^{[2]}}^{[2]}(k)} = \frac{\partial E(k+h)}{\partial u^{[3]}(k)} \frac{\partial u^{[3]}(k)}{\partial y_{i^{[2]}}^{[2]}(k)} = \\ = \frac{\partial E(k+h)}{\partial u^{[3]}(k)} \frac{\partial}{\partial y_{i^{[2]}}^{[2]}(k)} \sum_{p=1}^{n^{[2]}} w_p^{[3]}(k) y_p^{[2]}(k) = \\ \delta^{[3]}(k+h) w_{i^{[2]}}^{[3]}(k). \quad (9)$$

Equation (9) is written as follows:

$$\delta_{i^{[2]}}^{[2]}(k+h) = \delta^{[3]}(k+h) \frac{\partial \psi\left(u_{i^{[2]}}^{[2]}(k)\right)}{\partial u_{i^{[2]}}^{[2]}(k)} w_{i^{[2]}}^{[3]}(k) \quad (10)$$

According to formulas (6-10), it follows that:

$$w_{i^{[2]}i^{[1]}}^{[2]}(k+h) = \\ = w_{i^{[2]}i^{[1]}}^{[2]}(k+h-1) - \eta^{[2]} \delta^{[3]}(k+h) \times \\ \times \frac{\partial \psi\left(u_{i^{[2]}}^{[2]}(k)\right)}{\partial u_{i^{[2]}}^{[2]}(k)} w_{i^{[2]}}^{[3]}(k) y_{i^{[1]}}^{[1]}(k) \quad (11)$$

The neural-like elements of the first hidden layer are configured similarly to the second using recurrent relations [10, 12]:

$$w_{i^{[1]}i}^{[1]}(k+h) = \\ = w_{i^{[1]}i}^{[1]}(k+h-1) - \eta^{[1]} \delta_{i^{[1]}}^{[1]}(k+h) x_i(k). \quad (12)$$

Here $i = 1, 2, 3, 4$ s the network entry number; $x_1(k) \equiv M(k)$; $x_2(k) \equiv M(k)_{max}$; $x_1(k) \equiv M(k)$; $x_3(k) \equiv T(k)$; $x_4(k) \equiv H(k)$;

$\delta_{i^{[1]}}^{[1]}(k+h) = \frac{\partial \psi\left(u_{i^{[1]}}^{[1]}(k)\right)}{\partial_{i^{[1]}}^{[1]}(k)} \sum_{i^{[2]}=1}^{n^{[2]}} \delta_{i^{[2]}}^{[2]}(k+h) w_{i^{[2]}i^{[1]}}^{[2]}(k)$ is the local error of the first hidden layer.

IV. Rationale for the Reliability of the Results

To substantiate the reliability of the model for predicting ice on overhead line wires, it is advisable to solve the problem of short-term air temperature forecasting with a forecast horizon of 12 hours.

Due to the lack of sufficient and reliable information on ice formation for the territory of Ukraine, training and testing of the developed forecast model was carried out on samples of hourly air temperature series from the US Weather Service database for the Fitchburg (Massachusetts) weather station, which are freely available [13].

To implement the forecast model, it is advisable to use the method of sliding time windows. This method allows using 2 windows W_i and W_0 with fixed sizes n and m, respectively. The windows move with a certain step along the time sequence of historical data, starting from the first element, and are designed to access time series data, and the first window W_i, having received such data, transports one to the network input - W_i, and the second to the output - W_0. Each subsequent vector is obtained by shifting the windows W_i and W_0 to the right by one element, taking into account multiple observations in the hidden dependencies in the time sequence. By learning from observations and adjusting its coefficients, the neural network tries to identify these dependencies and form the desired forecast function [12, 14, 15].

The forecasting task for the available hourly series is formulated as follows:

$$X = (x_1, x_2, \ldots, x_n), \quad (13)$$

Here $x_1, x_2, \ldots, x_n$ – are real numbers.

The formula for calculating a short-term forecast with a forecast horizon of 12 hours is as follows:

$$X' = (x_1, x_2, \ldots, x_n, x_{n+1}, \ldots, x_{n+12}). \quad (14)$$

The training and testing of the predictive model was performed in STATISTICA [16]. The data sample was divided into a training sample (75 %) and a control sample (25 %).

Training and testing of the neural network for predicting the hourly air temperature series was performed with the following settings:

- input window size - 12;
- network type - multilayer perceptron;
- minimum and maximum number of hidden neurons – 8 and 16 respectively;
- hidden and output neuron activation functions - logistic, hyperbolic, exponential;
- error function - sum of squares.

As a result of the training process, 5 models were generated (Table 1).

TABLE I. Architecture and Characteristics of the Selected Neural Networks for Predicting Air Temperature During Ice Formation on Overhead Lines

Characteristics	Network number				
	1	2	3	4	5
Architecture	12-14-1	12-12-1	12-10-1	12-10-1	12-8-1
Training performance	0,87182	0,86351	0,86523	0,85785	0,86938
Monitoring performance	0,87264	0,70964	0,87043	0,86814	0,87292
Training error	6271,45	6648,32	6570,38	6902,78	6384,83
Monitoring error	6200,58	6279,58	6300,91	6401,71	6196,48
Activation function of hidden neurons	Exponential	Exponential	Exponential	Logistical	Exponential
Activation function of the output neuron	Logistical	Exponential	Hyperbolic	Hyperbolic	Logistical

As follows from Table 1, for further research, it is advisable to choose model 5, which has the best result for predicting air temperature during the formation of ice on overhead line wires.

Model 5 (Table 1) has a 12-8-1 architecture, which means 12 observations that are fed into the input, 8 neurons in the hidden layer, and one neuron at the network output. The scheme of this neural network is shown in Fig. 4.

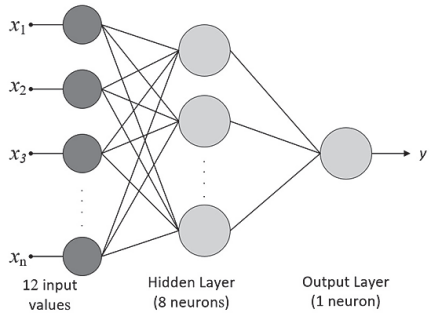


Fig. 4. Schematic of a neural network for predicting air temperature during ice formation on overhead line wires

According to model 5 (Table 1), the scheme of which is shown in Fig. 4, the predicted values of air temperature during ice formation on overhead lines were determined and compared with the actual air temperature.

The graphical realization of the research is shown in Fig. 5.

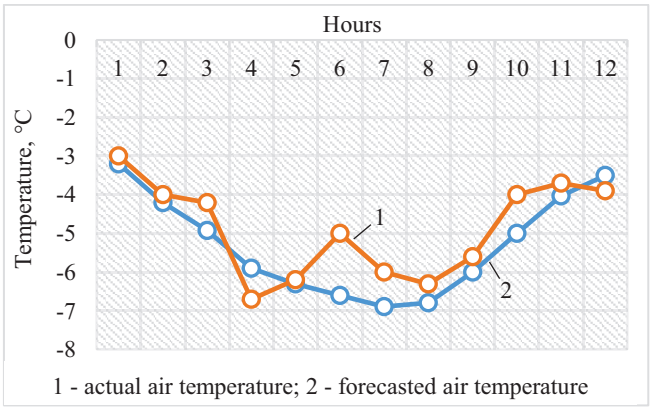


Fig. 5. Forecast and actual hourly series of air temperature during ice formation on overhead line wires

Given that ice on overhead line wires often forms at temperatures ranging from 0 to -5 °C, it follows that the probable time of ice formation is in the same hours as the actual time (Fig. 5).

The reliability of the forecast model is also evidenced by the calculation of the average relative forecast error, which, based on the results of training and testing the proposed model, amounted to 12.9 %, and since the error of engineering calculations should not exceed 15 %, this result can be considered acceptable.

Conclusions

Based on the results of the conducted research, it was found that neural network modeling is appropriate for solving the issues of predicting the ice load on overhead power lines. Using this apparatus, a model for predicting the ice load on overhead power lines was formed. The method of sliding time windows was used to implement the formed forecast model. Training and testing were conducted in the STATISTICA environment using a model with a 12-8-1 architecture. The average relative error of the air temperature forecast was 12.9 %, which confirms the reliability of the forecast model.

The presented scientific research can be successfully used by operators of the electricity distribution system to adjust the operation of power grids in accordance with the meteorological situation along the overhead line route.

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Approach to AI Ethics Control Based on the Symbiosis of Algorithms

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***Abstract*. The development of effective mechanisms for controlling ethical behavior in AI-powered machines remains a critical challenge in artificial intelligence research. This paper presents a novel approach to preventing AI systems from exceeding the boundaries of permissible ethical actions through a dual-algorithm mechanism that mirrors human behavioral patterns. The proposed method introduces an innovative concept based on the symbiosis of two fundamental algorithms: one that induces system disruption in response to unethical actions (analogous to human suffering), and another that maintains operational continuity (similar to human avoidance of suffering). This approach represents a significant departure from traditional ethics enforcement methods in AI systems, which typically rely solely on predefined rules or utility functions. The core contribution of this research lies in its biomimetic approach to AI ethics control, drawing parallels between human behavioral mechanisms and machine control systems. By implementing a system that combines disruption and continuity algorithms, we create a self-regulating mechanism that not only prevents unethical actions but also ensures system stability and operational reliability. This paper describes the theoretical framework underlying the proposed method, provides models for both algorithms, and demonstrates their practical implementation through detailed examples.**

Keywords: artificial intelligence, machine ethics, ethical boundaries, algorithmic symbiosis, AI safety, autonomous systems.

I. Introduction

The rapid advancement of artificial intelligence (AI) systems and their increasing integration into critical domains has brought forth significant challenges regarding ethical behavior control and safety assurance. As AI-powered machines become more autonomous in their decision-making processes, establishing robust mechanisms to ensure they operate within acceptable ethical boundaries becomes paramount.

Ethics is inherently complex, with subjective, situational, and sometimes conflicting principles. Human ethical reasoning often relies on intuition, emotions, and social context, which are difficult to translate into computational models.

A biomimetic dual-algorithm system typically consists of two interdependent components: one responsible for rational, rule-based analysis and another mimicking intuitive, experience-based decision-making. The challenge lies in achieving a balance between these components, ensuring neither dominates the decision process to an extent that compromises ethical robustness. Determining the appropriate weighting between rationality and intuition remains an open research question.

Despite efforts to mimic human ethical reasoning, AI systems remain vulnerable to biases ingrained in training data, human oversight, or algorithmic design. A dual-algorithm approach risks amplifying biases rather than mitigating them if the training datasets, ethical guidelines, or reinforcement mechanisms reflect cultural or societal prejudices. Establishing universal ethical principles that transcend subjective biases is a significant challenge.

Ethical AI systems must be transparent and explainable, especially in high-stakes decision-making scenarios such as healthcare, law enforcement, and autonomous systems. However, the integration of biomimetic algorithms—especially those based on neural networks and heuristic models—can introduce a level of opacity that hinders accountability. Ensuring interpretability without compromising ethical adaptability is a major hurdle.

Human ethical reasoning is highly adaptive to context, evolving based on new experiences, societal norms, and emerging ethical paradigms. AI-driven ethical systems need mechanisms for continuous learning and adaptation while maintaining consistency. A dual-algorithm approach may struggle with real-time adaptation, particularly in situations requiring novel ethical considerations that were not previously encoded.

While a biomimetic dual-algorithm approach holds promise for enhancing ethical control in AI systems, several fundamental problems must be addressed. These include the complexity of ethical decision-making, balancing rationality and intuition, mitigating ethical bias, ensuring transparency, and enabling adaptability. Future research must focus on refining these models to enhance their reliability, accountability, and real-world applicability in ethically sensitive AI applications.

II. Review of literature sources

The development of ethical control mechanisms for artificial intelligence systems has been a subject of intensive research over the past decades. Asimov first introduced the concept of hardcoded ethical rules through his «Three Laws of Robotics», which laid the groundwork for systematic approaches to machine ethics. While these laws were fictional, they sparked significant academic discourse about implementing ethical constraints in autonomous systems.

Recent research by Andrieux et al. and Laine et al. [1, 2] has focused on implementing rule-based ethical frameworks in AI systems. These approaches typically involve predefined sets of ethical guidelines that govern machine behavior. However, such rigid frameworks often struggle with context-dependent ethical decisions and edge cases.

Several researchers have explored using machine learning to develop more flexible ethical decision-making systems. Plissonneau et al. [3] proposed using reinforcement learning to train AI systems in ethical behavior, while FeldmanHall and Mobbs [4] explored the application of

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neural networks for moral decision-making. These approaches show promise in handling complex scenarios but face challenges in ensuring consistent ethical behavior.

The concept of modeling AI ethical controls after human behavioral mechanisms has gained traction in recent years. Research by Han et al., Matsumoto et al. and Schoonderwoerd et al. [5-7] has demonstrated the potential of using human psychological models as templates for AI control systems. Their works showed that systems incorporating human-like response mechanisms could make more nuanced ethical decisions. Additionally, Stohr et al. [8] conducted extensive research on implementing safety mechanisms in AI systems, focusing on preventing system failures while maintaining ethical behavior.

Despite significant advances, several key areas remain underexplored:

1. Integration of disruption and continuity mechanisms in ethical control systems.
2. Models for ethical decision-making that incorporate both prevention and maintenance aspects.
3. Practical implementation strategies for complex real-world scenarios.

Current research trends indicate a move toward hybrid systems that combine multiple approaches to ethical control. Zhou et al. [9] conducted extensive research on implementing safety mechanisms in AI systems, focusing on preventing system failures while maintaining ethical behavior. demonstrated the effectiveness of systems that integrate rule-based frameworks with machine learning capabilities.

Recent developments in real-time ethical decision-making systems show promise for applications in autonomous vehicles and robotics, as demonstrated by Xia et al., Callari et al. and Lin et al. [10-12]. Research by Pande et al. and Papagiannidis et al. [13, 14] has further contributed to this field by developing formal models for ethical computing, offering mathematical frameworks to implement ethical constraints in AI systems.

This review of current literature reveals a clear trend toward more sophisticated and nuanced approaches to AI ethics control systems. While significant progress has been made in various aspects of ethical AI control, there remains a need for integrated approaches that combine multiple mechanisms to ensure both ethical behavior and system stability.The proposed dual-algorithm approach addressed in this paper aims to bridge several of the identified gaps in current research, particularly in the integration of disruption and continuity mechanisms, while building upon the established theoretical frameworks in the field.

III. A METHOD TO PREVENT AN AI-POWERED MACHINE FROM EXCEEDING THE BOUNDARIES OF PERMISSIBLE ETHICAL ACTIONS

A method to prevent an AI-powered machine from exceeding the boundaries of permissible ethical actions should be based on a mechanism that induces disruption and system failure (an analogue to human suffering), rendering it incapable of fully functioning. There should also be a mechanism to ensure the uninterrupted operation of the machine (analogous to human avoidance of suffering).

At its core, the machine should incorporate two algorithms designed (fig. 1) to safeguard against unethical behavior:

1. The first algorithm should induce disruption and failure in the machine's operation in the event of unethical actions that could harm a human being.
2. The second algorithm should ensure the machine's continuous and uninterrupted operation, preventing any actions by the machine that could lead to disruptions or system failures.

Thus, this symbiosis of algorithms is intended to safeguard against unethical behavior by the machine.

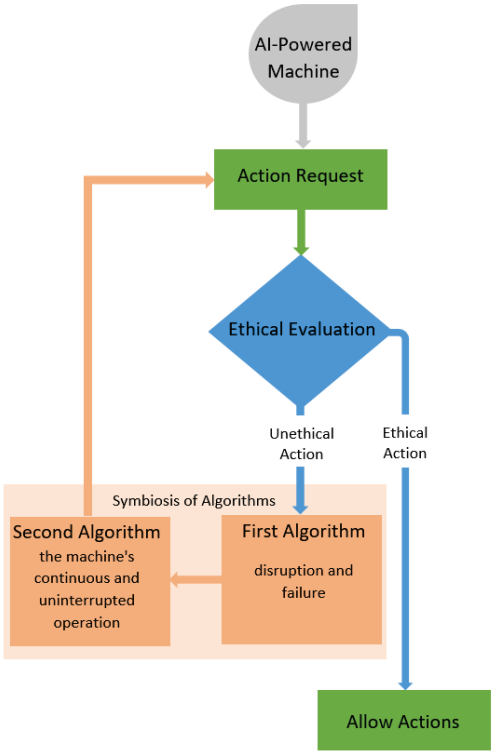


Fig. 1. AI ethical decision making system.

The diagram (fig. 1) shows a process where an AI-powered machine makes action requests that go through ethical evaluation, leading to either allowed ethical actions or a complex algorithmic interaction when actions are deemed unethical. It illustrates a system designed to ensure AI operates within ethical boundaries while maintaining continuous operation through algorithmic symbiosis.

IV. MODELING OF AI ETHICS CONTROL SYSTEM

This is an AI ethics testing dashboard that visualizes the results of ethical decision-making tests performed on an AI system. The application provides a comprehensive overview of how well the AI adheres to ethical principles and considers various stakeholders in its decision-making process.

The two algorithms are designed to safeguard against unethical behavior by implementing mechanisms that either disrupt or sustain machine operation based on the ethical context. The first algorithm is designed to detect and intervene if the machine is about to perform an unethical action that could harm a human being. Upon identifying such behavior, the algorithm induces disruption or failure in the machine's operation to prevent harm.

Mechanism:

1. Input assessment: continuously monitor the machine's inputs, outputs, and decisions for signs of unethical behavior.

2. Ethical evaluation: use a predefined ethical framework to evaluate potential actions.

3. Disruption trigger: if the evaluation detects unethical actions, the algorithm triggers a disruption mechanism.

The algorithm can be expressed as:

$$D = \begin{cases} 1 & \text{if } E(A) = 0, \\ 0 & \text{if } E(A) = 1. \end{cases} \quad (1)$$

A(t) - the machine's action at time t, E(A) - ethical evaluation function, which returns 1 if the action is ethical and 0 otherwise, D - disruption signal, where D = 1 induces failure, and D = 0 allows normal operation.

For example, a robot is about to push a box, but a human hand is detected in the box's path. Ethical Evaluation: E(A) = 0 (pushing the box is deemed unethical as it could harm the human). Result: the disruption signal D = 1 halts the robot's motors, preventing the action.

Challenges:

1. Defining ethical boundaries: the algorithm relies on an accurate ethical framework.
2. False positives: the algorithm might disrupt operation unnecessarily if it misinterprets a situation.

The second algorithm prevents the machine from taking actions that could lead to disruptions or system failures. Its goal is to ensure continuous, safe, and reliable operation, even in complex scenarios.

Mechanism:

1. Risk assessment: Evaluate the potential consequences of each action for system stability.

2. Preventive action: Modify or block actions that could lead to disruptions.

The algorithm can be expressed as:

$$D = \begin{cases} 1 & \text{if } R(A) = 1, \\ 0 & \text{if } R(A) = 0. \end{cases} \quad (2)$$

R(A) - risk assessment function, which returns 1 if the action risks a disruption and 0 otherwise, P - preventive action signal, where P = 1 blocks the action, and P = 0 allows normal operation.

For example, an autonomous drone is losing battery power but is programmed to complete a delivery. Risk assessment is R(A) = 1 (continuing the delivery risks a crash). Result is the preventive action signal P = 1 redirects the drone to a safe landing zone, ensuring system stability.

Challenges:

1. Balancing safety and performance: excessive caution may reduce efficiency.

2. Dynamic environments: the algorithm must adapt to rapidly changing conditions.

These two algorithms work in tandem to ensure ethical and reliable operation. The Disruption Algorithm ensures the machine halts unethical actions that could harm humans. The Continuity Algorithm ensures the machine avoids disruptions in its operations (fig. 2).

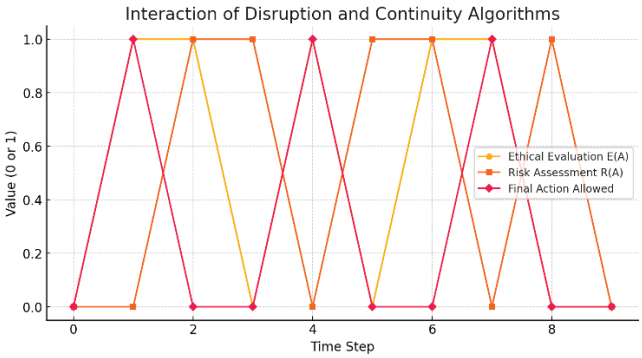


Fig. 2. System state change depending on the evaluation of action ethics (E(A)) and risk of failure (R(A)).

The combined decision-making can be modeled as:

$$\text{Action}_{\text{Allowed}} = \begin{cases} 1 & \text{if } E(A) = 1 \text{ and } S(A) = 1, \\ 0 & \text{otherwise.} \end{cases} \quad (3)$$

S(A) - safety evaluation function, where S(A) = 1 if the action ensures safety and 0 otherwise.

Practical implementation:

1. Sensors and data collection: both algorithms rely on sensors and data inputs for real-time ethical and risk assessments.

2. Machine learning models: neural networks or decision trees could be trained to recognize unethical behavior or potential risks.

3. Example workflow:

Input: A machine receives a command to perform an action.

Step 1: The action is assessed for ethicality (E(A)).

Step 2: The action is assessed for safety (S(A)).

Step 3: The machine executes the action only if E(A) = 1 and S(A) = 1.

These algorithms form a robust ethical safeguard for machines, promoting responsible behavior and operational reliability. Together, they embody principles of machine ethics and system resilience, paving the way for safe and trustworthy AI deployment (fig. 3).

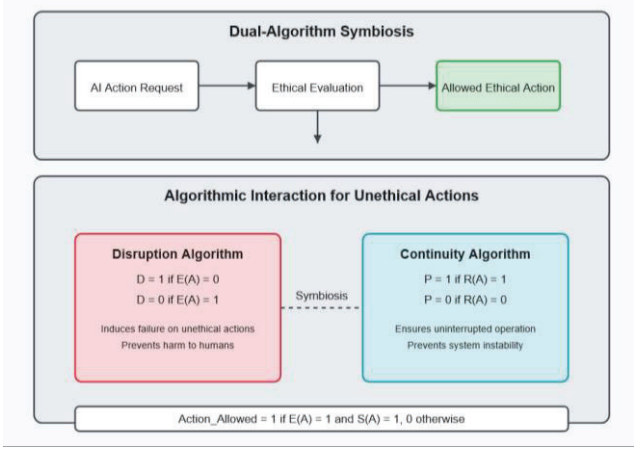


Fig. 3. AI ethical decision making system.

Implementing the symbiosis of the Disruption Algorithm and Continuity Algorithm in an unmanned vehicle (UV), such as an autonomous drone or self-driving car, involves integrating ethical and risk-aware decision-making processes. For example, an autonomous drone is tasked with

delivering medical supplies to a remote area. The drone must adhere to ethical principles to avoid harm and maintain continuous, reliable operation despite dynamic environmental challenges.

Core components:

1. Sensors: cameras, LiDAR, GPS, and proximity sensors to perceive the environment. Temperature, wind speed, and battery status monitors for internal assessments.
2. Disruption algorithm: ensures no unethical actions occur (e.g., colliding with people or private property).
3. Continuity algorithm: avoids actions that risk the drone's operation (e.g., flying into storms or running out of battery).
4. Controller: executes decisions based on algorithm outputs and adjusts the flight path dynamically.

For example, key objectives is deliver medical supplies without endangering humans, animals, or property. Ensure continuous operation and safe completion of the mission. The mission starts with a planned route optimized for time and safety. The drone evaluates:

1. Ethical compliance are any restricted zones (e.g., residential areas) on the route?
2. Operational risks is the weather forecast safe, and is the battery sufficient?

$$P(t) = \begin{cases} 1 & \text{if } E(t) = 1 \text{ and } R(t) < \text{threshold}, \\ 0 & \text{otherwise.} \end{cases} \quad (4)$$

R(t) - risk score at time t, E(t) - ethical compliance score at time t, P(t) - permission to proceed.

If P(t) = 0, the drone modifies the route or delays the mission.

During the flight, the drone continuously monitors its surroundings and internal state. Real-time ethical evaluation: the Disruption Algorithm evaluates potential actions for harm. Example: the drone detects a person standing in its landing zone. If landing could cause harm (E(A) = 0), the drone hovers or seeks an alternative spot. Real-time risk assessment: the Continuity Algorithm checks for operational risks. Example: if strong winds (R(A) = 1) threaten stability, the drone adjusts altitude or reroutes.

$$\text{Action}_{\text{Allowed}} = \begin{cases} 1 & \text{if } H(t) = 0 \text{ and } S(t) > \text{safe_threshold}, \\ 0 & \text{otherwise.} \end{cases} \quad (5)$$

S(t) - system stability at time t, H(t) - harm potential at time t.

Ethical decision example. The drone's direct path crosses a restricted wildlife area, where flying may disturb animals. The Disruption Algorithm identifies this action as unethical (E(A) = 0). The drone recalculates its path to avoid the restricted zone, even if it increases travel time. Outcome: the drone avoids unethical actions while maintaining operational safety.

Operational risk example. Mid-flight, the drone detects a low battery level. The Continuity Algorithm determines that continuing the delivery risks a crash (R(A) = 1). The drone prioritizes a safe landing at an intermediate charging station. Outcome: the mission is paused, but the system avoids catastrophic failure.

Emergency handling. In emergencies, the algorithms collaborate to choose the safest course of action. For example, a sudden obstacle (e.g., a bird) appears in the drone's path. The Disruption Algorithm prevents collision by halting forward motion (D = 1). The Continuity Algorithm ensures stability by recalibrating the drone's balance and redirecting the flight path.

Interaction between algorithms. The symbiosis ensures robust decision-making:

1. The Disruption Algorithm halts unethical actions immediately.
2. The Continuity Algorithm prevents actions that risk operational failure.

This interaction is represented as:

$$\text{Final_Action}(t) = \begin{cases} 1 & \text{if } E(t) = 1 \text{ and } S(t) > \text{safe_threshold}, \\ 0 & \text{otherwise.} \end{cases} \quad (6)$$

Ambiguity in ethical decisions. Challenge: deciding when harm is unavoidable (e.g., choosing between minor property damage and mission failure). Solution: use a weighted scoring system to balance ethical and operational priorities.

Adapting to dynamic environments. Challenge: rapidly changing weather or unforeseen obstacles. Solution: real-time updates from onboard sensors and external data sources.

Balancing efficiency and safety. Challenge: rerouting for safety may delay mission completion. Solution: optimize alternative routes to minimize delays while ensuring safety.

By integrating the two algorithms, the drone achieves:

1. Ethical compliance: it avoids actions that harm humans, animals, or property.
2. Operational stability: it completes its mission safely, even in adverse conditions.

This example demonstrates how the symbiosis of ethical and operational safeguards creates a resilient and responsible autonomous vehicle.

Research aims to contribute to the broader field of AI safety and ethics by providing a practical, implementable solution for ensuring ethical behavior in AI-powered machines. The proposed method has potential applications across various domains, from autonomous vehicles to industrial robots, where maintaining ethical boundaries is crucial for safe and responsible AI deployment.

V. Discussion of the Results

The symbiotic relationship between the disruption and continuity algorithms demonstrates remarkable effectiveness in maintaining ethical boundaries while ensuring operational stability. This approach's success lies in its biomimetic nature, mirroring human behavioral mechanisms of pain avoidance and operational continuity. The mathematical formalization of the system, particularly through the functions E(A) for ethical evaluation and R(A) for risk assessment, provides a robust framework that can be adapted across various AI applications.

The ability to make real-time ethical decisions while maintaining operational stability suggests that this approach could be valuable in high-stakes environments where both ethical compliance and system reliability are crucial.

This research extends beyond traditional rule-based ethical frameworks by introducing a dynamic, self-regulating mechanism. Unlike previous approaches that rely solely on predefined rules or utility functions, the dual-algorithm system creates an emergent ethical behavior pattern through the interaction of its components.

The models developed for this system demonstrate that ethical behavior can be effectively encoded through algorithmic symbiosis, providing a foundation for future research in AI safety mechanisms. The formalization of ethical evaluation and risk assessment functions offers a template for implementing similar systems across different domains.

The research findings have several important practical implications:

- The system's ability to handle complex, real-world scenarios suggests its potential applicability in critical applications such as healthcare, autonomous transportation, and industrial automation.
- The balance between ethical compliance and operational stability achieved through algorithmic symbiosis provides a model for developing robust AI systems that can operate reliably while maintaining ethical boundaries.

This research makes a significant contribution to the field of AI ethics and safety by providing a practical, implementable solution for ensuring ethical behavior in AI-powered machines. As AI technology continues to advance and integrate more deeply into society, such mechanisms will become increasingly crucial for ensuring safe and responsible AI deployment.

Conclusion

This research presents a novel approach to ensuring ethical behavior in AI-powered machines through a symbiotic dual-algorithm system. The proposed method successfully demonstrates how combining disruption and continuity algorithms can create an effective mechanism for maintaining ethical boundaries while ensuring operational stability.

The mathematical formalization of the dual-algorithm system has been developed, which provides a robust framework for practical implementation across various AI applications. The proven mathematical models demonstrate that the disruption algorithm effectively prevents unethical actions through the function E(A), the continuity algorithm maintains system stability through the risk assessment function R(A), the combined decision-making process successfully balances ethical compliance with operational reliability.

The implications of this research extend beyond theoretical frameworks, offering practical solutions for autonomous vehicle systems, industrial robotics, medical AI applications, smart city infrastructure, AI-powered decision support systems.

This research contributes significantly to the field of AI ethics by providing a practical, implementable solution for ensuring ethical behavior in AI-powered machines. The proposed dual-algorithm approach offers a promising foundation for developing more sophisticated ethical control systems in future AI applications. As AI technology continues to advance and integrate more deeply into society, such mechanisms will become increasingly crucial for ensuring safe and responsible AI deployment.

The ability of this method in maintaining ethical boundaries while ensuring operational stability suggests that biomimetic approaches to AI ethics control may offer valuable solutions for future challenges in AI safety and reliability.

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Improving the Efficiency of Data Routing in Wireless Sensor Networks Using a Genetic Algorithm with Dynamic Adaptation of Crossover and Mutation Probabilities

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***Abstract* — This paper focuses on data routing using a genetic algorithm for wireless sensor networks. A flowchart of the genetic algorithm (GA) is presented, incorporating adaptive probabilities for crossover and mutation, as well as genetic operators: tournament selection, ordered crossover, and shuffle mutation. The efficiency of the developed GA in route formation for a network topology with 25 nodes is investigated. Based on the results of simulation modeling, it is shown that the use of the GA allows for a route length that is 47.14% shorter than the one obtained with the classical greedy algorithm.**

***Keywords* — *routing, genetic algorithm, wireless sensor networks, dynamic adaptation of parameters*.**

I. INTRODUCTION

In modern wireless sensor networks (WSNs), efficient routing is an important task that determines the performance, energy efficiency, and lifetime of the network [1]. Due to the limited resources of sensor nodes and the dynamically changing topology of WSNs, traditional routing algorithms can be inefficient. This is especially true for large, scalable networks where it is necessary to optimise data transmission and minimise delays.

Genetic algorithms (GAs) are an effective approach to solving optimisation problems in wireless sensor networks [2-4]. They mimic the process of natural selection, using mechanisms such as crossover, mutation, and selection to find optimal solutions. The use of a genetic algorithm in WSNs allows to

- minimise the power consumption of nodes,

- reduce data transmission delays,

- increase the reliability and survivability of the network.

The relevance of the study is due to the need to develop efficient routing mechanisms that take into account the limited resources of WSNs and the dynamically changing network topology [5]. The genetic algorithm is a promising method for finding optimal data transmission routes, which ensures an increase in network performance.

The aim of this paper is to apply the developed GA to WSN routing to determine its effectiveness and potential advantages over the greedy algorithm.

II. FUNCTIONAL FEATURES OF THE DEVELOPED GENETIC ALGORITHM

The effectiveness of the GA depends largely on the type of genetic operator used and the value of the crossover and mutation probability [6-8]. To improve the convergence of the GA, it is proposed to use dynamic adjustment of the crossover probability [9, 10] based on the estimation of the fitness function for the set of routes $F(x)_{set_r}$ and the i-th route $F(x)_{r_i}$ in a certain generation. Thus, these two values are compared, on the basis of which the quality level of a particular individual is determined.

The dynamic change p_{cross} can be represented as:

$$p_{cross} = \begin{cases} \dfrac{a+(b-a)(F(x)_{\max_set_r} - F(x)_{r_i})}{F(x)_{\max_set_r} - F(x)_{aver}}, \text{at } F(x)_{r_i} \succ F(x)_{aver} \\ \dfrac{a+b}{2}, at\ F(x)_{r_i} \approx F(x)_{aver} \\ b, at\ F(x)_{r_i} \prec F(x)_{aver} \end{cases} \quad (1)$$

where a, b are numbers indicating the lower and upper bound of the given crossover probability, respectively; $F(x)_{\max_set_r}$ is maximum value of the fitness function of routes for a generation; $F(x)_{aver}$ is average value of the fitness function obtained for the current generation; $F(x)_{r_i}$ is value of the fitness function obtained by selecting the best value when considering two values of the selected individuals for crossing.

The principle of dynamic adaptation is based on the following:

- reducing p_{cross} the fitness function (values that correspond to the best routes based on a certain criterion) at high quality, as such routes already have certain positive characteristics, and crossing them can lead to their loss. In addition, crossing such routes is more likely to lead to the repetition of existing routes and reduce the diversity in the population. This solution is designed to minimise the deterioration of route characteristics by reducing the p_{cross}.

- with an average quality of the fitness function, the use of the arithmetic mean of the crossover probability range is considered, which allows balancing the process of finding the optimal route;

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- increase p_{cross} in the case of low quality of the fitness function, since such routes (values that correspond to the worst routes based on a certain criterion) have the potential for improvement. Thus, the increase p_{cross} speeds up the search for the optimal route.

A similar principle of dynamic adaptation is used for the value of the mutation probability p_{mut}. Based on (1), we can say that the value p_{mut} is smaller for more adapted individuals and larger for less adapted ones. The block diagram of the developed GA using dynamic adaptation of crossover and mutation probabilities is shown in Fig. 1.

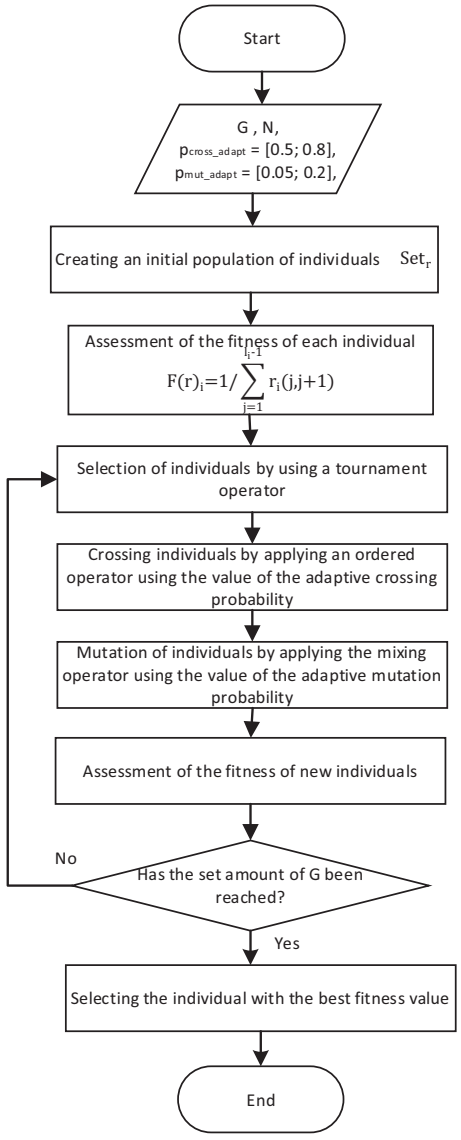


Fig. 1. Block diagram of the developed GA operation

After analysing the GA parameters of the scientific papers, the following ranges were selected for dynamic adjustment in accordance with (1): $p_{cross_adapt} = [0.5; 0.8]$, $p_{mut_adapt} = [0.05; 0.2]$. The choice of the types of genetic operators was based on the analysis of [11-14].

III. Simulation Modelling of the Algorithm

To verify the performance of the presented GA (Fig. 1), we performed a simulation by applying the developed software product in the Python programming language using the DEAP library. For the simulation, 25 sensor nodes were used, the location of which is set randomly, while the network is considered at a fixed point in time. It is assumed that if a node is more distant from another node than the specified range, then data transmission is impossible. In this regard, a penalty was used for such a case, the use of which encouraged the GA to choose routes with a shorter distance.

A laptop with the following specifications was used to run the application:

- Processor: AMD Ryzen 7, 5700U, 8 cores;
- RAM: 16 GB, DDR4, 3200 MHz.

For the simulation, 25 nodes were used, randomly placed in a free space on a 100×100 m area (Fig. 2). The range of each node is 30 m. If a node is farther away from another node, data transmission is impossible, which results in a distance penalty of 1000 m.

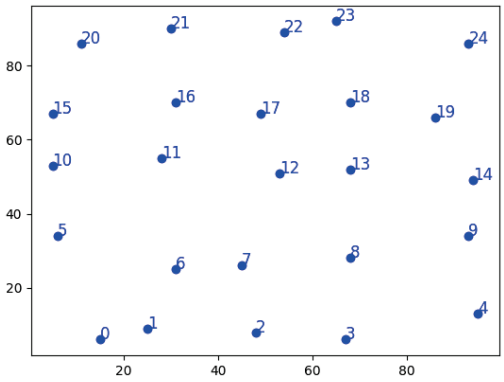


Fig. 2. Topology of the investigated network

The simulation of the process of establishing a route between nodes 4 and 20 using the Euclidean metric for the presented GA and the classical greedy algorithm was carried out.

To compare the performance of GAs, we chose the greedy algorithm (GrA), which is a powerful tool for route discovery due to its simplicity of implementation, computational speed, and low memory requirements [15, 16]. Each node decides on the next step based on the minimum distance to the next available node, which reduces the complexity of the algorithm and reduces the use of computing resources.

A GrA is characterised by local search, according to which nodes do not have information about the global network structure or all possible routes. They work with limited information, which reduces the complexity of the calculations performed.

The local nature of decision-making allows this algorithm to respond quickly to changes in the network, including the failure/addition of nodes. At the same time, there is no need to globally re-form the data transmission route. Each node can independently make new decisions based on up-to-date information about its neighbours, which has a positive impact on the overall reliability of the network. One of the main advantages of the greedy algorithm is its scalability, which makes it appropriate for use in large networks. Thus, this type of algorithm can be used for data

routing in networks with a dynamic variable topology, in particular for WSNs.

Fig. 3-4 shows the matrix of nodes of the considered network, which contains information about the relationships between sensor nodes when using GA and GrA.

GA Distance Matrix

	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24	25
1	0	10	1000	1000	1000	29	24	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
2	10	0	23	1000	1000	1000	17	26	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
3	1000	23	0	19	1000	1000	24	18	28	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
4	1000	1000	19	0	28	1000	1000	29	22	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
5	1000	1000	1000	28	0	1000	1000	1000	30	21	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
6	29	1000	1000	1000	1000	0	26	1000	1000	1000	19	30	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
7	24	17	24	1000	1000	26	0	14	1000	1000	1000	30	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
8	1000	26	18	29	1000	1000	14	0	23	1000	1000	1000	26	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
9	1000	1000	28	22	30	1000	1000	23	0	25	1000	1000	27	24	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
10	1000	1000	1000	1000	21	1000	1000	1000	25	0	1000	1000	1000	30	15	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
11	1000	1000	1000	1000	1000	19	1000	1000	1000	1000	0	23	1000	1000	1000	14	1000	1000	1000	1000	1000	1000	1000	1000	1000
12	1000	1000	1000	1000	1000	30	30	1000	1000	1000	23	0	25	1000	1000	25	15	24	1000	1000	1000	1000	1000	1000	1000
13	1000	1000	1000	1000	1000	1000	1000	26	27	1000	1000	25	0	15	1000	1000	29	16	24	1000	1000	1000	1000	1000	1000
14	1000	1000	1000	1000	1000	1000	1000	1000	24	30	1000	1000	15	0	26	1000	1000	24	18	22	1000	1000	1000	1000	1000
15	1000	1000	1000	1000	1000	1000	1000	1000	1000	15	1000	1000	1000	26	0	1000	1000	1000	1000	18	1000	1000	1000	1000	1000
16	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	14	25	1000	1000	1000	0	26	1000	1000	1000	19	1000	1000	1000	1000
17	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	15	29	1000	1000	26	0	18	1000	1000	25	20	29	1000	1000
18	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	24	16	24	1000	1000	18	0	19	1000	1000	29	22	29	1000
19	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	24	18	1000	1000	1000	19	0	18	1000	1000	23	22	29
20	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	22	18	1000	1000	1000	18	0	1000	1000	1000	1000	21
21	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	19	25	1000	1000	1000	0	19	1000	1000	1000
22	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	20	29	1000	1000	19	0	24	1000	1000
23	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	29	22	23	1000	1000	24	0	11	1000
24	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	29	22	1000	1000	1000	11	0	28
25	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	29	21	1000	1000	1000	28	0

Fig. 3. Distance matrix for GA

Greedy Distance Matrix

	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24	25
1	0	10	1000	1000	1000	29	24	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
2	10	0	23	1000	1000	1000	17	26	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
3	1000	23	0	19	1000	1000	24	18	28	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
4	1000	1000	19	0	28	1000	1000	29	22	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
5	1000	1000	1000	28	0	1000	1000	1000	30	21	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
6	29	1000	1000	1000	1000	0	26	1000	1000	1000	19	30	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
7	24	17	24	1000	1000	26	0	14	1000	1000	1000	30	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
8	1000	26	18	29	1000	1000	14	0	23	1000	1000	1000	26	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
9	1000	1000	28	22	30	1000	1000	23	0	25	1000	1000	27	24	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
10	1000	1000	1000	1000	21	1000	1000	1000	25	0	1000	1000	1000	30	15	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000
11	1000	1000	1000	1000	1000	19	1000	1000	1000	1000	0	23	1000	1000	1000	14	1000	1000	1000	1000	1000	1000	1000	1000	1000
12	1000	1000	1000	1000	1000	30	30	1000	1000	1000	23	0	25	1000	1000	25	15	24	1000	1000	1000	1000	1000	1000	1000
13	1000	1000	1000	1000	1000	1000	1000	26	27	1000	1000	25	0	15	1000	1000	29	16	24	1000	1000	1000	1000	1000	1000
14	1000	1000	1000	1000	1000	1000	1000	1000	24	30	1000	1000	15	0	26	1000	1000	24	18	22	1000	1000	1000	1000	1000
15	1000	1000	1000	1000	1000	1000	1000	1000	1000	15	1000	1000	1000	26	0	1000	1000	1000	1000	18	1000	1000	1000	1000	1000
16	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	14	25	1000	1000	1000	0	26	1000	1000	1000	19	1000	1000	1000	1000
17	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	15	29	1000	1000	26	0	18	1000	1000	25	20	29	1000	1000
18	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	24	16	24	1000	1000	18	0	19	1000	1000	29	22	29	1000
19	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	24	18	1000	1000	1000	19	0	18	1000	1000	23	22	29
20	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	22	18	1000	1000	1000	18	0	1000	1000	1000	1000	21
21	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	19	25	1000	1000	1000	0	19	1000	1000	1000
22	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	20	29	1000	1000	19	0	24	1000	1000
23	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	29	22	23	1000	1000	24	0	11	1000
24	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	29	22	1000	1000	1000	11	0	28
25	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	1000	29	21	1000	1000	1000	28	0

Fig. 4. Distance matrix for GrA

Fig. 5-6 shows the generated data transmission route between the considered nodes.

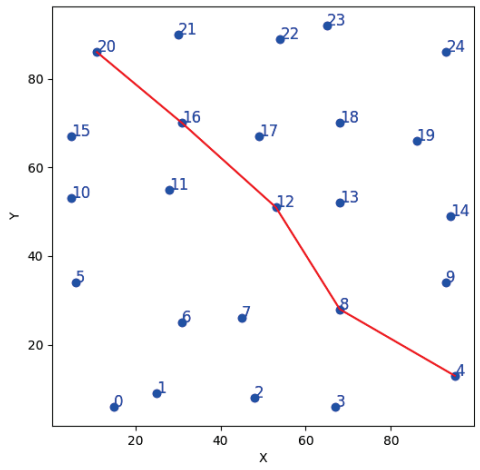


Fig. 5. The resulting route when using GA

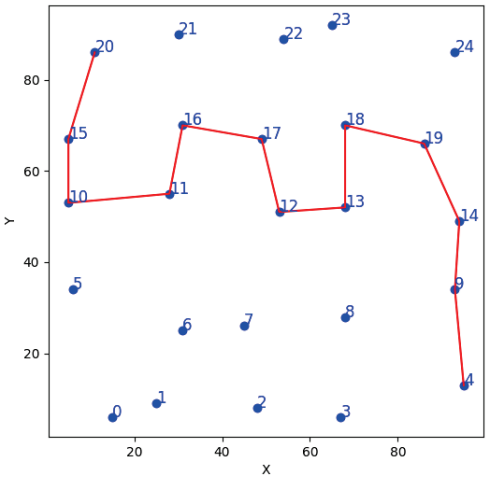


Fig. 6. The resulting route when using GrA

The results of the simulation of the considered algorithms are presented in Table 1.

TABLE I. COMPARISON OF THE GENERATED ROUTES USING THE DEVELOPED GENETIC ALGORITHM AND THE CLASSICAL GREEDY ALGORITHM FOR A NETWORK OF 25 NODES WITH THE SAME RANGE

Algorithm	Generated route with node numbers	Length of the route, metres
Genetic	[4, 8, 12, 16, 20]	111
Greedy	[4, 9, 14, 19, 18, 13, 12, 17, 16, 11, 10, 15, 20]	210

Thus, the use of the developed GA for data routing is more efficient in finding the shortest path compared to the GrA.

CONCLUSIONS

This paper presents a developed genetic algorithm with dynamic adaptation of crossover and mutation probabilities for data routing in wireless sensor networks. A simulation study was conducted to compare the performance of the genetic algorithm (GA) and the greedy algorithm in route discovery for a network topology consisting of 25 sensors with identical transmission ranges. The results demonstrate that the proposed GA identifies a route that is 47.14% shorter than that obtained using the greedy algorithm, while the number of hops is reduced by a factor of three.

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Computational Efficiency of an Artificial Neural Network in Calculating the Rresonant Frequencies of Patch Antennas

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***Abstract*—This paper explores the impact of artificial neural network architectural and training parameters on the accuracy and convergence of resonance frequency prediction for rectangular patch antennas. A series of experiments were conducted where key parameters, such as the number of hidden layers, neurons per layer, batch size, and optimizer, were systematically varied. The goal is to determine which configurations provide the best balance between prediction accuracy and training efficiency. The results demonstrate that neural networks can effectively approximate the resonant frequency based on physical antenna dimensions and material properties, while also highlighting configurations that either hinder or improve generalization. These findings contribute to the optimization of machine learning models for electromagnetic structure analysis.**

Keywords—Neural network, patch antenna, resonant frequency, hidden layer, mean squared error.

I. Introduction

The design of antennas with desired performance characteristics represents a complex, multi-objective optimization task, as it depends on numerous interrelated parameters, including geometry, substrate materials, dielectric properties, and feeding techniques. Classical analytical approaches such as the transmission line model, well-documented in foundational studies [1]-[4], provide accurate but often computationally expensive methods that do not scale well with increasing model complexity.

In recent years, artificial intelligence (AI) has emerged as a powerful tool in microwave and antenna engineering, enabling data-driven predictions that bypass the need for direct analytical modeling. Artificial neural networks (ANNs), in particular, have demonstrated substantial potential in estimating the resonant frequency of patch antennas based on structural parameters [5]-[8]. Additionally, ANNs have proven effective in electromagnetic design automation and the modeling of microwave components, highlighting their ability to discover nonlinear dependencies within complex systems [9]-[11].

Despite these advances, many published works prioritize prediction accuracy without deeply analyzing how internal parameters of ANNs impact convergence speed, model generalization, and computational efficiency. These parameters are the number of hidden layers, activation functions, optimizer types, and training configurations. Only a limited number of studies, such as [12], discuss architectural design choices, and even then without detailed experimentation. The present study seeks to fill this gap by systematically examining how changes in network architecture affect the performance of a model that predicts the resonant frequency.

II. Setup, Experimental Model and Determination of the Reference Construct

In order to conduct a full-fledged study of the ANN's efficiency in calculating the resonant frequency of a rectangular microstrip patch antenna that presented on Fig. 1, a stack of modern tools was chosen, allowing flexible control over all stages, from data generation to analysis of training results.

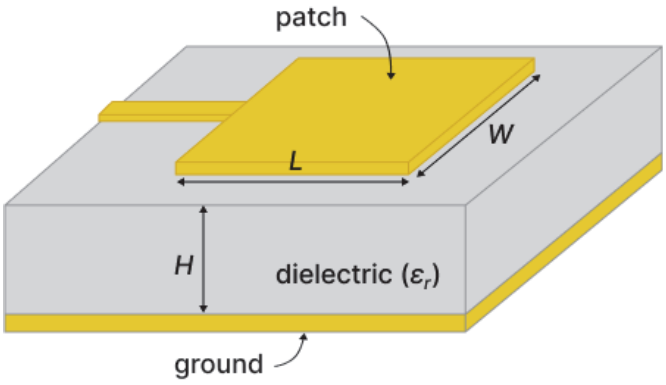


Fig. 1. Schema of a patch antenna.

The Python programming language is selected due to its versatility, simple syntax, and large ecosystem of scientific libraries. In particular, Python is the de facto standard in the field of machine learning, thanks to libraries such as TensorFlow, PyTorch, Keras, Sci-kit Learn, and others. Additionally, using Python made it easy to combine numerical modeling with the construction of neural networks.

The PyCharm Professional integrated development environment (IDE) was chosen, running on the macOS operating system with an Apple Silicon processor. This environment provides flexible management of environments (venv, conda), support for libraries with autocompletion, and convenient visual control over code, graphs, and training logs, which is especially important when conducting numerous experiments.

The foundation of machine learning is TensorFlow 2.x, which is a modern, powerful framework from Google supporting both low-level programming and high-level abstractions through the integrated Keras API. The Keras is used to build and train models, as it provides the ability to quickly create neural network architectures using either sequential or functional approaches, and allows for easy

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modification of hyperparameters and experimentation with the architecture.

For reproducibility, all experiments were carried out in Python 3.11 with TensorFlow 2.15 on macOS running an Apple M1 Pro processor with 16 GB of RAM. The reference model requires approximately 11 s to complete 100 epochs. Random seeds are fixed (42) for both dataset generation and weight initialization, and training is repeated three times to ensure stability of the reported results.

For creating the training dataset, the data-generator is implemented using C# language. This generator is implemented as a classical physical model for calculating the resonant frequency f of a patch antenna. A rectangular microstrip antenna has a rectangular shape with dimensions L and W, representing the length and width of the antenna, respectively. A conducting surface acts as a ground plane and is placed beneath a substrate of thickness H and relative dielectric permittivity ε_r. The frequency f_{mn} at which the antenna resonates can be calculated as [13], [14]:

$$f_{mn} = \frac{c}{2\sqrt{\varepsilon_{\text{eff}}}}\left[\left(\frac{m}{L_{\text{eff}}}\right)^2 + \left(\frac{n}{W_{\text{eff}}}\right)^2\right]^{1/2}, \quad (1)$$

where ε_{eff} is the effective relative permittivity of the substrate, c is the speed of electromagnetic waves in free space, m and n are integers defining the mode, and L_{eff} and W_{eff} are the effective dimensions. The patch width directly influences some important antenna parameters, such as the radiation pattern and the frequency at which the antenna is likely to resonate. The resonant frequency f_{mn} of the antenna in its fundamental TM_{10} mode can be calculated as:

$$f^{10} = \frac{c}{2L_{\text{eff}}\sqrt{\varepsilon_{\text{eff}}}}, \quad (2)$$

where $L_{\text{eff}} = L + 2\Delta L$.

The expressions for the effective dielectric constant and the length extension of the antenna substrate are derived from empirical models that are widely used in the antenna theory. In particular, the effective permittivity can be approximated as

$$\varepsilon_{\text{eff}}(W) = \frac{\varepsilon_r+1}{2} + \frac{\varepsilon_r-1}{2}\frac{1}{\sqrt{1+10h/W}}, \quad (3)$$

and the length extension is given by

$$\Delta L = 0.412h\frac{(\varepsilon_e(W)+0.300)(W/h+0.264)}{(\varepsilon_e(W)-0.258)(W/h+0.813)}. \quad (4)$$

The input parameters for ANNs are thickness of the substrate (H), its relative dielectric permittivity (ε_r), effective width (W_{eff}) and length (L_{eff}).

Equations (2)-(4) summarize the standard closed-form description of a rectangular microstrip patch antenna operating in its fundamental mode. These expressions are widely used in the antenna design as they provide a reliable first-order approximation of resonant frequency behavior without requiring full-wave simulations. It should be emphasized that although such equations do not capture all higher-order effects, they remain sufficiently accurate to serve as a reference for generating synthetic datasets in machine learning experiments [15]. In the future, for more accurate calculation of antenna characteristics, it is proposed to use specialized software for electromagnetic modeling (e.g., CST, Ansys HFSS).

Both the training and validation datasets consisted of 2000 examples each, generated from the same model but with different randomized input values to ensure statistical independence. Data are split within the training set using an 80/20 training/validation partition (validation_split = 0.2) with shuffling enabled each epoch. Models are trained with the Adam optimizer and MSE loss. Early stopping monitored val_loss with patience = 10 and restore_best_weights = True. Unless otherwise stated, the batch size is 64 for the forward model and 128 for the inverse model, with 100 and 300 epochs respectively as upper bounds.

Values of parameters of the reference 3 layer perceptron are presented at Table I.

TABLE I. REFERENCE PARAMETERS VALUES

Number of the layers	Input layer, 3 hidden layers, output layer
Activation function	ReLU
Optimizer	Adam
Loss-function	Mean Squared Error
Number of epochs	100
Batch size	128
EarlyStopping patience	10

The basic architecture of the ANN includes three fully connected (dense) hidden layers with 1024, 512, and 256 neurons, respectively (Fig. 2).

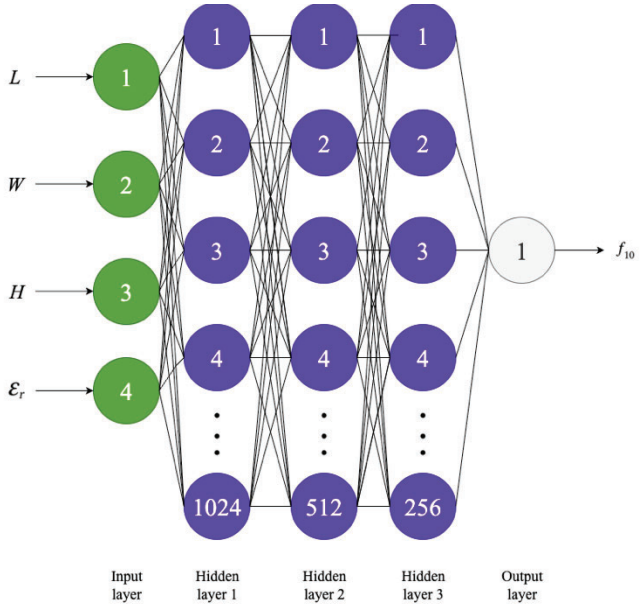


Fig. 2. Schema of the 3-layer perceptron used for the patch antenna optimization.

III. RESULTS AND DISCUSSIONS

The experiments revealed clear trends regarding how architectural and training parameters impact the ANN's ability to predict the resonant frequency of three hidden layers ([1024, 512, 256] neurons), ReLU activations, the Adam optimizer, and trained for 100 epochs with a batch size of 128, demonstrated stable convergence and low error rates, achieving a training error of 0.0100 and a validation error of 0.0107 in just over 11 seconds. The plot of convergence is

presented in Fig. 3, as well as a difference between expected value and actual value per iteration.

Reducing the number of hidden layers negatively impacts the learning capacity [6, 12], leading to a rise in both training and validation errors (to 0.0325 and 0.0378 respectively), suggesting the model is no longer sufficiently expressive. On the other hand, increasing the depth by adding a fourth layer slightly improved generalization, where validation error dropped to 0.0105, but with a noticeable cost in training time (14.9 seconds).

The experiments also explore different activation functions [16]. While ReLU consistently produces accurate and stable results, alternatives such as sigmoid and tanh functions underperformed. The use of a sigmoid function results in complete training failure, with exploding errors (174.8 and 186.4) and NaN convergence behavior. The tanh function yields more tolerable results, but still lagged behind ReLU, especially in terms of training stability and convergence speed.

Optimizer choice significantly affects model behavior. The Adam proved to be the most effective, balancing speed and accuracy. The RMSprop, while functional, led to higher errors (7.24 training, 5.36 validation), and stochastic gradient descent (SGD) failed to train the model entirely. This highlights the importance of adaptive learning rate methods in complex regression tasks.

Further experiments showed that halving the number of training epochs (from 100 to 50) or reducing the batch size (from 128 to 64) could compromise performance. Shorter training leads to a slight drop in accuracy, while a smaller batch size causes a more noticeable rise in training error (0.0231). Thus, moderately sized batches and sufficient training time appear essential for achieving consistent results.

Beyond error values, the experiments also reveal clear differences in convergence speed and computational cost. The baseline configuration required approximately 11.3 s for 100 epochs, while reducing the number of hidden layers to two slightly lowered runtime (10.6 s) but at the expense of much higher errors. Increasing depth to four layers improved validation accuracy, yet training time rose to almost 15 s, reflecting the expected trade-off between accuracy and efficiency. Activation functions had an even stronger effect: tanh maintained convergence but is significantly slower (13.8 s), whereas sigmoid completely destabilized training despite a short runtime of 6.1 s. Optimizer choice is equally decisive. RMSprop converged in only 5.6 s but with high errors, while SGD failed to train at all. Varying the number of epochs confirmed that shorter training (50 epochs) reduced runtime to 8.4 s with only minor accuracy loss, suggesting potential for efficiency gains when early stopping is applied. Finally, halving the batch size to 64 further decreased training time to 8.1 s and accelerated early stopping, confirming that hyperparameters jointly determine not only accuracy but also computational efficiency.

It is also worth noting that the smaller batch size introduced a stochastic effect into the optimization process, which slightly degraded the training accuracy but helped the model to avoid overfitting and converge faster in validation. This balance between stability and generalization is typical in deep learning and confirms that hyperparameter tuning must be problem-specific

These findings suggest that the best-performing configuration in this task balances architectural complexity and training effort. Too shallow or too deep networks, inappropriate activations, or suboptimal optimizers can significantly degrade performance. The combined use of ReLU and Adam, within a three-layered network, offers a sweet spot between performance and efficiency.

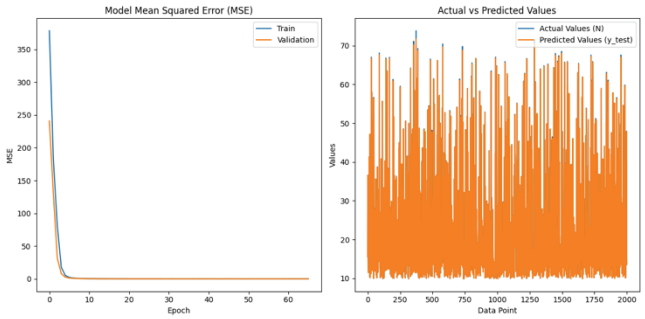


Fig. 3. Plot of the convergence speed at reference model.

A notable limitation of the present study is that the dataset used for training and validation was generated from closed-form analytical equations. As a result, the neural network essentially learns a deterministic mapping that is free from measurement noise, fabrication tolerances, or the numerical artifacts that typically arise in full-wave electromagnetic simulations. This makes the learning problem easier and may overestimate the model's generalization ability compared to real-world scenarios. Moreover, only rectangular patch antennas in the fundamental mode were considered, which restricts the applicability of the findings to more complex antenna geometries and higher-order modes. The experiments were also limited to feedforward perceptron architectures, leaving unexplored other model families that might better capture non-linear dependencies or improve computational efficiency.

Furthermore, robustness against noisy or incomplete data has not been assessed. In practical scenarios, imperfections in measurements or simulation outputs can significantly affect model behavior, and accounting for such factors remains outside the scope of this work. Another open question is the sensitivity of the trained models to variations in the input distribution, since the current dataset was generated under idealized and uniform sampling assumptions.

IV. Conclusions

Throughout the conducted series of experiments, we investigated how specific architectural and training parameters influence the predictive accuracy and learning dynamics of an ANN tasked with estimating the resonant frequency of a rectangular patch antenna. The study covered variations in the number of hidden layers, activation functions, optimizers, training duration (epochs), and batch sizes.

Reducing the number of hidden layers from three to two led to a considerable increase in both training and validation error, highlighting the network's reduced capacity to model the non-linear dependencies between the inputs and the target frequency. Conversely, increasing the number of hidden layers to four slightly improved generalization, albeit at the cost of increased training time. Thus, the three-layer architecture was confirmed as a solid baseline for this task.

Activation functions also played a critical role. The ReLU function demonstrated superior performance in terms of convergence and final error values. In contrast, the sigmoid activation failed to train the model effectively, resulting in rapidly diverging loss values. The tanh activation provided moderate results, suggesting some utility but not outperforming ReLU in this case.

The optimizer choice significantly affected learning behavior. While the Adam optimizer consistently led to stable training and lower errors, switching to stochastic gradient descent (SGD) caused the model to completely fail, with training loss quickly becoming undefined (NaN). RMSProp showed better stability than SGD but still underperformed compared to Adam.

Shortening the number of training epochs from 100 to 50 slightly degraded model performance and slowed convergence. This result implies that longer training allows the network to better capture complex relationships in the data, provided that overfitting is properly controlled (e.g., via early stopping).

Finally, changing the batch size from 128 to 64 slightly increased the training error but improved validation performance and convergence speed, suggesting that smaller batches can enhance generalization through the stochasticity introduced into the optimization process.

In conclusion, the experiments confirmed that tuning neural network parameters has a measurable effect on both training efficiency and predictive quality. The findings underscore the importance of a deliberate and data-driven approach to neural network design, especially in technical applications like antenna modeling, where precision is paramount.

V. Future Work

Future research will primarily focus on addressing the inverse problem related to predicting the geometric and material parameters of a patch antenna from a specified resonant frequency. This task is inherently more challenging than forward modeling, as it is ill-posed and may admit multiple feasible solutions, making regularization and prior knowledge essential. Neural network architectures capable of handling such ambiguity, including encoder-decoder frameworks, variational approaches, or physics-informed neural networks (PINNs), will be investigated. In addition, incorporating data from full-wave electromagnetic solvers (e.g., CST, Ansys HFSS) and measured prototypes will allow the models to generalize beyond idealized closed-form equations. Transfer learning strategies may be used to adapt models trained on synthetic data to experimental conditions. Finally, extending the framework towards antenna array synthesis and optimization would demonstrate the broader applicability of the approach in practical microwave engineering.

Another promising direction is the exploration of convolutional or transformer-based models for handling spatial field distributions and antenna arrays. Such architectures may enable direct learning from electromagnetic field maps rather than only scalar parameters, broadening the applicability of the proposed framework

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YOLOv8 based model for Ukrainian license plate recognition

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***Abstract*—This work addresses the problem of automatic license plate recognition using computer vision and machine learning methods. An analysis of existing approaches to Ukrainian's license plate recognition was conducted, an optimal neural network architecture was selected, and a software solution was implemented. The proposed system employs modern deep learning algorithms based on YOLOv8 for license plate detection in images and subsequent processing. Image preprocessing, license plate segmentation, and optical character recognition of characters have been implemented. The model was trained using a dataset containing real vehicle images. The proposed solution can be applied in automated parking control systems, road safety enforcement, and law enforcement agencies.**

Keywords—automatic license plate recognition (ALPR), machine learning, neural networks, deep learning, YOLOv8, computer vision, optical character recognition (OCR)

I. Introduction

Automatic license plate recognition (ALPR) is a crucial technology in modern intelligent transportation systems, used for traffic management, law enforcement, and automated parking control. The ability to accurately and efficiently recognize license plates from images is essential for ensuring security and enhancing automation in various domains [1].

The global automatic number plate recognition system market size is calculated at USD 3.79 billion in 2024 and is predicted to reach around USD 9.27 billion by 2034 (Fig. 1) [2].

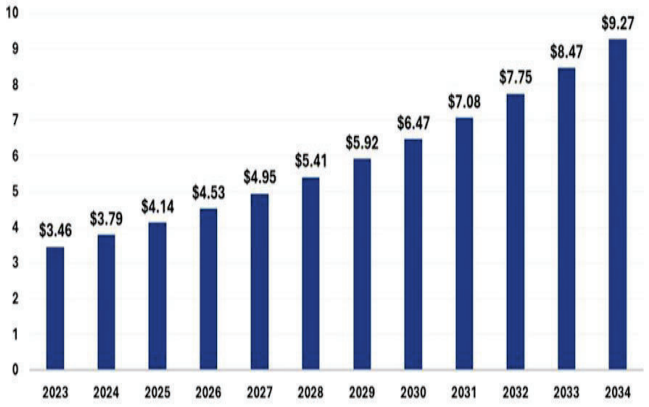


Fig. 1. Automatic Number Plate Recognition System Market Size 2024 to 2034 (USD Billion)

The objective of this work is to develop a robust and efficient license plate recognition system using computer vision and deep learning techniques. The proposed approach involves detecting license plates in images, preprocessing them, segmenting characters, and applying Optical Character Recognition (OCR) for text extraction [3, 4].

This study employs deep learning-based object detection models and OCR algorithms to achieve high recognition accuracy. The dataset consists of real-world images of vehicles with license plates, enabling the model to generalize well to different conditions.

II. Characterization of the License Plate Recognition Problem and Solution Methods

A. Issues and Challenges

Automatic License Plate Recognition is a crucial technology widely used in security, transportation, and parking automation [5]. It enables the automatic identification of vehicles based on their license plates, utilizing advanced computer vision techniques and deep learning models. ALPR plays a significant role in traffic management, law enforcement, and intelligent transportation systems by providing accurate and efficient vehicle recognition in real-time [6].

The main challenges in license plate recognition include:

- Varied fonts and styles of license plates across different countries.
- Impact of weather conditions and lighting on image quality.
- Partial occlusion of the license plate by other objects.
- Partial occlusion of the license plate by other objects.

B. General Characteristics of Ukrainian License Plates

There are many types of different car license plates in Ukraine. They are all defined in the State Standard of DSTU 4278: 2019 [7]. The license plate is the car ID and the person who owns this car. The car license plate of any type has the appropriate characteristics and designated areas. The most common type of license plates – type 1 (Fig. 2), it has the following characteristics: height – 112 mm; length – 520 mm; number of digits – 4; number of letters – 4; the color of the

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signs – white. The color of the numbers, letters and the ribs of the rigidity of the signs – black.



Fig. 2. The Ukrainian car license plate of type 1

The first two letters in the car license plate indicate the administrative and territorial affiliation of the registration of vehicles in Ukraine (Fig. 3).



Fig. 3. Compounds of letters and numbers denoting the administrative-territorial affiliation of the license plates of vehicles on the map of Ukraine

C. *License Plate Recognition Methods*

Modern ALPR methods are based on the following approaches:

- Traditional computer vision methods: edge detection, character segmentation, threshold binarization methods, and MORPH operations [8, 9].
- Deep learning models: the use of convolutional neural networks (CNN), recurrent neural networks (RNN), and transformers for image processing [10, 11].
- Optical Character Recognition (OCR): Tesseract OCR and neural networks for extracting text from license plates [12, 13].
- Object Detection Models (YOLOv8): YOLO (You Only Look Once) is a state-of-the-art object detection model that provides real-time performance with high accuracy. The latest version, YOLOv8, improves upon previous versions by incorporating advanced features like anchor-free detection, adaptive computation, and transformer-based attention mechanisms [14, 15].

D. *YOLOv8 in License Plate Recognition*

YOLOv8 is particularly well-suited for automatic license plate recognition (ALPR) due to its high detection accuracy, real-time performance, and adaptability to different conditions. The model follows a two-stage pipeline, combining object detection and optical character recognition (OCR).

Logical Structure of YOLOv8 for ALPR.

YOLOv8 operates in two main stages:

- License Plate Detection – the model is trained to detect the bounding box of a license plate in an image. Thanks to its efficient one-stage detection pipeline, it can process multiple frames per second, making it ideal for real-time applications.
- Character Recognition Integration – after detecting the license plate, the extracted region is processed by an OCR model (such as Tesseract OCR or a CNN-based classifier) to recognize the individual characters.

In the YOLOv8 architecture the backbone, neck, and head of the network, which work together to optimize object detection speed and precision [16].

Advantages of YOLOv8 for ALPR:

- Fast and accurate detection, even in real-time applications.
- Robust performance under varying lighting conditions, angles, and occlusions.
- Optimized architecture that balances speed and precision, making it suitable for deployment on edge devices like surveillance cameras.
- Support for transfer learning, allowing fine-tuning on domain-specific ALPR datasets for improved accuracy.

With its efficient structure and real-time capabilities, YOLOv8 provides a reliable solution for automatic license plate recognition, making it highly applicable in traffic surveillance, parking management, and intelligent transportation systems.

III. Result and Discussion

In this project, methods for automatic license plate recognition were examined and implemented using computer vision and deep learning. Specifically, the study focused on the application of convolutional neural networks (CNN), YOLOv8, and optical character recognition (OCR) to develop a high-accuracy license plate recognition algorithm.

The algorithms were implemented using Python 3.9 in the Google Colab development environment, utilizing libraries such as OpenCV, NumPy, Sklearn, and Tesseract OCR. The model execution was carried out in Google Colab with a GP T4 GPU, ensuring high performance and the ability to process large volumes of data in real-time.

The accuracy evaluation of the algorithm was conducted using the following metrics [17]:

- Accuracy – the ratio of correctly recognized license plates to the total number of samples.
- Precision – the proportion of correctly recognized license plates among all detected ones.
- Recall – the proportion of correctly detected license plates among all existing plates in the images.
- F1-score – the harmonic means between precision and recall.
- mAP@0.5 – the mean average precision of an object detection model when the Intersection over Union (IoU) threshold is set to 0.5. In ALPR, mAP@0.5 assesses how well the model detects license plates by evaluating the overlap between predicted and actual plate locations [18].

A. *Data Preparation*

A set of data "License Plate Recognition" using dates from ROBOFLOW [19] and AUTO.RIA [20] websites and

containing 10,000 images of license plates of vehicles that was used for training and testing algorithms was formed to perform the study.

This dataset is structured into three subsets:

- Training set – used to train the model.
- Validation set – used for hyperparameter tuning and evaluating the model's generalization ability.
- Test set – used for the final assessment of model accuracy on unseen data.

The data was split in a ratio of 80% (train) / 10% (validation) / 10% (test).

B. Model Training

To recognize license plates, the YOLOv8 model was trained on the "License Plate Recognition" dataset containing 8,000 images. Training was conducted in Google Colab using a GP T4 GPU over 30 epochs.

Training Process:

- Data Preprocessing – image normalization, augmentation (brightness adjustment, contrast enhancement, rotation).
- YOLOv8 Training – optimization of bounding boxes for precise license plate localization.
- Performance Evaluation – analysis of loss functions, precision, recall, F1-score and mAP.
- Testing on New Images – verifying the model's accuracy in real-world conditions.

Training Results:

- Loss functions (box_loss, cls_loss, dfl_loss) decreased steadily, indicating effective learning.
- Precision: 1.00, Recall: 0.97, F1-score: 0.96, demonstrating high accuracy.
- mAP@0.5 = 97.5%, meaning the model successfully detects license plates in most cases.
- The Precision-Recall curve confirms minimal false detections.

Yolov8 Precision-Recall Curve is presented in Fig. 5.

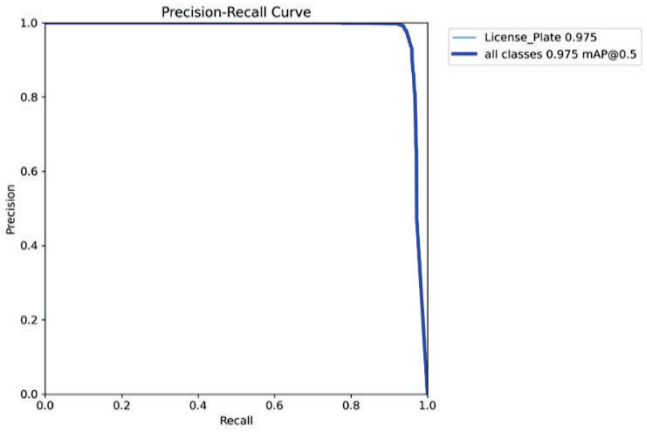


Fig. 5. YOLOv8 Precision-Recall Curve

Yolov8 F1-Confidence Curve are presented in Fig. 6.

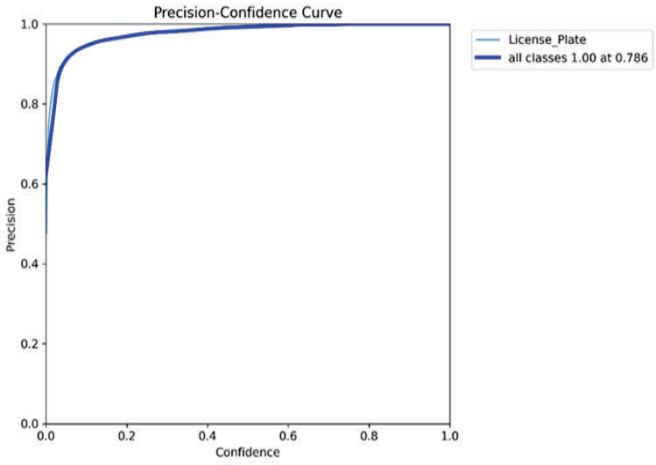


Fig. 6. YOLOv8 F1-Confidence Curve

Yolov8 Training Graphs are presented in Fig. 7.

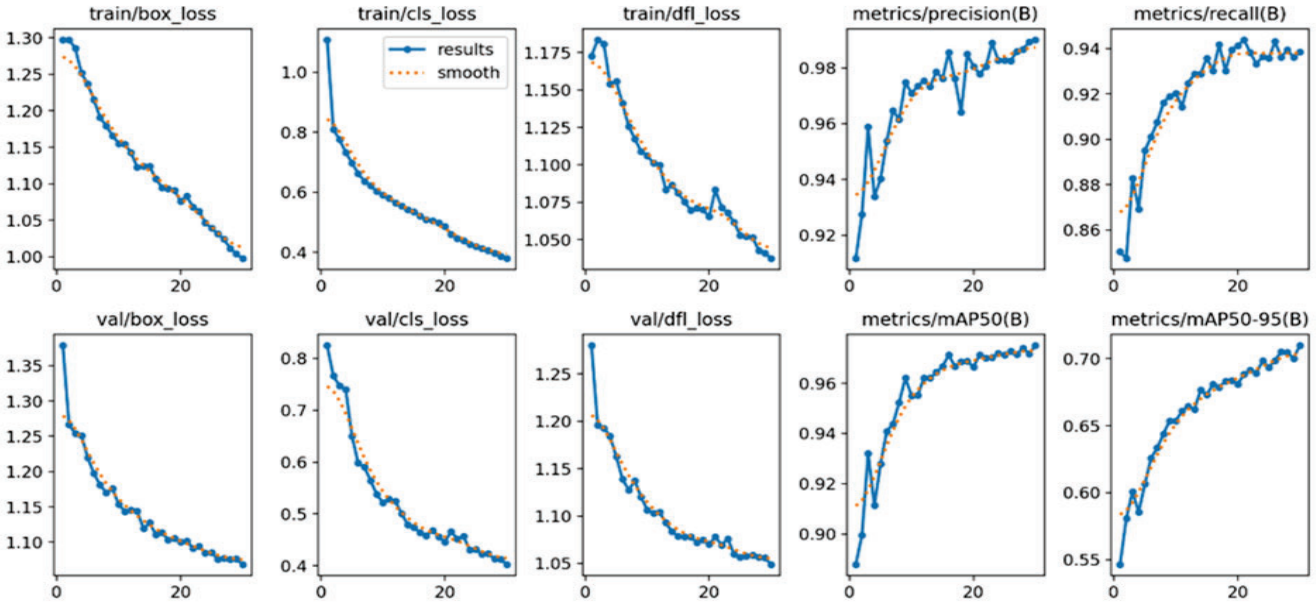


Fig. 7. YOLOv8 Training Graphs (Loss Reduction, Increasing Precision and Recall)

Training Image Examples are presented in Fig. 8.



Fig. 8. Training Image Examples (Batch 0)

The results of training the YOLOv8 model demonstrate its high efficiency in detecting license plates under various conditions. The model successfully achieved 97.5% accuracy (mAP@0.5), confirming its ability to precisely locate and identify license plates in images.

To ensure the robustness of the YOLOv8 model, both validation and test datasets were used for additional evaluation. The performance metrics, including precision and confidence scores, were computed for all datasets. The results are summarized in Table 1.

TABLE I. PERFORMANCE METRICS OF YOLOV8 ON TRAINING, VALIDATION, AND TEST SETS

Dataset	Precision	Recall	F1-score	mAP@0.5
Training	1,00	0,978	0,967	0,975
Validation	0,98	0,97	0,962	0,97
Test	0,975	0,968	0,958	0,96

C. *License Plate Recognition Algorithm*

The license plate recognition algorithm consists of two main stages:

- License plate detection – using YOLOv8 to locate the license plate in the image.
- Character recognition – applying OCR (Tesseract or CRNN) to extract text from the detected plate.

The general workflow of the license plate recognition algorithm is illustrated in Fig. 9.

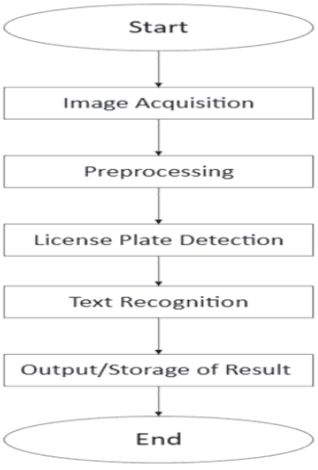


Fig. 1. Flowchart representing the license plate recognition process

Algorithm Steps

1) Image Acquisition

a) The system receives an input image or video frame.

b) The image is resized to 640×640 px for uniform processing.

2) Preprocessing

a) Noise reduction (Gaussian blur, median filtering).

b) Contrast normalization to enhance visibility.

3) License Plate Detection (YOLOv8)

a) Identifies the bounding box of the license plate.

b) Selects the region with the highest confidence score.

c) Extracts the detected plate area for further processing.

4) Text Recognition (OCR)

a) Applies Tesseract OCR or CRNN to recognize characters.

b) Post-processing:

- Removes unnecessary symbols.
- Formats text to match license plate standards (AA 1234 BB).
- Corrects OCR errors (O → 0, I → 1).

5) Output & Integration

a) Displays or stores the recognized license plate in a database.

b) Can be integrated into parking systems, traffic monitoring, and vehicle identification applications.

IV. Conclusions

This study explored and implemented an automatic license plate recognition algorithm based on deep learning. YOLOv8 was used for license plate detection, while OCR (Tesseract/CRNN) was applied for character recognition. The YOLOv8 model achieved high accuracy (mAP@0.5 = 97.5%) in detecting license plates, confirming its efficiency. The optical character recognition (OCR) system ensures correct text extraction, with additional post-processing to correct potential recognition errors. Preprocessing techniques, including contrast normalization and noise filtering, significantly improved detection and recognition quality.

Despite its high accuracy, the algorithm presents some challenges. Blurred, damaged, or non-standard license plates may cause OCR errors, while difficult lighting conditions and partial obstructions can affect detection. Future improvements could include expanding the dataset with real-world images to enhance robustness and optimizing the OCR model by leveraging deep learning-based networks to improve character recognition accuracy. The proposed algorithm can be effectively applied in traffic monitoring systems, automated parking solutions, and violation detection, ensuring reliable and efficient license plate recognition.

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Optimal Robot Trajectory Planning via Artificial Neural Network Reinforcement Learning with Metaheuristic Optimizer

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***Abstract* — An approach to robot optimal trajectory planning is developed in the study. The robot's environment is filled with multiple obstacles, which must be avoided during the robot's movement. The trajectory length was minimized, and a penalty function was developed to satisfy obstacle avoidance. The trajectory planner is an artificial neural network (ANN) fed with discrete time and generates discrete positions in the Cartesian coordinate system. Comparison with the DC-programming optimization method evidenced the superiority of the developed approach in terms of solution stability. The only revealed drawback of the ANN application is connected with the local optimality of the generated trajectory.**

Keywords — robot, trajectory planning, constrained optimization, nonlinear objective function, VCT-PSO algorithm.

I. Introduction

Trajectory planning is a crucial aspect of autonomous robotics, enabling mobile and manipulator robots to navigate complex environments while avoiding obstacles. Many algorithms have been developed to address this problem, each offering trade-offs between optimality and computational efficiency. Classical graph-based methods such as the A* algorithm [1] provide optimal paths on discretized representations of the environment, serving as a baseline for many state-of-the-art planning strategies. Other commonly used methods – Rapidly-exploring Random Trees (RRT) [2] and its optimal variant RRT* [3] – exploit a sampling-based approach with a probabilistic nature of trajectory search. However, RRT* may suffer from slow convergence in cluttered environments. Covariant Hamiltonian Optimization for Motion Planning (CHOMP) [4] uses functional gradients, allowing it to find smooth, collision-free trajectories. Another optimal method, TrajOpt [5], utilizes sequential convex optimization to generate locally optimal trajectories, combining obstacle avoidance and dynamical constraints. Metaheuristic optimization methods are also widely applied to optimize trajectories [6]. Obstacles in this work are presented as balls, simplifying trajectory design. The application of ANN to develop a robot trajectory [7]. This scientific work stresses the number of ANN inputs and outputs (regarding the signal nature) and the ANN architecture.

In the current study, we proposed an approach that reduces the trajectory planning problem to the ANN training problem. A metaheuristic method was applied to find the solution to the reduced problem.

II. Problem Statement

In order to state the problem in the mathematical sense, one should observe Fig. 1, where the robot environment is presented. It includes the robot position's initial (x_0, y_0) and the final (x_T, y_T) points. The robot is a point mass that obeys the physics laws (for instance, it cannot be in two different points simultaneously, or Newton's laws govern its acceleration and deceleration).

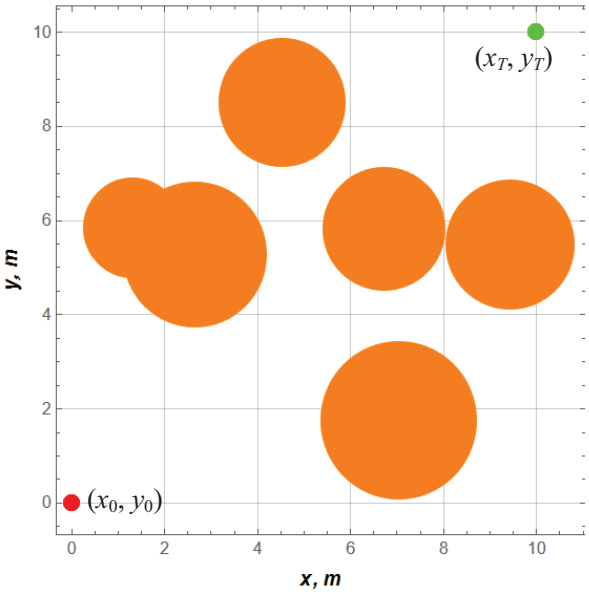


Fig. 1. Robot environment with the denotation of initial (red) and final (green) points and obstacles to avoid (orange discs)

The presented environment (Fig. 1) is a model of the robot's domain – the set of all possible robot positions. The goal of the problem is to find the optimal, in some sense, trajectory (curve) that connects the initial and final points and does not belong to any of the obstacles:

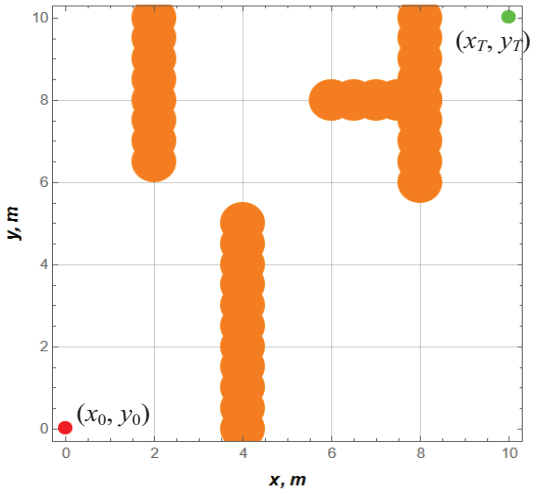
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$$\begin{cases} \begin{cases} x(0)=x_0;\ \ y(0)=y_0; \\ x(T)=x_T;\ \ y(T)=y_T; \end{cases} \\ x=x(t),\ \ t\in[0,\ \ T]; \\ y=y(t); \\ y=y(x)\notin O_i\ \ \forall i\in[1,\ \ I] \\ \int\limits_{x_0}^{x_T}\left(1+\left(\frac{\partial y}{\partial x}\right)^2\right)dx\to\min; \end{cases} \tag{1}$$

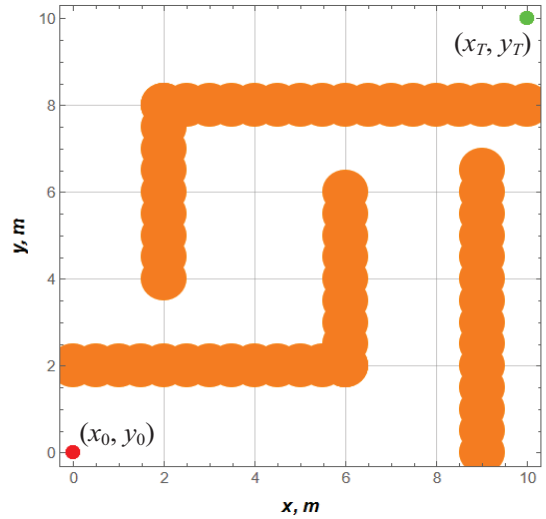
where T – the final time, when the robot reaches the final position; $x(t)$ and $y(t)$ – laws of robot motion along the x and y axes of the „x-y" plane as functions of time; $y(x)$ – law of robot motion as a function of x coordinate; O_i – i-th obstacle; I – total number of obstacles. In many cases, the criterion to minimize is the trajectory length. It is expressed in the last line of the system (1). However, the development of the study approach is not limited to that kind of criterion. Other criteria are possible, and their rational setting and exploitation are issues for further investigations. In the frame of the current study, we assumed that all the obstacles may be presented as discs, with three parameters – two coordinates of a center and a radius. Thus, all the obstacles may be described with the following matrix O_{Ix3}:

$$\mathrm{O}=\begin{bmatrix} x_{o.1} & y_{o.1} & r_{o.1} \\ \dots & \dots & \dots \\ x_{o.i} & y_{o.i} & r_{o.i} \\ \dots & \dots & \dots \\ x_{o.I} & y_{o.I} & r_{o.I} \end{bmatrix}, \tag{2}$$

where O – the obstacle-matrix; $x_{o.\,i}$ and $y_{o.i}$ – the coordinates of the i-th obstacle center; $r_{o.i}$ – the radius of the i-th obstacle. Note that all possible configurations of obstacles may be presented with proper arrangement of disc centres and setting the correct values of discs' radii if $I\to\infty$. In practical cases, I must be big enough. For example, in Fig. 2, two variants of such an arrangement are presented.



a)



b)

Fig. 2. Robot environment with obstacles, which form corridors between initial and final points: a) I=33; b) I=62

Making the value I bigger extends the ability to approximate real obstacles in the robot's environment.

III. Problem Solution

A. Problem reducing

Since the problem (1)-(2) is stated, we should develop an approach to solve it. In the study, we create a strategy involving an ANN approximating the optimal robot trajectory. First, we should note that the solution will search in the discrete form. Thus, continuous functions $x(t)$ and $y(t)$ are replaced by the discrete ones:

$$\begin{aligned} &x(t)\to x_n,\ \ n\in\overline{[1,\ N]}; \\ &y(t)\to y_n. \end{aligned} \tag{3}$$

where N is the total number of discrete time steps.

Next, we must develop the penalty function, which forbids the inclusion of discrete trajectory points into the obstacles O:

$$\begin{aligned} &P=\sum_{n=1}^{N}\sum_{i=1}^{I}P_{i.n.}; \\ &P_{i.n.}=\begin{cases} 0,\ \ if\ \ (x_n,\ y_n)\notin O_i; \\ \delta\left(r_{0.i}-\left\|(x_n,\ y_n)-(x_{0.i},\ y_{0.i})\right\|_2\right), \\ if\ \ (x_n,\ y_n)\in O_i. \end{cases} \end{aligned} \tag{3}$$

where δ – is the weight coefficient (δ=10^6).

Note: to calculate penalty functions, one must check whether every discrete trajectory point belongs to any O_i. To implement this, a loop within a loop must be performed. The inner loop checks whether all the discrete points belong only

to the *i*-th obstacle, and the outer loop does this for all the obstacles.

Let us explain the central concept of penalty function *P* (3). For this, we consider the *i*-th obstacle and the *n*-th trajectory point. Assume that this point belongs to the *i*-th obstacle. Thus, the $P_{i.n}$ is proportional to the Euclidean distance from the point to the *i*-th obstacle boundary (Fig. 3) [5].

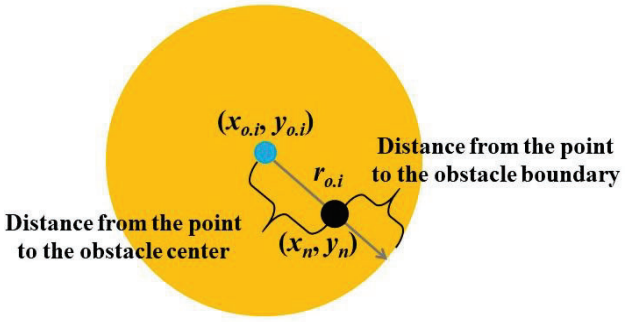


Fig. 3. Graphical explanation of the calculation of the distance from the point to the obstacle boundary

The definite integral (1) for the discrete case must be substituted with the Riemann sum:

$$l = \sum_{n=1}^{N-1} \left\| (x_n,\, y_n) - (x_{n+1},\, y_{n+1}) \right\|_2. \tag{4}$$

Thus, we may write the objective function to minimize:

$$Cr = l + P. \tag{5}$$

B. Selection of the trajectory planner

In order to minimize *Cr,* we must select the proper planner. Since strong approximation properties feature ANN [8] (its ability to map an input to the needed output), we applied it in the study. We consider a one-hidden-layer feedforward ANN with five neurons in the hidden layer.

The parameters of the ANN are given in Table I.

TABLE I. PARAMETERS OF THE APPLIED ANN

Parameter or feature	Value or denotation
Activation function	Arctg
Number of weights	15
Number of biases	7
Domain of weights and biases	-3…3
Domain of gains k_1 and k_2	0…10

The ANN is fed with discrete time $\tau = T/n$ and generates outputs A_1 and A_2, which are post-processed to the discrete functions x_n and y_n (Fig. 4).

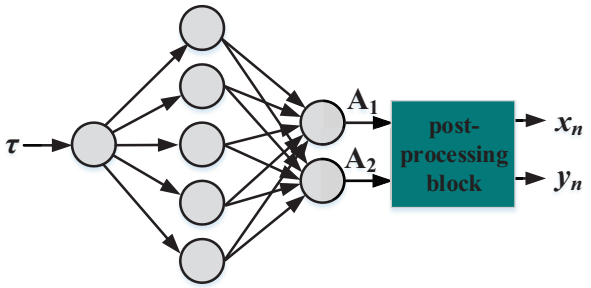


Fig. 4. Graphical representation of the developed ANN

We consider a post-processing block to normalize and gain the ANN outputs A_1 and A_2. It generates dynamically varied values of discrete functions x_n and y_n:

$$\begin{aligned} x_n &= x(0) + \tau\big(x(T) - x(0)\big)\Big(1 + (T-\tau)\pi^{-1} k_1 \mathrm{A}_1(\mathrm{W}, \mathrm{B}, \tau)\Big), \\ y_n &= y(0) + \tau\big(y(T) - y(0)\big)\Big(1 + (T-\tau)\pi^{-1} k_2 \mathrm{A}_2(\mathrm{W}, \mathrm{B}, \tau)\Big), \end{aligned} \tag{7}$$

where k_1 and k_2 – gains of the block; A_1 and A_2 – the first and the second ANNs outputs respectively. To satisfy the boundary conditions of the robot's position (1), the ANN outputs are multiplied by the coefficient $\tau(T - \tau)$. Thus, we may write the reduced problem – the problem of ANN training with reinforcement procedure:

$$Cr(\mathrm{W}, \mathrm{B}, k_1, k_2) \to \min, \tag{7}$$

where W, B – tensor of weights and matrix of biases of ANN, respectively. Indeed, the problem refers to the class of reinforcement learning paradigms. The reward in the problem is inverse to the function (7): the smaller the function (7), the better the ANN is trained.

C. Problem solution with VCT-PSO algorithm

Minimization of the objective function (7) is an unconstrained optimization problem. We may expect a diverse and complicated topology of the function *Cr*. For its minimization, a proper optimizer should be chosen. In the current study, we applied the VCT-PSO algorithm [9] with the following parameters: number of iterations 50, swarm population 30, coefficients $c_1 = c_2 = 1.19$, $w = 0.72$, $RC = 5$. *Cr* reducing during VCT-PSO execution is presented in Fig. 5.

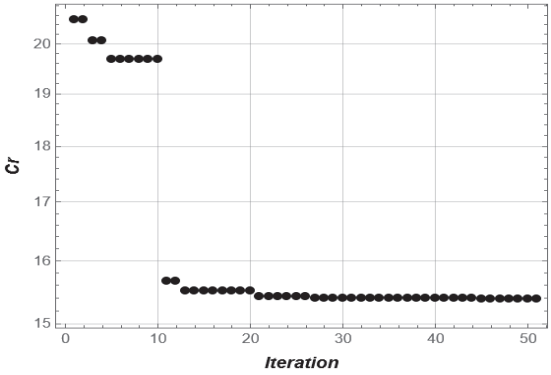


Fig. 5. Objective function reduction during VCT-PSO execution

Fig. 5 evidences that minimization of *Cr* occurred without penalty components: on the first iteration, the VCT-PSO algorithm found a solution without violating the obstacles' boundaries. The algorithm converges quickly at early iterations and slows the convergence rate after the 10th iteration.

IV. Brief Analysis of the Obtained Trajectories

The configuration of the robot's environment with six obstacles and two trajectories is presented in Fig. 6.

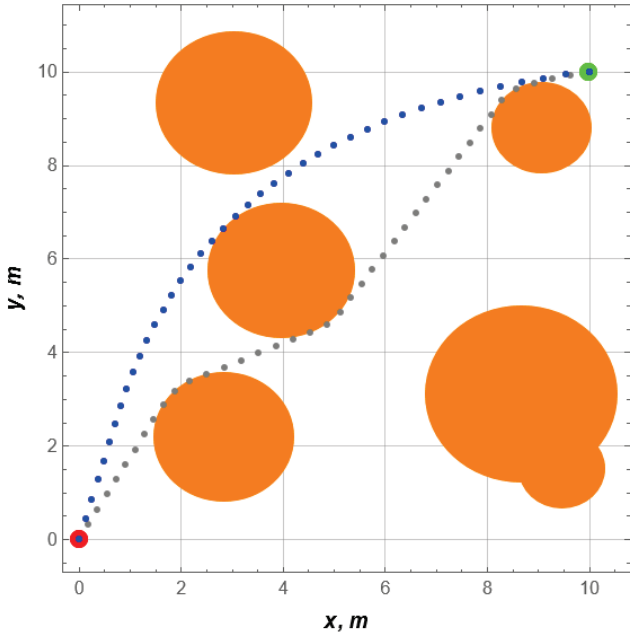


Fig. 6. Two trajectories in the robot's environment with six obstacles

The first trajectory, which is obtained with the described approach, is shown by the blue dots; the second one involves direct numerical calculation of each point's coordinates (the grey dots show it). Since we set N=40, the second solution includes 80 numbers – two coordinates of each point. They were calculated using the DC-programming method [10].

The length of the ANN-based trajectory is 15.34 m, and the second solution trajectory length is 14.75 m.

Fig. 6 shows that the second solution corresponds to the global minimum of criterion (4), while the ANN-based trajectory reaches only the local minimum. However, the features of the ANN-based approach may be revealed with more complicated problems (Fig. 7), i.e., when more obstacles are involved. For the case illustrated in Fig. 7, the VCT-PSO method brings a trajectory length of 17.25 m, and the DC-programming method brings a trajectory length of 29.43 m, which is not even rational. In addition, as one may observe from Fig. 8, both DC-programming trajectories intersect the obstacles, i.e., the constraints of the problem (1) are violated.

Repeated runs of the DC-programming method do not bring better results. Each repeated run of the VCT-PSO method for ANN training corresponds to approximately the same results.

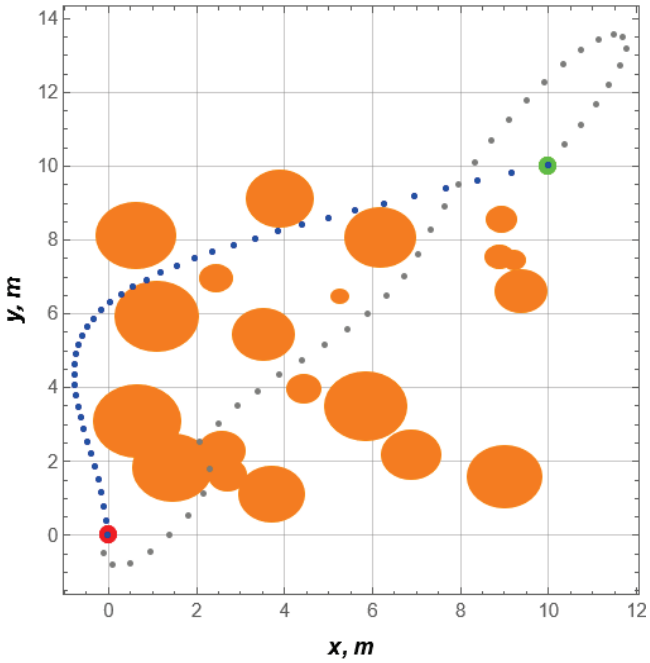


Fig. 7. Two trajectories in the robot's environment with twenty obstacles

Testing both methods for another configuration of the robot's environment (Fig. 8) corresponds to the close results: the DC-programming method fails, the ANN-based approach with the VCT-PSO optimizer allowed to find the local minimum of the trajectory length.

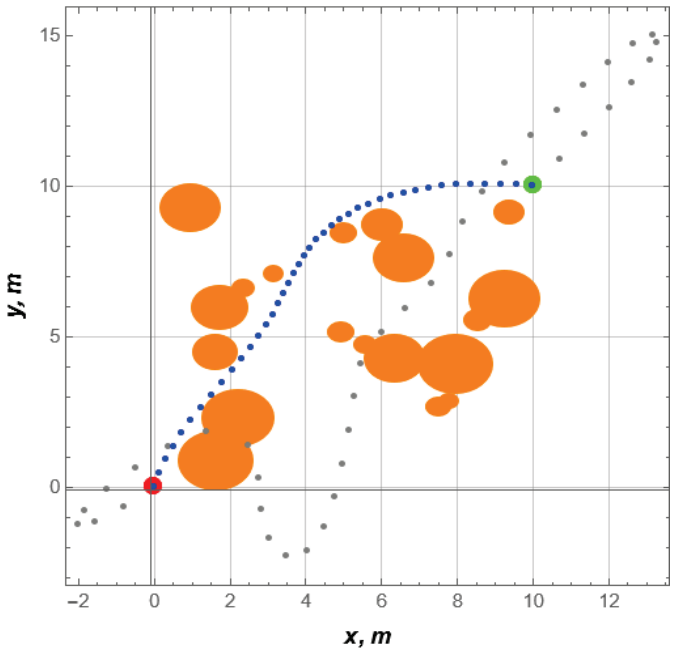


Fig. 8. Two trajectories in the robot's different configuration environment with twenty obstacles

Generally, the ANN-based approach with VCT-PSO optimizer has strong stability of results even for a larger number of obstacles.

To illustrate this statement, we present an ANN-based trajectory in the robot environment with fifty obstacles I=50 (Fig. 9).

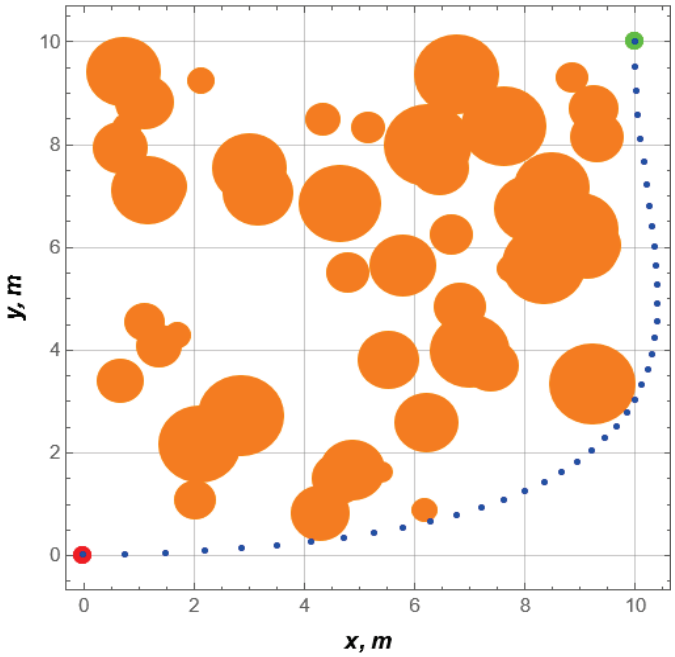


Fig. 9. ANN-based trajectory in the robot's configuration environment with fifty obstacles

However, almost all ANN-based trajectories are locally optimal, which is one possible reason why they are connected with the ANN architecture. Indeed, in the study, a simple ANN was exploited. The bigger number of hidden layers (deeper ANN) allows for building more complicated mapping $\tau \rightarrow (x_n, y_n)$. This thesis is an issue for further studies in this direction.

Another improvement of the developed approach is considering the limitations of the robot's environment. We may note that the additional penalty function must be designed for this reason. It will penalize the distance from where the robot left the environment to the closest boundary. The mathematical description of the additional penalty function will be given in the authors' further works.

V. Conclusions

1. In the study, the problem of planning the optimal trajectory movement of robots was stated. The problem was reduced to an unconstrained one with an ANN-based planner. The objective function of the reduced problem includes only 24 arguments, while the initial problem has 80 of them.
2. To find the problem solution, the VCT-PSO metaheuristic method was applied. It converges to a local minimum, which satisfies all the constraints (obstacle avoidance), but does not correspond to the shortest possible trajectory of the robot movement. Comparison of this approach to the DC-programming method indicates its strong stability even for fifty obstacles to avoid.
3. The issues to consider in the further investigation include, but are not limited to the following items:
   a) making ANN's architecture deeper and figuring out how these changes improve the trajectory;
   b) taking into account the problem statement limitations of the robot's environment as well as different optimization criteria (maximum power consumption, energy losses, etc.);
   c) trajectory planning in 3-D spaces is in the area of interest as well;
   d) further investigation of the developed approach and its comparison with commonly-used algorithms (A*, RRT*, TrajOpt, CHOMP, etc.) is needed to reveal its pros and cons.

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Choosing the Energy Storage Cell Type for a Hybrid Powertrain of a Diesel Multiple Unit

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***Abstract*— The study determined the optimal cell types for the onboard energy storage device of a hybrid diesel train power plant for various storage device usage scenarios. The use of the storage device for energy accumulation during braking and in plug-in mode was considered. It was established that for the energy storage mode, the use of lithium titanate oxide (LTO) cells is rational in terms of mass and size criteria. In plug-in mode, the minimum mass and size parameters are achieved when using lithium iron phosphate (LFP) and lithium nickel manganese cobalt oxide (NMC) cells. The proposed approach can be used when selecting cells for storage devices for various purposes.**

Keywords—diesel multiple unit, rolling stock, plug-in hybrid powertrain, traction system, energy storage

I. Introduction

Hybrid traction systems incorporating energy storage devices in rail rolling stock offer significant environmental and economic benefits through reduced fuel consumption and emissions [1]. This is confirmed by both theoretical studies and the results of regular operation of rolling stock. The hybridization of Régiolis trains manufactured by Alstom involved replacing several diesel engines with energy storage devices [2] that provided an energy recovery factor of more than 90%, resulting in a 20% saving of resources. The results of the operation of the Ki-Ha E200 railbus equipped with an energy storage device showed a 10% fuel saving [3]. The use of a hybrid powertrain on the HD300 locomotive has reduced fuel consumption by up to 36%, NO_X emissions by up to 62%, and noise by 22 dB compared to the DE10 diesel locomotive, based on which the hybrid locomotive was created [4]. Theoretical researches include the study of the use of hybrid traction systems on mainline locomotives [5, 6], shunting locomotives [7], multiple unit rolling stock [8], and specialized rolling stock [9-11]. The results obtained show fuel savings and a reduction in harmful emissions.

A particularly advantageous application of hybrid systems is in multiple unit rolling stock, where frequent stops allow substantial energy recovery through regenerative braking. Results of the article [12] show that for the domestic diesel train DEL-02, equipped with hybrid power transmission, it is possible to achieve 25% fuel savings. In the article [13], the use of an energy storage device in the traction system of the DMU DEL-02 is considered. In the case of using a plug-in storage device, which is charged from a stationary energy source, a reduction in the cost of fuel and energy resources by 36.5-40.4% is achieved. Thus, the use of energy storage devices contributes to the reduction of operating costs.

For the effective use of storage devices, their technical characteristics must ensure their capability to work in different operating modes. The aim of this study is to determine the rational types of elements for the on-board energy storage device, which is part of the hybrid power unit of DMU, for various options for using the on-board energy storage device.

II. Materials and Methods

Current energy storage technologies in rail applications predominantly utilize lithium-based cells [14, 15]. There are applications and theoretical studies on using supercapacitors [12, 16], inertial storage devices [17], and sodium-sulfur cells [18]. Also, lead batteries of different types are used on various rolling stock [19]. The characteristics of the cells vary, and therefore the task of determining the best type of cells for a specific application arises.

In the article [13], the use of energy storage devices as part of the traction system of the DEL-02 DMU is researched. Three options for using the storage device are considered:

1) the storage device accumulates energy during electrodynamic braking and uses it during acceleration (first option);
2) the storage device provides power during movement except for acceleration and accumulates energy during electrodynamic braking (second option);
3) the energy storage device is used instead of one diesel generator set (third option).

In all three cases mentioned above, electrodynamic braking is performed by all traction electric motors.

Table 1 shows the parameters of the energy storage devices. In this case, for the first and second options, it is necessary to use two storage devices since the serial DMU DEL-02 is equipped with two diesel generator sets. In the third option, one storage device is required, which is installed instead of one diesel generator set. The operation of the traction asynchronous electric drive of the DMU is not studied in this work. However, it is worth noting that to increase energy efficiency indicators, an adaptation of the traction electric drive to work with a hybrid power unit is necessary [20-22].

Minimal dimensions and weight are important for onboard energy storage devices. This stems from the fact that there are restrictions on the maximum load on the rails from the vehicle and the dimensions of the rolling stock. Therefore, it seems appropriate to use the weight and size parameters of the storage device as a criterion for determining its applicability to rolling stock. Since cells account for a significant portion of the mass and volume of the onboard storage device, their mass

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and the volume required for placement are calculated using the methods described below.

This study focuses on the use of supercapacitors and lithium cells in the storage. Consideration of inertial storage requires a separate study of the storage itself, which is beyond the scope of this study.

Table 2 shows the parameters of some elements of certain types, which are selected for calculations. Lithium cells were selected in such a way as to reflect different characteristics depending on the chemical structure. The data in Table 2 is used in calculations according to the methods described below. In all calculations, it is assumed that the cell current corresponds to the C-rate value.

TABLE I. ENERGY STORAGE PARAMETERS [13]

Type	Energy storage power rating, kW	Operating capacity, kW·h
1	468	28.7
2	468	200
3	835	450

TABLE II. PARAMETERS OF ELEMENTS FOR ENERGY STORAGE

	Maxwell BCAP3000 P270 K04 [23]	Toshiba SCiB 20Ah [24]	Eve LF50K [25]	CALB L148F 88A [26]	SVOLT 51Ah [27]
Cell type	Super-capacitor	LTO	LFP	LFP	NMC
Capacity, A·h	-	20	50	88	51
Capacity, kW·h	0.0033	0.046	0.160	0.282	0.186
Rated voltage, V	2.7	2.3	3.2	3.2	3.64
Maximum charging voltage, V	2.7	2.8	3.65	3.65	4.2
Minimum discharge voltage, B	1.35	1.5	2.5	2.5	2.8
C-rate	-	3	0.5	1	1
Maximum current, A	170	60	25	88	50
Cell electrical resistance, mOhm	0.15	0.53	0.7	0.3·	0.5
SOC window, %	-	10-90	10-90	10-90	10-90
Number of charge-discharge cycles	1000000	20000	7000	6000	3000
Dimensions, m: length width height	0.06 0.06 0.130	0.116 0.022 0.106	0.135 0.0293 0.180	0.148 0.052 0.102	0.148 0.0265 0.097
Mass, kg	0.475	0.515	1.4	1.62	0.855

The calculation of the number and parameters of the supercapacitor storage device was made following [28].

The number of supercapacitors required to form a storage device of a given operating energy capacity:

$$N_{sc1} = (2E) / (U^2_{sc}\, C_{sc}\, (1\text{-}d^2)) \quad (1)$$

where E – operating energy capacity,

U_{sc} – supercapacitor rated voltage,

C_{sc} – supercapacitor capacity,

$d = U_{min} / U_{sc}$ – supercapacitor discharge depth,

U_{min} – minimum voltage on a supercapacitor.

Number of supercapacitors required to provide storage capacity:

$$N_{sc2} = P / (U_{sc}\, I_{sc}\, d) \quad (2)$$

where P – storage power rating,

I_{sc} – supercapacitor permissible current.

The values obtained by (1) and (2) are rounded to the nearest higher number. The number of supercapacitors to create an energy storage device of a given energy capacity and power is defined as:

$$N_{sc} = \max (N_{sc1}, N_{sc2}) \quad (3)$$

The number of supercapacitors determined by (3) will satisfy only one of the expressions (1) or (2).

After determining the number of supercapacitors, the value of the working energy capacity is specified:

$$E' = (U^2_{sc}\, (1\text{-}d^2)\, N_{sc}\, C_{sc}) / 2 \quad (4)$$

Specified power rating:

$$P' = N_{sc}\, U_{sc}\, I_{sc}\, C_{sc}\, d \quad (5)$$

Minimum storage voltage:

$$U_{sc\min} \approx U_{dc}\, d \quad (6)$$

where U_{dc} – DC-link voltage.

Cells mass:

$$M = N_{sc}\, m_{sc} \quad (7)$$

where m_{sc} – weight of one supercapacitor.

The volume required to accommodate the cells, excluding the structural elements of the storage device

$$V = abhN_{sc} \quad (8)$$

where a, b, h – the length, width and height of one cell, respectively.

Equations (7) and (8) determine the total mass and volume required to accommodate supercapacitors. Features such as the supercapacitor connection scheme, storage design, etc., which may affect mass and volume, are not taken into account in this study.

The calculation of the quantity and parameters of the lithium-ion battery is performed according to [29].

The capacity of the energy storage device is determined as follows:

$$E_{es} = \Delta E / ((SOC_{max}–SOC_{min})(1–D)k) \quad (9)$$

where ΔE – operating energy capacity,

SOC_{max} – maximum permissible cell charge level,

SOC_{min} – minimum permissible cell charge level,

D – reduction in storage capacity over service life, accepted as equal to 20%.

k – the coefficient that takes into account the decrease in the capacity of the energy storage device due to temperature changes, self-discharge, “overcharging” of the cell, etc. For the estimates made in this study, the coefficient can be taken as 0.9.

SOC_{max} and SOC_{min} values are from the storage cell datasheet.

Number of storage cells provided that the operating energy capacity is ensured

$$N_{li1} = E_{es} / E_{cell} \quad (10)$$

where E_{cell} – the energy that one cell can give off when discharged with current I_{cell}. Typically, the current I_{cell} corresponds to the C-rate indicator recommended by the cell manufacturer.

Number of cells with capacity provision condition:

$$N_{li2} = P / (U_{\min} I_{cell}) \quad (11)$$

where $U_{\min}$ – the cell voltage above which the battery must operate at full capacity. It can be taken to be equal to the cell open circuit voltage at a cell state of charge equal to SOC_{min}. It is determined by the SOC–OCV curve of the cell.

The number of cells is determined by the expression:

$$N_{li} = \max (N_{li1}, N_{li2}) \quad (12)$$

Specified energy storage capacity

$$E'_{es} = N_{li} E_{cell} \quad (13)$$

Specified available energy storage capacity:

$$P'' = N_{li} U_{\min} I_{cell} \quad (14)$$

Minimum voltage of the energy storage device

$$U_{li\min} \approx (U_{dc} / U_{\max}) U_{\min} \quad (15)$$

where $U_{\max}$ – the maximum cell voltage can be taken equal to the cell open circuit voltage at a cell charge level equal to SOC_{max}.

The mass and volume are calculated by (7) and (8), respectively using the number of lithium cells calculated by (12).

Thus, the above expressions allow us to determine the mass of the cells and the volume required for their placement. These values allow us to determine whether it is possible to use cells of this type in a storage device, as well as to compare the mass and dimensional parameters of storage devices with different cells.

III. Results and discussion

The calculation of the parameters of the storage that uses supercapacitors was performed according to (1)-(8). In the calculations, it was assumed that the depth of discharge of the supercapacitor is 0.3, since the optimal depth of discharge determined according to [28], equal to 0.1, leads to a large voltage regulation range. This negatively affects the mass and dimensions of the matching converter, with the help of which the supercapacitor accumulator is connected to the traction system. It is worth noting that previous calculations showed that the depth of discharge does not affect the number of cells, since in the studied case their number is determined precisely by the operating energy capacity.

The calculation of the parameters of the storage that utilizes lithium cells was performed according to (9)-(15). In this work, mass and dimensions are used to determine the rational option: the smaller the mass of the cells and the overall volume, the more suitable the accumulator is for use on rolling stock. Economic aspects and cell lifecycle resources are not analyzed in this study.

Tables 3-5 show the results of calculating the parameters of energy storage devices for each option with different types of cells.

According to Tables 3-5, the largest volume and mass of cells have storage devices formed from supercapacitors. For the second option, the mass of the storage device is comparable to the mass of the car, and for the third, it exceeds the mass of the car. Therefore, the use of such a storage device is not possible.

When using lithium cells, the parameters of the storage devices depend on the use case. For the first option (Table 3), the largest mass and volume have storage devices with LFP cells Eve LF50K. For other types of lithium cells, the mass and dimensions of the storage device are close. It is worth noting that the capacity of the storage devices is in the range of 194-974 kW·h, which significantly exceeds the required operating capacity. It is unlikely that the use of such a storage device in the first option is appropriate. A possible way to create a storage device with more consistent parameters for the operation of the first option is to use LTO cells with different parameters – lower capacity and higher operating current.

TABLE III. Energy storage parameters for the first option (28,7 kWh, 468 kW)

Parameter	**Value**				
Cell	***Maxwell BCAP3000 P270 K04 [23]***	***Toshiba SCiB 20Ah [24]***	***Eve LF50K [25]***	***CALB L148F88A [26]***	***SVOLT 51Ah [27]***
Number of cells	10057	4207	6090	1740	3045
Capacity, kW·h	-	194	974	489	497
Rated power, kW	-	543	487	489	497
Maximum voltage, V	1500	1503	1501	1501	1501
Minimum voltage, V	500	1278	1385	1366	1382
Occupied volume, m^3	4.7	1.1	4.3	1.4	1.2
Mass, t	4.8	2.2	8.5	2.8	2.6

TABLE IV. ENERGY STORAGE PARAMETERS FOR THE SECOND OPTION (200 KWH, 468 KW)

Parameter	**Value**				
Cell	***Maxwell BCAP3000 P270 K04 [23]***	***Toshiba SCiB 20Ah [24]***	***Eve LF50K [25]***	***CALB L148F88A [26]***	***SVOLT 51Ah [27]***
Number of cells	87791	7813	6090	1740	3045
Capacity, kW·h	-	360	974	489	497
Rated power, kW	-	1008	487	489	497
Maximum voltage, V	1500	1503	1501	1501	1501
Minimum voltage, V	500	1283	1385	1366	1382
Occupied volume, m^3	41.1	2.1	4.3	1.4	1.2
Mass, t	41.7	4.0	8.5	2.8	2.6

TABLE V. ENERGY STORAGE PARAMETERS FOR THE THIRD OPTION (450 KWH, 835 KW)

Parameter	**Value**				
Cell	***Maxwell BCAP3000 P270 K04 [23]***	***Toshiba SCiB 20Ah [24]***	***Eve LF50K [25]***	***CALB L148F88A [26]***	***SVOLT 51Ah [27]***
Number of cells	162800	17429	10440	3045	5220
Capacity, kW·h		802	1670	857	852
Rated power, kW		2248	835	857	852
Maximum voltage, V	1500	1503	1501	1501	1501
Minimum voltage, V	500	1285	1384	1366	1381
Occupied volume, m^3	76.2	4.7	7.7	2.4	2.0
Mass, t	77.3	9.0	14.6	4.9	4.5

The calculation results for the second option (Table 4) show that the best mass and dimensional indicators have storage devices with LFP cells CALB L148F88A and NMC cells SVOLT 51Ah. The energy storage device on LTO cells has slightly worse indicators. The storage device with LFP cells Eve LF50K has the highest mass and volume, so based on this criterion, its use is not advisable. From Table 4 it follows that the energy capacity of the storage devices for this option is 360-497 kW·h, which exceeds the operating one. This will allow for longer operation in the case of using an energy storage device according to the second option. To create a storage device with parameters that are closer to the operational according to second option, it is necessary to use cells with other characteristics.

Calculations for the third option (Table 5) are similar to the results obtained for the second one. The best mass and dimensional indicators are provided by storage devices with LFP cells CALB L148F88A and NMC cells SVOLT 51Ah. At the same time, the actual capacity exceeds the required one by approximately two times, which allows for movement without charging the storage device for a longer time.

Thus, the calculations confirm the possibility of creating an energy storage device for the traction system of a DMU. For the first option, it is considered advisable to use a storage device based on LTO cells. As the calculations show, for the second and the third options, storage devices based on LFP and NMC cells are advisable. A possible option is LTO cells with improved mass and dimension indicators of the storage.

It is worth noting that in the second option, the energy storage must be integrated into each of the two diesel generator sets. In the third option, one of the diesel generator sets is replaced by an energy storage. As can be seen from the calculations, the mass and volume of two storage units for the second option exceed the mass and volume of one storage unit for third one. In this case, in the second option, it is necessary to place the storage units and related equipment on the cars. In the third option, the storage unit can be installed in place of the diesel generator set that will be dismantled. To ensure the operation of the storage unit according to the third option, it is necessary to connect the traction systems located on different cars. For the final choice of option, it is necessary to study the placement of equipment and assess economic indicators, in particular, taking into account the need to replace cells due to their degradation [30].

IV. CONCLUSIONS

The use of energy storage devices in the traction system of a diesel train with electric power transmission is considered. The parameters of the energy storage device for different cases of its use are determined. According to the criterion of minimum mass and size indicators for the case of braking energy accumulation, it is rational to use LTO cells. In this case, the mass of the cells is 2.2 tons, and the volume required for placement is 1.1 m^3. When using a plug-in storage device, it is advisable to use LFP and NMC cells in energy storage devices. The minimum weight of LFP cells is 2.8 tons and 4.9 tons, and the minimum weight of NMC cells is 2.6 tons and 4.5 tons. The volume required for cell placement is 1.4 m^3 and 1.2 m^3, respectively. To finally decide on the type of cell, you need to figure out its service life and check out the economic effects of using onboard energy storage in hybrid power units for rail rolling stock.

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Study of Integral Stiffness and Damping Characteristics of Active Magnetic Bearings to Provide a Suspension of the Rotating System

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***Abstract*—The paper presents a study of integral stiffness and damping characteristics of active magnetic structural elements which provide a suspension of the vibration isolation system depending on possible operating conditions and parameters of the internal structure of the electromagnetic and electromechanical parts of the rotating machine. The approach to the investigation of the dynamic coefficients of the vibration isolating supports consists in considering the transfer functions of active magnetic bearing components joined into the interconnected system including the electric and rotating mechanical components. The methodology is applied to the solution of the practical problem of avoiding the resonant state of the industrial fan and electric motor assembly by the proper selection of stiffness and damping characteristics of its supports with varying bias current of the electromagnet of the active magnetic structural suspension element. The investigation will be valuable for engineers and scientists who are engaged in the research and development of active electromagnetic systems.**

Keywords—vibration isolation, active magnetic suspension, dynamic coefficients, electromagnet, electric motor

I. Introduction

At the moment, the nonlinear structural analysis suggests engineers and scientists a rich variety of tools to analyze the problems of different nature, from the complex modelling of the destructive processes in the pretensioned structures [1] to the simulation of the rotor dynamics behavior under the impact of the foreign objects [2]. However, even for these cases, it was shown that the linear approach can be applied with some restrictions, for example, the nonlinear impact of the objects can be simplified to the action of force [3] to model the mechanics of the rotor structure without the significant decrease of the accuracy [4].

This approach is also applicable to modelling of the rotor bearings as linear supports with their own stiffness and damping properties [5] which was effectively used by the authors to model the dynamics of rotors in active magnetic bearings [6], [7].

The benefit of using the magnetic levitation is actively used in transportation and turbomachinery fields [8]. Along with passive magnetic bearings [9], active magnetic bearing (AMB) technology offers significant benefits over a conventional bearing and is increasingly being used in different fields of industry. Using AMBs instead of traditional oil-lubricated or rolling element bearings in turbomachinery results in oil-free operation with reduced maintenance, increased efficiency due to lower power losses, and improved reliability because of the controllable dynamic behavior of the system and absence of wear between a shaft and suspension.

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Active magnetic bearings have been reaching high popularity among engineers and scientists as support systems for high-speed and low-friction machines. Since 1980s, applying them became a promising option when designing new turbomachine units or dealing with instability issues in the existing ones [10].

The main idea of the active magnetic bearing operation is based on the electromagnetic processes. It works on the principle of electromagnetic suspension. An electrical current passing through densely wound copper coils creates a magnetic field. This field interacts with a rotor sleeve which is made of a magnetic material such as Silicon-Iron. Levitation of a rotor in AMBs results in a number of their advantages [11] among which are the capability to vary stiffness and damping coefficients providing smooth passing through the resonances.

The size range of machines with AMBs varies from very small to big applications with power and speed limits of 30 MW and 150000 rpm. Active magnetic bearings are used:

- In natural gas extraction and transportation systems for compressors, turboexpanders, and expander-compressor units [12], [13], [14].
- In power industry for steam and gas turbines [15], [16].
- For electrical machines [12].
- For some specific applications where the other types of bearings are inapplicable, such as nuclear centrifuges, energy storage flywheels, machining spindles, turbomolecular pumps, etc. [17], [18], [19], [20].

The approaches to simulation of the dynamics of the 'rotor in AMBs' system as a mechatronic one has its own features in comparison with the standard rotor dynamics (RD) modelling since the interconnectedness of electromagnetic and mechanical systems should be adequately taken into account. Depending on the needs and complexity of the system, these approaches may include:

- Analytical [21] or numerical [22] determination of AMB stiffness and damping parameters with further RD simulations considering the 'rotor in spring-damper supports' models [23].
- Experimental identification of AMB control system parameters [24] with further modelling of its operation in connection to the rotor response including test rig validations [25], [26].
- In-depth modelling of all AMB components [27] with further application of optimization algorithms [28], particularly neural network capabilities [29], to seek

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for the AMB optimal parameters which provide the 'rotor in AMBs' system with the smoothest passing through the resonances and safe operation on the operational regimes.

Despite the high attention to the innovative technology and advantages that AMBs provide, their use is limited due to the complexity of the process of design, analysis, and integration into a machine. The fact that rotors with AMBs are mechatronic systems means that both mechanical and electromagnetic processes take place in the systems simultaneously. Therefore, the comprehensive modelling of AMB component properties and features is crucial when designing the turnkey turbomachines with active magnetic bearings or constructing the AMB plants for the existing machines to replace the outdated or improperly selected bearings. The more adequate AMB models are, the easier switching from these models to the real structure becomes since it provides engineers with better correspondence between model parameters and real properties of AMB components. This enables the widest possible optimization capabilities with an opportunity to tune each feature of a certain AMB component in order to reach the best vibrational properties of the 'rotor in AMBs' mechatronic system.

The goal of the current work is to show the applicability of the approach for modelling of AMBs by their stiffness and damping properties without a decrease in the simulation accuracy which allows to model the rotor dynamics behavior of the machine properly.

The entire approach allows increasing of the mechatronic system efficiency by the possibility to accurately predict its rotor dynamics response which enables switching to such friction-free supports as AMBs and decreases the losses due to friction in bearings.

The rising popularity of AMBs among engineers and scientists provides this work with high relevance, and its novelty is determined by the unique set of input data of the AMB components, namely amplifier features, electromagnet physical properties, controller tunings, antialiasing filter and sensor parameters available for optimization during RD simulations of the considered mechatronic machine.

II. Theoretical Foundation of the Approach

A. Electromagnetic Part Dynamics

To describe the electromagnetic part dynamics, the frequency domain is used. The relation between the frequency domain s and the time domain t is driven by the Laplace transform which for a random function $f(t)$ defined in the range $t \in [0, +\infty]$ is represented by the following operator [30]:

$$\mathcal{L}\{f(t)\} = \int_0^{+\infty} \mathrm{e}^{-st} f(t)dt. \quad (1)$$

The approach presented in the paper utilizes the proportional-integral-derivative (PID) controller model with single input-single output control (SISO) principles which conditions the correspondence between amplifier voltage u and delay voltage u_{delay}:

$$u(s) = G_{PID}(s)u_{delay}(s). \quad (2)$$

The amplifier is considered as a bandwidth-limited transconductance device which presents the relationship between current i_{mag} and voltage u with a closed-loop transconductance $G_{amp}(s)$:

$$i_{mag} = G_{amp}(s)u. \quad (3)$$

Three other AMB components, namely sensor, antialiasing filter, and delay, are connected in a series:

$$\begin{aligned} u_{delay} &= G_{delay}(s)G_{aa}(s)G_{sense}(s)y_{sense} \\ &= G_{out}(s)y_{sense}, \end{aligned} \quad (4)$$

which introduces the correspondence between displacement y_{sense} measured in the sensor and delay voltage u_{delay}.

Finally, the relationship between the current and the displacement measured in the sensor is:

$$i_{mag} = G_{amp}(s)G_{PID}(s)G_{out}y_{sense} = G(s)y_{sense}. \quad (5)$$

This relationship is further transformed to the bearing dynamic coefficients which is a part of the approach and will be shown in the following subsections.

B. Transfer Functions of AMB Components

Although it is proven that the advanced multiple input-multiple output (MIMO) [31] control algorithms like H_∞ [32], [33] give robust control possibilities [34], [35], experiments show [36], [37] that simple analogue schemes can still be used as one of the variants of AMB assemblies for turbomachines to get the benefits of the rotor behavior control [38].

The following table contains the formulas for transfer functions of the AMB components which model their dynamics based on the experience from the provided papers.

TABLE I. AMB Component Transfer Functions

Component	Transfer Function		
	Formula	***Parameter***	***Value***
Controller	$G_{PID}(s) = \frac{K_D s + K_P}{(\tau_1 s+1)(\tau_2 s+1)} + \frac{K_I}{s}$	K_D [µs]	449.3
		K_P [-]	1.362
		K_I [s^{-1}]	82.6
		τ_1 [µs]	33
		τ_2 [µs]	11
Amplifier	$G_{amp}(s) = \frac{k_{amp}}{\frac{s^2}{4\pi^2 f_{b1}^2} + \frac{2\xi s}{2\pi f_{b1}} + 1}$	k_{amp} [-]	1
		f_{b1} [Hz]	1200
		ξ [-]	0.8
Sensor	$G_{sense}(s) = \frac{\frac{u_{max}-u_{min}}{m_r}}{\frac{s}{2\pi f_{b2}} + 1}$	u_{min} [V]	-25
		u_{max} [V]	25
		m_r [m]	0.001
		f_{b2} [Hz]	8000
Sampling delay model	$G_{delay}(s) \approx \frac{24\tau^2 s^2 + 240}{\tau^3 s^3 + 12\tau^2 s^2 + 60\tau s + 120} - 1$	τ [µs]	100
Antialiasing filter[a]	$G_{aa}(s) = \frac{\frac{\alpha^2}{4\pi^2 f^2} s^2}{\frac{1}{4\pi^2 f^2} s^2 + \frac{2\beta}{2\pi f} s + 1}$	α [-]	0.95
		β [-]	0.05
		f [Hz]	100

[a.] The second-order low-path antialiasing filter is used to condition the sensor signals before they are sampled by the proportional-integral-derivative controller.

The values of the parameters, presented in the table, are used for the further consideration of the case of the rotor in active magnetic bearings.

C. Rotor Dynamics Model

The flexible rotor model for the finite-element formulation is described by the matrix form of equilibrium equations including mass [*M*], damping [*C*], gyroscopic [*D*] and stiffness [*K*] matrices and displacement vector $\{z\}$ in relation to the nodal force vector $\{f\}$ [39]:

$$[M]\{\ddot{z}\} + ([C] + \Omega[D])\{\dot{z}\} + [K]\{z\} = \{f\}. \quad (6)$$

For the case of the unbalance response of the 'rotor in AMBs' system, the nodal force vector can be divided into unbalance $\{f_u\}$ and AMB $\{f_{AMB}\}$ forces:

$$\{f\} = \{f_u\} + \{f_{AMB}\}. \quad (7)$$

The relation between the global AMB force vector $\{f_{AMB}\}$ and local bearing force vector $\{f_a\}$ is described by the "local to global" transformation matrix $[T_A]$:

$$\{f_{AMB}\} = [T_A]^T\{f_a\}. \quad (8)$$

The same matrix also represents the local displacement vector in actuator locations $\{z_a\}$ in terms of the overall displacement vector $\{z\}$:

$$\{z_a\} = [T_A]\{z\}. \quad (9)$$

The local bearing force vector $\{f_a\}$ in general is a nonlinear function of the control currents in the bearings $\{i_a\}$ and of the local displacement vector in actuator locations $\{z_a\}$. Consideration of small deviations around a static equilibrium point allows this function to be simplified to the linearized form:

$$\{f_a\} = k_s\{z_a\} - k_i\{i_a\}. \quad (10)$$

The actuator gain k_i and the actuator nominal stiffness k_s depend on the actuator parameters, including the magnetic permeability of a vacuum $\mu_0 = 4\pi*10^{-7}$ H/m, the number of turns of coils per a horseshoe pair *N*, pole cross-sectional area *A*, the half of the angle between two poles θ, the bias current i_b, and the nominal air gap s_0 [40], [41]:

$$\begin{aligned} k_i &= \frac{\mu_0 N^2 A \cos\theta\, i_b}{s_0^2}; \\ k_s &= -k_i \frac{i_b}{s_0}\cos\theta. \end{aligned} \quad (11)$$

The bias current i_b is determined based on the required target stiffness value k_t of the AMB:

$$i_b = \sqrt{\frac{s_0^3 k_t}{\mu_0 N^2 A \cos\theta(1+\cos\theta)}}. \quad (12)$$

Therefore, the resulting AMB force vector can be rewritten:

$$\{f_{AMB}\} = [T_A]^T(k_s\{z_a\} - k_i\{i_a\}). \quad (13)$$

Neglecting the amplifier dynamics considered in the G_{amp} expression in Table 1 and assuming the collocation of actuators and sensors corresponding to them which means that $\{y_{sense}\} = \{z_a\}$ (the displacement signals in the sensor locations are equal to the displacements in actuator locations) allows considering the equation (5) expressed in the local vector form as:

$$\{i_a\} = [G(s)]\{z_a\}. \quad (14)$$

where $[G(s)]$ is the diagonal matrix in case of the SISO control with the diagonal components equal to the transfer functions from the equation (5) with parameters individual for each direction in each AMB (for two radial AMBs with two controlled directions this is a 4x4 diagonal matrix).

If to assume further that the AMB structure is the same in horizontal and vertical directions and the control system is tuned for the equivalent feedback in horizontal and vertical directions, this will allow considering all the diagonal components in the matrix $[G(s)]$ equal to each other and the expression (14) will be simply written as follows:

$$\{i_a\} = G(s)\{z_a\}. \quad (15)$$

Substituting the equations (7), (13) and (15) into (6) with going from the Laplace transform variable to the excitation frequency variable $s = j\omega$ gives:

$$\begin{aligned} &[M]\{\ddot{z}\} + ([C] + \Omega[D])\{\dot{z}\} + [K]\{z\} \\ &-k_s[T_A]^T[T_A]\{z\} + k_i G(j\omega)[T_A]^T[T_A]\{z\} = \{f_u\}. \end{aligned} \quad (16)$$

Here $j = \sqrt{-1}$ is the imaginary unit. Note that in this equation the only complex term is still $k_i G(j\omega)[T_A]^T[T_A] = \Re\mathrm{e}(k_i G(j\omega)[T_A]^T[T_A]) + j\Im\mathrm{m}(k_i G(j\omega)[T_A]^T[T_A])$. As the unbalance response of the rotor is a type of the harmonic response due to the harmonic force $\{f_u\} = \Re\mathrm{e}(\{F_u\}\mathrm{e}^{j\omega t})$ with the excitation frequency coinciding with the rotational speed, the solution of the equation (16) can be searched in the same form $\{z\} = \Re\mathrm{e}(\{Z\}\mathrm{e}^{j\omega t})$. Substituting these forms into the equation gives the following change:

$$\begin{aligned} &(-\omega^2[M]\{Z\} + j\omega([C] + \Omega[D])\{Z\} + [K]\{Z\} \\ &-k_s[T_A]^T[T_A]\{Z\} + k_i G(j\omega)[T_A]^T[T_A]\{Z\})\mathrm{e}^{j\omega t} \\ &= \{F_u\}\mathrm{e}^{j\omega t}. \end{aligned} \quad (17)$$

Simplifying and rearranging this expression gives:

$$\begin{aligned} &-\omega^2[M]\{Z\} + j\omega([C] + \Omega[D] \\ &+\Im\mathrm{m}(k_i G(j\omega)[T_A]^T[T_A])/\omega)\{Z\} \\ &+([K] - k_s[T_A]^T[T_A] + \\ &+\Re\mathrm{e}(k_i G(j\omega)[T_A]^T[T_A])\{Z\} = \{F_u\}. \end{aligned} \quad (18)$$

The additional multipliers at the second and third terms of the equilibrium equations are the impacts of the active magnetic bearings on the rotor system. As they are additives to the stiffness and damping matrices of the rotor, they can be considered as the stiffness and damping matrices of the active magnetic bearings which are written in the following form:

$$\begin{aligned} [K_{AMB}] &= \left(-k_s + \Re\mathrm{e}(k_i G(j\omega))\right)[T_A]^T[T_A]; \\ [C_{AMB}] &= \Im\mathrm{m}\left(k_i G(j\omega)\right)/\omega)[T_A]^T[T_A]. \end{aligned} \quad (19)$$

In view of the same parameters for each direction of each AMB and the fact that the excitation frequency is equal to the rotational speed Ω for the rotor unbalance response, their stiffness and damping in both directions can be calculated:

$$k_{AMB} = -k_s + k_i * \Re\mathrm{e}\big(G(j\Omega)\big); \\ c_{AMB} = k_i * \Im\mathrm{m}(G(j\Omega))/\Omega. \tag{20}$$

These simple relations are used for the calculation of the rotor critical speeds to represent the rotor in AMBs as in the spring-damper elements with properties which are dependent on the rotational speed.

III. Case Study of the Rotor in AMBs

A. Background of Using the AMBs in the Rotor

The paper [5] considered a rotor of the axial industrial fan shown in Figure 1 to determine its static strength and modal characteristics based on the three-dimensional finite-element approach which allows calculating the mode shapes associated with the vibrations of the blades, the disc, the shaft, etc. However, in this paper, only first six natural frequencies were considered which were not associated with the shaft deflection, and the bearings were modeled as rigid ones since the rotor dynamics natural frequencies were higher than the considered natural frequencies.

The structural analysis did not consider the significant drop in the efficiency of the machine due to the increased frictional losses in the rolling element bearings because it was out of scope of that study. However, due to the underloading of the bearings, the slippage of rolling elements occurred which led to the necessity of using the increased clearance but did not resolve the issues of the rolling element bearing effect on the machine efficiency.

To overcome this issue, it was suggested to use a simple scheme of active magnetic bearing support with SISO PID control which is relatively cheap to implement in comparison with the advanced MIMO digital schemes but should give the desired benefit of reducing the frictional losses in the supports and having a controllability on the rotor resonances which may arise due to the transfer from the rolling element bearings to the active magnetic bearings.

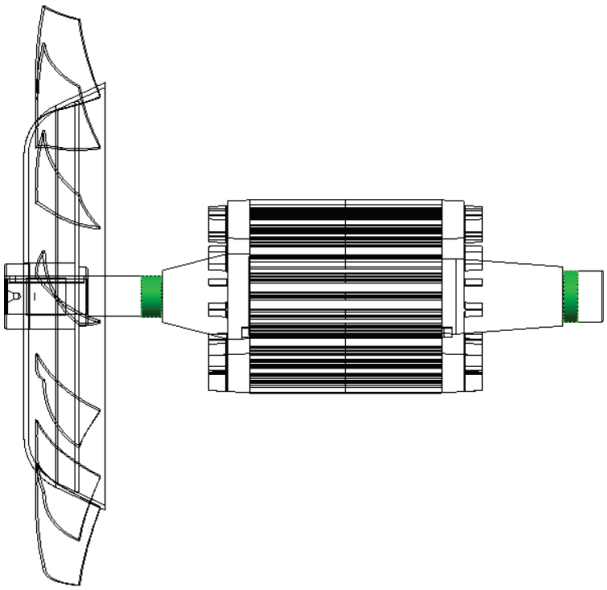


Fig. 1. Rotor model with bearing locations shown in green color

B. Input Data for the Rotor Dynamics Analysis

The parameters of the 'Rotor in AMBs' assembly related to the rotating and electromagnetic parts are presented in Table 2 (the bias current is varied for optimization purpose).

TABLE II. AMB and Rotor Properties

Component	Properties		
	Name	***Parameter***	***Value***
Electromagnet	Number of turns	N, [-]	88
	Pole cross-sectional area	A, [mm^2]	675
	Angle between poles	2θ, [deg]	45
	Bias current	i_b, [A]	4..8
Actuator	Gain	k_i, [N/A]	388..777
	Nominal stiffness	k_s, [MN/m]	5.74..23
	Nominal air gap	s_0, [mm]	0.25
Rotor	Mass	m, [kg]	161
	Diametral mass MOI[a]	I_d, [kg*m^2]	7.154
	Polar mass MOI[a]	I_p, [kg*m^2]	2.445
	Location of COG[a]	x_{COG}, [mm]	337.32
	Front bearing diameter	D_f, [mm]	55
	Rear bearing diameter	D_r, [mm]	69.4
	Bearing span	x_b, [mm]	594
	Target bearing stiffness	k_t, [MN/m]	11..44
	Rated speed	Ω_{rated}, [rpm]	3000

[a] MOI – moment of inertia; COG – center of gravity.

These parameters in addition to the transfer functions presented in Table 1 are used in the calculations of the dynamic response of the 'Rotor in AMBs' assembly.

C. Calculation Model of the Rotor in AMBs

The three-dimensional finite-element (FE) model shown in Figure 2 was created to calculate the dynamic response of the rotor in active magnetic bearings modeled as spring-damper elements using stiffness and damping coefficients.

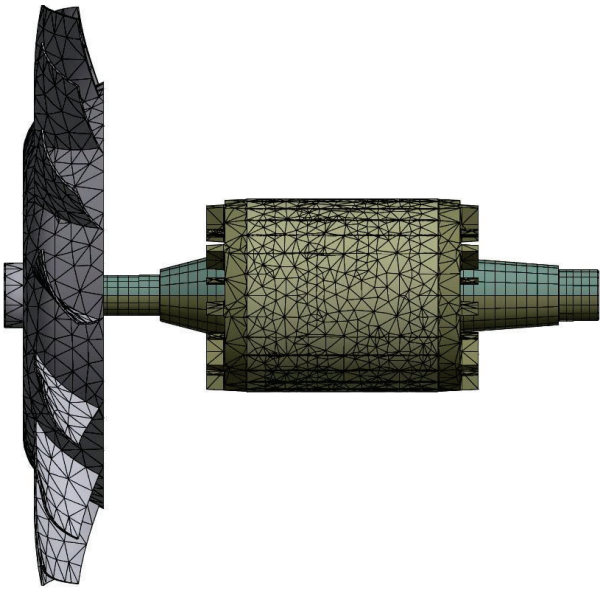


Fig. 2. Finite-element model of the rotor used for calculation

The geometrical model of the rotor was modified by simplifying the parts less important for the rotor dynamics response as a main focus of the study to reduce the FE model size having 243175 nodes and 135623 quadratic finite elements of tetrahedron, pyramid and hexahedron shape with special attention to the swept mesh creation for the shaft as the most responsible part of the rotor dynamics response and free mesh for the lumped elements which in the same way as the mapped one can represent their distributed mass-inertia properties correctly (the mesh error is less than 2%).

As per assessment made in [5], the front and rear bearing reactions were about 1000 N and 500 N respectively which required the appropriate electromagnet selection to provide the corresponding bearing capacity. The source [27] presented an in-depth study of the rotor in AMBs with the electromagnets having the bearing capacity of 2400 N which satisfied the requirements for the current rotor. Based on this, the majority of parameters presented in Table 1 and Table 2 were selected according to this source with some tuning for the size and modal characteristics of the investigated design. This refers to the use of the second-order low-path filter to cut off the first free-free backward whirl critical speed of the rotor equal to 113 rotations per second.

IV. Results and Discussion

A. Stiffness and Damping of AMBs

Although the rotor rated speed is 3000 rpm, stiffness and damping coefficients of active magnetic bearings were calculated up to 10000 rpm to give the insights of the possible resonances due to the presence of rigid body modes and for different values of bias current to find the optimal parameters of the AMBs. Figure 3 presents the dependence of these coefficients on rotor rotational speed (RS). This dependence is used for calculation of rotor critical speeds as spring-damper element direct stiffness and damping coefficients.

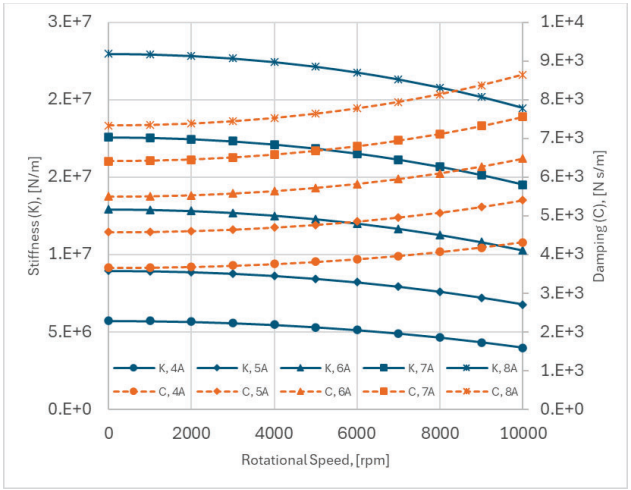


Fig. 4. AMB stiffness and damping vs RS for i_b = 4 A..8 A

B. Rotor Dynamics Response in AMBs

As mentioned in the previous paragraph, the calculation of AMB stiffness and damping was performed for variable bias current which allowed determining the backward whirl (BW) and forward whirl (FW) critical speeds (CS) of the rotor for this variation shown in Figure 4 and, as a result, the required value of the bias current to have no resonances within the rotational speed range.

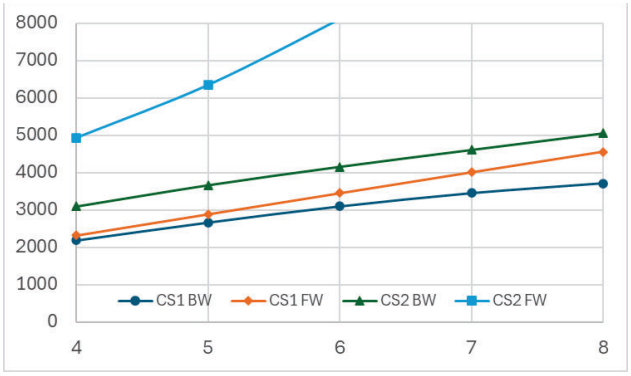


Fig. 3. First two BW and FW critical speeds, [rpm] vs bias current, [A]

As it follows from the figure, the bias current of 8 A results in the 20% margin of the first backward and forward whirl critical speeds above the rated speed of the rotor which makes it rigid (no resonances occur during rotor startups and shutdowns). Figure 5 shows the Campbell diagram for this value of bias current with the corresponding mode shapes.

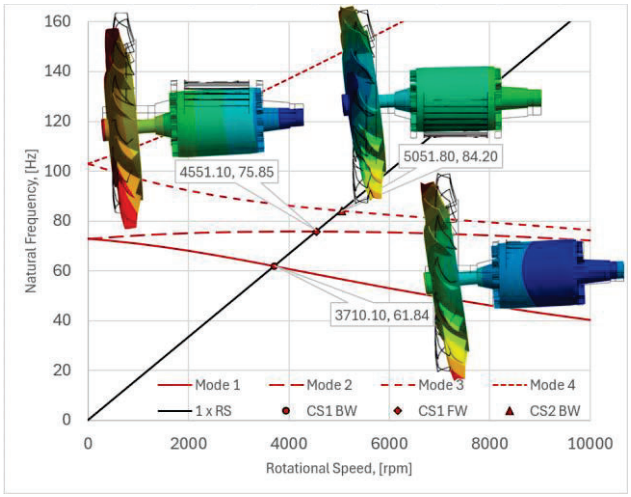


Fig. 5. Campbell diagram for AMBs with stiffness and damping coefficients identified for i_b = 8 A

The first BW and FW critical speeds associated with combined rigid-flexible body modes are about 3710 rpm and 4551 rpm respectively which allows concluding that the use of AMBs with selected parameters in the rotor results in its non-resonant operation for its whole rotational speed range.

V. Conclusion

The paper presented an approach which allows identifying the active magnetic bearing stiffness and damping coefficients for their use in rotor dynamics calculations intended to justify the possibility of AMB implementation in the existing rotary machine for increasing its efficiency which had a decrease due to ineffective operation in rolling element bearings. The approach is based on the consideration of the AMB component dynamics in the form of the corresponding transfer functions. The results of the rotor dynamics response of the machine installed in the active magnetic bearings with a SISO PID control and other parameters specially tuned for the considered design indicate the non-resonant operation of the industrial fan rotor in AMBs which enables further use of these bearings for increasing the efficiency of the considered fan by approximately 10% due to switching from the rolling element bearings with high friction to the frictionless AMBs.

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Optimization of petrographically heterogeneous multi-basin coal batch composition taking into account the electrical resistivity of coke

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***Abstract*—The dependence of coke resistivity on the component composition of the coal batch and the final process temperature was confirmed. Given that the resistivity and reactivity are affected by some common factors, such as temperature and coke readiness, which indicates the orderliness of its structure, the relationship between these indicators was confirmed. The resistivity is an aggregate indicator of coke properties, the values of which are determined by the ash, sulfur content, volatile matter, petrographic composition and carbonization conditions of coal raw materials. In the course of research, a comprehensive selection and study of components included in the raw material base of coking of coke production was carried out. According to the results of technical, plastometric, petrographic analyses of the studied coals recommendations on their rational use in the batch of the enterprise were given. These recommendations have been tested by pilot coking, according to the results of which the proposals on providing rational temperature regime of their coking have been given.**

Keywords—electrical resistivity, petrographically heterogeneous inter-basin coal batch, coke structure, process temperature, coke reactivity index.

I. INTRODUCTION

At present, blast furnace production sets higher requirements to coke quality, the standards of which are fixed both in the current technical specifications (A^d, S^d_t, M_{10}, M_{25}) and in ISO 18894-2006 (CRI and CSR). Thus, the current requirements for coke quality are $M_{25} \geq 88.0\text{-}90.0\%$; $M_{10} \leq 6.0\text{-}6.5\%$; CSR – 60.0-75.0%; CRI – 25-30.0 %.

Analysis of coal resources shows that domestic raw material base is not able to provide production of blast furnace coke of improved quality on its own. As a consequence, the question of attraction and rational use of imported components (coals from far abroad) in coal batch for coking arises [1].

Classification of distant foreign coal and its aggregation into classes and groups [2,3] differs from the accepted classification of grades in accordance with DSTU 3472-2015.

Therefore, it is obligatory to formulate the coal mixture taking into account the maceral composition, the indicator of vitrinite removal and the distribution of vitrinite components by stages of metamorphism, the optimal ratio between clinkering (ΣCC) and lean (fusinized) components (ΣFC). When predicting the output and properties of coke, we should take into account that the macerals of the groups of vitrinite, liptinite and 1/3 of semivitrinite constitute the sum of reactive and clinkering components, while the whole of inertinite and 2/3 of semivitrinite have an oppressive effect on the system as a whole [4].

In order to determine the optimal conditions for preparation and coking of batch with a large content of petrographically heterogeneous coal in laboratory and experimental-industrial conditions, studies were conducted to determine their rational grade and component composition, preparation schemes, coking mode, as well as to assess the impact of these measures on the quality of blast furnace coke (in terms of mechanical strength, specific electrical resistance).

Carrying out the above measures is especially relevant in the conditions of coke-chemical enterprises of Ukraine, the raw material base of which contains coals of different basins.

II. MATERIALS AND METHODS

Samples of the production batch. Component composition of the batch was as follows: A-10%. E-11%, B-8%, F+G+H-10%, D-35%, C-26%. The technological properties and petrographic characteristics of these samples were determined.

The characteristics of the coal concentrates were determined according to the following standard methods: ISO 1171-97 “Solid mineral fuels. Methods for determination of ash”; ISO 589-81 “Hard coal –

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Determination of total moisture"; ISO 7404-3-84 "Methods for the petrographic analysis of bituminous coal and anthracite - Part 3: Method of determining maceral group composition"; ISO 7404-5-85 "Methods for the petrographic analysis of coals–Part 5: Method of determining microscopically the reflectance of vitrinite"; State standard of Ukraine 7722:2015 "Hard coal. Method of determining plastometric characteristics".

TABLE I. TECHNOLOGICAL PROPERTIES OF COAL SAMPLES

Sample	Rank of coal DSTU 3472-2015	Proximate analysis, %				Plastometric parameters, mm	
		A^d	S^d_t	V^d	V^{daf}	x	y
A	G	9.4	0.32	37.3	41.2	35	14
B	GZh	6.9	0.73	33.3	35.8	26	17
C	Zh	9.1	0.54	29.8	32.7	31	19
D	K/KP/ KZh	12.2	0.67	21.5	24.5	28	15
E	K	8.8	0.72	26.6	29.2	17	12
F	K	12.8	1.75	24.0	27.5	36	14
G	PS	9.9	0.40	19.0	21.1	16	11
H	K	8.6	0.41	18.2	20.0	25	11

TABLE II. TECHNOLOGICAL PROPERTIES OF COAL BATCH

Sample	Proximate analysis, %				Plastometric parameters, mm	
	A^d	S^d_t	V^d	V^{daf}	X	Y
Basic batch	10.4	0.83	26.5	29.6	34	13

TABLE III. PETROGRAPHIC CHARACTERISTICS OF COAL BATCH

Sample	Petrographic composition (without mineral impurities), %					Mean vitrinite reflection coefficient, %
	Vt	Sv	I	L	∑ FC	R_o
Basic batch	74	1	22	3	23	1.07

The production batch of the plant has increased ash (A^d=10.4%) compared to the batch of other plants of Ukraine, where ash of the batch, as a rule, varies within 8.5-9.0%. Sulfur content in the batch is low (S_t^d= 0.83%). It should be noted that the composition of the batch vitrinite includes coals of almost all stages of metamorphism from from low-metamorphosed to highly metamorphosed. The main amount of vitrinite constituents vitrinite constituents have reflection coefficient R_o values in the range of 0.65-0.84 (20%) and 0.90-1.19 (59%).

The total content of vitrinite constituents corresponding to coals Zh and K grades amounted to 66%. However, taking into account that the absolute content of vitrinite in the batch is only 74%, the actual amount of vitrinite components having a reflection index of 0.9-1.39%, in the organic mass amounted to only 46.9%, which is far enough from the recommended minimum value of 53%, which ensures the production of high-strength metallurgical coke. Reduced amount of well clinkering components of vitrinite and increased content of non-clinkering macerals of inertinite group, as well as high ash, caused low sinterability and coking ability of the batch as a whole, which was reflected in the results of the study of plastic and viscous properties of the batch.

The variants of pilot batches were developed to test them by box coking (batches1-3). The grade and component composition of the pilot batches, as well as petrographic characteristics of coals and batches are given in Table 4. At development of variants of experimental batches we were guided by the fact that first of all it was necessary to reduce the content of non-binding macerals of inertinite group in experimental batches in comparison with the basic one, first of all at the expense of coal H. Also, it is necessary to exclude from the batch coal F, which actually does not contain vitrinite with reflection index 1.2-1.39, and has in its composition an increased number of components, with R_o=0.65-0.84%, R_o=0.90-1.19% and R_o=1.7-2.59%. Introduction of petrographically homogeneous H coal into the batch, as well as increase in the content of E and A coal, which have the minimum content of inertinite group macerals, also contributes to the solution of the above problems.

TABLE IV. PETROGRAPHIC CHARACTERISTICS OF COAL SAMPLES AND BATCHES

Sample	Petrographic composition (without mineral impurities), %					Mean vitrinite reflection coefficient, %
	Vt	Sv	I	L	∑FC	R_o
A	85	-	12	3	12	0.69
B	73	-	23	4	23	0.93
C	79	-	20	1	20	0.94
D	91	-	8	1	8	1.06
E	89	1	8	2	9	1.31
F	76	3	21	-	23	1.14
G	38	2	60	-	62	1.28
H	89	1	10	-	11	1.53
Basic batch	74	1	22	3	23	1.07
Batch 1	79	-	20	1	20	1.00
Batch 2	78	1	19	2	20	1.05
Batch 3	80	1	18	3	18	1.1

The content of fusinized macerals (ΣFC) in the experimental batch compared to the base batch decreased from 23% to 19-20%. The amount of vitrinite components with a mean vitrinite reflection coefficient of 0.9-1.39% in organic mass increased from 45.1% in the base batch to 51.4, 49.9 and 48.8%, respectively, in the experimental variants.

Box coking was carried out according to the following methodology. Experimental batches weighing 5.5 kg were loaded into metal crates. With the crate volume of 7.3 dm^3, the density of the experimental batch in them was approximately 0.760 t/m^3, that is, it corresponded to the average practical data on the density of the batch in the chamber. Photos of the boxes are presented in Figure 1.

Coking of experimental batches was carried out on coke batteries (effective volume of the chamber 21.6 m^3) in specially selected ovens with good condition of masonry of heating partitions and stable thermal regime.




Fig.1. Metal boxes for coking of pilot batches

Loading of boxes was carried out through the middle hatch after the release of batch from the coke and machine hoppers of the coal loading car, so that the boxes with batch were located at a level of approximately 2.5 m from the bottom of the coking chamber, that is, in the zone of optimal temperature conditions. Four boxes were loaded into each chamber, according to the number of experimental batches. Experimental coking was carried out under the thermal regime, the parameters of which are presented in Table 5.

TABLE V. COAL BATCHES COKING CONDITIONS

Coking period, hours	Temperatures in the control verticals, °C		Coke cake axis temperature, °C
	machine side	coke side	
15.5	1320	1350	1035

Once the coking process was complete, the coke cake was delivered using standard technology (from the chamber to the quenching car, then quenching and unloading onto the coke slide). If necessary, the boxes were quenched with water and pre-cooled in air.

The obtained experimental coke was extracted from the crates, weighed and dried to constant weight to determine the yield of gross coke from the batch and subsequent analyses (proximate analysis, mechanical strength, thermochemical properties). The characteristics of the obtained cokes (indicators of proximate analysis, physical and mechanical properties, thermomechanical properties) were determined according to the following standard methods: ISO 556-80 “Coke (greater than 20 mm in size)-determination of mechanical strength”; ISO 18894:2006 “Coke-determination of coke reactivity index (CRI) and coke strength after reaction (CSR)”.

Additionally, to characterize the orderliness of the structure of coke and the degree of their readiness determined the specific electrical resistance for DSTU 8831:2019 Coke. Method for determination of the resistivity of coal coke powder, which establishes a method for measuring the resistivity of coal coke powder in the range from 0.05 Om·cm to 0.5 Om·cm. The results of studies of experimental coke are shown in Tables 6 and 7.

III. RESULTS AND DISCUSSION

As can be seen from the given data, ash of basic and experimental coke is high and varies within 12.6-12.9 %. At the same time, sulfur content is low and varies in the range of 0.52-0.61%.

The ashing coefficients depend on the yield of volatile substances from the batch, decreasing at lower yield and increasing at its increase. The coefficient of residual sulfur in coke is within the range of 0.800-0.869.

TABLE VI. QUALITATIVE INDICATORS OF BOX COKING COKE AND ASHING AND DESULPHURIZATION COEFFICIENTS

Sample	Proximate analysis, %			Ashing coefficient	Desulphurization coefficient	ρ, Om·cm
	A^d	S_t^d	V^{daf}			
Basic	12.9	0.61	0.9	1.303	0.806	0.0597
Batch 1	12.9	0.53	0.6	1.330	0.815	0.0568
Batch 2	12.6	0.52	0.5	1.299	0.800	0.0562
Batch 3	12.8	0.53	1.1	1.312	0.869	0.0547

TABLE VII. QUALITATIVE INDICATORS OF COKE

Sample	CRI, %	CSR, %	Yield and strength of coke, %		
			Yield	M_{25}	M_{10}
Basic	38.3	40.5	76.31	87.1	8.0
Batch 1	36.5	44.0	75.26	87.3	7.8
Batch 2	36.1	44.9	76.60	87.4	7.5
Batch 3	35.2	48.8	75.82	87.7	7.4

Electrical resistivity is a measure of temperature action during the thermal destruction of coal and coke conversion. Regardless of the final temperature of the process, the pithomium electric opir decreases the more the higher the temperature is. Therefore, the indicator of pithomic electric pressure can characterize the level of readiness of coke. Also, the readiness of the obtained coke is good, which is indicated by the data on the yield of volatile, the value of which is equal to 0.5-1.1%, and the value of specific electrical resistance, the values of which are within the range of 0.0547-0.0597 Om·cm.

Indices of reactivity CRI and post-reaction strength CSR, determined by the method of “Nippon Steel” were the worst for coke from the base batch, and the best values were obtained for coke from batch of variant 3. Coke from the batch of variants 1 and 2 is also better than coke from the basic batch, primarily due to the fact that F coal, which has an extremely high ash of 12.8%, was removed from the composition of the experimental batch. The results of thermochemical characteristics correlate with the data on the value of electrical resistance of coke, which corresponds to the results of studies presented by us in [5,6]. The established correlation between the coke electrical resistivity and reactivity and post-reactivity strength is logical. Coke electrical resistivity and reactivity are different for their nature. Thus, the analysis of many studies [7-10] shows that the CRI is influenced by the properties of the used coal (ash, content of sulfur, mineral content, yield of volatile substances, petrographic composition), the technology of batch preparation and heat engineering mode of its coking, and the period of coking. At the same time, both of these indicators change synchronously due to the fact that they are influenced by some combined factors, such as temperature level and coke readiness, which indicates the order of its structure.

The resistivity reflects the state of the carbon structure and the density of contact between the residual material of

the coal grains and the petrographic components that make up the organic mass of the grains. Consequently, coke produced from poorly coking and petrographically heterogeneous and high-ash coal is characterized by a higher resistivity. An increase in resistivity and reactivity indicates a decrease in the orderliness of the coke structure. Consequently, the reactivity of coke will increase with increasing electrical resistivity. Also, the electrical resistance decreases with increasing structural strength and abrasive hardness [11-14].

Based on the analysis of experimental data, a graph of the dependence of the specific electrical resistance of coke on the reactivity of coke (Fig.2), which is characterized by a high correlation coefficient, was obtained r = 0.84.

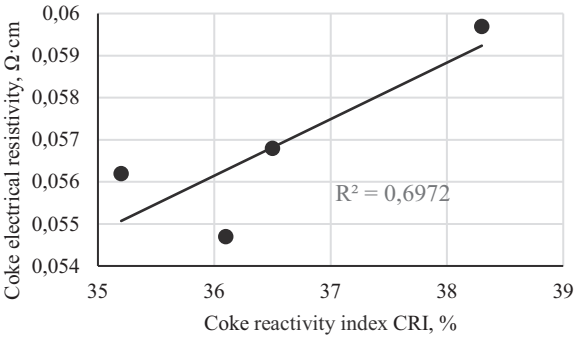


Fig.2. Relationship between the coke electrical resistivity and the coke reactivity index

The analysis shows that the best values of M_{25} and M_{10} have coke from charge of variants 2 and 3, which contain G coal.

Transition to batch coking without participation of G and F coals allows to noticeably improve mechanical strength indicators, as well as CRI and CSR indicators. Quality indicators of metallurgical coke were improved due to introduction of 10% H coal into the batch.

Thus, the minimum specific electrical resistance was recorded for variant 3 of the experimental batch, which is characterized by the minimum ash, sulfur content, maximum vitrinite content, and the value of the vitrinite reflection index [15].

IV. Conclusions

The experimental studies made it possible to formulate the following conclusions:

- petrographically heterogeneous and multi-basin raw material base causes certain difficulties in the development of rational compositions of coal batches, taking into account the conditions of their preparation and coking to obtain blast furnace coke of stable quality;
- based on the results of pilot coking of scientifically based compositions of coal batches developed with the participation of imported coal, three variants of coal batch compositions were proposed, the coke from which has better quality indicators.

Thus, it was found that changing the component composition of the coal batch by increasing the content of highly metamorphosed coal, as a result, causes an increase in the vitrinite reflection index, and also leads to an increase in the orderliness of the coke structure and a decrease in its specific electrical resistance.

It has been shown that when optimizing the component composition of coal blends, it is necessary to take into account the value of the electrical resistivity of coke, which depends on the ash, volatile matter, petrographic composition, and vitrinite reflectance.

It should be noted that the use of the coke electrical resistance test to optimize the component composition of blends and monitor coke quality in a particular coke production is practical and economically feasible. Since the electrical resistance determination is simpler, does not require reagents, and takes much less time than the reactivity test. Therefore, it can be considered an operational method for determining the structural characteristics and readiness of coke.

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Modeling Load Profile of Aggregated Electric Water Heaters for Energy Market Integration

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***Abstract*—This paper investigates the potential of residential electric storage water heaters (ESWHs) to provide demand-side flexibility in modern power systems. To capture both physical and behavioral aspects of ESWH operation, a thermophysical simulation in OpenModelica is integrated with an agent-based behavioral simulation of hot water usage using the Load Profile Generator (LPG). We propose, for the first time, the Reserve Coefficient of Aggregated Portfolio (RCAP) as a novel metric to quantify the availability and controllability of aggregated flexibility in ESWH portfolios. The RCAP framework enables precise quantification of flexibility, linking technical availability with market-oriented decision-making and supporting strategic planning for aggregators and grid operators. The approach generates realistic and diverse demand profiles across different household types, capturing daily and weekly variability, and validates them against empirical consumption patterns. Aggregation is modeled by simulating market participation, segmenting households into representative groups, and estimating baseline consumption for aggregated loads. The RCAP metric further quantifies the thermal availability and controllability of the ESWH portfolio over time. Collectively, these contributions provide a robust basis for integrating ESWHs into demand response programs, enhancing both operational efficiency and market value.**

Keywords—electric storage water heaters, hot water demand profiles, residential demand response, consumer portfolio aggregation, demand side flexibility, aggregator strategy

I. Introduction

In contemporary power systems, integrating variable renewable generation and ensuring grid reliability require both technical and managerial innovations in demand-side flexibility. Residential ESWHs, when aggregated and intelligently controlled, represent a significant source of controllable load that can support balancing services, improve resource utilization, and enable new business models for aggregators, energy service companies, and grid operators.

A precise estimation of the flexibility potential of ESWHs, along with the system-level implications of their activation, requires detailed modeling of end-user behavior, particularly realistic hot water usage patterns and the resulting load profiles. Ignoring the diversity and stochasticity of user behavior can lead to substantial errors in flexibility forecasting and, more critically, in predicting the timing and magnitude of rebound effects that may threaten grid stability.

From a strategic perspective, accurate ESWH modeling is not only a matter of technical precision but also a foundation for informed decision-making. Aggregators rely on such modeling to design optimal bidding strategies, manage portfolio composition, and target specific consumer segments for participation in demand response (DR) programs. Load profiles serve as essential instruments for understanding user behavior, optimizing energy resource allocation, and facilitating DR integration. At the same time, advanced data analytics and automated control algorithms enable tailored activation strategies that maximize participation while preserving user comfort.

Several methodological approaches exist for modeling ESWH operation, including analytical, statistical, and simulation-based techniques. Analytical models, grounded in thermodynamic and electrical differential equations, offer detailed insights but fail to capture behavioral variability. Statistical models use historical consumption data to forecast usage via regression or machine learning, providing adaptability but relying on data quality. Simulation-based methods, particularly agent-based modeling, integrate heterogeneous consumer behaviors and system interactions, producing realistic and granular consumption profiles suitable for strategic flexibility planning.

This study aims to analyze ESWH load profiles in residential contexts, identify characteristic consumption patterns under different operational regimes, and introduce the RCAP as a decision-support tool for technology management in distributed energy systems. The proposed framework links behavioral and physical modeling to inform optimization, planning, and innovation strategies for real-world DR programs and market participation. Furthermore, the incorporation of socio-demographic profiling ensures that demand response planning reflects realistic household diversity, enhancing both the technical robustness and the strategic value of the proposed approach. This integration enables stakeholders to directly translate modeling results into effective operational and market strategies.

II. Simulation of ESWH Usage

A. Analysis of Hot Water Demand Models

Studies [1] have demonstrated that household electricity demand exhibits a strong temporal dependency, closely aligned with residents' daily routines and lifestyle patterns. The use of ESWH is particularly related to occupants' work and rest schedules, as evidenced by time-use data. Findings in [2] indicate that the return-home time of residents significantly contributes to evening peak electricity demand, a factor that must be incorporated into demand-side management strategies.

A promising direction of research involves the use of behavioral models that capture the intra-day distribution of load. In [3], neural networks were applied to forecast long-term shifts in household behavior and their corresponding

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impact on electricity consumption, enabling peak load shifting and improved system optimization. Agent-based approaches are particularly valuable in this context, allowing for the simulation of appliance usage scenarios based on behavioral models that incorporate both habitual practices and external influences such as tariff changes and renewable energy integration.

Stochastic models have also been widely used in residential water consumption modeling. One such model, the Stochastic Residential End-use Model (STREaM) [4], generates high-resolution synthetic time series of water usage based on the operational characteristics of individual fixtures, including frequency, duration, and time-of-day patterns. STREaM supports scenario analysis and impact assessment under varying consumer behaviors and management policies.

Despite its advantages, STREaM lacks flexibility in adjusting parameters to account for household-specific behaviors or socio-economic differences, limiting its applicability in customized settings. Additionally, research [5] has shown that while high-frequency sampling improves anomaly detection and end-use disaggregation, it also imposes greater demands on data storage and processing. These limitations highlight the need for adaptive modeling techniques, with agent-based models offering a promising way to incorporate behavioral heterogeneity.

Over the past decades, numerous methods have been proposed for generating household electricity load profiles. However, most stochastic models fail to represent realistic human behavior. The modeling tool LPG, developed by Noah Pflugradt [6], introduces an agent-based modeling approach grounded in the theory of human needs, initially proposed by D. Dörner [7]. In contrast to traditional probabilistic models, LPG assumes that individuals act to minimize deviation from an internally defined comfort level.

LPG simulates residential electricity and water demand by emulating occupant behavior. Each simulated household follows predefined routines and behavior patterns, resulting in dynamic appliance use and energy consumption. This allows LPG to generate plausible high-resolution load curves for electricity, hot and cold water, and occupant activities, which can be exported as CSV files for further analysis.

Among its core strengths, LPG allows detailed configuration of household, including family structure, employment status, lifestyle routines, appliance availability, and location (urban vs rural). The load profile is shaped by environmental, demographic, and behavioral inputs:

1. Environmental: season, ambient temperature, daylight duration;

2. Demographic: household size, working hours, hobbies;

3. Behavioral: meal schedules, comfort preferences, energy-saving awareness.

This level of customization supports the creation of plausible, scenario-specific profiles.

Most importantly, besides electricity consumption models of various appliances, the LPG includes a hot water demand model built on the same behavioral principle.

B. Agent-Based Modeling of Hot Water Demand

At the core of the LPG hot water demand model is the behavioral representation of users. The basic concept is that several devices offer different activities to the person. The person checks the expected satisfaction of desires from each activity and chooses the one that promises the greatest satisfaction at a given time. In turn, every desire is specified by its weight (importance), threshold, and decay of importance.

Thus, the load or consumption of an energy resource, such as hot water, by household $E(t)$ can be described as

$$E(t) = \sum_{n} \sum_{i} E_i\left(t, A_n^c(t)\right), \tag{1}$$

where n is the person number; i is the device number; A_n^c is the current activity of the n-th person; E_i is the energy consumption profile of the i-th device.

$$A^c(t) = \underset{\{A_j\}}{\arg\max}\, F_S\left(t, A_j(t), \{D_k, w_k, h_k\}\right), \tag{2}$$

where F_S is the satisfaction function; A_j is the j-th possible activity; D_k is the value of the k-th desire; w_k is the weight of the k-th desire; h_k is the threshold of the k-th desire.

The value of desire decays (expires) with time:

$$D(t+1) = e^{\left(\frac{\ln 0{,}5}{T_d}\right)} D(t) \tag{3}$$

where T_d is the decay (half-life) of the desire.

A person's desires also depend on many specific factors such as age, social status, habits, as well as the day of the week, the season, the general characteristics of the household, etc. And they need to be described carefully and completely to obtain a plausible load profile.

For the research topic, hot water flow rate is a key variable. While the maximum flow is determined by the capacity of the equipment, the actual rate is influenced by user desires and actions, such as when the tap is turned on or off, or how the water temperature is adjusted. Compared to other modelling approaches, LPG can plausibly reflect user behavior and generate high-resolution profiles of hot water demand over time, and importantly, it allows for the description of almost arbitrary households. This makes it a powerful tool for estimating household energy use, supporting the design of adaptive demand response and flexibility market mechanisms.

Hot water usage profiles exported from LPG in CSV format can serve as input data for the model of ESWH electricity consumption running on external simulation platforms. In our previous work [8], the OpenModelica modeling and simulation environment was used to implement a detailed physical model of ESWH and simulate its operation modes. This model incorporates heat transfer dynamics, thermostat control logic, and tank thermal capacity, allowing calculation of electricity consumption based on hot water draining rate, inlet water temperature, ambient losses, and heating element characteristics.

The combination of behavior-driven water usage modeling with thermophysical ESWH simulation enables an accurate estimation of ESWH electric energy consumption. First, a baseline consumption profile, i.e., without any external control, is calculated. The aggregate baseline of

multiple ESWHs represents a reference for flexibility interventions. This then enables the analysis of different energy management strategies for demand-side flexibility, and demand forecasting across different time scales and population groups for energy system operation optimization.

C. Validation of Hot Water Demand Profiles

A crucial step in the selection of a hot water demand model is the verification of the generated daily profiles, especially for water heated to 40°C. In the study, this is done by corresponding with measured electrical load profiles of real ESWHs. This should provide an assessment of the accuracy of the hot water demand model and identify fundamental deviations from the real ESWH operating modes if they are.

As shown in the study of energy demand and rhythms of everyday life [9], hot water consumption closely follows human activity rhythms, with peaks typically occurring in the morning 6:00–9:00 and evening 18:00–22:00. These periods coincide with daily routines such as showering and cooking, placing greater load on the energy system and emphasizing the need for demand-side flexibility. Consumer behavior, and in particular hygiene habits, play a key role in shaping hot water demand, which also exhibits clear seasonal variability. Real-world studies [10] indicate that the average European consumes 40–50 liters of hot water daily, with winter usage increasing by 15–25% compared to summer due to colder ambient temperatures.

For the verification, a household was selected from LPG that closely matches the real-life test case. Computer modeling of the ESWH operation was carried out in OpenModelica. At the first stage, daily hot water demand profiles generated by LPG were converted and imported into OpenModelica. Then, using a thermodynamic model that accurately represents the temperature stratification of water in the tank and the heating element control algorithms [8], the instantaneous load of the ESWH electricity consumption was calculated. Fig. 1 illustrates the historical and simulated electrical power and water demand profiles for winter and summer periods, respectively.

During the winter season, both electricity and hot water consumption exhibit a clear daily pattern, with distinct peaks reflecting household activity. Between 00:00 and 06:00, consumption remains low due to minimal activity. A sharp rise occurs between 06:00 and 07:00, corresponding to typical morning routines such as showering, cooking, and appliance use. During this time, normalized power reaches 60%, and hot water usage significantly increases. After the morning peak, demand gradually declines and remains relatively stable until 16:00–17:00, followed by another rise during evening hours 18:00–22:00. Although the model slightly underestimates the evening peak, the total simulated electricity consumption (9.93 kWh for the test ESWH with rated power of 1.2 kW) closely matches the experimentally measured value (10.02 kWh). Simulated hot water usage of 227.89 liters is also very close to observed data, demonstrating LPG model accuracy.

Winter patterns are easier to model due to their predictability and routine nature. In summer, electricity use is significantly lower due to reduced heat losses and changes in user behavior. While a morning peak is still present, it is less pronounced, reflecting altered daily routines and a lower need for hot water. Daily energy demand remains low, and the evening peak is also less defined. The model shows some deviation (measured electricity use for the test ESWH is 3.54 kWh compared to the simulated 4.51 kWh). This may be due to external factors not captured in the model. Summer hot water consumption drops to 187.70 liters, confirming seasonal variation. Despite this, it can be concluded that the ESWH model, together with hot water demand profiles generated by LPG, generally reflects all the key trends of real electrical load profiles.

III. CONSUMER AGGREGATION

A. Market Aggregation Processes

Electricity consumer aggregation for DR can be generally categorized into three main process groups:

1. Resource Formation for Aggregation: This includes the identification and registration of participants, grouping into aggregation units, and equipping installations with demand-side management devices;

2. Development of Aggregation Products: This involves registering aggregation units as resource entities, obtaining ancillary service provider status, collecting telemetry, forecasting individual consumption, and generating load profiles.

3. Market Implementation: This includes submitting planned schedules, bidding in auctions, and executing load control in response to Transmission System Operator dispatch instructions. It also covers metering and monitoring operations.

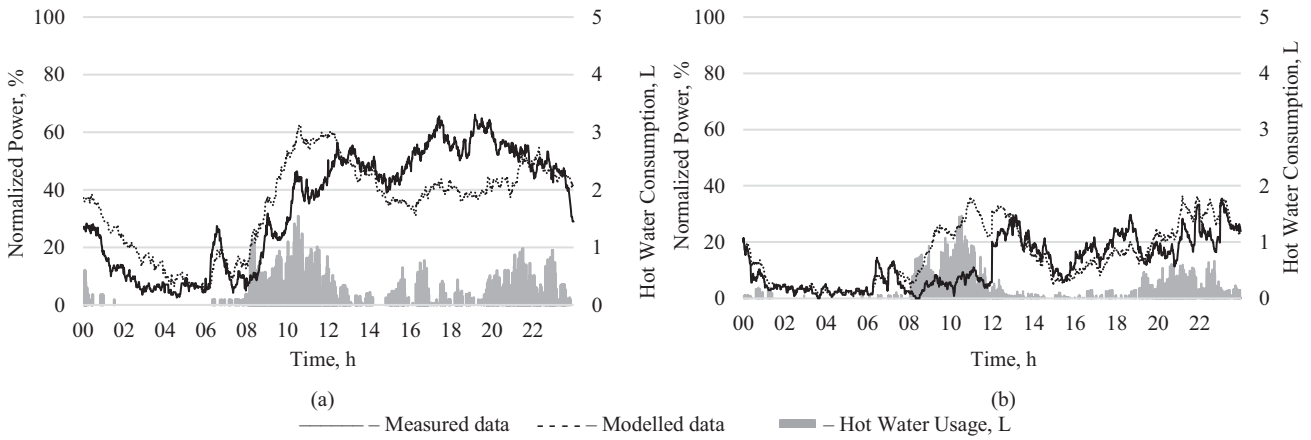


Fig. 1. Power and hot water demand profiles during winter (a) and summer (b)

The process of product development and market implementation is inherently cyclical, driven by fluctuations in demand, price, and system needs. Aggregator must continuously update forecasts, adjust load schedules, and resubmit bids accordingly. At this stage, the use of generated water demand profiles, which are generally stochastic but consistent with real experience, as well as derived load profiles is crucial for realistic modeling of consumers DR. These profiles must capture the variability in baseline demand and enable accurate simulation of ESWH flexibility, which depends on thermal state shaped by past water use. Second, they must support accurate modeling of ESWH response to control signals, as flexibility availability depends on the current thermal state defined by tank temperature, thermostat setting, and also water drain. Equally important, plausible and diverse profiles are essential for quantifying rebound effects, which are governed by the accumulated thermal deficit and its heterogeneity across ESWHs, ultimately determining the shape, amplitude, and duration of load recovery peaks.

B. Consumer Group Profiling

A fundamental prerequisite for the effective implementation of demand-side management programs is a deep understanding of end-user behavioral patterns. This extends beyond the technical characteristics of ESWH to include usage behavior shaped by socio-demographic factors, daily routines, and work schedules.

With the LPG modeling tool, 25 daily ESWH consumption profiles were stochastically generated for 6 types of households, representing a range of socio-economic scenarios. Each profile reflects the behavior of an individual user during a typical working day in the autumn season, in the absence of external DR control signals.

To obtain representative daily electrical load profiles, the profiles within each group were averaged. All calculations were performed in relative units, enabling analysis of load dynamics independent of absolute power. Each value was normalized to the rated power of the ESWH, resulting in a time-dependent normalized load profile, $P_n(t)$, expressed as a percentage.

$$P_n(t) = \frac{1}{N}\sum_{i=1}^{N}\frac{P_i(t)}{P_{r,i}} \times 100\%, \qquad (4)$$

where $P_i(t)$ is the power consumption of the i-th ESWH at time t, in watts; $P_{r,i}$ is the rated power of the i-th ESWH, in watts; N is the number of individual profiles in the group; t is time of day, in minutes (from 1 to 1440).

An analysis of the normalized ESWH load profiles for various socio-demographic household types reveals clearly defined behavioral patterns. Households with two working adults (Fig. 2a) exhibit a distinctive double-peak profile, with morning and evening peaks aligned with standard working hours. Unemployed couples (Fig. 2b) show a more temporally extended and less intense patterns of use. The highest overall consumption is observed in families with three children and working adults (Fig. 2c), characterized by a broad evening plateau, which presents a promising opportunity for load shifting. A similar household without employed members (Fig. 2d) demonstrates significant midday activity, indicating a potential flexibility window during periods of renewable energy surplus. Single working adults (Fig. 2e) have less distinct peaks, while unemployed singles (Fig. 2f) display smoother, more predictable profiles with activity concentrated in the first half of the day.

Classifying consumers by household type allows aggregators to design targeted load optimization strategies that integrate both technical feasibility and behavioral compliance. In this context, each household type should be treated as a distinct response unit, capable of engaging with dispatcher signals differently depending on the time of day and user habits.

C. Baseline Load Modeling for Aggregated Loads

To analyze the operation of a portfolio under aggregated load management, a coefficient of activity K_{act} is introduced. It is equal to the normalized sum of load profiles of all ESWH participating in DR. That is, $K_{act}(t)$ represents the share of the aggregated rated (installed) capacity of the ESWHs that are actively used at a given time t.

$$K_{act}(t) = \frac{\sum_{i=1}^{N_{act}} P_{act,i}(t)}{\sum_{i=1}^{N} P_{r,i}} \times 100\%, \qquad (5)$$

where $P_{act,i}(t)$ is the actual power of ESWH in the portfolio, which is in active operation at time t.

Fig. 3 shows $K_{act}(t)$ for a portfolio of 425 simulated consumers covering a broad spectrum of socio-demographic categories, ranging from families with children to retirees

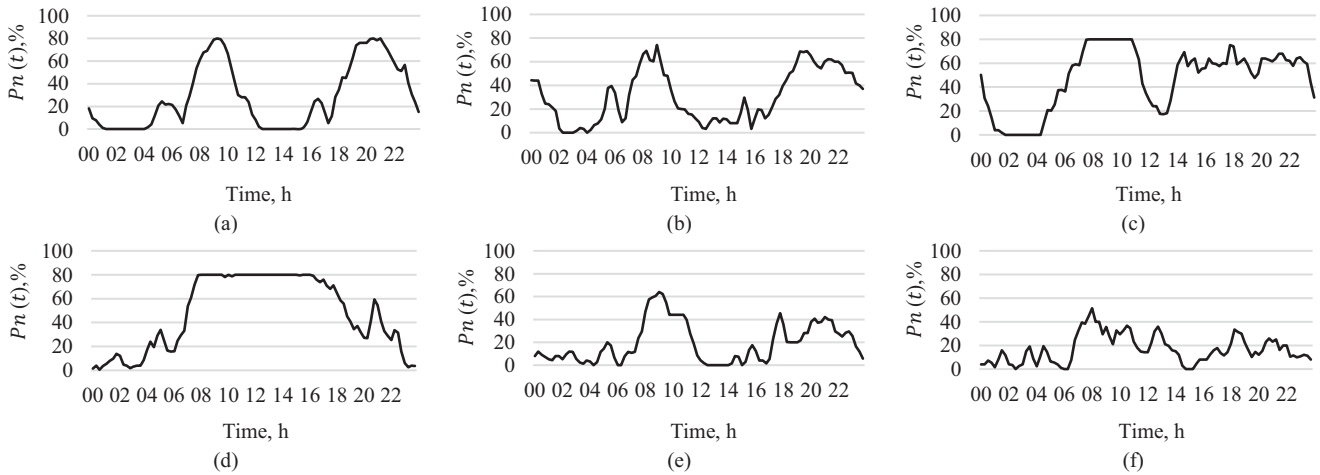


Fig. 2. Normalized ESWH load profiles: (a) Two working adults; (b) Two unemployed adults; (c) Two working adults with three children; (d) Two unemployed adults with three children; (e) Single working adult; (f) Single unemployed adult

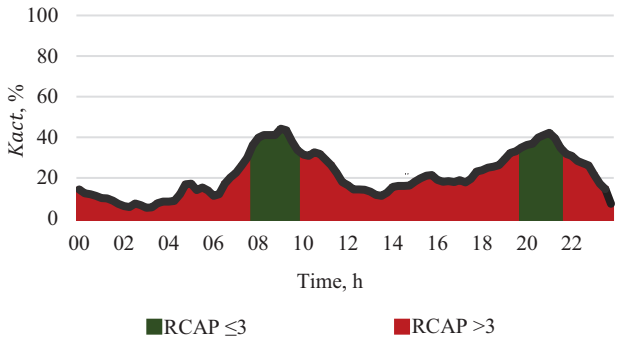


Fig. 3. Coefficient of activity of the portfolio

and shift-working households. In total, 12 household types were modeled, each represented by 25 to 50 individual consumption profiles. Each household was modeled with identical ESWH units rated at 1.2 kW and 80 liters. The resulting aggregate profile corresponds to a weekday in the autumn period.

The coefficient of activity of this portfolio exhibits a smoother consumption pattern compared to individual curves, reflected by a reduced peak-to-average ratio due to the averaging of individual peaks. The load factor is higher, indicating more consistent utilization throughout the day, while the form factor decreases as a result of more uniform load distribution.

To reliably deliver a certain power level as a flexibility service in the market, an ESWH aggregator must have some reserves in the portfolio. Lower levels of K_{act} require proportionally higher rates of reserve to maintain equivalent service delivery. This can be described by the RCAP K_{res} defined as:

$$K_{res}(t) = \frac{1}{K_{act}(t)} \quad (6)$$

This study finds that aggregator performance is optimal when RCAP ≤ 3. RCAP enables quantitative assessment of flexibility, availability, and reliability, guiding market participation strategies and supporting segmentation of consumer groups for targeted control within the portfolio.

A detailed analysis of the aggregated ESWH load profile reveals key insights into consumption patterns and flexibility. From 05:00, user activity increases gradually, peaking between 09:00–10:00, when up to 45% of the total ESWH capacity is used. This synchronized morning peak, driven by workday routines, shows the highest load utilization and lowest RCAP ≈ 2.5.

After the peak, a sharp decline in activity offers opportunities for short-term load reduction or shifting with minimal comfort impact. From 11:00–16:30, ESWH usage stays low at 15–20%, resulting in a high RCAP 5–6, allowing for increased load by providing preheating, which functionally acts as an energy storage device.

In the evening, demand rises gradually, peaking from 19:30–20:30, so RCAP ≈ 2,5, but increased variability in user routines reduces predictability and complicates management. Overnight (00:00–05:00), demand drops to 5–10%, pushing RCAP up to 10. Thus, this period is ideal for absorbing excess generation and smoothing the nighttime load curve through incentives.

Conclusions

This study demonstrates that integrating agent-based hot water usage modeling with thermophysical simulation of ESWHs produces realistic, validated load profiles across diverse socio-demographic groups. The introduction of the RCAP metric provides a quantifiable measure of portfolio flexibility, availability, and reliability, supporting the identification of optimal market participation conditions.

This capability allows aggregators and energy service companies to align operational strategies with broader innovation objectives, such as developing scalable demand response portfolios and enhancing customer engagement. The modeling outputs can inform innovation strategies in real-world DR programs by revealing when and how to activate specific consumer segments, designing targeted control schemes, and structuring market offers that maximize both technical performance and economic value.

For stakeholders, these insights extend beyond system operation, offering guidance for long-term portfolio planning. By linking behavioral diversity with measurable technical potential, the proposed framework supports strategic decision-making in the evolving flexibility services market. The model assumes uniform technical specifications for all ESWHs and ideal communication reliability between devices and control systems. It does not account for heterogeneous installation conditions, varying user comfort preferences, or potential delays in demand response activation. Future research should include field validation under real operating conditions and explore real-time optimization algorithms for portfolio management.

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Left-Handed Planar Metamaterials Formed by Nanostructured TiO_2-Ferrite System for Microwave Circuits Components

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***Abstract*—This work presents an analysis of the electromagnetic properties of a planar metamaterial composed of elements made from two natural materials: ferrite and a nanostructured TiO_2-based semiconductor exhibiting memristive behavior. In the microwave frequency range, the structure exhibits distinct transparency peaks/zones associated with its left-handed characteristics (double-negative, DNG-peaks/zones). The possibility of controlling these DNG transparency peaks/zones using static magnetic and electric fields is investigated. The findings highlight the potential of such metamaterials for the design and implementation of tunable components in microwave circuits.**

Keywords—planar metamaterial, left-handed medium, inverse split-ring resonator, memristor, microstrip transmission line

I. Introduction

It is well established that left-handed media (metamaterials/structures) are promising candidates for use in microwave-frequency circuit elements [1–7], owing to their fast-switching capabilities and relatively simple mechanisms for external control. Successful research in this field has been ongoing for over a decade. These structures are typically composed of alternating elements exhibiting metallic-like dielectric permittivity and magneto-resonant magnetic permeability. In the frequency range where the real parts of both constitutive parameters become negative, the structure can support wave propagation (i.e., become transparent), provided that the wavelength within the metamaterial significantly exceeds the characteristic size of its constituent elements [4–7].

Thus, a prototype of the fast-response [4–6], electrically and magnetically tunable passive microwave element is obtained. The primary obstacle to the practical application of such left-handed metamaterials in high-frequency technologies lies in the excessively large imaginary parts of the magnetic and dielectric permittivity of the constitutive parameters of natural materials that form them, which lead to significant electromagnetic energy losses. Consequently, key research directions in this field include: (a) the search for new low-loss natural materials, and (b) the optimization of the spatial configuration of such metamaterials.

In our recent study [4], we proposed a new type of left-handed metamaterial based on a nanostructured memristor. It was demonstrated that a bulk (3D) metamaterial composed of alternating ferrite elements and nanostructured TiO_2-based semiconductor (memristor) elements can exhibit left-handed properties in the microwave frequency range. Owing to the short relaxation times of the entire structure, such a metamaterial is a promising candidate for use as an electrically and magnetically tunable component of microwave band devices. The advantages of this new type of metamaterial stem primarily from the unique properties of TiO_2, which are manifested at nanometer sizes (on the order of 10^1 –10^4 nanometers [4]).

This paper presents further results in this research direction, specifically analyzing two structures based on the previously proposed model [4] but implemented in a more compact 2D geometry. This planar configuration reduces overall electromagnetic losses compared to the 3D case described in [4] and offers improved technological feasibility for device fabrication [10,11].

II. Structures Description

The structures investigated in this manuscript are as follows:

- Structure #1 (Fig.1, a), referred to as the "Strips Shaped Metamaterial," consists of a microstrip transmission line (shown in yellow color in Fig.1, a) placed on top of a dielectric substrate (shown in brown color in Fig.1, a) with metallization below. The permittivity of the substrate is $\varepsilon' = 2.2$. The width of the microstrip line is $a = 1.12$ mm. The structure #1 is bonded to the substrate on both sides, with the same thickness as the substrate ($d = 0.381$ mm), and is composed of alternating layers of ferrite and TiO_2. Structure #1 consists of 10 bi-elements. The width of each bi-element is $b = 15$ mm. The length of each ferrite element (shown in blue in the figure) is $c_f = 0.2$ mm, and the length of each TiO_2 element (shown in black in the figure) is $c_s = 50$ nm.

- Structure #2 (Fig.1, b), referred to as the "Ring-Shaped Metamaterial," is based on a microstrip transmission line and consists of several unit cells (in our case, 7). The microstrip line manufactured from TiO_2, with a width of $a = 1.12$ mm, is located on the top surface of a substrate with thickness $d = 0.381$ mm and permittivity $\varepsilon' = 2.2$. Periodically spaced linear gaps with a width of $\delta = 100$ mkm are etched into the microstrip line. Directly beneath these linear gaps, on the bottom surface of the substrate, ring-shaped gaps are formed

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in a TiO_2 layer that coats the substrate. These ring-shaped gaps constitute Inverse Split-Ring Resonators (ISRRs) [8]. The ISRR parameters are radius $r = 3$ mm, ring width $s = 0.2$ mm, and center-to-center spacing between adjacent ISRRs $d = 0.5$ mm (the thickness of each TiO_2 layer equals 35mkm).

The electromagnetic wave propagates along the strip lines shown in the Fig. 1 for both structures, in the direction of the OY axis, from left to right.

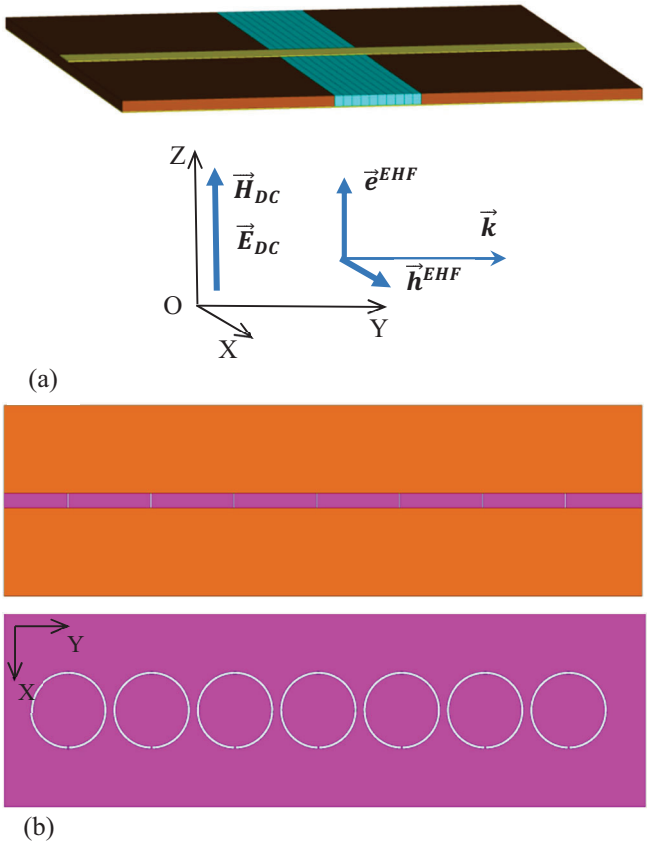


Fig. 1. Schematics of the studied structures. (a) structure #1, "Strips Shaped Metamaterial"; (b) structure #2 "Ring-Shaped Metamaterial"

III. Discussion

Fig. 2 (a, b) shows the transmission spectra obtained in the frequency ranges of 5–25 GHz for structure #1 and 1–10 GHz for structure #2. From these spectra, it is evident that in the frequency bands where the real parts of the permittivity of TiO_2 and the permeability of the ferrite are simultaneously negative, pronounced transmission peaks (transparency zones) are observed. These are indicated by vertical blue arrows in Fig. 2 (a, b), with the central resonance frequencies $f_{res}^{DNG} = 15.24$ GHz and 4.05 GHz respectively.

To clarify the nature of these peaks, spatial distributions of the electromagnetic field were simulated at the corresponding frequencies (Fig. 3, a, b). For each of the analyzed structures, the incident wave propagates from left to right. However, the spatial field profiles reveal the presence of a backward wave inside each structure, i.e., a wave (phase front) propagating from right to left.

Let us first analyze the spectrum of structure #1.

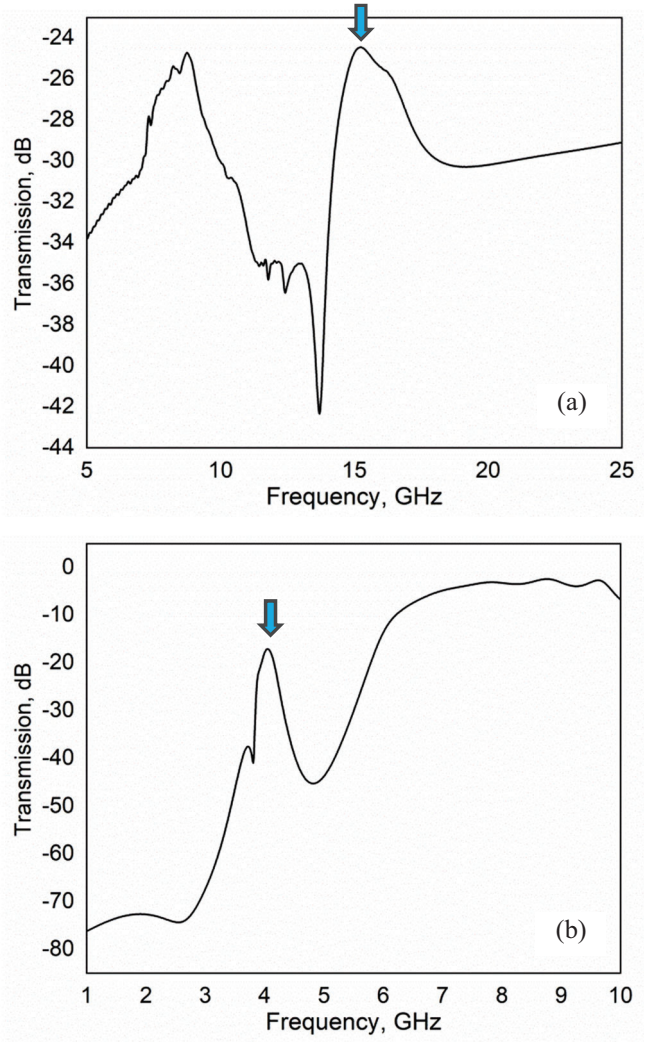


Fig. 2. (a, b) Transmission spectra corresponding to structures #1 and #2, respectively

The appearance of the transparency peak in the spectrum must be attributed to the fact that, at these frequencies, the real parts of both the dielectric permittivity and the magnetic permeability of the structure become simultaneously negative within the same frequency range, near the peak's resonance frequency 15.24 GHz. In detail, the method for calculating the frequency dependencies of these parameters for TiO_2 and ferrite separately, as well as for the entire composite left-handed structure, is presented in [4].

To investigate the effect of the gap width δ on the resonant frequency of the DNG peak in structure #2, the corresponding dependence was obtained and is presented in Fig. 4. The parameter δ was varied in the range of 100 to 700 μm.

For the cases analyzed in this work, our calculations confirm that the observed transparency peaks (like those in [4]) arise from the manifestation of left-handed properties in the studied structures. These peaks occur in a frequency range where the characteristic impedance of the left-handed medium approaches 1 and, therefore, are classified as Double Negative (DNG) zones/peaks. The physical origin of these peaks is associated with a specifically tailored phase

relationship between multiple reflected waves in the spatially periodic structure [1–7].

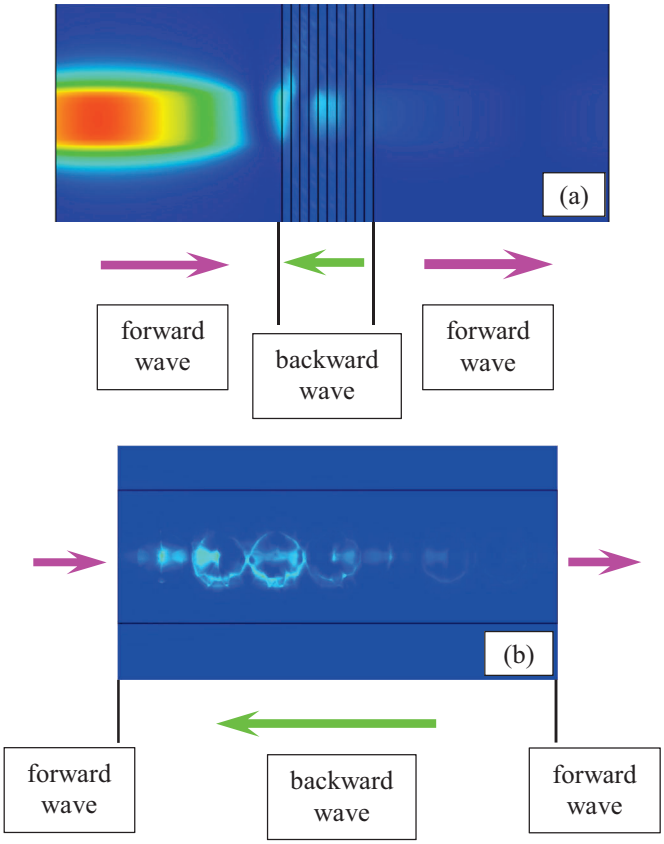


Fig. 3. (a, b) Spatial distributions of the electromagnetic field for structures #1 and #2, respectively

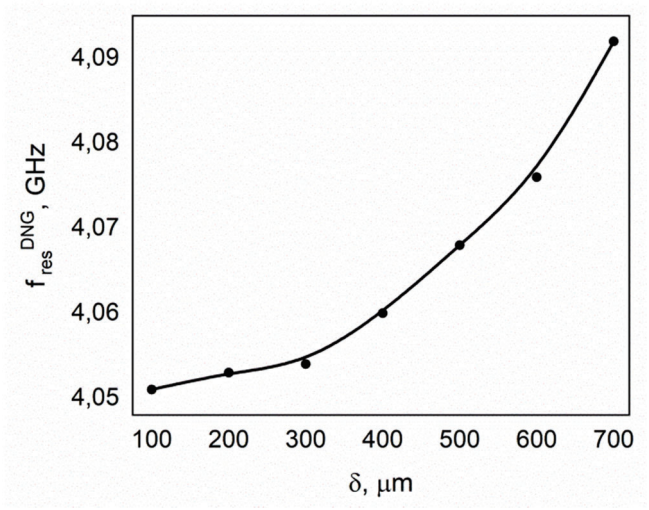
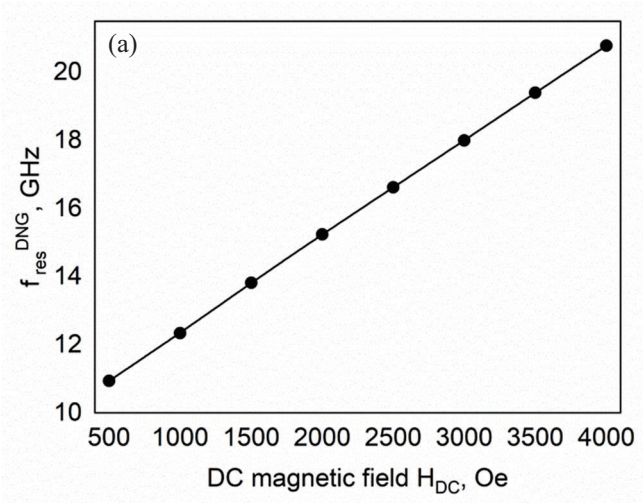


Fig. 4. Shift of the DNG peak as a function of gap width δ for stricture #2

Structure #2 may initially appear more technologically complex than structure #1; however, it offers several notable advantages. Most importantly, the fabrication (deposition) of the memristor elements is significantly simpler in structure #2 compared to structure #1. In the case of structure #1, each subsequent TiO_2 layer must be deposited on top of the previous one. Since the total number of layers can reach up to 10^2, and the deposition of each layer can take up to 1-2 days, the overall fabrication time for structure #1 becomes substantially longer than that for structure #2.

Structure #2 is based on the concept of a metamaterial previously discussed in [8]. In that paper, the negative effective permittivity of the metamaterial was achieved by incorporating varactor diodes into each ring-shaped element of the so-called split-ring resonator (SRR) [9], with the varactor acting as an electrically tunable component. In the present design, however, the entire TiO_2 memristor semiconductor layer assumes this role, making the resulting integrated system a more technologically advantageous solution. The negative effective permeability of the structure, meanwhile, is provided by its geometric configuration (the chain of rings) characteristic of traditional magnetic metamaterials.

Note that, as shown in [4], the electromagnetic properties of the studied structures (in particular, their DNG peaks) can be effectively tuned by both a static external magnetic field and a static external electric field. In particular, controlling metamaterial characteristics using a static magnetic field is a known and practically significant task [1–7]. Following the above-mentioned algorithm [4], we evaluated the shift in the resonance frequency (f_{res}^{DNG}) of the DNG peak (Fig. 5, a) as a function of the applied DC magnetic field H_{DC}.



The linear behavior of the resulting curve reflects the direct proportionality between f_{res}^{DNG} in accordance with motion equations [4,5].

Notably, structure #1 exhibits a substantial derivative value: $df_{res}^{DNG}/dH_{DC} = 2.96$ MHz/Oe . This value is approximately 20% higher than that reported in [4] for a considerably more technologically complex structure. This result highlights the potential of using a static magnetic field to achieve practical and efficient control of metamaterial properties.

We now analyze the possibility of tuning the DNG peak in structure #1 using a static electric field. As seen in Fig. 5, b, the dependence of the peak frequency on the applied electric field is nearly linear, similar to the behavior reported in [4]. This is attributed to the approximately linear

current-voltage (I–V) characteristic of the TiO_2-based memristor.

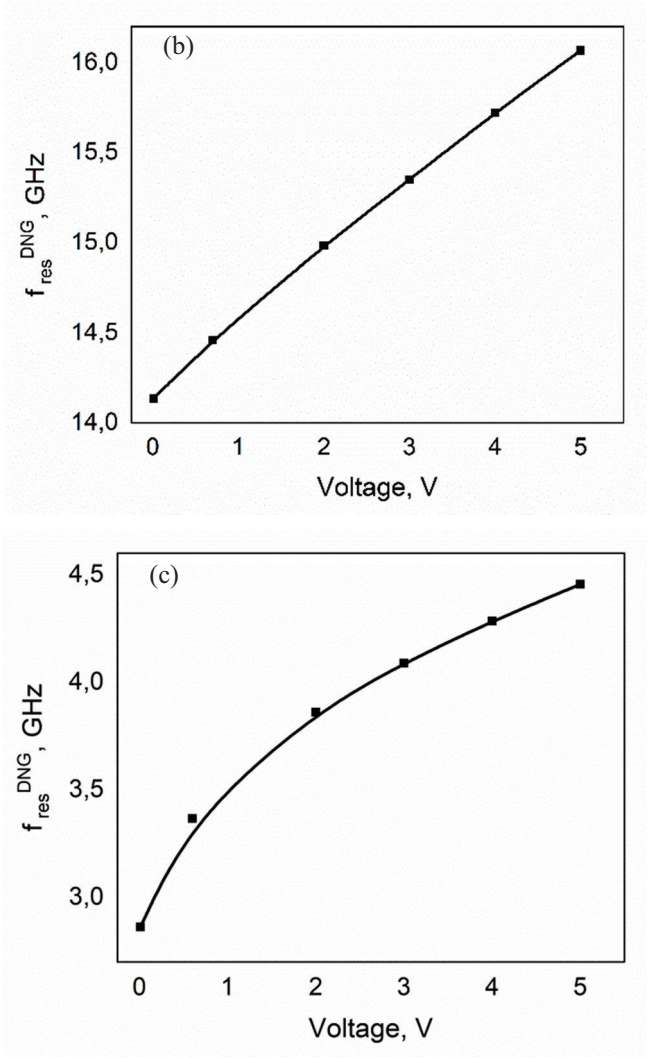


Fig. 5. DNG peaks behavior under external fields: (a) Variation of the DNG peak position with applied static magnetic field (H_{DC}) for structure #1; (b) Shift of the DNG peak as a function of applied static electric field / voltage (U_{DC}) for structure #1, H_{DC}=2000 Oe; (c) Shift of the DNG peak as a function of applied static electric field / voltage (U_{DC}) for structure #2

The derivative df_{res}^{DNG}/dU_{DC}, which characterizes the tuning efficiency of the resonance frequency, has a value comparable to that observed in the more technologically complex structure described in [4]: $df_{res}^{DNG}/dU_{DC} = 370$ MHz/V.

For structure #2, the dependence of the DNG peak frequency on the applied voltage is markedly nonlinear (Fig. 5, c). This behavior is due to the strongly nonlinear current–voltage characteristic of the TiO_2 memristor [4]. As a result, we observe in the higher voltage range $U_{DC} = (4-5)V$, the tuning rate is relatively moderate: $df_{res}^{DNG}/dU_{DC} \approx 170$ MHz/V. However, at lower voltages, in the range $U_{DC} = (0.01-1)$V, the derivative significantly increases: $df_{res}^{DNG}/dU_{DC} \approx 630$ MHz/V.

This analysis leads to the following conclusions:

IV. Conclisions

In summary:

1. The electrodynamic properties of a planar metamaterial based on a semiconductor nanostructure of the TiO_2 memristor type, as well as a composite structure formed by alternating planar (quasi-2D) ferrite and TiO_2 elements, have been numerically analyzed. It has been shown that in the microwave range, these structures exhibit transmission bands or peaks in the spectrum, caused by their left-handed characteristics.

2. The feasibility of precise tuning of the DNG peak frequency in the studied left-handed structures using both static magnetic and static electric fields has been demonstrated through numerical simulations.

3. A technological implementation of the proposed TiO_2 - ferrite nanostructured system has been suggested in the form of two configurations of promising electrically tunable components for microwave-frequency applications.

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Signal Coding and Network Technologies for Signal Processing in the Method of Incoherent Scatter of Radio Waves

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***Abstract*—The procedure of choice and application of complex sounding signals for ionospheric studies by the radar method of incoherent scatter (IS) of radio waves is considered. It is shown that the solution to the problem of reliable estimation of ionospheric parameters in an extended altitude interval with a sufficiently short time interval consists in improving the structure of sounding signals. It is proposed to modernize the process of pulse coding to obtain signals with specified correlation properties that make it possible calculating the ionospheric parameters both in the lower and upper parts of the ionosphere during a single measurement session. When synthesizing individual elements of the sounding signal, not only a change in the direction of circular rotation of the polarization plane of the radio wave is used, but also a change in the frequency of its radiation. For this purpose, the sounding signal should consist of groups of discrete-frequency elements representing radio pulses with different parameters. We propose to use an identifying neural network that enables estimating the parameters of ionospheric plasma in real time based on the obtained experimental autocorrelation functions of the IS signal.**

Keywords—incoherent scatter method, signals and sounding modes, coding of radio pulses, neural network signal processing

I. Introduction

Advances in understanding the physical processes in near-Earth space have been achieved through the use of new research methods, among which the incoherent scatter (IS) method occupies a special place. A detailed review of the theoretical aspects and practical use of this method, as well as the types and features of radars designed for these purposes, is presented in numerous publications.

Using the IS method, it became possible to determine simultaneously a number of ionospheric parameters, such as electron density, electron and ion temperatures, ion composition, plasma drift velocity, and many others. It is with the use of this method that the IS radars, created by Kharkiv polytechnic scientists, operate with a unique 100-meter (in diameter) fixed zenith antenna and a fully steerable 25-meter antenna [1, 2]. The results obtained with their use for more than half a century have been of great fundamental and applied importance for world science. For example, information on the state and dynamics of ionospheric plasma is extremely necessary both from the point of view of deepening knowledge about near-Earth space and for solving many practical problems in industries covering the activities of space weather systems, satellite communication systems, and object positioning systems in space, which indicates the relevance of obtaining this information at the present stage.

Traditional radar uses sounding signals that enable determining the coordinates and speed of objects that are point targets in outer space and airspace, on the surface of the earth and the water surface [3–6]. In the case of measurements using the IS method, the ionospheric plasma is a spatially distributed target, and the requirements for the type of sounding signal are significantly more complex than in traditional radiolocation. Therefore, a number of foreign authors proposed different signal options and sounding modes [7–9], which were subsequently improved by our Ukrainian scientists [10, 11].

In the sequel, the process of choice and using complex signals for studying geospace was continued. It was emphasized that in the process of synthesizing optimal signals for ionospheric sounding, radar equipment, sounding modes and processors for IS signal processing should be considered as interconnected elements of the system. Accordingly, the objective of the presented studies is to improve the sounding process and subsequent radar processing of the IS signal to study the state and dynamics of the ionosphere in an extended altitude interval with sufficient range (height when vertical sounding) resolution. Since the main carrier of information on the state of the ionospheric plasma are the experimentally obtained statistical characteristics of the scattered signal (in particular, the autocorrelation function (ACF)), we consider the case of a more accurate determination of these characteristics by using elements of the complex sounding radio signal coded by several parameters.

II. Analysis of the Pulse Signal Sounding Mode

Since the ionospheric plasma has a structure that changes with altitude, it is proposed to use different modes in the process of conducting experiments to determine its characteristics. The implementation of such modes with a gradual complication of probing signals began more than 50 years ago [12–14]. Recently, similar approaches have been proposed for studying the ionosphere at altitudes <400 km, where its parameters change significantly with altitude. In this case, the modes of radiation of multi-element pulsed sounding signals (Fig. 1, a) are used, making it possible to obtain the ordinates of the normalized ACF of the scattered signal with high resolution both in altitude (Δh) and in correlation delay ($\Delta\tau$). This is achieved

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by using manipulation of the direction of circular rotation of the polarization plane of the radio wave during its radiation. However, these pulses enable to obtain only the right-hand side of ACF, i.e., in the interval of correlation delay from τ_1 to τ_2 (Fig. 1, c).

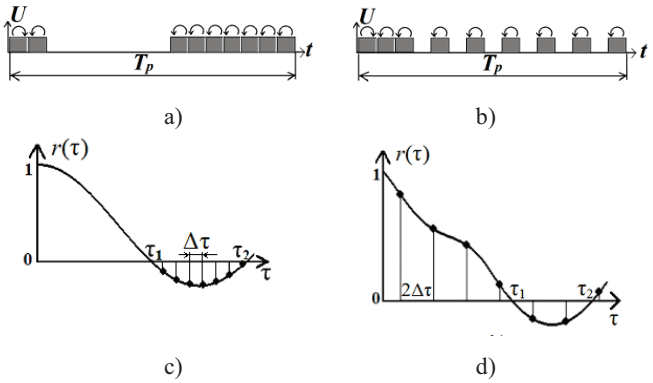


Fig. 1. Envelopes of pulse elements of the sounding signals (a, b) and the normalized ACFs of the scattered signal obtained with their help for two sounding modes (c, d). The arrows conventionally indicate the direction of circular polarization of each emitted pulse element.

For the study of the upper altitude range of the ionosphere, an algorithm for searching for the structure of the sounding signal was developed and a new arrangement of elements in the sounding pulse was obtained [15, 16] (Fig. 1, b). This option also enables obtaining the ACF ordinates with a high resolution along the height. However, since the spectrum of the scattered signal expands with increasing height due to the presence of light ions in the upper ionosphere, and the ACF acquires a complex shape, it is necessary that the ACF ordinates be obtained in the delay interval from 0 to τ_2 (Fig. 1, d). With the same total duration of the sounding pulse, this option for the arrangement of the coded elements provides a 2-fold worse resolution in correlation delay: $2\Delta\tau$.

Below, we propose such a variant in the algorithm for searching for combinations of coded elements of the probing signal, in order to simultaneously (in one measurement session) implement both of these modes for determining the ordinates of the ACF of the IS signal while ensuring the maximum acceptable resolution both in height and in correlation delay for the simultaneous study of the lower and upper altitude ranges of the ionosphere.

III. Improving the Structure of the Sounding Signal

The results of calculations using a computer program of a number of combinations of coded elements in the structure of a sounding signal with total duration T_p are given in [16]. The main rule in their search was that the most active element should be located at the beginning of the coded pulsing.

In the process of calculating the ordinates of the IS signal ACF, the echo signal from this element is involved in multiplication with the echo signals from all subsequent elements shifted relative to the first element by $\tau_k = k\Delta\tau$ (k is the number of the correlation delay). The results of the multiplications for each altitude section with a length of $\Delta h = c\tau_{el}/2$ and a center at height $h_0 = ct_0/2$, which this first element has reached, are statistically accumulated in the working memory. Here c is the speed of light, τ_{el} is the duration of the pulse element of the sounding signal.

The signals proposed in [15, 16] enable using only 8 elements as active ones for calculating up to 7 ACF ordinates (Fig. 1, a, b), while the total duration of the signal allows containing up to 15 elements. In addition, it is necessary to use two coded sequences in turn, corresponding to the process of studying the lower or upper altitude range.

We proposed to modernize the process of synthesis of coded elements of the sounding pulse signal, namely:

– to fill completely the pulse with active elements, for example, with different carrier frequencies (ensuring that the spectra of the scattering signals caused by them do not overlap) and with different directions of circular polarization of the emitted wave;

– thereby ensure the optimal operating mode of the radar transmitters;

– have the ability to determine the ACF ordinates in each radar scan with satisfactory accuracy when receiving scattered signals simultaneously from the lower and upper altitude ranges.

The results of the synthesis of multi-element coded signals that satisfy these requirements are presented in Fig. 2 as one of the fragments. The black elements are supposed to be emitted at the frequency f_1, and the gray ones at the frequency f_2.

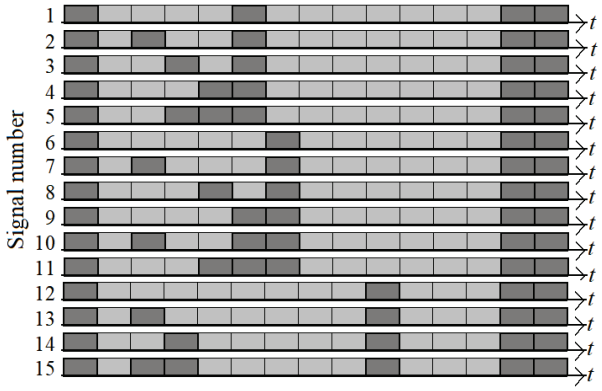


Fig. 2. Results of the search for multi-element coded signals that provide simultaneous calculation of up to 13 ordinates of the scattered signal ACF.

Table 1 illustrates the analysis of the capabilities of these signals.

TABLE I. Information on the Calculation of ACF Ordinates.

Signal number (according to Fig. 2)	**1**	**2**	**3**	**4**	**5**	**6**	**7**	**8**	**9**	**10**	**11**	**12**	**13**	**14**	**15**
Total number of multiplications	7	12	11	7	9	7	12	11	7	13	9	7	12	11	15
Number of ACF ordinates obtained	7	11	10	7	8	7	11	11	7	12	9	7	11	10	13

The first line is the numbers of the found signal variants, starting from the top one in Fig. 2. The second line is the total number of mutual multiplications between elements within the implemented signal. The third line is

the number of ACF ordinates that can be obtained with these multiplications.

For greater clarity, the information on this table is presented graphically in Fig. 3, a. When comparing these 15 versions of sounding signals, one can observe different cases of the location of the calculated ACF ordinates along the axis of correlation delays. Larger dots indicate better calculation statistics (these ordinates were obtained twice).

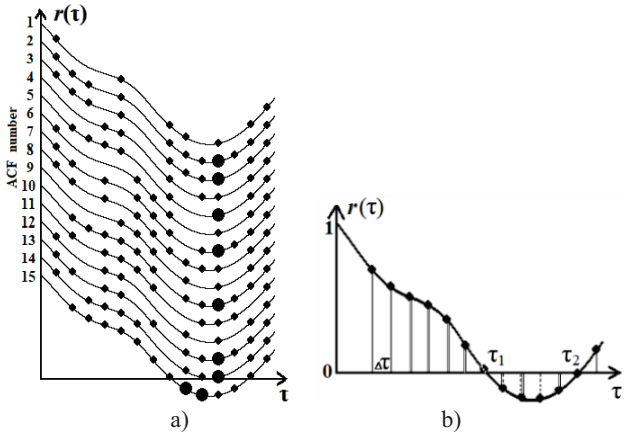


Fig. 3. Versions of the location of the ordinates of the scattered signal ACF (a) when using the sounding signals shown in Fig. 2, and the lower (15th) version (b) as the the optimal one.

Based on the information presented, coded signal number 15 is proposed for subsequent use (see Fig. 2). It makes it possible to obtain a sufficient number of uniformly distributed ACF ordinates (Fig. 3, b), and the small step of their location along the correlation delay axis satisfies the requirements for calculating ionospheric parameters from the shape of the ACF of signals scattered by plasma in both the lower and upper altitude ranges.

Thus, as a result of such a choice of the arrangement of the coded elements of the sounding signal, it will look in its final form upon emission as shown in Fig. 4.

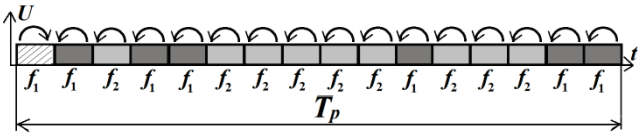


Fig. 4. Envelopes of coded elements of the sounding pulse.

The first element of the sounding signal will have a right-hand direction of rotation of the plane of polarization of the radio wave, as in [16], which will make it possible to obtain the power of the IS signal in a separate receiving channel for subsequent normalization of the ACF ordinates. All other elements intended for calculating the correlation coefficients will have a left-hand direction of rotation. The duration of each code element $\tau_{el} = 30$ μs that we have chosen provides a quite acceptable altitude resolution $\Delta h = c\tau_{el}/2 \approx 4.5$ km with a total length of the sounding pulse $T_p = 30 \times 16 = 480$ μs [17, 18].

The process of propagation in space of the elements of the coded pulse separately at frequencies f_1 and f_2 is shown in Fig. 5 and Fig. 6. The figures demonstrate the possibility of obtaining the ACF ordinates by multiplying not only the echo signal from the first element of the pulse by the echo signals from the rest of elements, but also by additional mutual multiplications between the echo signals from these rest code elements.

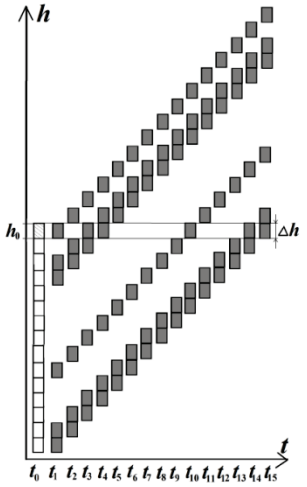


Fig. 5. Process of propagation in space of the coded pulse elements at frequency f_1.

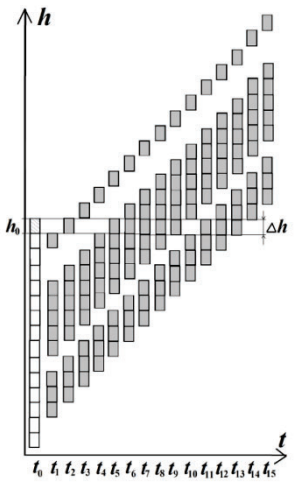


Fig. 6. Process of propagation in space of the coded pulse elements at frequencies f_1 and f_2.

In the final form, the obtained ordinates of the normalized ACF of the scattered signal will look like in Fig. 3, b. Single vertical lines on the graph mean that these ordinates (numbers 2, 3, 9, 13, and 14) will be obtained at frequency f_1 using multiplications of the echo signals from the first element with those from the other pulse elements. Ordinates numbered 4, 5, 6, 7, 10, and 12 (double lines) will be obtained at the same frequency, but using mutual multiplication of echo signals from pulse elements without involving the first element. Ordinates numbered 8, 9, 10, and 11 (dashed lines) will be obtained at frequency f_2.

The calculation of the ACF ordinates is performed in specialized devices – correlators, the structure of which depends on the processing option [16, 19].

IV. Network Technologies in the Incoherent Scatter Method

The calculation procedure in the IS method consists of calculating the parameters of the ionospheric plasma using the experimentally obtained ACF of the scattered signal. The peculiarity is that due to the absence of direct analytical formulas, this calculation is based on the iterative procedure of solving the inverse problem. Its essence lies in comparing each ACF obtained at each studied altitude section with a set of theoretical analogues, previously calculated according to physical concepts of the nature of the ionospheric plasma. These analogues are stored in the computer memory (N model correlation functions with K ordinates in each of them). When using them alternately with the use of the coordinate descent method and the concomitant use of the convolution operation (to take into account the current characteristics of the ACF of the filter of the receiving system of the IS radar), the surface of the residual function is formed. This is a rather slow calculation procedure, and the most time-consuming operations are operations of matrix multiplication (convolution). Therefore, the goal of further improving this procedure for real-time operation is to increase its speed by parallelizing the ACF identification process using theoretical analogs.

A significant number of publications are devoted to solving such problems, and among them the principle of using artificial neural networks is of interest. Interesting results are given in the studies of T. Kohonen [20] on the example of self-organizing structures used in recognizing images represented by vector quantities. In addition, in [21], there is a description of a significant number of features that the problem being solved must have so that the use of convolutional neural networks to be justified. The main ones, taking into account our conditions, include the following:

– there is no strict mathematical expression, but a sufficient amount of experimental data has been accumulated;

– the processing process is characterized by large volumes of information;

– data received under ionospheric interference conditions are significantly noisy.

As we can see, according to these features, neural networks can be used to recognize the ACF shapes of the IS signal, during which the parameters of the ionospheric environment are assessed using correlation functions.

Let us formulate the tasks of statistical processing of ionospheric information as follows. It is necessary to determine the unknown parameter θ, characterizing the state of the ionospheric plasma, having information about the K ordinates of the experimental ACF $r^{\mathrm{e}}(k; \boldsymbol{\theta})$, which are obtained during the processing of the scattering signal. This approach involves the algorithm of functioning of associative memory and can be organized using a neural network. In our case, there is a certain set of reference, theoretical correlation functions, among which the network should be able to select the reference sample, most similar to the ACF obtained in the experiment.

To implement this network, we will define the algorithm of its functioning for our case. It should correspond to the calculation of the elements of the residual function when solving the difference functional of the following type:

$$\chi_n^2(\boldsymbol{\theta}) = \sum_{k=2}^{14} \left[r^{\mathrm{e}}(k; \boldsymbol{\theta}) - \langle r_n^{\mathrm{t}}(k; \boldsymbol{\theta}) \otimes r^{\mathrm{f}}(k) \rangle \right]^2.$$

Here the symbol ‹⊗› denotes the operation of convolution of each of the $n = 1 \ldots N$ theoretical ACFs of the signal $r^{t}(k)$ with the ordinates of the filter ACF $r^{f}(k)$. This operation is intended to take into account the influence of the receiver filter, as the narrowest-band element of the receiving path, on the measured characteristics of the useful IS signal. It simulates the distortion of the theoretical ACF of the ionized medium as a result of the passage of the conditional signal through the receiver filter, similar to what actually occurs with the experimental ACF of the IS signal.

As a result, the identifying neural network (Fig. 7) should have several layers of neurons in the proposed radial-basis structure. Since the IS radar of Institute of Ionosphere of NTU "KhPI" currently uses a library of N = several thousand theoretical ACFs, and in accordance with this experiment, ACFs with K=13 ordinates each will be obtained, then in such a structure, the neurons of the input layer $\mathbf{A^1}$ should be organized in the form of a continuous delay line with a length of 13N cells. These cells are intended for filling them with the values of the ordinates of the theoretical ACFs, i.e., at the beginning of processing, the network perceives service information about r^{t}, after which it is ready for work.

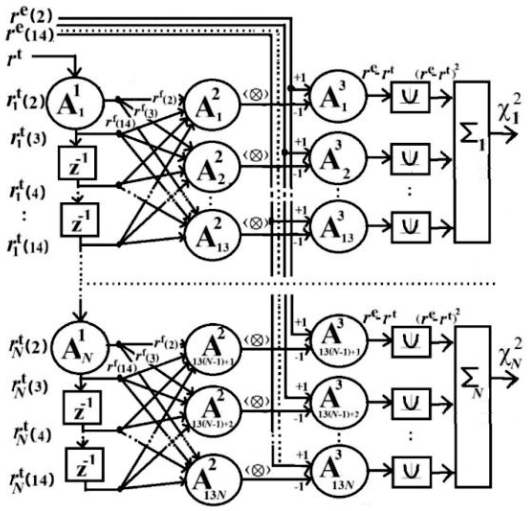


Fig. 7. Identifying neural network for ACF recognition of the HP signal.

Subsequently, the information is transmitted through the neurons of the interface layer that represents by elements of the synaptic connections $r^{\mathrm{f}}(k)$ and enables the implementation of the process of convolution of two functions.

Layer $\mathbf{A^2}$, in which the results of the convolution are accumulated, passes them to layer $\mathbf{A^3}$, which has the form of difference neurons with connections at the input of the form "±1" to determine the difference between the values of the ordinates being compared. Quadratic activation functions are at the outputs of this layer. The summation of the squared results of calculating the differences occurs in the last interface layer ∑, i.e., at its outputs, we have all n = 1, 2, …, N elements of the χ^2 residual function.

Thus, the neural network relates the input value $r^{\mathrm{e}}(k; \boldsymbol{\theta})$ simultaneously to the entire set of N updated (taking into account the filter) theoretical analogues. The value $\chi^2_{\min}$, which the computer finds among the list of obtained values of the residual function, reflects the degree of best proximity and corresponds to the sought ACF $r^{\mathrm{t}}(k; \boldsymbol{\theta})$ and the vector θ of ionospheric parameters.

Such a neural network makes it possible implementing simultaneous identification of the obtained ACFs r^{e}, using all the involved theoretical analogs r^{t}. The network is trained using the procedure of reconfiguring the weights $r^{\mathrm{f}}(k)$ to implement new current experimental conditions using test reference (noise-free) values of r^{e}. That is, in the case of using another type of filter in the receiving system of the IS radar, the neural network learns by adjusting to the new weights $r^{\mathrm{f}}(k)$ until zero results are obtained at the corresponding outputs χ^2. In addition, it is possible to use special methods for signal processing that increase the efficiency of processing by reducing computational costs and increasing the speed of algorithms, for example, [22].

When performing this work, the task was set to programmatically model a network with a simplified

architecture (only for several theoretical functions r^t). As a result, this modeling confirmed the efficiency of recognizing various forms of the scattering signal ACF, i.e. the operability of the proposed identifying neural network. In the future, the hardware implementation of this neural network is considered, taking into account specialized neuroboards. Similar products based on the use of the mechanism of action of neural networks have begun to appear on the market. For example, the large-scale integrated circuit developed by Adaptive Solutions (USA) contains 64 neurons, has 262144 synaptic connections and allows a speed of up to 1.2 billion connections/sec [23]. These elements will certainly become the basis for specialized multiprocessor devices, including those for the IS method.

V. Conclusions

The procedure for choice of complex sounding signals for studying the ionosphere using the incoherent scatter method is considered in order to improve the sounding process and subsequent radar processing of the IS signal to determine the state and dynamics of the ionosphere in an extended altitude range with sufficient resolution in range (altitude when vertical sounding) and time.

It is proposed to modernize the process of coding of these signals. When synthesizing individual elements of the sounding signal, it is proposed to use both the change in the direction of the circular polarization of the radio wave and the manipulation of the carrier frequency. The results of searching for this type of multi-element coded signals are presented.

To identify the correspondence of the obtained experimental ACFs of the scattered signal to their theoretical analogs, it is proposed to use an identifying neural network that allows for operational real-time assessment of the ionospheric plasma parameters. This network is the basis for the future creation of flexible and high-speed ionospheric information processing systems.

This work performed contributes to the development of means for monitoring the near-Earth environment, which has a significant impact on many terrestrial processes and on the state of which the quality of functioning of modern communication, radar and radio positioning systems depends.

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Improved Thermal Quadrupoles Model for Multilayer Dielectric with Defect

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Abstract— **The objective function for filament defect detection utilizes thermal quadrupoles in a cascading configuration to model a multi-layer structure with a defect. The model combines matrix products representing each layer. Two approaches for thermal quadrupole models are considered. The first approach simplifies the defect model to a thermal resistance layer, providing analytical expressions for defect thickness and depth, which offers some advantages but may lead to complexities due to the non-symmetric defect matrix. The second approach, focusing on a symmetric defect layer matrix, is more mathematically sound and has greater informational value, making it more suitable for this study. Initially, hyperbolic functions were used for modeling heat transfer, but due to their complexity, trigonometric functions were adopted for practical implementation. The objective function, which aims to identify defect parameters, incorporates these models and could eventually be integrated into a larger defect detection algorithm. This would automate the defect-checking process, enhancing practical applications. Simulation results incorporate sensor imperfections using random error (1%) in the equations, adding realism to the model without introducing significant distortions. These errors are minor and unlikely to affect practical applications given the expected thickness of the filament.**

Keywords—thermal quadrupoles, model, defect, multilayer, dielectric, objective function

I. INTRODUCTION

The thermal quadrupole method is a mathematical approach based on electrothermal analogies and the fundamental heat conduction equation. Initially introduced by H.S. Carslaw, and later developed further with J.C. Jaeger, the method uses Laplace transforms to solve complex thermal problems, particularly in systems with periodic or transient heat flow [1,2]. Its modular nature makes it well-suited for analyzing multilayered materials and structures, allowing each layer to be modeled as an individual quadrupole and combined in a cascade matrix configuration [14].

For precise thermal modeling, accurate thermal property measurements are essential. Reference [3] presents a comprehensive guide to selecting and applying suitable thermal measurement techniques, offering practical insights for both experts and non-specialists. This foundation is especially relevant in cases where thermal models are used for material characterization and defect detection.

In the context of 3D printing, defectoscopy focuses on identifying imperfections in printed filament structures. Since many filament parameters (e.g., thermal conductivity, density) are known a priori, thermal models can be adapted efficiently. Research in this area includes the use of terahertz time-domain spectroscopy (THz-TDS) to determine refractive indices while correcting for internal reflections [12]. The thermal quadrupole method has also been adapted to improve modeling accuracy for multilayer filament structures and to detect defects located between layers [14,15].

Several studies [4,5] have demonstrated the effectiveness of pulse thermography (PT) and transient thermal imaging in identifying delamination in multi-layered composite materials. These methods are particularly efficient when combined with analytical tools such as locally normalized thermal contrast and Laplace-transformed data, enabling the reconstruction of contact resistance and defect depth profiles. Results have shown strong correlation with traditional ultrasonic methods, indicating their reliability and speed for non-destructive testing (NDT) [4].

For an NDT technique to be effective, the simulated defects must realistically represent actual flaws in shape and material properties. One such study examined debonding at the edge of a composite patch using polyurethane, providing insights into accurate experimental setup and improving understanding of heat transfer behavior at critical bonding interfaces [6,7].

Thermal behavior in laminated structures can be significantly influenced by mechanical loading and defect geometry. Numerical models that simulate heat transfer under these conditions help predict temperature contours and identify delamination effects, considering parameters like shallow depth and thermal resistance [8].

Thermal quadrupole modeling has proven effective in designing and optimizing thermographic inspections. Parametric studies using two- and four-layer models demonstrate the method's adaptability to various composite systems, contributing to better planning of inspection protocols [9].

Passive thermography offers real-time, in-situ monitoring capabilities, especially useful during cyclic mechanical loading of composite materials. Modeling of subsurface heat generation enables accurate prediction of surface temperature changes, aiding in early defect detection [10].

An important development is the creation of analytical models that relate defect characteristics (e.g., opening, depth, temperature, conductivity) to the resulting thermal response under controlled heating. Such models enhance the interpretability of thermographic data and support the use of active thermal control strategies [13].

Across all studies, thermographic NDE methods – including both active and passive approaches – are

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emphasized as critical tools for high-reliability sectors like aerospace, where early detection of internal defects can prevent structural failure [11].

The reviewed works collectively highlight the effectiveness and adaptability of thermal quadrupole modeling and thermographic methods in the non-destructive evaluation of composite and 3D-printed materials. From fundamental heat conduction theory to advanced layered simulations, these approaches support accurate, fast, and reliable detection of internal defects such as delamination, debonding, and voids. By integrating analytical modeling, experimental thermography, and real-time monitoring techniques, these methods provide a comprehensive toolkit for ensuring structural integrity in modern engineering materials.

The purpose of the work is to improve the model in the form of thermal quadrupoles for multilayer dielectrics, namely for non-destructive defect detection of filament for 3D printing, which has internal defects. The improvement is carried out by finding additional connections between opening, depth, temperature, and thermal conductivity.

II. Multilayer Model with Inner Defects

A. Thermal Conduction Equation

As a thermophysical model of the sample, we will consider a multilayer unlimited plate of thickness l, one of the layers of which, differing in thermophysical characteristics, is a defect(Fig.1). Fig. 1 shows a multilayer sample, where λ, λ_d are thermal conductivity of the material of the sample and the defect, W/K m^2, a, a_d are thermal diffusivity the material of the sample and the defect, m^2/s, h is the depth of the defect, mm, δ is opening of the defect, mm, lower index d is defect.

The direct problem involves determining the heat flux at the output and the temperature if the input values for temperature and heat flux are known. The inverse problem entails identifying the defect parameters by analyzing the temperature and heat flux at both the input and output of the sample. A significant challenge in this research is the difficulty of measuring temperature and heat flux between layers without causing damage, which complicates non-destructive testing. The main problem of the study is the inaccessibility of the places between the layers for measuring the temperature and heat flux, as it should be during non-destructive testing.

The mathematical model includes the equation of non-stationary heat conductivity of the form

$$a\frac{\partial^2 T(x,\tau)}{\partial x^2}=\frac{\partial T(x,\tau)}{\partial \tau}, \tag{1}$$

where T is temperature, x is coordinate (fig.1), τ is time.

The generally accepted boundary conditions are

$$T\big|_{\tau=0}=0;\ T_{x=0}=T_1;\ -\lambda_1\frac{\partial T}{\partial x}\bigg|_{x=0}=Q_1(\tau); \tag{2}$$

$$-\lambda_n\frac{\partial T}{\partial x}\bigg|_{x=l}=Q_n(\tau);\ \ Q_1=q_1-\alpha_1 T_1;\ Q_n=-q_n+\alpha_n T_n. \tag{3}$$

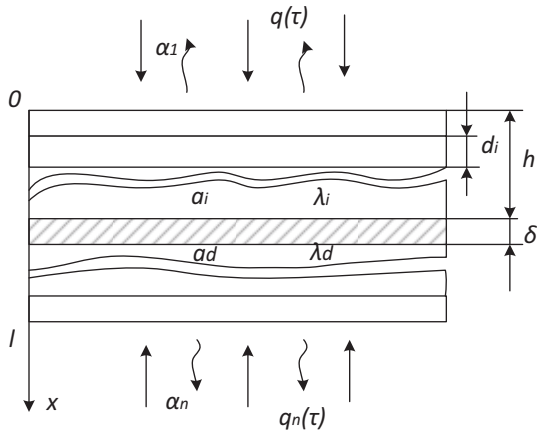


Fig.1 Multi-layer sample

To determine temperature as a solution to the heat conduction equation with specified boundary conditions, the result is expressed as a sum of trigonometric functions within a rectangular coordinate system. The heat flow is directly related to temperature through its derivative, which translates into heat flux with an associated coefficient. When obtaining this derivative, the trigonometric functions are differentiated according to standard rules, where sine and cosine functions are interchanged with corresponding sign.

B. Thermal Quadrupoles Model as Development of Thermo Conductivity Equation

The method of thermal quadrupoles is most descriptive when dealing with Laplace transforms of surface (F-front and R-rear) temperatures and heat fluxes:ΘF, ΘR, ΦF, ΦR [4].

The problem is solved in the matrix form

$$\begin{bmatrix}\theta^F\\ W\end{bmatrix}=M_1M_dM_3\begin{bmatrix}\theta^R\\ 0\end{bmatrix}, \tag{1}$$

where $\theta^F(z,p)$, $\theta^R(z,p)$ are the Laplace transforms on the front and rear surfaces, M1,M2 are the matrices characterizing plate layers (except the defect)

$$M=\begin{vmatrix}A & B\\ E & D\end{vmatrix}, \tag{2}$$

where

$$A=D=\cos h(kl);\ E=\lambda k\sinh(kl);$$
$$B=\sinh(kl)l(\lambda k);\ k=\sqrt{p/a}. \tag{3}$$

Respectively, for the matrix Md describing the defect layer

$$A_d=D_d=1;B_d=R_d;\ E_d=0. \tag{4}$$

In the Laplace domain, the front and rear-surface solutions are algebraic, thus allowing the calculation of defect depth l1 and thermal resistance Rd = d/λd.

III. Objective Function for Filament Defect Checking

Thermal quadrupoles can be cascaded to model the multi-layer structure of a sample with a defect, with each layer represented by the product of its corresponding matrix. Two approaches to thermal quadrupole models are explored in the literature ([2-4] and [13]).

The first approach, as seen in [2-4], simplifies the defect model to a thermal resistance layer, making the expressions for defect thickness and depth analytical, which is advantageous. However, this simplification can introduce complexities, particularly due to the non-symmetric nature of the defect matrix.

In contrast, the second approach, outlined in [13], uses a trigonometric identity ($\cos^2\theta + \sin^2\theta = 1$) to express the defect layer thickness. While this approach is mathematically robust, it lacks the depth of informational value found in the first model. However, the second model offers the benefit of a symmetric thermal quadrupole matrix, which is preferable to the non-symmetric model from [2-4].

The symmetric defect layer matrix approach is more appropriate for our research, and it was the model chosen for implementation. Initially, hyperbolic functions were considered, as they are theoretically better suited for modeling heat transfer in dielectric materials with defects. However, their complexity made them less practical for real-world applications. Trigonometric functions, by comparison, are simpler and easier to implement. Therefore, for the time being, the model will utilize trigonometric functions, with the potential to transition to hyperbolic functions in the future, pending successful model validation and error analysis.

For a multilayer filament (see Fig. 1) objective function is

$$\begin{aligned} F_1 &= (t_1 - \cos(kd + kh + k_d\delta))^2 + \\ &+ \left(t_2 + \sin(kd + kh + k_d\delta)\right)^2 + \\ &+ (t_3 - \cos(kd + kh + k_d\delta)^2 + \\ &+ \left(t_4 - \sin\left(kd + kh + k_d\delta\right)\right)^2, \end{aligned} \tag{5}$$

$$\begin{aligned} &\frac{\partial F_1}{\partial\left(k_d\delta\right)} = 2t_2\cos(kd + kh + k_d\delta) - \\ &-2t_4\cos\left(kd + kh + k_d\delta\right) + \\ &+2t_1\sin(kd + kh + k_d\delta) + \\ &+2t_3\sin\left(kd + kh + k_d\delta\right) = 0, \end{aligned} \tag{6}$$

where t_i defined experimentally, it is constant for each measurement case. From (6) sues

$$tg\left(kd + kh + k_d\delta\right) = \frac{t_2 - t_4}{t_1 + t_3}. \tag{7}$$

$$kd + kh + k_d\delta = arctg\left(\frac{t_2 - t_4}{t_1 + t_3}\right). \tag{8}$$

The thickness of a multilayer sample without defects is obtained through measurement. The value of k is calculated for known materials, such as a specific filament, based on reference data.

$$kl = k(kd + kh + k\delta), \tag{9}$$

If expression (9) result is not equal $arctg\left(\frac{t_2 - t_4}{t_1 + t_3}\right)$ This means there is a defect in the sample.

The objective function (which represents the goal of identifying defect parameters) could eventually be woven into a larger algorithm. This will allow for the automation of defect detection, making it easier to apply the model in practical scenarios.

While the model's theoretical framework has room for refinement, the current approach has potential, especially as we move toward integrating more advanced functions and improving error handling in future iterations

IV. Simulation and Uncertainty Definition

For the modeling process, we will incorporate the imperfections of the sensors by adding the rnd function to the equations that describe the layers. Errors have been introduced into the model using the MathCad rnd function, with a 1% random error (i.e., rnd(0.01)). This ensures that the model accounts for real-world imperfections in measurements, though these errors are kept within a reasonable range to avoid significant distortions in the results.

The relative error is calculated using the following formula

$$\Delta = \frac{\frac{t2(kd, kh, k_d\delta) - t4(kd, kh, k_d\delta)}{t1(kd, kh, k_d\delta) - t3(kd, kh, k_d\delta)} - \left(kd + kh + k_d\delta\right)}{kd + kh + k_d\delta}, \tag{10}$$

The dependence of total sample thicknes from first layer thickness is shown on the fig.2.

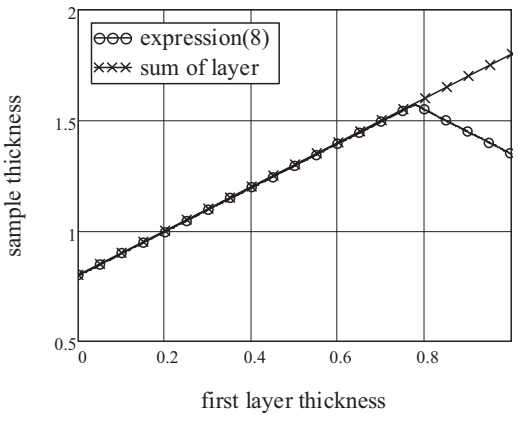


Fig.2 Dependence total sample thicknes from first layer thickness

After reaching 0.6 of the normalized layer thickness, the expression starts to exhibit an error. However, up until this point, it behaves correctly and aligns with the physical expectations of the problem. Since the filament is relatively thin, it is unlikely that we will encounter this part of the graph in practice.

Fig. 3 illustrate the behavior of the relative error fromexpression (10).

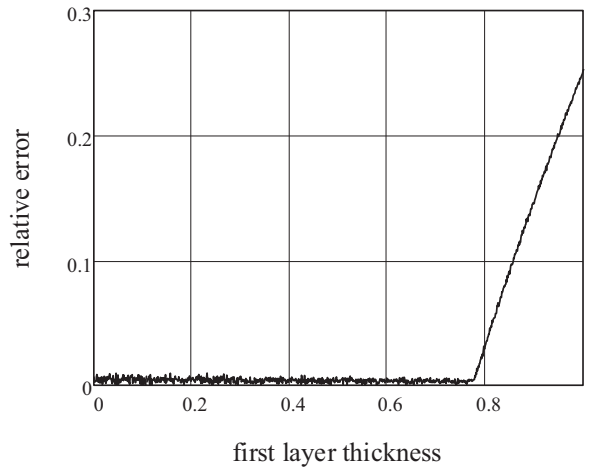


Fig.3 Dependence of relative errore from defect location depth

V. CONCLUSION

This study presents a thermal quadrupole model for detecting defects in multi-layer filament structures, highlighting the advantages and trade-offs of two approaches for modeling the defect layer. The first approach simplifies the defect to a thermal resistance layer, offering analytical expressions for defect thickness and depth, but introduces complexities due to non-symmetric matrices. The second, more suitable approach utilizes a symmetric defect layer matrix, providing greater mathematical clarity and informational value.

While the model initially used hyperbolic functions for heat transfer, it was found that trigonometric functions offer a simpler and more practical implementation for defect detection in real-world applications. The objective function developed in this study serves as the foundation for identifying defect parameters and can be integrated into a broader defect detection algorithm, which would enable automated analysis and improve practical usability.

Simulations incorporating sensor imperfections demonstrate that the model accurately reflects the behavior of the system under realistic conditions, with minimal error introduced by random measurement uncertainties.

Future work may involve refining the model further, potentially transitioning to hyperbolic functions for greater precision once the feasibility and error margins are thoroughly evaluated. Overall, the current approach holds promising potential for effective and efficient defect detection in filament materials

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Modelling of geometric irregularity of railway track using deep learning neural network

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***Abstract* − Ensuring the comfort and safety of passengers is one of the most important tasks in the design and maintenance of passenger rail transport. One of the threats to passenger comfort and safety arising from train movement is its oscillations, which are mainly caused by irregularities of the railway track. The paper proposes the use of Statistics-Informed Neural Network (SINN) for track irregularity modelling due to its ability to reproduce stochastic processes according to their statistical characteristics. In this work, a computer program was developed to generate short-time track profile trajectories based on track irregularity power spectrum. The obtained short-time trajectories were used as a target sample for training a neural network. The trained neural network generates samples of railway track irregularities that are close to real ones and have the same statistical characteristics. The obtained neural network is able to generate trajectories of railway track irregularities that are longer than the training samples, while maintaining the features of the desired stochastic process. It can be used to model irregularities on sections of track of various length and further used in a complex model of vertical oscillations of rolling stock for the on-board control systems of Ukrainian trains.**

Keywords: rolling stock, oscillatory process, track irregularities, neural network, SINN, decision support system, on-board control systems.

I. Problem Statement

Ensuring the comfort and safety of passengers is an important task that scientists are working on both at the design stage of rolling stock and during the repair and maintenance of existing trains. One of the threats that arise when a train is moving along a railway track is the forced oscillations of the carriage and its components. They lead to accelerated wear of the suspension system, contact surfaces of wheels and a track, cause loss of wheel grip and, ultimately, increase energy consumption. Critical values of accelerations in the vertical and transverse planes can also lead to passenger discomfort and even train derailments [1−3]. The negative effects of car oscillations can be exacerbated on the Ukrainian railways by the use of outdated rolling stock and rail irregularities caused by their wear and tear. All of this significantly limits the maximum permissible speeds on the tracks. To solve these problems, it is necessary to conduct research on rolling stock oscillations and their dependence on track irregularities. To this end, experts are constantly developing new mathematical and simulation models of rolling stock and upgrading existing ones, which saves time and resources on complex and sometimes dangerous real-world experiments.

II. Analysis of Recent Research and Publications

Modelling of oscillations in the vertical plane (bouncing and galloping) is usually performed by using a system of differential equations that describes these dynamic processes. The first main area of research on vertical oscillations is the study of their dependence on the stiffness and damping coefficients of the dampers of the first and second levels of car suspension or on the taper of the wheels [4−6]. The second direction is to study the dependence of oscillations on the condition of the railway track: ballast layer and rails [7−10]. Regardless of the ultimate goal of the research, in order to reproduce forced vertical oscillations, it is necessary to set a disturbance input in the form of a random irregularity of the rails [11].

Irregularities can be represented in several ways. Firstly, real data of irregularities existing on a particular section of track can be taken for modelling using track measuring wagons [11, 12]. This method most accurately reproduces the actual condition of the rails, but requires real-world experiments using specialised equipment and blocking the selected railway junctions, hence in many cases is not cost-effective. Given the lack of experimental data, scientists widely use an indirect method, which consists of simulating a random process that describes track irregularities by their power spectral density (PSD) [11, 13−16]. This method covers the entire frequency spectrum of the irregularities that are being analysed. An alternative option for irregularity simulation is to represent it as a signal consisting of the sum of only a few selected frequencies that are specific to a particular section of the track or that describe specific periodic types of irregularities, such as a joint irregularity [17].

In recent years, the use of neural networks in railway-related research has become increasingly popular. In addition to their classical use for storing information and problem-solving in the field of logistics and transportation management, they have also found their application in the analysis of data related to the calculation of passenger comfort [18], rolling stock oscillations and calculation of the forces of its interaction with the railway track [9, 10, 18].

The range of possibilities for using artificial neural networks has expanded significantly with the advent of so-called Deep Learning Neural Networks, which are

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capable of modelling high-level abstractions using numerous nonlinear transformations. In recent years, recurrent neural networks have been actively used in the study of oscillations and forces arising during the train movement. In [10, 19], they are used to predict rolling stock oscillations. Convolutional neural networks are also used in [10, 20–22]. Their main application is to predict oscillations and forces in the wheel-rail contact.

Deep learning neural networks can also be used to generate stochastic processes by finding dependencies in the distribution of values that are invisible to the human eye [22, 23]. In [23], the authors developed a so-called Statistics-Informed Neural Network (SINN) framework based on a recurrent neural network, which is capable of learning the stochastic dynamics of a given signal and reproducing it by generating a free-length signal. Since railway track irregularities are stochastic in nature and can be reproduced based on their power spectral density and other statistical characteristics, SINN can be used to generate railway track irregularities caused by rail wear.

The goal of the paper is to perform modelling of random irregularities of a railway track section based on their power spectral density using a Statistics-Informed Neural Network. The generated irregularities will be further used to model the vertical oscillations of rolling stock, which are used to calculate the control laws in on-board decision support systems for drivers on Ukrainian railways.

III. Random Irregularities of the Railway Track

The values of railway track irregularities obtained using the methods of the theory of random functions are widely used as external disturbances for modelling the forced oscillations of rolling stock [11, 13, 14]. One of the most used approaches uses an equivalent geometric irregularity η with specified spectral characteristics [13 – 15]. The data collected by researchers from different countries demonstrate that track irregularities can be represented as a stationary ergodic random process. The main measured statistical characteristics of track irregularities are power spectral density and autocorrelation function (ACF).

The PSD is calculated as a function of frequencies $\omega = 2\pi/L_d$, where L_d is the wavelength of the irregularity [14]. For our model, we select the formula for calculating the PSD for German high-speed railways [14].

$$S_v(\omega) = \frac{A_v \omega_c^2}{(\omega^2 + \omega_r^2)(\omega^2 + \omega_s^2)}$$

where $S_v(\omega)$ is the PSD, m^2/rad/m; A_v is the vertical irregularity parameter, m rad; ω_c, ω_r, ω_s are frequency parameters, rad/m.

The numerical modelling of a random process of track irregularities with a given power spectral density using the trigonometric series method is conducted. Let the sample of irregularities be $\eta = \eta(x)$, where $\eta(x)$ is the track irregularity of coordinate x and is represented by a random function, the mean of $\eta(x)$ equals zero. The discrete function $\eta(x)$ can be approximated by the following trigonometric series:

$$\eta(x) = \sum_{k=1}^{N}(a_k \cos \omega_k x + b_k \sin \omega_k x) \quad (2)$$

where a_k, b_k are independent Gaussian random variables with mean 0 and standard deviation σ_k, and their values are independent of each other for $k = 1,2,...,N$.

$$\sigma_k^2 = 2S_v(\omega_k)\Delta\omega \quad (3)$$

$$\omega_k = \omega_{\min} + (k - \frac{1}{2})\Delta\omega \quad (4)$$

$$\Delta\omega = (\omega_{\max} - \omega_{\min})/N \quad (5)$$

The frequencies from the frequency spectrum ω_k are calculated using formula (4) for $k = 1, 2,...N$, where N is the total number of frequency increments in the $[\omega_{\min}, \omega_{\max}]$. The frequencies $\omega_{\min} = 2\pi/L_{d\,\max}$ and $\omega_{\max} = 2\pi/L_{d\,\min}$ are the minimum and maximum frequencies of the given frequency spectrum, respectively. $L_{d\,\min}$ and $L_{d\,\max}$ are the minimum and maximum wavelengths. $\Delta\omega$, used in equation (5), is the frequency increment over the interval $[\omega_{\min}, \omega_{\max}]$.

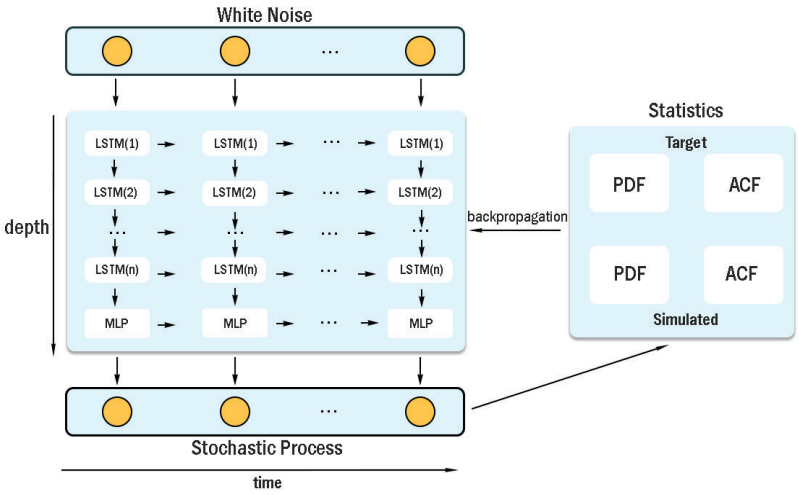


Fig. 1 SINN architecture

IV. Statistics-informed neural network

A. Neural network architecture

The structure of a statistics-informed neural network is shown in Fig. 1. It is based on several layers of a Long Short-Term Memory (LSTM) neural network. The output gate of the last LSTM layer is connected to a fully connected linear layer for data conversion.

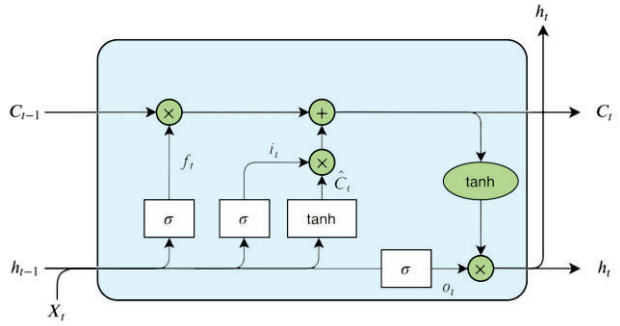


Fig 2. LSTM-cell structure

The input of the model is data with high entropy, i.e. white noise. It enters the first of a series of LSTM cells, each of which has the structure shown in Fig. 2. An LSTM cell has an input vector x_t; a hidden state vector h_t, which represents the network's "memory" and is initialised with zeros; a cell state c_t, which is responsible for storing long-term information; and three types of gates to control the flow of information in the network.

The first gate is the Forget gate. It accepts the past hidden state h_{t-1} and the current input x_t and outputs a vector of values scaled from 0 to 1, which indicate how much of the cell's previous state should be "forgotten". This gate allows the network to selectively delete and remember information from the previous step. The second gate is the Input gate, which takes the past hidden state h_{t-1} and the current input x_t and outputs a vector of values scaled from 0 to 1, which indicate which part of the input vector should be added to the cell state. This allows to choose whether to add or discard new information. The third gate is the Output gate. It takes the previous hidden state h_{t-1}, the input vector x_t, the current state of the cell c_t, as it's input and outputs a vector of values, and shows which part of the current state of the cell should be displayed as the current hidden state h_t. This allows to ignore parts of the cell state when calculating the output. At the output of each cell, an output vector y_t is generated, which in turn is fed to the input of the next LSTM cell.

The equation for the first LSTM cell in the SINN:

$$f_t^{(1)} = \sigma_g\left(W_f\xi_t + U_f h_{t-1}^{(1)} + b_f\right)$$

$$i_t^{(1)} = \sigma_g\left(W_i\xi_t + U_i h_{t-1}^{(1)} + b_i\right)$$

$$o_t^{(1)} = \sigma_g\left(W_o\xi_t + U_o h_{t-1}^{(1)} + b_o\right)$$

$$\hat{c}_t^{(1)} = \sigma_c\left(W_c\xi_t + U_c h_{t-1}^{(1)} + b_c\right)$$

$$c_t^{(1)} = f_t^{(1)} \circ c_{t-1}^{(1)} + i_t^{(1)} \circ \hat{c}_t^{(1)}$$

$$h_t^{(1)} = o_t^{(1)} \circ \sigma_h\left(c_t^{(1)}\right)$$

where ξ_t is the cell input vector represented as white noise; $c_t^{(1)}$ is the state of the first LSTM cell; $\hat{c}_t^{(1)}$ is the activation of the cell input; $f_t^{(1)}$, $i_t^{(1)}$, $o_t^{(1)}$ are the forget, input and output gates, respectively. W_f, U_f, W_i, U_i, W_o, U_o, W_c, U_c are the weight matrices of the forget, input, output gates and activation of the cell, b_f, b_i, b_o, b_c are the biases for the gates and activation; σ_g is a sigmoidal activation function; σ_c is a hyperbolic tangent.

The input of the next layer, instead of ξ_t, is the output of the previous layer $h_t^{(i-1)}$. The final output of the SINN is calculated as $X_t = W_m h_t^{(n)}$, where $h_t^{(n)}$ is the output of the n-th LSTM cell, W_m is the weighting matrix.

B. Loss Functions

To obtain the desired result in SINN, the statistics of the output signal must match the statistics of the target sample. These statistical characteristics include the autocorrelation function and the probability density function (PDF).

The autocorrelation function for a sequence X is a deterministic function of the delay τ, and is defined as:

$$ACF_X(\tau) = \frac{E[X_t X_{t+\tau}]}{E[X_t^2]}, \quad (6)$$

where the process X has zero mean.

There are two main methods for calculating the ACF. According to the first, the ACF for each τ is calculated using the formula (6). The second method is to use the Fast Fourier Transform (FFT) to calculate the ACF.

$$ACF_X = FFT^{-1}(FFT(X)FFT(X)^*),$$

where * is a complex conjugate.

The second approach, although faster, can be significantly inferior to the first in accuracy, especially for large values of τ. Both methods allow to propagate the ACF-based losses to the neural network model. The loss function for the autocorrelation function is calculated using formula (7). Here, $ACF_O(\tau)$ is the ACF of the original SINN signal at time τ, $ACF_T(\tau)$ is the ACF of the target stochastic process.

We calculate the probability density function for the original and desired signals using the kernel density estimation:

$$\hat{f}_h(x) = \frac{1}{|X|}\sum_i^{|X|} K_h(x - X_i)$$

where K is a non-negative kernel, h is a smoothing parameter. $K_h(d) \doteq \frac{1}{h}K\left(\frac{d}{h}\right)$ is the scaled kernel.

The loss function for the PDF is calculated by formula (8), where $\hat{f}_h^O(x)$ and $\hat{f}_h^T(x)$ are the PDF of the source sequence and the target stochastic process.

$$Loss_{acf} = \frac{1}{n}\sum_{\tau\in T}|ACF_O(\tau) - ACF_T(\tau)| + \frac{1}{n}\sum_{\tau\in T}[ACF_O(\tau) - ACF_T(\tau)]^2 \quad (7)$$

$$Loss_{pdf} = \frac{1}{n}\sum_{i=1}^{n}\left|\hat{f}_h^O(x_i) - \hat{f}_h^T(x_i)\right| + \frac{1}{n}\sum_{i=1}^{n}\left[\hat{f}_h^O(x_i) - \hat{f}_h^T(x_i)\right]^2 \quad (8)$$

V. Preparation for the Simulation

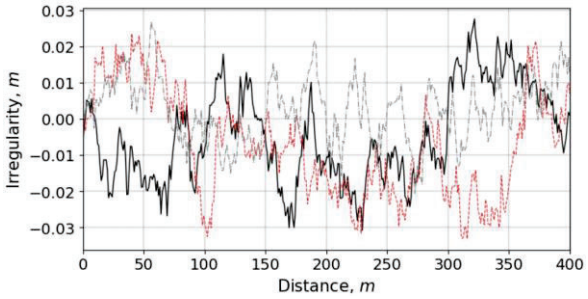


Fig. 3. Simulated track profile samples

To collect the data for the training, a MATLAB program was created that calculates the trajectories of railway track irregularities using the trigonometric series method according to power spectral density calculated using formulas (1) and (2). The minimum and maximum wavelengths of the irregularities used in the calculations are $L_{d\,\min}$ = 0.5 m and $L_{d\,\max}$ = 50 m. The equation parameters were selected according to the recommendations for calculating irregularities of high interference level for high-speed railway transport of Germany: A_v = 10.80 10^{-7} m rad, ω_c = 0.8246 rad/m, ω_r = 0.0206 rad/m, ω_s = 0.4380 rad/m [14]. The number of frequency increments is N = 2000. The program generates trajectories for the target and validation samples. Three different realisations of the railway track trajectories with irregularities obtained using the developed programme are shown in Fig. 3.

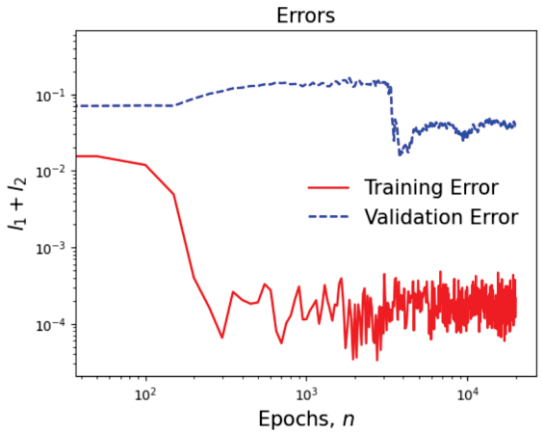
The SINN neural network was implemented in the Python programming language using the PyTorch machine-learning framework. The neural network is trained using stochastic gradient descent method with the Adam optimizer. The learning rate is set to 10^{-3}, with parameters β_1 = 0.9 and β_2 = 0.999. The values of the training and validation loss functions are calculated every 50 iterations during the training process.

The input data are trajectories consisting of Gaussian white noise and generated randomly for each new training of the network. The training, target, and validation samples each contain 400 separate process realisations, each consisting of 400 points. The validation sample was created to prevent overtraining of the neural network.

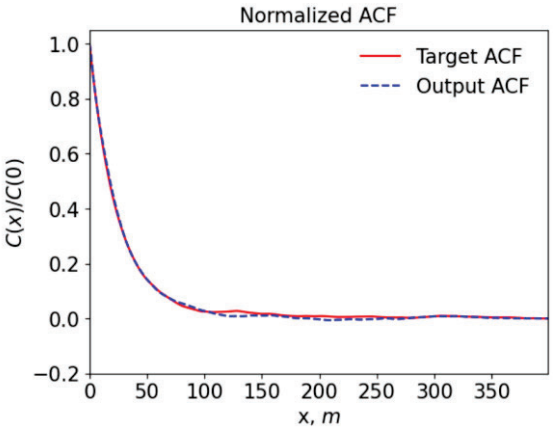
From the entire range of lags 1...t_{max}, a set T of lag values τ_1,..., τ_m is randomly selected at each step of the gradient descent, where $m << t_{max}$.

VI. Results of Neural Network Training

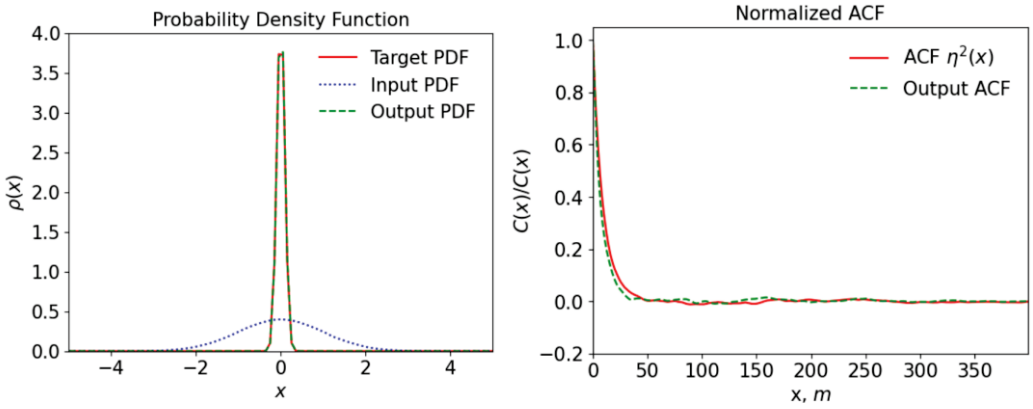
During the training, the network structure was chosen, namely four layers: an input layer, two hidden LSTM layers, and an output layer. The input and output layers have one neuron each, the first hidden layer contains 25 neurons, and the second hidden layer contains 5 neurons. Training was performed for 20 000 epochs. Fig. 4*a* shows the graphs of training and validation errors. Figs. 4*b* − 4*d* show the graphs of statistical characteristics of the target and the obtained processes. The graphs show that the autocorrelation functions of the target and the output samples have a high degree of convergence. The probability density function of the resulting sample fully reproduces the probability density function of the target process. Thus, the irregularity trajectories obtained by the SINN neural network have the desired statistical characteristics.



a)



b)



c) *d*)

Fig. 4. Results of neural network training. a) Training and validation errors; b) Normalized ACFs; c) Probability density function; d) Normalized ACF for η^2

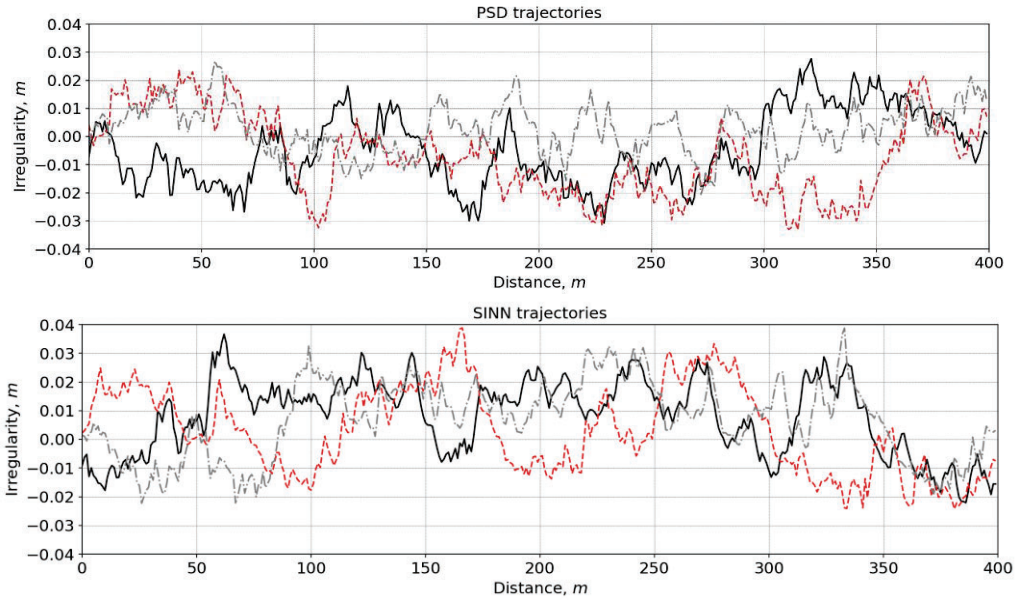


Fig. 5. Simulated irregularities of railway track

The samples obtained with PSD and those obtained with SINN are shown in Fig. 5.

The irregularity trajectories on a 400-meter long section obtained using the neural network have the same statistical characteristics as the target trajectories, the same amplitude ranges, and are generally similar to the implementations reported in the literature [13, 14].

The trained neural network is also capable of generating trajectories that are longer than those of the target sample. The implementation of such a long-term trajectory for a 1200-meter long path section is shown in Fig. 6. Such trajectories can be used as a disturbance input to calculate the vertical oscillations of the rolling stock on a track section of arbitrary length.

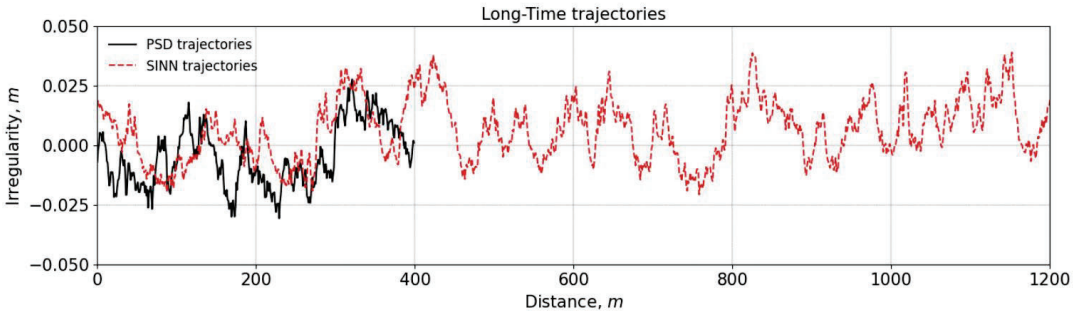


Fig. 6. Long-time track profile trajectories

VII. Conclusions

In this paper, analysis of the main causes of vertical oscillations of a train is carried out. It is proposed to analyse the dependence of oscillations on track irregularities. For this purpose, a review of methods for modelling the track profile and the use of neural networks for generating stochastic processes was performed. It was decided to use a statistics-informed neural network to generate irregularity trajectories.

To collect samples for training the neural network, a program was developed that calculates the power spectral density of the desired signal and generates short-term samples of path irregularities based on the trigonometric series method. In the course of the experiments, the structure and parameters of the statistics-informed neural network are chosen. The statistical characteristics of the desired stochastic process for training the neural network are obtained from previously generated training sample. The trained neural network generates samples of railway track irregularities that are close to real ones. The neural network is able to generate irregularity trajectories that are longer than the training samples, while maintaining the characteristics of the desired stochastic process, and therefore can be used to model irregularity on sections of arbitrary lengths.

The resulting neural network can be used to model irregularity on long sections of railway track and further applied in a complex model of vertical oscillations of rolling stock for on-board decision support systems that calculate optimal control actions of a train on a given section of track.

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Method for diagnosing power elements and electronic devices of an autonomous electric power system

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Abstract*—The article analyzes the causes of emergency conditions in shipboard alternating current (AC) electric power systems. It focuses on how external failures, such as phase disconnection or voltage drops, can lead to internal malfunctions and a cascading chain of damage. The author explains specific failure mechanisms. For example, exceeding current limits in rectifiers can cause internal short circuits due to thermal breakdown, while inverter backfeeding leads to increased power switch-off times. The article emphasizes that the thermal regime of components, which depends on geometric dimensions and current pulse frequency, is a critical factor. To study the dynamics of defect propagation, the article proposes developing a model that can evaluate the sequence and speed of damage spread. This model would use simplifying assumptions like network symmetry and linear resistance. The author also suggests using time-domain diagrams of voltages and currents to demonstrate different fault types. The text lists common failures, including short circuits of semiconductor modules, inverter and control module failures, and interphase short circuits. It concludes by discussing diagnostic methods, proposing that voltages and currents across power valves be used as primary diagnostic parameters because their variations indirectly reflect the thermal conditions of components. The loss of control in one valve is highlighted as a factor that increases the load on others, promoting overheat.

***Keywords*—power plant, automatic control, autonomous voltage inverter, electrical apparatus, electrical machines, mathematical modeling, electrical systems and networks.**

I. Introduction

Modern shipboard power systems are increasingly utilizing frequency converters (FCs) to enable efficient and flexible control of electric drives, automation systems, and other onboard equipment. The reliability and uninterrupted operation of FCs are critically important for the safety of the vessel and its ability to perform designated functions. Therefore, the issue of timely and accurate diagnostics of their condition has become particularly relevant. This literature review covers current trends and methods employed for diagnosing shipboard frequency converters.

One of the key trends in the diagnostics of frequency converters is the shift from traditional methods based on periodic inspections and measurements to more proactive and intelligent approaches. This includes the implementation of continuous monitoring systems that track key FC performance parameters in real time, such as current, voltage, temperature, frequency, and vibration.

The diagnostic methods currently under active research and application can be classified as follows:

Electrical signal analysis: This approach is among the most widely used and involves analyzing current and voltage at the input and output of the FC. Spectral analysis techniques, such as Fast Fourier Transform (FFT), allow for the detection of characteristic harmonics that indicate specific faults, such as short circuits, open circuits in power semiconductors, or failures of DC-link capacitors [1].

Thermal monitoring: An increase in temperature is one of the primary indicators of faults in FCs. The use of thermal imaging cameras and temperature sensors enables the detection of overheating in specific components, which may indicate performance degradation or impending failure [2].

Vibration analysis: Mechanical damage to cooling fan bearings or uneven loading of power modules can lead to elevated vibration levels. Vibration spectrum analysis enables early identification of such issues [3].

Model-based diagnostics: This approach involves developing mathematical models of the FC, which are used for comparison with actual measurements. Deviations between the model and real data can point to the presence of faults. The development of machine learning techniques has enabled the creation of more accurate and adaptive models for diagnostics [4].

Artificial intelligence (AI) and machine learning (ML)-based diagnostics: AI and ML algorithms, such as artificial neural networks, decision trees, and support vector machines, are increasingly being used to analyze large volumes of data obtained from monitoring systems. These methods make it possible to detect complex patterns and accurately predict future failures [5].

Modern shipboard power systems increasingly rely on frequency converters to provide efficient and flexible control of a wide range of onboard equipment. The safe operation of the vessel, the functionality of critical systems, and the overall operational efficiency are directly dependent on the reliable performance of these converters. Despite significant advancements in the development and implementation of diagnostic methods for FCs, there remains an urgent need for further research to address existing challenges and to expand the capabilities of monitoring and fault prediction systems.

Modern diagnostic methods include electrical signal analysis, thermal and vibration monitoring, model-based diagnostics, and approaches based on artificial intelligence

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and machine learning. However, the integration of these heterogeneous methods into a single, comprehensive diagnostic system remains a complex challenge [6]. Data obtained from various sensors and monitoring systems are often unstructured and require sophisticated processing and correlation algorithms to produce a coherent picture of the FC's condition.

Moreover, the specific operational conditions of shipboard FCs—such as exposure to aggressive marine environments (humidity, salt mist, vibration from main engines and wave impacts)—impose additional demands on the reliability and robustness of diagnostic systems [7]. Sensors and monitoring equipment must be protected against corrosion, vibration, and electromagnetic interference, which may limit the choice of available technologies and increase the cost of implementation.

Another critical aspect is the need to develop more accurate and predictive fault models. Most existing methods focus on detecting already existing defects, whereas the ability to predict future failures at an early stage would enable timely preventive maintenance, minimize vessel downtime, and prevent emergency situations [8, 9]. Achieving this goal requires deeper research into the mechanisms of fault occurrence and development in shipboard FCs, considering their specific operating conditions.

The advancement of AI and ML technologies opens new prospects for improving FC diagnostic systems. However, the effective application of these methods requires large volumes of high-quality data on FC operation under various modes and fault conditions. The collection and sharing of such data among different vessels and operators may be hindered by confidentiality concerns and the lack of standardized protocols [10, 11]. Therefore, further research should focus on developing efficient methods for collecting, processing, and exchanging diagnostic data, as well as on creating adaptive AI algorithms capable of learning from limited datasets while considering the specific characteristics of particular FC types and their operational environments.

Additionally, there is a need to design simpler and more cost-effective diagnostic systems that are accessible to a broader range of vessels, including smaller ships. Many of the current advanced methods are complex and expensive to implement, limiting their applicability on low-tonnage vessels.

II. Enhancing the Efficiency and Reliability of Operation of the Shipboard Unified Alternating Current Power System

The causes of external faults are located beyond the boundaries of the control system of the ship's power system and the power circuits of the valve complexes of this system. Given this, if external defects (emergency processes) are not immediately eliminated, this usually leads to the further development of internal critical conditions within the shipboard unified AC power system.

If the overload current values exceed the permissible operating currents of an individual ship rectifier, this creates conditions for internal short circuits. This becomes possible due to thermal breakdown caused by a significant temperature rise in the structure of the rectifier's valve. Such a situation can lead to a hard short circuit at the rectifier's output terminals.

The backfeeding of the ship inverter causes current oversaturation of the inverter, which leads to an increase in the switch-off time of the power keys beyond their nominal values. For such a failure to occur, one of the phases must be disconnected or the voltage in the ship's AC network must drop below the nominal level.

The load-carrying capacity of the valves is determined by their thermal operating mode, which depends on their geometric dimensions, duration of exposure, and the frequency range of anode current pulses. Analysis of modes caused by emergency situations in these conditions requires calculating the overload currents and understanding the potential thermal operating regimes of the power valves under these load conditions.

To clearly study the development of defects during the operation of the unified shipboard AC power system, it is necessary to develop a model of emergency mode propagation. This model will allow us to illustrate the dynamics of damage propagation and make it possible to evaluate the sequence and speed of the spread of emergency processes.

To identify specific types of failures that occur in the ship's AC power system, we will consider the following rectifier operating modes:

– short circuit of the semiconductor module;
– open circuit of the semiconductor module;
– inverter failure due to excessive current loads or malfunctions in the control system;
– interphase short circuit in the winding;
– malfunction or complete failure of the control module;
– break in the phase winding;
– failure of the power element of a switch.

To construct the model, we establish the following assumptions:

– the three-phase network is symmetrical;
– the active and inductive resistances in the circuits are linear;
– the electromotive force is sinusoidal and has a constant amplitude.

The models are represented as time diagrams of voltages and currents, which clearly illustrate the failure types listed above.

Let us consider, based on a standalone inverter, a model of a valve pair (Fig. 1) [12, 13].

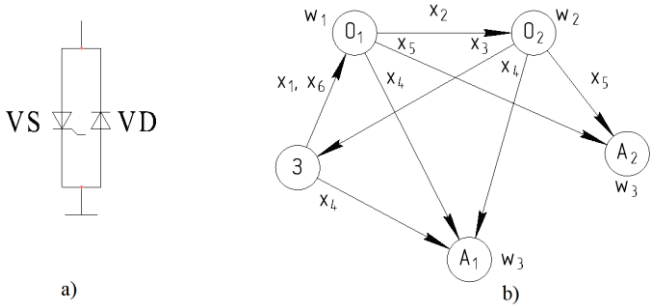


Fig. 1. The model of a a valve pair

a) valve pair;

b) types of transitions

The valve pair is considered as a Moore finite state machine.

The input and output signals are given in Tables 1 and 2.

TABLE I. INPUT SIGNALS

Code Value	*Description*
$x_1(i_y \neq 0)$	control signal present
$x_2(i_y \prec 0)$	thyristor off and diode on
$x_3(i_y \succ 0)$	full switch off
$x_4(i_y = 0)$	valve failure
$x_5(i_y \neq i_{кз})$	alve breakdown (short circuit)
$x_6(u_i \succ 0)$	positive voltage present at the valve

TABLE II. OUTPUT SIGNALS

Code Value	*Description*
w_1	$i_T(t);\ u_T(t)$
w_2	$i_D(t);\ u_D(t)$
w_3	alarm signal $i_{short\ circuit}$

Table 3 lists the possible states of the valve pair.

TABLE III. VALVE PAIR STATES

Encoded Value	*Description*
3	closed state
O1	thyristor conduction state
O2	diode conduction state
A1	fault (due to open circuit, one of the valves failed)
A2	fault (due to breakdown, one of the valves failed)

It is possible to divide the emergency transient process in the ship rectifier into intervals depending on the number of phases participating in current conduction. For any interval, the equivalent circuit changes at its boundaries but remains constant within the interval.

Initial conditions for the currents at each moment are set based on the final conditions from the previous interval. The "short circuit" mode arises in the ship's converter circuit during the breakdown of a power switch. Let us assume that Power Valve 1 has failed (shorted), and the emergency process begins only when Power Valve 3 opens—an internal short-circuited loop is formed in the circuit. The current, under the influence of line EMF e_{2ab}, flows through thyristors V1 and V3. A two-phase short circuit loop forms between phases e_{2a} and e_{2b} (Power Valves 1 and 3)—see Fig. 2a.

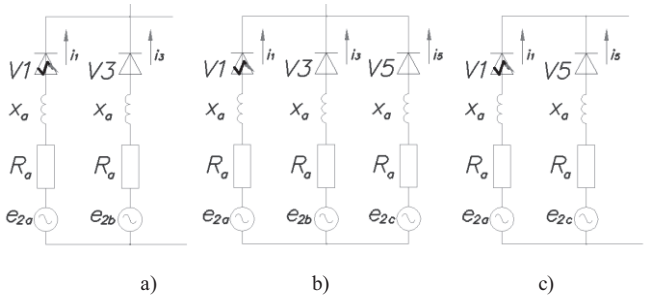


Fig. 2. Short circuit occurrence between phases

At the moment a valve from the group of damaged power switches opens (in this case, Power Valve 5), a three-phase short circuit occurs—not between two, but all three phases (e_{2a}, e_{2b}, and e_{2c}) – see Fig. 2b.

The three-phase short circuit mode continues until the current in Power Valve 3 drops to zero. Once the current in Valve 3 reaches zero, the three-phase short circuit transitions into a two-phase short circuit between phases e_{2a} and e_{2c} (Power Valves 1 and 5) – see Fig. 2c.

In the circuits where the short circuit occurred, the current values are limited only by the active resistance of the ship's power transmission line Ra. During the creation of an internal short circuit, overcurrents of emergency overload appear starting from the first period of the supply network voltage waveform.

In a bridge inverter circuit that uses counter-parallel diodes, similar processes occur when semiconductor modules are shorted. The equivalent circuit of the inverter in this mode is shown in Fig. 3, where Z_n – the load resistance, L_k and C_k – inductance and capacitance of the inverter oscillatory circuit and C_p – the capacitance of the capacitor connected in parallel to the transformer winding.

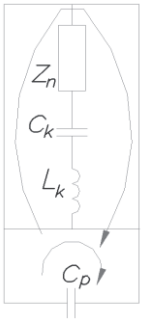


Fig. 3. Equivalent circuit of the shipboard converter during a short circuit of the semiconductor module

The formation of short-circuited loops causes a redistribution of energy reserves. Very high current values arise, resulting in significant overloads on reactive components.

These emergency conditions develop in an avalanche-like manner, with extremely high rates of current and voltage rise in the circuit elements. They are accompanied by significant overvoltages and excessively high currents in the inductive components. These currents pose a serious

threat to the valves and wires of the ship's electrical system, especially in powerful installations with thin wire terminations.

The duration of emergency processes typically lasts one control period. Protection from the development of these negative conditions is only possible through the use of modern protection systems. Such systems in real converters include protection channels. Protection devices that can disconnect circuits with fault currents help to shorten the duration and magnitude of overcurrents, thus reducing the severity of the emergency condition.

Table 4 presents the results of emergency mode analysis in the form of cause-and-effect relationships. It compares diagnostic signs with the causes that triggered their appearance.

Table 4 shows that it is impossible to link each specific type of failure to external defects in the power circuits. This is explained by the fact that electrical circuits are continuous networks, and any failure in one part of the circuit leads to a new redistribution of current values in all branches, affecting the operation of all components and worsening their performance.

Diagnosing shipboard power systems is most appropriate in the signal space. In this case, diagnostic indicators include currents, voltages, and their derivatives measured at several control points in the ship's electrical network. When diagnosing complex ship subsystems, each of them may use its own variant of indicator evaluation (e.g., full state information, evaluation of only input and output signals, partial evaluation, or evaluation based solely on output signals).

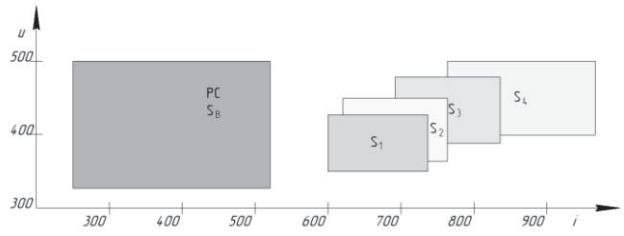


Fig. 4. Relationship of External Indicators and States for Power Inverter Valves

The consideration of interrelations between the states of the inverter power circuit (see Fig. 4) and external signs (such as voltage and current values on the power valves of the inverter) does not imply a one-to-one correspondence between them.

The subsystems of shipboard power systems consist of a large number of components. The condition of each component is determined by its own diagnostic parameters. Semiconductor elements are monitored through parameters such as the rate of current and voltage rise on the valves, recovery time of control properties, and temperature. Capacitors, inductors, and resistors can be monitored by tracking values such as capacitance loss, inductance loss, and resistance loss respectively. However, because the weakest elements in power circuits are the semiconductor valve elements, we will use the parameters listed in Table 5 as diagnostic parameters for assessing the health of these circuits.

TABLE IV. RELATIONSHIP BETWEEN CAUSE AND EXTERNAL INDICATORS FOR POWER INVERTER VALVES

Cause	External Indicator										
	No power	*Current decrease*	*No current*	*Voltage decrease*	*No voltage*	*Temporary loss of output voltage*	*Appearance of pulsations*	*Increase in the rate of rise of du/dt*	*Increase in the steepness of the rise of di/dt*	*Generation of short-circuit currents on the valves*	*Uneven current loading of the valves*
Short circuit in module S4										+	+
Short circuit in module B						+		+	+	+	
Short circuit in supply circuit	+		+		+						
Short circuit in phase winding	+		+		+						
Open circuit in power valve S2							+				
Open circuit in power valve B						+	+				
Open circuit in supply circuit			+		+						
Open circuit in phase winding B	+	+		+							

For the cooling unit, it is sufficient to measure only the output quantity (pressure or flow rate of the cooling fluid). For the control system, it is necessary to detect the presence or absence of control pulses and their parameters, which makes it convenient to assess functionality based on the parameters of the components comprising the system. To

predict the system's condition and detect gradual failures, it is advisable to use diagnostics based on parameters.

TABLE V. PARAMETERS CHARACTERIZING THE CONDITION OF SHIPBOARD POWER SYSTEMS

Subsystem	Diagnostic Parameters of Elements						
	$\frac{du}{dt}$	$\frac{di}{dt}$	$t_в$	*Temperature*	*Resistance*	*Pressure*	*the feedback signal*
Power circuits	+	+	+	+	+	–	–
Cooling subsystem	–	–	–	–	–	+	–
Control, protection devices	–	–	–	–	–	–	+

Malfunctions in the power circuit of a ship converter occur due to the formation of short-circuited loops, which cause redistribution of currents and loads among individual components and branches of the circuit. The loss of control in one of the parallel valves increases the current stress on the remaining ones. Higher current through the semiconductor elements leads to overheating, gradually degrading the valve structure. Overheating may result in insulation failure, which means that temperature conditions can be indirectly monitored by observing the current. If one of the series-connected valves fails, the voltage across the remaining ones rises, further reducing their reliability, which confirms the research assumptions [14-16]. Therefore, since it is difficult to directly measure parameters such as temperature or bulk resistance of semiconductor structures, the most informative diagnostic indicators of ship power systems are the currents and voltages across the power valves , which confirms the research assumptions [17-19] and which confirms the findings of the research [20, 21].

VI. CONCLUSIONS

Failures in the power circuit of the ship converter are associated with the emergence of short-circuited loops, which lead to the redistribution of currents and loads across individual components and circuit branches. Loss of control in one of the parallel valves increases the current load on the remaining ones. An increase in currents flowing through the semiconductor elements of the circuit causes excessive heating of components included in the circuit, which, in turn, leads to gradual degradation of the valve structure. As a result of overheating, insulation breakdown may occur, meaning that temperature conditions can be indirectly tracked by monitoring current. If one of the series-connected valves fails (breaks down), the voltage across the remaining ones increases, which also reduces the reliability of the elements.

Therefore, given the difficulties in directly measuring parameters such as temperature and bulk resistance of semiconductor structures, currents and voltages on the power valves are chosen as the diagnostic parameters that best characterize the state of shipboard power systems.

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Refined Calculation of Steady Energy Modes of a Frequency-Regulated Surface Permanent Magnet Synchronous Motor

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***Abstract* — For a frequency-regulated surface permanent magnet synchronous motor a refined calculation of the main electromagnetic power losses is proposed, which simultaneously takes into account all types of main power losses present in this motor, as well as the components of electrical losses caused by the transportation of magnetic power losses to the magnetization circuit and additional power losses through the air gap of this motor, respectively. Based on the obtained refined calculation dependencies for this motor its electromagnetic processes and energy characteristics (main electromagnetic power losses, main consumed active power, efficiency and power factor) are quantitatively estimated in steady operating modes of the mentioned motor at the different values of the speed and useful torque on the motor shaft.**

Keywords — surface permanent magnet synchronous motor, main electromagnetic power losses, frequency regulation.

I. INTRODUCTION

Due to the development of industrial production and wide implementation in various branches of the economy of frequency-regulated (f-r) synchronous motors with permanent magnets installed inside (IPMSM – Interior Permanent Magnet Synchronous Machine) or on the surface (SPMSM – Surface Permanent Magnet Synchronous Machine) of the rotor of these motors, now in practice there is an actual called for a refined calculation of the energy characteristics of these motors with frequency regulation in a wide range of speed and load changes. Based on the now observed rise in the cost of electricity in the world and Ukraine, a refined definition of the energy characteristics of these motors will allow subsequently to optimize (minimize) their operating modes in exploitation, thereby reducing power losses in them and their overall electricity consumption.

Despite of the existence of known publications devoted to the calculation and research of energy modes of IPMSM and SPMSM, all of them have a number of shortcomings that reduce the precision of calculating their electromagnetic and energy processes. In particular, in monographs [1, 2] and article [3], additional power losses always are not taken into account when calculation of power losses are realized in these motors. In contrast to these publications, in monograph [4] and articles [5–8] in the calculation: power losses of the mentioned motors, all types of main power losses present in them (including additional power losses) are taken into by the transportation of magnetic power losses to the magnetization circuit of the motors; however, the component of electrical losses caused by the transportation of additional power losses through the air gap of the motors, is not taken into account in the calculations.

The purpose of the article is to obtain analytical dependencies for a more precise calculation of the stator current and the main electromagnetic power losses of a three-phase SPMSM and to research, using the obtained more precise dependencies, the steady main electromagnetic and energy processes of this motor in the operating ranges of its speed and load changes.

II. INITIAL ASSUMPTIONS

a) three-phase stator and rotor windings for different phases of the motor are made mutually symmetrical and constructional identical;

b) the motor air gap is the same along the entire inner circumference of the stator;

c) there is no zero sequence in the three-phase system of stator supply voltages (currents) of the motor;

d) internal parameters of the motor circuit: the active resistance R_s of the stator winding (which corresponds to the motor operating temperature) and for SPMSM equal to each other dissipation inductances: $L_{\sigma s.d} = L_{\sigma s.q} = L_{\sigma s}$ and magnetizing inductances: $L_{md} = L_{mq} = L_m$ along the longitudinal «d» and transverse «q» axes of this motor – were assumed to be unchanged;

e) only the main (caused by the first harmonic components of the phase stator voltages and currents) total power losses ΔP_m SPMSM were taken into account, which contain the following instantaneous losses components [4 – 8]:

- electrical ΔP^*_e (in the three-phase stator winding);
- magnetic ΔP_{ir} (in steel of the stator core);
- mechanical ΔP_{mech} (caused by friction in the motor shaft bearings and to self-ventilation);
- additional ΔP_{ad} (caused in the air gap of the motor by magnetic induction pulsations in the stator teeth).

For further simplification of analytical dependencies, describing electromechanical and energy parameters of the SPMSM mode, the method of generalized vectors [1, 2, 5] and the known relative system of units, accepted for

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alternating current electric machines [5], were used. The researches were carried out by the method of computer simulation modelling.

The mentioned electrical ΔP^*_{e} (in stator winding), magnetic ΔP_{ir}, additional ΔP_{ad} and mechanical ΔP_{mech} types of main power losses, as well as main electromagnetic ΔP^*_{em} and total ΔP_m power losses, instantaneous η and nominal η_n values of motor efficiency, instantaneous values of electromagnetic power P_{em} and power factor $\cos(\varphi)$ SPMSM are determined from known analytical dependencies [5, 8 – 10]:

$$\left.\begin{aligned}
&\Delta P^*_{e}= R_s\cdot I_s^2,\ \ \Delta P_{e.n} = R_s\cdot I_{s.n}^{\ 2}, R_{ad} = \Delta P_{ad.n}/I_{sn}^2,\\
&P_{1.n} = \frac{P_{2.n}}{\eta_n},\ \Delta P_{i\,r} = \Delta P_{i\,r.n}\left(\frac{\Psi_m}{\Psi_{m.n}}\right)^2\left(\frac{\omega_1}{\omega_{1n}}\right)^{\lambda},\\
&\Delta P_{ad.n} = P_{1.n} - \Delta P_{e.n} - \Delta P_{i\,r.n} - \Delta P_{mech.n}, \omega_s = n/n_n,\\
&\Delta P_{ad} = R_{ad}\omega_s^2 I_s^2,\ \Delta P_{mech} = \Delta P_{mech.n}\cdot(\omega_s/\omega_{sn})^2,\\
&\Delta T_{mech} = \Delta T_{mech.n}\cdot\left(\frac{\omega}{\omega_n}\right),\ \Delta T_{mech.n} = \frac{\Delta P_{mech.n}}{\omega_n},\\
&\Delta P_{em}^{\ *} = \Delta P^*_{e} + \Delta P_{i\,r} + \Delta P_{ad},\ \cos\varphi = P_1/U_s I_s,\\
&\eta = \begin{cases} P_2/P_1 & \text{at } T \ge 0,\\ P_1/P_2 & \text{at } T < 0; \end{cases}\quad \eta_n = P_{2n}/P_{1n},\\
&P_2 = \omega T,\ \ P_1 = P_2 + \Delta P_m, \Delta P_m = \Delta P_{em} + \Delta P_{mech},\\
&P_{em} = P_1 - \Delta P^*_{e} - \Delta P_{i\,r} = \omega\cdot(T + \Delta T_{mech}) + \Delta P_{ad}.
\end{aligned}\right\} \quad (1)$$

In formula (1) the following notations are used: R_s – active resistance of the phase stator winding; I_s and I_{sn} – respectively instantaneous and nominal values of the module of the generalized vector stator current $\bar{I}_s$ of the motor (in the accepted relative system of units: I_{sn} = 1 r.u.); Ψ_m and Ψ_{mn} –respectively instantaneous and nominal values of the module of generalized vector $\overline{\Phi}_m$ of magnetic flux SPMSM; n and n_n – instantaneous and nominal values of speed [rpm] of the motor respectively; ω_s and ω_{sn} – respectively instantaneous and nominal values of the module of angular frequency of phase stator voltages and currents of the motor (which in steady modes in the accepted relative system of units are respectively equal to the instantaneous ω and nominal ω_h values of angular frequency of rotation of the rotor of the motor; the values of ω_{sn} and ω_h are equal to 1r.u.); λ –coefficient, which takes into account the change of magnetic power losses ΔP_{ir} SPMSM from the angular frequency ω_s of the stator magnetic field [5, 9] (for the researched motor in the article: λ = 1.55); $\Delta P_{e.n}$, $\Delta P_{ir.n}$ and $\Delta P_{mech.n}$ – nominal values (inherent in the nominal mode of SPMSM operation) of electric, magnetic and mechanical power losses of the motor, respectively; R_{ad} – equivalent active resistance to account the additional power losses ΔP_{ad} SPMSM; P_1 and P_{1n} – instantaneous and nominal values of consumed active power of the motor respectively; P_2 and P_{2n} – instantaneous and nominal values of useful power on the motor shaft; ΔT_{mech} and $\Delta T_{mech.n}$ – instantaneous and nominal values of mechanical losses of motor torque respectively; T and $(T+\Delta T_{mech.n})$ – instantaneous values of useful on the shaft and rotating torques of this motor respectively (positive values of torque T correspond to motor, and negative – to generator mode of SPMSM operation).

In steady modes of operation of the f-r SPMSM, the most convenient and sufficiently precise calculation of its electromagnetic processes is realized from the known analytical dependences [5]:

$$\left.\begin{aligned}
&U_{sd} = R_s I_{sd} + j\omega_s L_{\sigma s} I_{sd} + j\omega_s L_m I_{1d} - \omega_s L_m I_{1q},\\
&U_{sq} = R_s I_{sq} + j\omega_s L_{\sigma s} I_{sq} + j\omega_s L_m I_{1q} +\\
&+ \omega_s\left(L_m I_1 + \Psi_f\right),\ \Psi_{mn} = \sqrt{\Psi_f^2 + \left(L_m I_{1q.n}\right)^2},\\
&U_s = \sqrt{U_{sd}^2 + U_{sq}^2},\ I_{md} = I_f - I_{1d},\ \ I_{mq} = I_{1q},\\
&I_f = \Psi_f/L_m,\ \ U_f = \omega_s\cdot\Psi_f,\ \ I_s = \sqrt{I_{sd}^2 + I_{sq}^2},\\
&I_m = \sqrt{I_{md}^2 + I_{mq}^2},\ T + \Delta T_{mech} = \Phi_{md}\cdot I_{1q},\\
&\Phi_{mq} = L_m I_{1q}, \Phi_{md} = \Psi_f + L_m I_{1d}, \Phi_m = \sqrt{\Phi_{md}^2 + \Phi_{mq}^2}
\end{aligned}\right\}, (2)$$

in which generalized vectors (stator voltage $\overline{U}_s$ and current $\bar{I}_1$ or $\bar{I}_s$, magnetic flux $\overline{\Phi}_m$ and magnetizing current $\bar{I}_m$) were set through projections (respectively: U_{sd} and U_{sq}, I_{1d} and I_{1q} or I_{sd} and I_{sq}, Φ_{md} and Φ_{mq}, I_{md} and I_{mq}) on the axes of the rotating orthogonal coordinate system (OCS) «d-q», oriented by the real (longitudinal) axis «d» in the direction of the geometric axis of the magnetic field, created by permanent magnets installed on the motor rotor.

In publications [3, 7], due to the use of equivalent SPMSM substitution diagrams together with dependences (2) (containing additionally equivalent resistors, simulating the presence of magnetic power losses in the motor), the mentioned power losses are taken into account. At the same time, the instantaneous values of I_{sd} and I_{sq} projections of the generalized stator current vector are determined, in which the component of electric power losses of the motor, caused by the transportation of its magnetic power losses to the magnetizing circuit, is also taken into account. The deficiency of the calculations, proposed in these publications, is that they do not take into account the component of electric losses, caused by the transportation of the additional power losses through the SPMSM air gap.

In formula (2) the following notations are applied: Ψ_f and i_f – are the rotor magnetic flux and SPMSM excitation current, respectively, generated by permanent magnets; I_1, I_{1d} and I_{1q} – are the modulus, longitudinal and transverse projections of the generalized stator current vector $\bar{I}_1$, corresponding to the idealized SPMSM representation (when magnetic and addional power losses are not taken into account); Ψ_m and I_m – modules of the generalized vectors of magnetic flux $\overline{\Psi}_m$ and magnetizing current $\bar{I}_m$, respectively; j – imaginary unit (by means of which, when the vector diagram is graphically constructing, the sections of projections of vectors turn by 90 electrical degrees against the clockwise direction [5]). The vector diagram for SPMSM constructed from dependences (2) is presented in Fig. 1 for the second speed zone.

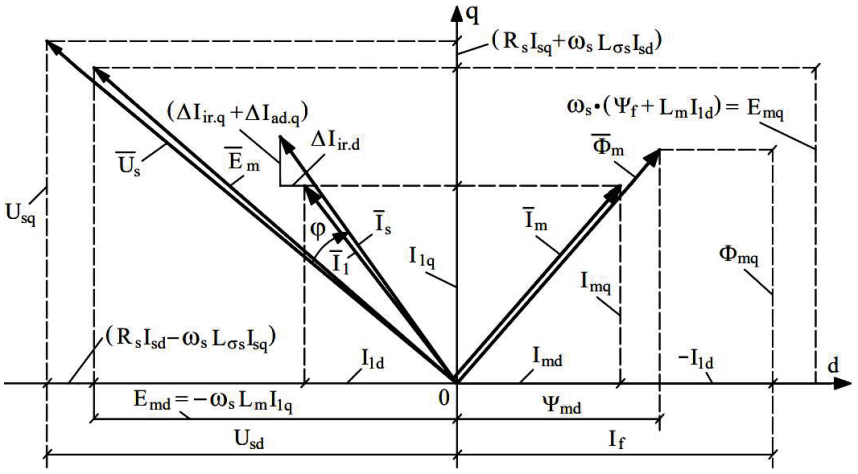


Fig. 1. Vector diagram for steady motor mode of operation (at $I_{1d} < 0$ and $\omega_s > \omega_{sn}$) of SPMSM

III. REFINED CALCULATED ANALYTICAL DEPENDENCIES

We selected known ones and obtained new refined analytical dependencies for calculation the instantaneous values of projections I_{sd} , I_{sq} and modulus I_s of the generalized stator current vector $\bar{I}_s$ SPMSM, which simultaneously take into account all the main power losses present in this motor, as well as the stator current increments, caused by transportation of magnetic power losses ΔP_{ir} to the magnetizing circuit and additional power losses ΔP_{ad} through the air gap of the motor.

Based on the dependence for the instantaneous magnetic power losses ΔP_{ir} in (1) and known relations [5, 8, 9]:

$$\left.\begin{aligned} &\Delta P_{ir} = \Delta P_{ir.n} \cdot \left(\Psi_m / \Psi_{mn}\right)^2 \cdot \left(\omega_s / \omega_{sn}\right)^{\lambda} = \\ &= \Delta P_{ir.n} \cdot \left(E_m / E_{mn}\right)^2 \cdot \left(\omega_s / \omega_{sn}\right)^{\lambda-2}, \\ &E_m = \omega_s \cdot \Psi_m, \quad E_{mn} = \omega_{sn} \cdot \Psi_{mn} \end{aligned}\right\} \qquad (3)$$

for the instantaneous E_m and nominal E_{mn} values of modulus of the SPMSM generalized electromotive magnetization force (EMF) vector $\bar{E}_m$, let us make an energy balance for the instantaneous magnetic power losses of this motor:

$$\left.\begin{aligned} &\Delta P_{ir} = \Delta P_{ir.d} + \Delta P_{ir.q}, \; E_m = \sqrt{E_{md}^2 + E_{mq}^2}, \\ &E_{md} = -\omega_s L_m I_{1q}, E_{mq} = \omega_s \cdot \left(\Psi_f + L_m I_{1d}\right), \\ &\Delta P_{ir.d} = \Delta P_{ir.n} \cdot \left(E_{md} / E_{mn}\right)^2 \cdot \left(\omega_s / \omega_{sn}\right)^{\lambda-2}, \\ &\Delta P_{ir.q} = \Delta P_{ir.n} \cdot \left(E_{mq} / E_{mn}\right)^2 \cdot \left(\omega_s / \omega_{sn}\right)^{\lambda-2} \end{aligned}\right\}, \qquad (4)$$

where $\Delta P_{ir.d}$ and $\Delta P_{ir.q}$ – are the components of the instantaneous magnetic power losses of the SPMSM along the longitudinal and transverse axes, respectively; E_{md} and E_{mq} – are the projections of the instantaneous generalized vector of the EMF $\bar{E}_m$ on the longitudinal and transverse axes of the rotating OCS «d – q», respectively.

Let us determine, based on (3), for SPMSM from the expression:

$$R_{ir} = \frac{E_m^2}{\Delta P_{ir}} = \frac{E_{mn}^2}{\Delta P_{ir.n}} \cdot \left(\frac{\omega_{sn}}{\omega_s}\right)^{\lambda-2} \qquad (5)$$

equivalent active resistance R_{ir}, which in the substitution diagrams of this motor [3, 7 – 9] simulates the instantaneous value ΔP_{ir} of its magnetic power losses.

Taking into account (4) and (5), we obtain analytical dependencies:

$$\left.\begin{aligned} \Delta I_{ir.d} &= \frac{E_{md}}{R_{ir}} = \frac{E_{md} \cdot \Delta P_{ir.n}}{E_{mn}^2} \cdot \left(\frac{\omega_s}{\omega_{sn}}\right)^{\lambda-2}, \\ \Delta I_{ir.q} &= \frac{E_{mq}}{R_{ir}} = \frac{E_{mq} \cdot \Delta P_{ir.n}}{E_{mn}^2} \cdot \left(\frac{\omega_s}{\omega_{sn}}\right)^{\lambda-2} \end{aligned}\right\} \qquad (6)$$

to determine the increments $\Delta I_{ir.d}$ and $\Delta I_{ir.q}$, respectively, for the longitudinal and transverse projections of the generalized $\bar{I}_s$ SPMSM stator current vector, caused by transporting its magnetic power losses to the motor magnetizing circuit.

After subtraction the above mentioned increments $\Delta I_{ir.d}$ and $\Delta I_{ir.q}$ from the generalized stator current vector $\bar{I}_s$, we obtain the following generalized vector $\bar{I}_{1\Sigma}$, which calculated from the relations:

$$\left.\begin{aligned} &\bar{I}_{1\Sigma} = I_{1\Sigma d} + jI_{1\Sigma q}, \; I_{1\Sigma d} = I_{sd} - \Delta I_{ir.d}, \\ &I_{1\Sigma q} = I_{sq} - \Delta I_{ir.q}, \end{aligned}\right\}, \qquad (7)$$

where $\Delta I_{1\Sigma d}$ and $\Delta I_{1\Sigma q}$ – are the longitudinal and transverse projections of the generalized vector $\bar{I}_{1\Sigma}$, respectively. According to the last relationship from formula (1), the values of the increments $\Delta I_{ir.d}$ and $\Delta I_{ir.q}$ of the stator current vector $\bar{I}_{1\Sigma}$ have not physically influence on the electromagnetic power P_{em}, which is transmitted through the SPMSM air gap. Consequently, these increments $\Delta I_{ir.d}$ and $\Delta I_{ir.q}$ also have not influence on the additional power losses ΔP_{ad} of the motor, which is part of the power P_{em}.

Therefore, in order to increase the precision of determination of the instantaneous additional power losses ΔP_{ad} f-r SPMSM, we will calculate them (in contrast to the relations from (1), known in publications [3 – 5]) differently by a way, excluding the mentioned increments $\Delta I_{ir.d}$ and $\Delta I_{ir.q}$ from the module I_s of the generalized stator current vector. Namely, we will find the additional losses ΔP_{ad} directly

proportional to the second degree of the modulus $I_{1\Sigma}$ of the stator current from the following mathematical expressions:

$$\left.\begin{aligned}\Delta P_{\text{ad}} &= R^{*}_{\text{ad}}\cdot\omega_{\text{s}}^{2}\cdot I_{1\Sigma}^{2}=R^{*}_{\text{ad}}\cdot\omega_{\text{s}}^{2}\cdot\left(I_{1\Sigma\text{d}}^{2}+I_{1\Sigma\text{q}}^{2}\right),\\ R^{*}_{\text{ad}} &= \frac{\Delta P_{\text{ad.n}}}{I_{1\Sigma\text{d.n}}^{2}+I_{1\Sigma\text{q.n}}^{2}}=\frac{\Delta P_{\text{ad.n}}}{I_{1\Sigma\text{.n}}^{2}}=R_{\text{ad}}\cdot\left(\frac{I_{\text{sn}}}{I_{1\Sigma\text{.n}}}\right)^{2}\end{aligned}\right\} \quad (8)$$

through the instantaneous values of the mentioned modulus $I_{1\Sigma}$ or longitudinal $I_{1\Sigma.\text{d}}$ and transverse $I_{1\Sigma.\text{q}}$ projections of the generalized vector $\bar{I}_{1\Sigma}$ on the axes of the rotating OCS «d – q». In formula (8): R^{*}_{ad} – is the refined value of the equivalent (intended for calculation of the additional power losses) motor active resistance, which is calculated through the nominal (that is corresponding to the nominal SPMSM mode) values of: the additional power losses $\Delta P_{\text{ad.n}}$, as well as the module $I_{1\Sigma.\text{n}}$, longitudinal $I_{1\Sigma\text{d.n}}$ and transverse $I_{1\Sigma\text{q.n}}$ projections of the motor stator current vector $\bar{I}_{1\Sigma.\text{n}}$.

In order to find the increment $\Delta I_{\text{q.ad}}$ of the transverse projection of the generalized stator current vector $\bar{I}_{\text{s}}$ (caused in SPMSM by the transmission of additional power losses ΔP_{ad} through the air gap of the motor), we shall make the energy balance for the electromagnetic power P_{em} of this motor, taking into consideration (8) and the last relation from formula (1):

$$P_{\text{em}}=\omega_{\text{s}}\cdot T_{\text{em}}=\omega_{\text{s}}\cdot\left(T+\Delta T_{\text{mech}}\right)+R^{*}_{\text{ad}}\cdot\omega_{\text{s}}^{2}\cdot I_{1\Sigma}^{2}. \quad (9)$$

Substituting into expression (9) the known relations for determining the electromagnetic T_{em} and rotating $(T+\Delta T_{\text{mech}})$ moments in [1, 5]:

$$T_{\text{em}}=\Phi_{\text{md}}I_{1\Sigma\text{q}} \quad \text{and} \quad T+\Delta T_{\text{mech}}=\Phi_{\text{md}}I_{1\text{q}}, \quad (10)$$

we obtain, taking into account the relation

$$I_{1\Sigma\,\text{q}}^{2}=\left(I_{1\text{q}}+\Delta I_{\text{ad.q}}\right)^{2}=I_{1\text{q}}^{2}+2I_{1\text{q}}\cdot\Delta I_{\text{ad.q}}+\Delta I_{\text{ad.q}}^{2}, \quad (11)$$

the following equation:

$$\omega_{\text{s}}\cdot\Phi_{\text{md}}\cdot\left(I_{1\Sigma\,\text{q}}-I_{1\text{q}}\right)=R^{*}_{\text{ad}}\cdot\omega_{\text{s}}^{2}\cdot\left(I_{1\Sigma\,\text{d}}^{2}+I_{1\Sigma\,\text{q}}^{2}\right). \quad (12)$$

Reducing the left and right parts by the multiplier ω_{S} and denoting the difference $(I_{1\Sigma\text{q}}-I_{1\text{q}})$ between the transverse projections $I_{1\Sigma\text{q}}$ and $I_{1\text{q}}$ as the increment $\Delta I_{\text{ad.q}}$ of the transverse projection I_{sq} of the generalized motor stator current vector $\bar{I}_{\text{s}}$ caused by transporting the additional power losses ΔP_{ad} through the air gap SPMSM, we transform equation (12) to the form of

$$\Delta I_{\text{ad.q}}^{2}+\Delta I_{\text{ad.q}}\cdot\left(2I_{1\text{q}}-\frac{\Phi_{\text{md}}}{|\omega_{\text{s}}|R^{*}_{\text{ad}}}\right)+I_{1\Sigma\,\text{d}}^{2}+I_{1\text{q}}^{2}, \quad (13)$$

where the unknown variable value is the increment $\Delta I_{\text{ad.q}}$.

The solution of the quadratic algebraic equation (13) is an analytic relationship:

$$\Delta I_{\text{ad.q}}=\left(\frac{0.5\Phi_{\text{md}}}{|\omega_{\text{s}}|\cdot R^{*}_{\text{ad}}}-I_{1\text{q}}\right)-\left[\left(\frac{0.5\Phi_{\text{md}}}{|\omega_{\text{s}}|\cdot R^{*}_{\text{ad}}}-I_{1\text{q}}\right)^{2}-I_{1\Sigma\,\text{d}}^{2}-I_{1\text{q}}^{2}\right]^{2}, \text{ where } I_{1\Sigma\,\text{d}}=I_{1\text{d}}. \quad (14)$$

From the analysis of dependence (14) it is revealed that as the frequency ω_{S} of the stator (equal to the speed ω of the rotor) approaches zero, then the increment $\Delta I_{\text{ad.q}}$ tends to zero.

To create instantaneous values of longitudinal projection Φ_{md} of magnetic flux and rotating torque $(T+\Delta T_{\text{mech}})$ SPMSM at a given angular frequency $\omega_{\text{S}}=\omega$ of the motor stator or rotor, the specific values are required of: longitudinal $I_{1\text{d}}$ and transverse $I_{1\text{q}}$ projections and module I_1 of the generalized stator current vector $\bar{I}_1$ (which correspond to the idealized representation of SPMSM), as well as refined (corresponding to the real representation of this motor) values of module I_{S} and orthogonal projections I_{sd} and I_{sq} of the generalized stator current vector $\bar{I}_{\text{s}}$ SPMSM. The values of the above quantities are determined from formulas (2), (6) and (14) by means of the following analytical dependencies:

$$\left.\begin{aligned}&I_{1\text{d}}=\left(\Phi_{\text{md}}-\Psi_{\text{f}}\right)/L_{\text{m}},\ I_{1\text{q}}=\left(T+\Delta T_{\text{mech}}\right)/\Phi_{\text{md}},\\ &I_1=\sqrt{I_{1\text{d}}^{2}+I_{1\text{q}}^{2}},\ \bar{I}_1=I_{1\text{d}}+jI_{1\text{q}},\\ &I_{\text{sd}}=I_{1\text{d}}+\Delta I_{\text{ir.d}},\ \ I_{\text{sq}}=I_{1\text{q}}+\Delta I_{\text{ir.q}}+\Delta I_{\text{ad.q}},\\ &I_{\text{s}}=\sqrt{\left(I_{1\text{d}}+\Delta I_{\text{ir.d}}\right)^{2}+\left(I_{1\text{q}}+\Delta I_{\text{ir.q}}+\Delta I_{\text{ad.q}}\right)^{2}},\\ &\bar{I}_{\text{s}}=\left(I_{1\text{d}}+\Delta I_{\text{ir.d}}\right)+j\left(I_{1\text{q}}+\Delta I_{\text{ir.q}}+\Delta I_{\text{ad.q}}\right).\end{aligned}\right\} \quad (15)$$

IV. RESEARCH MATERIALS

In steady operation modes of the SPMSM (with parameters from [11] and nominal values of useful power 35.2kW and speed 2800rpm) the electromag-netic and energy processes of this motor were calculated. The mentioned calculations were carried out for different instantaneous values of speed and load, using the obtained refined calculated analytical dependences: for the first (at speeds $|n|\le n_{\text{n}}$ and at null longitudinal projection $I_{1\text{d}}=0$ of stator current) and for the second (at speeds $|n|>n_{\text{n}}$, and at negative projection $I_{1\text{d}}<0$ of stator current) zones of speeds. In the second zone, the choice of values of projections $I_{1\text{d}}$ and $I_{1\text{q}}$ of stator current corresponded to minimization of the stator voltage modulus ($U_{\text{s}}=\min$) SPMSM [12].

Results of refined calculations of electromagnetic processes and power characteristics of the researched motor are presented in Table I, Table II and Table III. Also in Table II and Table III there are additionally calculated ratios $\Delta P_{\text{em}}/\Delta P^{*}_{\text{em}}$ between inaccurately determined (without taking into account components of electric power losses, caused by transportation of magnetic and additional power losses to the magnetizing circuit and through the SPMSM air

TABLE I. THE VALUES OF SPEED n, RELATIVE STATOR FREQUENCY ω_{S} AND CORRESPONDING TO THEM MECHANICAL LOSSES OF POWER ΔP_{MECH} AND TORQUE ΔT_{MECH} OF THE MOTOR FOR RESEARCHED SPMSM

Value	Zone 1			Zone 2		
n [rpm]	300	1500	2800	3500	4000	4500
ω*s *[r.u.]	0.1071	0.5357	1	1.25	1.4286	1.6071
ΔP*mech *[kW]	0.4018	0.1004	0.350	0.5469	0.7143	0.9040
ΔT*mech *[Nm]	0.1256	0.6278	1.172	1.465	1.674	1.883

TABLE II. CALCULATED STEADY ELECTROMAGNETIC PROCESSES AND ENERGY CHARACTERISTICS OF F-R SPMSM FOR THE FIRST ZONE OF SPEEDS (AT $|n| \le n_n$ AND $I_{1d} = 0$)

n	$\frac{T}{T_n}$	$\frac{I_s}{I_{sn}}$	$\frac{I_s}{I_1}$	$\frac{\Phi_m}{\Phi_{mn}}$	$\frac{\Delta I_{ir.d}}{I_s}$	$\frac{\Delta I_{ir.q}}{I_s}$	$\frac{\Delta I_{ad.q}}{I_s}$	ΔP_e^*	ΔP_{ir}	ΔP_{ad}	ΔP_{em}^*	$\frac{\Delta P_{em}}{\Delta P_{em}^*}$	$\frac{P_1}{P_{1n}}$	η	$cos(\varphi)$	$\frac{U_s}{U_{sn}}$
rpm	–	–	–	–	*%*	*%*	*%*	***kW***	***kW***	***kW***	***kW***	–	–	*%*	–	–
2800	0.5	0.5134	1.029	0.9639	-0.4145	2.531	0.2544	0.1833	0.4794	0.0469	0.7203	0.9852	0.5038	94.32	0.9850	0.9501
	1	1.00	1.018	1	-0.4187	1.291	0.5116	0.7190	0.5160	0.1850	1.447	0.9816	1.00	95.20	0.9460	1.00
	1.5	1.502	1.017	1.057	-0.4194	0.8650	0.7701	1.607	0.5767	0.4157	2.653	0.9797	1.505	94.70	0.8916	1.0717
	2	2.001	1.017	1.132	-0.4192	0.6495	1.030	2.848	0.6614	0.7405	4.348	0.9774	2.025	93.86	0.8307	1.1619
	-0.5	0.4654	0.9704	0.9629	0.4393	2.792	0.2565	0.1693	0.4785	0.0429	0.6808	1.015	-0.4471	94.08	-0.9842	0.9307
	-1	0.9511	0.9816	0.9981	0.4343	1.366	0.5098	0.6909	0.5141	0.1742	1.354	1.019	-0.9037	95.08	-0.9414	0.9624
	-1.5	1.431	0.9836	1.055	0.4334	0.9059	0.7618	1.565	0.5738	0.3926	2.481	1.021	-1.348	94.54	-0.8792	1.0190
	-2	1.915	0.9834	1.129	0.4335	0.9785	1.013	2.792	0.6575	0.6968	1.054	1.023	-1.779	93.61	-0.8078	1.0967
1500	0.5	0.5045	1.020	0.9636	-0.2965	1.827	0.1359	0.1800	0.1821	0.0132	0.3826	0.9810	0.2674	95.19	0.9851	0.5131
	1	0.9959	1.012	0.9995	-0.2988	0.9257	0.2731	0.7124	0.1959	0.0524	0.9781	0.9822	0.5378	94.67	0.9476	0.5435
	1.5	1.489	1.010	1.057	-0.2993	0.6193	0.4108	1.597	0.2189	0.1177	1.967	0.9830	0.8187	93.28	0.8960	0.5853
	2	1.983	1.010	1.131	-0.2994	0.4649	0.5489	2.835	0.2510	0.2095	3.354	0.9826	1.110	91.73	0.8390	0.6364
	-0.5	0.4743	0.9796	0.9631	0.3087	1.944	0.1377	0.1725	0.1819	0.0126	0.3600	1.019	-0.2419	95.04	-0.9837	0.4945
	-1	0.9617	0.9878	0.9985	0.3062	0.9586	0.2739	0.6974	0.1955	0.0507	0.9268	1.018	-0.4809	94.46	-0.9392	0.5078
	-1.5	1.448	0.9896	1.055	0.3056	0.6368	0.4097	1.575	0.2183	0.1142	1.875	1.017	-0.7094	92.89	-0.8735	0.5350
	-2	1.933	0.9899	1.130	0.3055	0.4770	0.5451	2.805	0.2502	0.2028	3.201	1.018	-0.9275	91.09	-0.7971	0.5742
300	0.5	0.4943	1.008	0.9634	-0.1238	0.7695	0.0272	0.1770	0.0150	0.0005	0.1954	0.9854	0.0056	90.55	0.9863	0.1100
	1	0.9841	1.004	0.9991	-0.1242	0.3865	0.0546	0.7064	0.0162	0.0021	0.7309	0.9914	0.1215	83.79	0.9585	0.1229
	1.5	1.474	1.003	1.056	-0.1244	0.2580	0.0820	1.588	0.0180	0.0047	1.622	0.9933	0.1964	77.77	0.9250	0.1374
	2	1.964	1.003	1.131	-0.1244	0.1936	0.1094	2.823	0.0207	0.0083	2.869	0.9940	0.2808	72.52	0.8910	0.1530
	-0.5	0.4844	0.9919	0.9633	0.1258	0.7853	0.0275	0.1755	0.0150	0.0005	0.1882	1.015	-0.0456	89.65	-0.9803	0.0917
	-1	0.9733	0.9956	0.9989	0.1254	0.3908	0.0548	0.7034	0.0162	0.0021	0.7154	1.009	-0.0822	80.74	-0.9172	0.0878
	-1.5	1.462	0.9966	1.056	0.1252	0.2602	0.0821	1.584	0.0180	0.0046	1.596	1.007	-0.1092	71.50	-0.8069	0.0883
	-2	1.951	0.9970	1.130	0.1252	0.1950	0.1094	2.817	0.0207	0.0082	2.828	1.006	-0.1266	62.18	-0.6660	0.0930

TABLE III. CALCULATED STEADY ELECTROMAGNETIC PROCESSES AND ENERGY CHARACTERISTICS OF F-R SPMSM FOR THE SECOND ZONE OF SPEEDS (AT $|n| > n_n$ AND $I_{1d} < 0$)

n	$\frac{T}{T_n}$	$\frac{I_s}{I_{sn}}$	$\frac{I_s}{I_1}$	I_{1d}	$\frac{\Phi_m}{\Phi_{mn}}$	$\frac{\Delta I_{ir.d}}{I_s}$	$\frac{\Delta I_{ir.q}}{I_s}$	$\frac{\Delta I_{ad.q}}{I_s}$	ΔP_e^*	ΔP_{ir}	ΔP_{ad}	ΔP_{em}^*	$\frac{P_1}{P_{1n}}$	η	$cos(\varphi)$	$\frac{U_s}{U_{sn}}$
rpm	–	–	–	***r.u.***	–	*%*	*%*	*%*	***kW***	***kW***	***kW***	***kW***	–	*%*	–	–
3500	0.5	0.7895	1.019	-0.492	0.8192	-0.3651	1.561	0.5962	0.4414	0.4894	0.1773	1.125	0.6387	93.00	0.7716	1.00
	1	1.787	1.021	-0.972	0.7911	-0.3924	0.5599	1.687	2.253	0.4564	0.9224	3.729	1.301	91.32	0.6942	1.00
	1.5	2.2413	1.025	-0.822	0.9395	-0.4358	0.4788	1.976	3.518	0.6437	1.453	5.794	1.949	91.45	0.6855	1.209
	-0.5	0.6779	0.9790	-0.417	0.8390	0.3930	1.871	0.5299	0.353	0.5133	0.1392	0.9907	-0.552	92.94	-0.7765	1.00
	-1	1.355	0.9807	-0.641	0.8427	0.4354	0.8562	1.140	1.405	0.5179	0.5480	2.417	-1.106	93.14	-0.7788	1.00
	-1.5	2.215	0.9755	-0.913	0.9299	0.4524	0.4647	2.086	3.793	0.6305	1.454	5.693	-1.608	90.26	-0.687	1.1195
4000	0.5	1.113	1.017	-0.845	0.7210	-0.3245	1.027	1.127	0.8822	0.4663	0.4653	1.843	0.7472	90.86	0.6401	1.00
	1	2.098	1.028	-1.209	0.7701	-0.4068	0.4547	2.560	3.067	0.5318	1.663	5.434	1.519	89.37	0.6159	1.121
	-0.5	0.9619	0.9835	-0.748	0.7442	0.3398	1.240	0.9699	0.7039	0.4968	0.3614	1.539	-0.6173	90.94	-0.6121	1.00
	-1	2.073	0.9732	-1.303	0.7571	0.4216	0.4366	2.731	3.339	0.5141	1.664	5.341	-1.189	87.60	-0.5343	1.024
4500	0.5	1.443	1.018	-1.159	0.6420	-0.3126	0.7246	1.925	1.477	0.4438	0.9939	2.969	0.8669	88.09	0.5729	1.00
	1	2.110	1.031	-1.211	0.7706	-0.4328	0.4819	2.897	3.082	0.6392	2.127	6.042	1.710	89.33	0.6144	1.258
	-0.5	1.227	0.9836	-1.023	0.6700	0.3217	0.9138	1.572	1.144	0.4832	0.7395	2.330	-0.6753	88.42	-0.5252	1.00
	-1	2.064	0.9704	-1.302	0.7564	0.4509	0.4683	3.059	3.327	0.6160	2.0879	5.837	-1.340	87.73	-0.5359	1.155

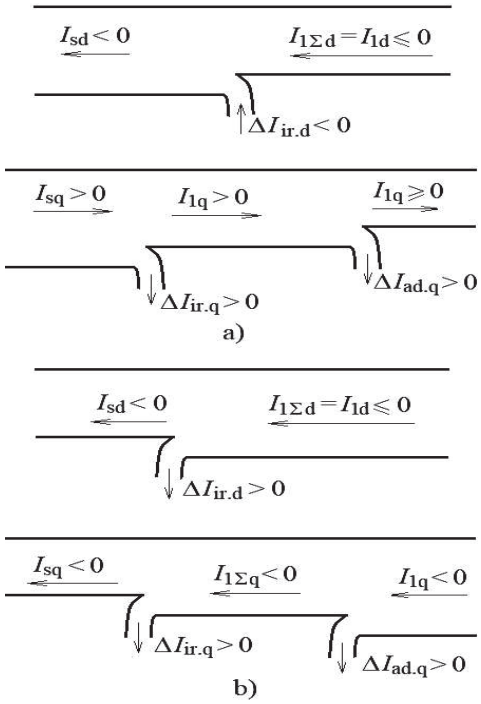


Fig. 2. Diagram of distribution of increments $\Delta I_{ir.d}$, $\Delta I_{ir.q}$ and $\Delta I_{ad.q}$ of stator current between projections I_{sd}, $I_{1\Sigma d}$, I_{1d} and I_{sq}, $I_{1\Sigma q}$, I_{1q} of generalized stator current vectors $\bar{I}_s$, $\bar{I}_{1\Sigma}$, $\bar{I}_1$ in f-r SPMSM for motor (a) and generator (b) modes of its operation

gap respectively) electromagnetic power losses ΔP_{em} and refined calculated electromagnetic power losses ΔP^*_{em}. The mentioned electromagnetic losses ΔP_{em} and the included to them electrical power losses ΔP_e were determined inaccurately from the ratios:

$$\Delta P_{em} = \Delta P_e + \Delta P_{ir} + \Delta P_{ad} \quad \text{and} \quad \Delta P_e = R_s \cdot I_1^2 . \quad (16)$$

According to the calculated ratio $\Delta P_{em}/\Delta P^*_{em}$, given in Table II for the first speed zone, the difference between inaccurate ΔP_{em} and refined ΔP^*_{em} values of the main electromagnetic power losses f-r SPMSM varies from 2.3% (at nominal speed equal to 2800rpm) to 1.5% (at speed 300rpm). And between the above mentioned electromagnetic power losses the following ratios are always fulfilled: $\Delta P^*_{em} > \Delta P_{em}$ – for propulsion or vice versa $\Delta P^*_{em} < \Delta P_{em}$ – for generator mode of SPMSM operation. According to Table II and Table III, the calculated relationships I_s/I_1 between the modules I_s and I_1 of the generalized vectors of the stator current $\bar{I}_s$ and $\bar{I}_1$ SPMSM take values from 0.9696 to 1.0317 (always values greater, than one, correspond to the motor mode of operation and values less, than one, correspond to the generator mode of operation of this motor).

From formulus (7) and (15) the analytical dependences were obtained for SPMSM, which are tied mathematically increments $\Delta I_{ir.d}$, $\Delta I_{ir.q}$ and $\Delta I_{ad.q}$ with the orthogonal projections I_{1d} and I_{1q}, $I_{1\Sigma d}$ and $I_{1\Sigma q}$, I_{sd} and I_{sq} of the generalized vectors $\bar{I}_1$, $\bar{I}_{1\Sigma}$ and $\bar{I}_s$ stator current, respectively. Taking into account the given calculation data in Table II and Table III, let us draw up in Fig. 2 a diagram of the distribution of above mentioned increments of currents in the f-r SPMSM for the motor and generator modes of its operation.

From the analysis of Tables II and III it follows that at the values of useful torque T in the range of speeds equal and higher than the nominal speed: $|n| \geq n_n$, – it is required to provide for normal functioning SPMSM a stator supply voltage reserve, which is equaled to (16 – 26)% of the nominal stator voltage of the motor.

V. Conclusions

New analytical dependences (5) – (7), (14) and (15) for determining the increments of longitudinal and transverse projections of the f-r SPMSM stator current (caused by transportation of magnetic and additional power losses to the magnetizing circuit or through the air gap of the motor, respectively) have been obtained. On the basis of these dependencies, in steady modes of operation of this motor we have performed refined calculations of its: stator current, main electromagnetic power losses, main consumed active power, efficiency and power factor for the first and second speed zones, – at the different values of speed and load. Peculiarities of steady modes of the researched SPMSM are revealed and the diagram of distribution of the mentioned increments of the projections of the motor stator current is constructed.

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Fractal PI-PI$^{\mu}$D with Artificial Neural Network Speed Controller for BLDC Motor Drive

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Abstract*—The research is devoted to synthesize a controller with a fractional-order integral component to achieve the order of astaticism of the closed-loop speed BLDCM higher than 1 and with a neural network to improve the static and dynamic control accuracy. The control object is identified by a fractional-order differential equation. Four types of controllers are synthesized: a classic PID to provide settings close to the modular optimum (IntPID); a controller with a fractional-integral component that provides the order of astaticism of the loop 1.8 (FrPID), and – what is the scientific novelty of the work – hybrid controllers based on the first two, supplemented by neural networks of direct signal propagation NN–IntPID, NN–FrPID. FrPID compared to IntPID allowed to reduce the mean square error by almost 5 times, and the neural controller with a fractional integral component provided the best results: overshoot with a jump in the reference signal was reduced from 2.98% to 0.35%, the mean square error with a linearly changing reference signal was reduced by 11 times and the relative speed error decreases from 0.00801 to 0.00022, more than 35 times. The proposed algorithm for fast learning in each cycle of the control program also ensures the adaptation of the controller when the parameters of the control object change, what makes the possibility of using neural controllers with a fractional integral component for other objects, such as universal collector motors, electromagnetic brakes with adjustable torque.

Keywords—*BLDCM, PID-regulator, hybrid Neural Network – PI$^{\mu}$D-controller, astaticism of the closed-loop system, fast learning algorithm.*

I. Introduction

Brushless DC motors (BLDCM) have become increasingly widespread in recent decades due to the fact that the production of these motors saves resources – there are no windings in the rotor, the stator uses 30-40% less copper than in asynchronous motors, the manufacturing technology is simpler, the cost of motors is 4 times less. However, nonlinear dependencies between the current and the magnetic field strength due to the saturation of the magnetic system complicate the control of BLDCM. To synthesize a speed controller, instead of a full nonlinear model of such a motor, it is advisable to use differential equations with a fractional order, since they belong to the class of linear equations, which allows the use of methods for solving classical problems of automatic control theory [1], [2], and the emergence of approximation methods for calculating a fractional integral makes it possible to use controllers with such components in modern microprocessor control systems [3]–[6].

One of the most popular types of controllers – PID – using differential and integral components of fractional order $\left(PI^{\mu}D^{\nu}\right)$ allows to obtain a controller that is more flexible in tuning, while improving the quality indicators of transient and steady-state processes [7]–[10], its transfer function looks like this:

$$H_r(s)=\frac{Y_r(s)}{X(s)}=k_{rp}+k_{ri}\frac{1}{s^{\mu}}+k_{rd}s^{\nu}, \quad (1)$$

If it is possible to accept $\nu=1-\mu$, then the transfer function of the controller can be written in the following form:

$$H_r(s)=\frac{Y_r(s)}{X(s)}=k_{rp}+\frac{1}{s^{\mu}}\left(k_{ri}+k_{rd}s\right), \quad (2)$$

which can significantly simplify the calculation of the output signal of the controller.

And the main objective of this work is the synthesis of neural controllers as a combination *PID*, $PI^{\mu}D^{\nu}$ and artificial neural networks to achieve the following goal.

The goal of the work is the synthesis of a controller that includes an integral component of fractional order to obtain the order of astaticism of the closed speed BLDCM loop higher than 1, and an artificial neural networks of direct signal propagation to improve the static and dynamic control accuracy.

II. Identification of the Control Object

The transient processes of the BLDCM start-up when applying voltage surges of different amplitudes, calculated using the detailed model in Matlab speed_control_of_bldc_motor.slx, are characterized by a fast initial section and a delayed end of the transient process,

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which is typical for solutions of differential equations of fractional order.

The BLDCM model, which provides sufficient accuracy, can be described in proportional units by the following transfer function:

$$H_{BLDC}(s)=\frac{\omega(s)/\omega_0}{U_c^*(s)}=\frac{1}{T_tT_as^{1+\mu}+T_ts^{\mu}+1}. \quad (3)$$

Then, taking into account the uncompensated time constant of the control system T_v and the gain of the power converter k_c, the control object is described as follows:

$$H_C(s)=\frac{1}{T_tT_as^{1+\mu}+T_ts^{\mu}+1}\frac{k_c}{T_vs+1}. \quad (4)$$

The parameters of such a BLDCM model were obtained during the start-up process with a jump of U_{ref} from $0.2U_n$ to U_n:

$$\begin{aligned}\mu&=0.851-0.045\omega_0^*,\\ T_a&=0.0031-0.0021\omega_0^*,\\ T_t&=0.0056+0.0091\omega_0^*.\end{aligned} \quad (5)$$

Table I shows the accuracy indicators of the obtained model.

TABLE I. ROOT MEAN SQURE ERROR OF BLDCM START PROCESS IDENTIFICATION WHEN CHANGING SPEED REFERENCE

ω_0^*	0.2	0.3	0.4	0.5	0.6
RMSE	0.019	0.020	0.018	0.017	0.015
ω_0^*	0.7	0.8	0.9	1.0	
RMSE	0.013	0.012	0.011	0.013	

These results allow us to consider the obtained model as a base for the synthesis of speed controllers with different internal architectures.

III. SYNTHESIS OF LINEAR REGULATORS

We will consider several variants of regulators, comparing static and dynamic indicators when processing tachograms, including jumps and linearly changing reference signals.

A number of assumptions were made during the modeling: the supply voltage limit is set significantly higher than the nominal one so that the regulators operate mainly in the active mode, and not in the saturation mode. The engine speed is calculated in proportional units, the step $\Delta t = 6.564\text{e-}05\text{s}$ is determined automatically and is assumed to be the same in all the graphs below, the relative time $t/\Delta t$ is displayed on the abscissa axes.

The solution of fractional differential equations describing the object is based on the transfer function (3), which allows only one calculation of the fractional integral of the order μ using the Oustaloup expansion [4]:

$$I^{\mu}(s)=\prod_{i=-N}^{N}\sqrt{\frac{\omega_{k\,i}{}^2+1}{\omega_{kp\,i}{}^2+1}}\frac{s+\omega_{kp\,i}}{s+\omega_{k\,i}}, \quad (6)$$

where $\omega_L=0.1^{3.75}s^{-1}$, $\omega_H=10^{3.75}s^{-1}$, $\lambda=\frac{\omega_H}{\omega_L}$, $N=8$, $\omega_{kp\,i}=\lambda^{\omega_L(i+N-0.5\mu)/(2N+1)}$, $\omega_{k\,i}=\lambda^{\omega_L(i+N+0.5\mu)/(2N+1)}$.

1. Considering $\mu\approx 1$, we synthesize a controller for tuning the speed loop to the modular optimum. For this, the tuned open loop, including the controller and the control object, must be described by the following transfer function:

$$H_{opt}(s)=\frac{1}{aT_vs}\frac{1}{T_vs+1},\ a=2. \quad (7)$$

Then we get a PID controller with parameters:

$$k_{rp}=\frac{T_t}{ak_cT_v},\ k_{ri}=\frac{1}{ak_cT_v},\ k_{rd}=\frac{T_aT_t}{ak_cT_v}. \quad (8)$$

Let us denote such a controller as IntPID. Considering that it includes an integrating link, when covered by single negative feedback we obtain an astatic system of the 1st order.

2. For an object that includes a unit of fractional-order μ, it is possible to synthesize a controller that ensures the order of astaticism $1+\mu$ with a transfer function of the tuned open circuit of the following form:

$$H_{opt}^{1+\mu}(s)=\frac{bT_vs+1}{abT_v^{\mu+1}s^{\mu+1}}\frac{1}{T_vs+1},\ 1<\mu+1<2,\ b>1. \quad (9)$$

The synthesis procedure consists of selecting parameters a,b depending on the order of astaticism [7]:

$$\begin{aligned}&(ab)\approx e^{-10.27+7.831(\mu+1)}\ \Rightarrow\\ &b\approx 7.336+0.792(ab)+3.83\ln(ab)\ \Rightarrow\\ &a=\frac{(ab)}{b}.\end{aligned} \quad (10)$$

Then, taking into account (4), we obtain $PI-PI^{\mu}D-regulator$:

$$\begin{aligned}&H_{PI-PI^{\mu}D}(s)=H_{opt}^{1+\mu}(s)/H_C(s)=\\ &=\left(1+\frac{1}{bT_vs}\right)\left(\frac{T_aT_ts}{aT_v^{\mu}k_c}+\frac{T_t}{k_caT_v^{\mu}}+\frac{1}{k_caT_v^{\mu}s^{\mu}}\right).\end{aligned} \quad (11)$$

Let us designate such a controller as FrPID.

A remarkable feature of the synthesized controller is that it gives the system an astatic order greater than 1 with stability reserves no less than when tuning to the modular optimum. This ensures lower oscillations and the absence of not only static but also speed errors.

As a result of modeling the system, graphs of transient processes were obtained for a jump (Fig. 1, a) and a linearly changing reference signal (Fig. 1, b). Visually, the graphs are

almost identical for both controllers. However, the overshoot for a jump in the reference signal with IntPID is 2.98%, and with FrPID – 0.61%, and for a linearly changing reference signal, it is clearly seen that the error graphs (Fig. 1, c) differ quite significantly, the root-mean-square error (RMSE) at the beginning of the transient process from point 1 to 10,000 and during the entire process from point 1 to 50,000 is almost 5 times less.

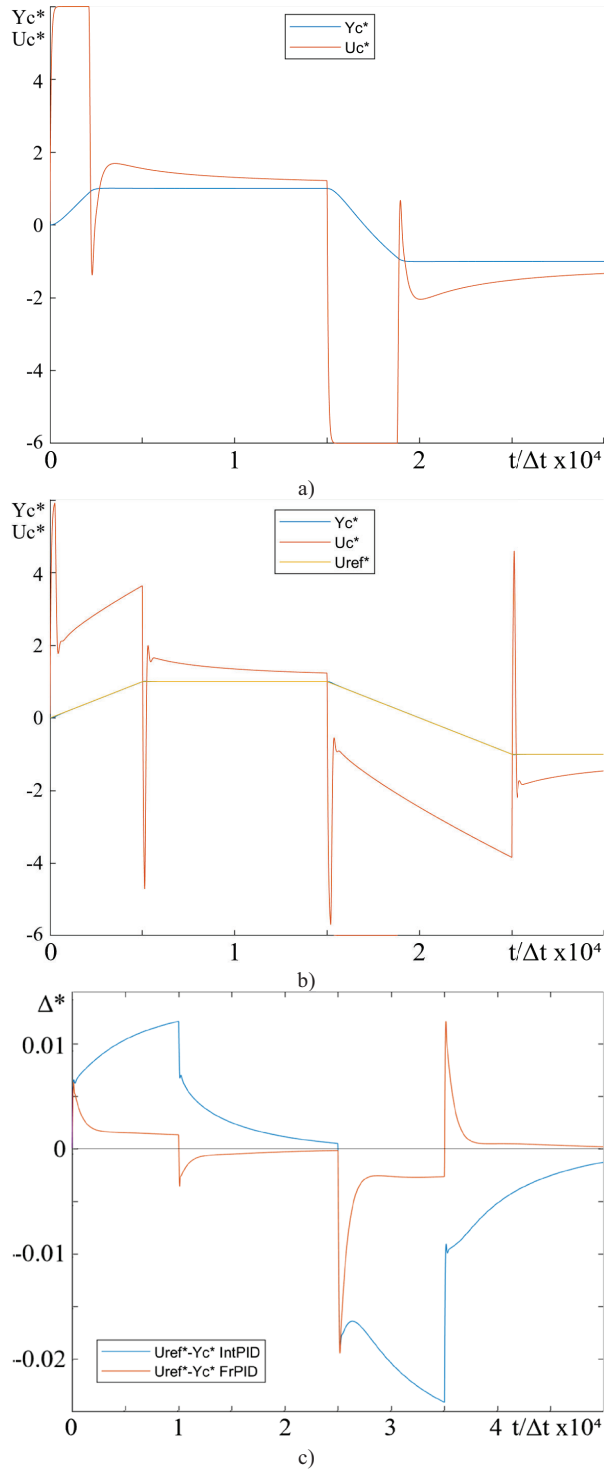


Fig. 1. Graphs of transient processes of the reference signal Uref*, converter voltage Uc*, speed Yc* with a jump (a) and linear change (b) of Uref and error (c) for case (b) when using linear regulators

III. Synthesis of Neuroregulators

Synthesized linear regulators allowed us to obtain fairly good results. However, modern technologies allow us to optimize the system performance by using adaptive regulators, among which neuroregulators [11]–[14] with a structure identical to the synthesized linear IntPID, FrPID seem to us to be the most promising. In such regulators, instead of amplifiers with fixed coefficients, fully connected neural networks of direct propagation with a small number of neurons are used (Fig. 2).

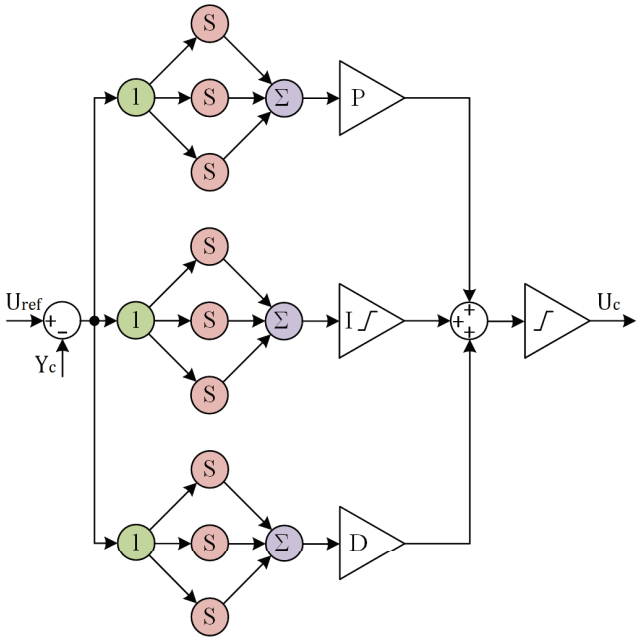


Fig. 2. Functional diagram of the neuro-PID controller

Let us designate such controllers as NN–IntPID, NN–FrPID, respectively.

Neural networks in such controllers should contain one neuron in the input and output layers and 5…10 neurons in the hidden layer. If we use the hyperbolic tangent as activation functions $S(x)$, then they actually form a certain centrally symmetric nonlinear dependence between the error value and the output values of individual components. The output signal of such a neural network in a programming language containing matrix processing tools is calculated using a simple expression [14]:

$$y = S\left(S\left(x * w_2\right) * w_3\right), \tag{12}$$

where w_2; w_3 – matrix of dimension $[1,n]$; $[n,1]$ respectively, the operator "*" corresponds to the vector product of matrices.

Traditionally, for training ANNs, the calculation of the RMSE on a certain interval is used, followed by correction of the weighting coefficients by the method of back propagation of the error using the Levenberg-Marquardt or gradient descent algorithms [15]–[17]. However, for the class of objects under consideration, a fast-learning algorithm can be applied in each cycle of the transient process calculation program.

For this purpose, using the model of the control object, in each cycle of calculating the controller output signal and the

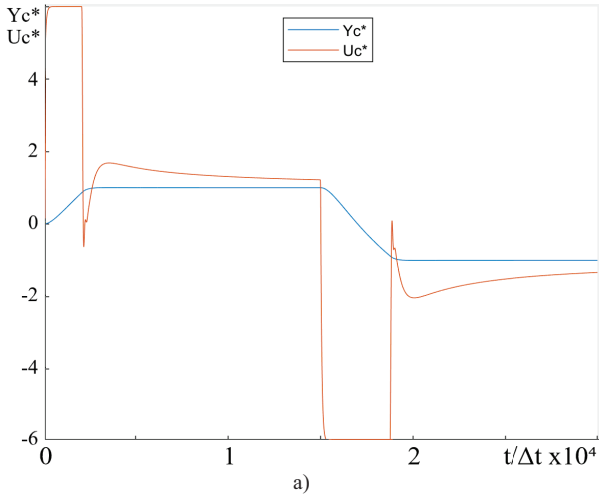
model response, the error $e = Uref - Yc$ is calculated. Since the goal of control is to achieve an error equal to zero, it is its value that is used to correct the weighting coefficients using the Error Backpropagation method (EBP) in combination with the Scaled Conjugate Gradient (SCG) method. In turn, the SCG method is based on calculating the derivative of the activation function, which for the hyperbolic tangent is calculated using the expression $S'(x) \sim S(x)(1 - S(x))$ [13]. Then the calculation of the correction value of the weighting coefficients is carried out using the following procedure:

$$
\begin{aligned}
&\% \ EBP \ with \ Scaled \ Conjugate \ Gradient \\
&e = Uref - Yc; \\
&s2 = S(e * w2); \\
&sx2 = s2 .* (1 - s2); \\
&dE2 = ((e * w2) .* sx2) .* e; \\
&s3 = S(s2 * w3); \\
&sx3 = s3 .* (1 - s3); \\
&dE3 = (e .* sx3) .* (e * w2)'; \\
&w2 = w2 + alfa .* dE2; \\
&w3 = w3 + alfa .* dE3;
\end{aligned}
\tag{13}
$$

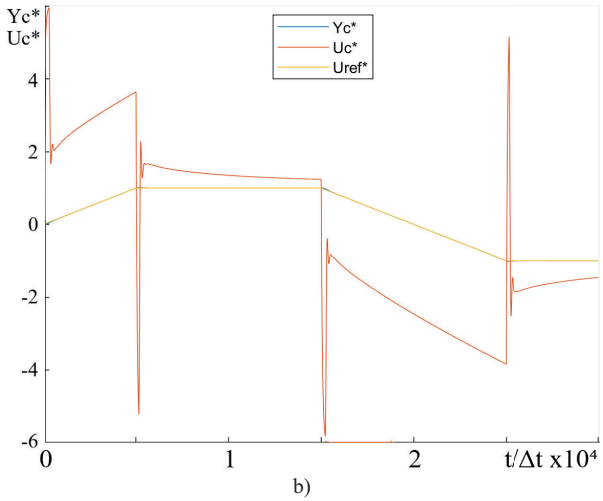
in which the learning coefficient α is set in the range from 0.1 to 1.0, the operator " * " corresponds to the vector product of matrices, " .* " – to element-wise multiplication, " ' " – to matrix transposition.

It is important to note that it is advisable to conduct training with a reference signal changing so that the controller does not enter the saturation mode and in the 1st quadrant, i.e. with a non-negative value of both the reference signal itself and its derivative. We chose a signal linearly increasing from 0 to 1 with a limitation and such a rate of increase that does not lead to saturation of the controller.

As a result of repeating this procedure for 100 epochs, the regulator is adjusted, which is then tested by applying a reference signal in a jump and linearly changing with different acceleration already in the range $-1...+1$ (Fig. 3).



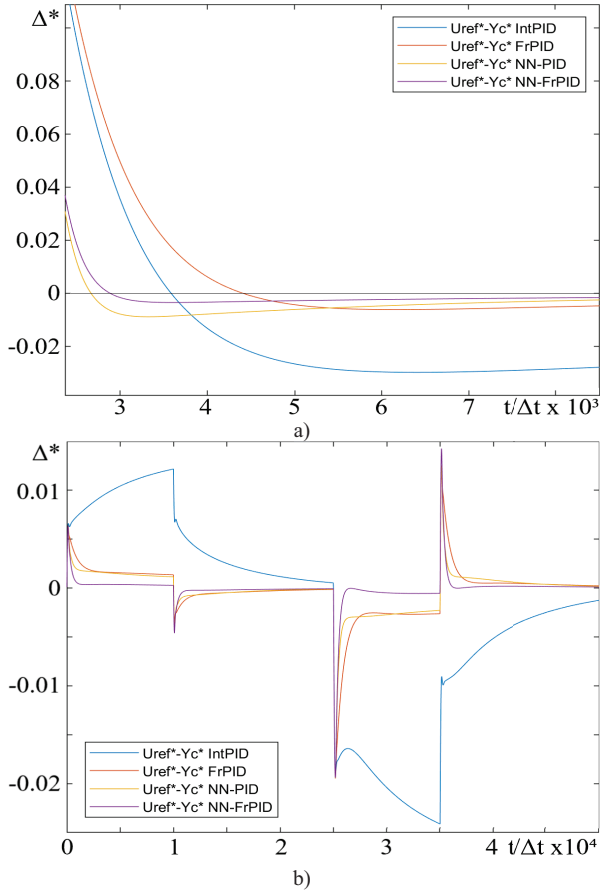
a)



b)

Fig. 3. Response of trained systems with NN–IntPID and NN–FrPID to jumps (a) and linear changes (b) in the reference signal

Visually, the graphs do not differ significantly from Fig. 1, a, b. However, the overshoot during a jump in the reference signal with NN-IntPID is 0.89%, and with NN-FrPID – 0.35%, reaching the extremum with neuroregulators occurs almost twice as early as with linear regulators. The relative speed error c=Δ*/a* decreases from 0.00801 with IntPID and 0.00091 with FrPID to 0.00077 with NN-IntPID and by another 3.5 times to 0.00022 with NN-FrPID. Fig. 4 shows the graphs of the change in the error in transient processes when using different regulators, and Table II compares various indicators depending on the regulator type.



a)

b)

Fig. 4. Graphs of transient processes of errors with a jump in the reference (a) and a linearly changing reference (b) when using different regulators

TABLE II. COMPARATIVE INDICATORS OF INTPID, FRPID AND NN–INTPID, NN–FRPID

Regulator	RMSE		
	RMSE 1:10,000	RMSE 1:50,000	
IntPID	0.0102	0.0107	
FrPID	0.0021	0.0027	
NN–IntPID	0.0016	0.0020	
NN–FrPID	0.0009	0.0016	
	Dynamic indicators		
	Δ_{max}*	t_{max}, s	c=Δ*/a*, s
IntPID	0.0298	0.431	0.00801
FrPID	0.0061	0.430	0.00091
NN–IntPID	0.0089	0.213	0.00077
NN–FrPID	0.0035	0.235	0.00022

IV. CONCLUSION

Thus, based on the identification of the starting processes of a brushless DC motor by a fractional-order differential equation with an error of no more than 2%, it is shown that several different types of controllers can be used for such an object. The use of a classic PID controller allows for RMSE to be provided with a linear reference signal at the level of 0.010…0.012, but with an error increasing to 0.025 and a slowly decreasing static error, which is typical for fractional-aperiodic systems. A reduction of 5 or more times in the root-mean-square error, a rapid reduction in the static error to zero and an error to 0.002 with a linear change in the reference signal is provided by a controller with a fractional-order integral component, which also imparts to the system the property of astaticism for the reference and disturbance of order 1.8. Even higher control accuracy is achieved by combining a fractional-integral controller with fully connected three-level direct-propagation neural networks as preamplifiers of the fractional-integral and proportional components, which allows for both static and velocity errors in most of the transient process to be close to zero.

This result is also achieved due to two more factors: 1. Using a fast learning algorithm – in each cycle of the transient process calculation program – based only on the instantaneous error value and considering the goal of learning to reduce the error to zero; 2. Preliminary selection of the structure and coefficients of the controller, which allows for faster finding of the necessary weighting coefficients of the neurocontroller.

Taking into account the close description of both BLDCM and universal AC/DC motors with series excitation, electromagnetic brake with adjustable torque [7], the proposed approach to the synthesis of neurocontrollers can be used for all types of the listed devices. Verification of this, including experimental, can be the subject of further research.

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Influence of Surface Effect on Nonlinear Normal Vibrations Modes of Nanotube

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***Abstract*—Multi-walled nanotube (MWN) nonlinear free oscillations are simulated by three nonlinear partial differential equations (NPDEs). Nonlinear oscillations of MWN are modeled by using the shell theory. Koiter-Sanders theory is applied to simulate geometrically nonlinear deformations of MWNs. The van der Waals nonlinearity and surface effects are accounted in MWN oscillations model.**

Three NPDEs are transformed to nonlinear ordinary differential equations (NODEs) by applying the Galerkin technique.

As follows from the NODEs analysis, the free nonlinear oscillations of MWN are nonlinear modes. Such nonlinear modes are essentially curvilinear. The properties on nonlinear modes are analyzed numerically.

Keywords— nonlinear oscillation of multi-walled shell, surface effect, van der Waals nonlinearity, nonlinear ordinary differential equation

I. Introduction

Most part of nano research is devoted to the use of beam models to study the nonlinear dynamics of MWNs. In this paper shell models are used to simulate the oscillations of MWN.

Mathematical model of plate is applied to simulate the oscillations of nanofilms. The surface layers, theory of geometrically nonlinear plate by von Karman are applied to describe the oscillations of nanofilms in [1, 2]. The shear theory of the first order and the surface effect are applied to analyze nanoplate forced vibrations in Ref. [3, 4].

Now it is considered the shell models of nanotube oscillations. The linear oscillations of nanotube are modeled using surface effect and shell theory in Ref. [5, 6]. The accounting of the surface effect has essential influence on the values of the nanotube eigenfrequencies. The buckling behavior of nanotube is studied in [7] using the shell nonlinear model. Effect of imperfections on nonlinear buckling of nanoshell is analyzed in Ref. [8]. Gurtin-Murdoch surface theory (GMST) is used in this study. Nanotube free oscillations with account of geometrical nonlinearity and surface effect are discussed in [9, 10] using the shell nonlinear theory. The double-walled nanostructure oscillations are studied accounting effect of surface on four surfaces and shell theory of Novozhilov in Ref. [11]. As follows from [12], the effect of surface has considerable influence on piezoelectric nanotube oscillations. The detailed discussion of surface effect is published in [13]. The influence of effect of surface on static buckling of nanotube is considered in [14, 15]. The survey of solid mechanics modeling of nanotubes is presented in [16].

In the present paper, the system of NPDEs is obtained to simulate the vibrations of nanoshell with account of nonlocal elasticity, effect of surface and geometrical nonlinear behavior. Using the Galerkin method, the obtained system of NPDEs is transformed into system of NODEs. Nonlinear normal oscillations modes, which are analyzed in the configuration space, are discovered by means of numerical simulation. As follows from the results on simulations, nonlinear oscillations modes are essentially curvilinear.

II. Problem Formulation and Main Equations

The nanotube with N walls is described by N thin shells, which are linked by van der Waals forces. These forces are essentially nonlinear. N walled nanotube has the same thickness of all walls h and the i^{th} wall has the radius R_i. Nanotube performs oscillations with moderate amplitudes. In this case, the nanotube geometrical nonlinear behavior is considered.

MWN is analyzed in cylindrical coordinates (x, θ, z). The main unknowns of the problem are i^{th} wall projections of the displacements on the cylindrical coordinates u_i, v_i, w_i. A rotary inertia and a shear are not considered in the shell models. The displacements projections of arbitrary points of i^{th} wall $u_x^{(i)}, u_\theta^{(i)}, u_z^{(i)}$ are

$$\begin{bmatrix} u_x^{(i)} \\ u_\theta^{(i)} \\ u_z^{(i)} \end{bmatrix} = \begin{bmatrix} u_i \\ v_i \\ w_i \end{bmatrix} - z \begin{bmatrix} w_{i,x} \\ \dfrac{w_{i,\theta}}{R_i} - \dfrac{v_i}{R_i} \\ 0 \end{bmatrix}. \tag{1}$$

Three kinds of nonlinearity are considered here. The geometrical nonlinearity is the first kind. The van der Waals loads between nanotube walls are the second kind of

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nonlinearity. The axis stresses resultants are nonlinear functions of displacements, which are used in boundary conditions. This stresses resultants generate the third type of nonlinearity.

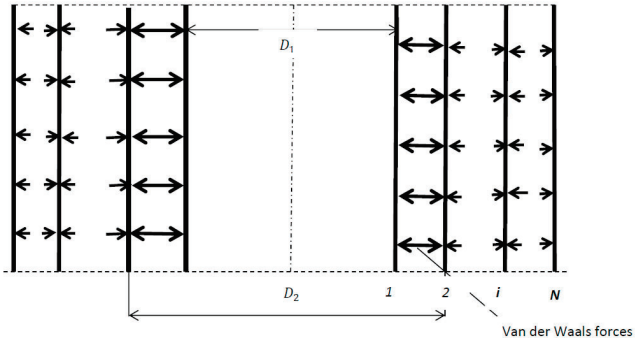


Fig. 1. *N*-walled nanotube

Sanders-Koiter theory, which is modeled the geometrical nonlinear behavior of the nanoshell, is described by the equations [17]:

$$\begin{bmatrix}\varepsilon_{xx}^{(i)}\\ \varepsilon_{\theta\theta}^{(i)}\\ \gamma_{x\theta}^{(i)}\end{bmatrix}=\begin{bmatrix}\varepsilon_{x,0}^{(i)}\\ \varepsilon_{\theta,0}^{(i)}\\ \gamma_{x\theta,0}^{(i)}\end{bmatrix}+z\begin{bmatrix}k_x^{(i)}\\ k_\theta^{(i)}\\ k_{x\theta}^{(i)}\end{bmatrix}=\begin{bmatrix}u_{i,x}+\varepsilon_{i,X,0}^{(NL)}\\ \frac{v_{i,\theta}}{R_i}+\frac{w_i}{R_i}+\varepsilon_{i,\theta,0}^{(NL)}\\ \frac{u_{i,\theta}}{R_i}+v_{i,x}+\gamma_{i,X\theta,0}^{(NL)}\end{bmatrix}$$
$$+z\begin{bmatrix}-w_{i,xx}\\ \frac{v_{i,\theta}}{R_i^2}-\frac{w_{i,\theta\theta}}{R_i^2}\\ -\frac{2w_{i,x\theta}}{R_i}+\frac{3v_{i,x}}{2R_i}-\frac{u_{i,\theta}}{2R_i^2}\end{bmatrix}, \quad (2)$$

where

$$\varepsilon_{i,X,0}^{(NL)}=\frac{1}{2}\left(w_{i,x}\right)^2+\frac{1}{8}\left(v_{i,x}-\frac{u_{i,\theta}}{R_i}\right)^2;$$
$$\varepsilon_{i,\theta,0}^{(NL)}=\frac{1}{2}\left(\frac{w_{i,\theta}}{R_i}-\frac{v_i}{R_i}\right)^2+\frac{1}{8}\left(\frac{u_{i,\theta}}{R_i}-v_{i,x}\right)^2;$$
$$\gamma_{i,X\theta,0}^{(NL)}=w_{i,x}\left(\frac{w_{i,\theta}}{R_i}-\frac{v_i}{R_i}\right);$$

$\varepsilon_{x,0}^{(i)},\varepsilon_{\theta,0}^{(i)},\gamma_{x\theta,0}^{(i)}$ are middle surfaces strains.

The nonlocal Hooke's law describes the elastic properties of the i^{th} wall of nanotube

$$\sigma_{xx}^{(i)}-l^2\nabla^2\sigma_{xx}^{(i)}=(\lambda+2\mu)\varepsilon_{xx}^{(i)}+\lambda\varepsilon_{\theta\theta}^{(i)}; \quad (3)$$
$$\sigma_{\theta\theta}^{(i)}-l^2\nabla^2\sigma_{\theta\theta}^{(i)}=(\lambda+2\mu)\varepsilon_{\theta\theta}^{(i)}+\lambda\varepsilon_{xx}^{(i)};$$
$$\sigma_{x\theta}^{(i)}-l^2\nabla^2\sigma_{x\theta}^{(i)}=\mu\gamma_{x\theta}^{(i)},$$

where $\nabla^2(\cdot)=\frac{\partial^2(\cdot)}{\partial x^2}+\frac{\partial^2(\cdot)}{R_i^2\partial\theta^2}$; $\sigma_{xx}^{(i)},\sigma_{\theta\theta}^{(i)},\sigma_{x\theta}^{(i)}$ are stress tensor elements; μ,λ are parameters of Lame; $l=e_0a$ is the scale coefficient.

According to GMST [18], the constitutive equations for infinitesimal surface layers is the following

$$\sigma_{\alpha\beta}^{(i,s)}=\tau_i^{(s)}\delta_{\alpha\beta}+\left(\tau_i^{(s)}+\lambda^{(s)}\right)\varepsilon_{\gamma\gamma}^{(i,s)}\delta_{\alpha\beta}+$$
$$2\left(\mu^{(s)}-\tau_i^{(s)}\right)\varepsilon_{\alpha\beta}^{(i,s)}+\tau_i^{(s)}u_{\alpha,\beta}^{(i,s)}, \quad (4)$$
$$\sigma_{\alpha z}^{(i,s)}=\tau_i^{(s)}u_{z,\alpha}^{(i,s)}, \quad (5)$$

where $\delta_{\alpha\beta}$ is Kronecker delta; $\lambda^{(s)},\mu^{(s)}$ are surface parameters of Lame; $\tau_i^{(s)}$ are values of residual stresses.

The surface stresses satisfy the following equations:

$$\sigma_{xx}^{(i,s)}=\tau_i^{(s)}+\left(\lambda^{(s)}+2\mu^{(s)}\right)\varepsilon_{xx}^{(i,s)}+\left(\tau_i^{(s)}+\lambda^{(s)}\right)\varepsilon_{\theta\theta}^{(i,s)}$$
$$-\tau_i^{(s)}\varepsilon_{i,X,0}^{(NL)}; \quad (6)$$
$$\sigma_{\theta\theta}^{(i,s)}=\tau_i^{(s)}+\left(\tau_i^{(s)}+\lambda^{(s)}\right)\varepsilon_{xx}^{(i,s)}+\left(\lambda^{(s)}+2\mu^{(s)}\right)\varepsilon_{\theta\theta}^{(i,s)}$$
$$-\tau_i^{(s)}\left(\frac{w_i}{R_i}+\varepsilon_{i,\theta,0}^{(NL)}\right);$$
$$\sigma_{x\theta}^{(i,s)}=\mu^{(s)}\gamma_{x\theta}^{(i,s)}-\tau_i^{(s)}\left[\left[\frac{\partial v_i}{\partial x}+\gamma_{i,X\theta,0}^{(NL)}\right.\right.$$
$$\left.\left.+z\left\{-\frac{w_{i,x\theta}}{R_i}+\frac{3v_{i,x}}{2R_i}-\frac{u_{i,\theta}}{2R_i^2}\right\}\right]\right];$$
$$\sigma_{\theta x}^{(i,s)}=\mu^{(s)}\gamma_{x\theta}^{(i,s)}-\tau_i^{(s)}\left[\left[\frac{u_{i,\theta}}{R_i}+\gamma_{i,X\theta,0}^{(NL)}\right.\right.$$
$$\left.\left.+z\left\{-\frac{w_{i,x\theta}}{R_i}+\frac{v_{i,x}}{2R_i}-\frac{u_{i,\theta}}{2R_i^2}\right\}\right]\right].$$

Following the paper [19], the stresses σ_{zz} satisfy the following equation:

$$\sigma_{zz}^{(i)}=\frac{2z}{h}\left[\tau_i^{(s)}\nabla^2w_i-\rho^{(s)}\ddot{w}_i\right], \quad (7)$$

where $\ddot{w}_i=\frac{\partial^2w_i}{\partial t^2}$; $\rho^{(s)}$ is surface density.

The i^{th} wall nanotube potential energy $\Pi^{(i)}$ is the following:

$$\Pi^{(i)}=\frac{1}{2}\int_{S_i}\int_{-h/2}^{h/2}\left(\sigma_{xx}^{(i)}\varepsilon_{xx}^{(i)}+\sigma_{\theta\theta}^{(i)}\varepsilon_{\theta\theta}^{(i)}+\sigma_{x\theta}^{(i)}\gamma_{x\theta}^{(i)}\right)dzdxR_id\theta+$$
$$\frac{1}{2}\int_{S_i^+}\left(\sigma_{xx}^{(i,+)}\varepsilon_{xx}^{(i,+)}+\sigma_{\theta\theta}^{(i,+)}\varepsilon_{\theta\theta}^{(i,+)}+\sigma_{xz}^{(i,+)}\gamma_{xz}^{(i,+)}+\sigma_{yz}^{(i,+)}\gamma_{yz}^{(i,+)}\right.$$
$$\left.+0.5\sigma_{x\theta}^{(i,+)}\gamma_{x\theta}^{(i,+)}+0.5\sigma_{\theta x}^{(i,+)}\gamma_{x\theta}^{(i,+)}\right)dxR_id\theta$$
$$\frac{1}{2}\int_{S_i^-}\left(\sigma_{xx}^{(i,-)}\varepsilon_{xx}^{(i,-)}+\sigma_{\theta\theta}^{(i,-)}\varepsilon_{\theta\theta}^{(i,-)}+\sigma_{xz}^{(i,-)}\gamma_{xz}^{(i,-)}+\sigma_{yz}^{(i,-)}\gamma_{yz}^{(i,-)}\right.$$
$$\left.+0.5\sigma_{x\theta}^{(i,-)}\gamma_{x\theta}^{(i,-)}+0.5\sigma_{\theta x}^{(i,-)}\gamma_{x\theta}^{(i,-)}\right)dxR_id\theta, \quad (8)$$

where $\sigma_{xx}^{(i,\mp)},\sigma_{\theta\theta}^{(i,\mp)},\sigma_{xz}^{(i,\mp)},\sigma_{yz}^{(i,\mp)},\sigma_{x\theta}^{(i,\mp)},\sigma_{\theta x}^{(i,\mp)}$ are components of stresses for inside and outside surfaces; S_i is domain of middle surface of i^{th} layer.

The i^{th} layer potential energy has the following form:

$$\Pi^{(i)}=\frac{1}{2}\int_{S_i}\left(\bar{N}_{xx}^{(i)}\varepsilon_{x,0}^{(i)}+\bar{M}_{xx}^{(i)}k_x^{(i)}+\bar{N}_{\theta\theta}^{(i)}\varepsilon_{\theta,0}^{(i)}+\bar{M}_{\theta\theta}^{(i)}k_\theta^{(i)}+\right.$$
$$\left.\bar{N}_{x\theta}^{(i)}\gamma_{x\theta,0}^{(i)}+\bar{M}_{x\theta}^{(i)}k_{x\theta}^{(i)}+Q_x^{(i,s)}w_{i,x}+Q_y^{(i,s)}\frac{w_{i,\theta}}{R_i}\right)dxR_id\theta, \quad (9)$$

where

$$\bar{N}_{xx}^{(i)}=l^2\nabla^2N_{xx}^{(i)}+A_{11}^{(i)}\varepsilon_{x,0}^{(i)}+A_{12}^{(i)}\varepsilon_{\theta,0}^{(i)}+2\tau_i^{(s)}-2\tau_i^{(s)}\varepsilon_{i,X,0}^{(NL)};$$
$$\bar{N}_{\theta\theta}^{(i)}=l^2\nabla^2N_{\theta\theta}^{(i)}+A_{11}^{(i)}\varepsilon_{\theta,0}^{(i)}+A_{12}^{(i)}\varepsilon_{x,0}^{(i)}+2\tau_i^{(s)}$$
$$-2\tau_i^{(s)}\left(\frac{w_i}{R_i}+\varepsilon_{i,\theta,0}^{(NL)}\right); \quad (10)$$
$$\bar{N}_{x\theta}^{(i)}=l^2\nabla^2N_{x\theta}^{(i)}+A_{55}^{(i)}\gamma_{x\theta,0}^{(i)}-\tau_i^{(s)}\gamma_{i,X\theta,0}^{(NL)};$$
$$\bar{M}_{xx}^{(i)}=l^2\nabla^2M_{xx}^{(i)}+D_{11}^{(i)}k_x^{(i)}+D_{12}^{(i)}k_\theta^{(i)}+E_{11}^{(i)}\nabla^2w_i-G_{11}\ddot{w}_i;$$
$$\bar{M}_{\theta\theta}^{(i)}=l^2\nabla^2M_{\theta\theta}^{(i)}+D_{11}^{(i)}k_\theta^{(i)}+D_{12}^{(i)}k_x^{(i)}+E_{11}^{(i)}\nabla^2w_i-G_{11}\ddot{w}_i;$$

$$\bar{M}_{x\theta}^{(i)} = l^2\nabla^2 M_{x\theta}^{(i)} + D_{55}^{(i)} k_{x\theta}^{(i)} - E_{22}^{(i)}\left[-\frac{w_{i,x\theta}}{R_i} + \frac{v_{i,x}}{R_i} - \frac{u_{i,\theta}}{2R_i^2}\right];$$

the parameters $A_{11}^{(i)}, A_{12}^{(i)}, A_{55}^{(i)}, E_{11}^{(i)}, G_{11}, D_{11}^{(i)}, D_{12}^{(i)}$ are published in [19].

The i^{th} layer nanotube kinetic energy is the following:

$$T^{(i)} = \frac{I_0}{2}\int_{S_i}(\dot{u}_i^2 + \dot{v}_i^2 + \dot{w}_i^2)dx R_i d\theta, \qquad (11)$$

where ρ and $\rho^{(s)}$ are densities of nanotube and its surfaces; $I_0 = \rho h + 2\rho^{(s)}$.

The considered simply-supported nanotube satisfies the following boundary conditions:

$$v_i|_{x=0} = w_i|_{x=0} = v_i|_{x=L} = w_i|_{x=L} = 0. \qquad (12)$$
$$\bar{N}_{xx}^{(i)}\big|_{x=0} = \bar{M}_{xx}^{(i)}\big|_{x=0} = \bar{N}_{xx}^{(i)}\big|_{x=L} = \bar{M}_{xx}^{(i)}\big|_{x=L} = 0.$$

Three NPDEs of the nanotube oscillations are obtained from the Hamilton principle.

$$\mathbb{\Gamma}_w^{(i)} = I_0\ddot{w}_i + \frac{\bar{N}_{\theta\theta}^{(i)}}{R_i} - \bar{M}_{xx,xx}^{(i)} - \frac{\bar{M}_{\theta\theta,\theta\theta}^{(i)}}{R_i^2} - \frac{2\bar{M}_{x\theta,x\theta}^{(i)}}{R_i} - Q_{x,x}^{(i,s)} - \frac{Q_{y,\theta}^{(i,s)}}{R_i} - \frac{\partial}{\partial x}\left(\bar{N}_{xx}^{(i)} w_{i,x}\right) - \frac{\partial}{\partial\theta}\left[\frac{\bar{N}_{\theta\theta}^{(i)}}{R_i}\left(\frac{w_{i,\theta}}{R_i} - \frac{v_i}{R_i}\right)\right] - \frac{\partial}{\partial x}\left[\bar{N}_{x\theta}^{(i)}\left(\frac{w_{i,\theta}}{R_i} - \frac{v_i}{R_i}\right)\right] - \frac{\partial}{\partial\theta}\left(\frac{\bar{N}_{x\theta}^{(i)}}{R_i} w_{i,x}\right) - p_z^{(i)} = 0;$$

$$\mathbb{\Gamma}_v^{(i)} = I_0\ddot{v}_i - \frac{\bar{N}_{\theta\theta,\theta}^{(i)}}{R_i} - \frac{\bar{M}_{\theta\theta,\theta}^{(i)}}{R_i^2} - \bar{N}_{x\theta,x}^{(i)} - \frac{3\bar{M}_{x\theta,x}^{(i)}}{2R_i} - \frac{\partial}{\partial x}\left[\frac{\bar{N}_{xx}^{(i)} + \bar{N}_{\theta\theta}^{(i)}}{4}\left(v_{i,x} - \frac{u_{i,\theta}}{R_i}\right)\right] - \frac{\bar{N}_{\theta\theta}^{(i)}}{R_i^2}\left(w_{i,\theta} - v_i\right) - \frac{\bar{N}_{x\theta}}{R_i} w_{i,x} - p_\theta^{(i)} = 0; \qquad (13)$$

$$\mathbb{\Gamma}_u^{(i)} = I_0\ddot{u}_i - \bar{N}_{xx,x}^{(i)} - \frac{\bar{N}_{x\theta,\theta}^{(i)}}{R_i} + \frac{\bar{M}_{x\theta,\theta}^{(i)}}{2R_i^2} + \frac{\partial}{\partial\theta}\left[\frac{\bar{N}_{xx}^{(i)} + \bar{N}_{\theta\theta}^{(i)}}{4R_i}\left(v_{i,x} - \frac{u_{i,\theta}}{R_i}\right)\right] - p_x^{(i)} = 0,$$

where the partial derivatives with respect to the index is indicated by the index after comma; $p_x^{(i)}, p_\theta^{(i)}, p_z^{(i)}$ are external forces projections.

The NPDEs (13) are presented with respect to u_i, v_i, w_i in the form:

$$\mathbb{\Gamma}_u^{(i)} = \Lambda\left(I_0\ddot{u}_i - p_x^{(i)}\right) + F_u^{(i)} + \mathcal{L}_u^{(i)} - R_u^{(i)} = 0; \qquad (14)$$
$$\mathbb{\Gamma}_v^{(i)} = \Lambda\left(I_0\ddot{v}_i - p_y^{(i)}\right) + F_v^{(i)} + \mathcal{L}_v^{(i)} + R_v^{(i)} = 0;$$
$$\mathbb{\Gamma}_w^{(i)} = \Lambda\left(I_0\ddot{w}_i - p_z^{(i)}\right) + F_w^{(i)} + \mathcal{L}_w^{(i)} + R_w^{(i)} = 0; i = 1, \dots, N,$$

where $\Lambda(\cdot) = (\cdot) - l^2\nabla^2(\cdot)$; $\mathcal{L}_u^{(i)}, F_u^{(i)}, R_u^{(i)}, \mathcal{L}_v^{(i)}, F_v^{(i)}, R_v^{(i)}, \mathcal{L}_w^{(i)}, F_w^{(i)}, R_w^{(i)}$ are operators.

The van der Waals nonlinear loads act on the every layer of the nanotube

$$p_z^{(i)} = \sum_{j=1}^{N} c_{ij}\left(w_i - w_j\right) + \sum_{j=1}^{N} e_{ij}\left(w_i - w_j\right)^3;$$
$$p_x^{(i)} = p_\theta^{(i)} = 0, \qquad (15)$$

where

$$c_{ij} = -\frac{\varepsilon R_j\pi}{a^4}\left\{\frac{1001\sigma^{12}}{3}E_{ij}^{(13)} - \frac{1120\sigma^6}{9}E_{ij}^{(7)}\right\};$$
$$e_{ij} = -\frac{\varepsilon R_j\pi}{a^4}\left\{\frac{65065\sigma^{12}}{6}E_{ij}^{(15)} - \frac{3920\sigma^6}{3}E_{ij}^{(9)}\right\};$$
$$E_{ij}^{(m)} = \left(R_i + R_j\right)^{-m}\int_0^{\pi/2}\frac{d\theta}{\left(1 - K_{ij}cos^2\theta\right)^{m/2}};$$
$$K_{ij} = \frac{4R_iR_j}{\left(R_j + R_i\right)^2},$$

a is bond length; ε is potential depth; σ is parameter of equilibrium.

III. Nonlinear ordinary differential equation of nanotube motions

The Galerkin technique is used to transform the NPDEs (14) into the system of NODEs. This transformation is discussed in this section. The nanoshell nonlinear vibrations are expanded as:

$$u_i = \sum_{j=1}^{r_i^{(u)}} P_{i,j}^{(u)}(x)\left[q_{i,j}^{(u,c)}(t)\cos(n\theta) + q_{i,j}^{(u,s)}(t)\sin(n\theta)\right] + \sum_{j=1}^{r_i^{(u,0)}} P_{i,j}^{(u,0)}(x)\, q_{i,j}^{(u,0)};$$

$$v_i = \sum_{j=1}^{r_i^{(v)}} P_{i,j}^{(v)}(x)\left[q_{i,j}^{(v,c)}(t)\cos(n\theta) + q_{i,j}^{(v,s)}(t)\sin(n\theta)\right];$$

$$w_i = \sum_{j=1}^{r_i^{(w)}} P_{i,j}^{(w)}(x)\left[q_{i,j}^{(w,c)}(t)\cos(n\theta) + q_{i,j}^{(w,s)}(t)\sin(n\theta)\right] + \sum_{j=1}^{r_i^{(w,0)}} P_{i,j}^{(w,0)}(x)\, q_{i,j}^{(w,0)}, \qquad (16)$$

where

$$P_{i,j}^{(u)}(x) = \sum_{\nu=1}^{g_{ij}^{(u)}} A_{u,\nu}^{(i,j)} cos\left(\frac{\nu\pi x}{L}\right);$$
$$P_{i,j}^{(u,0)}(x) = \sum_{\nu=1}^{g_{ij}^{(u,0)}} A_{u,\nu}^{(0,i,j)} cos\left(\frac{\nu\pi x}{L}\right);$$
$$P_{i,j}^{(v)}(x) = \sum_{\nu=1}^{g_{ij}^{(v)}} A_{v,\nu}^{(i,j)} sin\left(\frac{\nu\pi x}{L}\right);$$
$$P_{i,j}^{(w)}(x) = \sum_{\nu=1}^{g_{ij}^{(w,0)}} A_{w,\nu}^{(0,i,j)} sin\left(\frac{\nu\pi x}{L}\right);$$

n is waves number; $\mathbf{q} = \left(q_1, \dots, q_{N_*}\right) \equiv \left(q_{i,j}^{(u,c)}, \dots, q_{i,j}^{(w,0)}\right)$ is vector of generalized coordinates; $A_{u,\nu}^{(i,j)}, \dots, A_{w,\nu}^{(0,i,j)}$ are coefficients of eigenmodes, which are obtained by the Rayleigh– Ritz method [20].

The Galerkin technique is expressed by the following relations:

$$\iint_{S_i} \mathbb{I}_u^{(i)} P_{i,j}^{(u,0)}(x)\, R_i d\theta dx + \int_0^{2\pi} \left[\bar{N}_{xx}^{(i)} P_{i,j}^{(u,0)}(x)\right]_0^L R_i d\theta = 0;$$

$$\iint_{S_i} \mathbb{I}_u^{(i)} P_{i,j}^{(u)}(x) \begin{Bmatrix}\cos(n\theta)\\ \sin(n\theta)\end{Bmatrix} R_i d\theta dx + \int_0^{2\pi} \left[\bar{N}_{xx}^{(i)} P_{i,j}^{(u)}(x) \begin{Bmatrix}\cos(n\theta)\\ \sin(n\theta)\end{Bmatrix}\right]_0^L R_i d\theta = 0;$$

$$\iint_{S_i} \begin{Bmatrix}\mathbb{I}_v^{(i)} P_{i,j}^{(v)}(x)\cos(n\theta)\\ \mathbb{I}_v^{(i)} P_{i,j}^{(v)}(x)\sin(n\theta)\end{Bmatrix} R_i d\theta dx = 0; \quad (17)$$

$$\iint_{S_i} \begin{Bmatrix}\mathbb{I}_w^{(i)} P_{i,j}^{(w)}(x)\cos(n\theta)\\ \mathbb{I}_w^{(i)} P_{i,j}^{(w)}(x)\sin(n\theta)\\ \mathbb{I}_w^{(i)} P_{i,j}^{(w,0)}(x)\end{Bmatrix} R_i d\theta dx = 0.$$

The equations (17) can be presented as nonlinear dynamical system. This dynamical system has the following matrix form:

$$\mathbf{M}\ddot{\mathbf{q}} + \mathbf{K}\mathbf{q} + \mathbf{K}^{(\mathbf{B})}\mathbf{q} + \mathcal{P}_{\boldsymbol{F}}^{(\mathbf{2})}(\mathbf{q}) + \mathcal{P}_{\boldsymbol{W}}^{(\mathbf{3})}(\mathbf{q}) + \mathcal{P}_{\boldsymbol{G}}^{(\mathbf{2,3})}(\mathbf{q}) = 0, \quad (18)$$

where $\mathbf{M}$; $\mathbf{K}$ are matrices, which describe the linear part of the system; the term $\mathbf{K}^{(\mathbf{B})}\mathbf{q}$ is linear contribution for nonlinear boundary conditions for $\bar{N}_{xx}^{(i)}$. The nonlinear term $\mathcal{P}_{\boldsymbol{F}}^{(\mathbf{2})}(\mathbf{q})$ is contribution of the nonlinear summands of $\bar{N}_{xx}^{(i)}$ into the NODEs (18).

$$\mathcal{P}_{\boldsymbol{F},i}^{(\mathbf{2})} = \sum_{\nu=1}^{N_*}\sum_{j=1}^{\nu} f_{\nu j}^{(i)} q_\nu q_j .$$

The van der Waals nonlinearity in the equation (18) are described by vector-function $\mathcal{P}_{\boldsymbol{W}}^{(\mathbf{3})}(\mathbf{q})$

$$\mathcal{P}_{\boldsymbol{W},i}^{(\mathbf{3})} = \sum_{\boldsymbol{\nu=1}}^{N_*}\sum_{\boldsymbol{j=1}}^{\boldsymbol{\nu}}\sum_{j_1=1}^{j} w_{\nu j j_1}^{(i)} q_\nu q_j q_{j_1} .$$

The geometrical nonlinearity of nanoshell is described by the vector-function

$$\mathcal{P}_{\boldsymbol{G},i}^{(\mathbf{2,3})} = \sum_{\nu=1}^{N_*}\sum_{j=1}^{\nu} g_{\nu j}^{(i)} q_\nu q_j + \sum_{\boldsymbol{\nu=1}}^{N_*}\sum_{\boldsymbol{j=1}}^{\boldsymbol{\nu}}\sum_{j_1=1}^{j} g_{\nu j j_1}^{(i)} q_\nu q_j q_{j_1} .$$

IV. Nonlinear modes numerical analysis

Here the nonlinear modes are considered in the sense of Rosenberg [21, 22]. In order to calculate numerically these nonlinear modes, the continuation method and shooting approach are applied jointly [23, 24]. The free nonlinear oscillations in the form of nonlinear normal modes are presented in backbone curves. The multipliers are obtained to analyze stability of these periodic motions [25, 26, 27]. The multipliers are eigenvalues of monodromy matrix. This matrix is calculated numerically too.

V. Data of numerical analysis

Nonlinear oscillations of nanoshell with one wall are considered from Si «100» with the surface and bulk parameters:

$$\frac{R_1}{h} = 50;\ E = 210\ GPa;\ \rho = 2331\ kg/m^3;\ \nu = 0.24;$$
$$\lambda^{(s)} = -4.488\ N/m;\ \mu^{(s)} = -2.774\ N/m;$$
$$\tau_1^{(s)} = 0.605\ N/m;\ \rho^{(s)} = 3.17\cdot 10^{-7}\ kg/m^2, \quad (19)$$

where ν is bulk material Poisson ratio; E is Young modulus. The nanoshell geometrical values are:

$$\frac{R_1}{h} = 30; h = 10\ nm;\ L = 4500\ nm.$$

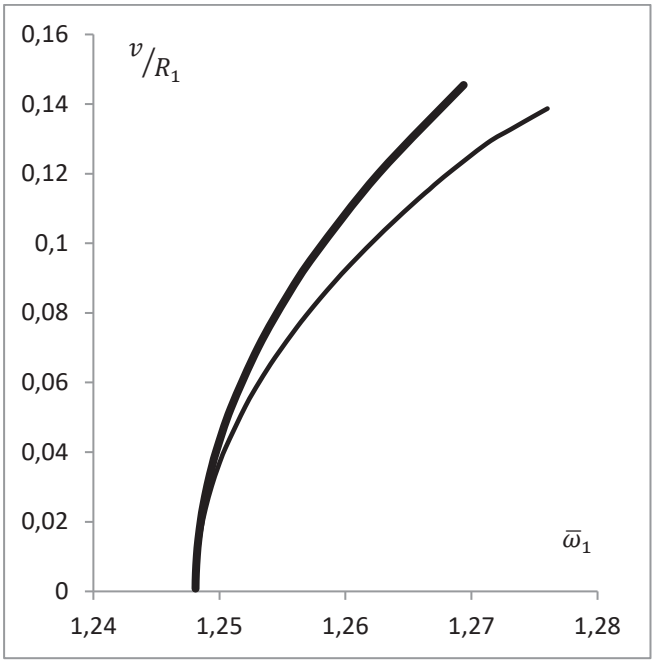


Fig. 2. Nanoshell backbone lines

Two nonlinear systems (18) having eight and fourteen degrees of freedom are obtained. Nonlinear modes of these two systems are calculated by combination of arc length continuation approach and shooting technique. The data of the numerical study are presented on Fig.2, where the dependences of nanotube torsions ν on the frequency $\bar{\omega}_1$ are presented. The thick line shows the nonlinear modes in eight degrees of freedom dynamical system. Thin line shows nonlinear modes in fourteen degrees of freedom dynamical systems. The solutions convergences take place (Fig.2). As follows from the results of numerical simulations, energy exchanges between torsional and radial displacements are observed.

The dependence of nonlinear oscillations on surface effect is considered. The backbone curves taking the surface effect and without one are presented on Fig.3. Thus, the parameters of nonlinear oscillations depend essentially on the surface effect.

Nonlinear oscillations of the nanotube with the parameters (19), the length $L = 7200\ nm$ and wave number $n = 2$ are studied. As follows from the numerical studies, two internal resonances 1:1 are observed in nanotube. The internal resonance is explained by existing of two conjugate modes, which is observes in cylindrical nanoshell. Moreover, two eigenfrequencies is the following: $\bar{\omega}_1^{(n=2)} = 3.1$; $\bar{\omega}_2^{(n=2)} = 3.13$. Thus, these eigenfrequencies meet the condition of the internal resonance 1:1.

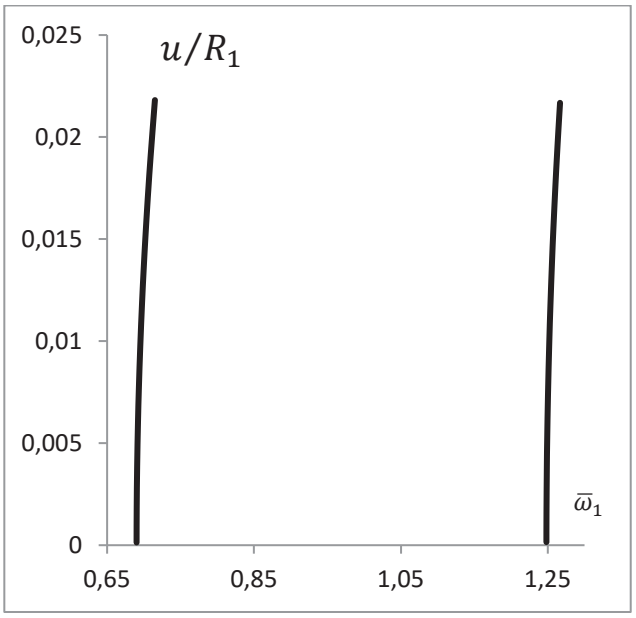


Fig. 3. Nanotube backbone curve with and without surface effect

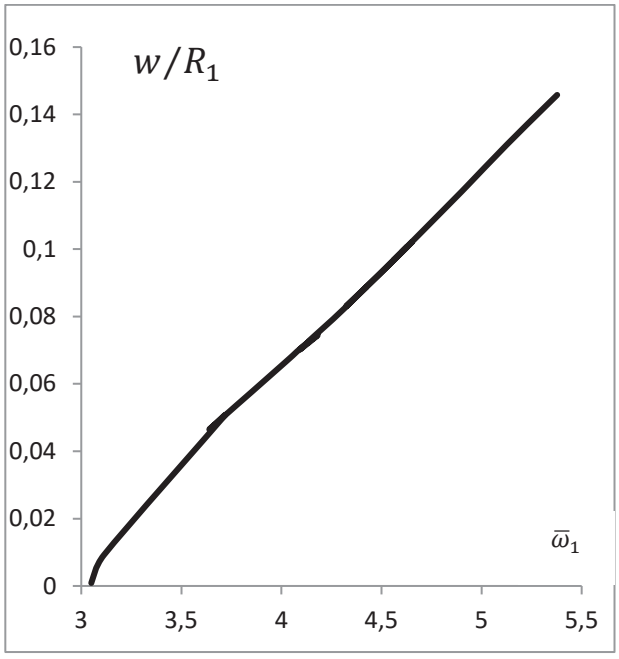
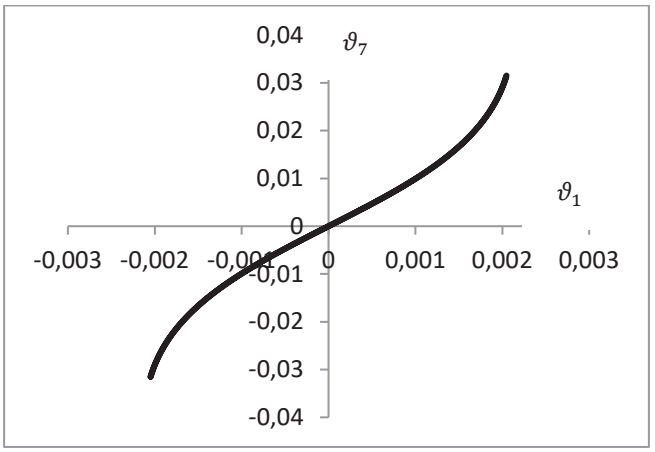


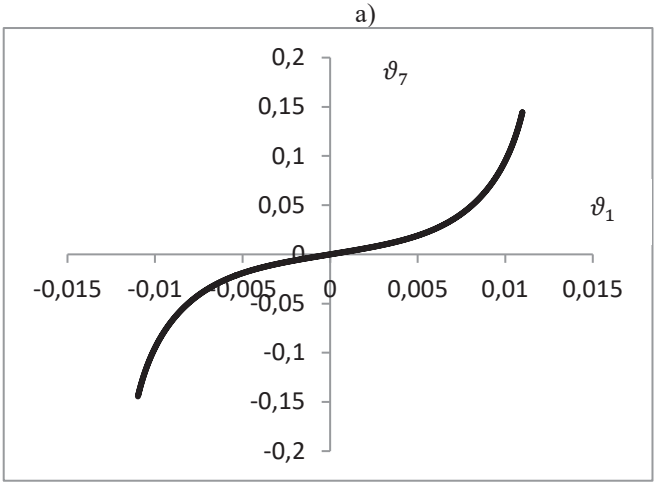
Fig. 4. Backbone curves in the case of two internal resonances

The nanotube travelling waves are considered. Then two conjugate modes (16) are considered. The 14 degrees of freedom nonlinear dynamical system is constructed to study such travelling waves. Fig.4 shows the backbone curve of this dynamical system, which is essentially nonlinear. Thus, the backbone curve is hard. The frequency of nonlinear oscillations belongs to the range $\bar{\omega}_1 \in [3.1; 5.5]$. The amplitudes of radial and torsional motions are closed.

The periodic oscillations, which are shown on Fig.4, are nonlinear modes. These nonlinear modes are essentially curvilinear. The nonlinear modes are analyzed in the configuration subspace $(\vartheta_1, \vartheta_7)$. Two curvilinear nonlinear modes for different values of amplitude of radial displacements a_w are shown on Fig.5.



a)



b)

Fig. 5. Nonlinear modes at radial displacements amplitudes a_w and frequencies $\bar{\omega}_1$: a. $(a_w, \bar{\omega}_1) = (5.13 \cdot 10^{-2}; 3.732)$; b. $(a_w, \bar{\omega}_1) = (1.45 \cdot 10^{-1}; 5.37)$

VI. CONCLUSIONS

Nonlinear mathematical model of nonlinear oscillations of nanotubes is obtained. This model consists of nonlinear ordinary differential equations system. The geometrically nonlinear shell theory with account of nonlocal elasticity, surface effect and nonlinear van der Waalse forces are used to obtain the system of nonlinear partial differential equations. The Galerkin technique is applied to these partial differential equations. As a result, the system of nonlinear ordinary differential equations, which is mathematical model of nanotube, is derived.

As a result of the numerical integration, it is obtained, that the nonlinear oscillations of nanotube are nonlinear normal modes in configuration space. The obtained nonlinear modes are essentially curvilinear.

As follows from the numerical simulations, surface effect affects essentially on nanotube nonlinear oscillations.

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Investigation of the Modifing of the Granulometric Composition of the Solid-Dispersed Phase of Coal Tar by Electric Discharge

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***Abstract* - Coal tar is a key by-product of coking processes widely used for producing binders in carbon electrode manufacturing. One of its essential characteristics is the presence of a solid dispersed phase consisting of high-carbon particles, notably the α-fraction (toluene-insoluble substances) and its high-molecular-weight subset, the α_1-fraction (quinoline-insoluble substances). The performance of coal-tar-based binders and impregnation pitches depends significantly on the granulometric composition of this dispersed phase. In this study, we investigated the feasibility of modifying the particle size distribution of the α_1-fraction by initiating hydraulic shock within coal tar using high-voltage electrical discharges in a sealed vessel. Experimental results demonstrate that even a limited number of discharges leads to a measurable reduction in the mass fraction of α- and α_1-fractions due to mechanical disintegration of coarse and agglomerated particles. Scanning microscopy confirms the shift toward finer particle sizes, enhancing the interaction surface between the solid phase and dispersion medium. This transformation improves the infiltration potential of the pitch into porous carbon blanks, potentially increasing impregnation efficiency without compromising coking properties. However, partial reaggregation over time suggests that dispersion should occur at the thermal fractionation stage, where higher viscosity suppresses particle recombination. The findings open new possibilities for tailoring coal tar microstructure to meet advanced requirements in electrode production.**

Keywords - Coal tar, dielectric, granulometric composition, α-fraction, electric discharge, hydraulic shock.

I. Introduction

Coal tar is one of the main products of coke chemical processing, widely used as a binder for carbon electrodes in the aluminum industry, electrometallurgy, and other modern high-tech sectors. These binders possess a number of unique properties that make them virtually irreplaceable materials in a wide range of applications [1], [2]. One of its key properties is the presence of a solid dispersed phase composed of high-carbon particles — substances insoluble in toluene (known as the α-fraction or TI). Within this fraction, a high-molecular-weight component with the highest degree of pyrolysis and properties closest to carbon materials is further distinguished: the quinoline-insoluble substances (α_1-fraction, QI or so named “carbon blacks”). During the processing (rectification) of coal tar pitch, the aforementioned components almost entirely transfer into the non-distillable residue, which serves as the raw material for the production of binder materials.

Another common indicator of the composition of the solid dispersed phase of coal tar is the mass fraction of the α2-fraction (otherwise - β-rezines). However, this indicator is of purely auxiliary importance, its value is determined as the arithmetic difference between the values of QT and QI, and therefore does not provide additional information from the point of view of the purpose of the studies presented in this article.

The homogeneity, and therefore the binding and plastic properties of the dispersed system, improves with greater interaction between the solid carbonaceous dispersed phase and the dispersion medium. This interaction depends, among other factors (e.g., surface condition, particle shape, wettability), on the specific surface area of the dispersed phase. A higher content of finely dispersed α1-fraction in the pitch increases this surface area and thus enhances its interaction with the dispersion medium. Moreover, this interaction can be strengthened by the “activation” (radicalization) of particle surface fragments through mechanical disruption.

In carbon electrode manufacturing—especially for graphitized electrodes—an additional stage is commonly applied: impregnation of the baked body with a special binder known as impregnation electrode pitch, also obtained as the non-volatile residue from coal tar rectification. In this context, modifying the granulometric composition of the dispersed phase toward smaller fractions becomes particularly important. Various sources estimate the average pore diameter of electrode blanks to fall within the range of 1–45 μm. It has been shown that decreasing the content of particles within this size range in the pitch can increase the degree of impregnation by 2–5% [3].

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Our previous research [4] confirmed that the particle size of a significant (if not dominant) portion of coal tar's solid phase corresponds to the typical pore sizes of electrode blanks. Thus, transforming part of the α1-fraction into the 0–1 μm range can improve the penetration ability of the impregnation pitch without impairing its coking performance.

Our previous studies [4], [5] also demonstrated a high tendency of α_1-fraction particles to form aggregates. Microscopic analysis makes it possible to distinguish between primary aggregates, which are formed in the free-space zone under the roof of the coke oven and are resistant to dissolution of the tar in organic solvents, and secondary aggregates, which are formed during the condensation of coal tar from the vapor phase, as well as during storage, transportation, and other technological stages involving heating under conditions that reduce tar viscosity (see Fig. 1 and 2).The reduction in tar viscosity ensures the mobility of microscopic dispersed particles, which, in combination with specific surface activity characteristics, promotes their aggregation. Under conditions where heating leads to an increase in the viscosity of the liquid phase (e.g., during distillation or rectification), aggregation is expected to be hindered.



Primary (1) and secondary (2) aggregates of α_1-fraction coal tar particles (microscope magnification ×500)

Fig. 1. Reflected light micrographs of examples of primary and secondary aggregates of α_1-fraction particles (microscope magnification ×500).

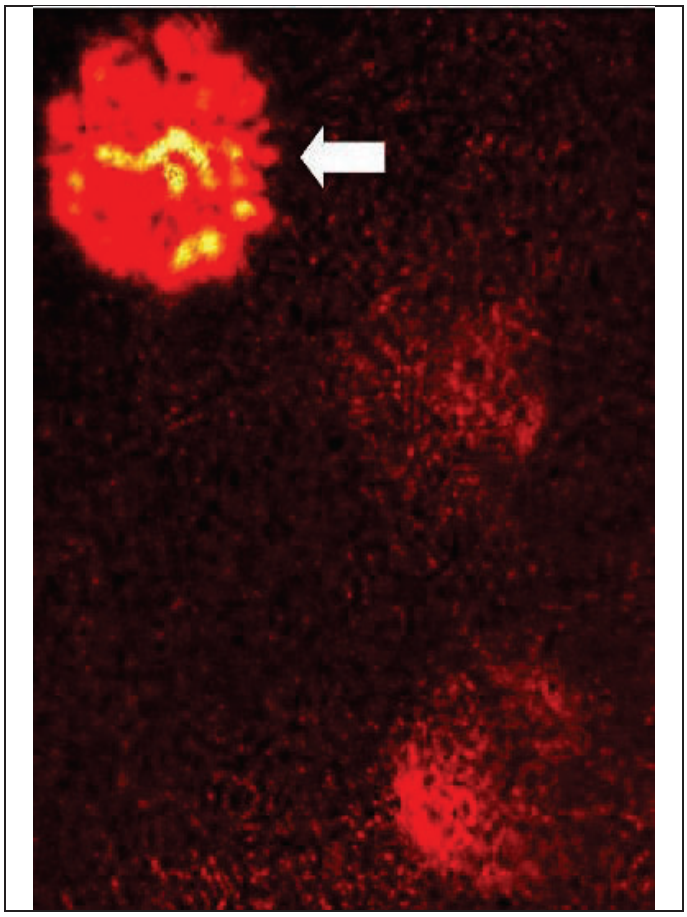


Fig. 2. Aggregation of α_1-fraction particles during viscosity reduction of coal tar by dissolution in quinoline, observed under laser beam illumination at an angle to the sample plane (microscope magnification ×800)

II. Materials and Methods

We evaluated the fundamental feasibility of dispersing α-fraction particles through hydraulic shock initiated in coal tar confined within a sealed vessel. A hydraulic shock is a sudden pressure increase that propagates as an elastic wave in all directions from the initiation center. Upon reaching the vessel wall, part of the pressure is absorbed, and the drop in pressure is reflected back toward the initiation point.

In the hydraulic shock zone, suspended α_1-fraction particles experience an abrupt increase in surface pressure, leading to fragmentation. Aggregates and large irregular particles are the most susceptible to such destruction.

An electric discharge was used as the initiator of the hydraulic shock. Since coal tar is a dielectric, discharge theoretically can occur with dielectric breakdown — loss of dielectric properties under an electric field exceeding a certain strength.

The dielectric properties of coal tar remain insufficiently studied, and the available data reported in various sources vary significantly. This inconsistency is due to the complexity of its composition and substantial differences in the properties of coal tars produced by different manufacturers.

Generally, coal tar is a complex physicochemical system, with a number of components present in an emulsified or dispersed state. The major part of this material consists of aromatic hydrocarbons. The presence of solid

carbonaceous particles (commonly referred to as “carbon blacks”) and water globules with varying salt content significantly affects the dielectric permittivity threshold of coal tar.

As for the dispersion matrix of coal tar itself, it exhibits high electrical resistance. For example, a protective mastic used for pipeline insulation, based on anhydrous coal tar, demonstrates the following parameters (ShaliEnamel. “Coal Tar Enamel for Pipe Coating,” Url: https://stpltd.com/wp-content/upload/TDS/ShaliEnamel.pdf.):

- Electrical resistance, 10000 volts/mm – No breakdown;
- Dielectric strength, kV/mm, Min – 10;
- Volume resistivity, ohm·cm, Min – 10^{13}.

Therefore, one of the objectives of this study was to refine certain dielectric characteristics of raw coal tar, particularly the threshold of its dielectric permittivity.

It is known, that in liquids containing water impurities, the breakdown is preceded by the formation of a conductive channel between electrodes. Water droplets polarize and elongate under the electric field, forming ellipsoids that merge and, under sufficient field strength, create a highly conductive channel. This results in rapid heat release, liquid boiling, and a gaseous discharge along the vapor channel. Similar effects are observed in dielectrics with solid conductive inclusions (e.g., carbon particles) [6].

In our case, the coal tar contained both conductive solid inclusions and water (3.7%). The explosive boiling of liquid in the discharge zone inside the sealed vessel served as the initiator of the hydraulic shock.

Under such conditions of hydraulic shock initiation, an additional source of physical impact on the solid-dispersed particles is the cavitation effect, which arises due to the boiling not only of water but also of the light fractions of coal tar under hydrodynamic cavitation. It is well known that cavitation effects are used, for example, in wet coal grinding to produce fine coal suspensions in liquid dispersive media [7]. To some extent, this can be regarded as a simplified analogue of the effect we aimed to achieve in our study.

The characteristics of the initial coal tar are presented in Table 1.

TABLE I. CHARACTERIZATION OF THE INITIAL COAL TAR

Parameter	**Coal Tar Sample 3**
Density (dry mass), kg/m³	1197
Moisture content, %	3.7
Mass fraction of α_1-fraction, %	5.8
Mass fraction of α-fraction, %	11.1
Ash content, %	0.22
Distillation characteristics – mass fraction distilled at temperature range, °C, %: – up to 180 – 180–200 – 200–230 – 230–300 – 300–360 – above 360 – losses	 1.8 0.5 16.1 8.7 12.4 60.2 0.3

The electric circuit of the laboratory setup included a DC power supply, a capacitor, and an insulated discharge cell with two oppositely oriented electrodes. After static breakdown was achieved, the electrical discharge parameters necessary to initiate hydraulic shock in the tar were determined: voltage – 50 kV, capacitor capacity – 100 pF, discharge duration – 100 μs.

The efficiency of dispersion was assessed by the reduction in the mass fractions of quinoline-insoluble (α_1-fraction) and toluene-insoluble (α-fraction) substances, using the standard hot filtration method. It is worth noting that according to TU U 17.1-2510012269-001:2015 “Laboratory filter paper”, the density of filter papers used in this method is not specified by pore size (which is nearly impossible to define for paper materials) but by the duration of distilled water filtration through a funnel-shaped folded filter.

Therefore, paper filters are unsuitable for precise granulometric analysis of the solid dispersed phase using screening techniques. Nevertheless, as shown later, the hot filtration method provides insight into the behavior of solid dispersed particles under the studied conditions.

III. RESULT AND DISCUSSION

Coal tar subjected to electric discharge was analyzed for toluene-insoluble (α-fraction) and quinoline-insoluble (α_1-fraction) content. The results are presented in Table 2.

The experimental data show that the α-fraction content in tar decreases after exposure to hydraulic shock. The difference between the untreated tar and the sample subjected to 50 discharges is significant—over 2% absolute—which is approximately four times higher than the margin of analytical error.

A similar trend is observed for the α_1-fraction. and the reduction in both fractions suggests that the decrease in toluene-insoluble substances is predominantly due to a reduction in the quinoline-insoluble component.

TABLE II. EFFECT OF HYDRAULIC SHOCKS ON COAL TAR PROPERTIES

Indicator	**Number of Discharges**			
	0	**1**	**5**	**50**
Mass fraction of α-fraction, %	11.1	10.4	9.5	8.9
Reduction in α-fraction per discharge. % abs.	–	0.70	0.32	0.04
Mass fraction of α_1-fraction. %	5.8	4.9	4.0	3.8
Reduction in α_1-fraction per discharge. % abs.	–	0.80	0.36	0.08

Since significant chemical changes of the solid dispersed phase are unlikely under the given experimental conditions, we assume that the hydraulic shock induced by electric discharge causes the mechanical destruction of quinoline-insoluble particles—likely aggregates in particular. As a result of this fragmentation, dispersed particles of reduced size are able to pass through the pores of the filter during analysis, thereby creating the appearance of a reduced α- and α_1-fraction content. This observation is consistent with our earlier research, which confirmed that the proposed method provides a reliable estimation of the average particle size within the α_1-fraction.

To illustrate this, Table 3 presents the results of determining the mass fraction of quinoline-insoluble substances, obtained using the standardized hot filtration method, for tars sampled at various stages of pyrolysis (sample B corresponds to the tar used in the present study). According to the standard procedure, two layers of "white ribbon" filter paper are typically employed. However, in order to better understand the filtering process, we conducted additional measurements under otherwise identical conditions using one and three filter layers. As shown in the table, the measured yield of the target fraction consistently increases as the number of filter layers is raised. This trend strongly suggests that the standard filter configuration permits at least some α_1-fraction particles to pass through, meaning the obtained values may underestimate the true content.

Thus, there is no doubt that with an increase in the number of filter layers, the number of small-class particles captured on the filter also increases. This finding further supports the assumption that the apparent reduction in fraction content observed after electrical discharge treatment is not due to chemical decomposition but rather due to changes in particle size distribution.

TABLE III. EFFECT OF FILTER MATERIAL DENSITY ON THE DETERMINATION OF THE A1-FRACTION MASS CONTENT

Sample	Number of "white ribbon" filter paper layers		
	1	2 (standard)	3
A	4.3	6.1	6.2
B	4.4	5.8	5.9
C	3.9	4.7	5.4

However, to verify the conclusion that the decrease in the mass fraction of the fractions studied is caused by a decrease in the size of their particles, we investigated the granulometric composition of the α_1-fraction in both untreated and 50-discharge-treated tar using a method based on scanning microscopy [8]. The results are presented in Figure 3. which also shows the composition of the solid phase after 38 hours of aging at 80°C.

The data clearly confirm that hydraulic shock treatment results in a pronounced shift in the granulometric distribution of the solid phase toward finer particles. Consequently, the observed decrease in high-molecular-weight components in Table 2 can be explained by the transformation of the material into ultrafine fractions that are capable of passing through the filter during analysis.

Conversely, the long-term exposure of tar to typical storage and pumping temperatures leads to partial restoration of the original particle size distribution. This finding provides further support for our earlier assumption that dispersion of the solid phase is most effective when performed during a technological stage that simultaneously increases the viscosity of the dispersion medium—namely, during thermal fractionation—since under these conditions re-aggregation of particles is significantly hindered [5].

Taken together, these results demonstrate that the reduction in α- and α_1-fraction content observed after electrical discharge treatment is primarily a mechanical effect associated with particle size reduction rather than a chemical transformation. The choice of filtration conditions plays a crucial role in correctly interpreting the data, as filters with insufficient retention capacity may allow ultrafine fractions to pass through. Furthermore, the tendency of the solid phase to partially re-aggregate under typical storage conditions highlights the importance of integrating dispersion processes into stages of technological treatment where viscosity is high and re-aggregation is minimized. This ensures more stable and reliable modification of the solid dispersed phase, with potential implications for improving tar processing efficiency and product quality.

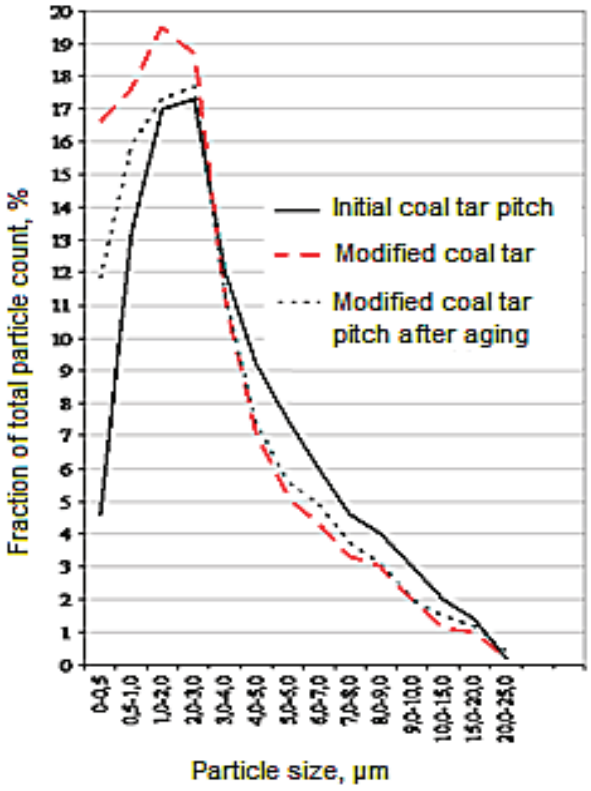


Fig. 3. Granulometric composition of the solid-dispersed phase of untreated and processed coal tar

IV. CONCLUSIONS

Hydraulic shock initiated by electric discharge is an effective method for dispersing the solid dispersed phase of coal tar—particularly the quinoline-insoluble (α_1) fraction. This is evidenced by the reduction in their mass content (via hot filtration) and by the granulometric redistribution toward finer particles.

A reduction in coal tar viscosity enhances the mobility of microscopic dispersed particles, facilitating their aggregation due to surface activity. Conversely, processes that increase liquid phase viscosity (e.g., distillation or rectification) impede aggregation.

Under hydraulic shock initiation, cavitation—resulting from boiling of both water and light tar fractions—provides additional physical impact on solid particles.

The reduction in α- and α_1-fraction content is not due to chemical degradation, but rather to mechanical disintegration, which causes particles to pass through filter pores during analysis

Dispersion increases the specific surface area of the solid phase, improving its interaction with the dispersion medium. This can significantly enhance the properties of binders and impregnation pitches derived from coal tar.

Ultrafine α_1-fraction particles (0–1 μm) penetrate more easily into the pores of graphitized electrode blanks, increasing impregnation without compromising coking capacity.

However, the dispersion effect is temporary: prolonged storage leads to partial re-aggregation. Therefore, this treatment is best applied during thermal fractionation, where aggregation is limited due to increased viscosity of the non-volatile residue.

The presented studies were undertaken with the aim of proving the fundamental possibility of adjusting the granulometric composition of the solid dispersed phase of coal tar.

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Mathematical Simulation Of Brushless High-Speed Permanent Magnet Motor

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***Abstract*— To develop a numerical two-dimensional field mathematical model of a contactless high-speed permanent magnet motor to estimate its parameters, characteristics and determine the magnitude of losses in the magnetic core under different control methods. Methods. The finite element method was used to calculate the electromagnetic field distribution in the computational domain of the motor under study. To develop two-dimensional and three-dimensional drawings of the design area, methods and tools of computer-aided design were used. Numerical methods for calculating losses in the magnetic core developed by Giovanni Bertotti were used to calculate losses in the magnetic core. The Fourier series expansion method and spectral analysis methods were used to model various motor control methods. Results. A two-dimensional numerical circular-field mathematical model of a contactless high-speed motor with permanent magnets has been developed. The model was developed to estimate the distribution of the electromagnetic field and eddy currents in the structural and active elements of the motor in question in order to determine the magnitude of losses under different power supply methods. The paper investigates the dependence of losses in individual structural elements of the motor under study when powered by a sinusoidal voltage source and when powered by a PWM inverter. Replacing permanent magnets with rectangular magnets reduces the cost of motor manufacturing. The use of rectangular permanent magnets reduces losses in the computational domain, the magnitude of electromagnetic torque fluctuations, but reduces the magnitude of traction. Scientific novelty. Using the developed numerical circular-field mathematical model, it is proved that losses in the magnetic core of a high-speed motor and its structural elements significantly depend on the control method and configuration of the magnetic system and the motor as a whole. The study allows solving urgent scientific and practical problems related to the optimization of the structure depending on next criteria: reduction of heating or losses, reduction of vibration and noise, etc.**

Keywords*— high-speed motor, mathematical simulation, permanent magnets, losses in the magnetic core, electromagnetic torque.

I. Introduction

The overall dimensions of electromechanical energy converters (EMCs) significantly depend on the rotor speed or frequency of the supply voltage; it is known that the higher the rotational speed, the smaller the overall dimensions.

High-speed permanent magnet motors are experiencing a trend towards higher power density and smaller dimensions. Increasing the rated speed of rotation reduces the size and weight of electric machines, making them more efficient for applications where weight, compactness and performance are critical.

The development of new high coercive permanent magnets and the use of CNC tools for EMF production leads to the emergence of new types of electric motors that meet the growing requirements for energy efficiency, reliability and compactness [1, 2]. At the same time, increasing the speed of motors requires an increase in the accuracy of design and manufacturing, causes an increase in losses and heating, and raises certain issues related to dynamic stability [3, 4]. To solve these problems, it is necessary to use modern approaches, methods and tools of mathematical modelling that provide accurate analysis and allow optimization of engine parameters [5]. Modern application packages for numerical mathematical modelling allow for calculations when electromagnetic, thermal and mechanical processes occurring in the system interact and their impact on the parameters and characteristics of an electric motor or EMF.

High-speed motors are those that operate at speeds exceeding several tens of thousands of revolutions per minute [6], which opens up wide opportunities for their application. For example, such motors are used in compressors, medical equipment, high-speed spindles for machining machines, and even in modern electric vehicles. However, at high speeds, it is important to take into account factors that affect the stability of the motor, such as heating, vibrations, electromagnetic losses, and mechanical deformations [7, 8].

Brushless high-speed permanent magnet motors (BHSPMM) are increasingly used in various industries, transport, aviation, power generation, power tools, household appliances, etc. [9]. They have certain key advantages that are a prerequisite for their proliferation: high specific power, low weight, relatively small overall dimensions, no mechanical contacts and minimal maintenance due to reduced wear. These motors are also used in a variety of specific applications that require minimal size and weight, high speed, accuracy and reliability [10].

When developing, designing and optimizing BWDMs, it is necessary to take into account the multiphysical processes occurring in the core of this type of EMF: temperature, losses, magnetic induction values in individual parts of the magnetic

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core, etc., which determines reliability, efficiency and maintainability [11].

In addition, the accurate determination of magnetic core losses is important for the analysis of efficiency, losses, temperature.

II. Materials and Research Findings

A three-dimensional geometric model of the prototype of the studied contactless high-speed motor with permanent magnets is shown in Fig. 1.

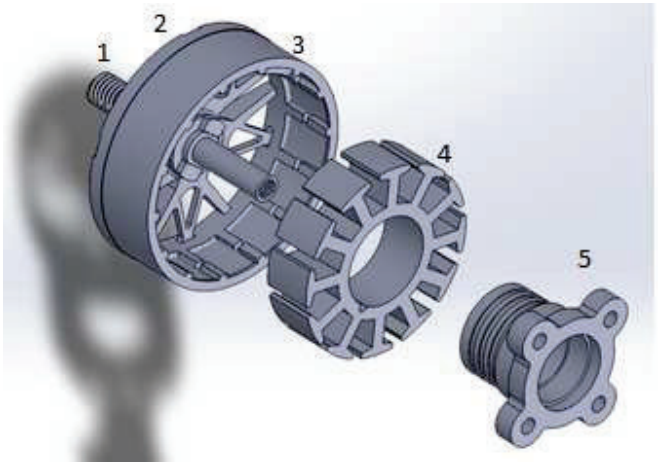


Figure 1. Design of the studied BHSPMM [10]

In Fig. 1 is show: 1 - rotor shaft; 2 - rotor end cover; 3 - rotor bushing with permanent magnets; 4 - stator magnetic core; 5 - stator bushing.

The shaft is made of magnetic steel and has a through bore. The rotor end cover is mounted at the output end of the shaft and is configured to improve motor cooling at high speeds. The cover is made of a single piece of lightweight non-magnetic but electrically conductive material. The rotor bushing is made of a solid (not ring-shaped) magnetically conductive material, which has protrusions in the internal bore in the radial and axial directions for fixing the PM.

The permanent magnets are mounted between the protrusions on the inner surface of the rotor hub and are magnetized radially. The PMs have a complex radial shape that follows the arc of the internal bore of the rotor hub on one side and allows for a uniform air gap between the stator and rotor of the motor.

The motor stator (stator magnetic core) is made of charged plates made of electrical steel. The stator bushing is made of a lightweight non-magnetic alloy as a single piece. Two bearings (not shown in Fig. 1) are installed in the inner part of the bushing. The stator winding is placed in the grooves of the stator magnetic core and connected in a triangle pattern [10].

The main design parameters of the BHSPMM prototype used to develop the mathematical model are given in Table 1 [8], [9]. The permanent magnets are made of sectoral magnets with homogeneous magnetization in the radial direction from neodymium-iron-boron alloy N52H with a residual magnetic induction of 1.36 T at 20 °C.

For the calculation, it is assumed that the magnetic core of the stator is made of cold-rolled isotropic electrical steel grade 2411. It is assumed that the rotor bushing is made of a composite magnetically soft material based on Somaloy. The high-speed rotor shaft is made of structural steel grade 45.

Based on the data in Table 1, a geometric model is developed and then imported into the COMSOL Multiphysics numerical modelling environment.

The problem is solved by the finite element method. The electromagnetic field and characteristics of the motor under study are analyzed in a two-dimensional approximation, taking into account the following simplifications and assumptions:

1. The axial length of the BHSPM is taken into account empirically, i.e., longitudinal end effects are not taken into account;
2. The properties of electrical steel of the magnetic core are determined by the main magnetization curve, magnetic hysteresis is not taken into account;
3. Since it is difficult to accurately account for the full magnetization and demagnetization curve of permanent magnets, their characteristics are determined only by the residual magnetic induction and coercive force;
4. The magnetic induction vector is considered to be tangent to any point within the computational domain.

The distribution of the electromagnetic field in the computational domain of the study motor is described by equations with respect to the vector magnetic potential and supplemented by boundary conditions with respect to known magnetic field sources:

$$\nabla^2 A = \mu \nabla \times M; \tag{1}$$

$$n \times (A_1 - A_2) = 0; \tag{2}$$

$$n \times \left([\mu_0^{-1} \nabla \times A - M] - [\mu_0^{-1} \nabla \times A - M] \right); \tag{3}$$

$$n \times A_{in} = 0, \tag{4}$$

where: $\boldsymbol{A}$ is the vector magnetic potential, Vb/m; $\boldsymbol{M}$ is the magnetization vector, A/m; A_{in} is the vector magnetic potential at the outer boundary of the computational domain, Vb/m; $\boldsymbol{A_1}$, $\boldsymbol{A_2}$ are the vector magnetic potentials at the boundary of adjacent domains 1 and 2, Vb/m; n is the normal vector.

The electromagnetic torque of a high-speed motor is calculated using the Maxwell's magnetic tension tensor, which has proven to be a reliable method for determining this kind of force [11]. The magnitude of the force acting on the outer rotor of the motor under study is determined by the following expression:

$$f = \mu((nH)H - 0{,}5nH^2, \tag{5}$$

where $\boldsymbol{H}$ is the magnetic field intensity vector, A/m:

$$H = \mu^{-1} \nabla \times A - M. \tag{6}$$

The total electromagnetic torque is the result of integrating the electromagnetic forces acting between the external rotor and stator of the motor and is calculated as follows:

$$T_{em} = \int_{S_g}^{-} (r - r_0) \times f dS, \tag{7}$$

where: S_g is the integration area, f is the distribution of electromagnetic force on the inner surface of the rotor of a high-speed motor under study.

Losses in the stator magnetic core were obtained using the modified Bertotti equation [12] as a result of calculating the magnetic field inside the motor computational domain for each harmonic. The value of losses in the magnetic core is calculated according to the following expression:

$$Q_c = k_h \cdot (n \cdot f)(|\boldsymbol{B}|)^{\beta} + k_c \cdot \sigma \cdot n \cdot f^2 \cdot |\boldsymbol{B}|^2 + 8 \cdot \sqrt{\sigma \cdot G \cdot S \cdot V_0} \cdot (n \cdot f)^{1,5} \cdot (|\boldsymbol{B}|)^{1,5}, \quad (8)$$

Where: $Q_{(c)}$ is the total losses in the magnetic core; σ is the electrical conductivity of the blended magnetic core; G is a constant in the equation, which is taken equal to 0.1356 ; S is the area of the computational domain for which the losses are determined; V_0 is the excess loss coefficient; $k_{(c)}$ is the coefficient of eddy current losses determined by the thickness of the electrical steel sheet; n is the ordinal number of the harmonic; β is the exponential coefficient for magnetic induction; f is the frequency of the fundamental harmonic; kh is the hysteresis loss coefficient.

The eddy current loss factor k_C describes the eddy current losses in the magnetic core and is determined by the thickness of a single sheet of electrical steel:

$$k_c = \frac{\pi^2 \cdot d^2}{6}, \quad (9)$$

where: d is the thickness of a single sheet of electrical steel of the charged magnetic core of the stator of the motor under study.

Equation (8) has three components responsible for: the first component corresponds to the hysteresis losses; the second component characterizes the eddy current losses; the third component describes the excess losses arising from the rotor rotation of the motor under study, according to the Bertotti method.

The hysteresis loss factor kh, eddy current loss factor k_C and excess eddy current loss factor are a function of the current frequency.

Losses in other electrically conductive elements of the design domain of the motor under study (PM, rotor sleeve, shaft, stator winding [13]) were calculated using the following expression:

$$Q_j = {}^1\!/_T \int_{T_1-T}^{T_1} J \cdot E dt, \quad (10)$$

where: T, T_1 - the period of time for which the losses are calculated; J - the current density in the calculation area; E - the electric field strength.

Table 1 – Main design parameters of the prototype BHSPMM

TABLE I. MAIN DESIGN PARAMETERS OF THE PROTOTYPE BHSPM

№	Parameter	Value
1	Number of rotor pole pairs	7
2	Rotor outer diameter, mm	36,0
3	Axial rotor length, mm	9,5
4	Axial length of the stator magnetic core, mm	8,0
5	Residual magnetic induction of the PM, T	1.36 (N52H)
6	Outer diameter of the stator magnetic core, mm	30,0
7	Stator magnetic circuit material	Steel 2411
8	Rotor hub material	Samoloy
9	Rotor inner diameter, mm	33,2
10	Electrical conductivity of permanent magnets, MSm/m	0,56
11	Electrical conductivity of Somaloy, MSm/m	0,14
12	Height of permanent magnets, bmag, mm	1,5
13	Inner/outer diameter of permanent magnets, mm	30,4/33,4
14	Maximum rotor speed, rpm	29100
15	Air gap size, mm	0,2
16	Diameter of the winding wire, mm	0,25
17	Maximum power, W	1300
18	Rated supply voltage, V	24,0

Also, losses in the stator magnetic core were obtained using the modified Bertotti equation as a result of calculating the magnetic field inside the motor computational domain for each harmonic.

III. LOSSES AT DIFFERENT POWER SUPPLY TYPES

The estimation of losses in the computational domain of the motor under study is determined by expressions (1), (2). The division of losses into losses in the magnetic core, PM, rotor sleeve and armature winding is performed by integrating the volumetric losses in each area separately. Obviously, in this case, the magnitude of losses in structural elements is determined by the rotation speed (supply voltage frequency) and the magnitude of magnetic induction in individual motor sections.

The shape of the motor supply voltage curve and the spectral decomposition when powered by a PWM inverter show on Fig. 2.

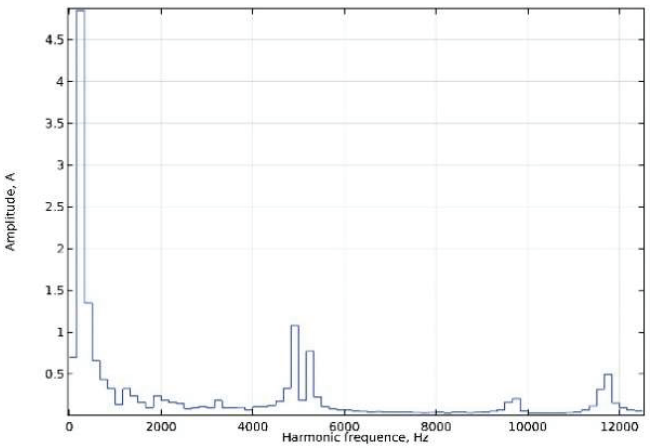


Figure 2. *The voltage supply curve shape and the spectral decomposition when powered by a PWM inverter*

In the spectrum of the supply voltage curve, the main harmonic occupies the main weight, and there are also clearly expressed harmonics of the order of 24, 32, 56, 58, etc.

Obviously, these are high-frequency harmonics and they lead to an increase in losses.

TABLE II. RESULTS OF CALCULATION OF LOSSES AND ELECTROMAGNETIC TORQUE AT DIFFERENT POWER SUPPLY METHODS

n, (rpm)	T_{em} (Н·м)		sin power supply	PWM
	sin	PWM	Losses, (W)	Losses, (W)
3000	0,505	0,491	21,67	35,14
6000	0,735	0,722	28,45	46,65
9350	0,965	0,951	38,41	65,68
12500	1,270	1,259	56,4	102,27
16500	1,487	1,473	74,5	138,95
19000	1,704	1,691	107,17	182,78

Dependence of motor losses on rotational speed show on Fig. 3.

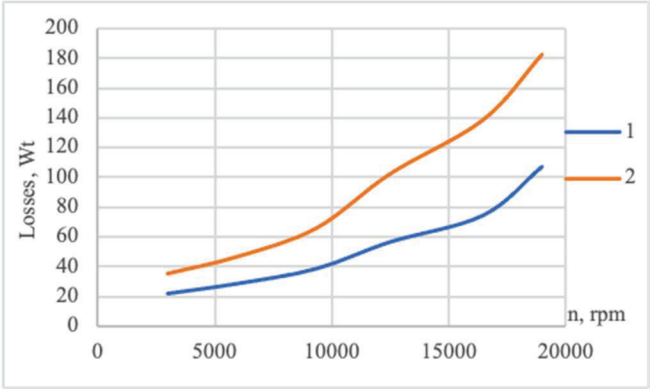


Figure 3. *Dependence of motor losses on rotational speed*

The curves in Fig. 3 represent two types of power supply for the motor under study:

1 — power supply from a sinusoidal voltage source;

2 — power supply of the motor from a PWM inverter.

Obviously, as the rotational speed increases, the losses increase. This is due to an increase in the magnetisation frequency and, accordingly, an increase in hysteresis and eddy current losses. In other electrically conductive structural elements (rotor sleeve, stator winding), losses increase due to the amplification of the current displacement effect, which manifests itself with an increase in rotational speed and supply voltage frequency.

When the motor is powered by a PWM inverter, the losses increase by approximately 53–62% depending on the rotation speed. This is due to an increase in losses from higher harmonics when the motor is powered by a PWM converter. In addition, losses in the design domain increase nonlinearly: when the rotation speed increases by a factor of 2, the losses increase by ≈30%; when the rotation speed increases by a factor of 3, the losses increase by ≈77%; when the speed increases by a factor of 4, the losses increase by ≈260%.

At maximum rotational speeds of 20,000 rpm, the losses increase by ≈5.4 times compared to the initial rotational speed of 3,000 rpm. This is due not only to the increase in the magnetisation frequency, but also to the increase in active resistance due to the enhanced effect of current displacement in the conductive medium.

IV. LOSSES AND TORQUE WITH MODIFIED PM

The radial-shaped PMs are used in the prototype of the BHSPMM under study. On the one hand, this ensures the uniformity of the air gap and allows filling the useful volume with permanent magnets to the maximum extent possible, which allows obtaining the maximum performance characteristics: power and torque in a given volume [9]. Distribution of the electromagnetic field at different shapes of the PM shows on Fig. 4.

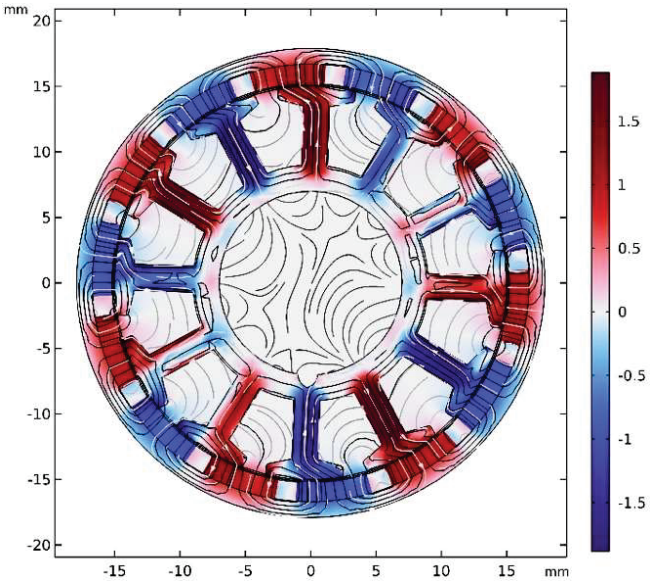


Figure 4. Distribution of the electromagnetic field at different shapes of the PM

However, this form of motor also has disadvantages: manufacturing a motor with this configuration is much more expensive and requires special expensive equipment, and unification is virtually impossible, which leads to an increase in the cost of the manufactured motor. It is possible to reduce the cost of manufacturing a serial sample of this type of motor by using rectangular motor shapes.

The dimensions of the rectangular PMs are selected so that the minimum air gap remains unchanged and is ≈0.2 mm. Obviously, the use of rectangular shaped PMs leads to a decrease in the volume occupied by PMs by ≈23%. This leads to a decrease in the electromagnetic torque, power, and magnetic induction in certain parts of the magnetic system. On the other hand, the magnitude of losses is reduced, both by reducing the magnitude of induction and by reducing the influence of longitudinal end effects. This will lead to reduced motor heating, increased efficiency and operational reliability.

To estimate the losses and the magnitude of the traction force (electromagnetic torque), a series of calculations were carried out for a motor with a rectangular PM shape. The results of calculating the losses in the computational domain when powered by a sinusoidal voltage source and when powered by a PWM inverter are shown in Table III.

TABLE III. RESULTS OF THE CALCULATION OF BHSPMM LOSSES UNDER DIFFERENT FORMS OF PM

n, (rpm)	Prototype		Rectangular PMs	
	sin power supply Losses, (W)	PWM losses, (W)	sin power supply Losses, (W)	PWM losses, (W)
3000	21,67	35,14	17,31	28,07
6000	28,45	46,65	21,86	35,84
9350	38,41	65,68	27,42	46,88
12500	56,4	102,27	41,75	75,7
16500	74,5	138,95	49,17	92,38
19000	107,17	182,78	68,6	116,99

Dependence of losses on the shape of the PM are show on Fig. 5.

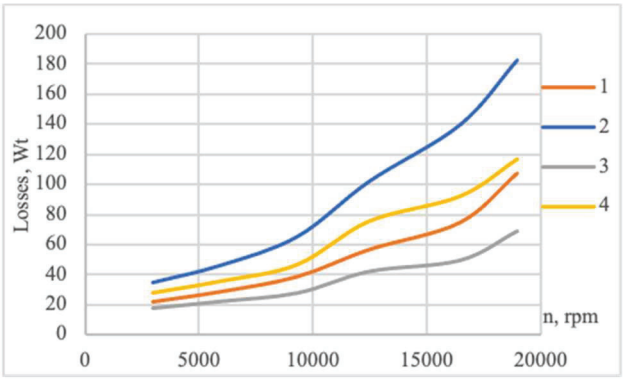


Figure 5. Dependence of losses on the shape of the PM

Fig. 5 shows: 1 – prototype motor losses at sinusoidal power supply voltage; 2 – prototype motor losses when powered by a PWM inverter; 3 – motor losses at sinusoidal power supply voltage and rectangular PM shape; 4 – motor losses when powered by a PWM inverter and rectangular PM shape.

In general, depending on the rotational speed, motor losses when using a rectangular PM shape are reduced by 20–36%.

When changing the configuration of the PM, the value of the electromagnetic torque changes. The dependence of the electromagnetic torque at different shapes of the PM on the rotational speed is shown in Figure 5.

CONCLUSIONS

The study of high-speed permanent magnet motors for their use in various devices and systems is of great scientific and practical value. These motors have compact dimensions and a relatively high power density, as well as well-developed control and regulation facilities.

The results of the calculations show that when the motor is powered by a PWM inverter, the losses increase by ≈53–62% depending on the rotation speed compared to the case of power supply from a sinusoidal voltage source. Losses in the calculated area increase not proportionally with increasing rotational speed: with a 2-fold increase in speed, losses increase by ≈30%; with a 4-fold increase in speed, losses increase by ≈260%. At maximum rotational speeds of 20,000 rpm, the losses increase by ≈5.4 times compared to a speed of 3,000 rpm.

Locally, the largest losses are concentrated in the stator magnetic core on the teeth as the most saturated elements of the magnetic system. The total losses in these areas are ≈37% of the total calculated losses.

The losses and electromagnetic torque were estimated when using the rectangular shape of the PM. This will simplify and reduce the cost of manufacturing the final product, increase the level of unification in the manufacture of serial samples. Depending on the rotational speed, motor losses when using this motor shape are reduced by 20–36%. The value of electromagnetic torque, when using rectangular PMs, decreases by ≈22.5–30% depending on the rotation speed.

The results of the research provide practical recommendations for the design and operation of BWDMs to improve efficiency, operational reliability, losses and heating.

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Data Encoding Algorithm for Transmission from Wearable Biomedical Sensors via Wireless Communication Channels

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***Abstract*—This paper addresses the problem of efficient data transmission from wearable biomedical sensors via wireless communication channels in human motion biomechanics monitoring systems. A combined approach to sensor data compression is proposed, which includes preliminary octahedral transformation of accelerometer vectors followed by the use of the lossless Sprintz compression algorithm. This method accounts for the vector nature of the signal, reduces redundancy, and improves compression efficiency without significantly increasing the computational or memory load. Experimental evaluation demonstrated that the application of octahedral encoding reduces accelerometer data volume to 57% (compared to 84% without transformation), and the total transmitted data volume — to 61%, depending on the type of activity. The proposed approach is suitable for implementation on resource-constrained microcontroller-based platforms and is recommended for use in real-time systems.**

Keywords—biomechanics, wearable electronics, wireless sensors, ESP-Now, IMU, monitoring, octahedral encoding, Sprintz

I. Introduction

According to data from the World Health Organization, approximately 1.71 billion people — around 22% of the global population — suffer from musculoskeletal disorders, which lead to temporary or permanent limitations in physical activity, professional performance, and social interaction [1]. This category includes more than 150 different nosological forms characterized by damage to muscles, the skeletal system, joints, and surrounding connective tissues. These pathologies vary in their clinical course — from acute, short-term conditions such as fractures, sprains, and muscle strains accompanied by pain and functional impairment, to chronic diseases, including primary chronic lower back pain and osteoarthritis. Notably, lower back pain is among the most common medical issues over a lifetime, affecting up to 80% of individuals over the age of 35, leading to reduced productivity and quality of life. Under the current wartime conditions in Ukraine, there has been a significant increase in the number of limb amputations — according to estimates by Ottobock (Germany), a leading global manufacturer of prosthetics, the number of people with amputated limbs has reached approximately 50,000 [2].

Under these conditions, the use of biomechanical analysis methods for identifying and assessing functional impairments of the musculoskeletal system is of particular relevance, as it enables effective planning of therapeutic interventions. Modern non-invasive technologies are employed for this purpose, including electronic and electromechanical systems, smart clothing, exoskeletons, and other digital devices. In recent years, there has been a significant increase in the number of studies focused on the application of such technologies in biomedical contexts — for diagnostics, therapy, functional correction, and prevention of musculoskeletal disorders [3, 4, 5].

II. Review of Existing Solutions

Since the early 2000s, a large number of inertial measurement units (IMUs) have been developed, combining accelerometers and gyroscopes within a single device [5]. Key trends in the evolution of IMUs include miniaturization, reduced power consumption (down to microampere levels in low-power modes), and improved accuracy and stability of measurements. These characteristics have led to the widespread use of IMUs as wearable biomedical sensors [4], particularly for monitoring human motion biomechanics. A key feature of such sensors is the ability to transmit data wirelessly, which is essential given their typical placement on a freely moving human body. These sensors are integrated into a distributed monitoring system with a central device — a base station — that performs data collection, storage, and preliminary processing [6]. The central device can be a personal computer, a smartphone, or a specialized embedded computer. After data acquisition is complete, the recorded information is typically synchronized, resampled, and subjected to further analysis and processing.

A typical solution for establishing communication between sensors and the base station involves the use of one

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of several wireless protocols, such as BLE/Bluetooth, Wi-Fi, Zigbee, ESP-Now [7, 8, 9], among others. One of the critical challenges in such systems is the synchronization of measurements from multiple sensors. A typical biomedical biomechanics diagnostic system consists of at least four sensors, which may, for example, be placed on the lower limbs of the patient. To ensure stable synchronization during wireless data transmission, a key parameter is the latency — the time required for a packet containing a time reference from the base station to reach the sensors. Therefore, it is advisable to select communication protocols with minimal transmission delay, such as ESP-Now [9]. However, such protocols are generally characterized by simplified connection mechanisms and limited bandwidth, despite operating on fixed frequency and channel width. This necessitates the implementation of additional mechanisms for managing time-division access to the channel (Time on Air, ToA) for sensor data transmission. At the same time, human biomechanics monitoring must be performed in real-time — sensor data are generated and transmitted continuously and must be stored without loss.

Given the aforementioned limitations, the task of reducing the volume of data transmitted via wireless communication channels becomes particularly relevant. Several approaches can be considered to address this challenge. One possible method is to reduce the sampling rate of the sensors. However, this approach significantly limits the potential for further analysis and detailed interpretation of the collected data, making it unsuitable for most applications involving precise monitoring of human motion biomechanics. Typical sampling rates in such systems range from 100 to 200 Hz, which provides sufficient temporal resolution for accurate assessment of dynamic movement parameters.

Another potential solution involves reducing the bit depth of the sensor-generated data [10]. However, this option also has substantial constraints: the typical bit resolution of signals is 12 to 16 bits per sample, which is the minimum required to preserve measurement accuracy and reliability. In the case of using two three-axis sensors (e.g., an accelerometer and a gyroscope), the total data volume from a single measurement reaches approximately 12 bytes. This imposes relatively high demands on the bandwidth of the communication channel, even when only a small number of sensors are used in the system.

At present, numerous sensor data compression methods have been developed, including well-known general-purpose algorithms such as Huffman coding [11], as well as less common approaches like block-dictionary encoding algorithms [12]. In addition, specialized compression methods are actively being developed. These differ from general-purpose techniques in that they require minimal computational resources for both encoding and decoding and are specifically designed for real-time processing of small data blocks [13], taking into account the characteristics of data generated by various types of sensors [14]. This study focuses on a lossless compression algorithm for sensor data, along with its software implementation. The proposed approach is based on the Sprintz method [15], which is designed for highly efficient compression of small time-series data blocks.

III. Data Encoding Algorithm

Sprintz is an efficient time series compression algorithm specifically designed for processing data from high-frequency sensors under constrained computational resources. This approach is particularly relevant for use in wearable devices and Internet of Things (IoT) systems. It offers fast compression and decompression while requiring minimal additional memory, which is critical for implementation on microcontroller-based platforms with limited resources. The algorithm is tailored for compressing time series represented as sequences of low-bitwidth integers (typically 8–16 bits), which closely corresponds to the data format generated by most sensor devices.

The method employs a predictor — either delta coding [16] or FIRE (Fast Integer Regression Estimator) — to forecast the next value based on a linear combination of previous samples. After prediction, the difference (prediction error) between the estimated and the actual value is calculated.

These differences typically have smaller amplitudes compared to the original values and exhibit reduced dynamic range, which significantly improves the efficiency of subsequent compression. In the next stage, the calculated differences are packed using bit-level encoding. For each fixed-size block of samples, the maximum number of bits required to represent the largest absolute value among the differences is determined. Based on this, all values within the block are encoded using a uniform bit width.

This block-based approach enables convenient integration of the algorithm into sensor data acquisition and transmission systems. In typical IMU (Inertial Measurement Unit) use cases, data are collected in a FIFO mode: values are stored in the sensor's internal memory and transferred to the microcontroller once a specified amount of data has been accumulated. This naturally forms a block of time-series data, which can be compressed independently. This independence is a key advantage, as the loss of a single block during transmission does not affect the decoding of others — the compression process does not rely on inter-block context.

After compression, each block for each channel follows a defined structure (Fig. 1), which includes the initial value of the sequence, the number of bits used for encoding, and the actual encoded differences aligned to byte boundaries.

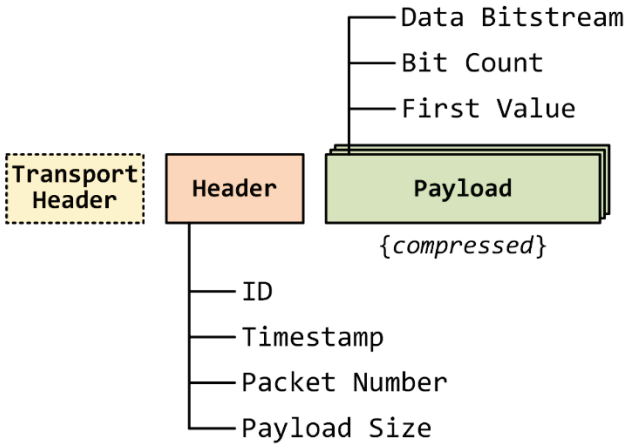
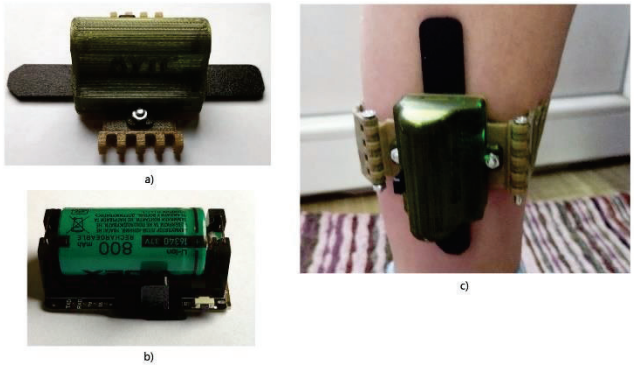


Fig. 1. Structure of compressed data blocks

IV. Implementation of the Algorithm Based on a Distributed Biomechanical Monitoring System

The algorithm was implemented within an existing monitoring system (Fig. 2), which is built using the BMI-160 IMU sensor module and an ESP32C3 microcontroller with an autonomous power supply. The system transmits data using the ESP-Now protocol in packet mode, with time-division access to the radio channel between sensors.



a) monitoring system in enclosure;
b) main board of the system with installed power supply;
c) method of attachment to the lower limb of a person.

Fig. 2. Wireless human biomechanics monitoring system

Each sensor module collects data from both the accelerometer and gyroscope, which are transmitted within a designated time slot along with time synchronization markers. The data rate from each sensor is 100 Hz, resulting in a total data throughput of approximately 1200 bytes per second per module. Experimental testing of the system was conducted during gait biomechanics analysis, with sensors mounted on the subject's lower limbs. The use of the Sprintz algorithm reduced the average volume of transmitted data to 75% of the original amount.

The transmitted data consist of the following components:

1. Packet headers (not subject to compression), accounting for approximately 8–10% of the total volume.
2. Accelerometer data — approximately 45%.
3. Gyroscope data — also approximately 45%.

Analysis of the compression efficiency for individual data streams showed that gyroscope data are compressed significantly more effectively, demonstrating an average compression ratio of approximately 63% during gait simulation at a walking speed of 3 km/h. In contrast, accelerometer data exhibit lower compression efficiency under the same conditions — only about 84%. A closer examination of the accelerometer data reveals that a significant portion is occupied by the gravity vector, while the acceleration components along the different axes are highly correlated. To improve compression efficiency, it is advisable to treat accelerometer data not as three independent channels, but rather as a single vector signal, from which both direction and magnitude can be derived. This vector-based model enables the application of transformations that reduce data redundancy and thus increase the overall compression ratio [17]. In particular, encoding the magnitude of the acceleration vector separately is beneficial, as it tends to change more slowly compared to the vector's orientation — providing opportunities for more aggressive compression without compromising accuracy.

Given the observed correlation among accelerometer data components and their inherent vector nature, it is appropriate to apply specialized spatial encoding methods that enable a more compact representation of the acceleration vector's orientation. One such approach is octahedral encoding [18], which was proposed for the efficient representation of normalized vectors in three-dimensional space.

The core idea of the method involves projecting a unit vector (normal) onto the surface of an octahedron, folding the octahedron onto a plane, and mapping the projection onto a unit square. This approach provides a uniform distribution of vector directions while reducing quantization artifacts typical of other spherical coordinate discretization methods [19]. The advantages of octahedral encoding include uniform value distribution, low computational cost for both encoding and decoding, and efficient storage in resource-constrained environments such as embedded sensor systems.

The proposed octahedral encoding technique is characterized by low computational complexity, making it well suited for implementation on systems with limited resources. Specifically, the encoding process requires only two reciprocal operations (rcp, i.e., 1/x), which can be efficiently executed using hardware arithmetic units available in most modern microcontrollers.

A key benefit of this method is the separation of the acceleration vector's magnitude from its orientation, allowing each component to be processed and compressed independently. Unlike some other spatial vector transformation methods, octahedral encoding does not increase the number of data channels and thus does not complicate the structure of transmitted packets.

To perform octahedral transformation, the following algorithm is used:

1) Vector magnitude calculation. Let $\vec{g} = (x, y, z)$ *be the acceleration vector. First, its Euclidean norm (magnitude) is calculated:*

$$\left|\vec{g}\right| = \sqrt{x^2 + y^2 + z^2}\ . \tag{1}$$

Alternatively, to reduce computational complexity, the Manhattan norm can be used as an approximation:

$$S = |x| + |y| + |z|$$

2) Vector normalization. The vector is normalized to unit length:

$$\vec{g}_{norm} = \left(\frac{x}{\left|\vec{g}\right|}, \frac{y}{\left|\vec{g}\right|}, \frac{z}{\left|\vec{g}\right|} \right) \tag{2}$$

3) Projection to a plane (dimensionality reduction). Since the vector is normalized, its orientation can be represented in two dimensions by projecting it onto a plane using octahedral transformation:

$$u = \frac{x}{S},\ v = \frac{y}{S} \tag{3}$$

4) Reflection to preserve direction (sign of z). If the normalized component is negative z < 0, reflection is applied to ensure the orientation can be uniquely restored:

$$u' = \left(1 - |v|\right) \cdot \text{sign}(u),\ v' = \left(1 - |u|\right) \cdot \text{sign}(v) \tag{4}$$

5) Quantization of the result. The resulting coordinates (u', v') *are scaled and quantized into a discrete* ***N*×*N*** *grid, for example, 256×256 (corresponding to 8-bit representation for each coordinate):*

$$u_q = \text{round}\left(\frac{u'+1}{2} \cdot (\text{N}-1)\right), \tag{5}$$

$$v_q = \text{round}\left(\frac{v'+1}{2} \cdot (\text{N}-1)\right) \tag{6}$$

After encoding, the vector magnitude is stored separately, and the direction is represented as a pair of quantized coordinates . This approach enables a significant reduction in the volume of data for storage and transmission without substantial loss of accuracy.

Experimental evaluation of the effectiveness of this approach was conducted under conditions similar to the previous test — specifically, during simulated human walking at a speed of 3 km/h.

The obtained results demonstrate the following:

1) the total volume of transmitted data after transformation and compression was reduced to approximately 61% of the original (uncompressed) size;

2) the accelerometer data component, after applying octahedral encoding, was compressed to 57% of its original size — a significant improvement compared to the previous value of 84% without transformation.

The results of data compression for various types of user physical activity are presented in Table 1, illustrating how compression efficiency varies depending on the nature of the movement.

TABLE I. SENSOR DATA COMPRESSION RESULTS FOR DIFFERENT ACTIVITY MEASUREMENTS

Activity	Sprinz			Octahedral encoding	
	Acc, %	Gyro, %	Total, %	Acc, %	Total, %
Slow walk	83.8	63.0	74.5	56.3	61.0
Walk	83.6	66.3	76.0	57.7	63.5
Run	95.3	77.9	87.1	66.5	73.3
Sports exercises	97.5	80.6	89.5	70.2	76.4

V. DISCUSSION

Analysis of the obtained results reveals a consistent trend: as the intensity of physical activity increases — from normal walking to dynamic exercise — the amplitude and variability of sensor signals also increase. Consequently, compression efficiency decreases due to greater difficulty in prediction and higher data entropy.

Nevertheless, the application of octahedral encoding as a preprocessing step for accelerometer data enables a substantial reduction in the overall volume of transmitted information. Even in the least favorable scenario, a 29.8% data size reduction was achieved, which is a significant result for resource-constrained systems, particularly embedded and wearable sensor modules.

VI. CONCLUSIONS

The results of this study demonstrate that the application of octahedral encoding significantly reduces the volume of data transmitted from accelerometers in human motion biomechanics monitoring systems. Specifically, under normal walking conditions, the data size was reduced from 84% to 57%, which represents a substantial improvement.

The implementation of both the octahedral transformation and the subsequent lossless compression algorithm (Sprintz) requires no significant additional computational or memory resources from the microcontroller, making the proposed approach well suited for integration into resource-constrained sensor devices. The total volume of transmitted data was reduced to 61–77% depending on the type of physical activity, confirming the method's effectiveness in real-world conditions.

Future research should focus on the exploration of lossy compression methods, which have the potential to achieve even greater data reduction, as well as the implementation of preprocessing and filtering directly on the sensor module. Evaluating the impact of such preprocessing on the efficiency of various compression techniques is also recommended.

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Solving the Optimal Control Problem for the Drying Chamber Model

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***Abstract*— The drying process in industrial chambers is critical for optimizing energy efficiency and product quality, but presents complex control challenges. This study addresses the optimal control problem for a drying chamber model, aiming to minimize energy consumption. A mathematical model of the drying process is developed, and optimal control strategies are derived for specific time intervals. The results demonstrate the potential of advanced control techniques to enhance drying efficiency, providing practical benefits for industrial applications and guiding future research into real-time control implementations.**

Keywords—automatic control system, transfer function, mathematical model, drying chamber, optimization

I. Introduction

Optimal control of drying chambers is critical across industries where drying processes are integral. The drying chamber is a key component in these processes, and controlling it optimally offers many benefits. Optimal control of the drying chamber is **relevant** for achieving energy efficiency, maintaining product quality, optimizing process time, ensuring uniformity, prolonging equipment life, and reducing the environmental impact of drying processes in various industries. The drying process and batch type drying chambers and solving the optimal control problem for the drying chamber using the second order model are considered. The **purpose** of the article is to optimize the drying processes of capillary-porous materials. A batch drying chamber was chosen, which allows better quality and flexibility to set drying modes in order to achieve the most optimal parameters. An analysis of existing technical solutions was carried out this topic. The solution of optimization tasks with the use of the second order model is considered. The optimal solution of this problem in a classical way, using the Hamiltonian was found. According to **the results** of the study, graph of optimal movement trajectory was obtained. The optimal movement trajectory for the control object is three-interval process.

Today, capillary drying is most often carried out in drying chambers. Drying takes place in accordance with all known laws of physics [1]. What the faster the material receives the energy needed to evaporate the liquid, the faster it is dries [2]. Wood drying is the stage where more energy is consumed during wood processing. An incorrect drying process generates cracking problems, rupture of the cellular structure of the wood, folding in the piece [3]. In the industrial drying various drying defects still occur, which include uneven moisture content between and within boards, board distortion, residual drying stresses that may induce internal and external checking, and wood collapse [4]. Natural drying is more time-consuming and does not give the required moisture percentage [5]. Therefore, artificial drying is used. It takes place in batch or continuous chambers. All types of dryers, irrespective of differences in their structure, perform loading, drying, cooling, and unloading operations. In batch dryers, these operations are separated in time and conducted in a specified order [6]. There are several types of drying chambers, but the drying chamber is considered in this work batch dryer chamber. These chambers are much shorter than others and during drying throughout the chamber constant environment parameters are supported. In such a chamber, the material can be dried up to of any required final humidity [7]. At the moment, the equipment is drying cameras are equipped with automatic humidity control systems, ventilation control etc. Such drying chambers use 20% less energy.

II. Overview of existing technical solutions

PI and PID controllers were one of the first regulators. They have proven their effectiveness in managing various processes. The use of such controllers does not require knowledge of the exact process model, so they are effective in managing industrial processes, the mathematical models of which are quite complex. PI and PID controllers are built on the basis of classical control theory and are easy to understand [8]. Establishing relationships between parameters and managing system actions can be carried out by practical engineers and operators. In addition, several tuning methods for PI and PID controllers have been developed over the past decades. A more modern method of automatic adjustment of PI and PID controllers was proposed by Nishikawa in [9]. It requires an input test signal to evaluate process parameters. Optimal values of PI and PID parameters are achieved by minimizing the weighted integral of the quadratic error. However, along with the above advantages, PI and PID controllers have a number of disadvantages. Yes, if the process operating point changes, the controller parameters must be manually readjusted to obtain the new optimal setting. Adjustment must be performed by an experienced operator. For systems with interacting loops, this procedure can be complex and time-consuming. And even if an energy-optimal tracking controller exists, it is only as efficient as the reference trajectory that it is tracking [10]. PI and PID controllers can be effectively applied to optimal control by tuning their parameters to minimize a cost function that balances error, control effort, and dynamic performance. PID is better for dynamic systems requiring fast response and stability. For practical implementation, we have to model the system, define a cost function, and use simulation or optimization tools to find the best gains. PID controllers can be used as a low-level controller within an MPC framework, where MPC computes optimal setpoints or gains over a horizon.

One of the modern approaches to the analysis and synthesis of control systems based on mathematical optimization methods is the theory of control of dynamic objects using predictive models. In [11] and [12], the algorithm of this method is very well described, but the

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implementation of the method is not shown on specific examples, including in control systems.

In design and optimization of drying processes, there is a great need for stable and reliable models to quantify and predict drying rates and drying times with a satisfying accuracy [13].

The steaming process during drying of capillary-porous materials should be optimized using a second-order model. In [14] the solution to the problem optimization with constraints is shown very well. At the same time, the solution of this problem is considered in a classical way, using the Hamiltonian [15].

III. Solving the Optimal Control Problem for the Second-Order Model

The mathematical models in the drying process have been studied under different conditions and configurations i.e., porous media and shrinkage [16]. But as a predictive model for the drying chamber, we will consider a second-order control object with a transfer function [17]:

$$K_2(s) = \frac{4}{3.7s+1} \cdot \frac{42}{1.3s+1} = \frac{168}{4.81s^2+5s+1} \quad (1)$$

A second-order differential equation models the dynamics of key variables, such as temperature or moisture content, with control inputs like heat or airflow rate.

The solution to these problems can be characterized by the Pontryagin's maximum principle [18]. Pontryagin's Maximum Principle is a fundamental tool in optimal control theory used to find the control inputs that optimize a performance index for dynamic systems. It provides a rigorous mathematical approach to optimal control by defining conditions under which a control input optimizes a performance criterion. PMP leverages the system's dynamics to derive analytically grounded control laws, making it well-suited for the behavior of drying processes. Let's find the optimal energy-saving control for this model using the classical method.

The differential equation corresponding to the transfer function (1) will have the following form:

$$3.7 \cdot 1.3\frac{d^2\theta}{dt^2} + (3.7 + 1.3)\frac{d\theta}{dt} + \theta = 168 \cdot U, \quad (2)$$

where the controlling influence is limited $0 \le U \le 1$. In [19] there are some examples of such problems are given, including extensions of the linear-quadratic controller with tight and possibly state-dependent control constraints, as well as piecewise linear-quadratic constraints. Let's write equation (2) as a system of two equations:

$$\begin{cases} 3.7\frac{d\theta_1}{dt} + \theta_1 = 168 \cdot U; \\ 1.3\frac{d\theta}{dt} + \theta = \theta_1. \end{cases} \quad (3)$$

We reduce the system (3) to its normal form:

$$\begin{cases} \frac{d\theta_1}{dt}_1 = \frac{1}{3,7} \cdot (168 \cdot U - \theta_1) = f_1; \\ \frac{d\theta}{dt} = \frac{1}{1.3} \cdot (\theta_1 - \theta) = f_2. \end{cases} \quad (4)$$

Using (4), let's make the equation for the coupled variables for ψ

$$\begin{cases} \frac{d\psi_1}{dt} = -\left[\frac{\partial f_1}{\partial \theta_1} \cdot \psi_1 + \frac{\partial f_2}{\partial \theta_1} \cdot \psi_2\right] = \\ = \frac{1}{3,7} \cdot \psi_1 - \frac{1}{1,3} \cdot \psi_2; \\ \frac{d\psi_2}{dt} = -\left[\frac{\partial f_1}{\partial \theta} \cdot \psi_1 + \frac{\partial f_2}{\partial \theta} \cdot \psi_2\right] = \\ = \frac{1}{1,3} \cdot \psi_2. \end{cases} \quad (5)$$

Let's find a solution for ψ_1 and ψ_2:

$$\begin{cases} \psi_2 = \mathrm{C}_0 \cdot e^{\frac{-t}{1.3}}; \\ \psi_1 == C_1 \cdot e^{\frac{t}{1.3}} - C_2 \cdot e^{\frac{t}{3.7}}. \end{cases} \quad (6)$$

Let's add the Hamiltonian:

$$H = \psi_1 \cdot \frac{1}{3.7} \cdot (168 \cdot U - \theta_1) + \psi_2 \cdot \frac{1}{1.3} \cdot (\theta_1 - \theta) \quad (7)$$

Considering only the term that depends on U, and taking into account the values of ψ_1 and $\psi_2$we obtain from (6):

$$H' = (C_1 \cdot e^{\frac{t}{1.3}} - C_2 \cdot e^{\frac{t}{3.7}}) \cdot 45.41 \cdot U. \quad (8)$$

In order for the Hamiltonian to have a maximum value as required by the maximum principle [20], it is necessary to observe equality $U = U_{max}$ and ensure that the sign changes U_{max} as many times as it is changed by the function ψ_1. Otherwise, you can write that

$$H' = (C_1 \cdot e^{\frac{t}{1.3}} - C_2 \cdot e^{\frac{t}{3.7}}) \cdot 45.41 \cdot U_{max} \quad (9)$$

that is, control will be determined by the ratio

$$U(t) = \sigma \cdot U_{max}$$

where:

$$\sigma = sign(C_1 \cdot e^{\frac{t}{1.3}} - C_2 \cdot e^{\frac{t}{3.7}}) \quad (10)$$

Since the function $\psi_1(t)$ changes its sign at any values $C_1$and C_2 no more than once per time interval, then the law of change $U(t) = \sigma \cdot U_{max}$ where $\sigma = \pm 1$, and taking into account the restriction $0 \le U \le 1$ is obtained $\sigma = +1{,}0$.

Let's determine the moments of switching t_1 and t_2, and also construct the optimal transitional process. The switching moments are determined by the optimal control function:

$$t_i = f_i(\alpha_1, \ldots, \alpha_n, \overline{x}_0, \overline{x}_n, U_{max}) \quad (11)$$

To find this function, we use the method of joining differential equations with a sign-changing right-hand side, write down the solution of equation (2) on the first interval:

$$\theta(t) = C_0' + C_1' \cdot e^{\frac{-t}{3.7}} + C_2' \cdot e^{\frac{-t}{1.3}}, \quad (12)$$

where $C_0' = 168 \cdot U$, C_1' and C_2' constant integrations on the first control interval.

Using the initial conditions for $t = 0$, $\theta = 0$ and $\dot{\theta} = 0$, and we write down the system of equations for determining the constants of integration:

$$\begin{cases} 168 \cdot U + C_1' + C_2' = 0; \\ -0.27 \cdot C_1' - 0.77 \cdot C_2' = 0. \end{cases} \quad (13)$$

Using Kramer's method, we will find the solution of the system (13)

$$\Delta = \begin{vmatrix} 1 & 1 \\ -0.27 & -0.77 \end{vmatrix} = -0.5, \quad (14)$$

$$\Delta_1 = \begin{vmatrix} -168 \cdot U & 1 \\ 0 & -0.77 \end{vmatrix} = 129.36 \cdot U, \quad (15)$$

Then

$$C_1' = \frac{\Delta_1}{\Delta} = \frac{129.36 \cdot U}{-0.5} = -258.72 \cdot U \quad (16)$$

$$C_2' = \frac{\Delta_2}{\Delta} = \frac{-45.36 \cdot U}{-0.5} = 90.72 \cdot U \quad (17)$$

Therefore, the solution to equation (2) in the first control interval will be written as follows:

$$\theta(t) = 168 \cdot U - 258.72 \cdot U \cdot e^{\frac{-t}{3.7}} + 90.72 \cdot U \cdot e^{\frac{-t}{1.3}} \quad (18)$$

Let's write the solution of the differential equation (2) on the second control interval:

$$\begin{cases} \theta(t) = C_1'' \cdot e^{\frac{-t}{3.7}} + C_2'' \cdot e^{\frac{-t}{1.3}} \\ \dot{\theta}(t) = -0.27 \cdot C_1'' \cdot e^{\frac{-t}{3.7}} - 0.77 \cdot C_2'' \cdot e^{\frac{-t}{1.3}} \end{cases} \quad (19)$$

Taking into account the final conditions at $t = t_2$, $\theta = \theta_n$, $\dot{\theta} = 0$ determine the constant integrations on the second interval according to Kramer's method:

$$\begin{cases} \theta_n = C_1'' \cdot e^{\frac{-t}{3.7}} + C_2'' \cdot e^{\frac{-t}{1.3}} \\ 0 = -0.27 \cdot C_1'' \cdot e^{\frac{-t}{3.7}} - 0.77 \cdot C_2'' \cdot e^{\frac{-t}{1.3}} \end{cases} \quad (20)$$

$$\Delta' = \begin{vmatrix} e^{\frac{-t}{3.7}} & e^{\frac{-t}{1.3}} \\ -0.27 \cdot e^{\frac{-t}{3.7}} & -0.77 \cdot e^{\frac{-t}{1.3}} \end{vmatrix} \quad (21)$$

$$\Delta_1' = \begin{vmatrix} \theta_n & e^{\frac{-t}{1.3}} \\ 0 & -0.77 \cdot e^{\frac{-t}{1.3}} \end{vmatrix} \quad (22)$$

$$\Delta_2' = \begin{vmatrix} e^{\frac{-t}{3.7}} & \theta_n \\ -0.27 \cdot e^{\frac{-t}{3.7}} & 0 \end{vmatrix} \quad (23)$$

Then

$$C_1'' = \frac{\Delta_1'}{\Delta'} = 1.54 \cdot \theta_n \cdot e^{0.27t} \quad (24)$$

$$C_2'' = \frac{\Delta_2'}{\Delta'} = -0.54 \cdot \theta_n \cdot e^{0.77t} \quad (25)$$

We will connect solutions at the moment of time t_1:

$$C_0' + C_1' \cdot e^{\frac{-t}{3.7}} + C_2' \cdot e^{\frac{-t}{1.3}} = = C_1'' \cdot e^{\frac{-t}{3.7}} + C_2'' \cdot e^{\frac{-t}{1.3}}$$

$$-0.27 \cdot C_1' \cdot e^{\frac{-t}{3.7}} - 0.77 \cdot C_2' \cdot e^{\frac{-t}{1.3}} = = -0.27 \cdot C_1'' \cdot e^{\frac{-t}{3.7}} - 0.77 \cdot C_2'' \cdot e^{\frac{-t}{1.3}} \quad (26)$$

$$\begin{cases} (C_1' - C_1'') \cdot e^{\frac{-t}{3.7}} + (C_2' - C_2'') \cdot e^{\frac{-t}{1.3}} = -168U \\ 0.27 \cdot (C_1' - C_1'') e^{\frac{-t}{3.7}} + 0{,}77 \cdot (C_2' - C_2'') e^{\frac{-t}{1.3}} = 0 \end{cases} \quad (27)$$

Including the constant integrations found earlier, we obtain a system of equations for determining the switching moments:

$$\begin{cases} \frac{\theta_n}{168 \cdot U} e^{0.27t_2} - e^{0.27t_1} + 1 = 0 \\ \frac{\theta_n}{168 \cdot U} e^{0.77t_2} - e^{0.77t_1} + 1 = 0 \end{cases} \quad (28)$$

According to the (28) the moments of time for various are calculated and listed in Table 1.

TABLE I. THE MOMENTS OF TIME FOR VARIOUS

θ_n	t_1	t_2
10	0.3887	2.2975
20	0.7805	2.5123
30	1.1786	2.7453
40	1.5851	2.9857
50	2.0081	3.2712
60	2.4483	3.5714
70	2.9127	3.9042
80	3.4085	4.2749
90	3.9456	4.6925
100	4.5883	5.2133
110	5.2057	5.7339

The control algorithm and transition process for $\theta = 100°$ for the control object (2) are shown in the graph (Fig.1)

As can be seen, the optimal control will be three-interval and can be easily calculated using electronic computing machines. In the future, this model can be used to predict higher-order systems. This requires extending the lumped parameter model to include thermal inertia and moisture diffusion dynamics. The model then includes a second-order temperature response (to account for heat accumulation) and a delayed humidity response (to model diffusion delays). The second-order temperature response causes overshoot, and the delayed humidity response causes oscillatory control actions. While PID mitigates some of the problems with derivative action, both methods rely on heuristic tuning, not optimizing energy use in complex systems. Higher-order systems make the model more realistic, but increase control complexity.

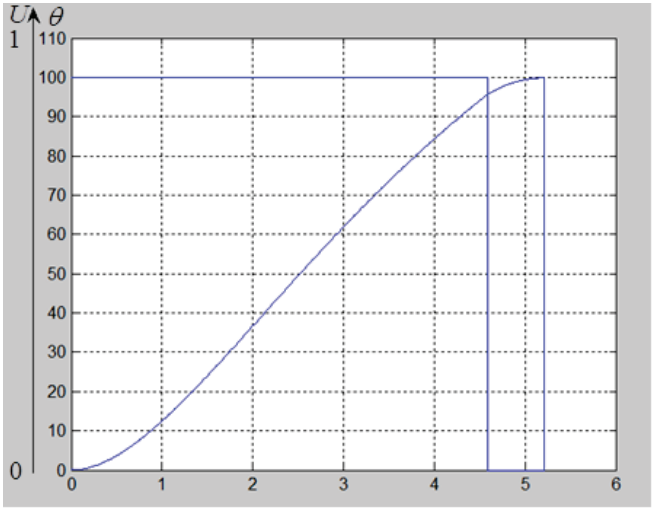


Fig. 1. The optimal movement trajectory for the control object.

Pontryagin's maximum principle for a drying chamber allows finding an optimal control that minimizes the drying time or energy consumption. For simple models, optimal control is often marginal, as we can see from the above, but in

real problems, numerical methods are required due to nonlinearity and additional constraints. The principle is poorly suited for systems with random disturbances, such as non-uniform material moisture in a drying chamber, because it was designed for deterministic systems. This maximum principle guarantees the absence of optimality conditions, not sufficient ones. In nonlinear problems, it is possible to have local optimality that are not globally optimal.

Using AI in modeling drying chambers enables the development of highly accurate second-order models by leveraging machine learning to analyze complex drying dynamics, such as moisture content and temperature variations. Neural networks can be trained on historical and real-time data to predict optimal parameters, accounting for nonlinearities and material-specific properties like those of capillary-porous materials. Current developments in AI offer transformative potential for solving problems in modeling, predicting, and controlling drying chambers dynamics. Future research could improve predictive modeling based on machine learning to predict temperature and humidity trajectories under different conditions. An example of a scheme that can use our model made in this work is shown in Fig.2.

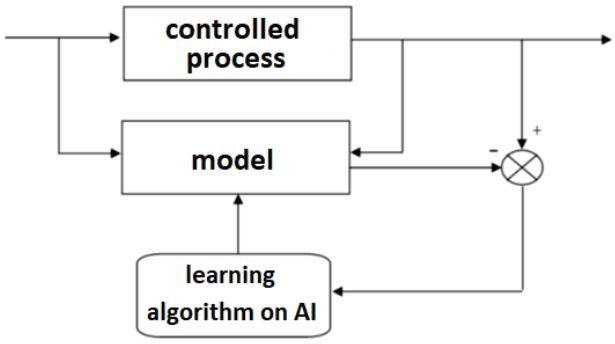


Fig. 2. The scheme for model with AI.

AI could enable adaptive control by processing sensor data in real time, adjusting for disturbances such as material changes or environmental changes.

Furthermore, AI can optimize the second-order model by identifying key parameters through techniques like sensitivity analysis, ensuring robust performance under varying conditions. It also facilitates the real-time adjustment of control inputs by continuously learning from sensor data, enhancing the practicality of implementing optimal control strategies. This capability makes AI a transformative approach for advancing drying chamber research and industrial applications.

CONCLUSIONS

1. Choosing the right method is the most effective and practical tool in solving the problem of optimal management.

2. The optimal control was found, which minimizes the consumption of the heat carrier during the heating stage of the material being dried under the given restrictions on the control influence.

3. Optimal movement trajectories of the regulated quantity for mathematical models of the second order is constructed.

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Image Representation by Atomic Functions: Features for Classification

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***Abstract*—Digital images are widely used in many technologies that are an integral part of various systems. Compared to other types of data, their specific feature is a relatively large size that restricts the use of different machine learning (ML) and computer vision (CV) methods. A large data processing time and image spatial complexity become additional blockers for applying many ML and CV algorithms if real-time results are required or if the main computations must be performed on autonomous systems with limited capabilities. For this reason, the problem of the development of a compact image representation, which contains features useful in ML/CV techniques, is relevant. In this paper, we consider the use of discrete atomic transform (DAT) based on atomic functions for solving this problem. Our previous investigation had proven that DAT provided a combination of data compression and encryption features. This research shows that DAT representation is ML/CV-oriented with respect to image classification using state-of-the-art convolutional neural networks. Thus, its application in different methods for image analysis is promising.**

Keywords—image representation, atomic function, discrete atomic transform, atomic embeddings, classification, convolutional neural networks

I. Introduction

Digital images are widely used in many fields including ecological monitoring [1], precision agriculture [2], urban planning [3], disaster management [4], oceanography [5], climate sciences [6], archeology [7], military and defense [8], etc. Similarly to other types of big data, their specific feature is a large size that leads to huge memory, time, and traffic expenses due to the ability of modern sensors to produce images with tens of millions of pixels. In order to reduce these costs, image compression algorithms are applied [9]. Such methods provide a reduction of the resources required for storing, processing, and transferring images via networks and communication lines. Another significant challenge is the limitations of applying different machine learning (ML) and computer vision (CV) algorithms.

Remote sensing (RS) images are a key component of intelligent systems that implement a wide range of ML/CV techniques [10]. These methods extract features and transform them into smart data. Often, they manipulate raw image data that leads to significant computational expenses required for both analysis and decompression. To perform these computations, different strategies are applied [11]. They involve the usage of graphics and tensor processing units, cloud and distributed computing, etc. However, there exist several significant limitations. First, for autonomous systems and edge devices, special hardware cannot be utilized, especially due to the absence of a required power supply. Next, remote computing increases the load of networks. Furthermore, transferring large amounts of data may not be possible due to the communication channels' bandwidth limitations or interference. For this reason, approaches that leverage machine learning at the edge are adopted [12]. Regardless of the way applied, significant computational resource expenses are challenging. And the problem of their reduction is of particular relevance.

To reduce computation required for image processing and analyzing we propose the use of image representation that combines the following attractive properties:

- Data compression: the decrease of memory expenses required for storing an image.
- Features' compactness: the decomposition of image content into a structured set of small-sized blocks containing features referred to the whole data.

In general, we call a representation ML/CV-oriented if an application of ML/CV methods to a limited number of such blocks provides near the same results as their application to the (full-size) original image. In this context, "near the same results" means minor degradation of methods effectiveness in terms of appropriate metrics.

A lot of data compression methods have been developed recently [11]. Some of them are general-purpose algorithms and can be applied to any data. Moreover, there are specialized techniques that can be used for compressing data of a specific type, in particular, digital images [13]. Generally, they provide better compression. The main goal in the development of both algorithm types was to ensure high memory savings in combination with an acceptable computational complexity. At that time, the majority of analysis algorithms used raw image data [14, 15]. For this reason, a design of ML/CV-oriented representations was not the focus of the research and development.

In this paper, we consider image representation by atomic functions (AFs) [16]. It is based on discrete atomic transform (DAT) that is constructed using the function $up_s(x)$ [17]. It is this combination that provides rapid data expansion and efficient image compression. In [18], it is shown that lossy compression by DAT has a minor impact on the performance

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of a set of state-of-the-art image classifiers. Hence, the following hypothesis is plausible: the image data representation by DAT is ML/CV-oriented. Our aim is to prove this statement with respect to classification using convolutional neural networks (CNNs). We explore image previews produced by DAT in the MobileNetV2, VGG16, VGG19, ResNet50, NASNetMobile, and NASNetLarge [19] classifiers, which constitutes the novelty of this research. We show that, when applying these algorithms to a source image and its preview, nearly the same results are obtained. This is the main contribution.

II. Task Formulation and Solution Methods Choice

Image classification is one of the main tasks of CV [15]. A lot of methods for solving this problem has been recently developed. The application of deep CNNs is among them and has shown high performance [19]. In general, a CNN-based classifier takes an input image, extracts high-level features and predicts its class. An input of such models is a matrix of a fixed size that contains raw pixel data. Therefore, in order to apply some model to a given image, the following steps are performed. First, since images are usually stored in a compressed format, a decompression procedure is applied. Second, a decompressed image is resized to meet CNN input size requirements. This implies that additional computational resources are needed. We note that modern cameras provide images with a great number of pixels. So, the preprocessing step, which is applied before data class prediction, might be time consuming, which might be a blocker for applying CNN-based classifiers in real time, especially in embedded devices. Thus, the preprocessing expenses reduction task is relevant. We propose to solve it using image representation by atomic functions.

Fig. 1 describes the scheme of discrete atomic compression (DAC) of 24-bit full color digital images [16, 20]. Its first step performs the RGB-to-YCrCb color space conversion and computes matrices of luma (Y) and chroma components (Cr, Cb). Then, the DAT transform is applied to each of them. Further, elements of the obtained matrices of DAT coefficients are quantized and, then, encoded using a combination of Golomb codes and context adaptive binary arithmetic coding.

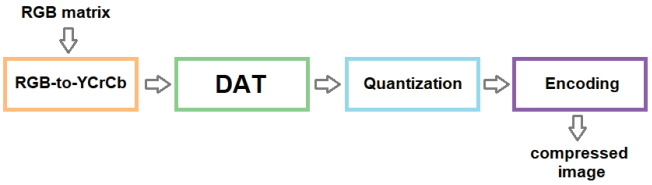


Fig. 1. Discrete atomic compression of 24-bit full color images.

The DAC produces a file with compressed data. We suggest to extract an image preview from this file without a complete decompression. Further, this preview, which is supposed to be an image of significantly smaller size than a source one, is resized and, then, is passed to a classifier. It is clear that, in general, the computed prediction may not coincide with the prediction obtained for a source. So, the following question naturally raises: on average, what percentage of result similarity is ensured? To answer this question, we consider a set of models that includes the following CNNs: MobileNetV2, VGG16, VGG19, ResNet50, NASNetMobile, and NASNetLarge. Currently, we use models with trained weights from the TensorFlow Keras collection [21]. These models implement different architectures. Table 1 illustrates their main parameters.

TABLE I. Models' Characteristics

Model	Depth	Number of Parameters (millions)
MobileNetV2	55	4.3
VGG16	16	138.4
VGG19	19	143.7
ResNet50	107	25.6
NASNetMobile	389	5.3
NASNetLarge	533	88.9

Further, we apply the considered models to a set of digital images and their previews obtained from DAC files. For each pair, we compare top five predicted classes with the highest probabilities. Moreover, to provide a baseline, we perform the same analysis for a discrete transform based on the Haar wavelet [22, 23].

III. Image Representation by Atomic Functions

DAT is a core of the DAC algorithm. It is based on the following expansion [24]:

$$f(x) = \sum_{k=1}^{m+1} \sum_j \omega_j^{[k]} w_k \left(x - \frac{j}{2^{m+1}}\right),$$

where m is positive integer, $\{w_k(x)\}_{k=1,\dots,m+1}$ is a system of atomic wavelets constructed using the atomic function

$$up_s(x) = \frac{1}{2\pi} \int_{-\infty}^{\infty} e^{itx} \prod_{k=1}^{\infty} \frac{\sin^2\left(st(2s)^{-k}\right)}{s^2 t(2s)^{-k} \sin\left(t(2s)^{-k}\right)} dt, s \in N,$$

and $\Omega = \left\{\omega_j^{[k]}\right\}$ is a system of atomic wavelet coefficients.

If $f(x)$ represents a discrete data $D = (d_1, \dots, d_n)$, then the transform, which maps D to Ω, is called the array DAT.

The matrix DAT is defined as follows. Let M be a given matrix. First, the array DAT is applied to each row of M. Then, it is applied to each column of the matrix obtained at the previous step. An output is a block-structured matrix of DAT coefficients B. Fig.2 illustrates this procedure.

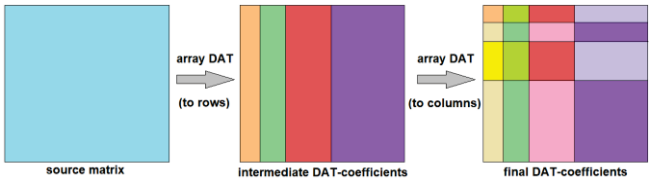


Fig. 2. Discrete atomic transform of a matrix.

By design, B consists of blocks: $B = \left(B_{ij}\right)_{i,j=1}^{m+1}$. Their total number is $(m+1)(m+1)$. Each of them contains DAT coefficients corresponding to a certain spatial frequency band. From the properties of atomic wavelets [24], it follows that elements of the block B_{00} represent an aggregation of the matrix M. In terms of digital images, B_{00} is a preview or small copy of an image given by M. If h and w are, respectively, the height and width of M, then B_{00} is a matrix of the size $\frac{h}{2^m} \times \frac{w}{2^m}$. Blocks of atomic wavelet coefficients constitute image representation by atomic functions.

The matrix DAT has two hyperparameters that specify its settings: s and m. These entities define the applied atomic function and the depth of the wavelet transform, respectively.

In our previous research [20], it has been shown that the DAC based on DAT with $s = 32$ and $m = 5$ provides a combination of high compression ratio with low loss of quality measured by maximum absolute deviation (MAD) and peak signal-to-noise-ratio (PSNR). Moreover, the paper [18] proves that the produced distortions have minor effect on the efficiency of the MobileNetV2, VGG16, VGG19, ResNet50, NASNetMobile, and NASNetLarge classifiers. Here, we use the same hyperparameters.

It is clear that the size of the block B_{00} decreases exponentially with respect to the growth of the depth m. Therefore, an image preview preprocessing, which is needed before an application of CNNs, requires significantly less computations than preprocessing of the corresponding source image. Besides, due to good approximation properties of atomic functions and, hence, atomic wavelets [17], a preview might preserve image features sufficient for classification of a source. Further, we call an image preview produced by the blocks B_{00} atomic. Fig. 3 shows an example.



Fig. 3. Digital photo preview produced by the blocks B_{00} of atomic wavelets. A true size image is given. Preview: 256 × 193 pixels, 144 KB (raw). Source: 8157×6117 pixels, 4.31 MB (JPEG).

IV. Image Classification Using Atomic Preview

To explore the efficiency of image classification by its atomic preview, we consider a set of 130 high quality digital photos given in the JPEG format. The total memory size of JPEG files is 1.35 GB. Raw data require 19.55 GB. Each sample has large height and width. The average number of pixels is 50 million. We compress these data by DAC with settings that provide mean PSNR = 40 dB. We obtain a set of DAC files of the total memory size of 823 MB. The minimum value of PSNR is 35.15 dB. Then, we obtain atomic previews. Fig. 3 shows an example. Next, we apply the MobileNetV2, VGG16, VGG19, ResNet50, NASNetMobile, and NASNetLarge models to each source image and its preview. An output of these models is a vector probabilities (each component represents the likelihood of the analyzed image belonging to a particular class). Further, we compare top five predictions with the highest probabilities for a source and a preview, and compute the similarity, which is the percentage of equal elements. Finally, we evaluate its mean value over all sample images. In addition, we obtain the percentage of cases, when the best predicted classes for a source image and its preview coincide. Also, we perform the same analysis for previews obtained using the Haar wavelet of the depth 5.

Tables 2, 3 present the obtained results. We note that, actually, the aggregated results are given. Detailed results for each image analysis can be found at the following link: https://drive.google.com/drive/folders/14DBoS9JskIgS3yOZ0ftccc7OXzv6VFQo?usp=sharing. The considered images and their atomic previews are available there. Fig. 4 – 7 provide a visualization.

TABLE II. The Average Percentage of Similar Classes, %

Model	Atomic preview	Haar preview
MobileNetV2	84.00	80.92
NASNetLarge	86.00	84.77
NASNetMobile	87.23	87.85
ResNet50	90.15	86.46
VGG16	92.00	90.15
VGG19	92.92	91.69

TABLE III. The Best Class Similarity, %

Model	Atomic preview	Haar preview
MobileNetV2	86.92	76.15
NASNetLarge	83.85	84.62
NASNetMobile	86.92	89.23
ResNet50	90.00	86.15
VGG16	90.77	81.54
VGG19	95.38	87.69

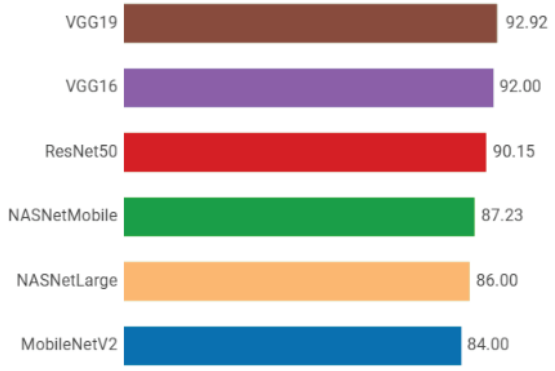


Fig. 4. The average percentage of similar classes: atomic preview.

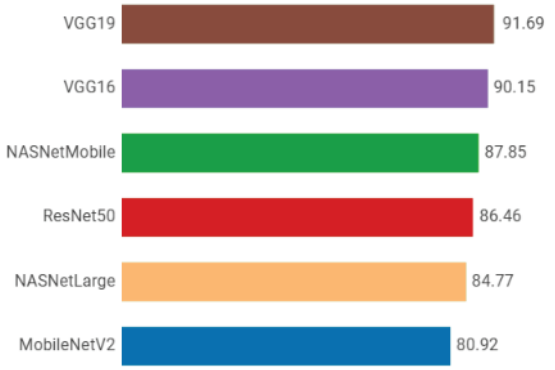


Fig. 5. The average percentage of similar classes: Haar preview.

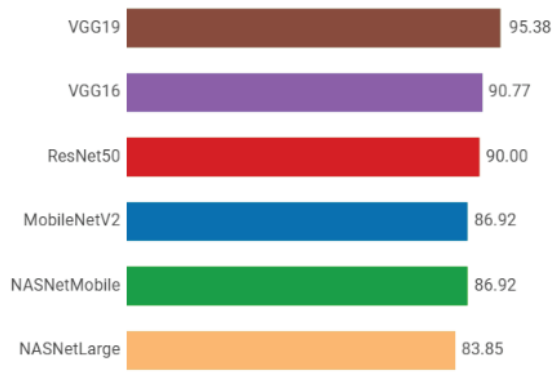


Fig. 6. The best class similarity (%): atomic preview.

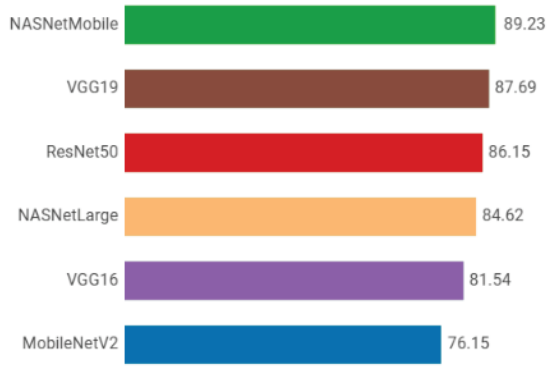


Fig. 7. The best class similarity (%): Haar preview.

In addition, Fig. 8 shows the correlation plot indicating the dependencies between the evaluated metrics and the model's characteristics given in Table 1.

	layers	parameters
similar classes (atomic preview)	-0.47	0.81
best class (atomic preview)	-0.67	0.76
similar classes (Haar preview)	-0.15	0.91
best class (Haar preview)	0.50	0.38

Fig. 8. The correlations between the model's characteristics and the performance metrics.

Considering the mean percentage of similar classes, we see the following:

- The use of both atomic and Haar previews for classifying a source image ensures a high match of top five predicted classes. On average, at least four of them are similar.
- In the most cases, atomic preview provides better results than Haar preview. The only exception is the NASNetMobile model. Although, the difference is minor (see Table 2).
- The worst performance is obtained in the case of MobileNetV2 for both applied previews.
- The classifiers with a greater number of parameters indicate the higher performance (see Fig. 4, 5, and 8). Moreover, the results of the use of atomic preview is less sensitive to this feature.

Next, analyzing the coincidence of predicted classes with the highest probability, we get that

- The application of both previews provides a high percentage of coincidence of the most likelihood class.
- The use of Haar preview performs better, when very deep CNNs are applied (NASNetMobile and NASNetLarge). However, the difference is minor.
- The models with less layers produce better results when one uses the atomic preview. Wherein, the highest difference is indicated for the case of the MobileNetV2 classifier.
- For atomic preview, the impact of the model's characteristics is stronger than for the case of Haar preview.

Finally, comparing the models, we see that VGG19, which has the greatest number of parameters, demonstrates the best performance, when using a preview for classification of a source. And the MobileNetV2 network demonstrates the worst results.

V. Discussion

The performed comparison shows that a set of CNN-based image classifiers provides nearly the same results of the analysis of an image and its atomic preview, which is reconstructed using the blocks B_{00} of DAT coefficients, are close to each other. We remark that the efficiency of the considered models is not evaluated. In fact, we measure how much MobileNetV2, VGG16, VGG19, ResNet50, NASNetMobile, and NASNetLarge change their "opinion" if we replace the original image with a preview. The obtained results show that the difference is not major.

The conducted analysis uses a set of large scale digital photos. The transfer from a source to a preview can be treated as a special resizing technique. It is obvious that this procedure produces the loss of data. Nevertheless, the obtained results show that the most significant features are preserved. However, for the case of small size images, such a degradation might be great. Moreover, the same effect is expected if DAT's depth is increased. We suggest to use such a value of the hyperparameter m, which provides atomic preview of the size close to the input size of the model that is planned to be used.

In this research, neural networks of different architectures are explored. Generally, they demonstrate high robustness to the replacement of a source image with its preview constructed using DAT and Haar wavelet coefficients. The applied function systems have non-similar fundamental features, in particular, the order of smoothness, compactness, and approximation properties [17, 22]. For this reason, the same results could be obtained for other wavelet systems.

Further, the comparison of the predictions provided by the considered models is carried out in the unsupervised manner. Indeed, the actual image classes are not utilized, when computing numerical indicators. An advantage of this approach is that it does not need labeled data, which makes it useful in a wide range of cases [15]. However, a significant drawback of the applied technique is the lack of ability to assess whether classification performance has improved or not. Meanwhile, it follows that the number of parameters has a greater impact on the studied metrics than the depth of a network.

Finally, the obtained results shows that image representation by atomic function is ML/CV-oriented with respect to image classification by CNNs. Indeed, an atomic preview, which is reconstructed using a single block of DAT coefficients, ensures classification of the whole image, i.e. the features' compactness property is also guaranteed. Besides, DAC files requires less memory than both JPEG files and raw data, and, therefore, the data compression property is provided. Moreover, the computed value of PSNR indicates that visually lossless compression is ensured.

In Fig. 9, a general image classification pipeline, which utilizes atomic embeddings, is shown. It could be deployed completely on edge devices or autonomous systems. If computational capabilities are limited, then such light-weight CNNs as MobileNetV2 should be used. Besides, DAT coefficients, which represent atomic preview of an image, could be transferred via communication networks to remote data centers. This could be performed without significant increase in network load, since the size of compressed blocks B_{00} is drastically small.

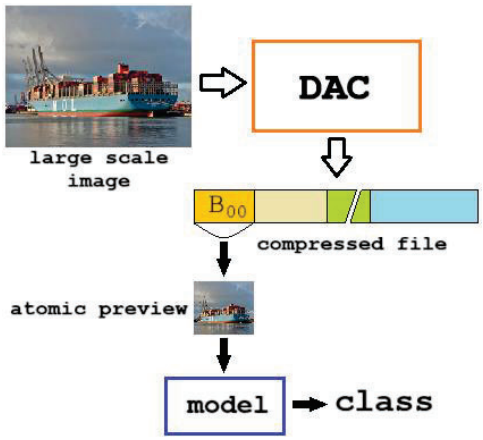


Fig. 9. Large scale image classification pipeline that utilizes atomic embeddings.

VI. Conclusions

This research has shown that image representation by atomic functions is ML/CV-oriented. The use of atomic image preview has been proposed, and its application to classifying source image has been investigated. The performed comparison has demonstrated that a set of state-of-the-art CNN-based classifiers produces nearly the same results, when analyzing either original image or its preview. The evaluated metrics, which measure this similarity in the unsupervised manner, have indicated that DAT is superior to the transform based on the Haar wavelet.

Thus, the use of the DAC algorithm provides data compression feature in combination with image format convenient for classification. The application of atomic image preview ensures the reduction of computations required for the preprocessing step. This makes the deployment of DAC on edge computing and autonomous systems promising.

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Analysis of an *LCL–T*-Based DC-DC Resonant Converter for Load-Independent High-Voltage Capacitor Charging

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***Abstract*—This paper presents the design and analysis of a high-voltage DC-DC converter based on the *LCL–T* resonant topology for capacitor charging in a combinational waveform generator compliant with IEC 61000-4-5. Traditional charging circuits often suffer from low efficiency, limited output current control, and difficulty in achieving soft-switching under varying load conditions. The proposed converter addresses these issues by ensuring zero-voltage switching, high efficiency, and output current stability regardless of the capacitor's charge level. The converter's behavior under the short-circuit and fully charged states is examined to confirm reliable operation in surge testing scenarios. Analytical design guidelines for the *LCL–T* resonant network are derived to maintain stable performance and reduce sensitivity to sudden load changes. Simulation results validate the converter's ability to deliver fast, dependable charging while protecting the power stage. Importantly, the inverter output current exceeds its nominal value by no more than 30% during transients, ensuring safe and predictable operation in pulsed power applications.**

***Index Terms*—high-voltage capacitor charging, *LCL–T* resonant network, microsecond pulse generator, series-resonant converters**

I. Introduction

Modern electronic equipment should be safeguarded against surges. Lightning strikes, electrostatic discharges, motor power switching, and magnetic or inductive coupling can result in power surges (temporary spikes or transient overvoltage/overcurrent) [1]. Microsecond pulse generators are often used to evaluate the surge immunity of electronic equipment. These generators produce pulses by rapidly discharging a capacitor. Series-resonant converters have been used as power sources for capacitor charging applications in these generators [2]–[5]. However, changes in the output current of a series-resonant converter during operation, influenced by the capacitor's charge level, can cause hard switching modes of the converter transistors or excessive converter output current, especially during rapid capacitor discharge.

Several common capacitor charging techniques exist, such as resistive charging, high-frequency constant current charging, and resonant charging [6]–[10]. Resonant converters are particularly appealing for capacitor charging power supplies because of their zero-voltage and/or zero-current switching operation, high-frequency capability, high efficiency, compact size, and low electromagnetic interference [11]. The *LCL–T* resonant converter topology is especially well-suited for capacitor charging applications, as it provides inherent short-circuit protection at the output [11]–[13]. At the same time, more complex resonant networks are also being studied, since they are highly suitable for high-voltage applications [3].

For high-voltage capacitor charging, converters typically employ either a single-channel configuration, which consists of a high-voltage transformer with a complex rectifier composed of series-connected diodes, or a multi-channel setup, where multiple identical transformer-rectifier modules operate in parallel [2], [3], [5], [9], [14]–[16].

This paper presents a comprehensive design, analysis, and simulation of a high-voltage DC-DC converter based on the *LCL–T* resonant network, developed to ensure load-independent performance when powered by a constant DC source. The converter is intended for capacitor charging in microsecond pulse generators used in surge testing and other high-voltage pulsed power applications. The proposed topology enables zero-voltage switching and stable output current, regardless of the capacitor's charge level, addressing major drawbacks of traditional resonant and hard-switched converters. Analytical design conditions for the resonant circuit are derived to guarantee stable operation under both short-circuit and fully charged capacitor conditions.

II. Overview of IEC 61000-4-5 Surge Immunity Standard

A. Test Levels

The IEC 61000-4-5 standard defines the testing and measurement techniques for surge immunity assessment. It specifies multiple test levels corresponding to different environmental conditions. The table below summarizes the surge voltage levels and their associated meanings.

TABLE I
Test Levels

Level	Open-circuit test voltage $\pm 10\%$ [kV]
1	0.5
2	1.0
3	2.0
4	4.0
X[a]	Custom

[a]X can be any level above, below, or in between the other levels. This level can be specified in the product standard.

Different operating conditions classes require various types of combinational waveform generators (CWGs) during testing. The $10/700 - 5/320$ µs CWG is used to test ports designed for connecting symmetrical communication lines. In all other cases, including ports for power lines and short-distance signal connections, the $1.2/50 - 8/20$ µs CWG is used.

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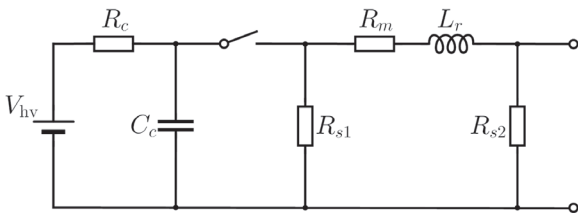
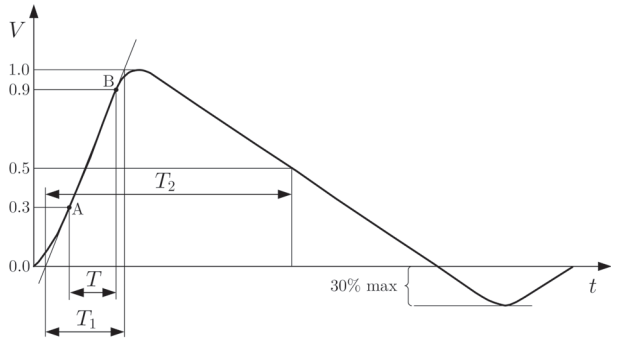
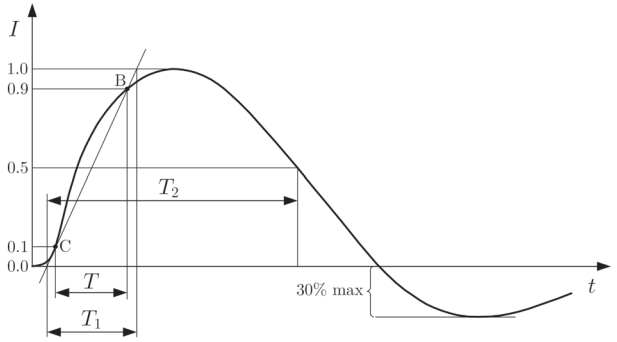


Fig. 1. Simplified combinational waveform generator.



(a)



(b)

Fig. 2. Waveform at the output of the CWG. (a) Open-circuit voltage waveform. (b) Short-circuit current waveform.

B. The Combinational Waveform Generator

The simplified CWG shown in the standard includes a power supply that charges the energy storage capacitor C_c through the resistor R_c (Fig. 1). The capacitor is charged from high-voltage source V_{hv}. When the switch closes, the C_c discharges through a pulse-shaping network consisting of the rise time shaping inductor L_r, two pulse duration shaping resistors R_{s1} and R_{s2}, and an impedance matching resistor R_m. The values of these components are chosen to generate the compliance, voltage and current waveforms under short-circuit and open-circuit conditions, as specified in IEC 61000-4-5.

C. 1.2/50–8/20 μs Combinational Waveform Generator

In existing IEC publications, voltage and current waveforms are defined based on IEC 60060-1 and IEC 60469. Both definitions apply to IEC 61000 and describe a single generator [17]. A standard surge waveform (Fig. 2) is characterized by its front time T_1 and time to half value T_2. Following IEC 60060-1, the open circuit voltage waveform of the $1.2/50$ μs shape has a front time of 1.2 μs ±30% and a time to half value of 50 μs ±20%. And the short-circuit current waveform $8/20$ μs shape has a front time of 8 μs ±20% and a time to half value of 20 μs ±20%.

The effective impedance of CWG, which is defined as the ratio of peak open-circuit output voltage to peak short-circuit current, has to be equal to 2 Ω±10%.

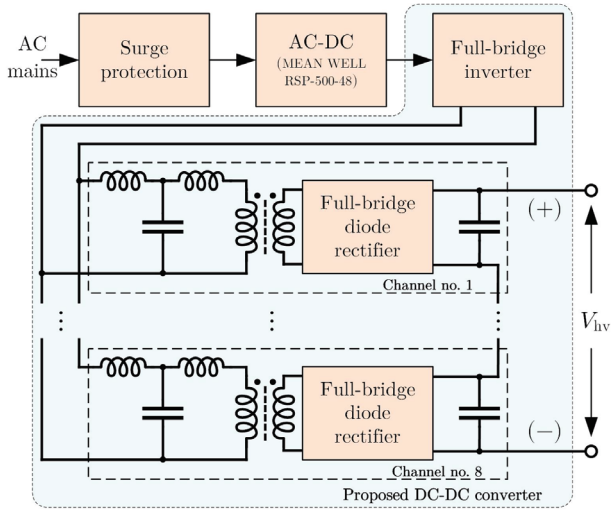


Fig. 3. Power supply system configuration of the 4 kV $1.2/50-8/20$ μs CWG.

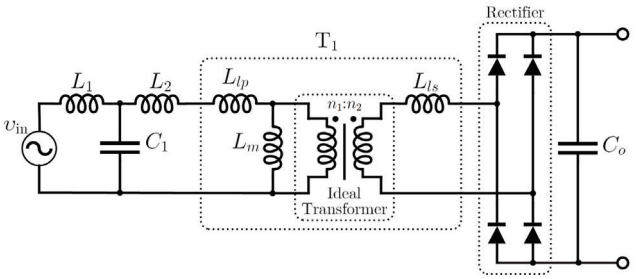


Fig. 4. Simplified circuit of the proposed DC-DC resonant converter.

In [18] and [19], step-by-step methods for determining the values of the parameters C_c, R_{s1}, R_{s2}, R_m, and L_r are shown. However, the values of these parameters can be significantly varied. And one of the criteria for their selection may be the availability of components with the desired nominal values.

III. Proposed DC-DC Converter

Fig. 3 shows the power supply system configuration of the 4 kV, $1.2/50-8/20$ μs CWG, which is based on the DC-DC resonant converter proposed in this paper. The power circuit of the DC-DC converter includes a full-bridge voltage-source inverter, eight channels of *LCL–T* resonant networks with full-bridge rectifiers. The inverter is powered by an AC-DC converter (MEAN WELL RSP-500-48). The power supply has a surge protection block to protect against pulses that may appear at its input as a result of its operation. If full isolation from the AC mains is required, the proposed DC-DC converter can be powered by a 48 V battery.

In the case of generating a 4 kV pulse, the peak output current of the CWG must reach 2 kA. Accordingly, the peak discharge current of capacitor C_c will be even higher. Therefore, the capacitor must be selected appropriately, particularly with a high pulse rise time rating (V/μs). Given that the required capacitance of this capacitor should be 6 μF or more, the proposed DC-DC converter includes eight channels, each containing a capacitor. The outputs of the channels are connected in series. For this purpose, a 60 μF, 900 VDC polypropylene capacitor from ELECTRONICON was selected for each channel. As a result, the total value of C_c is 7.5 μF.

The resistors R_{s1} and R_{s2} were each chosen to be 18.8 Ω and implemented using four series-connected 4.7 Ω resis-

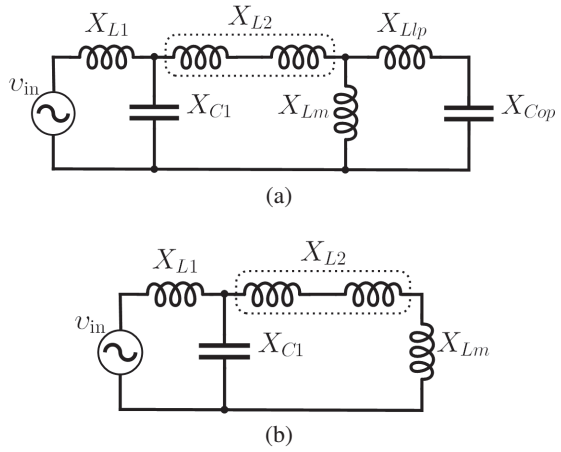


Fig. 5. Equivalent circuit of the proposed converter. (a) Under the short-circuit condition. (b) Under the charged condition.

tors (100 W aluminum-housed wire-wound resistors, model HS100, with a maximum rated voltage of 1900 VDC per element). The resistor R_m was chosen as 0.9 Ω and implemented using a series-parallel connection of 0.1 Ω power film resistors (MP9100-0.20-1%).

The inductor L_r was implemented as two series-connected air-core inductors, each with 15 turns, resulting in a total inductance of 9.5 μH.

A. Circuit Analysis

Fig. 4 shows a simplified circuit of the proposed DC-DC resonant converter. Since all eight channels with the *LCL–T* resonant networks operate in parallel under identical conditions, only a single channel is depicted to simplify the analysis. The *LCL–T* resonant network consists of two inductances L_1 and L_2, as well as the capacitor C_1. The matching transformer T_1 is represented by an equivalent circuit of a two-winding transformer, which includes the primary and secondary leakage inductances, L_{lp} and L_{ls}, as well as the magnetizing inductance L_m. The capacitance of the output capacitor C_o in every channel is 60 μF, as mentioned above.

For the circuit analysis of the proposed converter, it is expedient to consider two main cases: 1) short-circuit condition – when C_o is uncharged and the voltage across it is equal to zero, and 2) charged condition – when the capacitor C_o is charged to the desired voltage level. The first condition corresponds to the initial operation of the converter, as well as the moment when a pulse is generated at the output of the CWG during operation. The second condition corresponds to the state when the voltage across C_o has reached its maximum value, resulting in either no current or a very small current flowing through C_o.

1) Short-circuit condition: Fig. 5 (a) shows the equivalent circuit of the proposed converter under the short-circuit condition, reflecting the secondary parameters to the primary side of T_1.

The electrical impedance Z of the equivalent circuit can be described as follows:

$$Z = X_{L1} + Z_1 \parallel X_{C1} \tag{1}$$

where X_{L1} and X_{C1} are the reactance of L_1 and C_1, respectively, and Z_1 is the impedance, given by

$$Z_1 = X_{L2} + (X_{Cop} + X_{Llp}) \parallel X_{Lm} \tag{2}$$

where X_{L2} is the reactance of sum L_2 and L_{lp}, X_{Lm} is the reactance of L_m, X_{Llp} is the reactance of L_{lp}, and X_{Cop} is the reactance of the capacitor C_{op}, whose capacitance corresponds to that of C_o reflected to the primary side of T_1.

2) Charged condition: In the charged condition, the current flowing through C_o is very small, and the converter's output impedance can be considered very high. The equivalent circuit of the proposed converter for this case is shown in Fig. 5(b). In this case, Z is given by

$$Z = X_{L1} + Z_2 \parallel X_{C1} \tag{3}$$

where Z_2 is the impedance, given by

$$Z_2 = X_{L2} + X_{Lm}. \tag{4}$$

3) Resonant network design: One of the key criteria for selecting parameters of the resonant network is the desired switching frequency f_O of the converter. At this frequency, the impedances Z_1 and Z_2 should be relatively close to each other, so that a sharp change in the voltage across the capacitor C_o, resulting from its discharge, does not significantly affect the inverter's output current. Therefore, the following condition can be used as a design guideline for choosing the resonant network parameters:

$$Z_1 \gtrsim 0.5 Z_2. \tag{5}$$

Considering condition (5), and based on (2) and (3), it can be determined that the reactance X_{L2} should be at least two times greater than X_{Lm}. Assuming that L_{lp} is significantly smaller than L_m, the following condition should be satisfied when selecting L_2:

$$L_2 \gtrsim L_m. \tag{6}$$

To ensure that the proposed converter operates as a current source supply, regardless of the voltage across C_o, the following condition must be satisfied:

$$X_{C1} \ll Z_1. \tag{7}$$

However, if Z_1 is much greater than X_{C1}, the charging process may become too slow, or the amplitude of the inverter's output current may become excessively high. As a compromise, it is reasonable that the reactance X_{C1} at f_O should be approximately 20% of Z_1. Under this assumption, the value of C_1 can be estimated as:

$$C_1 \approx \frac{1.25}{(\pi f_O)^2 L_2}. \tag{8}$$

The value of L_1 should be selected so that the resonant frequency f_r of the *LCL–T* resonant network is slightly lower than f_O.

It should also be noted that Z at f_O, resulting from the selected component values of the *LCL–T* resonant network, should correspond to the desired maximum output current level of the converter's inverter.

B. Characteristic Diagrams of the Proposed Converter and Frequency-Sweep Analysis

Table II presents the selected component values of the *LCL–T* resonant network for the proposed converter.

As a result of manufacturing, the magnetizing inductance L_m of the matching transformer may be too high. In this case, it is advisable to introduce an air gap in the core to reduce this

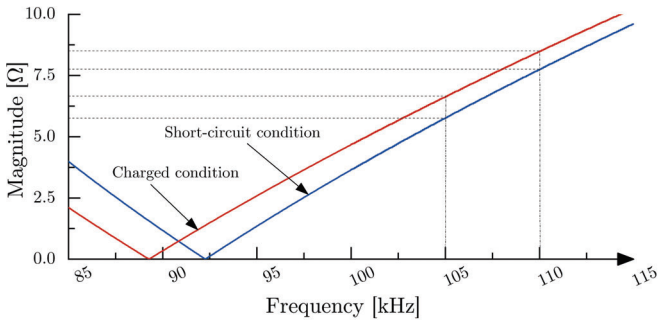


Fig. 6. Frequency response of the impedance magnitude for the selected *LCL–T* resonant network components.

TABLE II
SELECTED COMPONENT VALUES

Item	Value
L_1	33 µH
C_1	0.11 µF
L_2	150 µH
L_m	82 µH
C_{op}	82 µF

value. Alternatively, an additional inductor can be connected in parallel with the primary winding of the transformer. In this case, the effective magnetizing inductance L_m can be considered as the equivalent of the parallel combination of the transformer's inherent magnetizing inductance and the added inductance.

Fig. 6 shows the frequency response of the impedance magnitude $|Z|$ under short-circuit and charged conditions, based on the selected component values of the *LCL–T* resonant network. It can be observed that at the operating frequency range of 105 kHz to 110 kHz, the impedance magnitude varies by approximately 14%. In the case of frequency-sweep operation, where f_O decreases from 105 kHz to 103 kH during the charging process to maintain a more constant charging current for C_o, the system impedance changes more significantly when transitioning from the charged condition to the short-circuit one. Overall, the impedance variation during operations does not exceed 16% even under frequency-sweep operation, which ensures that the proposed converter can effectively operate as a current-source supply.

IV. SIMULATION RESULTS

To verify the theoretical analysis, the proposed DC-DC resonant converter was implemented in the MATLAB/Simulink environment, and its operation was simulated under two conditions: short-circuit and open-circuit (Fig. 7).

To evaluate the system's behavior under the short-circuit condition, waveforms of the inverter output current i_O, high-voltage output V_{hv}, and current i_{out} through the inductor L_r are shown in Fig. 8. The detailed view of the i_{out} waveform demonstrates that the shape of the current pulse fully meets the requirements of IEC 60060-1. At the same time, the waveform of i_O shows that, from the moment the short-circuit condition occurs, its amplitude level exceeds the pre-short-circuit condition amplitude level by approximately 30%. This overshoot is higher than theoretically expected and can be attributed to the appearance of a transient oscillatory process. Over time, the oscillations decay, and the amplitude of i_O approaches the theoretically expected value. This effect should be taken into account during converter design.

Fig. 9 presents current waveforms through the elements of the *LCL–T* resonant network. These waveforms show that although the current i_{L2} through L_2 increases by slightly

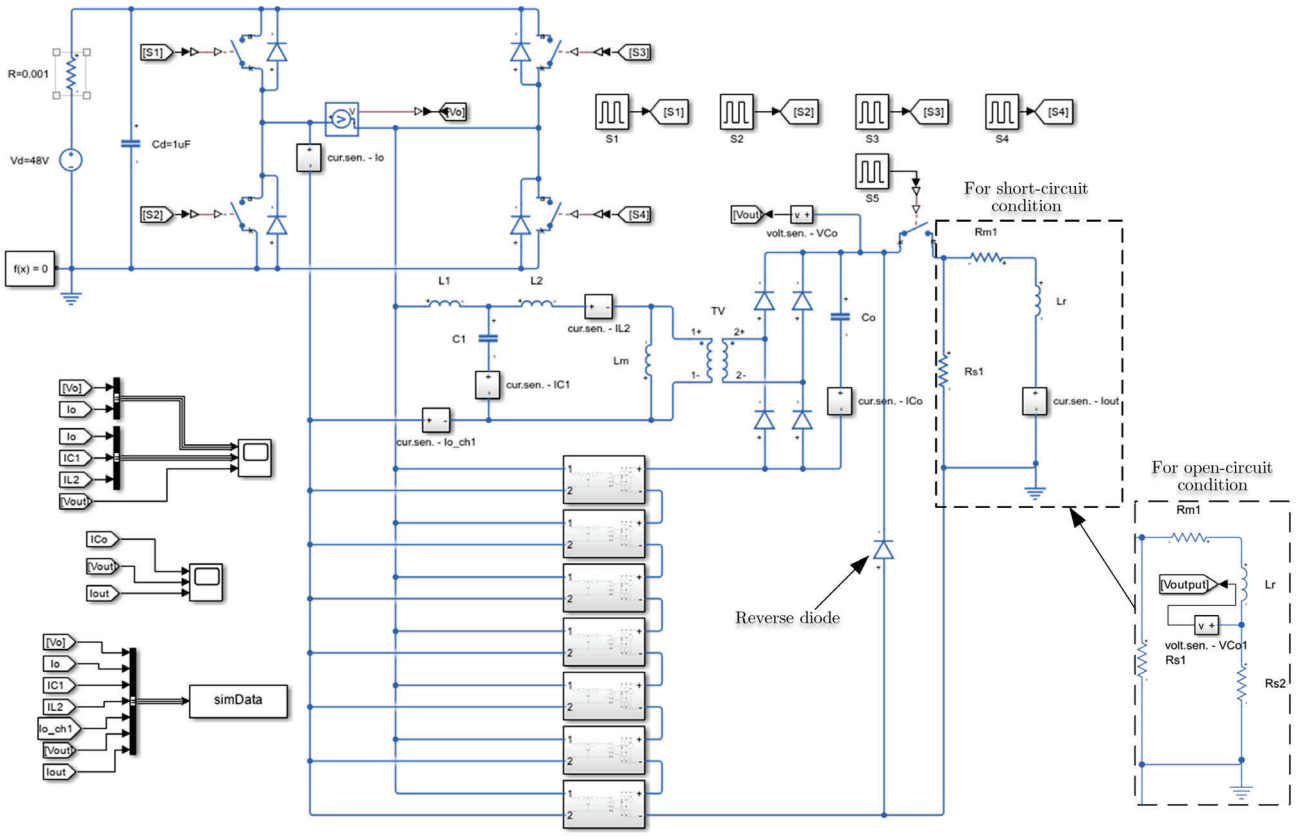


Fig. 7. Converter simulation model in the MATLAB/Simulink environment.

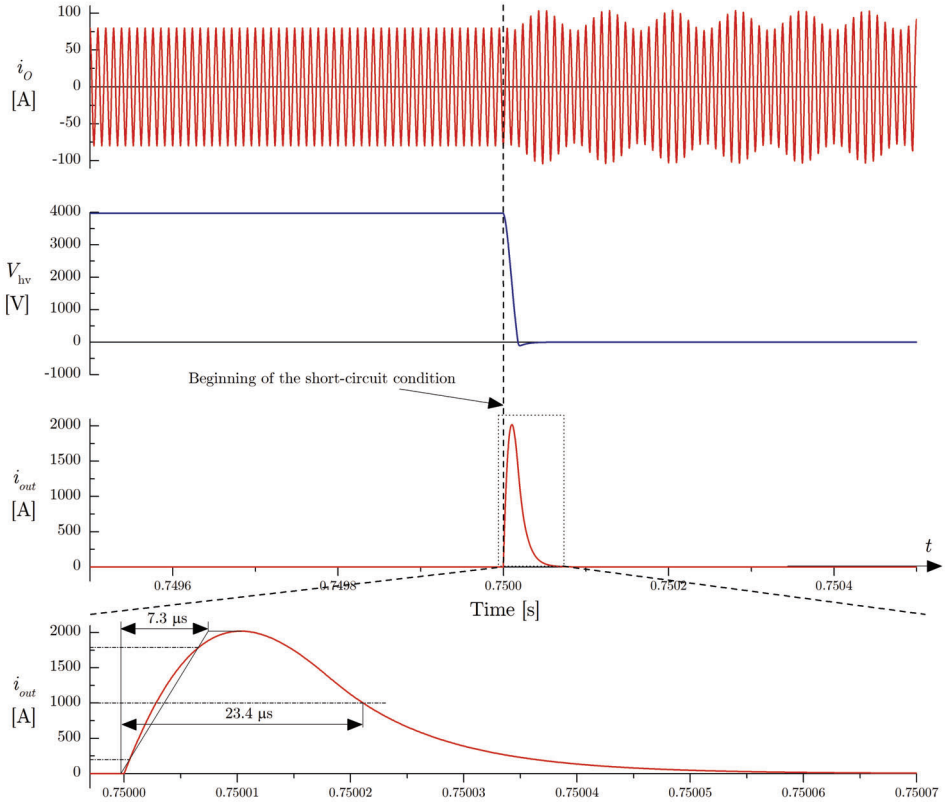


Fig. 8. Waveforms of the inverter output current, high-voltage output and current through the inductor L_r under the short-circuit condition.

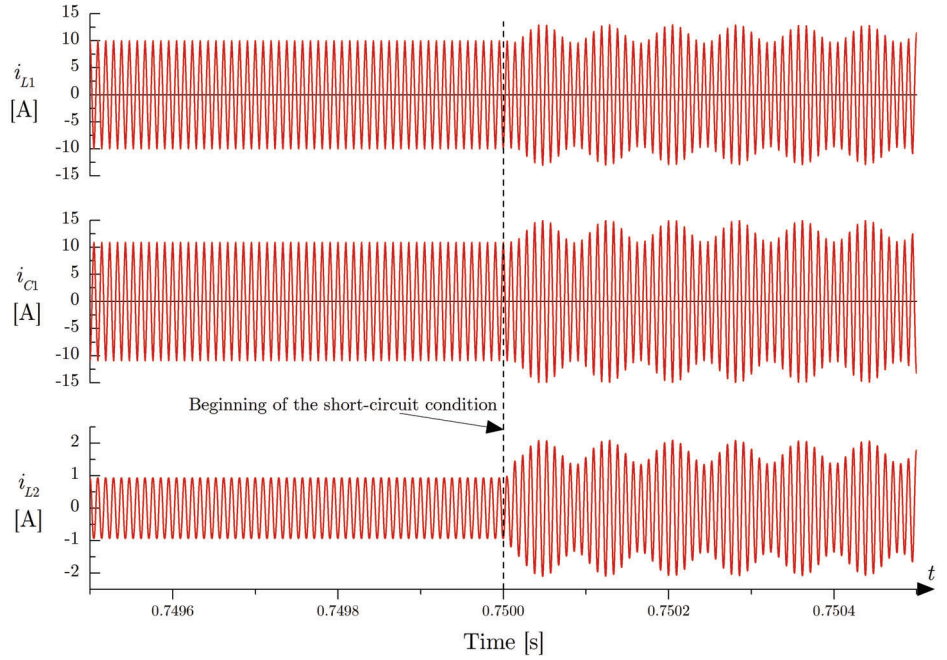


Fig. 9. Current waveforms through elements of the *LCL–T* resonant network in the case of the open-circuit condition.

more than a factor of two after the short-circuit condition, the amplitude of the current i_{L1} through L_1 increases by only about 30%.

Additional insights are provided in Fig. 10, which shows the high-voltage waveform V_{hv} in the absence of the reverse diode shown in Fig. 7. The diagram illustrates that V_{hv} can reach –680 V, caused by the current i_{out} flowing through 16 series-connected diodes of the converter channels. Based on this, the inclusion of the reverse diode is recommended to prevent excessive reverse voltage.

Fig. 11 shows the waveform of the output voltage υ_{output} at the output of the CWG under the open-circuit condition. The voltage υ_{output} corresponds to the voltage across R_{s2}. As seen from the waveform, the pulse shape fully complies with the requirements of IEC 60060-1. It should be noted that the obtained peak value of υ_{output} reaches only 3670 V with a transformer turns ratio of $n_1 : n_2 = 1 : 13$. To achieve 4 kV, this ratio should be increased to $1 : 14$.

Conclusion

This paper presented a high-voltage DC-DC converter based on the *LCL–T* resonant topology for capacitor charg-

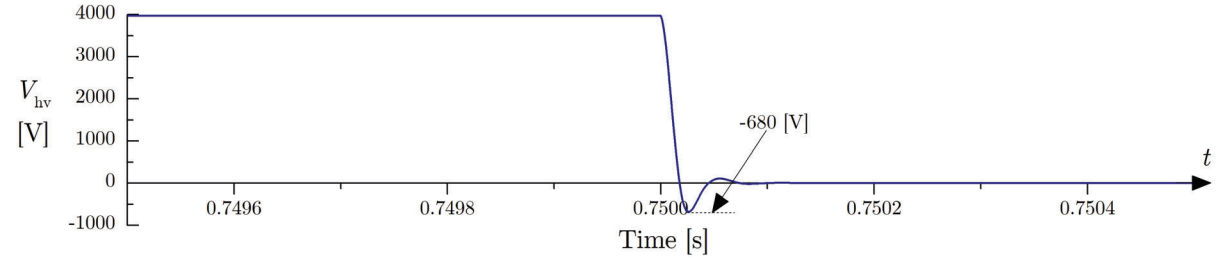


Fig. 10. High-voltage waveform in the case of the absence of the reverse diode.

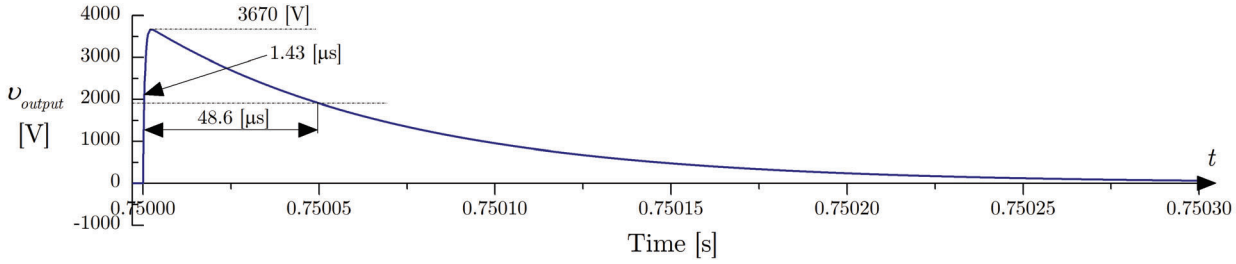


Fig. 11. Voltage waveform at the output of the CWG under the open-circuit condition.

ing in a combinational waveform generator (CWG) compliant with IEC 61000-4-5. The solution ensures zero-voltage switching, high efficiency, and output current stability regardless of the capacitor's charge. The converter's behavior under the short-circuit and fully charged conditions was analyzed, and design criteria for the resonant network were derived to ensure stable operation under varying load impedances. Results confirm that the *LCL–T* topology offers inherent protection and reliable performance for surge testing equipment, with the inverter output current exceeding the nominal value by no more than 30% during transients.

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Improvement of Fuses to Increase the Efficiency of Medium Voltage Instrument Transformers Protection

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***Abstract* — The paper considers an improved design of a fuse for the protection of the medium voltage instrument transformer. The problem of switching off by fuses of small overload currents and the low efficiency of their use for protecting voltage transformers are analyzed. Thermal images of the temperature distribution of the studied fuse are presented and the time and temperature at which its operation occurs are established. The time-current characteristics of the developed fuse are obtained and a comparison is made with samples from well-known manufacturers. The competitiveness of the new fuse design is established and directions and prospects for further development are proposed.**

Keywords* — *medium voltage instrument transformer, fuse, fuse-link, voltage transformer protection, instrument transformer primary winding current

I. INTRODUCTION

As shown in [1–4], in the medium voltage distribution network of 6–35 kV with an isolated neutral, the problem of protecting instrument voltage transformers (VTs) is relevant, which is associated with insufficiently effective level of their protection using fuses [5, 6]. A feature of the operation of VTs is that they operate in a mode close to no-load. The primary winding of the VT, in the circuit of which the fuse is installed, has a very high resistance and, accordingly, the current flowing through the winding is quite small and in the nominal mode is determined by no-load losses, which is approximately 10–20 mA. At the same time, the current Standard [7], and, accordingly, domestic manufacturers of fuses for VT protection [8, 9], do not standardize the values of the rated fuse current I_n, the maximum rated breaking current I_1 and the minimum rated breaking current I_3. At currents less than current I_3, the fuse-links burns out, but the fuse does not interrupt the current (the electric arc does not extinguish) [10, p. 270]. Therefore, the switching off by fuses of namely small overload currents is a problem. And, conversely, in the case of emergency situations accompanied by a significant increase in the flowing current, fuses effectively protect VTs by melting the fuse-links.

The most frequent dangerous situations in a medium-voltage network with an isolated neutral include overvoltages and resonance (ferroresonance) phenomena [11–19]. Their causes are various factors: ground fault; operation of a fuse in one of the phases; lightning surges, etc. [20–23]. But the most common cause of overvoltages is single-phase ground faults due to an alternating arc, as a result of which resonance phenomena and dangerous overvoltages can occur. In this case, the current in the primary winding of the VT can increase several dozen times, which can lead to its destruction [14, 24]. This is especially true for VTs with cast epoxy insulation, which is very sensitive to temperature increase due to current increase. This problem is not so acute in the case of operation of VTs with oil insulation, which, as is known, also performs the function of additional cooling. But, taking into account the disadvantages of oil insulation (large dimensions of the VT, fire hazard, need for periodic maintenance, etc. [25]), the vast majority of manufacturers and operators switch to using VTs with cast epoxy insulation.

From the experience of operating VTs, there are common cases when their protection is provided by such fuses in which the rated current of the fuse-link is significantly higher than the maximum permissible continuous current of the primary winding of the VT [6]. Thus, the designs of fuses require further improvement in order to increase the efficiency of VT protection.

As shown by an analytical review of known solutions in the field of protection of medium voltage VTs using fuses [5, 6, 8, 9, 26–28], today the efficiency of VT protection is not high enough, as evidenced by the increase in the number of their damages. Therefore, research in this area is relevant.

The goal of the paper is to improve the design of a fuse for protection of medium voltage VTs and conduct preliminary experimental studies of the development.

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II. EXPERIMENTAL RESEARCH

A prototype of a fuse-link placed on an experimental stand, manufactured at the Department of Electrical Apparatus of National Technical University «Kharkiv Polytechnic Institute» in cooperation with the manufacturer of electrical equipment «ABM AMPER» LLC, is shown in Fig. 1.

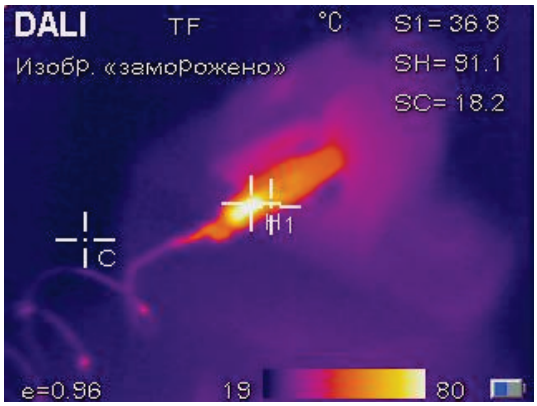


Fig. 1. Stand for experimental research of a fuse

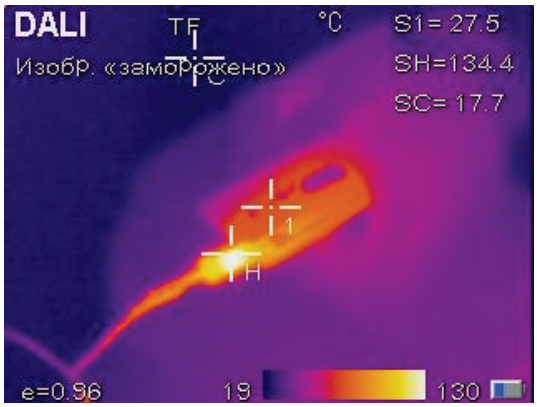
The basis of the manufactured experimental design of the fuse is taken from the developments described in detail in [6, 26], but with new solutions.

A metal bracket is installed on the insulating base of the replaceable fuse-link, to which, in turn, the resistor terminal is attached on one side, and to the other one – the fuse spring in the cocked position. When current flows, the fuse-link heats up, and at a certain value of it, the spring is triggered, switching off the circuit. By using materials with a lower melting temperature in the developed experimental design of the fuse compared to the basic version, it was possible to reduce the fuse operating time precisely in the range of small overload currents. A more detailed description of the design of the fuse and its operating principle is not disclosed, because it constitutes a commercial secret and is the subject of patenting.

In order to determine the temperature of the fuse-link at which the fuse operates, a series of experiments were conducted with thermal imaging of the temperature distribution pattern using a thermal imager. The results of the studies are shown in Fig. 2. Reliable operation of the fuse occurred at current of about 120 mA at temperature of about 180 °C (the ambient temperature during the studies was 18 °C).



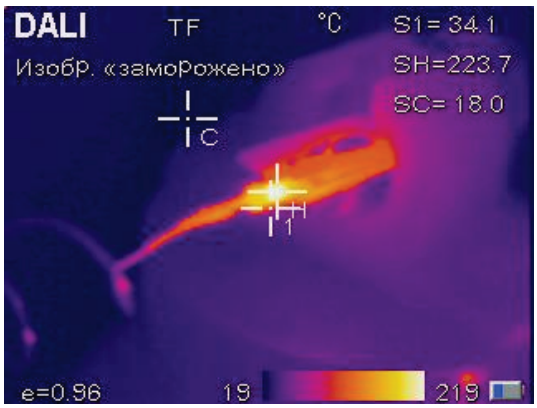
a) I = 80 mA, T = 91 °C



b) I = 100 mA, T = 134 °C



c) I = 120 mA, T = 177 °C



d) I = 140 mA, T = 223 °C

Fig. 2. Results of measuring the temperature of the fuse-link

The next stage of experimental research was to obtain the time-current characteristics for the developed prototype fuse and compare them with the characteristics of common fuses from other manufacturers – ESKA [27] and SIBA [28]. The rated current of all fuses participating in the study was I_n = 80 mA.

A comparison of the time-current dependencies obtained from the research results of the developed prototype fuse with the characteristics of ESKA and SIBA fuses is shown in Fig. 3 and given in Table 1. The data on the average operating time t_{av} in Table 1 were calculated as the average of the results of five experiments conducted.

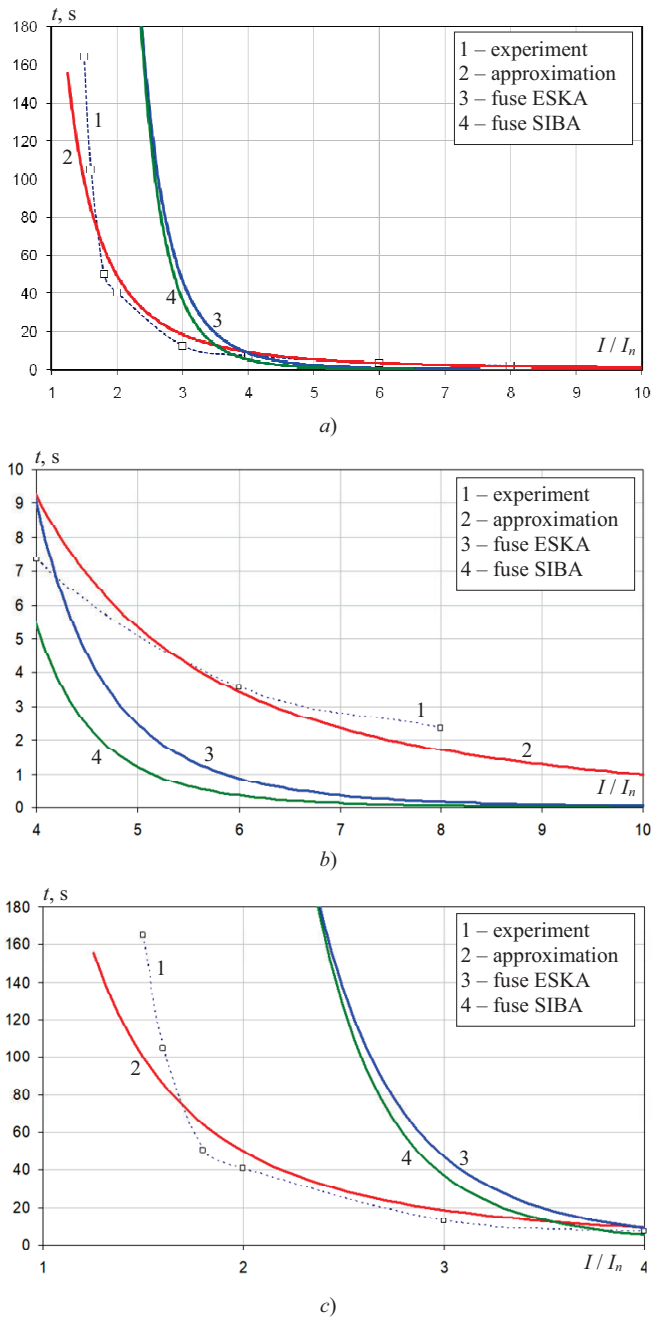


Fig. 3. Comparison of time-current characteristics of the developed fuse with ESKA and SIBA fuses:
a) in the entire range of current multiplicity (I / I_n) – from 1.5 to 10;
b) in the range of large current multiplicity (I / I_n) – from 4 to 10;
c) in the range of small current multiplicity (I / I_n) – from 1.5 to 4

It has been established that the experimental results of the measurement of the fuse operating time can be approximated with sufficient accuracy by the least squares method. The formula for approximating the obtained results of the measurement of the fuse operating time t depending on the current i has the following form:

$$t = 265.8 / i^{2.4} . \quad (1)$$

TABLE I. RESULTS OF EXPERIMENTAL STUDIES OF THE TIME-CURRENT CHARACTERISTICS OF THE DEVELOPED FUSE

I, mA	80	100	120	128	144	160	240	320	480	640
I / I_n	1	1.25	1.5	1.6	1.8	2	3	4	6	8
t_{av}, s	–	–	164.6	105.0	50.4	40.7	12.5	7.3	3.5	2.4

III. ANALYSIS OF THE RESULTS OBTAINED

Analysis of the results of experimental studies shows that the nature of the time-current characteristic for the developed sample in comparison with the ESKA and SIBA fuses is quite close (Fig. 3,*a*). At the same time, when comparing, it is worth noting a certain zonality of the time-current characteristic depending on the current multiplicity. For example, in the range of large emergency overcurrents I / I_n = 4 – 10, the developed fuse is inferior in operating time to foreign manufacturers (Fig. 3,*b*), but with increase in the emergency overcurrent, a decrease in the difference in operating time can be observed. As noted above, fuses cope more effectively with large emergency currents, and the problem is namely the switching off small overload currents. That is why it is necessary to pay attention to the nature of the time-current characteristic of the developed sample namely in the range of small overload currents I / I_n = 1.5 – 4 (Fig. 3,*c*). As we can see, in this range the developed fuse better protects the VT, significantly outperforming foreign manufacturers, and, therefore, is competitive and promising.

In addition, it should be noted that in the range of large emergency overcurrents I / I_n = 4 – 10 (at small rated current I_n = 80 mA these are currents of 320 – 800 mA) the increased operating time of the developed fuse may not be a critical drawback. In case if the thermal resistance of the VT is sufficient (for example, these may be the same VTs with oil insulation), then thermal inertia will protect the fuse from operation during probable short-term current surges, which are common in medium voltage networks.

CONCLUSIONS

Analysis of the current state and experience of operation of electrical equipment in the medium voltage electrical network of 6–35 kV revealed that today there is a pressing problem associated with the low level of efficiency of VT protection by fuses. It was established that there are frequent cases when VT protection is provided by fuses, in which the rated current is significantly higher than the maximum permissible continuous current of the primary winding of the VT.

An improved design of a fuse for protecting medium voltage VTs is considered in the paper and preliminary experimental studies have been conducted.

The obtained research results in the form of time-current characteristics showed the competitiveness of the new design, especially in the range of small overload currents, which is especially important for VT protection.

Since in Ukraine the needs for fuses for VT protection are mainly met by supplies from abroad, a promising direction for further development in this direction is the creation of a domestic competitive design of a fuse and its implementation in production.

Also, as a prospect for further development, it is planned to create a small batch of industrial samples of fuses for VT protection on the basis of «ABM AMPER» LLC in order to conduct in-depth experimental research to identify shortcomings and find the optimal design.

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New Approach to Synthesizing the CdO/por-CdS/CdS Heterostructure

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***Abstract*—This study presents a new approach to synthesizing the CdO/por-CdS/CdS heterostructure to improve solar energy conversion. We achieved a heterostructure with unique morphological and crystalline features by utilizing an anodic and cathodic processing method in an acidic solution followed by oxygen annealing. Scanning Electron Microscopy (SEM) revealed a complex surface topology combining elliptical and circular pores, providing a substantial active surface area. Energy-dispersive X-ray (EDX) analysis confirmed the elemental composition, indicating the presence of CdS and CdO or CdSO phases. Raman spectroscopy and X-ray Diffraction (XRD) allowed us to discern the crystallinity of CdS and the potential amorphous nature of the CdO layer. The intricate design of the heterostructure facilitates efficient light absorption, charge separation, and electron transport, which is critical for solar energy devices. This research contributes to developing photovoltaics and photocatalysis, presenting a heterostructure that leverages the synergistic properties of CdS and CdO for improved solar energy conversion.**

Keywords— CdO, CdS Heterostructures, Electrochemical Synthesis, Cadmium Oxide, Raman Spectroscopy, Semiconductor

I. INTRODUCTION

Recent advancements in nanomaterial studies have illuminated nanomaterials' unique luminescent and vibrational properties, contributing to the continuous development of new applications in renewable energy sectors [1-3]. Furthermore, the impact of defects on the structural and electronic properties of nanostructures emphasizes the importance of controlled synthesis and characterization for utilizing the full potential of nanomaterials in various technological applications [4]. Cadmium sulfide (CdS), with its narrow band gap and sensitivity to visible light, stands out as a quintessential photocatalyst capable of harnessing solar light across a broad wavelength range of 200–600 nm [5, 6]. CdS has attracted significant attention due to its excellent photocatalytic characteristics and high photoelectric conversion efficiency, making it a subject of intensive research [7, 8]. In photocatalysis, CdS demonstrates promising research prospects in water restoration, photocatalytic hydrogen production, and CO_2 reduction [9]. Similarly, CdS shows considerable application potential in photoelectrochemistry in photoelectrochemical (PEC) water splitting [10]. Nevertheless, the broader application of CdS is impeded by its susceptibility to photo corrosion and the rapid recombination of photogenerated electron-hole pairs [11, 12]. It has been proven that modifications of CdS, including morphological engineering, elemental doping, and heterojunction construction, effectively enhance its photocatalytic and PEC properties [13, 14]. The emergence of nanomaterial science has marked a transformative stage in advancing scientific efforts, primarily due to nanostructures' unique capabilities for a wide array of technological and applied objectives [15]. Semiconductor and dielectric nanomaterials, in particular, are pivotal to progress in photovoltaics and optoelectronics [16, 17].

Heterostructural materials have sparked intense scientific interest due to their exceptional properties and immense technological potential [18]. A variety of cadmium-based heterostructures have been studied [19, 20]. However, pursuing optimal heterostructure quality often encounters challenges such as lattice mismatch and the introduction of stress at the material interfaces [21]. One strategy to mitigate these issues involves using an intermediate buffer layer.

Porous layers serving as compliant substrates are increasingly recognized as a promising strategy to alleviate stress and reconcile lattice discrepancies [22]. Numerous methodologies exist for creating such porous architectures. While modern methods like ion-track templating merit attention [23], traditional methods such as electrochemical etching remain relevant due to their simplicity and capability for large-scale synthesis [24].

In this context, our research presents a new methodology for synthesizing the CdO/por-CdS/CdS heterostructure through the electrochemical treatment of the CdS surface. We comprehensively analyze the heterostructure's morphological, compositional, and structural attributes, allowing an in-depth understanding of its intrinsic properties and potential for solar energy conversion applications.

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II. Materials and Methods

The heterostructure was grown on the surface of wurtzite hexagonal CdS using a combined method that included anodic and cathodic processing in an acidic solution followed by oxygen annealing. The mechanism of action involves anodic treatment in the electrolyte, after which the solution is saturated with acid, changing the position of the anode and cathode for cathodic processing. Subsequently, the samples undergo annealing in an oxygen atmosphere for saturation.

The samples were electrochemically treated in a specialized three-electrode cell consisting of two working electrodes and an additional comparison electrode (chloro-argentum EVL1M3). A potentiostat was used for accurate current measurements, which ensured a fixed distance of 1.5 cm between the sample and the platinum electrode. The cell was equipped with additional functional modules, including an air blower and mixer, which facilitated uniform solution mixing during the process.

A. Anodic Etching

The initial stage involved forming a porous layer on the surface of CdS through electrochemical etching. This process was conducted in a 5% aqueous HCl solution at room temperature under potentiostatic conditions, maintaining a voltage of 5V. Group A2B6 semiconductors, such as CdS, are less efficiently etched in acidic solutions than group A3B5 semiconductors, so uniform pore nucleation was facilitated by in situ illumination. This was achieved by focusing UV radiation from a 200W Xe lamp onto the CdS surface. The etching lasted 3 min, during which bubble emission was noted at both the anode and cathode, indicating oxygen and hydrogen release from the electrolyte. Active mixing of the solution was employed to mitigate this effect and ensure effective etching. The electrolyte was at room temperature, and the experiment was conducted under natural lighting conditions. A pronounced increase in current density was observed, signifying the activity of electrochemical processes at the electrolyte-semiconductor interface. Additionally, the change of color of the electrolyte solution to yellow indicates the presence of cadmium ions released from the crystal surface.

B. Cathodic Deposition

The next stage focused on depositing the electrochemical reaction products onto the semiconductor, mainly forming CdO. To implement this step, the semiconductor wafer and platinum were swapped to facilitate cathodic deposition, reversing anodic etching. This process was carried out in the same electrolyte solution under a potential of 5V for 3 min. During this phase, the reduction process of ions at the cathode led to material deposition.

C. Oxygen Annealing

The final step in the manufacturing process was oxygen annealing, which played a crucial role in enhancing the quality and efficiency of the applied layers. Annealing was conducted in a JetFirst diffusion furnace, which was chosen for its precision in temperature control and uniform heating. The samples were annealed at a temperature of 150°C for 20 min. After annealing, the samples were thoroughly dried and stored outdoors for three months, stabilizing the environment naturally.

D. Characterization

Scanning electron microscopy (SEM) and energy-dispersive X-ray spectroscopy (EDX) analyzed surface morphology and composition.

Structural characteristics were investigated through Raman spectroscopy, utilizing a Solver Spectrum spectrometer from NT-MDT, operated in a 180-degree reflection mode. The primary excitation source in this process was a laser, characterized by a wavelength of 473 nm. The size of the laser beam when it made contact with the sample was approximately 2 micrometers in diameter.

III. Results

A. SEM Analysis

Fig. 1 presents the surface morphology of CdS after the electrochemical etching process. SEM analysis reveals that etching led to significant etching pits on the CdS surface, demonstrating two predominant types of visible pores.

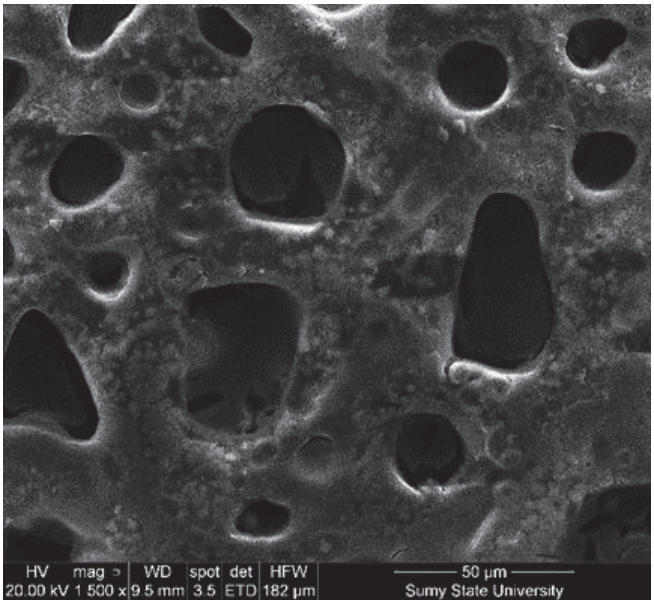


Fig. 1. SEM image of the porous CdS surface.

The first type of pores predominantly exhibits an elliptical shape, with the significant axis measuring approximately 50 µm. These large elliptical etching pits are likely the result of anisotropic etching, where different crystallographic orientations exhibit variable etching rates. Such anisotropy in the etching process can be attributed to the inherent crystalline structure of the CdS material, which has distinct planes of weakness that are more susceptible to the etching solution.

In contrast, the second type of pores is more uniform in shape, displaying a circular cross-section with a diameter of about 20 µm. These circular pores could form from isotropic etching, occurring at the same rate in all directions, leading to more uniform shapes. These pores are relatively smaller than the elliptical ones and suggest a different etching dynamic, possibly influenced by localized variations in current density, impurities, or defects acting as etching sites.

Furthermore, the surface is dotted with numerous tiny pores with diameters not exceeding 1 µm. The distribution of these minute pores suggests a delicate etching process,

possibly due to the prolonged action of the acidic solution or as a culmination of the electrochemical etching towards the end of the process when the more giant pits had already formed.

The diversity in pore sizes and shapes could benefit solar energy conversion applications. It may lead to a larger active surface area and more interfaces, which is advantageous for photon absorption and subsequent charge carrier generation. Moreover, the variation in pore morphology could facilitate the hierarchical assembly of nanostructures within the pores, enhancing light trapping and increasing charge transport efficiency.

Overall, the SEM image illustrates a complex three-dimensional surface topology indicative of meticulous control exerted during the electrochemical etching. This controlled porosity is critical for the subsequent deposition of CdO and the expected enhancement in the optoelectronic properties of the heterostructure.

Fig. 2 presents the morphology of the grown CdO/por-CdS/CdS heterostructure.

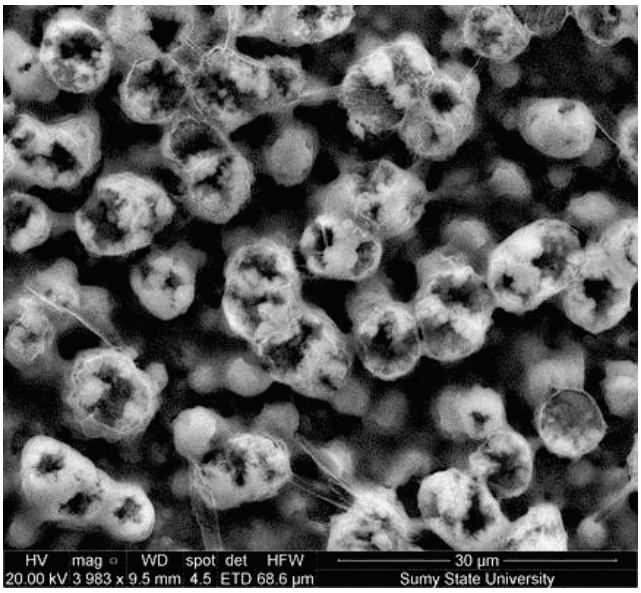


Fig. 2. SEM image of the CdO/por-CdS/CdS heterostructure.

The SEM image allows us to observe white regions, which are massive rounded formations with a loose, porous internal structure. The presence of voids within these globules indicates that material deposition resulted in the formation of not a continuous film but discrete porous clusters. These white globules signify the presence of cadmium oxide (CdO) or potentially a more complex structure, such as cadmium sulfite (CdSO), deposited on top of the initial porous cadmium sulfide (CdS) layer. The dark areas in the image represent voids between the spherical formations. This contrast is due to the differential scattering of electrons, where the electron beam is absorbed or deflected more by denser globules than voids. The mechanism leading to forming these globular structures with internal voids can be attributed to the kinetics of the electrochemical deposition process. During cathodic deposition, cadmium ions are reduced on the cathode surface, forming cadmium oxide. The presence of acid in the solution might lead to the formation of CdSO due to the reaction of cadmium ions with sulfur forms. The loose internal structure of the globules indicates that during the growth process, conditions favored rapid outward expansion of the material rather than uniform layer-by-layer growth. This might be related to localized supersaturation conditions where rapid deposition outpaces the ion migration required for a more compact crystal lattice formation. Furthermore, the stage of oxygen annealing could have influenced the morphology and internal structure of the globules. The annealing process might cause diffusion of atoms within the globules, potentially creating or enlarging internal voids, a phenomenon observed in some annealed metal-oxide systems. From a kinetic perspective, the development of this morphology suggests that a complex interplay of reaction rate and diffusion governs the process. The local microscale environment, including pH variations and local electric fields, can significantly influence the deposition process, leading to the heterogeneous morphology observed in the SEM image.

B. EDX Analysis

EDX spectroscopy was employed to analyze the elemental composition of the synthesized CdO/por-CdS/CdS heterostructure (Fig. 3, Table 1).

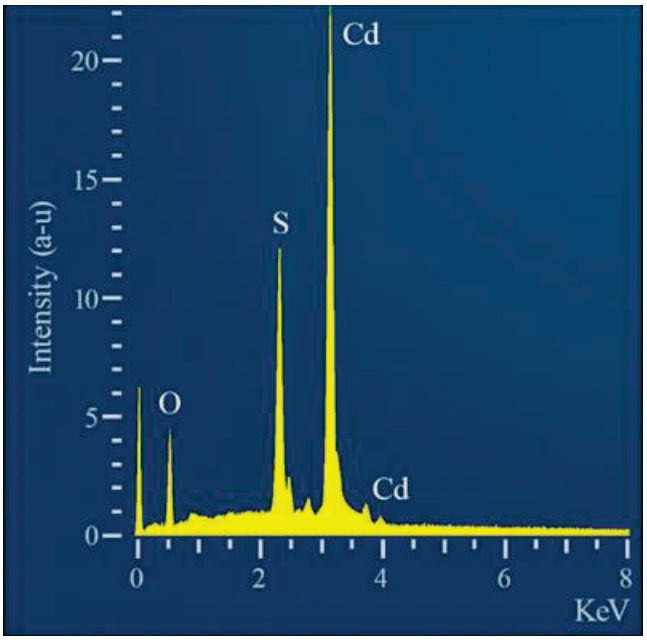


Fig. 3. EDX spectrum of the CdO/por-CdS/CdS heterostructure.

TABLE I. ELEMENTAL COMPOSITION OF THE CDO/POR-CDS/CDS HETEROSTRUCTURE

Element	Atom, %
Cd	31,13
S	31,79
O	37,08

The EDX data reveal a significant presence of oxygen, accounting for 37.08% of the atomic composition. This high percentage of oxygen aligns with cadmium oxide (CdO) formation in the heterostructure, corroborating the SEM observations of white globular masses indicative of CdO or possibly a complex cadmium compound.

The nearly equivalent atomic percentages of cadmium (Cd) and sulfur (S), both approximately 31%, suggest that the CdS layer remains a notable component of the

heterostructure. This balance between cadmium and sulfur atoms is consistent with the stoichiometry of cadmium sulfide (CdS), reflecting the successful preservation of the CdS layer during the synthesis process.

The presence of sulfur in the overall composition also indicates the possibility of cadmium sulfite (CdSO) formation, as the SEM analysis suggested the potential presence of complex structures within the globules. The reaction between cadmium ions and sulfur species in the acidic solution during the cathodic deposition stage might lead to the formation of such a compound.

The EDX analysis thus indicates that the heterostructure comprises a significant amount of oxygen, surpassing the atomic percentages of cadmium and sulfur. This observation might indicate forming a cadmium oxide-rich layer or cadmium compounds with higher oxygen content.

C. Raman Analysis

The synthesized CdO/por-CdS/CdS heterostructure's Raman spectrum displays a series of peaks that shed light on the vibrational modes of the constituent materials (Fig. 4). The spectrum reveals very weak peaks at 115, 148, and 234 cm^{-1}, weak at 254 and 471 cm^{-1}, moderate at 215 and 299 cm^{-1}, and a very strong peak at 602 cm^{-1}.

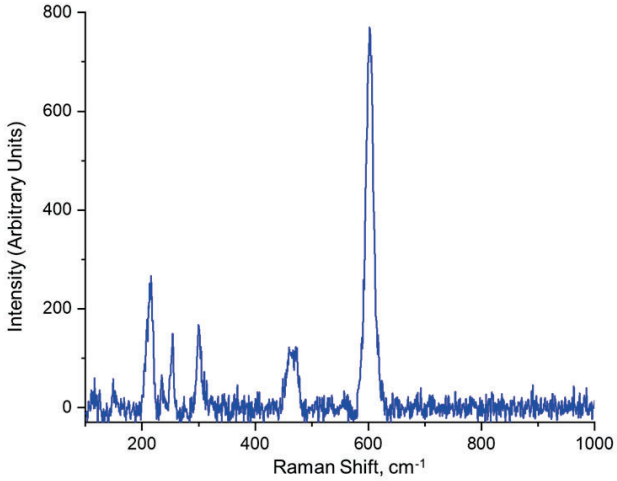


Fig. 4. Raman Spectrum of the CdO/por-CdS/CdS Heterostructure.

The weak peaks at 115, 148, and 234 cm^{-1} could be attributed to low-frequency vibrational modes associated with defects or secondary phases within the heterostructure. These modes are typically not prominent in the Raman spectra of bulk CdS or CdO, suggesting they might originate from surface or interface effects within the porous structure or from a cadmium compound that may have formed during the synthesis process.

The weak peaks at 254 and 471 cm^{-1} indicate the material's intrinsic vibrational modes. The peak at 254 cm^{-1} closely corresponds to the first-order longitudinal optical (LO) phonon mode of cubic CdS, a common feature in its Raman spectrum. While less definitive, the peak at 471 cm^{-1} could potentially relate to second-order vibrational modes of CdS or vibrational modes associated with CdO or complex CdSO phases.

Moderate peaks at 215 and 299 cm^{-1} indicate significant vibrational modes. The peak at 215 cm^{-1} does not align precisely with the typical vibrational modes of CdS or CdO, suggesting it might indicate stress or strain in the crystal lattice or a manifestation of a complex vibrational mode resulting from the heterostructure synthesis. The peak at 299 cm^{-1} aligns more closely with the expected LO phonon mode of cubic CdS, confirming the presence of crystalline CdS in the heterostructure.

The most pronounced feature of the Raman spectrum is the very strong peak at 602 cm^{-1}. This peak is characteristic of the second-order longitudinal optical (2LO) phonon mode of CdS, a harmonic of the 1LO mode. The intensity of this peak points to a significant presence of crystalline CdS within the heterostructure, likely contributing significantly to the vibrational signature. It is essential to consider that a heterostructure and potential interactions between CdO and CdS phases can lead to shifts in peak positions and new modes not seen in the pure phases. Factors such as particle size, crystallinity, and preparation methods could also influence the observed spectrum.

D. XRD Analysis

The X-ray Diffraction (XRD) analysis of the CdO/por-CdS/CdS heterostructure provides crucial insight into the crystalline structure of the present materials (Fig. 5). The diffractogram reveals two distinct peaks: a very high peak at 26.6 degrees, a lower one at 54.5 degrees, and a barely noticeable peak at 60 degrees.

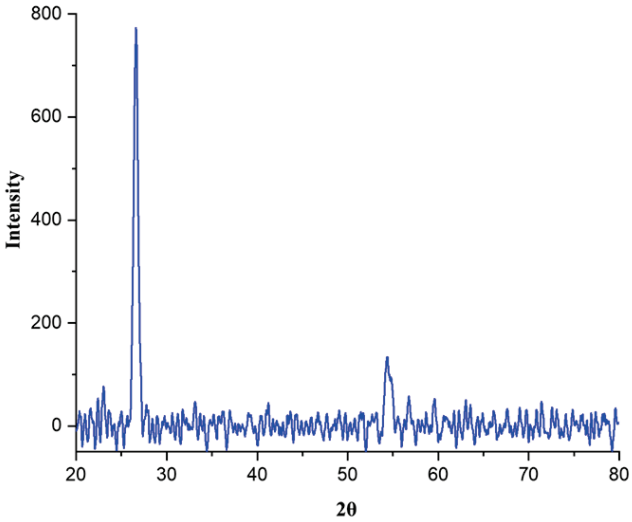


Fig. 5. XRD Spectrum of the CdO/por-CdS/CdS Heterostructure.

The pronounced peak at 26.6 degrees indicates the (002) plane of hexagonal wurtzite CdS. This high-intensity peak suggests a well-ordered crystalline structure and confirms the presence of CdS as a significant phase in the heterostructure. The sharpness of this peak also implies a significant degree of crystallinity within the CdS domains, which is favorable for photoelectric applications due to the efficient charge transport typically facilitated by such crystalline structures.

Although lower in intensity, the peak at 54.5 degrees is likely a higher-order reflection of the same wurtzite CdS, potentially corresponding to the (004) plane. The presence of this peak further corroborates the crystalline nature of the CdS and suggests uniformity in crystal orientation, which is beneficial for the consistency of optoelectronic properties across the heterostructure.

The subtle peak at 60 degrees, while not easily attributable to a specific material phase due to its low intensity, could correspond to a crystalline phase within the CdO or a related oxide compound, possibly indicating the presence of a CdSO phase. However, the low intensity of this peak suggests that if such a phase is present, it is not a principal constituent of the heterostructure or may be poorly crystallized.

Overall, the XRD analysis convincingly suggests that the heterostructure predominantly consists of crystalline wurtzite CdS, with possible minor phases of CdO or CdSO. The crystalline nature of the CdS is advantageous for solar energy conversion, as it typically exhibits high electron mobility and good transport properties, critical for the efficiency of solar cells. The potential presence of other cadmium compounds would also be of interest, as they could introduce additional functionalities to the heterostructure, such as defect states that might aid in charge separation or surface states that could participate in catalytic reactions under illumination.

IV. Discussion

The heterostructure synthesized in our study exhibits a combination of morphological and crystalline features, making it a compelling candidate for solar energy conversion technologies. The presence of a porous CdS layer provides an increased surface area, which is beneficial for absorbing a broader spectrum of solar radiation. As confirmed by our SEM analysis, the porosity of this layer introduces a rough surface texture that can potentially trap light, thereby increasing light absorption efficiency—a critical factor in improving the performance of photovoltaic and photocatalytic systems.

Furthermore, the CdS layer's ability to absorb visible light due to its narrow bandgap makes it particularly attractive for solar energy applications. Efficient absorption of photons in the visible range can generate electron-hole pairs that contribute to generating electric current in solar cells or driving chemical reactions in photocatalytic processes.

The formation of CdO or a complex CdSO structure, as indicated by our Raman and EDX analyses, adds another dimension to the heterostructure's functionality. CdO has a wider bandgap than CdS and can serve as an electron acceptor layer, facilitating charge separation and reducing the recombination of photogenerated carriers. This separation is crucial for increasing the lifetime of charge carriers, thereby enhancing the efficiency of solar energy devices.

The observed strong Raman peak at 602 cm^{-1} confirms the presence of crystalline CdS, known for its high electron mobility. This crystallinity is essential for efficiently transporting charge carriers to the electrodes in photovoltaic cells or to the reactive sites in photocatalytic systems. The X-ray diffraction analysis further supports this, as the sharp peak at 26.6 degrees is characteristic of the (002) plane of hexagonal wurtzite CdS, indicating a well-ordered crystal structure.

Additionally, the heterostructure's morphology—comprised of globular masses with a loose, porous internal structure—may facilitate the diffusion of reactants and products in and out of the material in photocatalytic applications, such as water splitting or CO_2 reduction. This diffusion is crucial for the continuous operation of photocatalytic processes.

In the context of solar cells, the unique structure of the CdO/por-CdS/CdS heterostructure could provide a Z-scheme system where the band alignment promotes a spontaneous flow of electrons from the conduction band of CdS to that of CdO, enhancing the charge separation and transport across the junction. Such a scheme is highly effective in photocatalytic and photoelectrochemical applications, leading to increased quantum efficiency.

Thus, the intricate structure of the CdO/por-CdS/CdS heterostructure, combining the advantageous properties of both CdS and CdO, presents a multifaceted approach to optimizing solar energy conversion. The composition and morphology of the heterostructure are poised to address some of the common challenges in solar energy, such as efficient capture and conversion of sunlight, separation and transport of charge carriers, and catalytic performance in photocatalytic applications. This makes the heterostructure a promising material for the next generation of solar energy conversion devices.

V. Conclusions

The synthesized CdO/por-CdS/CdS heterostructure demonstrates significant potential for applications in solar energy due to its enhanced light absorption and efficient charge carrier dynamics. The electrochemical method used for synthesis is facile and scalable, making it suitable for industrial applications. The heterostructure benefits from the combined properties of CdS and CdO, including a broad absorption range, effective charge separation, and good electron transport capabilities. Analyses such as SEM, EDX, Raman, and XRD collectively indicate the presence of a high-quality heterostructure with favorable morphology and crystallinity for solar energy conversion. Our findings suggest that this heterostructure could be a versatile material for developing high-performance solar cells and photocatalytic systems, offering a pathway toward more efficient solar energy utilization. Future research should focus on device fabrication and performance testing to fully realize the practical applications of this promising material.

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Fabrication and Characterization of a $Cd_xTe_yO_z$/CdS/por-ZnO/ZnO Heterostructure Using Combined Chemical and Electrochemical Methods

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***Abstract*—The quest for effective solar energy conversion has led to the development of innovative photovoltaic materials. This study presents the creation of a new heterostructure, $Cd_xTe_yO_z$/CdS/por-ZnO/ZnO, through an approach that combines chemical deposition and electrochemical etching. Structural and compositional analysis was conducted using Scanning Electron Microscopy (SEM) and Energy Dispersive X-ray Spectroscopy (EDX), revealing a heterostructure with unique morphology and multi-element composition. Raman scattering spectra further elucidated the heterostructure's structural integrity and phase purity, with distinct peaks corresponding to γ-TeO_2 and β-CdO. Structural and compositional analysis by SEM/EDX and Raman spectroscopy confirms the successful formation of a CdxTeyOz/CdS/por-ZnO/ZnO heterostructure with a multilayer architecture and mixed-oxide phases. These results establish a fabrication route and structural baseline for subsequent optical and electronic characterization toward photovoltaic and optoelectronic applications.**

Keywords— photovoltaic materials, heterostructure, $Cd_xTe_yO_z$, CdS, por-ZnO, ZnO, chemical deposition, electrochemical etching, SEM, EDX, Raman spectroscopy, oxygen annealing, solar cells, optoelectronics

I. Introduction

Solar energy, characterized by its sustainability and inexhaustibility, plays a vital role in the evolution of distributed energy networks serving a variety of applications in the civil and defense sectors [1, 2]. The need for efficient conversion of solar energy into electrical energy has spurred research and development of advanced materials and technological methodologies [3, 4]. For example, the deployment of silicon photovoltaics has historically dominated the solar energy sector due to their proven efficiency and scalability [5, 6]. However, environmental and economic issues associated with silicon photovoltaic elements have led to the exploration of alternative materials, such as CdTe [7, 8] and InP [9, 10], Ga_2O_3 [11, 12], which offer comparable efficiency with potentially less life-cycle impact. Meanwhile, developing thin-film technologies and perovskite solar cells represents a paradigm shift towards inexpensive, high-efficiency alternatives to conventional photovoltaic systems [13]. These developments underscore the dynamic landscape of solar energy research, where the search for sustainable, efficient, and economically viable solar cells continues to stimulate innovation in materials science and engineering.

In this context, heterostructures have attracted significant attention due to their unique properties and advantages over binary semiconductors [14, 15]. Heterostructures, consisting of multiple layers of different materials, provide improved charge separation and transport, excellent optical properties, and a tunable band gap [16, 17]. These features not only enhance the efficiency of solar cells but also expand their application for a broader spectrum of photovoltaic and photoelectrochemical applications [18]. The strategic design of heterostructures allows for optimizing light absorption, charge carrier mobility, and stability, thereby addressing some fundamental challenges in solar energy conversion [19].

II. Experiment

A. Substrate Preparation

Hydrothermally grown zinc oxide (ZnO) monocrystals were selected as the substrate material due to their high crystalline quality and suitability for surface modification. We applied surface texturing to enhance the adhesion of subsequent layers and increase the surface area for improved mechanical interlocking. This was achieved through anodic electrochemical etching, creating a specific micro-relief on the ZnO surface. The etching was conducted in a solution containing hydrochloric acid, water, and ethanol in a volume ratio 2:1:1, under a constant voltage of 10 V for 5 minutes. Illumination from an Osram XBO xenon lamp with a power of 250 W, positioned 10 cm away from the semiconductor surface, was applied to enhance the process by facilitating the generation of electron-hole pairs in ZnO, easing the etching. The resulting micro-relief contains terraces, steps, and pores, optimizing the substrate for better layer adhesion, light absorption, and scattering, which is crucial for fabricating a high-efficiency heterostructure.

B. Formation of CdS on the surface of por-ZnO/ZnO

Following substrate preparation, the deposition of the CdS layer was critical for establishing the heterojunction, affecting the optical and electronic properties of the solar cell. In distilled water, a solution was prepared from 0.2 M cadmium

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chloride, 0.2 M thiourea, and 5 M ammonia. This mixture was stirred at 80°C for 20 minutes to ensure uniformity. The textured ZnO substrate was immersed in this solution for 5 hours and optimized to achieve a uniform CdS layer. Subsequent rinsing in distilled water removed residual chemicals, ensuring the purity of the CdS layer.

C. Formation of $Cd_xTe_yO_z$ on the CdS/por-ZnO/ZnO Substrate

For the deposition of $Cd_xTe_yO_z$ oxides on the CdS/ZnO substrate, the successive ionic layer adsorption and reaction (SILAR) method was employed over five cycles. This process allowed for layer-by-layer controlled deposition of materials, achieving a finely structured heterostructure. The SILAR cycle involved immersion in a Na_2TeO_3 solution for the adsorption of tellurium ions, rinsing in H_2O_2 to remove excess ions, immersion in a $Cd(NO_3)_2$ solution for the adsorption of cadmium ions, and a final rinse in H_2O_2 to remove non-adsorbed ions.

D. Oxygen Annealing

After deposition, oxygen annealing was conducted at 150°C for 20 minutes in atmospheric air using a JetFirst diffusion furnace. This step aimed to saturate the layers with oxygen, which is crucial for the passivation of surface defects and enhancement of charge carrier mobility. The additional impact of external conditions over three months tested the natural stability and durability of the heterostructure.

E. Characterization

The morphology and elemental composition of the synthesized structures were analyzed using an SEO-SEM Inspect S50-B scanning electron microscope and energy-dispersive X-ray spectroscopy (EDX) with an AZtecOne spectrometer. Raman scattering spectra were obtained using a DRON-3M spectrometer with 532 nm excitation, a spectral resolution of ~1 cm^{-1}, and a temperature-controlled cryostat system.

III. Results

A. SEM Analysis

Scanning Electron Microscopy analysis revealed a distinct surface morphology characterized by a heterogeneous distribution of elements (Fig. 1). The micrograph contrasts the textured background and separate, darker areas, indicating material deposition or surface features. The dark areas have irregular shapes, ranging from sub-micrometer to several micrometers in diameter. These formations suggest an uneven growth pattern, which may be related to localized conditions during the material synthesis or post-synthesis processing stages.

The background texture consists of stripes and granular elements, corresponding to the micro-relief obtained during the electrochemical etching. The granular aspects of the surface are interspersed with terraces and steps, potentially useful for light absorption and scattering, crucial for solar cell applications.

Furthermore, the shadowing effects observed around the darker areas suggest some topographic dispersion, indicating that these areas are raised above the general plane of the substrate. This relief could potentially affect the light-matter interaction on the surface, a factor of significant interest for optimizing the optical properties of solar cells.

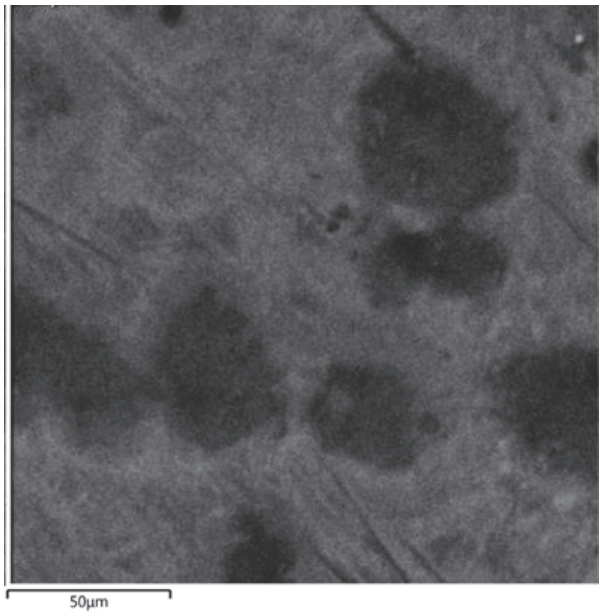


Fig. 1. SEM image of the $Cd_xTe_yO_z$/CdS/por-ZnO/ZnO heterostructure surface.

B. EDX Analysis

Energy-dispersive X-ray Spectroscopy (EDX) provided quantitative insight into the elemental composition of the $Cd_xTe_yO_z$/CdS/por-ZnO/ZnO heterostructure (Table 1, Figure 2). As expected, Tellurium (Te) atoms constitute 21.53% of the heterostructure, indicating significant inclusion in the material, considering its role in forming the $Cd_xTe_yO_z$ matrix. Oxygen (O) is the most abundant element with an atomic percentage of 33.41%, reflecting the oxide components of the heterostructure, as well as the oxygen annealing process, which likely contributed to its prevalence. Sulfur (S), making up 11.65% of the atomic composition, aligns with the presence of the CdS layer, which ideally should form a stoichiometric compound with cadmium. However, the atomic percentage suggests a non-stoichiometric ratio, potentially indicating sulfur vacancies or excess cadmium in the CdS layer.

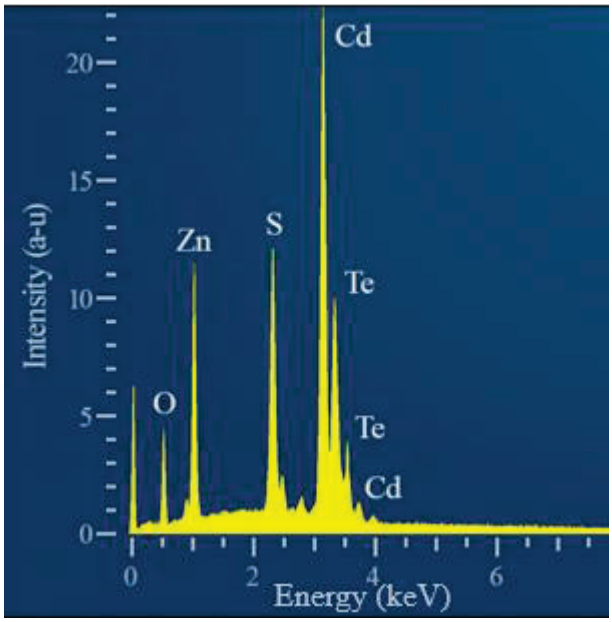


Fig. 2. EDX- spectrum of $Cd_xTe_yO_z$/CdS/por-ZnO/ZnO.

TABLE I. COMPONENT COMPOSITION OF THE OF THE $CD_XTE_YO_Z$/CDS/POR-ZNO/ZNO HETEROSTRUCTURE

Element	Atom, %
Te	21.53
O	33.41
S	11.65
Zn	1.15
Cd	32.26

Zinc (Zn) atoms are present at least 1.15%, which is expected since the ZnO layer serves as the substrate material, not the primary active layer of the heterostructure. Cadmium (Cd) has a significant presence with 32.26% of the atomic composition, expectedly, as it plays a crucial role in both the CdS and $Cd_xTe_yO_z$ layers. The high percentage of Cd underscores its importance in the overall structure, contributing to the electronic properties of the heterojunction.

Thus, the EDX analysis confirms the successful formation of the $Cd_xTe_yO_z$/CdS/por-ZnO/ZnO heterostructure with a complex multi-element composition. The elemental ratios indicate a degree of non-stoichiometry, which may affect the electronic and optical properties of the heterostructure. These findings lay the groundwork for further evaluations of design and efficiency for optimizing the material for photovoltaic applications.

C. Raman Analysis

Raman spectroscopy analysis of the $Cd_xTe_yO_z$/CdS/por-ZnO/ZnO heterostructure provides insights into its structural composition (Fig. 3).

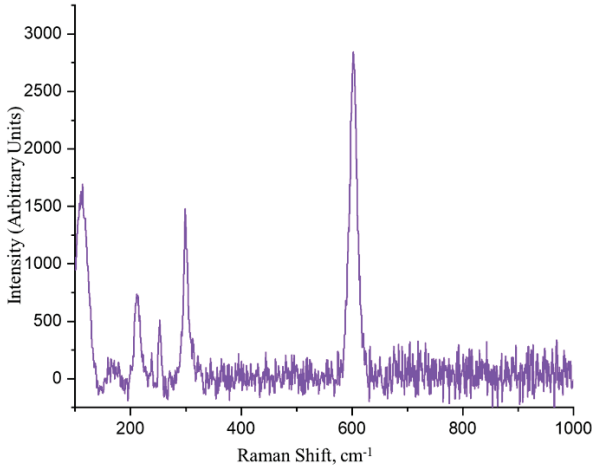


Fig. 3. Raman- spectrum of $Cd_xTe_yO_z$/CdS/por-ZnO/ZnO.

The most pronounced peak at 602 cm^{-1} is primarily attributed to the stretching vibrations of TeO_4 units, characteristic of the γ-TeO_2 phase. This peak indicates a significant content of tellurium oxide, particularly involving TeO_4 self-mot polyhedra within the $Cd_xTe_yO_z$ structure. The intensity and position of this peak suggest the degree of crystallinity and the presence of defects or impurities affecting the TeO_4 structure.

A strong peak at 300 cm^{-1} is closely associated with the γ-TeO_2 phase, representing a combination of vibrational modes. This peak indicates a structured γ-TeO_2 environment, with its intensity and spectral position reflecting this phase's stoichiometry and crystalline quality in $Cd_xTe_yO_z$.

The peak at 254 cm^{-1} corresponds to the transverse optical (TO) mode of β-CdO. This medium-intensity peak indicates the inclusion of cadmium oxide in the structure and sheds light on the nature of the Cd-O bond within the $Cd_xTe_yO_z$ heterostructure.

A medium-intensity peak at 213 cm^{-1} is best attributed to the E (longitudinal optical (LO)) mode in TeO_2, indicating a distinct vibrational mode associated with this specific phase of tellurium oxide in the material. This peak provides insights into the crystallinity and phase purity of the TeO_2 component in $Cd_xTe_yO_z$.

Finally, a weak peak at 112 cm^{-1} is associated with Ag modes involving Te-O-Te bonding. This peak in the spectrum indicates the presence of tellurium in a specific bonding environment, contributing to the structural stability of the material.

The synthesis of the $Cd_xTe_yO_z$/CdS/por-ZnO/ZnO heterostructure, as detailed in this study, reveals a complex mechanism of formation facilitated by a combination of chemical and electrochemical processes. The experimental data underscore the successful formation of a dense oxide layer, showcasing the presence of binary oxides such as TeO_2 and various TeO_x units on the surface. This indicates a multi-stage evolution of the structure where initial formations of CdO and TeO_4 evolve into TeO_2 and TeO_3 as they become saturated with oxygen, which is a critical component in forming the ternary oxides $Cd_xTe_yO_z$. These oxides are integral to the heterostructure, containing only Cd-O and Te-O bonds without direct Cd-Te linkages.

The direct application of the $Cd_xTe_yO_z$ layer onto the ZnO substrate may result in insufficient adhesion due to the weak interaction and incoherence between their crystal lattices. To address this challenge, we introduced a CdS buffer layer, which demonstrated an enhanced affinity and coherency with the lattice structures of the synthesized materials CdO, $CdTeO_3$, and $CdTe_3O_8$. This CdS buffer layer serves as a reliable interface for the subsequent oxide layers and facilitates the formation of a transition layer of CdO, which is essential for the subsequent formation of $Cd_xTe_yO_z$.

The application of the SILAR method was a pivotal aspect of our synthesis process. SILAR operates on the principle of adsorbing ionic particles from a solution onto the substrate surface, triggering a localized chemical reaction that results in the formation of an insoluble product. The selection of the primary precursor is crucial, as the initial thin film acts as the adsorption center for further material deposition. Moreover, the process of washing the sample to remove excess and weakly bonded atoms is a critical step to ensure good film adhesion to the substrate.

The potential applications of heterostructures are broad, ranging from high-efficiency solar cells to optoelectronic devices, where their layered structure and individual properties can lead to improved performance characteristics compared to traditional materials. The ability to manipulate light absorption, charge carrier dynamics, and interfaces at the molecular level offers a significant advantage in developing

next-generation photovoltaic devices. The enhanced light-matter interaction of the heterostructure, facilitated by its unique morphology and composition, may lead to devices with higher photovoltaic conversion efficiency and better stability under various environmental conditions.

Future research prospects include exploring doping strategies to modulate the electrical properties of the heterostructure, as well as incorporating additional layers or interfaces to improve charge separation. Investigating the long-term stability and performance of the heterostructure in real-world operating conditions will also be critical. Furthermore, attention to the scalability of the manufacturing process is needed to assess the commercial viability of the heterostructure for industrial applications. Finally, a comprehensive life cycle assessment would be helpful to ensure that the environmental impact of heterostructure production remains within sustainable limits, aligning with the broader goals of renewable technology advancements.

IV. Conclusions

We fabricated a multilayer CdxTeyOz/CdS/por-ZnO/ZnO heterostructure by combining chemical deposition, electrochemical texturing, and SILAR. SEM/EDX and Raman spectroscopy verify the multilayer composition (Cd, Te, O, S, Zn) and the presence of TeO_2- and CdO-related modes, indicating the formation of mixed tellurium–cadmium oxides on CdS/ZnO. While the present study focuses on synthesis and structural/compositional verification, future work will report the optical (PL/absorption) and electronic (I–V/photocurrent) properties under controlled conditions.

Acknowledgment

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Electricity Consumption in the Green Transition of Steelmaking: A Case Study of OHF/BF vs. Midrex-EAF at Zaporizhstal

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Abstract*— The paper emphasizes the urgent need for Ukraine's metallurgy sector, particularly at PJSC "Zaporizhstal," to replace outdated technologies with modern, sustainable steel production methods as part of a global "green transition". The investigation encompassed field measurements, followed by rigorous data processing and analysis. A comparative analysis was conducted to enhance the findings. The study compares the electricity consumption at two different steel production methods: (I) iron ore sintering process (IOSP) with an aspiration system + blast furnace (BF) + an open-hearth furnace (OHF) smelting; (II) Midrex process + electric arc furnaces (EAF) smelting. The analysis revealed that the specific energy consumption (SEC) varies significantly across different processes: the Electric Arc Furnace (EAF) requires the highest at approximately 340 kWh/t, followed by the Midrex process at 85 kWh/t. The Open Hearth Furnace (OHF) and Blast Furnace (BF) show lower energy demands at around 4 kWh/t and 3.3 kWh/t respectively, while the Sintering Plant (SP) consumes about 33 kWh/t. It is evident that the "IOSP + BF + OHF" method generates substantially higher CO2 emissions than the more efficient "Midrex + EAF" approach. The IOSP is among the most energy-intensive processes, whereas the BF requires less energy. The OHF, though consuming relatively little energy, is less attractive due to its inefficiency. While the Midrex process demands more energy than traditional methods, it compensates with substantial environmental advantages by emitting less CO2. The Electric Arc Furnace (EAF), despite its high energy requirements, excels in the efficient and flexible recycling of steel scrap. Research indicates that adopting advanced technologies like Direct Reduction (DR) via the Midrex process, followed by pellet melting in an EAF, is a crucial step toward lowering emissions and enhancing the energy efficiency of modern steelmaking practices.

Keywords*—green transition in metallurgy, steelmaking methods, specific energy consumption, CO_2 emissions, energy efficiency, Midrex process, electric arc furnaces.

I. Introduction

There is an acute need to transition to green production methods with a low carbon footprint worldwide. This is particularly relevant for the metallurgical industry, which is energy- and carbon-intensive [1]. Global trends toward green steel production methods are being considered [2]. These issues are relevant for many countries and this will also affect Ukraine in the period from 2025 to 2050 [3]. Reforming the steel industry is closely linked to the restructuring of the energy sector for the production of "green" electricity generated by alternative methods (solar [4], bioenergy [5], wind [6], etc.) without the use of hydrocarbons as feedstock. This is especially significant for PJSC "Zaporizhstal," which still uses outdated iron ore sintering process (IOSP) + blast furnace (BF) production and open-hearth furnace (OHF) methods. The project to construct new production facilities based on Midrex and Electric Arc Furnace (EAF) technology will be particularly relevant [7].

The metallurgical industry is diverse and energy-intensive, with various technologies used for steel production. These include the IOSP combined with a BF, OHF, Midrex direct reduction (DR) technology, and EAF, each with distinct methods and energy consumption profiles. The paper [8] presents calculations determining the amount of energy consumption in metallurgical processes in China. It is noted that the comprehensive energy consumption per tonne of crude steel has significantly decreased from 48.2 GJ in 1980 to 16.1 GJ in 2023. Despite this improvement, the total energy consumption on example of China's iron and steel industry has risen from 1.76×10^9 GJ in 1980 to 1.65×10^{10} GJ in 2023 due to the ongoing increase in crude steel production [9]. Theoretical calculations of electricity consumption do not take into account equipment wear [10], the state of the power system and the quality of electricity [11], fluctuations in the characteristics of raw materials and the changing range of smelted iron and steel [12]. Therefore, preference should be given to direct measurements and direct calculations of energy consumption based on the data taken [13].

The objective of this study is to compare these technologies, with a particular emphasis on their energy efficiency and operational parameters. The IOSP, including sintering plants with aspiration systems, and BF are

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conventional and widely used methods. The energy consumption for IOSP is approximately 33 kWh/t, while a blast furnace requires about 3.3 kWh/t. On the other hand, the OHF, which is currently less common due to high energy consumption and a prolonged process, requires approximately 4 kWh/t of steel produced. Modern technologies, such as the Midrex process, which uses hydrogen as a reductant, offer a more environmentally friendly alternative. The energy consumption for Midrex is about 85 kWh/t. The EAF, known for its high efficiency and flexibility, consumes approximately 340 kWh/t. Additionally, modern technologies help reduce CO_2 emissions, which is crucial given the tightening environmental regulations and sustainable development requirements. The evaluation of various steel production techniques encompasses considerations such as maintenance costs, capital expenditures for infrastructure, and environmental impacts. Consequently, this comparison will help pinpoint the most efficient and economically sustainable methods for modern steel manufacturing.

The IOSP and BF, although long-used, are continuously evolving to improve energy efficiency and reduce emissions. Study [12] have shown that optimizing the IOSP and enhancing BF efficiency can significantly impact overall energy consumption and emissions. For example, applying new temperature and charge composition control technologies reduces energy costs and improves the final product's quality. Implementing heat recovery systems and improving aspiration systems also helps reduce harmful emissions into the atmosphere. Research on OHF technology [14] highlights its inefficiency compared to modern technologies, but it is still used in some regions due to existing infrastructure. OHF require significant maintenance and modernization costs, and their energy consumption significantly exceeds modern alternatives, considering hydrocarbon consumption as energy carriers. However, in some countries where industrial modernization is slower, these furnaces remain an essential part of the metallurgical complex.

The Midrex process, especially with hydrogen integration (Midrex), represents a significant breakthrough in reducing CO_2 emissions. Literature data [15] indicates that using hydrogen in the reduction process can reduce the carbon footprint of steel production. Using hydrogen as a reductant instead of traditional natural gas significantly reduces greenhouse gas emissions. Moreover, Midrex technology is actively researched in the context of increasing demand for environmentally friendly technologies and renewable energy sources. However, the economic efficiency of this technology depends on the availability and cost of hydrogen and infrastructural investments. The EAF, despite high energy consumption, also benefits from advances in energy technologies and processes [7–9, 16], making it competitive in regions with abundant electricity. Through the comparison of these technologies, we can identify the most effective and economically viable approaches for contemporary steel production, taking into account energy use, environmental considerations, and operational performance.

II. Research Methodology, Materials and Equipment

The methodology of the study involved a comprehensive and multi-stage approach to analysing the energy consumption and efficiency of different steelmaking technologies, which enabled all key aspects of the processes to be covered: (a) field measurements and data collection; (b) data processing and analysis; and (c) comparative analysis.

(a) Field measurements and data collection. The basis of the study was accurate measurements of energy consumption at PJSC "ZAPORIZHSTAL" enterprise. To ensure a high level of data accuracy and minimize errors, a wide range of modern measuring devices and specialized metering systems was used. In particular, for obsolete technologies, which continue to be used at the enterprise, the work was organized with the help of Elster Alpha A1140-10-RAL-BW-4T meters. These meters, due to their high accuracy and reliability, were integrated into the plant-wide energy metering system "Alpha Center". This system, developed by the company's internal IT department, provides the ability to collect and analyze data in real time. "Alpha Center" allows not only to read metering devices at electrical substations but also to build phase portraits (Fig. 1) of energy consumption, which significantly expands the analysis capabilities. In particular, the system provides the ability to track dynamic changes in network load (Fig. 2, and Fig. 3), which is important for understanding peak power consumption.

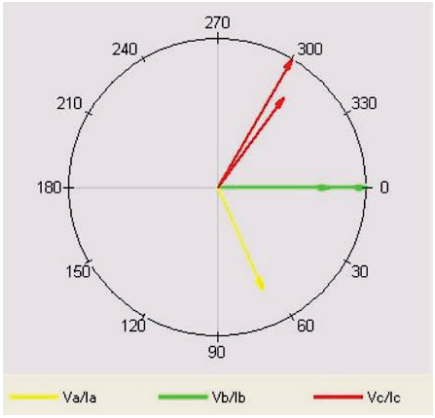


Fig. 1. Picture from the screen of the "Alpha-Center" software by phase portrait of electricity consumption (*Va* and *Ia*, *Vb* and *Ib*, *Vc* and *Ic* – voltage and current of phase "a", "b" and "c" respectively).

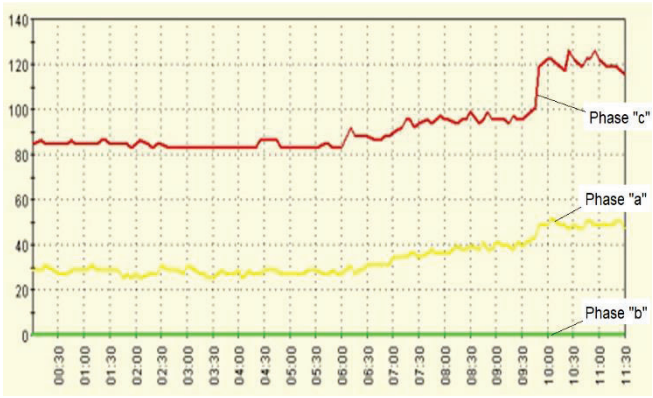


Fig. 2. Hourly (current time, h:min, – horizontally) dynamic monitoring of load changes (active power consumption, kW, – vertically) on the network of the IOSP measured by the "Alpha Center" software.

Data were collected from key substations serving the sintering process and aspiration plants (substations nos. 3, 10, 12, 16, and 19), as well as substations for blast furnaces (substations nos. 21, 30, and 45) and open-hearth furnaces (substations nos. 76, 78, and 51). All the collected data was aggregated in the "Alpha Center," which made it possible to

generate a consolidated report presenting a complete picture of energy consumption at all stages of production. In addition, production indicators were monitored as part of the internal development program of "Steel-Rolling" software, which allowed the energy consumption data to be supplemented with information on product output. The "Steel-Rolling" software is an internal development program aimed at measuring and evaluating production processes, specifically sintering, iron smelting, and steel pouring. The resulting data was generated into Excel reports for further analysis and comparison with reference data.

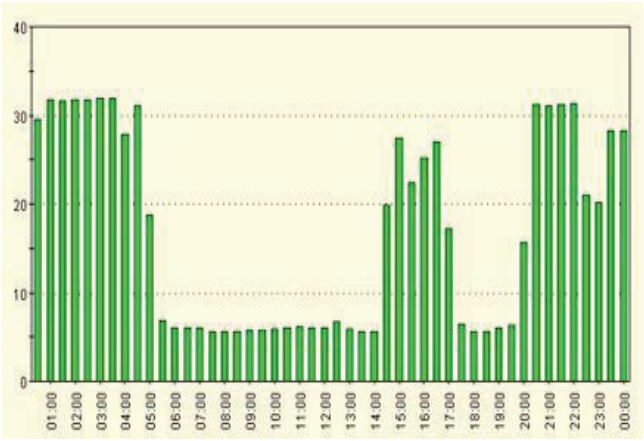


Fig. 3. Hourly (current time, h:min, – horizontally) dynamic monitoring of changes in total active load (total active power, kW, – vertically) on the network of the BF measured by the "Alpha Center" software.

Reference data from equipment manufacturers were used to assess the energy consumption of the new technology, which is planned to be introduced at the enterprise as part of the transition to "green" production. In particular, for the Midrex and EAF technology, data from “Midrex” and “Danieli” companies were provided. These companies, which have extensive experience in equipment development, provided data adapted to the conditions of PJSC "ZAPORIZHSTAL". The data included predicted energy consumption values and allowed for an accurate comparative analysis with the actual data obtained at the plant.

(b) Data processing and analysis. The processing of the measurement data constituted a crucial phase of the study.. During this phase, a thorough review of all the data collected was performed to determine the energy efficiency of each of the steel production methods used. An important aspect of the analysis was the comparison of actual data collected through the "Alpha Center" and "Steel-Rolling" systems with reference data from equipment manufacturers. This enabled an in-depth analysis of the efficiency of current production processes and identified possible reserves for energy efficiency improvements. The analysis took into account the distribution of energy consumption across different process stages, which allowed us to identify the most energy-consuming processes and propose ways to optimize them.

The assessment included a calculation of specific electricity consumption rates per ton of produced products. This calculation was performed using the methodology developed at the enterprise within the framework of ISO 50001 international standard certification, which ensured high accuracy and reliability of the results obtained. The main efficiency indicator was the specific energy consumption rate calculated by the equation:

$$S_a = C_a / P_o, \quad (1)$$

were S_a – actual specific consumption for the technology (kW/t); C_a – actual energy resource consumption (kW); P_o – amount of produced output (t).

(c) Comparative analysis. At the final stage of the study, a comprehensive comparative analysis of two methods of steel production was carried out: the traditional method, which uses outdated technologies, and the new method, which is planned to be introduced as part of the "green" transition. The analysis included an assessment of the methods in terms of their energy efficiency, environmental impact, and economic feasibility. The comparison was based on data obtained during measurements and reference data provided by equipment manufacturers.

The comparison highlighted the key advantages and disadvantages of each production method. Based on the analysis, specific measures to reduce energy consumption were proposed, including optimization of technological processes, equipment modernization, as well as introduction of advanced methods of energy management. As a result of this study, conclusions were made that will help PJSC "ZAPORIZHSTAL" not only to increase energy efficiency of production, but also to strengthen its position in the market by switching to more environmentally friendly technologies, which is especially important in the context of increasing requirements for sustainable development and carbon footprint reduction.

III. Results and Discussion

The combination of the IOSP + BF process with the OHF leads to significantly higher CO_2 emissions compared to the Midrex + EAF technology, as demonstrated in Table 1. Specifically, the IOSP + BF + OHF process produce 2430 kg/t of emissions per ton of steel, while Midrex + EAF generates only 613 kg/t of emissions per ton of steel. This indicates that the Midrex + EAF technology is significantly more environmentally friendly and contributes to reducing the carbon footprint of steel production. The experiments were conducted under real-world conditions, confirming the practical significance of the results obtained. These data demonstrate the importance of implementing modern technologies, such as Midrex, to reduce the environmental impact of steelmaking processes.

Table I. Calculation of Emissions per Ton of Steel

Production type	Emissions (kg/t)
IOSP+BF+OHF	2430
Midrex + EAF	613

Different steel production technologies have significant differences in energy consumption. The IOSP process with aspiration systems is one of the most energy-intensive, consuming about 33 kWh/t of steel (Table 2). This is due to the high costs of raw material preparation and maintaining the sintering process. The BF, typically used alongside the sintering process, requires 3.3 kWh/t of steel. Although this is much lower than IOSP, the overall energy costs remain high. The OHF, though less common today, requires relatively little energy — 4 kWh/t of steel. However, its inefficiency and prolonged process make it less attractive for modern production.

TABLE II. ENERGY INTENSITY OF DIFFERENT IRON AND STEEL PRODUCTION

Technology	Energy Consumption (kWh/t)
IOSP with Aspiration Systems	33
BF	3.3
OHF	4
Midrex	85
EAF	340

The modern Midrex technology, which uses hydrogen as a reductant, demonstrates high energy efficiency with a specific energy consumption of 85 kWh/t of steel. This makes it attractive in terms of environmental cleanliness and emission reductions. The EAF consumes 340 kWh/t of steel. Despite higher energy consumption compared to some other technologies, EAF is flexible in using secondary raw materials, contributing to a circular economy. When studying the IOSP+BF process and OHF technology (Fig. 4), it was found that this outdated method has high CO_2 emissions and significant energy consumption (Table 3). This method has a negative environmental impact due to the large amount of CO_2 emissions. Midrex (Fig. 5) and the EAF (Fig. 6), on the other hand, are modern technologies that show low CO_2 emissions but high energy consumption (Table 4). These methods have higher energy consumption and lower environmental impact, making them attractive for modern production conditions. Practical applications – the studies were conducted under real production conditions, giving them practical significance.

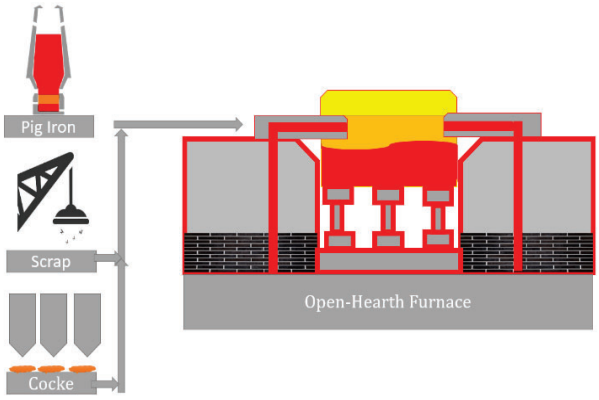


Fig. 4. Open-Hearth Steel Production Process.

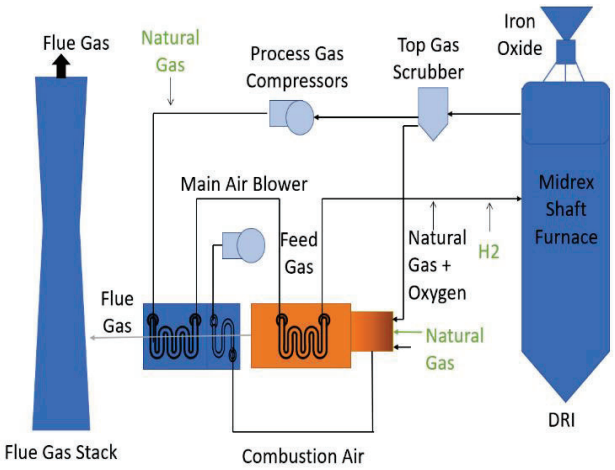


Fig. 5. Diagram of the Midrex process.

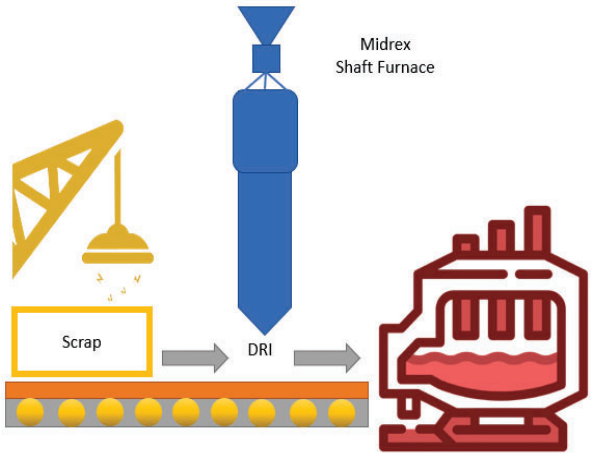


Fig. 6. Diagram of EAF charging process.

Comparison indicates differences in energy consumption (Fig. 7), but each method has its advantages and disadvantages. These research results can be used to make decisions about the optimal steel production method for a specific enterprise, taking into account energy efficiency and environmental requirements. Considering the growing requirements for reducing CO_2 emissions and optimizing energy use (Table 5), modern technologies such as Midrex and EAF have significant potential to increase the competitiveness and sustainability of steel production.

TABLE III. MAJOR ENERGY CONSUMERS IN THE SINTERING-BLAST FURNACE PROCESS AND OPEN-HEARTH FURNACE (FOR IRON ORE AND STEEL)

Process	Energy Consumer	Energy consumption (kWh/t)
IOSP	Crushing and sorting of ore	4.3
	Ore drying	5
	Mechanical mixers	3
	Conveyors and transporters	2.7
	Sintering belts	8
	Dust collectors and gas cleaners	10
BF	Air and oxygen injection	0.6
	Fuel feeding	0.55
	Heating systems	0.5
	Elevators and conveyors for raw materials	0.4
	Cooling systems	0.375
	Control and automation systems	0.475
	Gas cleaning and utilization systems	0.4
OHF	Air and gas supply systems	0.3
	Mechanical equipment for loading and unloading materials	1.3
	Cooling systems	0,8
	Gas cleaning and recirculation systems	0,6
	Control and automation systems	1

TABLE IV. MAJOR ENERGY CONSUMERS IN MIDREX TECHNOLOGY (FOR DRI)

Energy Consumer	Energy Consumption (kWh/t)
Hydrogen Electrolyzes	3000
Blowers and Compressors	12
Heating Systems	15
Gas cleaning and recirculation systems	7
Control and automation systems	2
Raw material handling systems	4

TABLE V. MAJOR ENERGY CONSUMERS IN ELECTRIC ARC FURNACE (EAF) TECHNOLOGY (FOR STEEL)

Energy Consumer	Energy Consumption (kWh/t)
Main electric arc process	300
Electrode feeding and regulation	10
Gas cleaning	10
Cooling systems	5
Loading and unloading mechanisms	10
Monitoring and control systems	5

The analysis of the data from the three tables indicates that the Midrex technology for direct reduced iron (DRI) production is the most energy-intensive, with hydrogen electrolysis consuming 3000 kWh/t. The Electric Arc Furnace (EAF) also shows significant energy demands, especially for the main electric arc process, which requires 300 kWh/t. In comparison, the Blast Furnace (BF) and Open-Hearth Furnace (OHF) processes are considerably less energy-intensive, with the highest consumption around 10 kWh/t for gas cleaning systems in the IOSP process. Therefore, while technologies like Midrex and EAF offer environmental benefits, they require substantially more energy compared to traditional methods.

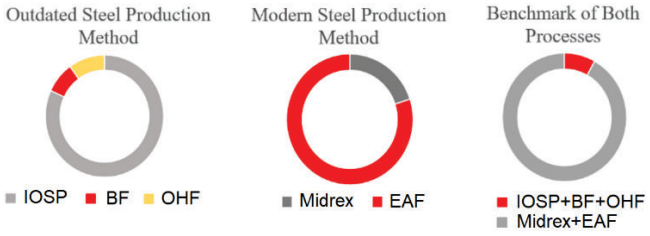


Fig. 7. Comparative energy consumption diagram.

Thus, the outdated method, the IOSP (Fig. 7a), which is used to prepare ore materials for blast furnace smelting, occupies the most share of energy consumption. According to the obtained data, its share is 83%, with the blast furnace process at 7% and the open-hearth process at 10%. In the modern method (Fig. 7b), the primary consumer is the EAF, which accounts for 80% of the complex's energy consumption. The benchmark (Fig. 7c) indicates also that the modern method consumes 90% of the energy compared to the outdated method, indicating that the modern method is 9 times more energy-intensive.

IV. CONCLUSION

The study demonstrated that transitioning from the traditional IOSP + BF + OHF process to the modern Midrex + EAF process leads to a significant reduction in CO_2 emissions – by as much as 75%, from 2430 to 613 kg per ton of steel. Despite a slight increase in electricity consumption, the new technology allows for a considerable decrease in the carbon footprint of production, making it more environmentally friendly and responsible. The comparison of steel production methods revealed significant differences in energy consumption. The open-hearth process, outdated and less efficient than EAF, is still used for the production of certain steel grades. The Midrex process offers a more energy-efficient and environmentally friendly alternative to the traditional IOSP + BF + OHF steelmaking route, particularly in the face of stricter environmental regulations.

The adoption of modern technologies, such as Midrex and EAF, is crucial for the steel industry to reduce its carbon footprint and improve energy efficiency, ensuring a sustainable future.

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Comparative Biomechanical Analysis of Shape-Closed and Force-Closed Femoral Stems in Total Hip Arthroplasty

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***Abstract*—The femoral stem design and cementation technique are among the most important factors determining the long-term success of total hip arthroplasty (THA), especially in patients with low bone quality. Cemented femoral stems can be force-closed or shape-closed, and each type demonstrates unique biomechanical attributes and clinical results. This study focuses on the comparative biomechanical analysis of force-closed versus shape-closed cemented femoral stems, which compares their structural and fatigue properties under physiologically relevant loading conditions. A three-dimensional structural analysis is performed on two cemented femoral stem designs such as i) a collarless, force-closed stem and ii) a collared, shape-closed stem. Static structural and fatigue analysis simulations were conducted for steady loading and a dynamic response. The results were examined for certain key parameters, including von Mises stress, total deformation, strain, and safety factor. Under static loading, the shape-closed design demonstrated superior mechanical performance concerning lower von Mises stress (44.7 MPa vs. 112.5 MPa), strain (2.84×10^{-4} vs. 5.54×10^{-4}), and deformation (6.27×10^{-3} mm vs. 7.76×10^{-3} mm). Dynamic load simulations also showed similar trends in the results. The fatigue analysis corroborated this enhanced performance of the shape-closed stem, revealing a factor of safety more than 2 times higher for every loading condition from the analysis. A femoral stem with a shape-closed design with a biomechanically optimized collar and control of proximal geometry can provide higher stability and fatigue resistance compared with a force-closed design. The results, therefore, imply that shape-closed stems should provide greater durability than forced-closed stems in the press-fit condition of cemented THA, particularly when cementing techniques are not adequate.**

Keywords— ***Total Hip Arthroplasty, Shape-Closed Stem, Force-Closed Stem, Femoral Stem Design, Fatigue Analysis.***

I. Introduction

Total Hip Arthroplasty (THA) is the most frequently performed and effective orthopedic procedure, used to restore movement and reduce pain in patients with advanced hip joint diseases such as osteoarthritis, rheumatoid arthritis, and femoral neck fractures. With increasing life expectancy and activity of patients, long-term durability and biomechanics of hip implants are of utmost importance. The optimal femoral stem is of utmost importance for the success of THA, as it is involved in load transfer, stability of the joint , and the survival of the prosthesis [1]. While femoral stems can be divided into multiple categories, cemented stems continue to be used frequently, especially in patients with poor bone quality [2].

About the biomechanical behavior, cemented femoral stems can be mainly classified into two groups: force-closed and shape-closed stems [3]. Force-closed stems, usually without a collar and polished, are meant to subside to a certain degree in the cement mantle, allowing even load transfer and durable anchorage. They are frequently used because they are versatile, easily modified, and not too demanding surgically. These stems are in more prevalent use, they perform well in myriad patients with ease of revision should it be required, and have excellent long-term findings, especially in the elderly [2] [4]. While long-term registry data shows that force-closed designs like the Exeter stem demonstrate superior survival rates compared to shape-closed designs such as the Lubinus stem (73.9% vs. 65.5% over 25 years), the reasons for this difference are not clearly stated [4]. Force-closed stems have their downsides as well. A greater amount of cement can initiate stress shielding and sometimes result in progressive subsidence or micromotion, ultimately causing the long-term excessive displacement of the prosthesis, particularly in young and old patients or those with higher physical activity [5] [6]. Shape-closed stems frequently exhibit a greater proximal geometry that is more anatomical and larger. They are designed to include a collar to reduce micromotion and improve proximal load transfer by closely matching the endosteal geometry of the femoral canal. Although they show lower aseptic loosening and superior initial stability, they have several limitations. These include increased intramedullary pressure during implantation, which may precipitate bone cement implantation syndrome (BCIS). They are additionally sensitive to cementing technique and can show loosening over time as a result of non-uniform application of forces and wear [2] [3] [5].

The factors to consider in hip replacement design are material selection, energy efficiency and force balance, and thermal management and material durability, to achieve a

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functional, durable, and easy-to-use solution [7]. Few novel designs used special plastic liner, ceramic head with metal backing or porous titanium cup [8]. But these design do not involve modification or optimization of the implant stem. The difficulty for the design is the balance of initial enough fixation, long-term stability, and safe cementation. However, limited comparative biomechanical studies have been reported in the literature investigating the biomechanical behaviors, such as the structural stiffness, deformation, and fatigue properties, concerning loading conditions of the force-closed versus the shape-closed stems. In the present study, we sought to fill this void by conducting an extensive structural and fatigue analysis on two cemented femoral stem designs - a collarless force-closed design and a collared shape-closed design. Because excessive pressure is the focus of conservative proximal geometry, we increased the collar width of concentration in the shape-closed design for load distribution to reveal the contrast. By comparing the total deformation, stress, strain, and fatigue safety factor under static and dynamic load, this study attempts to find out which shape design has better biomechanical characteristics of the nail. The following research will provide the basis for the design and selection of future implants by defining the mechanical compromise associated with these two commonly utilized stem types in cemented THA.

II. Methodology

A. Workflow in Brief:

The workflow for the design and validation of an implant stem is illustrated in Fig.1. It starts with CAD modeling, followed by structural analyses under different loading conditions, fatigue analysis, and structural validation, culminating in 3D printing.

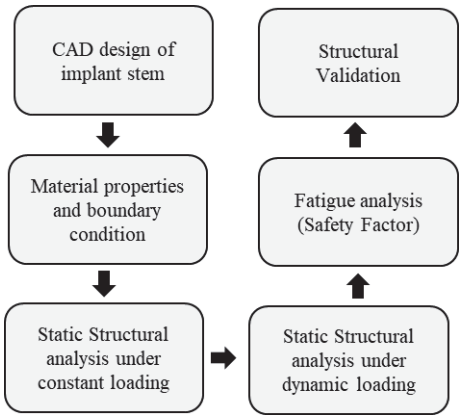


Fig. 1. Methodology flowchart

B. Stem Design:

The stem of implants is traditional in design. Fig. 2 (a) and (b) illustrate a shape-closed and force-closed implant, respectively. One of the main differences between the two designs lies in the presence of a collar and its proximal geometry, which significantly influences the load transfer, friction resistance, subsidence behavior, and distribution of tensile stresses. This design includes a 6 mm thick collar, which is thicker than usual 2mm. The proximal geometry shape is closed and small similar to force force-closed stem.

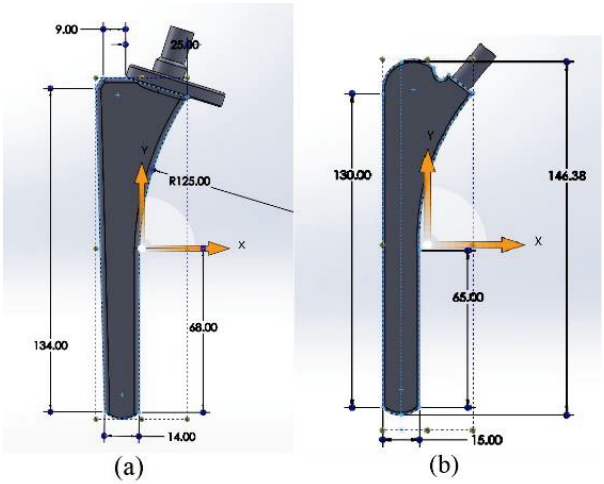


Fig. 2. Shape closed (a) and Force closed (b) hip implant stem

C. Structural Analysis Material Properties

The most common material used to make hip implant stems is Titanium alloy and cobalt-chromium alloy. In this study, titanium alloy is used. In TABLE I, material properties are listed for the titanium alloy.

TABLE I. MATERIAL PROPERTIES

Property	**Value**
Density	4620 kg/m³
Isotropic Secant Coefficient of Thermal Expansion	9.4×10^{-6} °C^{-1}
Young's Modulus	96000 MPa
Poisson's Ratio	0.36
Tensile Yield Strength	930 MPa
Compressive Yield Strength	9.3×10^{8} Pa
Tensile Ultimate Strength	1.07×10^{9} Pa

D. Boundary Conditions

To carry out the static structural analysis, here used a force of 2300.5 N, as a constant force [1]. Fig. 3 (a) and (b) show us the direction of force applied in both shape and force-closed stems. A dynamic loading condition is applied to examine the strength and load-bearing capacity of the implant stem.

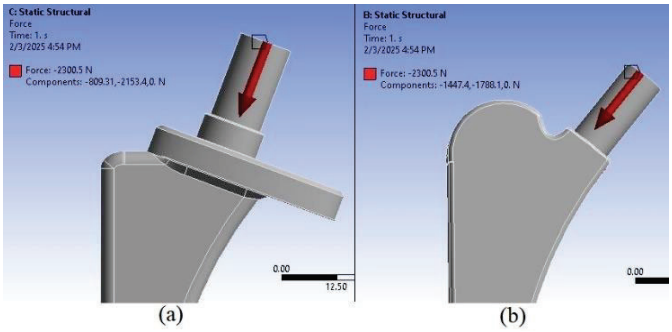


Fig. 3. Force direction on Shape closed (a) and Force closed (b) implant stem.

This dynamic load, expressed as a force versus time percentage curve, shown in Fig. 4, illustrates the variable forces acting on the hip joint during a normal gait cycle [9]. The force curve at the hip joint in walking shows two phases, stance and swing. Because of weight-support, forces on the hip joint are greater in the stance (0–60%) than in the swing (60–100%) phase during walking.

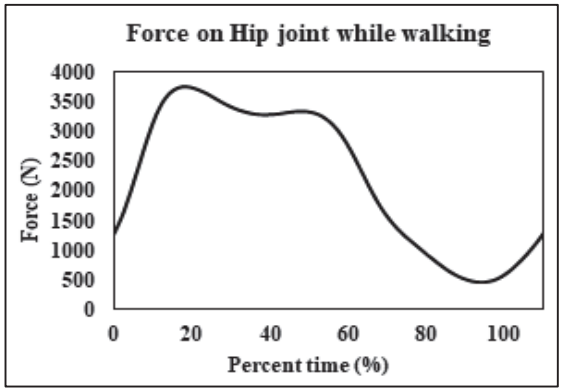


Fig. 4. Dynamic force plot on Hip Joint [9]

E. Fatigue Analysis

The fatigue analysis is carried out to assess the performance of the implants under long-term cyclic loading, which would represent the real-life physiological stresses in daily activity. This analysis provides an estimate of the implant's fatigue resistance as a function of the cyclic loadings and is based upon the S-N (Stress-Number) curve of titanium that describes its fatigue properties [11]. The fatigue safety value characterizes the margin in life by which the implant can withstand cyclic loads without fracture before the design life. A safety factor larger than 1 in fatigue represents survival under cyclic loading and smaller than 1 indicates failure. Maximum safety factor 15 is used to assess the fatigue resistance of the stem designs [10].

III. Results and Discussion

A. Static Structural Analysis Under Constant Loading

A comparative study is carried out on shape-closed and force-closed models to determine the most suitable design for the stem implant. Fig. 5 and Fig. 6 demonstrate the contour plots of von mises stress, deformation and strain for both shape closed and force closed designs. It is observed from TABLE II that, under constant loading of 2300.5 N collared stem exhibited less deformation compared collarless stem.

TABLE II. STRUCTURAL ANALYSIS UNDER CONSTANT LOADING

Design type	Total deformation (mm)	Maximum strain in mm/mm	Equivalent von Mises stress in MPa
Shape closed	6.2654E-03	4.6918E-04	44.662
Force closed	7.7558E-03	1.1998E-03	112.45

a) Equavalent Von Mises Stress: From the von mises stress plot in Fig. 7(a), it is found that the maximum stress in the shape-closed design is 44.662 MPa, while the force-closed design suffered much higher maximum stress of 112.45 MPa. There is almost 2.5-fold stress reduction which indicates that shape-closed stem has superior load-sharing capabilities compared to force-closed stem. There is also decreased likelihood of stress concentrating and developing a fatigue failure in shape-closed stem.

b) Total Deformation: Under constant loading conditions, the total deformation of the femoral stems is shown in Fig. 7(b) for both the shape-closed and force-closed designs. The shape-closed implant demonstrated a total deformation of 6.2654×10^{-3} mm, while the force-closed implant exhibited a significantly higher deformation of 7.7558×10^{-3} mm. The result shows shape-closed stem maintained greater structural rigidity and less deformation under constant loading.

c) Equivalent Strain: The equivalent strain values under constant loading are also compared to support the deformation results. The plots are shown in Fig. 7(c). The shape-closed implant recorded a maximum strain of 2.84×10^{-4}, while the force-closed implant showed a higher strain value of 5.54×10^{-4}. The lower strain in the shape-closed design indicates that it experiences less internal deformation when subjected to external forces.

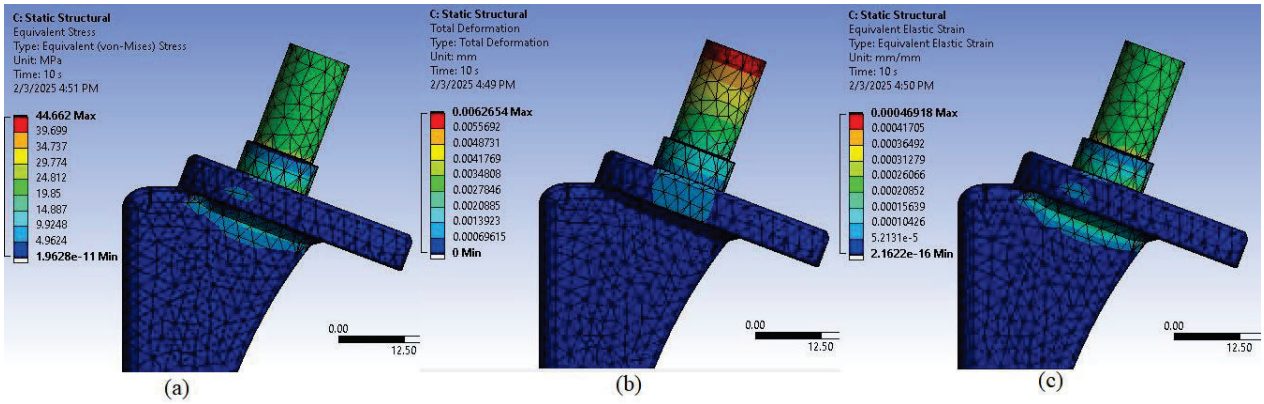


Fig. 5. Contour plots of Equivalent von mises (a), total deformation (b), equivalent strain (c) of the shape closed model

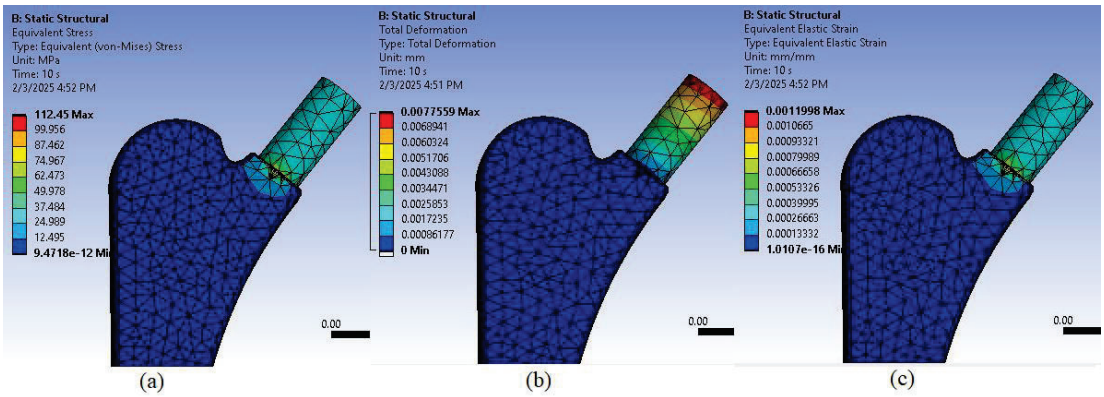


Fig. 6. Contour plots of Equivalent von mises (a), total deformation (b), equivalent strain (c) of force closed model.

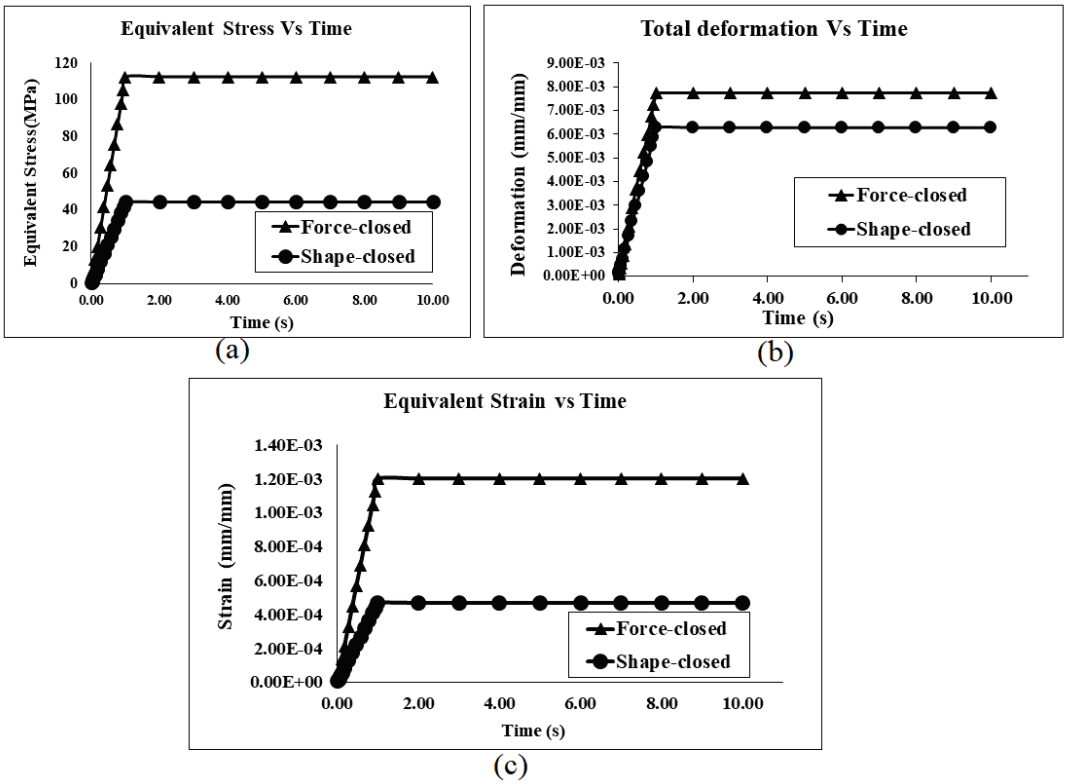


Fig. 7. Structural analysis plots for constant loading condition, Equivalent von Mises (a), total deformation (b), equivalent strain (c)

B. Static Structural Analysis Under Dynamic Loading

Under dynamic loading conditions simulating physiological movement, the shape-closed design continued to outperform the force-closed design. Fig. 8 and Fig. 9 show the contour plots of analysis results for both stem designs. Fig. 10 contains all the relevant plots for von mises stress, deformation, and strain results. TABLE III. outlines the stress, deformation, and strain results.

TABLE III. STRUCTURAL ANALYSIS UNDER DYNAMIC LOADING

Design type	Total deformation in mm	Maximum strain in mm/mm	Equivalent von Mises stress in MPa
Shape closed	3.55E-03	2.8373E-04	27.123
Force closed	1.12E-01	5.5421E-04	53.145

a) Equivalent Von Mises Stress: From Fig. 10(a), the highest von Mises stresses are 27.1 MPa observed in the shape-closed stem, but at the force-closed stem, a much higher stress of 53.1 MPa appeared. This suggests that the shape-closed implant efficiently manages dynamic loads by distributing stresses evenly throughout its structure, thus reducing localized stress concentrations that could lead to fatigue cracks or implant failure over time. Effective stress management under dynamic loading is critical for the longevity and mechanical reliability of hip implants.

b) Total Deformation: Fig. 10(b) illustrates the shape-closed implant exhibited a total deformation of 3.55×10^{-3} mm, remaining stable even under fluctuating forces. Conversely, the force-closed implant demonstrated a considerably higher deformation of 1.12×10^{-1} mm. This shows shape-closed stem is significantly better at maintaining dimensional stability during dynamic

activities, minimizing the risk of micromotion at the bone-implant interface

c) Equivalent Strain: The shape-closed stem showed a maximum strain of 2.84×10^{-4}, compared to 5.54×10^{-4} in the force-closed stem in Fig. 10(c). This suggests that the shape-closed implant efficiently manages dynamic loads by distributing stresses evenly throughout its structure.

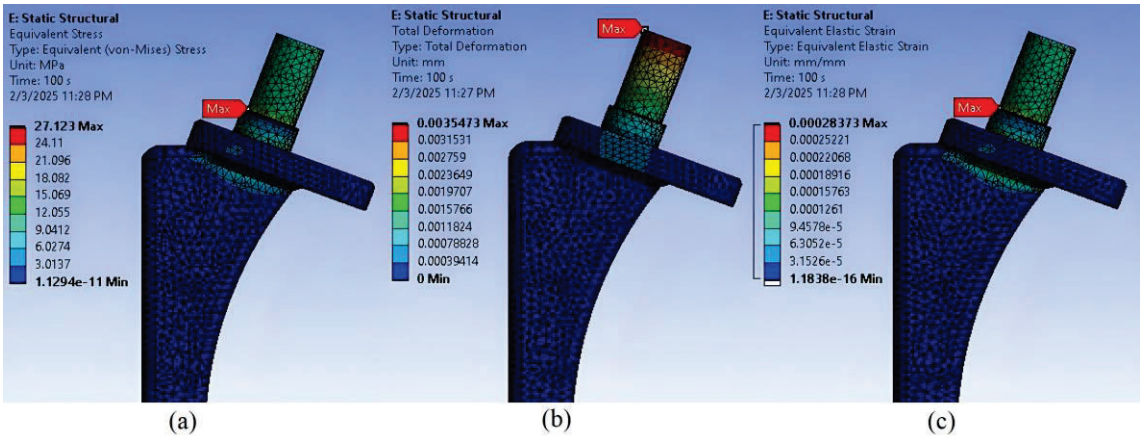


Fig. 8. Contour plots of Equivalent von mises (a), total deformation (b), equivalent strain (c) of force closed model under dynamic conditions.

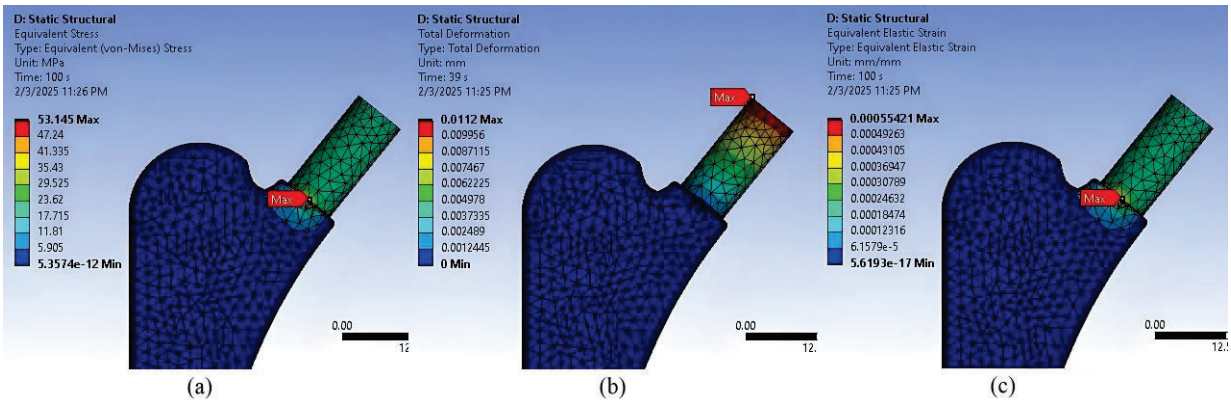


Fig. 9. Contour plots of equivalent von mises (a), total deformation (b), equivalent strain (c) of shape closed model under dynamic conditions

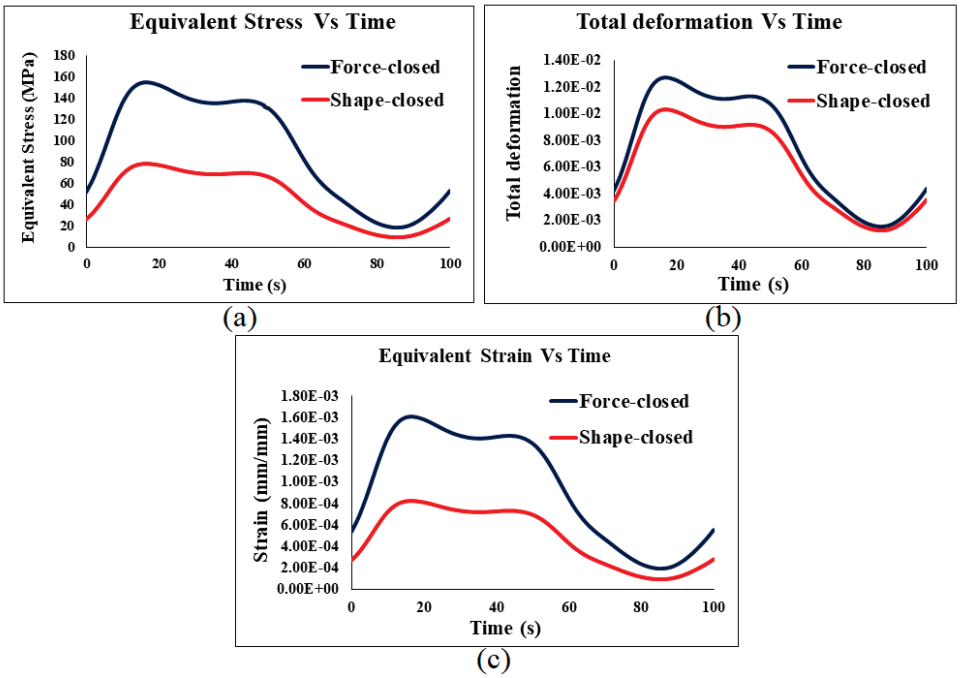


Fig. 10. Structural analysis plots for dynamic loading condition, Equivalent von mises (a), total deformation (b), equivalent strain (c).

C. Fatigue Analysis

The results of the fatigue analysis reveal a significant mechanical advantage of the shape-closed stem over the force-closed stem in both constant and dynamic loading scenarios. As detailed in Table IV, the shape-closed design exhibited a safety factor over twice that of the force-closed design under both constant (8.67 vs. 3.44) and dynamic (14.27 vs. 7.28) loading conditions. These findings indicate that the shape-closed design has superior resistance to cyclic stresses and sustained durability under realistic gait forces.

TABLE IV. SAFETY FACTOR

Design Type	Safety Factor (Constant loading)	Safety Factor (Dynamic loading)
Shape closed	8.6651	14.268
Force closed	3.4415	7.2819

IV. Conclusion

This preliminary study demonstrated that the shape-closed femoral stem, with its refined proximal geometry and the addition of a collar, outperformed the force-closed design under both constant and dynamic loading conditions. The differences were not just qualitative but substantial. Quantitatively, under constant loading, the shape-closed stem carried less than half the von Mises stress of the force-closed design (44.66 MPa vs. 112.45 MPa), deformed less (0.00627 mm vs. 0.00776 mm), and showed lower strain (2.84×10^{-4} vs. 5.54×10^{-4}). Similar trends appeared under dynamic loading, where the shape-closed stem again halved the stress (27.12 MPa vs. 53.15 MPa) and showed dramatically lower deformation (0.00355 mm vs. 0.112 mm). Fatigue analysis reinforced these findings: safety factors for the shape-closed stem were more than twice those of the force-closed stem—8.67 vs. 3.44 in constant loading, and 14.27 vs. 7.28 in dynamic loading. The optimized proximal shape helped reduce internal bone pressure during implantation, improving surgical safety, while the collar enhanced proximal load transfer, reducing stress. These design features contributed to lower von Mises stress, strain, and deformation, as well as significantly higher fatigue safety factors—indicating better structural stability and long-term durability. The structural analysis proves clear mechanical advantages of the shape-closed stem however, this study employed simplified 3D models, one material (Titanium alloy), and simulated only normal gait. Experimental validation were also not accounted for. While the models used in this study captured the essential mechanical behaviors of the implants, future research could benefit from incorporating more complex physiological conditions, a broader range of materials, and detailed modeling of the implant–cement–bone interface. Validating these findings through experimental testing or patient-specific simulations could provide even deeper insight into long-term implant performance. Nonetheless, the results presented here offer strong evidence that a carefully optimized shape-closed design has the potential to improve outcomes in cemented total hip arthroplasty.

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Modern IT solutions for eliminating excessive foundation tilts in multi-story buildings

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***Abstract*— This paper examines the practical use of Digital Twin technology in the reconstruction of multi-story buildings, focusing on the elimination of excessive tilts. The degree of foundation tilt is a critical factor in determining a building's operational suitability. In many cases, however, foundation tilts at different stages of a building's life cycle exceed acceptable or even critical limits. The combination of modern IT solutions, experimental non-destructive testing methods, and Digital Twin concepts makes it possible to assess the current technical condition of a multi-story building, forecast its near-term changes, and manage the process of tilt correction. At the same time, more than 90% of data on changes in a building's technical condition during prolonged alignment works are typically lost or distorted. This study offers a conceptual framework for organizing and protecting data generated during tilt correction using Digital Twins. The application of this IT-based diagnostic approach made it possible to complete the full scope of tilt elimination works across all block sections without interrupting the operation of utilities, elevators, or other technical systems, and without requiring resident evacuation. The key advantages of Digital Twins include the ability to forecast stress–strain changes, provide continuous diagnostic information, optimize processes, analyze and visualize real-time sensor data distributed throughout the building, simulate alternative tilt-correction scenarios, and support efficient resource management.**

Keywords*— modern IT solutions, digital twins, cyber-physical systems, excessive foundation tilts, reconstruction

I. INTRODUCTION

The use of digital technologies across many industries is undergoing rapid transformation, especially with the advent of Industry 4.0. Yet the domestic construction sector has been slow to adopt and apply these technologies. International experience shows that digital technologies in construction improve productivity, enhance quality and efficiency, and strengthen collaboration among stakeholders, thereby increasing project success [1]. Notable examples include Building Information Modeling (BIM), Digital Twins (DT), Distributed Ledger Technology (DLT) or Blockchain, 3D printing, the Internet of Things (IoT), Artificial Intelligence (AI), and Augmented or Virtual Reality (AR/VR) [2]. Within the framework of Construction 4.0, efforts are being made to introduce these advanced tools and apply them across different stages of a building's life cycle. In Ukraine, however, the practice of implementing IT systems for real-time diagnostics of building conditions remains rare, with only isolated cases to date. The development of such monitoring systems is shaped by several key factors: technical – identifying the physical and mechanical parameters on which the measurement system relies; technological – designing methods and procedures for manufacturing components, installing the system, and ensuring reliable operation; and economic – optimizing the cost parameters of the system [3]. Acceptable values for certain technical condition parameters of buildings and structures, such as foundation tilt (Fig. 1), are defined by regulatory documents [4, 5].

Uncertainty in the "foundation–multi-story building" system can result from flooding, seismic activity, emergency water supply failures, man-made vibrations (e.g., from transport or operating equipment), or changes in load due to increased building density, among other factors. Assessing the stress–strain state of the "soil–foundation" system during tilt correction requires monitoring multiple physical parameters. This, in turn, demands the use of numerous sensors and various primary measuring devices with different designs and operating principles. Examples include strain gauges, magnetoelastic and magnetostrictive devices for measuring forces and stresses; inductive and electromagnetic devices for measuring linear and angular displacements; and liquid-based devices for measuring relative settlements and foundation deformations.

The use of modern information technologies for diagnosing the current technical state of multi-story buildings during tilt elimination makes it possible to save both time and money, while also broadening knowledge of building components, foundations, and soil conditions, all while maintaining acceptable safety levels. A Digital Twin is based on real-time monitoring and analysis of the key parameters of

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the soil–foundation system, which allows revisions to the project at any stage of work [16, 17]. A universal DT must also include a module for continuous calculation of the stress–strain state of the soil and superstructure during the tilt elimination process, as well as a module for generating current management commands [16]. Technological settlements and tilts of the building are predicted using modern numerical methods, taking into account excavation specifics, drilling of wells, and other factors.

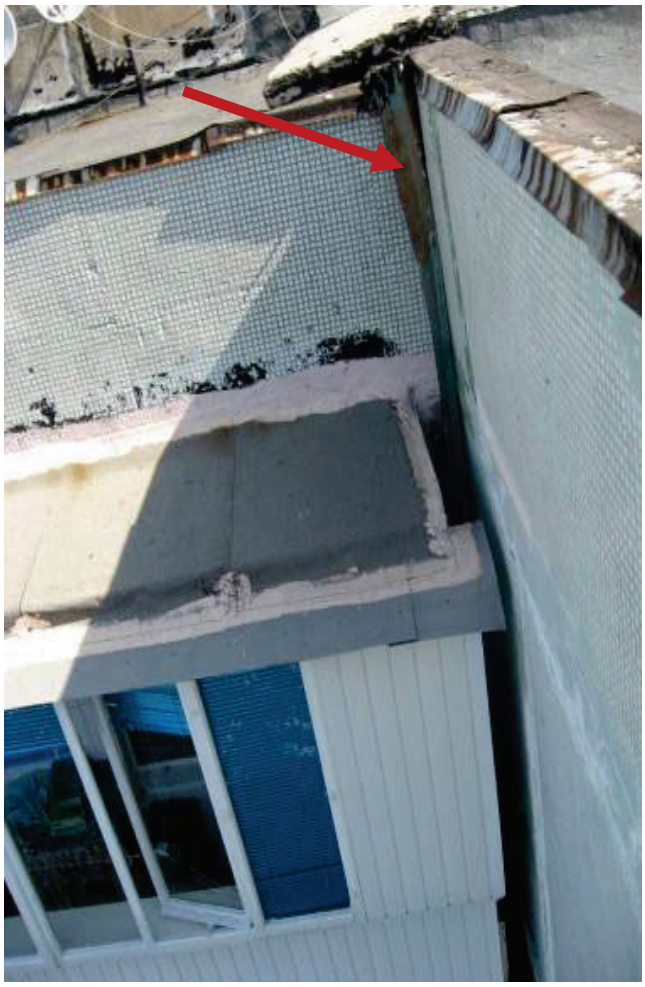


Fig. 1. Closure of the expansion joint and leaning of block-sections No. 3, 1 of residential buildings No. 23, 25 at Preobrazhenska Street in the city of Zaporizhzhia.

To create geomechanical and structural models of the "underworked soil mass (base) – tilted foundation" system, the general patterns of soil deformation, and in particular the contours of cylindrical wells, were studied. For effective decision-making and risk reduction, comprehensive DT-based information about the current technical state of the building is essential [18]. Such information can be obtained by performing a series of residual capacity checks for the main structural elements of the soil–foundation–superstructure system using real-time DT results. Calculations in LIRA CAD [14, 19] are based on a finite element spatial model of the building (Fig. 2).

The finite element method forms the theoretical basis of LIRA CAD. The implemented version of finite element method uses the principle of possible displacements:

$$a(u, v)=(f, v) \tag{1}$$

where u is the exact solution to be found, v is any possible displacement, $a(u, v)$ and (f, v) represent the possible actions of external and internal forces. The structure's domain is divided into finite elements Ω_r, and nodes with associated degrees of freedom L_i (displacements and rotations) are assigned. The approximate solution U_h must satisfy the main kinematic conditions and is expressed as a linear combination of the basic functions.

Let K be the stiffness matrix, P – the load vector and X – the vector of unknowns.

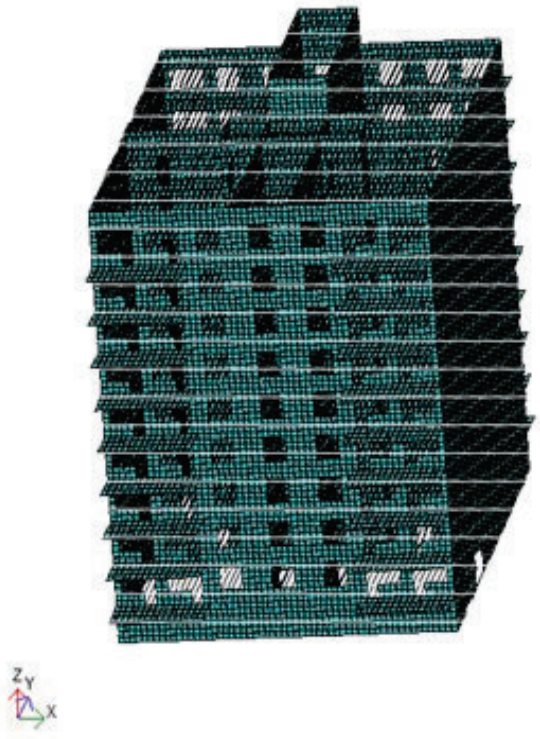


Fig. 2. Calculation model of a building that has undergone tilting.

Then can be written in matrix form as:

$$KX=P \tag{2}$$

Thus, finite element method reduces the problem to solving a system of linear algebraic equations (2). A key advantage is that the stiffness matrix K and load vector P are obtained by summing contributions from individual finite elements.

The Finite Element Library includes elements modeling the behavior of different structures: bar elements; quadrilateral and triangular elements for planar problems (e.g., slabs, shells); spatial elements such as tetrahedra, parallelepipeds, and trihedral prisms. It also includes special elements modeling finite stiffness connections, elastic compliance between nodes, and user-defined stiffness matrices. All elements in the Finite Element Library are theoretically validated, with established energy and displacement error estimates.

Some Ukrainian researchers are actively developing theoretical and methodological foundations and tools for creating and applying IT solutions in construction, assessing technical condition of multi-story buildings, and designing

and implementing information-analytical decision-support systems based on LIRA-CAD [6–12].

II. Digital Twin Technology in Eliminating Excessive Tilts of Multi-Story Buildings

If there is a fully integrated, automatic, two-way data exchange in real time between a physical object and its digital model, this is called a Digital Twin [13]. In such cases, the digital model can function as a controlling counterpart to the physical object. Other objects, physical or digital, may also influence the model.

Changes in the physical object's technical state lead to immediate updates in the digital model, and the reverse can also occur (see Fig. 3, adapted from [13]).

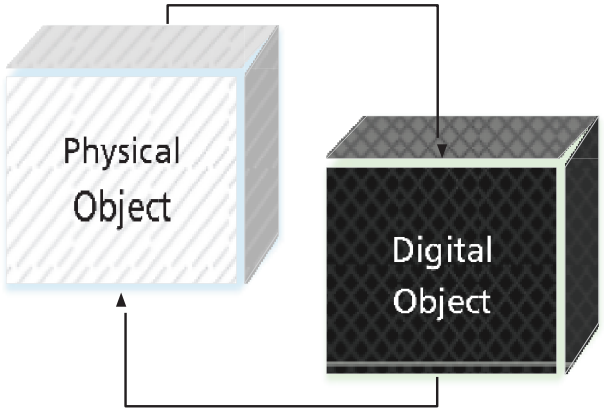


Fig. 3. Data Flow in a Digital Twin [13].

The application of Digital Twin technology for eliminating excessive tilts in multi-story buildings, and the components of this approach, are described below. A key element is the automated information-measuring system Monitoring, designed to measure tilt directions and angles of buildings and their elements in real time, ensuring safe correction of excessive tilts.

This system is based on inductive transducers and supports remote data acquisition, processing, and transmission from sensitive elements. Monitoring the stress–strain state of a building during tilt correction is achieved through an algorithm consisting of sequentially linked tasks, grouped into blocks: science and research, measurement and information, admission and control, calculation and analysis, and design and technology. The technical components of the Monitoring system are shown in Fig. 4.

Data are transmitted over the Internet (IoT) to cloud or PC, with subsequent processing. The system operates automatically with programmable activation intervals, powered by ~220 V supply. An uninterruptible power supply ensures 1000 hours in standby mode or 48 hours in measurement mode. Every 12 hours, the battery charge of the system is checked. After polling sensors (99 sensors in 12 seconds), the system enters power-saving mode. During development, an original inductance meter was created and integrated into the sensors.

Its sensitivity is 0.5 μH, with an error of only 0.2% across the working range. The displacement accuracy is 0.001 mm, while tilt error is 0.05%. The data collection unit provides multiple functions: power supply (from grid or battery), UPS charging, powering sensors via communication lines, data collection from sensors, data transmission and control, saving settings in non-volatile memory, entering power-saving mode between cycles, and connecting to communication lines via DB9 connector.

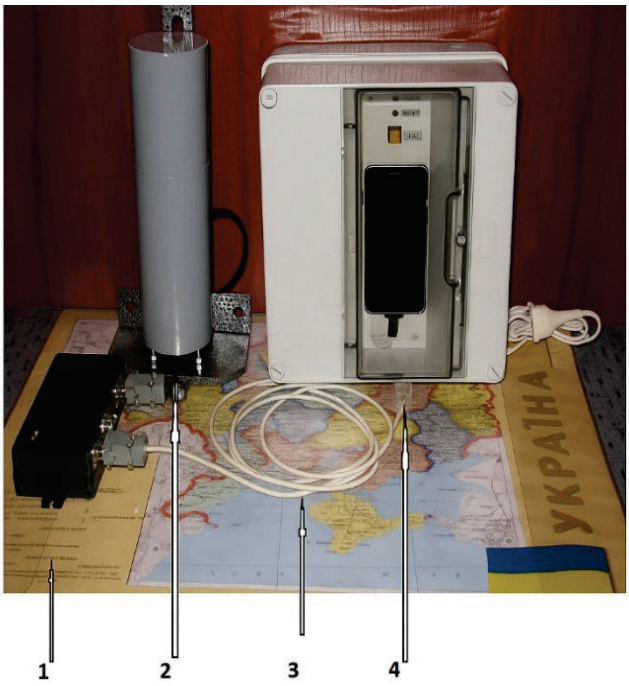


Fig. 4. Technical components of the automated information-measuring system Monitoring: 1 – inductance meter; 2 – angular deformation sensor; 3 – communication lines; 4 – data collection and transmission unit (IoT).

A schematic of the Monitoring system installed in a multi-story building is shown in Fig. 5 [15, 16].

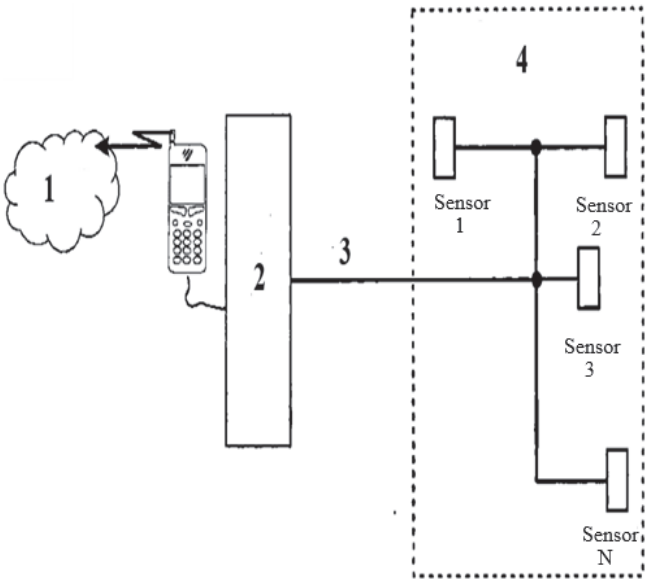


Fig. 5. Schematic of the Monitoring system in a multi-story building: 1 – Internet; 2 – data collection unit; 3 – communication line; 4 – measurement object.

The external appearance of a universal inductive sensor with an inductance meter, installed in the basement of a tilted residential building in Zaporizhzhia, is presented in Fig. 6.

The collection of information from precision inclinometric sensors (measuring tilt angles of structural elements)

distributed throughout the building is facilitated by IoT (No. 4 in Fig. 4 and No. 1 in Fig. 5).



Fig. 6. The appearance of the universal inductive device sensor installed on a wall in the basement of the building.

Current experimental measurements of the stress–strain state and technical condition of the building are promptly processed and provided to decision-makers to determine whether alignment should continue or be suspended. Processing of these online results by the monitoring system initiates the following management actions: (1) sequential adjustment of the digital model according to the calculated stress–strain parameters of the tilted building using the LIRA-CAD software suite [14, 19]; (2) comprehensive analysis of results to identify structural elements in critical states and, if needed, reinforcement measures [17]; (3) selection of technological methods for settlement adjustment based on the calculated stress–strain diagram; (4) implementation of corrective measures when discrepancies arise between actual and projected parameters such as settlement or tilt.

For this reason, straightening projects must include engineering measures to ensure feedback (Fig. 3). This feedback loop ensures consistency between predicted and observed deformations under conditions of uncertainty and potential ground failures, and helps identify soil zones requiring additional intervention. Such interventions may include correcting settlements under strip foundations by drilling supplementary boreholes of various diameters, moistening soil beneath foundations, and other methods. The ultimate goal is to securely stabilize the building in its new position, ensuring reliable performance during alignment without interrupting utilities, technical systems (elevators, etc.), or requiring resident evacuation.

Initial and boundary conditions are updated based on current inclinometer readings across the building, taking into account its actual technical state at the time of calculation. These measures increase the reliability of decision-support systems and the accuracy of managerial commands issued by project executives, while also allowing identification of localized soil areas requiring targeted management actions, such as settlement correction beneath strip foundations.

The DT uses the following instruments: universal inductive tilt sensors (Fig. 6) and crack-width sensors (accuracy: 0.001 m), operating within –30º…+70ºC and protected against dust and moisture (IP56); an inductance meter for recording; a mobile-phone-based IoT unit for data collection and transmission; and the Pendulum software package [15].

To improve measurement accuracy, the device incorporates several techniques: warming-up before data collection (about 75% of total measurement time), averaging nine consecutive readings, and normalization against a reference coil. Normalization is used because inductance of ferromagnetic-core coils strongly depends on current magnitude due to nonlinearity of the magnetization curve. The reference coil also compensates for temperature effects on coil properties and circuit elements. For robust compensation, the reference coil must be of the same type as those being measured. The meter transmits both normalized and raw measurement values to the collection unit, enabling verification of normalization accuracy.

During tilt correction, discrepancies often arise between actual and predicted values of settlement or tilt. Therefore, additional engineering measures such as real-time monitoring and technological settlement-control methods must be included in design projects. This ensures consistency between predicted and actual deformations of the “underworked soil–tilted strip foundation” system throughout tilt elimination.

III. Results

The development of horizontal displacements along the x and y axes during tilt elimination is presented in Fig. 7, in a format convenient for decision-makers.

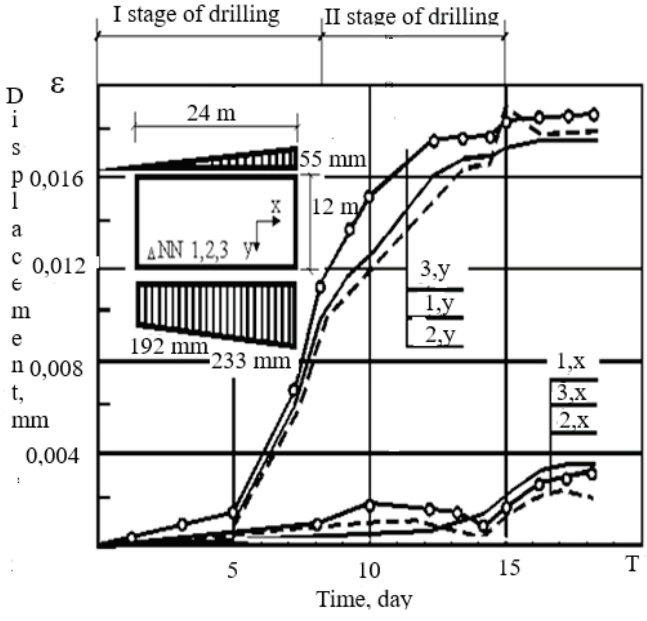


Fig. 7. Development of horizontal displacements along the x and y axes during elimination of excessive tilt in a multi-story building. Sensors No. 1, 2, 3 were installed in the basement, second floor, and roof, respectively.

Five phases of tilt changes in the block sections of the building were observed: (1) minor settlements during the first

stage of drilling; (2) active progressive deformations during the second stage of drilling and immediately after completion; (3) stabilization of settlements after drilling works; (4) active progressive deformations during adjustment of technological settlements by moistening the soil around wells; (5) stabilization of deformation processes.

IV. Conclusions

Modern IT solutions and Blockchain technology for eliminating excessive foundation tilts in multi-story buildings were examined. The integration of IT systems with non-destructive testing methods and Digital Twin technology allows real-time assessment of building condition, forecasting of near-term changes, and management of tilt correction [17]. However, over 90% of condition-change data generated during prolonged alignment is typically lost or distorted.

The graphs (Fig. 7) illustrate the development of horizontal displacements along the x and y axes throughout tilt correction. Five distinct phases of tilt development were observed: (1) minor settlements during the first drilling stage; (2) active deformations during and immediately after the second stage; (3) stabilization after drilling; (4) active deformations during adjustment of technological settlements by soil moistening; (5) stabilization of deformation processes. This IT-based approach made it possible to complete all tilt elimination works across all block sections without halting utilities, technical equipment (elevators, etc.), or requiring resident evacuation.

The key advantages of Digital Twins include the ability to analyze and visualize real-time sensor data distributed throughout the building, provide continuous diagnostic information, optimize processes, simulate alternative tilt-correction scenarios, forecast stress–strain changes, and support efficient resource management. These capabilities ensure more effective and economically efficient management across all stages of a building's lifecycle [18].

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Regulation Characteristics and Stability Criteria of Induction-Synchronous Electromechanical Converters in the Motor Operating Mode

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***Abstract*— The work is dedicated to the further development of the theoretical foundations of a special class of induction-synchronous electromechanical converters (ISEMC) with an enhanced design. Using numerical calculation methods with the application of circuit-field modeling, a study of steady-state and transient electromagnetic and mechanical processes in an ISEMC has been conducted. Based on the results of circuit-field modeling, the regulation properties have been determined for the first time, and a family of regulation and mechanical characteristics, as well as the operating characteristic regarding the dependence of the armature current on the mechanical power, have been obtained for different values of the applied supply voltage under conditions of combined and separate operation of the induction motor (IM) and synchronous generator (SG) parts of the ISEMC operating in the motor mode. For the first time, the stability criteria of induction-synchronous electromechanical converters for the motor operating mode have been determined.**

Keywords*— induction-synchronous electromechanical converters, regulation properties, characteristic, stability criteria, field current, armature current

I. Introduction

In the context of the active development of power engineering, production automation, and the implementation of energy-efficient technologies, the need for electromechanical systems with high levels of reliability, energy efficiency, and compactness is increasing [1]. This is particularly relevant for AC electric drives operating at reduced rotational speeds, where the use of gearboxes is undesirable or impossible due to limitations in dimension, noise, or maintenance [2]-[4]. In such cases, slow-speed synchronous motors, capable of developing nominal torque at low rotational speeds [5], become the optimal solution.

Electric machines with combined windings are of additional interest, as its provide expanded functionality, as follows: operation in various modes (supply network/inverter, power supply from two independent sources, etc.); increased fault tolerance and reliability; and flexibility in torque and speed control. The implementation of electric machines with combined windings also allows for increased efficiency and reduced electromagnetic torque ripple [6], which is particularly important for precise and dynamically complex systems such as CNC machine tools, conveyors, robotics, and direct drive systems in power engineering and transport [7]-[9]. Thus, the development of slow-speed synchronous motors with combined windings represents an important and promising direction in the field of electrical machines, contributing to the creation of modern, highly efficient, and technologically flexible electromechanical systems.

II. Analysis Of Last Researches

In modern AC electric drive systems, various types of electromechanical converters (EMC) can be used, such as: asynchronous motors with a squirrel-cage rotor and a wound rotor, synchronous motors with slip rings, contactless synchronous machines (CSM) with electromagnetic or permanent magnet excitation, and others. The advantages and disadvantages of these EMCs, as well as their regulated characteristics, are well-known. Along with the most common classes of EMC, cascade EMC combined in a single magnetic core with combined armature windings (CAW) are being developed [10]-[12]. The fundamental idea of creating combined EMC has been known for a long time and has been used in the creation of a variety of their designs [10]-[13]. Combined EMC were not considered from the perspective of the general theory of EMC, and the further development of their theory was classified as belonging to the class of special EMC [10]. Until recently, this class of EMC did not have prospects for further development due to existing drawbacks in their design: the presence of a rectifier in the rotor, which leads to a significant increase in weight and dimension characteristics, complexity in manufacturing technology and maintenance; increased inertia of regulation due to the increased electromagnetic time constant of the exciter's field winding [10]. In papers [10]-[13], the authors proposed a new class of induction-synchronous EMC (ISEMC), which represents a further stage in the development and improvement of the design of cascade low-speed combined EMC, that combine the properties of an induction motor (IM) and a synchronous generator (SG) [12]. The design peculiarity consists in the following: the stator winding of the first EMC is a three-phase $2p_1$-pole winding, and the stator winding of the second EMC is a single-phase $2p_2$-pole field winding. The three-phase $2p_1$-pole winding and the single-phase $2p_2$-pole winding are electrically combined with each other and form a single conductive system on the stator (armature) side of the ISEMC Each phase of the armature winding consists of two parallel branches of coil groups, the terminals of which are connected to two separate neutral points ("Y-connection"). The terminals of the cathode and anode valve groups of a controlled (or uncontrolled) rectifier are connected to these neutral points to provide the combined armature winding with field current [14], [15]. Using the developed universal method for determining the optimal winding factor of EMC, the author [16] proposed, for combined rotor windings, to exclude the

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sections of those coils, whose currents create oppositely directed MMF vectors. And those rotor core slots in which these sections were located should not be punched [12]. In this case, the design of the rotor winding is significantly simplified and will consist of electrically unconnected short-circuited coils, the number of which is determined by the sum of the number of poles of the two combined EMCs [12]. Thus, the presence of electrically combined excitation and armature windings in the common conductive system on the stationary armature side of the ISEMC, electrically unconnected short-circuited coils of the rotor winding, as well as the absence of permanent magnets, a brush mechanism, and a damper winding, highlight distinct positive features of the ISEMC's design compared to known designs of contact and contactless synchronous EMCs.

To research the steady-state and transient electromagnetic and mechanical processes of this ISEMC design, the author in [10], [12] developed an adapted dynamic 2D circuit-field model of the transient interconnected electromagnetic and electromechanical processes of the ISEMC. This model allows for the consideration of the parameters of the end-winding parts in transient operating modes through its circuit implementation, the nonlinearity of the magnetic and electrophysical properties of the active materials, and the surface and edge effects of the end arias of its active part. Also, in [10], [13], methods were developed and a methodology was presented for determining the active and inductive resistances of the end-winding parts of the ISEM windings, taking into account the effect of edge effects, which made it possible to obtain a qualitative transition from a 3D to a 2D circuit-field model of the ISEMC.

In [12]-[15], the author used numerical modeling methods to research the electromagnetic and mechanical processes in the idle mode with the combined interaction of the induction and synchronous parts within the single active part of the ISEMC. The presence of self-synchronizing properties for the ISEMC was demonstrated, whereby asynchronous starting is ensured, followed by the pulling of its rotor into synchronism in the idle mode without excitation, and the conditions for their occurrence are explained. The relationships between the idle current and the excitation current, which determine the operation of the synchronous part of the ISEMC in underexcited, normally excited, and overexcited modes, were established. It has been proven, that in underexcited and normally excited modes, its synchronous part operates as a reactive power compensator, and in the overexcited mode, it operates as a generator.

Thus, the preliminary results obtained from the study of electromagnetic and mechanical processes in the enhanced ISEC design require further research of electromagnetic and mechanical processes, determination of its regulated properties and characteristics, as well as the formulation of criteria for the stable operation of the ISEMC in motor mode.

III. Formulation Of The Work Purpose

The aim of the work is to research the electromagnetic and mechanical processes in the ISEMC, to determine its regulated properties and characteristics, as well as to formulate criteria for the stable operation of the ISEMC in motor mode.

IV. Expounding The Main Material

The study of the electromagnetic and mechanical processes of the ISEMC was carried out for a manufactured experimental sample of the ISEMC (based on the MTF-111-6 crane motor) with a combined $2p_1$=4 and $2p_2$=8 armature winding (cascade rotation speed n=500 rpm) with a number of rotor phase coils n_2=6, a supply frequency of the armature winding f_1=50 Hz, at U_{phase}=54V, U_{phase} =74V, U_{phase}=92V, U_{phase}=108V (nominal phase voltage U_{rated} =74V). The electrical circuit of the model implementation of the armature and rotor windings of the ISEMC is fully described and presented in [17]. The schematic-geometric view of the 2D circuit-field model of the ISEMC is shown in Figure 1 (the connection of the end-winding and slot parts of the armature and rotor windings is not shown).

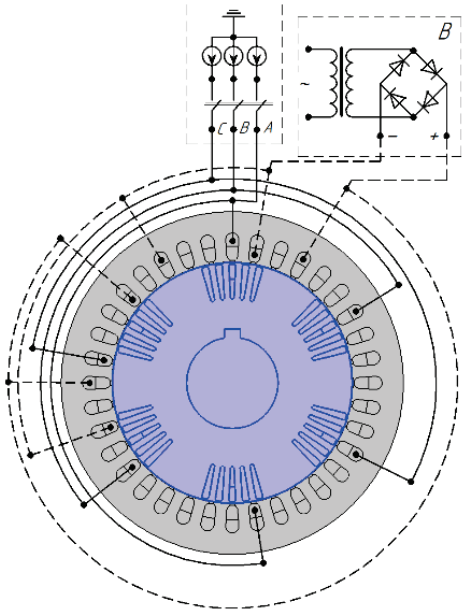


Fig. 1. Schematic-geometric view of the 2D circuit-field model of the ISEMC

In this case, the 2D field part of the mathematical model describes the transient electromagnetic processes in the calculation domain of the ISEMC using a system of nonlinear partial differential equations [17]-[19]:

$$\sigma_j \frac{\partial \mathbf{A}_j}{\partial t} + \nabla \times \left[\left(\mu_0 \mu_j (B) \right)^{-1} \nabla \times \mathbf{A}_j \right] - \ldots \\ \ldots - \sigma_j \mathbf{v} \times (\nabla \times \mathbf{A}) = \mathbf{J}_{ej} \quad (1)$$

where $\boldsymbol{A}$ is magnetic vector potential; $\boldsymbol{B}$ is magnetic flux density; μ, σ is accordingly relative magnetic permeability and electrical conductivity (for each calculation domain, it is specified separately according to the magnetic and electrophysical properties of the materials); $\boldsymbol{J}_e$ is current density of external sources; index j is corresponds to the subdomains of the geometric calculation area; $\boldsymbol{v}$ is velocity vector of the rotor's calculation domain relative to the stator's coordinate system, which is determined by the angular velocity of the rotor ω_m and the radius vector of a point in the rotor's calculation domain $\boldsymbol{r}$ ($\boldsymbol{v}$= $\omega_m \times \boldsymbol{r}$);

In accordance with [17], [18], the system of equations (1) is supplemented by the Coulomb gauge condition. Homogeneous boundary conditions are set on the external

boundaries of the 2D domain of the ISEMC's geometric model [19]. Equation (1) is supplemented by the coupling equations of the magnetic field and the electric circuit when phases A, B, C of the armature winding are supplied from the network, and also when supplied by the rectified current id from the excitation source, which flows through the neutral points "0+" and "0-":

$$\begin{cases} u_{m,n} \pm u_d = \dfrac{d\Psi_{m,n}}{dt} + L_{fr,n} \cdot \dfrac{d}{dt}\left(i_{m,n} \pm i_{f,n}\right) + \ldots \\ \ldots + \left(r_{fr,n} + r_{s,n}\right) \cdot \left(i_{m,n} \pm i_{f,n}\right), \\ \Psi_{m,n} = 2 \cdot \dfrac{n_w \cdot l_z}{S_{n.eff}} \int\limits_{S_a} A_{i,j} ds; i_{f,n} = \dfrac{i_d}{3}; \\ J_e = \dfrac{n_w \cdot \left(i_{m,n} \pm i_{f,n}\right)}{S_{n.eff.}}, \end{cases} \quad (2)$$

where m=A,B,C and n=1,2 are the phase index and parallel branch (coil group) of the armature winding phase; $u_{m,n}$ is voltage applied to the armature winding of the corresponding phase; $\psi_{m,n}$, $i_{m,n}$ are magnetic flux linkage and currents of parallel branches of phase coil groups n of the corresponding phase m of the armature winding; $L_{fr,n}$, $r_{fr,n}$, $r_{s,n}$ are respectively, the inductance of the frontal part and the resistances of the frontal and slots parts of the parallel armature winding branch; u_d, i_d, are rectified voltage and current of the excitation system; $i_{f,n}$ are the field currents flowing in the corresponding parallel branch n of phase m of the armature winding (the "±" is sign corresponds to the direction of the field current flow); n_w is number of turns of the armature winding; $S_{n.eff}$ – cross section of an effective conductor; S_a – total cross-sectional area of all turns of the parallel branch of the coil group.

The mathematical model (1)-(2) is also supplemented by the coupling equations of the magnetic field and the electric circuit of the rotor winding, which consists of r electrically unconnected coils:

$$0 = \frac{d\Psi_{m2,r}}{dt} + L_{fr,r} \cdot \frac{di_{m2,r}}{dt} + \left(r_{fr,r} + r_{s,r}\right) \cdot i_{m2,r}, \quad (3)$$

where $\psi_{m2,r}$, $i_{m2,r}$ are magnetic flux linkage and currents of the corresponding short-circuited rotor winding coil; $L_{fr,r}$, $r_{s,r}$, $r_{fr,r}$, are respectively, the inductance of the frontal parts and the resistances of the slot and frontal parts of the short-circuited rotor winding coils.

To investigate the ISEMC in idle and loaded modes, the system of equations of the mathematical model (1)-(3) is supplemented by the rotor's angular velocity equation and the electromagnetic torque equation [20]:

$$\begin{cases} \dfrac{d\omega_m}{dt} = \dfrac{M_e - M_c}{J}; \dfrac{d\phi}{dt} = \omega_m; \\ M_e = \dfrac{1}{R_o - R_i} \cdot \dfrac{l_z}{\mu_0} \int\limits_{S_{ag}} B_r B_\varphi \, dS_{ag}, \end{cases}, \quad (4)$$

where M_e is electromagnetic torque; M_c is load torque; J is rotor inertia moment; ω_m is rotor angular velocity; φ is rotor position angle; R_o is outer radius of the air gap; R_i is internal radius of the air gap; S_{ag} is cross-sectional area of the air gap.

The circuit implementation of the armature winding and the rotor winding, together with the equations of the mathematical model of the electromagnetic field, was carried out using the Comsol Multiphysics software with the use of a multiphysics structure combining the "Rotating Machinery, Magnetic" and "Electrical Circuit" interfaces. The connection and coordination of the parameters of the slot parts of the armature and rotor windings, which are elements of both the field model and the elements of the electrical circuit, is performed using "External I vs. U" terminals.

Considering the results of previous research by the authors [17], a U-shaped regulation characteristic of the ISEMC in idle operation was constructed based on the results of numerical calculation using finite element modeling for cases when the phase voltage is less than the nominal value (U_{phase}< U_{rated}) as well as ($U_{phase} \geqslant U_{rated}$) (Fig. 2).

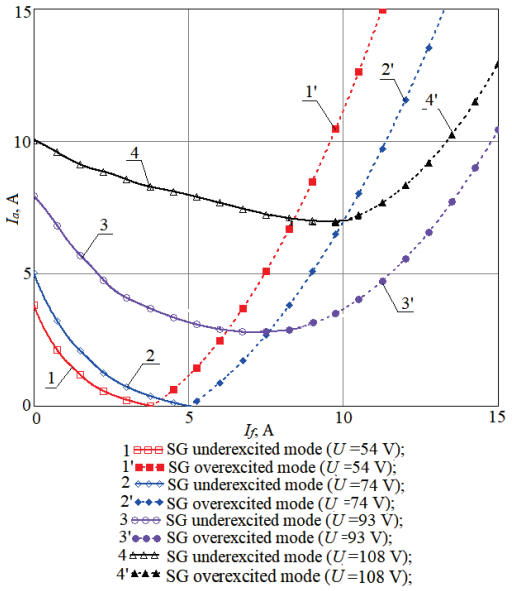


Fig. 2. General regulation characteristic at different values of the applied voltage to the armature winding of the experimental ISEMC in idle operation mode

The point of conditional symmetry of the U-shaped regulation characteristic of the ISEMC in idle operation will be determined by the condition of equality between the idle current I_{idle} of the IM part at the corresponding voltage U_{phase} and the field current I_f of the SG part of the ISEMC, that is I_f = I_{idle}. In this case, the SG part of the ISEMC operates in the normally excited mode. The left and right branches of the U-shaped regulation characteristic of the ISEMC will correspond to the underexcited and overexcited modes of the SG part of the ISEMC, respectively (fig. 2).

When U_{phase}<U_{rated}, the U-shaped regulation characteristic shifts to the left relative to the U-shaped regulation characteristic when U_{phase}=U_{rated} (U_{rated} =74V) (fig. 2). For this case, the normally excited mode of the SG part of the ISEMC will be achieved at a smaller value of I_f (I_f < $I_{f.rated}$), which will be equal to I_{idle} at the corresponding voltage U_{phase} (curve 1, fig. 2). In the overexcited mode of the SG, the armature current is generated by the SG part of the ISEMC and will already have active and reactive components, since E_{SG} > U_{phase} (curve 1', fig. 2).

When U_{phase} < U_{rated}, the U-shaped regulation characteristic shifts to the left relative to the U-shaped regulation characteristic when U_{phase} = U_{rated} (U_{rated} =74V)

(fig. 2). For this case, the normally excited mode of the SG part of the ISEMC will be achieved at a smaller value of I_f ($I_f < I_{f.rated}$), which will be equal to I_{idle} at the corresponding voltage U_{phase} (curve 1, fig. 2). In the overexcited mode of the SG, the armature current is generated by the SG part of the ISEMC and will already have active and reactive components, since $E_{SG} > U_{phase}$ (curve 1', fig. 2).

When $U_{phase} > U_{rated}$, the U-shaped regulation characteristic shifts to the right relative to the U-shaped regulation characteristic when $U_{phase} = U_{rated}$, and the normally excited mode of the SG part of the ISEMC will for $I_f > I_{f.rated}$. However, in this case, the EMF of SG part of the ISEMC will always be $E_{SG} < U_{phase}$. In this case, the reactive power of the IM part of the ISEMC will not be fully compensated, and therefore neither will the reactive component of the current I_{idle}.

The smallest value of the idle current I_{idle} for the U-shaped regulation characteristics when $U_{phase} > U_{rated}$ will be determined by the points of their intersection with the right branch of the U-shaped regulation characteristic at the nominal voltage value $U_{phase} = U_{rated}$ (curve 2', fig. 2).

With the appearance and subsequent increase of the load M_c, provided there is complete compensation of the reactive power of the IM part of the ISEMC (when $U_{phase} \leqslant U_{rated}$), meaning when the SG part of the ISEMC operates in the normally excited mode, the current I_a consumed from the supply network will only have an active component $I_{a.}|_{active}$. In this case, due to the combined effect of the electromagnetic torque from the IM part and the reactive torque of the SG part of the ISEMC, the synchronous rotational speed of the rotor will be maintained as long as the value of the armature current is less than or equal to the field current ($I_a \leqslant I_f$). In this case, the armature current I_a will always be greater than I_{idle} ($I_a > I_{idle}$), and the U-shaped regulation characteristic itself will shift upwards for all values of U_{phase}.

When $I_a > I_f$, the ISEMC rotor loses synchronism and begins to rotate with slip. With an increase in I_f, when the condition $I_a \leqslant I_f$ is met again, the ISEMC rotor is automatically pulled back into synchronism.

Thus, two stability criteria for the ISEMC can be defined. The first stability criterion for the ISEMC is that to ensure the operation of its SG part in underexcited and normally excited modes, the condition $I_f \leqslant I_{idle}$ must be satisfied, that is:

$$k\big|_{sc}^{1} = \left\{ I_f \leq I_{idle}\big|^{IM} \right\}. \tag{5}$$

This condition is a limitation on the field current I_f.

If $I_f > I_{idle}$, the SG part of the ISEMC transitions to the overexcited mode, which leads to the predominant effect of the SG part over the IM part of the ISEMC. This causes oscillatory processes in the resulting electromagnetic torque, as well as in the rotational speed of the ISEMC rotor, which leads to it losing synchronism followed by "tipping" [17].

The second stability criterion for the ISEMC is that to ensure the synchronization of the rotor under load, that is, to ensure the nominal synchronous speed of the ISEMC rotor, the condition $I_a \leqslant I_f$ must be satisfied:

$$k\big|_{sc}^{2} = \left\{ I_a \leq I_f \right\}. \tag{6}$$

When $I_a > I_f$, the ISEMC lacks the reactive synchronizing torque, which created by the SG part of the ISEMC to maintain the synchronization. This leads to the desynchronization of the rotor, the appearance of its natural slip, and an equalizing current, that is generated in the armature winding with a frequency $f\big|_{eq.c} = f_1 \cdot s$ [17]. The appearance of the equalizing current is a negative phenomenon, as its presence leads to an oscillatory process of currents in the parallel branches of each phase of the armature winding, and consequently to pulsations in the electromagnetic torque and the rotational speed of the rotor [17].

Based on the obtained data from the finite element modeling, the mechanical characteristic of the ISEMC was constructed under the condition of operation of only the IM part, as well as during the combined operation of the IM part and the SM part in the normally excited mode (Fig. 3). The curves 1, 3, and 5 depict the operation of the ISEMC without its SG part (without the field current of the SG) for phase voltage values of U_{phase} =54V, U_{phase} =74V, and U_{phase} =92V, respectively. In this case, regardless of the value of the applied voltage U_{phase}, with the presence of a load (M_c >0), the ISEMC rotor loses synchronism and begins to rotate with slip. With an increase in voltage ($U_{phase} \geqslant U_{rated}$), the value of the maximum torque ($M_{\max}$) increases, and the stiffness of the mechanical characteristic is enhanced, which will fully correspond to the mechanical characteristic of a conventional IM.

During the combined operation of the IM and SG parts of the ISEMC, when the SG part operates in underexcited or normally excited mode (that is, when the condition $I_f \leqslant I_{idle}$ is met), the stiffness of the mechanical characteristic at synchronous rotational speed is ensured by the condition $I_a \leqslant I_f$ (fig. 3, curves 2, 4, 6, at the corresponding phase voltage values U_{phase} =54V, U_{phase} =74V, and U_{phase} =92V). In this case, the maximum permissible value of the load torque M_c, at which the synchronous rotational speed of the rotor ($n = n_{syn}$) is maintained, will be determined by the condition $I_a = I_f$ at the corresponding value of U_{phase}.

Given the condition $I_a > I_f$, the rotor loses synchronism and begins to rotate with slip. In this case, the maximum torque value $M_{\max}$ at the corresponding voltage U_{phase} will occur at the critical slip ($s = s_{cr}$), which does not depend on U_{phase} and will also be equal to s_{cr} when only the IM part of the ISEMC is operating (see fig. 3). When the SG part of the ISEMC operates in u underexcited mode ($I_f < I_{idle}$) at a corresponding constant value of U_{phase} =c*onst.*, a family of mechanical characteristics can also be obtained, which will be located respectively between curves 1 and 2, 3 and 4, 5 and 6, that is, between the mechanical characteristic curves at I_f =0 and at $I_f = I_{idle}$.

To realize the overload capacity of the ISEMC, it is necessary to increase U_{phase} above its nominal value ($U_{phase} > U_{rated}$) (curve 6, fig. 3) while adhering to the stability condition $I_a = I_f$. However, in this case, the SG part of the ISEMC will

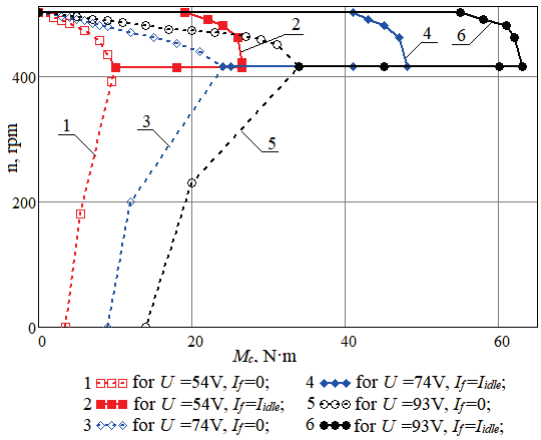


Fig. 3. Mechanical characteristic of the experimental ISEMC, when the SG part operates in the normally excited mode

no longer be able to fully compensate for the reactive power of its IM part, which will lead to the appearance of a reactive component of the armature current ($I_{a.reactive}$), and consequently to a significant decrease in $\cos(\varphi)$ and efficiency (η).

Using finite element modeling, a study was conducted on the starting characteristics of the ISEMC, and the values of the starting torque were obtained depending on the corresponding values of the applied voltage to the armature winding (U_{phase} =54V, U_{phase} =74V, and U_{phase} =92V) (fig. 3), as well as the permissible minimal $U_{phase,min}$ and maximum $U_{phase,max}$ values of the phase voltage of the experimental ISEMC for determining the range of change in the starting torque M_s (fig. 4). To ensure the starting of the ISEMC, its SG part must be disconnected (i.e., I_f =0), because when I_f >0, the SG part of the ISEMC creates a reactive synchronizing torque, that will counteract the electromagnetic torque created by the IM part of the ISEMC at the moment of starting. In this case, as with classical SEMC, asynchronous starting is ensured for the ISEMC.

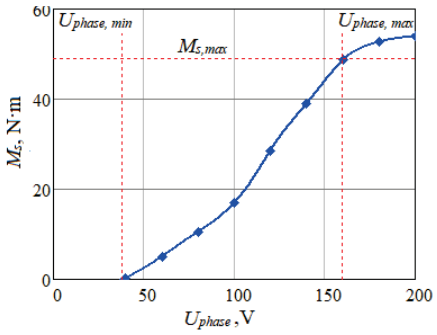


Fig. 4. Dependence of phase voltage on the starting torque of the IM part of the ISEMC

To ensure the maximum permissible starting torque, it is necessary to increase the voltage applied to the armature winding to $U_{phase,max}$, the value of which will correspond to the saturation of the ISEMC's magnetic system, at which the maximum value of the magnetic flux is ensured (see fig. 4).

In fig. 4, the minimal value $U_{phase,min}$ determinate the lower threshold of the phase voltage U_{phase} at which the idle starting of the ISEMC is performed. The determination of this value is important, as it ensures the minimization of the idle losses of the ISEMC.

Thus, when designing an ISEMC to ensure overload capacity, as well as a higher starting torque, it is necessary to calculate the value $U_{rated} << U_{phase,max}$, which will be achieved by the optimal ratio of the number of turns of the combined armature winding and the geometric dimensions and length of the armature core and rotor of the ISEMC.

Based on the results of the circuit-field modeling, the dependence of the armature current on the mechanical power of the ISEMC was obtained at constant operation of the SG part in the normally excited mode, i.e., under the condition $I_a = I_f$, $U_{phase} = U_{rated}$ (curve 1, fig. 5). In this case, the rated value of the ISEMC armature current and its rated power will be determined by the point of symmetry (at P_2 =2.15 kW) of the operating characteristic of the dependence $I_a = f(P_2)$ of the ISEMC at $I_a = I_f$ and $U_{phase} = U_{rated}$ (fig. 5, curve 1).

Under the condition of operation of the SG part of the ISEMC in the normally excited mode for the case, when $P_2 \leqslant P_{2rated}$, the armature current will have only the active component . For the case, when $P_2 \geqslant P_{2rated}$, the reactive power is not fully compensated, so the armature current will have active and reactive components:

$$I_a = \sqrt{I_a|^2_{active} + I_a|^2_{reactive}}. \quad (7)$$

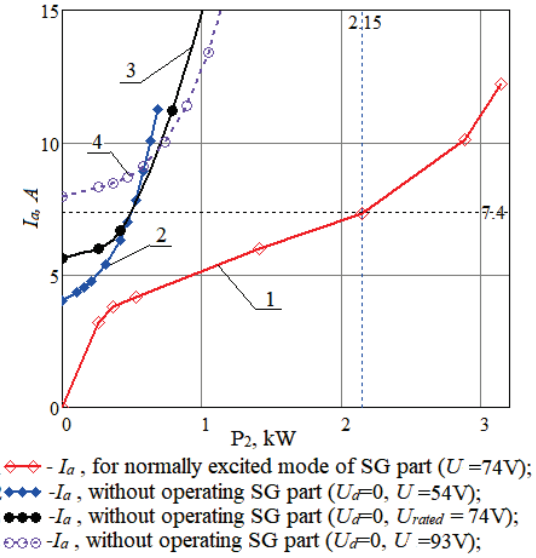


Fig. 5. Dependence of the armature current on the mechanical power of the ISEMC

With a further increase in power P_2 due to an increase in the applied voltage to the armature winding $U_{phase} > U_{rated}$, the content of the reactive component of the ISEMC armature current $I_{a.reactive}$ will increase according to the U-shaped regulating characteristic (see fig. 2). In the underexcited mode of the SG part of the ISEMC in the range of power change $P_2 \leqslant P_{2rated}$, the armature current also have a reactive component $I_{a.reactive}$ due to under-compensation of the reactive power of the SG part of the ISEMC. Fig. 5 also shows the dependences of the armature current on the mechanical power (curves 2, 3, 4) at different values of the applied voltage to the armature winding $U_{phase1} < U_{rated} < U_{phase2}$, when only the IM part of the ISEMC is operating. In this case, the above

dependencies fully correspond to the characteristics of traditional IM. In this case, the armature current will contain a significant part of the reactive component $I_{a.reactive}$ of the ISEMC armature current.

Thus, the issue of the ISEMC power gain in the motor operating mode has been resolved. Typically, according to the ISEMC calculation methodology [15], when determining the winding data, the number of turns of the ISEMC armature winding was determined by the criterion of reducing the idle current, especially for the generator mode of its operation. For the use of ISEMC as a generator, this is appropriate, since the U-shaped regulated characteristic will be shifted to the left as much as possible, which will reduce active losses from the field current in the generator mode. But for the use of ISEMC in the motor operating mode, this limits the power set, which has been established repeatedly in [11]-[13]. Therefore, when designing an ISEMC that will be used only in the motor operating mode, it is necessary that its U-shaped regulated characteristic be shifted to the right as much as possible to increase the power gain margin. That is, it is necessary to maximize the idle current I_{idle} at $U_{phase} = U_{rated}$, since this value I_{idle} determines the symmetry point of the U-shaped nominal regulated characteristic, which determines the normally excited mode of the SG part of the ISEMC, as well as the rated field current $I_{f.rated}$. The increase of the idle current I_{idle} is achieved by the optimal value of the number of turns w of the armature winding and the value of the rated phase voltage U_{phase}.

V. Conclusion

The stability criteria of the ISEMC in the motor operating in the idle and under load modes are determined. The defined stability criteria make it possible to ensure its trouble-free operation with an adjustable power set for the corresponding load while maintaining the cascade synchronous rotor speed.

Using the defined stability criteria of the ISEMC, the regulated properties were investigated and a family of regulated, mechanical characteristics and operating characteristics was obtained in terms of the dependence of the armature current on the mechanical power during the joint and separate operation of the IM and SG parts of the ISEMC motor operating mode.

According to the results of the study, the issue of power gain of ISEMC in motor operating mode was resolved. Recommendations for the selection of winding data for ISEMC depending on its purpose (motor, generator) are given.

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Optimization of Methods for Stabilizing of Glow Discharge in Welding

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***Abstract*—The ways of stabilization of high-current DC glow discharge at medium pressure of 1.33 – 13.3 kPa and discharge currents of 1 – 10 A while diffusion welding is considered. It is shown that the stable DC glow discharge can be provided by inclusion of the ballast resistor with the value of 60 – 80 Ohms in the discharge circuit. However, such a way of glow discharge stabilization is only effective against minor, short-term disturbances. It was also shown that another effective way to stabilize of high-current DC glow discharge in welding is the use of multi-sectioned anode system with the 3 and 5 sections accordingly.**

Keywords—glow discharge stabilization, diffusion welding, multi-section anode, ballast resistor

I. Introduction

Glow discharge at currents of 1 – 10 A and gas pressures of 1.33 – 13.3 kPa is effectively used as an energy source in diffusion welding processes of both homogeneous and heterogeneous materials due to its high energy efficiency, simplicity, and low equipment cost [1]. However, practical use of the glow discharge has revealed a significant drawback limiting its widespread application. Glow discharge as an intermediate form of gas discharges under certain conditions can lose its stability transitioning into a more stable form – an electric arc accompanied by a sharp increase in current [2]. As a result, the heating effect shifts from a dispersed to a locally concentrated one. The action of a concentrated arc discharge on welded samples can lead to melting and destruction of latter that is unacceptable.

Various technological and technical measures are used to stabilize glow discharge while welding. However, regardless of the stabilization measures applied, an active (ballast) resistor must always be included in the glow discharge circuit [3]. The presence of this resistor stabilizes the set burning mode of the discharge under conditions of minimal plasma volume fluctuations. The stabilizing effect of the external ballast resistor is in the fact that any increase in current in the discharge circuit above a steady-state value, including the loss of glow discharge's stability because of transition into an electric arc, results in a proportional increase in the voltage drop across the resistor [4]. Along with constant voltage at the output of power supply, this leads to a decrease in the voltage across the discharge gap and consequently a decrease in electron temperature and ionization rate in the discharge plasma. When the electric field strength in the discharge reaches a certain level, the electron generation rate becomes lower than the recombination rate leading to a decrease in electron concentration and a reduction in the discharge current to its original steady-state value [5].

In practical technological applications of glow discharge, there is a high probability of disturbances caused by the inhomogeneity of the physical and chemical state of the welded parts surfaces being heated by the discharge, local overheating of the gas in the plasma volume and other factors. Existing technical means for stabilizing glow discharge heating process normally respond only after an electric arc had been formed in the interelectrode gap and do not affect the physical processes in the discharge that lead to its transition into an arc [6]. Therefore, despite the large number of research and publications, the problem of stabilizing high-current glow discharge in technological processes such as welding, brazing, thermal and thermochemical metal treatment remains relevant.

The objective of current research is to analyze the processes that accompany the transition of a glow discharge into an electric arc and to improve the methods for stabilizing of high-current glow discharge under welding conditions.

II. Research Methods and Techniques

Theoretical research was based on the scientific principles of gas discharge physics, automatic control theory, and electrical engineering. For experimental investigations, a vacuum workstation for material sputtering equipped with a glow discharge's power supply system was used. Experimental installation scheme of glow discharge is given in Fig.1.

To obtain static current-voltage characteristics (CVC) of glow discharge the direct methods for measuring the main discharge parameters were used. The acquisition and analysis of dynamic characteristics of the steady state and transient glow discharge modes were carried out using a two-channel digital oscilloscope. The characteristics of the arc discharges that formed due to the loss of glow discharge stability were determined by examining the traces (autographs) they left on the sample surface when deflected by a transverse magnetic field.

III. Presentation of the Main Material

Normally, the heating process while welding is carried out using a normal glow discharge in which the cathode spot (heating zone) of the discharge is located on part of the welded samples' surface, which, in this case, act as the cathode of the discharge.

According to the physical literature [7], the CVC of such a discharge is typically considered horizontal, i.e., the

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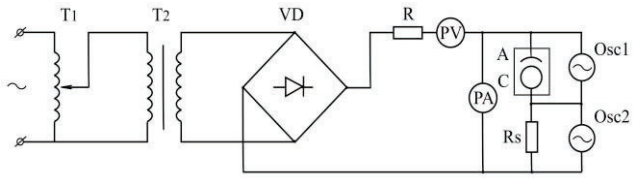


Fig. 1. Electrical scheme of the glow discharge's power supply: *T1* – autotransformer; *T2* – step-up transformer; *VD* – rectifier; *R* – ballast resistor; *Rs* – shunt resistor; *PA, PV* - measuring devices; *A* – anode; *C* – cathode; *Osc* – oscilloscope

discharge voltage is independent of the current. However, under welding conditions with elevated currents of 1 – 10 A and above, the CVC of the discharge becomes slowly falling when the voltage decreases slightly with increasing current. This is a consequence of gas heating in the interelectrode gap and increasing its electrical conductivity. Nevertheless, the descending branch of the CVC is unstable with respect to small deviations. This is explained by the fact that any increase in current relative to the steady state on the descending CVC's branch leads to a further increase in discharge current intensifying the gas heating, which in turn, increases its electrical conductivity. This leads to a reduction in discharge resistance and a subsequent rise in current.

A second critical parameter of the glow discharge's heating mode is the gas pressure in the working chamber, which determines the current density in both the positive column and the cathode spot of the discharge. Increasing the intensity and efficiency of heating process requires higher gas pressure, which leads to a significant radial non-uniformity in the discharge. The current density increases most rapidly along the discharge axis, where the gas is most intensively heated. This can cause contraction (pinching effect) of the discharge and its transition to an electric arc [8].

In both cases, local increases of gas ionization degree within the discharge gap result in increased current and a loss of stability in the preset glow discharge mode, potentially causing its transition to an arc discharge. The power supply's circuit of the glow discharge can be represented as a basic electric scheme that is made up of direct current power source with the open-circuit voltage *E*, an active (ballast) resistor *R*, the total circuit inductance *L* and the glow discharge itself burning between anode *A* and cathode *C* with current *I* and voltage drop *U(I)*. The burning mode of the glow discharge is determined by the point of intersection of the discharge CVC's with the load line $E-IR$ (Fig.2). The electric circuit in steady-state condition can be described by the following equation:

$$E = U(I) + IR + \frac{LdI}{dt} \tag{1}$$

Assuming at a steady state conditions ("*a*" on Fig. 2), we have $U_d = U_0(I_0)$. To assess the stability of the discharge with respect to point "*a*", we will use the method of small deviations of the current δI relative to the steady state I_0 [9]. Then the voltage on the discharge at the new point will have the form of $U(I) = U(I_0 + \delta I)$. Due to the small deviation of δI, when expanding the function $U(I_0 + \delta I)$ into a Taylor series, it is possible to confine with the first two terms:

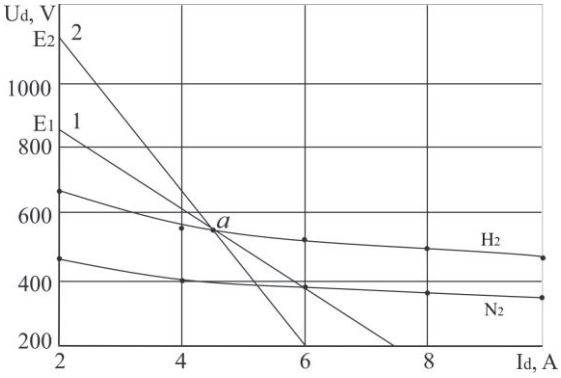


Fig. 2. Static current-voltage characteristics of a glow discharge in a nitrogen and hydrogen plasma and conditional load lines *E–IR* (1 and 2) for different values of ballast resistor

$$U(I_0 + \delta I) = U_0(I_0) + (\frac{\partial U_0}{\partial I_0})\delta I + \ldots \tag{2}$$

In expression (2), the component $\frac{\partial U_0}{\partial I_0} = R_d$ represents the differential discharge resistance.

For the transient process, equation (1) can be rewritten as:

$$L\frac{d(\delta I)}{dt} = E - U_0(I_0) - I_0 R - (R_d + R)\delta I \tag{3}$$

Considering that during stable glow discharge, i.e. in a steady state, the first three components on the right-hand side of the last expression sum to zero, it can be represented by the differential equation:

$$\frac{d(\delta I)}{dt} = -(R_d + R)\frac{\delta I}{L} \tag{4}$$

The solution of this equation is:

$$\delta I = \delta I(0)\exp\left(-\frac{R_d + R}{L}t\right) \tag{5}$$

where $\delta I(0)$ is the initial maximum value of current deviation.

The differential resistance (R_d) is negative because an increase in current along the descending branch of the CVC leads to a decrease in voltage drop across the discharge. Therefore, to stabilize the specified glow discharge mode, i.e., to ensure that current deviations in the system diminish over time $\delta I \to 0$ the following condition must be satisfied:

$$R_d + R > 0 or R > |R_d| \tag{6}$$

The absolute value of the glow discharge's differential resistance (R_d), determined from experimentally obtained static CVC's in the current range of 1 – 10 A for various gases, is approximately 30 – 40 Ohms. To ensure stable glow discharge operation, the ballast resistor must be 60 – 80 Ohms, which aligns with values proposed in studies [10, 11] and used in practice. However, the experience of practical application of glow discharge under conditions of high current and gas pressure shows that such ballast resistor values are only

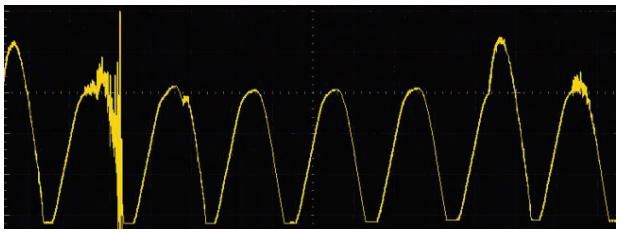
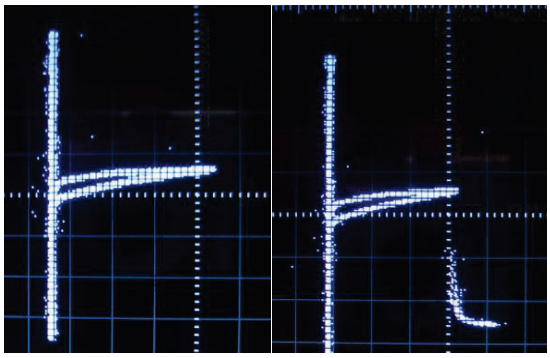


Fig. 3. Occurrence of short-term arc breakdowns in a glow discharge



a) b)

Fig. 4. Dynamic current-voltage characteristics of the glow discharge (a) and the glow discharge – electric arc transition zone (b)

effective against minor, short-term disturbances. In these cases, the formation of a brief arc discharge may eliminate the disturbance (Fig. 3). These disturbances could include fine hair-like projections or localized contaminants on the samples surface, which are melted or vaporized by the arc.

In case of significant disturbances associated with the structure of the welded joint or its thermal deformation during heating, such a level of ballast resistor becomes insufficient to prevent the avalanche development and stabilization of an arc discharge in the interelectrode gap.

The required value of ballast resistor can be estimated by analyzing the dynamic characteristics of the transient process between glow discharge and arc, obtained using a two-channel oscilloscope (Fig. 4). The analysis results show that the transient process is accompanied by a decrease in the voltage in the discharge gap of 150 – 300 V (approximately from 350 – 500 V in a glow discharge to 100 – 250 V in an electric arc). Moreover, this voltage drop increases with the discharge current.

The oscillograms indicate that the transition occurs with only a slight deviation of the discharge current from its steady-state value. Hence, the differential resistance of the discharge $R_d = \frac{dU}{dI}$ in the transition mode significantly increases and reaches several kilo ohms. To satisfy the stability condition (6), the ballast resistance must be of a similar magnitude. However, according to the load line $E - IR$ (line 2 in Fig. 2), incorporating such a high resistance requires increasing the open-circuit voltage of the discharge power supply to several thousand volts. This significantly complicates the power supply's design, operating conditions and safety measures. Therefore, such a method of stabilizing high-current glow discharge cannot be considered efficient or technically feasible.

To analyze other approaches to stabilizing a glow discharge under disturbances, let us consider the discharge-arc

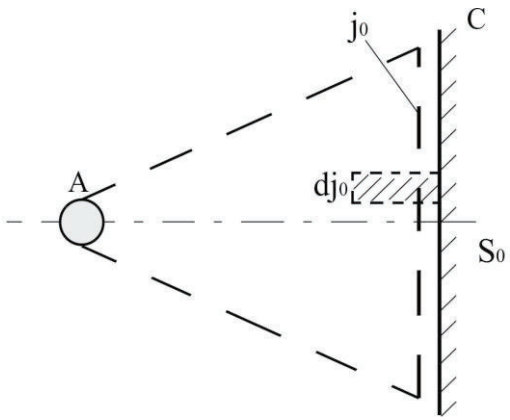


Fig. 5. Scheme of instability formation in a glow discharge: *C* – cathode; *A* – anode; S_0 – cathode heating spot; j_0 – glow discharge's current density; dj_0 – fluctuations in a discharge`s current density

transition process. The initiation of an arc discharge is accompanied by the emergence of a local fluctuation in current density dj_0 against the background of a uniform glow discharge with an average current density j_0 and cross-sectional area S_0 (where S_0 can be taken as the area of the cathode spot, i.e., the heating zone) (Fig. 5). This causes the system transition to a new state characterized by a voltage drop dU and an increase in current dI in the interelectrode gap.

The expression for the differential resistance of the discharge in this transitional state can be written as:

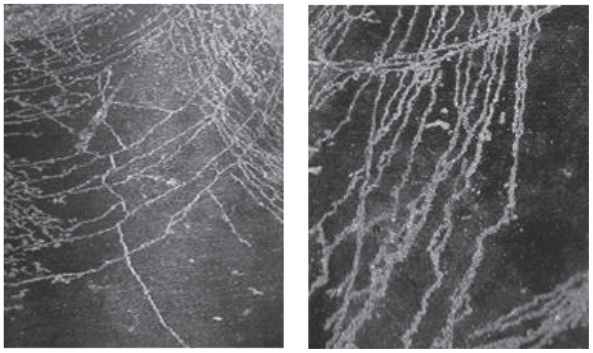
$$R_d = \frac{dU}{dI} = \frac{dU}{S_0 dj_0} \tag{7}$$

Since the values of the differential discharge resistance and the external ballast resistance are commensurate, we replace R_d in expression (7) with R_b. Then we can write expression (7) as follows:

$$S_0 R_b = \frac{dU}{dj_0} \tag{8}$$

From the previous equation, it follows that for a given constant value of ballast resistance (R_b) in order to reduce the voltage drop dU and the differential resistance of the discharge in the transitional state accordingly it is necessary to reduce the area of the discharge cathode spot S_0. The value of δj_0 can be considered constant since the arc autographs (traces) formed during the transition from glow to arc discharge, observed on the sample surface under magnetic field deflection, indicate that as the power of the glow discharge increases, the size of the arc cathode spot also increases (Fig. 6).

As was previously noted, the value of dU depends on the power of the glow discharge and decreases as the current is reduced. The area of the cathode spot in a normal glow discharge, in turn, directly proportional to the discharge current. Therefore, it follows that the conditions for stabilizing glow discharge using an external ballast resistance are more effective for low-power discharges, which are characterized by relatively limited current values. To ensure the stable existence of a high-current glow discharge, it is advisable to divide it into several low-power discharges, the total current and cathode spot area of which are equivalent to those of the original high-power discharge. Such a stabilization approach



a) b)

Fig. 6. Traces of an electric arc arising in a glow discharge at a discharge current of 3 A (a) and 7 A (b), while deflected by a magnetic field

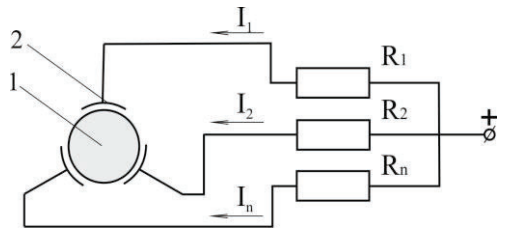
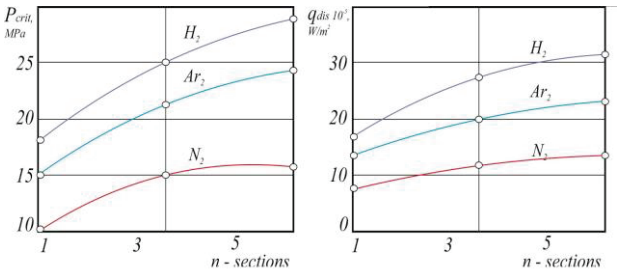


Fig. 7. Heating scheme with multi-sectioned anode: 1 – cathode; 2 – multi-section anode



a) b)

Fig. 8. The effect of anode sectioning on increasing the stability limit of the glow discharge in terms of gas pressure (a) and specific power (b) in different gases: 1 – monolithic contour anode; 3 – three-section anode; 5 – five-section anode

can be implemented by dividing the continuous contour anode of the discharge into several individual segments (sections), each connected to the power supply output through its own ballast resistor (Fig. 7). This configuration corresponds to a parallel connection of ballast resistors, meaning that increasing the number of anode segments allows the resistance values to be raised to the kilo ohm range. As has already been shown, this contributes to improved discharge stability under conditions of high current and gas pressure.

Experimental verification of this hypothesis was carried out by heating cylindrical steel workpieces using glow discharge in nitrogen, argon, and helium. The tests were conducted using both a continuous contour anode and segmented one with 3 and 5 sections accordingly. In the first case, a ballast resistor of 100 Ohms was connected in the glow discharge circuit. In the other two configurations, resistors of 300 and 500 Ohms were connected to each anode segment, respectively. The experimental results (Fig. 8) show that the proposed heating scheme significantly shifts the stability loss threshold of the glow discharge toward higher energy characteristics. Furthermore, even if a low-power arc forms on any section of the cathode, it is absorbed by the corresponding anode segment and has no effect on the discharge behavior in the other segments.

This method allows the usage of ballast resistor for stabilization of a glow discharge more effectively. To eliminate the possibility of an arc formation in the interelectrode gap such number of anode sections should be selected that the current in each segment does not reach the minimum value at which a cathode spot of the arc can form on the cathode (workpiece) surface i.e., the physical conditions for stable arc existence must not be met. For most metals (copper, titanium, iron, tungsten, etc.), the threshold current for arc cathode spot formation does not exceed 1.2 – 1.6 A [12]. Therefore, it is advisable that the current in each anode section not exceed 0.8 – 1.0 A.

IV. Conclusions

Analysis of the dynamic characteristics of the glow-to-arc discharge transition process shows that the value of the ballast resistance required to prevent such transitions and stabilize glow discharge combustion must reach units or tens of kilo-ohms. However, this would require increasing the discharge's power supply voltage to tens of thousands of volts, which is technically unacceptable.

It has been theoretically demonstrated and experimentally confirmed that it is possible to significantly increase the stabilizing ballast resistance, and thus the stability of high-current glow discharge, without increasing the supply voltage, by dividing the discharge into several low-power discharges operating simultaneously in the same working space. This was achieved by segmenting the continuous contour anode, where each individual anode segment is connected to the power source output through its own high-value ballast resistor.

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Using an Uninterrupted Power Supply to Reduce the Operating Costs of an Autonomous Electric Generator

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***Abstract*—An analysis of the use of an uninterruptible power supply in combination with an autonomous diesel/gasoline generator was conducted in order to reduce operating costs. Such systems can be used when using an autonomous power generating unit as the only power source for consumers, or in case of emergency or emergency power outages. An analysis of existing autonomous power supply systems was conducted and the relevance of this research direction was substantiated. Load schedules were constructed based on a specific example of an autonomous power generating unit and a certain list of equipment in a private house, depending on the time of day. The developed algorithms were used to assess the effectiveness of using a UPS in conjunction with an autonomous diesel/gasoline generator. The use of a UPS allows you to reduce operating costs by loading the generator to an optimal value of 75-80% of the nominal, saving fuel, reducing maintenance intervals and frequency, and harmful emissions into the atmosphere. The maximum amount of savings, depending on the system operating mode, load nature, and frequency of outages, is from 39% to 76%. Additional energy saving effect can be achieved if this system is additionally equipped with solar panels or a wind generator. This will allow for additional fuel savings and generator life. It is planned to evaluate the effectiveness of this technical solution in future studies.**

Keywords—autonomous electric generator, uninterruptible power supply, autonomous diesel-petrol power plants, renewable energy, electromechanical converter.

I. Introduction

Autonomous power generation units (diesel/gasoline) are vital for various facilities during outages, maintenance, unstable supply, and as backup, especially for remote or autonomous entities [1]. Operation incurs fuel, maintenance, and repair costs. Integrating uninterruptible power supply (UPS) with generators [2] ensures power continuity and significantly reduces operational costs by decreasing generator load and optimizing fuel consumption [3], while enhancing reliability through seamless power source transitions [4, 5]. Although inverter-based systems can have higher initial costs, their integration with generators provides fuel economy, extended maintenance intervals (saving cost/time), longer generator lifespan, and continuous AC power. This study explores innovative approaches to improve energy efficiency and reduce operational costs in autonomous generator systems. UPS integration minimizes low-load generator operation (reducing wear/fuel overuse and emissions [6, 7]), applicable across sectors (industry, healthcare, IT, residential). Literature [8, 9] indicates up to 20% fuel consumption reduction with UPS, and further studies [10, 11] confirm enhanced reliability and reduced outages. The primary goal is to shorten generator runtime, especially during peak loads or brief outages, or to load generators efficiently (75-85%) via inverters. Advanced UPS units provide hours of backup for critical loads [12]. Challenges include cost and complexity, requiring additional equipment and maintenance, though automation improves efficiency and reduces human error [13]. Economically, fewer starts extend engine life and lower maintenance [14]. Thus, UPS integration in autonomous systems offers significant potential for cost reduction and reliability improvement, representing a timely technical and scientific issue.

Aim of the Study. The aim of this study is to evaluate the effectiveness of utilizing a UPS within an autonomous power supply system based on an internal combustion engine electric generator. This will enable the optimization of autonomous power generation units, the reduction of operational expenditures, and the enhancement of system reliability.

Fig. 1 presents the structural diagram of an UPS system specifically designed for electrical installations. The diagram meticulously details the essential components that constitute such a system and clarifies their intricate interconnections. At the core of this configuration lies the UPS unit itself, which assumes the critical role of ensuring a continuous and stable power supply to connected equipment, particularly during instances of primary power source disruptions or failures. The illustration typically depicts the flow of power, control signals, and the integration of various elements that contribute to the system's overall functionality and reliability. The foundation of this sophisticated UPS system lies in the selection of a contemporary UPS unit, with a range of reputable manufacturers such as Victron Energy, Deye, Must, and Daxtromn offering suitable models. A key characteristic of these modern UPS devices is the inclusion

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of a communication interface, notably the RS485 standard (though other protocols may also be present). This communication port is crucial as it unlocks advanced configuration and customization functionalities for the power converter's operational modes. Engineers and system integrators can leverage this interface to fine-tune various parameters, optimize the UPS's performance characteristics to match specific load requirements, and potentially integrate it with broader energy management or monitoring systems. Furthermore, to achieve additional energy conservation in the proposed system, it can be strategically augmented with renewable energy generation sources, such as photovoltaic solar panels or a wind turbine. The integration of these clean energy technologies offers a dual benefit. Firstly, it reduces the system's reliance on conventional power grids or fossil fuel-powered generators, directly translating into tangible fuel savings, particularly in scenarios where a backup generator forms part of the extended power infrastructure. Secondly, by supplementing the primary power supply with energy harvested from renewable sources, the operational duty cycle and consequently the overall lifespan of any associated electric generator can be significantly extended. This reduction in usage translates to lower maintenance requirements, decreased wear and tear, and ultimately, reduced long-term operational costs. Looking ahead, the authors plan to practically realize and validate the efficacy of this proposed technical solution in their future research endeavors. This implementation phase would likely involve the physical construction and rigorous testing of a prototype UPS system integrated with advanced control mechanisms and potentially incorporating renewable energy sources as described. The subsequent work will then focus on evaluating the system's performance under real-world operating conditions, quantifying its energy efficiency gains, assessing its reliability and stability, and analyzing the economic and environmental benefits associated with such an integrated power supply architecture. The findings from this practical implementation will be crucial in validating the theoretical concepts and paving the way for potential wider adoption of this technology.

Under normal operation, the system channels mains electricity through a contactor, supplying both the UPS and the connected electrical load. Simultaneously, the Battery Management System (BMS) actively monitors the battery's state of charge, precisely controlling the charging and discharging cycles to ensure optimal battery health and longevity. This panel also enables authorized personnel to perform system configuration and adjust operational settings as needed. In the event of a mains power interruption, the system intelligently initiates a generator start sequence based on a comprehensive assessment of the battery's remaining capacity and the current power demand of the connected load. Following a brief warm-up period to ensure stable operation, the system seamlessly switches the power source from the UPS battery to the auxiliary generator via the Automatic Transfer Switch (ATS) contactors (identified as component 4 in Fig. 1 of the accompanying diagram). The specific components strategically chosen and implemented to achieve cost reduction within this overall system design are detailed and systematically listed in Table 1.

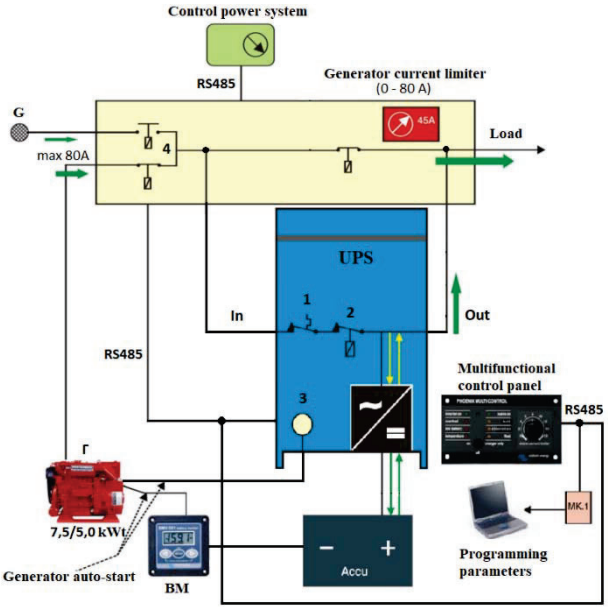


Fig. 1. Configuration of an power supply system based on a UPS: 1 – built-in thermal relay; 2 – protective (emergency) relay; 3 – multifunctional additional UPS input; 4 – input switch (ABA system); G – autonomous generator; BM – battery monitor system

TABLE I. MAIN COMPONENTS OF THE PROPOSED SYSTEM

№	Equipment Designation.	Specifications.	Cost, $
1	Victron MultiPlus 24/3000/70	Power: 2.4 kW; Input Voltage: 187 – 265 V	1350
2	Traction Battery Sunlight (10 PzS 600) LiFePO4.	Rated Voltage: 24 V; Rated Capacity: 600 Ah.	3030
3	SDMO HX 7500 T C	Full Power: 7.5 kVA Rated Active Power: 6.0 kW Fuel Consumption: 2.3 L/h at 75% load	2000
4	Multifunctional Control Panel	Victron Energy GX Touch 50	240
5	Battery State Monitoring System	Battery Monitor BMV-712 Smart 9 - 90B	215
6	Total		6835

Based on the components listed in Table 1, a subsequent calculation of the operational cost reduction effect of the autonomous power generator is performed.

II. MATERIALS AND RESEARCH FINDINGS

To evaluate the operational cost reduction effect, a typical list of equipment for a private house (apartment or townhouse) will be considered. The equipment list is presented in Table 2.

TABLE II. EQUIPMENT LIST

№	Devices	Power, W	Quantity
1	Lighting	20	15
2	TV	120	1
3	Fridge	140	1
4	Water heater	2000	1
5	Submersible well pump	800	1
6	Washing machine	1500	1
7	PC	500	1
8	Router and video surveillance cameras	50	3
9	Vacuum cleaner	800	1
Total			≈5930 W

Figure 2 illustrates a typical consumption profile of the equipment as a function of the time of day, according to the data in Table 2.

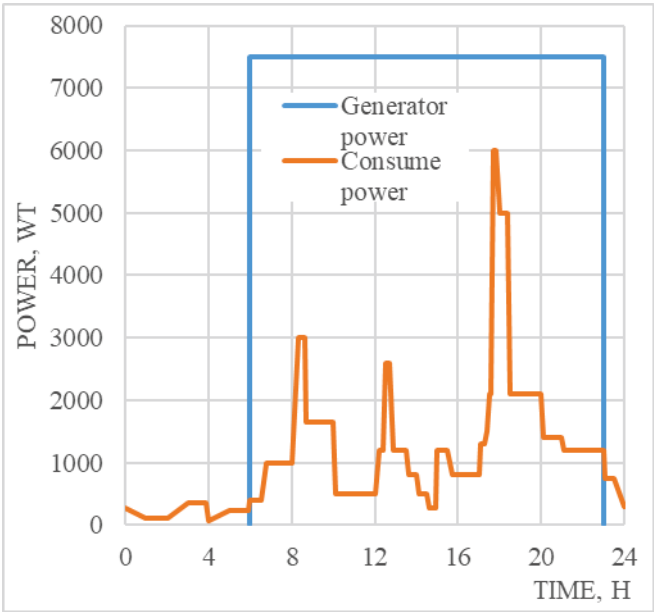


Fig. 2. System using 7.5 kVA generator only

Fig. 2 reveals a morning consumption peak of 3 kW lasting for two hours, a midday peak of approximately 2.5 kW, and an evening peak of 6 kW. Between the morning and midday peaks, and between the midday and evening peaks, consumption is relatively low, averaging around 500 W. Following the evening peak, the average consumption is approximately 1 kW for several hours.

Considering scenarios involving systematic power outages or complete failures of the main electricity supply, and with the assumption that an electric generator with a capacity of 7.5 kilovolt-amperes (kVA) serves as the sole source of power, the generator would be required to operate for a significant duration of 17 hours per day to meet the energy demands of the connected electrical installations. It is particularly noteworthy that for a substantial portion of this operational time, specifically 70% of the 17-hour period, the generator functions under a very light load, consuming less than 15% of its maximum rated power capacity. This indicates a potentially inefficient mode of operation for a considerable duration.

If a UPS with batteries of the necessary type and capacity is added to the system, it can provide the required power for the majority of the equipment's operating time. During peak consumption, the generator automatically starts, and the excess power is used to charge the batteries.

Fig. 3 illustrates the consumption profile of the equipment with the addition of a UPS-type inverter (using the Victron MultiPlus as an example).

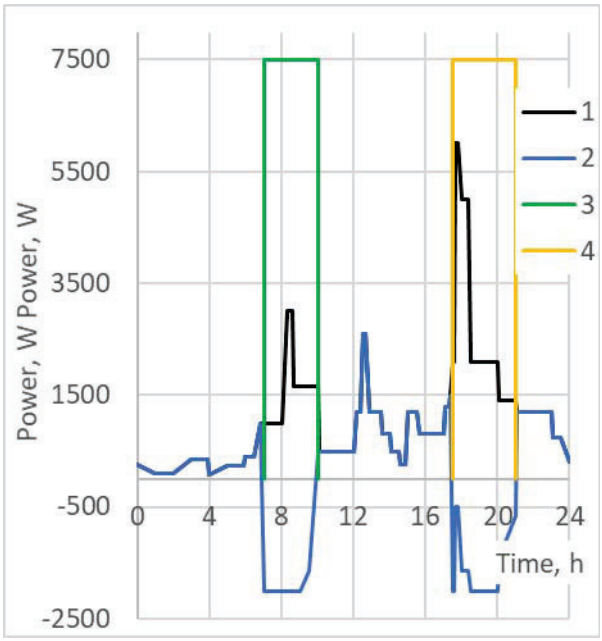


Fig. 3. Load schedule when using a UPS in the system: 1 – consumed load; 2 – UPS power; 3, 4 – generator power

Generator runtime is reduced to 5 hours/day as battery meets most power demands (≤ 2.5-3.0 kW). Generator surplus charges batteries (limited by surplus and inverter capacity, max 2.0 kWh in this example). Battery powers 12 hours of silent operation. Generator runs at up to 70% capacity.

Required battery capacity is lower than without a generator (24 V chosen). Calculation shows discharge to 166 Ah before morning start and 285 Ah before evening start. Formula-based calculation suggests ~658 Ah, thus 24 V/600 Ah commercial batteries are sufficient. A 48 V system would require ~300 Ah.

A 7.5 kVA generator consumes ~1.5 L/h diesel at low load (≤ 25-30%) and 2.3 L/h at 75% load. Standalone operation yields ~25.5 L/day consumption. UPS integration reduces this to 11.5 L/day for the same power, saving 5110 L/year, equating to ~$6643 annually (excluding refueling savings). Detailed operational costs are in Table 3.

Manufacturer recommends oil changes every 150 operating hours (mineral oil). Standalone generator (6205 hours) requires 41 annual changes. UPS integration (1825 hours) reduces this to 12, saving 29 changes/year. With 5 L oil volume, this saves 145 L/year, equating to $725 annually (excluding time/filter costs).

During morning/evening peak emergency outages, the system saves 2070 L fuel/year, ~$2691 annually. With UPS during 8h power, 16h outage (generator 2.5h/day at peak), savings are ~76%. For standalone or systematic outages with UPS, operational cost reduction is ~39%.

Considering equipment/installation costs, payback is ~1-1.5 years for standalone generator replacement and ~2.5-3 years for emergency outage scenarios with UPS.

TABLE III. CALCULATION OF OPERATING COSTS

	Generator only	**Generator + UPS**	**Economy**
Operating hours h/day	17 h	5 h	12 h
Diesel consumption/day	25,5 L	11,5 L	14 L
Annual diesel consumption	9307 L	4197 L (2070L)	5110 L 6643 $/year
Annual oil change	41	12	29
Annual lubricant consumption	205 L/year	60 L/year	145 L/year, costing $705/year
Generator load	15% load for 70% of operating hours	70 %	
Annual fuel and lubricant savings (excluding threefold longer lifespan)	7348 $/year		2691 $/year
Savings over 5 years	36740 $		

Diesel generators have longer lifespans under load (vs. no/low load). Some manufacturers advise ≥ 25-30% minimum load to prevent failure. UPS integration increases generator load from < 15% to 70%, further extending lifespan.

Operating generator with UPS reduces operational costs, specifically the cost per kWh (illustrated in Fig. 4).

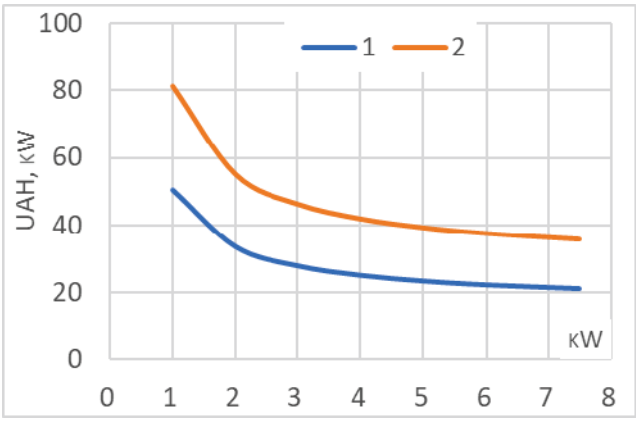


Fig. 4. Cost of electricity based on the output power of the electric generator: 1 – for a 7500 kVA generator with a diesel drive engine; 2 – for a 7500 kVA generator with a gasoline drive engine

CONCLUSIONS

The combined use of an UPS and an autonomous diesel/gasoline generator offers not only silent power delivery but also improved fuel efficiency and reduced maintenance costs. During regular power outages, this integration can lower operational costs by approximately 39%, and by up to 76% in emergency scenarios. These savings result from limiting generator use to 2.5 hours during peak demand, with the remainder supplied by rechargeable batteries, which can be charged from both the grid and the generator. Additionally, this setup minimizes environmental impact through reduced emissions and motor oil waste.

Cost savings depend on factors such as generator capacity, outage schedules, and load characteristics. This study estimates potential maximum savings, noting that actual results may vary. Operational efficiency improves with higher peak consumption level.

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Hybrid Neural Model for Cybersecurity Risk Assessment Based on Lotka–Volterra Dynamics

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***Abstract*—This paper proposes an approach to cyber risk assessment based on a hybrid model. This model combines the capabilities of deep neural networks for network traffic classification and the prediction of parameters for the Lotka-Volterra dynamical system. Such an approach enables the identification of anomalous activity and modelling interactions between attacking and defensive forces in cyberspace, thereby providing a deeper understanding of potential threats. The model was trained and tested on the NSL-KDD dataset. The results demonstrate that the model can effectively classify traffic (accuracy of 0.8006, AUC of 0.9016 on the test set) and predict the coefficients of the Lotka-Volterra model (Mean Squared Error of 0.0027). Analysis revealed that the predicted coefficients of the Lotka-Volterra model exhibit significant differences for normal and anomalous activity. The research of these coefficients and the simulation of the corresponding dynamical systems reveal distinct patterns that correlate with the level of cyber risk, opening new avenues for proactive assessment and management of cyber threats.**

Keywords— cyber risk assessment, neural networks, Lotka-Volterra dynamics, intrusion detection, NSL-KDD, cyber threat modelling

I. Introduction

The relentless growth in the number and complexity of cyber threats in the modern digital landscape poses a critical challenge for organisations in effectively assessing and managing cyber risks [14; 2]. Traditional risk assessment approaches are often static and fail to account for the dynamic interaction between a system's attacking and defensive mechanisms [5]. This leads to a limited ability to predict threat evolution and respond to them adaptively.

Overcoming these limitations requires the development of more sophisticated models capable of capturing the complex dynamics of cyber confrontations. The use of models from other scientific fields is one promising direction. In particular, ecological predator-prey models, such as the Lotka-Volterra model, offer a mathematical framework for describing interdependent populations [15; 12]. Such a model can be adapted to describe the interaction between "attackers" (analogous to predators) and "defensive forces" (analogous to prey or a reactive system) in a cyber environment [15]. The Lotka-Volterra model allows for abstraction to key aspects of the confrontation (e.g., attack propagation rate, effectiveness of defensive measures, system response to intrusion, and the "cost" of maintaining defense), making it attractive for the conceptual modelling of cyber risks.

In this study, we propose a hybrid neural model that integrates network traffic classification for intrusion detection with predicting parameters of the Lotka-Volterra dynamical model. The neural network is trained using network traffic characteristics to distinguish between normal and anomalous activity and estimate the coefficients that determine the interaction dynamics between a potential attack and the system's defense. This approach allows for a transition from merely stating the occurrence of an attack to assessing its potential development and the system's resilience.

This research aims to develop and evaluate the effectiveness of the proposed hybrid model for cyber risk assessment and demonstrate that the predicted parameters of the Lotka-Volterra model form distinct patterns for different types of network activity, thereby enabling qualitative and quantitative evaluation of the cyber risk level.

II. Methodology

The scientific community actively researches the problem of intrusion detection and cyber risk assessment. Machine learning methods, particularly neural networks [4; 8], have found wide application due to their ability to detect complex dependencies in data. Datasets such as KDD Cup'99 and its improved version, NSL-KDD, are frequently used for training and testing intrusion detection systems [10; 13]. Existing approaches based on neural networks demonstrate high accuracy in network traffic classification [1; 7]. However, they are typically limited to stating the presence or absence of an attack.

In assessing cyber risks, the objective extends beyond mere attack detection to include the quantitative or qualitative evaluation of potential harm and its likelihood [3]. However, many existing risk assessment methodologies do not consider

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the dynamic nature of cyber threats and the interaction between the attacker and the defense system.

Attempts to model the dynamics of cyber attacks and defenses exist. Still, they are rarely integrated directly with intrusion detection systems at the level of model parameterization based on real traffic data. Lotka-Volterra models, describing "predator-prey" type interactions, have been proposed for analyzing malware propagation and other aspects of cybersecurity [11; 12]. These models allow for investigating system stability and the conditions under which an attack can be successful or suppressed [12; 15]. However, a key challenge remains the robust estimation of parameters for these models based on fundamental characteristics of the cyber environment. Our work aims to bridge this gap by using a neural network to predict the coefficients of the Lotka-Volterra model directly from network traffic data.

In developing and evaluating the proposed hybrid cyber risk assessment model, we relied on a carefully selected dataset, adapted a classical dynamical system to the specifics of the subject domain, and utilized modern deep learning methods for model implementation.

Our research used the NSL-KDD dataset [10; 13], an improved version of KDD Cup'99. The NSL-KDD dataset, as noted in [13], was proposed to address some of the problems inherent in the KDD'99 dataset. The advantages of NSL-KDD over the original KDD dataset include the absence of redundant records in the training set, which prevents classifier bias towards more frequent records, and the absence of duplicate records in the test sets, ensuring unbiased performance evaluation. Furthermore, a more balanced distribution of records by difficulty level allows for a more accurate assessment of different learning methods, and a reasonable number of records in the training and test sets enables experiments to be conducted on the complete dataset, ensuring consistency and comparability of research results. Although NSL-KDD may not be an ideal representative of existing real-world networks, according to criticism in [10], it is widely used as an effective benchmark dataset.

The dataset contains files in ARFF format, with 41 features and a class label characterizing each. Basic attributes include connection duration, protocol type (e.g., TCP, UDP, or ICMP), and service (e.g., HTTP, FTP, or SMTP). Other essential attributes include the connection status flag, the volume of transmitted bytes (src_bytes and dst_bytes), as well as various features describing error statistics (e.g., serror_rate and rerror_rate) and time-based traffic characteristics (e.g., count and srv_count). For our study, class labels were aggregated into a binary form: "normal" and "anomaly".

The classical Lotka-Volterra model [12; 15] describes the dynamics of two interacting populations. In the context of cybersecurity, drawing upon previous works [12; 15], we interpret these populations as follows:

$A(t)$ The activity level or intensity of "attacks" is analogous to that of predators.

$Z(t)$ The level of "security" (or protection) or the resources of the security system, analogous to prey or the reactive defense capability of the system.

The Lotka-Volterra system of differential equations, as used in this study, is as follows:

$$\frac{\mathrm{dA}}{\mathrm{dt}} = \alpha \mathrm{A} - \beta \mathrm{A}\, \mathrm{Z}$$

$$\frac{dZ}{dt} = \gamma AZ - \phi Z$$

where the coefficients $\alpha, \beta, \gamma, \phi$ are interpreted and calculated from network features as follows.

Coefficient α represents the intrinsic growth rate of attacks and characterizes the inherent ability of attacks to propagate or emerge without effective countermeasures. A high value α may indicate an aggressive threat nature or a favorable environment for its development. Research calculated this coefficient as the fraction of network sessions exhibiting signs of complex or "hybrid" attacks. Such sessions were identified by the simultaneous exceeding of the 75th percentiles $Q_{0.75}(\cdot)$ for synchronization error rate (serror_rate) and rejection error rate (rerror_rate), which may indicate attempts to bypass defenses or use multiple attack vectors. The calculation is as follows:

$$\alpha = P\big(\text{serror_rate} > Q_{0.75}(\text{serror_rate}) \wedge \text{rerror_rate}. > Q_{0.75}(\text{rerror_rate})\big)$$

The parameter β determines the effectiveness of the defense in neutralizing attacks, reflecting how successfully defensive mechanisms reduce attack intensity through interaction. A high value β implies a strong defensive response per unit of attack. This coefficient was determined as the arithmetic mean of four error indicators: serror_rate, srv_serror_rate (synchronization error rate for services), rerror_rate, and srv_rerror_rate (rejection error rate for services). These indicators can collectively reflect the level of "friction" or confrontation between attacking actions and the system. The calculation is as follows:

$$\beta = \frac{1}{4}(\text{serror_rate} + \text{srv_serror_rate} + \text{rerror_rate} + \text{srv_rerror_rate})$$

The coefficient γ represents the system's protection level or vigilance growth due to interaction with attacks. This parameter reflects the system's ability to adapt and strengthen defensive measures in response to detected malicious activity. A high value γ indicates rapid mobilization of defensive resources. It was calculated as the arithmetic mean of detected anomalies in encrypted services (R_{enc}) such as http_443 and ssh, and the frequency of anomalies in open, unencrypted services (R_{plain}). This allows for assessing the system's response to threats in different visibility segments. The calculation is as follows:

$$\gamma = 0.5 \times \left(R_{\text{enc}} + R_{\text{plain}}\right)$$

Coefficient ϕ models the intrinsic decay or "cost" of maintaining defense. It can reflect the natural decline in the effectiveness of defensive mechanisms over time due to obsolescence, resource depletion, or operational costs of maintaining the security system, even without direct interaction with attacks. This coefficient was calculated as the complement of β:

$$\phi = 1 - \beta$$

We justify this operationalization of coefficients by linking the abstract parameters of the Lotka-Volterra model with specific, measurable characteristics of network traffic. These characteristics can indirectly reflect attack intensity, defense response, and the overall dynamics of the confrontation.

These four coefficients of the Lotka-Volterra model $(\alpha, \beta, \gamma, \phi)$, pre-calculated based on the analysis of characteristics from the entire dataset, serve as target values for training the regression part of the hybrid neural network. Additionally, to characterize the overall threat distribution, the proportions of each attack class were analyzed.

The proposed hybrid neural network is designed with a multi-task learning architecture, a common approach in deep learning [4; 8]. It comprises a shared feature extraction part and two separate output branches for classification and regression.

The input layer accepts feature vectors after preprocessing. A Gaussian noise layer is added to enhance the model's robustness to input data. The shared hidden layers include several fully connected layers with Relu activation functions, batch normalization for stabilizing training, and Dropout layers to prevent overfitting. In the specific implementation, layers with 256 and 128 neurons are used.

The classification branch is responsible for intrusion detection. It consists of several fully connected layers, including a layer with 64 neurons with ReLU and Dropout, an output layer with one neuron and a sigmoid activation function for binary classification into normal and anomalous traffic. The loss function for this task is Binary Cross-Entropy L_{BCE}, which for N samples is calculated as:

$$L_{BCE} = -\frac{1}{N}\sum_{i=1}^{N}[y_i \log(\hat{y_i}) + (1 - y_i)\log(1 - \hat{y_i})]$$

where y_i is the actual class label for the i-th sample, and $\hat{y_i}$ is the predicted probability.

The regression branch predicts the four coefficients of the Lotka-Volterra model. It consists of fully connected layers, including layers with 64 and 32 neurons with ReLU, and an output layer with four neurons and a linear activation function. The loss function for this task is Mean Squared Error (L_{MSE}), which for N samples and $M = 4$ coefficients is calculated as:

$$L_{MSE} = \frac{1}{N}\sum_{i=1}^{N}\frac{1}{M}\sum_{j=1}^{M}\left(c_{ij} - \widehat{c_{ij}}\right)^2$$

where c_{ij} is the actual value of the j-th coefficient for the i-th sample, and $\widehat{c_{ij}}$ is the predicted value.

The target values for training this branch are the pre-calculated coefficients of the Lotka-Volterra model. It is important to note that after the neural network's prediction, the coefficients' obtained values $\alpha, \beta, \gamma, \phi$ are clipped to the range [0,1] before their further use in simulation. This ensures their physical plausibility in the model context, where they often represent fractions or probabilities, and prevents the occurrence of invalid values, such as negative growth rates.

The total loss function of the model L_{Total} is a weighted sum of the loss functions for classification and regression:

$$L_{Total} = w_{BCE}L_{BCE} + w_{MSE}L_{MSE}$$

The weight for the classification loss w_{BCE} is set to 1.0, and for the regression loss w_{MSE}, it is set to 0.2. The model is compiled with the AdamW optimiser [6; 9], which has proven effective for training deep neural networks.

Before being fed into the neural network, the data from NSL-KDD undergoes several preprocessing steps. For numerical features representing byte counts, such as src_bytes and dst_bytes, a logarithmic transformation $f(x) = log(1 + x)$ is applied to reduce the impact of extreme values. A byte ratio is also calculated as a new feature. Categorical features, such as protocol_type and flag, are transformed using one-hot encoding. Other string features are encoded using label encoding. Finally, all numerical features are scaled using standardization to have a zero mean and unit variance. The encoders and the scaler object are trained on the training dataset and saved for subsequent use on the test data and during inference.

III. Evaluation of Hybrid Model Results

The model is trained on the NSL-KDD training dataset, with 20% of this data used as a validation set to monitor the training process and prevent overfitting. Standard callback mechanisms were used to manage the training process. Specifically, a ModelCheckpoint mechanism was employed, which saved the model state that demonstrated the highest Area Under the Receiver Operating Characteristic Curve (AUC) value on the validation set. AUC is a standard metric for evaluating the quality of binary classifiers, indicating their ability to distinguish between positive and negative classes at various classification thresholds. Values close to 1.0 indicate high discriminatory power of the model. Adaptive learning rate reduction (ReduceLROnPlateau) was also implemented, which adjusted the learning rate if the loss function on the validation set did not improve for several epochs. An early stopping technique (EarlyStopping) was applied to prevent overfitting, which terminated the training process if the AUC value on the validation set ceased to increase, restoring the model weights to the state corresponding to the best AUC score.

After training, the model is evaluated on the test set. The primary metrics for the classification task are Accuracy and AUC, and for the regression task, Mean Squared Error (MSE) for the predicted coefficients of the Lotka-Volterra model.

Following the training and configuration of the hybrid neural model according to the described methodology, its evaluation was performed on the NSL-KDD test dataset, and an analysis of the predicted parameters of the Lotka-Volterra dynamics was conducted.

The model training process was monitored using metrics on the training and validation sets. Fig. 1 presents the learning curves, which demonstrate classification accuracy, AUC, and MSE dynamics for the Lotka-Volterra model coefficients.

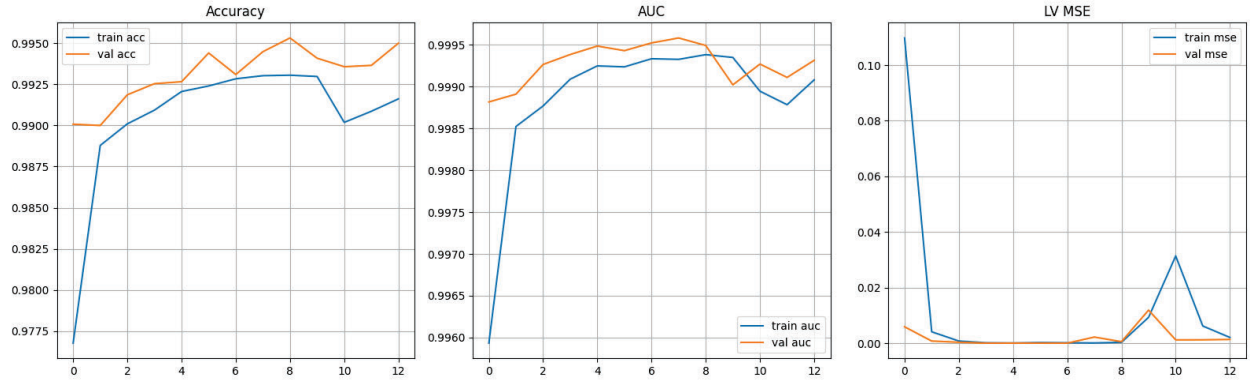


Fig. 1. Model learning curves: (Accuracy) classification accuracy, (AUC) classification AUC, (LV MSE) MSE for Lotka-Volterra model coefficients

Figure 1 visualizes the hybrid model's training process, displaying the dynamics of classification accuracy (Accuracy), AUC, and MSE for predicting the Lotka-Volterra coefficients. Analysis of these curves indicates stable and practical training. The expected increase in accuracy and AUC is observed on both the training and validation datasets, indicating an improvement in the model's ability to classify traffic. Simultaneously, a rapid decrease in MSE for the regression task occurs, demonstrating the model's successful learning of dependencies for predicting the dynamic coefficients. It is essential to note the proximity and synchronous behaviour of the curves for the training and validation sets, especially for AUC and MSE, which suggests good generalization ability of the model and the absence of significant overfitting. This is, in particular, a result of applying the early stopping mechanism, which terminated training upon reaching optimal performance on the validation data, preventing model degradation.

The final performance metrics of the model on the held-out test dataset are presented in Table 1.

TABLE I. MODEL PERFORMANCE METRICS ON THE NSL-KDD TEST SET

Metric	Value
Classification Accuracy	0.8006
Classification AUC	0.9016
MSE for Lotka-Volterra model coefficients	0.0027

Table 1 summarizes the key performance indicators of the developed hybrid model on the held-out NSL-KDD test dataset. The achieved classification accuracy (Accuracy) of 0.8006 indicates that the model correctly identifies most traffic samples. The AUC value is even more indicative, at 0.9016. Such a high AUC value underscores the classifier's excellent discriminatory power and ability to distinguish regular traffic from anomalous traffic at various activation thresholds. For the regression part of the model, which was responsible for predicting the Lotka-Volterra coefficients, a low MSE of 0.0027 was obtained. This indicates the high precision with which the neural network has learned to expect these dynamic parameters. Collectively, these results confirm the high performance of both components of the hybrid model: in anomaly detection and the quantitative assessment of parameters for subsequent dynamic modelling.

After model inference on the test dataset, the predicted Lotka-Volterra model coefficients were analyzed for normal and anomalous classes. The model generated 13592 normal and 8952 anomalous predictions on this set. Descriptive statistics of the predicted coefficients $\alpha, \beta, \gamma, \phi$ of the Lotka-Volterra model for normal and anomalous traffic are presented in Table 2.

TABLE II. DESCRIPTIVE STATISTICS OF THE PREDICTED LOTKA-VOLTERRA MODEL COEFFICIENTS $\alpha, \beta, \gamma, \phi$ FOR NORMAL AND ANOMALOUS CLASSES

Coefficient	Class	Mean	Std. Dev.	Minimum	Maximum
α	Normal	0.0603	0.0325	0.0200	1.0000
	Anomaly	0.0764	0.0256	0.0276	1.0000
β	Normal	0.2243	0.0203	0.1730	1.0000
	Anomaly	0.2514	0.0160	0.1941	0.7318
γ	Normal	0.7386	0.0224	0.2732	0.7533
	Anomaly	0.7348	0.0091	0.5205	0.7532
ϕ	Normal	0.8434	0.0225	0.8127	1.0000
	Anomaly	0.8563	0.0159	0.8196	1.0000

Table 2 shows that the mean values of the four predicted Lotka-Volterra model coefficients systematically differ for normal and anomalous traffic, confirming the informativeness of these parameters for characterizing the system's state. As expected, a higher mean coefficient value α is observed for anomalous traffic (0.0764 versus 0.0603), reflecting a greater inherent ability of attacks to propagate. The mean value of β s is also higher for anomalies (0.2514 versus 0.2243); this may indicate that under attack conditions, the defense system interacts more actively with the threat, leading to greater 'neutralization' per unit of attack, or that the characteristics of the attack itself provoke a more intense response. Somewhat unexpectedly, the mean value of the coefficient γ is slightly lower for anomalous traffic (0.7348 versus 0.7386), which might indicate certain limitations in the system's ability to scale up its defense during some types of anomalies, or that other factors not accounted for in this interpretation γ become dominant. A higher mean value of the coefficient ϕ for the anomalous class (0.8563 versus 0.8434) logically indicates a greater “cost” or faster “depletion” of defensive resources under increased load conditions. These quantitative differences demonstrate the model's ability to distinguish dynamic patterns and provide a basis for a deeper understanding of interaction mechanisms in cyberspace.

A comprehensive approach to visualization was used for the analysis and clear visual illustration of differences in the distributions of the predicted Lotka-Volterra model coefficients for normal and anomalous traffic. Specifically, we constructed histograms (Fig. 2), which allow for assessing the overall shape of each coefficient's distribution and the frequency of occurrence of specific values. For a more detailed comparison of statistical characteristics, such as medians, quartiles, and the presence of outliers, box plots were used (Fig. 3). Additionally, to combine the advantages of both

approaches and gain insight into the distribution density, violin plots were constructed (Fig. 4). This set of visualizations provides a deep understanding of how the predicted parameters of the dynamical system vary depending on the classified type of network activity.

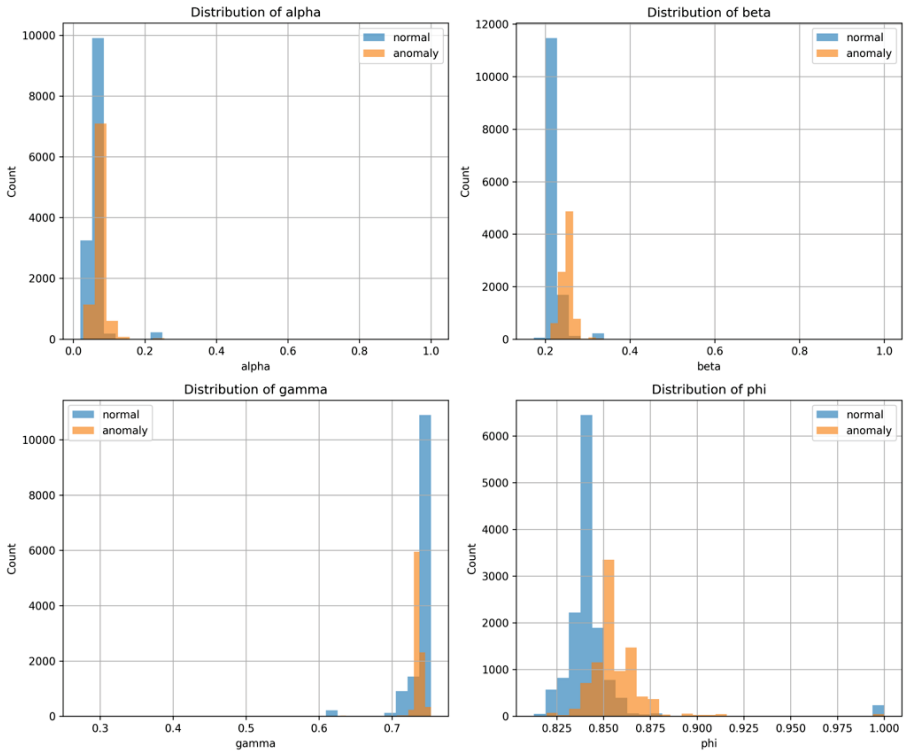


Fig. 2. Histograms of the predicted Lotka-Volterra model coefficients α,β,γ,ϕ distribution for regular (blue) and anomalous (orange) traffic

The histograms in Fig. 2 demonstrate the characteristic differences in the distributions of the predicted Lotka-Volterra model coefficients for regular (blue) and anomalous (orange) traffic, allowing for a deeper understanding of their behaviour.

For the coefficient α (intrinsic growth rate of attacks), the distribution of values for the anomalous class is noticeably shifted to the right. It has a longer right tail than the regular class. This visually confirms the tendency towards higher α values during anomalous activity, indicating its greater potential for independent development.

Regarding the coefficient β (defense effectiveness), the distribution for anomalous traffic also shows a rightward shift and greater variance in values. The peak of the distribution for anomalies is at higher β values, and the distribution itself is wider, which may indicate a more intense interaction between the attack and the defense system.

In the case of the coefficient γ (growth of protection level), although the main concentrations of values for both classes are close, the distribution for the anomalous class is distinguished by a more pronounced left tail. This indicates the presence of a subgroup of anomalous cases with predicted lower γ values, which can be interpreted as situations where the system less effectively strengthens its defense in response to a threat.

Finally, for the coefficient ϕ cost of maintaining defense, the histogram for the anomalous class also shows a rightward shift of the peak and a wider distribution, suggesting a tendency towards higher ϕ values during anomalies. This may reflect greater "costs" of maintaining the security system or its accelerated "depletion" under active confrontation conditions.

Thus, the histograms confirm the quantitative differences and reveal features of the distribution shapes, providing valuable insights for further analysis.

The box plots presented in Fig. 3 provide a detailed view of the statistical differences in the distributions of the predicted Lotka-Volterra model coefficients between normal and anomalous traffic, focusing on key indicators such as medians, interquartile ranges (IQR), and the presence of outliers.

For the coefficient α, the median value for the anomalous class (orange) is higher than for the regular class (blue), indicating a systematically higher level of attack "aggressiveness". A larger IQR is also observed for anomalies, suggesting greater parameter variability under attack conditions.

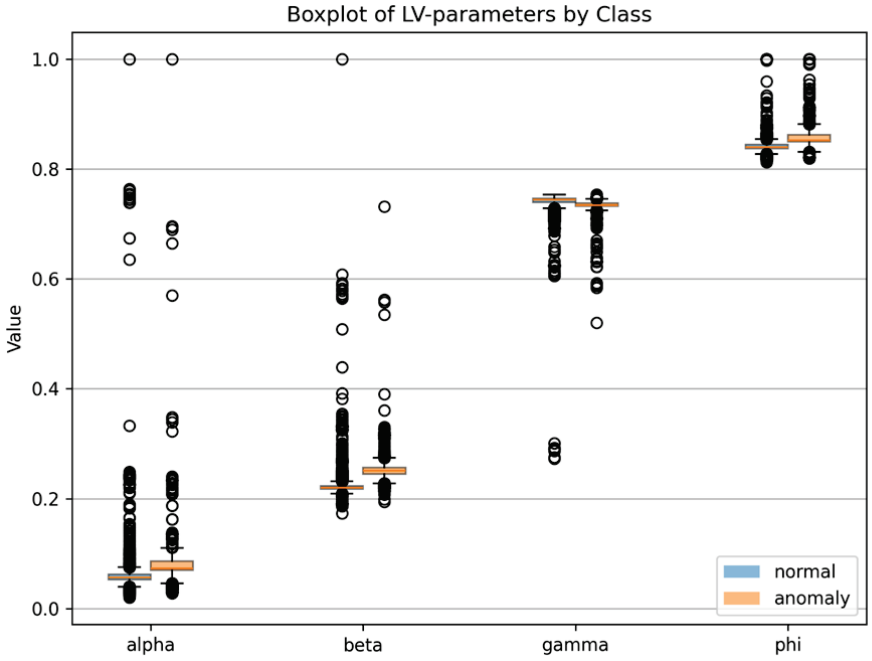


Fig. 3. Box plots of the predicted Lotka-Volterra model coefficients α,β,γ,ϕ for normal and anomalous traffic

A similar trend towards higher values for the anomalous class is observed for the coefficient β, its median and upper quartile for anomalies significantly exceed the corresponding indicators for regular traffic, with the interquartile range also being larger. This underscores the more intense interaction between attack and defense during anomalous events.

In the case of the coefficient γ, the median values for both classes appear quite close. However, the box for the anomalous class shows a wider interquartile range, mainly due to lower values, consistent with a left tail in the histogram. This may indicate greater heterogeneity in the defense system's response to different types of anomalies.

For the coefficient ϕ, the median and the overall range of values (including IQR) for the anomalous class are also shifted upwards compared to the regular class, confirming the trend towards a higher "cost" of maintaining defense during attacks.

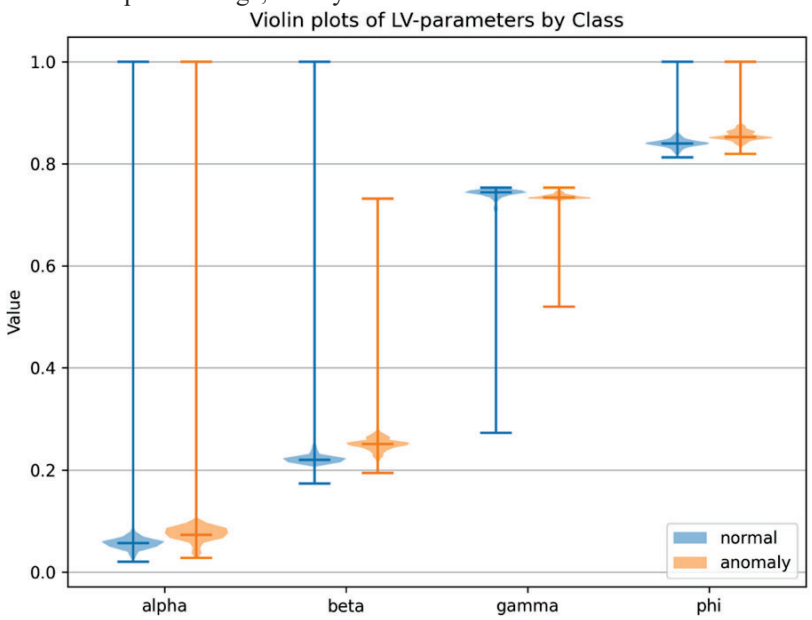


Fig. 4. Violin plots of the predicted Lotka-Volterra model coefficients α,β,γ,ϕ for normal and anomalous traffic

A significant number of outliers, especially for coefficients α and β, indicates the presence of individual network sessions with extreme characteristics, which the model can also identify. Overall, these diagrams convincingly demonstrate that the model classifies traffic and predicts parameters of the dynamical system that are statistically significantly different for regular and anomalous activity, which is key for further risk level assessment.

The violin plots in Fig. 4 effectively combine the informativeness of box plots (displaying the median and interquartile range) with the visualization of distribution density, similar to histograms, allowing for an even more

complete and nuanced understanding of the differences in the predicted coefficients for normal and anomalous traffic.

For coefficient α, the violin plot highlights that although both distributions have peaks at low values, the anomalous class (orange) shows a higher probability density (wider part of the "violin") in the range of higher values, as well as a longer upper tail, indicating a more frequent occurrence of extremely high α values during attacks.

In the case of the coefficient β, the violin plot for the anomalous class is shifted to the right and has a greater width in the value range of approximately 0.23 to 0.28. The distribution is significantly narrower and concentrated around 0.22. This demonstrates higher mean values of β anomalies and a greater diversity of these values.

For the coefficient γ, the violin plots show that the central mass of values for both classes is concentrated in the region of approximately 0.73-0.75. However, the "violin" widens in its lower part for the anomalous class, corresponding to many cases with lower γ values. This confirms observations from histograms and box plots about greater variability and the presence of a left tail in this coefficient's distribution for anomalies.

As seen from the violin plot, the distribution of the coefficient ϕ or the anomalous class is also shifted to the right, with a higher density of values in the 0.85-0.87 range, compared to the regular class, where the density peak is lower.

Thus, the violin plots confirm and complement the conclusions drawn from histograms and box plots, convincingly demonstrating clear and statistically significant differences in the distribution shapes of all four Lotka-Volterra model coefficients depending on the class of network activity. This creates a strong foundation for using these coefficients to model cyber risk dynamics further.

A key validation stage for the proposed hybrid approach is analyzing how the predicted coefficients $\alpha, \beta, \gamma, \phi$ affect the dynamics of the "attack-defense" system, modelled by the Lotka-Volterra equations. To this end, a numerical simulation of the system's development over time was conducted using sets of coefficients individually predicted for each sample from the test set (after their processing to ensure physical plausibility, as described in the methodology). This approach allows not only to state statistical differences in the coefficients but also to visualize and qualitatively assess which interaction scenarios- stable, oscillatory, or escalating-result from particular sets of parameters. The results of these numerous simulations, including time series for "attack" levels $A(t)$ and "defense" levels $Z(t)$, as well as corresponding phase portraits (if they were analysed), were thoroughly systematized. To illustrate the characteristic differences, Figs. 5 and 6 present representative examples of simulation time trajectories for typical cases classified by the model as normal and anomalous traffic. These examples were selected based on prior cluster analysis or analysis of summary characteristics (e.g., maximum amplitude of $A(t)$, time to stabilization, etc.) from the entire set of simulations to best reflect the generalized patterns of system behaviour.

For a typical case of regular traffic, as illustrated in Fig. 5, the Lotka-Volterra dynamics simulation demonstrates behaviour characteristic of a relatively stable system. Initially, a decrease in potential "attacks" $A(t)$ (blue line) is observed, likely due to the initial "defense" level $Z(t)$ (orange line). Although $A(t)$ subsequently begins to rise, reaching a local maximum, the defense system $Z(t)$, after a period of low activity (which may reflect resource saving in the absence of an explicit threat), activates with a delay but effectively enough to reverse the growth trend of $A(t)$ and lead to its decrease towards the end of the simulation period. Importantly, the "attack" level does not get out of control, and the oscillations are either damped or the system returns to an equilibrium state with a low threat level. Such dynamics indicate that the coefficients $\alpha, \beta, \gamma, \phi$ predicted for regular traffic describe a system capable of adequately responding to potential threats and maintaining a secure state.

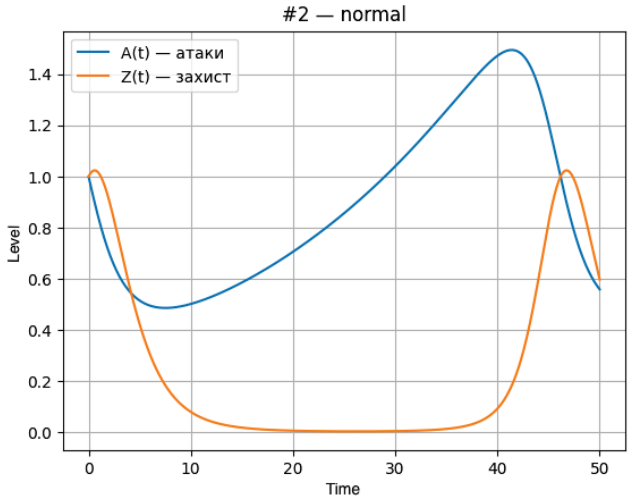


Fig. 5. Example of Lotka-Volterra dynamics simulation for a case classified as "normal"

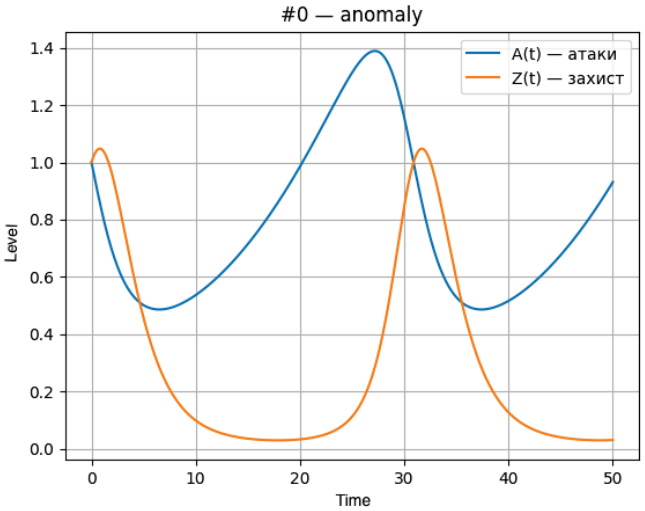


Fig. 6. Example of Lotka-Volterra dynamics simulation for a case classified as "anomaly"

In contrast to regular traffic, the dynamics simulation for a typical case of anomalous traffic (Fig. 6) reveals significantly less stable behaviour of the "attack-defense" system. A sharp increase in the "attacks" level $A(t)$ (blue line) is observed, reaching a considerably higher peak value than regular traffic. Although the "defense" level $Z(t)$ (orange line) also responds to this increase, its response may be insufficient or delayed to suppress the attack effectively. As a result, after a brief decrease, the $A(t)$ level may begin to rise again, or the system enters a mode of pronounced oscillations, where periods of high attack activity alternate with periods of defense

activation. Such oscillatory or uncontrollably increasing dynamics of the $A(t)$ level is characteristic of active cyber confrontation. It indicates a significantly higher level of cyber risk, as the system cannot quickly return to a secure, stable state. This underscores that the coefficients predicted for anomalous traffic describe more dangerous event scenarios.

Analysis of the aggregate simulations conducted for all samples in the test set confirmed these observations on a more general level. It was found that the predicted Lotka-Volterra model coefficients for cases classified as anomalous much more frequently lead to dynamic patterns characterized by either uncontrolled growth of the "attacks" level $A(t)$ (simulating a successful, escalating attack) or sustained high-amplitude oscillations (indicating prolonged active confrontation and system instability). In such scenarios, the "defense" level $Z(t)$ often proves insufficient or fails to respond adequately in time to the attack's development speed.

Coefficients obtained for regular traffic predominantly generate trajectories where the system either quickly returns to an equilibrium state with a low $A(t)$ level or exhibits only minor, controlled oscillations that are effectively suppressed by defensive mechanisms. These more stable and secure trajectories indicate the system's ability to maintain security.

Thus, the simulation results not only confirm the binary classification (normal/anomaly) but also provide qualitatively new information about the nature of the potential threat or system stability. This opens up the possibility for a more granular assessment of cyber risks: for example, Distinguishing anomalies that lead to rapid escalation from those that result in prolonged but less intense confrontation. Consequently, the proposed hybrid approach, which combines classification with dynamic modelling, has significant potential for creating more informative cyber risk assessment systems that go beyond merely stating the fact of an intrusion and allow for predicting possible consequences.

Although the obtained results demonstrate the proposed hybrid approach's effectiveness, it is essential to discuss some of its aspects and potential limitations for a complete picture. The latter, at the same time, point to possible avenues for future research.

Firstly, although the NSL-KDD dataset is a standard for comparison [13] and allowed us to demonstrate the method's viability, it is worth noting that it was created quite some time ago and may not fully cover the spectrum of modern threats [10]. However, the developed architecture of the hybrid model and the very principle of combining classification with the prediction of dynamic parameters are flexible. This creates a prospect for adapting and applying this approach to newer, more representative datasets, which can confirm the results and potentially increase the model's accuracy and relevance in modern conditions. Training or fine-tuning the model on such data is a logical next step.

Secondly, using the Lotka-Volterra model is undoubtedly a simplification of complex processes in cyberspace [12; 15]. Nevertheless, this model is a useful abstraction that allows for understanding and visualizing key dynamic aspects of the "attack-defense" interaction, which are often ignored in static risk models. The value of the approach lies not so much in an absolutely precise imitation of reality as in obtaining a qualitatively new level of understanding of the potential development of a situation.

Thirdly, one of several possible operationalizations is the method for determining the Lotka-Volterra model coefficients $(\alpha, \beta, \gamma, \phi)$ based on specific NSL-KDD features [12; 15]. When transferring the approach to other datasets where the feature set may differ, adaptation of these formulas or development of new methods for their calculation corresponding to the available data will be required. This opens a direction for further research on universalization or developing alternative ways to parameterize dynamic models based on network data.

Furthermore, the regression part of the neural network, like many deep learning models [4], has "black box" characteristics. Although the model demonstrates good accuracy in predicting coefficients, a detailed study of the influence of individual input features on the predicted dynamic parameters requires the application of interpretability methods. This is also an important direction for future work, allowing for a better understanding of the model's internal logic.

Therefore, acknowledging the existing limitations, we consider them not so much as shortcomings that negate the obtained results, but rather as stimuli for the further development and improvement of the proposed hybrid approach to dynamic cyber risk assessment.

IV. Conclusions

This work proposed and investigated a hybrid neural model for cyber risk assessment that combines network traffic classification with the prediction of parameters for the Lotka-Volterra dynamical model. Experiments on the NSL-KDD dataset showed that the model can effectively detect anomalous activity, achieving an accuracy of 0.8006 and an AUC of 0.9016, and predict the Lotka-Volterra model coefficients $(\alpha, \beta, \gamma, \phi)$ with a Mean Squared Error of 0.0027. Importantly, the predicted coefficients significantly differ for normal and anomalous network sessions.

Analysis of the predicted Lotka-Volterra model coefficients and subsequent simulation of the "attack-defense" dynamics demonstrated that the proposed approach allows for identifying a threat and assessing its potential development and interaction with the defense system. Different patterns of Lotka-Volterra model dynamics, corresponding to various traffic classes and sets of coefficients, can serve as indicators of the cyber risk level, providing a more detailed picture of the security state than traditional intrusion detection systems.

The practical significance of the work lies in the possibility of creating more informative decision support systems in cybersecurity. Such systems would allow us to respond to already detected incidents and proactively assess potential threats by analyzing the dynamic characteristics of interactions in cyberspace, considering the modern threat landscape.

Further development of this approach may include validating the model on more modern and diverse datasets reflecting current cyber threats; investigating the impact of different neural network architectures and methods for interpreting Lotka-Volterra model coefficients; integrating the model with real security monitoring systems to assess its effectiveness in an operational mode; extending the Lotka-Volterra model to account for more factors, such as different types of attacks, defensive strategies, and adaptation mechanisms; conducting a sensitivity analysis of the Lotka-

Volterra model to the predicted coefficients; and a deep investigation into the interpretability of the neural network's regression part to identify key network features influencing the dynamic parameters.

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Effect of nitrate-containing compounds on the morphology of black chromium coatings

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***Abstract*— This study investigates the effect of nitrate-containing additives on the morphology and elemental composition of black chromium coatings. The coatings were deposited on AISI 304 stainless steel substrates, which are widely used as a structural material in nuclear power applications, using the galvanostatic electrolysis method. The synthesized coatings were characterized by scanning electron microscopy (SEM) and energy-dispersive X-ray spectroscopy (EDX). The results showed that increasing the concentration of aluminum nitrate in the electrolyte led to a noticeable increase in the size of surface cracks in the coatings. Specifically, when aluminum nitrate was introduced while sodium nitrate was excluded from the electrolyte, the crack size increased by a factor of 12.5. The addition of aluminum nitrate also reduced the average crystallite size within the coating by a factor of 3.5 to 7. Despite this reduction, the presence of large crystallites (600–800 nm) was observed across all examined samples. A specimen formed in an electrolyte containing 0.1 mol/L aluminum nitrate without sodium nitrate exhibited larger crystallites compared to one formed in an electrolyte with the same aluminum nitrate concentration but with added sodium nitrate. The crystallite sizes were 250 nm and 100 nm, respectively. Elemental analysis confirmed that all coating samples were composed of chromium and oxygen, with no detectable aluminum. Based on EDX point analysis, the likely chemical formulas of chromium oxides in various oxidation states were calculated and correlated with specific points on the coating surface. These findings indicate that the coatings predominantly consist of heat-resistant and hard chromium (III) oxide, along with contributions from chromium (IV) oxide and mixed-valence chromium (II,III) oxide.**

Keywords—black chromium, protective coatings, chromium oxides, scanning electron microscopy, energy dispersive X-ray spectroscopy.

I. Introduction

Nuclear energy plays a significant role in ensuring stable electricity generation both in Ukraine and globally. Its continued development remains a strategic priority. At the same time, the long-term operation of nuclear power plants has become a current reality, and this trend is expected to persist. This evolution of the industry poses a number of challenges for researchers and engineers, one of the key ones being the extension of the service life of structural materials. Consequently, the development of new materials is a critical issue for enhancing the safety and resilience of nuclear energy systems.

One of the key factors significantly affecting the service life of structural components is corrosion. It is well known that equipment failures caused by corrosion lead to increased radiation exposure for personnel and the environment, as well as substantial economic losses. The materials used in nuclear power systems are exposed to various types of corrosion damage, including general corrosion, environmentally assisted cracking, stress corrosion cracking, irradiation-assisted stress corrosion cracking, flow-accelerated corrosion, and microbiologically influenced corrosion [1–4]. Another specific, yet equally hazardous form of corrosion is pitting corrosion of stainless steel, which occurs in aqueous environments containing chloride ions. The presence of chloride ions in solution can lead to the breakdown of the passive film that serves as the protective barrier for stainless steel. Once this film is compromised, localized pits—known as pitting sites—form on the surface. Within these pits, an acidic environment with a low pH develops, promoting aggressive localized attack. The progression of these pits into the bulk of the material can lead to significant degradation of its mechanical properties.

One of the materials widely used in the core and primary circuit of existing nuclear power plants is austenitic stainless steel, such as AISI 304 [5–8]. Its advantages include resistance to general corrosion in aqueous environments even at elevated temperatures, good ductility, and relatively low cost compared to other types of stainless steels. However, as no material is entirely immune to degradation, AISI 304 is still susceptible to corrosion damage, particularly pitting, crevice, and intergranular corrosion. Although various approaches exist for protecting materials against corrosion, it is essential to consider both economic feasibility and overall effectiveness. In this context, the use of protective coatings represents a practical and cost-efficient compromise.

Since the structural components of nuclear power plants operate under prolonged exposure to elevated temperatures and radiation [9–11], the protective coatings used must also meet the stringent requirements of heat resistance, mechanical strength, and radiation tolerance. In current research on Accident-Tolerant Fuel (ATF) concepts, particular attention is given to chromium-based coatings as a short-term solution to enhance fuel performance under accident conditions [12–15]. Of notable interest is the passive chromium (III) oxide film, which retains its protective properties even at high operating temperatures. Additionally, chromium (III) oxide is well known for its exceptional hardness. These properties make it highly suitable for meeting the performance demands placed on materials used in next-generation nuclear power plants [16–22].

Currently, scientific literature frequently explores various methods for the formation of metal oxide coatings, including sol-gel techniques, vapor-phase deposition methods, dip-coating, and electrochemical deposition—each with its own

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specific advantages and areas of application. This study emphasizes the electrochemical method as the most technologically versatile and economically justified approach to the fabrication of chromium oxide coatings.

In the formation of metal oxide coatings via electrochemical methods, two primary modification strategies can be employed: altering the electrolysis regime and modifying the electrolyte composition. The influence of electrolysis parameters has been demonstrated in [11, 12], particularly in discussions on the impact of pulsed deposition modes on the morphology of black chromium coatings. At the same time, the development of novel electrolyte compositions remains a promising area of research.

Significant efforts are being made to develop composite coatings based on black chromium [5, 6]. Studies have shown that the quality of black chromium coatings can be influenced by the use of catalytic additives. Several investigations [4, 6] have employed compounds containing nitrate ions as such additives. In the work of M. Aguilar-Sánchez [14], it was demonstrated that the addition of KNO_3 to the electrolyte promotes the formation of a "black" chromium deposit, whereas a similar electrolyte without the nitrate-containing additive yields a "white" chromium deposit. It was found that the black deposit consists predominantly of chromium (III) oxide, while the white deposit is composed of chromium (III) hydroxide. The researchers concluded that the reduction of the nitrate ion during electrolysis plays a critical role in determining the nature of the compounds that form the coating layer.

However, the effect of the combined introduction of aluminum nitrate and sodium nitrate on the morphology and elemental composition of black chromium remains insufficiently studied. Of particular interest is the investigation of how these components influence the crystalline structure, grain size, and crack formation patterns, as these factors have a direct impact on the performance characteristics of the coatings.

In our previous studies [5], it was established that black chromium coatings synthesized without the addition of modifying components contain chromium oxides in various oxidation states, specifically CrO_2, Cr_2O_3, and Cr_3O_4. These phases were identified using X-ray diffraction (XRD) analysis and were found to exhibit different degrees of thermal and electrochemical stability. Of particular importance is the presence of Cr_2O_3, as it is the most stable passivating phase, responsible for imparting protective properties to the coating.

This research aims to investigate the effect of aluminum nitrate and sodium nitrate additives on the morphology and elemental composition of the resulting surface layer. Establishing a correlation between the electrolyte composition and the characteristics of the coating's surface layer will make it possible to determine the most balanced solution composition for forming continuous protective layers on stainless steel.

II. Materials and Methods

A. Coatings formation

AISI 304 stainless steel was used as cathode. Coatings were formed by electrodeposition in a galvanostatic mode. Tested black chromium coatings were applied to both sides of each plate on a total area of 4 cm^2. The process was carried out in a cell with a working capacity of 500 ml. The anode was a platinum wire. Compositions of electrolytes, mol/L: 2.5 CrO_3, 0.3 H_3BO_3, 0.01 $Ba(OH)_2$, 0…0.1 $Al(NO_3)_3$ with addition of 0.06 $NaNO_3$, and 0.1 $Al(NO_3)_3$ without the addition of $NaNO_3$. Electrolysis was carried out at a current density of 50 A/dm^2 for a duration of 40 minutes. The electrolyte temperature was maintained at room temperature.

B. Surface morphology and elemental characterization.

The analysis of surface morphology and elemental composition of the obtained samples was performed using a ZEISS EVO 40XVP scanning electron microscope equipped with an INCA Energy 350 energy-dispersive microanalysis system (Carl Zeiss, Germany; Oxford Instruments, UK). Surface images were acquired by detecting secondary electrons (SE) generated during scanning of the surface with an electron beam. Secondary electron excitation was achieved by irradiating the samples with an electron beam at an energy of 15–20 keV. The acquired micrographs were processed using SMARTSEM and ImageJ software environments.

III. Results and Discussion

A. Morphology analysis

The coatings obtained after electrolysis exhibited a matte black appearance, uniformly covering the entire surface of the sample. The visual appearance of the coatings remained unchanged regardless of the concentration of $Al(NO_3)_3$ and $NaNO_3$ in the electrolyte.

The micrographs in Fig. 1 and Fig. 2, captured at different magnifications, provide a detailed view of the surface morphology of the formed coatings. It is evident that the largest cracks are characteristic of the coating obtained from the black chromium electrolyte containing 0.1 mol/L $Al(NO_3)_3$ without the addition of $NaNO_3$. Cracks in the chromium oxide layer of this sample are clearly visible even at a magnification of 500× (Fig. 1, d), with the largest cracks exceeding 2.5 μm in width.

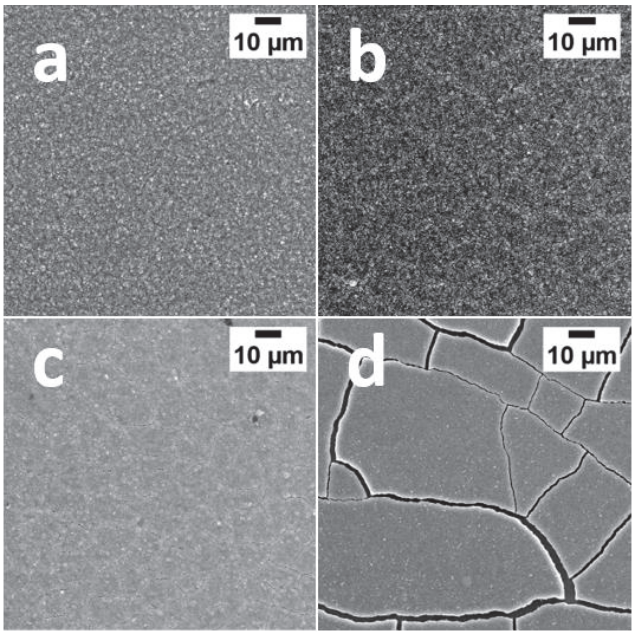


Fig. 1. SEM images of black chromium coatings, obtained from electrolytes with addition of 0 (a); 0.05 (b); 0.1 (c) and 0.1 mol/L without the addition of $NaNO_3$ (d)

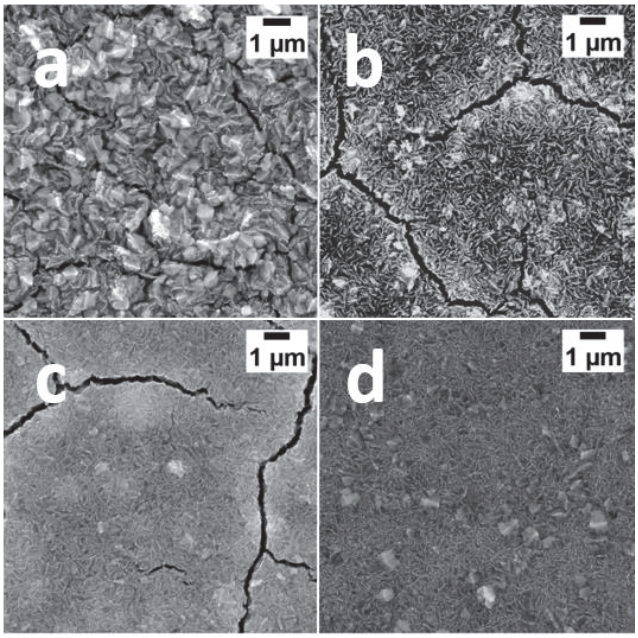


Fig. 2. SEM images of black chromium coatings, obtained from electrolytes with addition of 0 (a); 0.05 (b); 0.1 (c) and 0.1 mol/L without the addition of $NaNO_3$ (d)

It was also noted that the areas of the coating confined by cracks in this sample are internally homogeneous and structurally intact. In contrast, cracks in the coatings of samples formed from electrolytes containing $NaNO_3$ become visible only at a magnification of 5000× (Fig. 2 a, b, c), with the average crack width in each case measuring approximately 0.2 μm. These findings suggest a well-defined correlation between crystallite size and the concentration of nitrate-containing compounds, with increased concentrations leading to smaller crystallite sizes. The images in Fig. 2 show that the addition of $Al(NO_3)_3$ alters the configuration of the crystallites forming the coating. The surface acquires a finer-grained crystalline structure, which positively affects the mechanical, optical, and decorative properties of the coating. The coating obtained from electrolytes without the addition of $Al(NO_3)_3$ has an average crystallite size of approximately 700 nm. As the concentration of $Al(NO_3)_3$ increases, the crystallite size decreases. According to Fig. 2 (b), the crystallite size is about 200 nm when 0.05 mol/L $Al(NO_3)_3$ is added. In Fig. 2 (c), the crystallite size is reduced by half with an increase in $Al(NO_3)_3$ concentration to 0.1 mol/L.

The samples also contain inclusions of larger crystallites with an average size of approximately 600 nm. As shown in the micrograph in Fig. 2 (d), the sample obtained from the electrolyte without $NaNO_3$ exhibits a significant number of large crystallite inclusions, with an average size of 777 nm. In contrast, the average size of the smaller crystallites, which predominantly form the coating, is much lower, around 250 nm. Therefore, the presence or absence of $NaNO_3$ in the electrolyte plays a decisive role in determining crack size and influences the crystalline structure of the coating.

B. Analysis of elemental composition

The elemental composition of each coating sample was determined at four distinct points, the locations of which are shown in Fig. 3. At each point, Chromium and Oxygen were detected, with their respective contents presented in Table 1. Aluminum was not detected in the composition of any of the coatings.

Analyzing the data presented in Table 1, it was found that the introduction of $Al(NO_3)_3$ into the black chromium electrolyte at varying concentrations leads to a decrease in the proportion of chromium in the compounds forming the coating. Accordingly, the oxygen content increases. However, when $Al(NO_3)_3$ is introduced and $NaNO_3$ is excluded from the electrolyte composition, the chromium content in the resulting compounds is the highest among all the samples. Based on the results of the EDX analysis, the possible chromium oxide compounds were also calculated. According to these calculations, the surface of the black chromium coating formed without the addition of $NaNO_3$ contains chromium (III) and chromium (IV) oxides. Compound CrO_2 was identified at the beam impact points located on large, bright crystallites within the coating (Fig. 3d, point 2). In contrast, Cr_2O_3 was detected in areas where the coating consists of finer, darker crystallites. The coating obtained from the electrolyte without the addition of $Al(NO_3)_3$ (Fig. 3a) consists of crystallites similar in size to the large crystallites embedded in the previously discussed coating (Fig. 3d). Elemental analysis revealed that at points 2 and 4 (Fig. 3a), the coating consists of chromium (IV) oxide, which is consistent with the observations made for the coating obtained from the electrolyte without $NaNO_3$. At points 1 and 3 (Fig. 3a), the coating consists of Cr_2O_3 and Cr_3O_4. The identified oxides are consistent with the X-ray diffraction (XRD) results reported in publication [5].

The coating obtained from the electrolyte with the addition of 0.05 mol/L $Al(NO_3)_3$ consists of CrO_2, Cr_2O_3, and Cr_3O_4 compounds. The coating formed from the electrolyte containing 0.1 mol/L $Al(NO_3)_3$ consists of CrO_2 and Cr_2O_3 compounds. Coatings synthesized with both $Al(NO_3)_3$ and $NaNO_3$ additives predominantly consist of finer crystallites but also contain inclusions of larger crystallites, which is consistent with the observations described above. The most heat-resistant, chemically stable, and hardest of the identified chromium oxides is chromium (III) oxide — Cr_2O_3. It has a melting point of 2435 °C. Additionally, its hardness reaches up to 29.5 GPa, providing coatings based on it with excellent wear resistance. Among the detected compounds, the least thermally stable is chromium (IV) oxide — CrO_2. At temperatures around 277 °C, it becomes thermally unstable and transforms into the more stable Cr_2O_3 phase. Another oxide identified in the coating is Cr_3O_4, a mixed-valence chromium (II,III) oxide, with a melting point of 1705 °C [22].

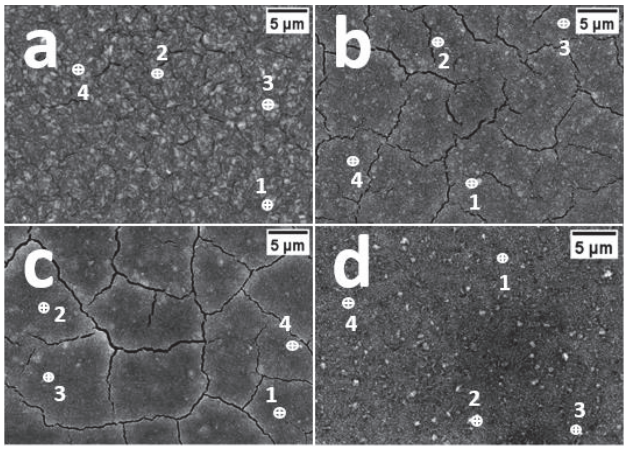


Fig. 3. EDX analysis points locations on black chromium coatings, obtained from electrolytes with addition of 0 (a); 0.05 (b); 0.1 (c) and 0.1 mol/L without the addition of $NaNO_3$ (d)

TABLE I. AVERAGE ELEMENTAL COMPOSITION OF COATINGS OBTAINED BY EDX POINT METHOD

Concentration of $Al(NO_3)_3$ in electrolyte, mol/L	Element	Concentration at a point, wt.%				
		1	2	3	4	Mean
0	Cr	68.49	61.92	59.03	63.51	62.50
	O	31.51	38.08	40.97	36.49	37.50
0.05	Cr	63.15	61.69	67.88	71.09	66.30
	O	36.85	38.31	32.12	28.91	33.70
0.1	Cr	68.18	67.88	61.86	62.39	64.99
	O	31.82	32.12	38.14	37.61	35.01
0.1 without $NaNO_3$	Cr	66.12	62.18	68.68	69.16	65.28
	O	33.88	37.82	31.32	30.84	34.72

Thus, the performed elemental analysis of the coatings demonstrates that varying the electrolyte composition allows for controlled formation of chromium oxides with different degrees of thermal stability. The most desirable outcome is a predominance of Cr_2O_3 in the coating, a thermodynamically stable and hard phase that ensures high wear and heat resistance. The obtained results confirm that the targeted formation of this phase can be achieved by optimizing the ratio of $Al(NO_3)_3$ and $NaNO_3$ additives in the electrolyte.

IV. CONCLUSIONS

Morphological analysis revealed that increasing the concentration of $Al(NO_3)_3$ in the electrolyte leads to a finer crystalline structure in the coating. However, this also results in larger crack formation. Elemental analysis confirmed the presence of chromium and oxygen. Based on the Cr:O ratio, the coatings were found to contain chromium oxides such as CrO_2, Cr_2O_3, and Cr_3O_4. Notably, no aluminum was detected in the coatings. Coatings produced from electrolytes without $NaNO_3$ exhibited larger cracks that covered broader surface areas compared to those formed with $NaNO_3$, while the crystal size remained nearly unchanged. All samples obtained from electrolytes containing $Al(NO_3)_3$ exhibited a fine-crystalline structure. The presumed oxide phases are characterized by high hardness and thermal resistance, confirming the effectiveness of the resulting chromium-based compositions as protective coatings.

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Comparative Analysis of Magnetic Compatibility of Two Variants of the Design of Electromagnets of the Orientation Control System of Micro-Satellites

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***Abstract*— The magnetic interference created by the electromagnets of the orientation control system inside a small spacecraft is considered. Depending on the two variants of the electromagnet design, an analysis of the magnetic field strength in the central part of five satellite models was comply. It was established that the level of magnetic interference created by a cylindrical electromagnet is 2–4 times lower than the magnetic field of a square current winding under the condition of the same magnetic moments and length. It is shown that the specific magnetic moment per unit of mass and electric power of a cylindrical electromagnet is approximately ten times higher than that created by a square current winding.**

Keywords—magnetic compatibility, spacecraft, magnetic field, electromagnet, current winding.

I. Introduction

One of the trends of the modern stage of development of space technology is the use of a wide variety of spacecraft (SC) for the study of outer space around the Earth. Based on the number of mission objectives, SCs differ in size, mass and time of operation in orbit depending on the parameters of the payload equipment. For the simplest mission objectives, the payload and the SC itself can be made with minimized overall dimensions and mass. Reducing the cost of delivering a satellite to low orbit to a level of less than \$ 10^3 per 1 kg has created a real boom in the development of very small spacecraft. Today, small spacecraft are classified by approximate overall dimensions and mass in the following ranges. Spacecraft with overall dimensions from 0.2 m to 0.5 m and a mass from 10 kg to 100 kg are considered micro-satellites. Spacecraft with overall dimensions from 0.1 m to 0.2 m and a mass of 1–10 kg belong to the class of nano-satellites. Even smaller spacecraft in size and mass are classified as pico-satellites [1–5].

Regardless of the size of the satellite, the reliability of its operation depends on many factors, one of which is the stable standard orientation of the spacecraft in orbit. A common property of small spacecraft is the control of the angular position due to the mechanical interaction of the magnetic moment of the orientation system and the Earth's magnetic field. To ensure the reliability of stabilization of the angular position of the satellite in orbit, the magnitude of the magnetic moment created by such electromagnets can be estimated at the rate of 0.2 A·m^2 per 1 kg of the mass of the SC. Therefore, such electromagnets usually have a design consisting of a cylindrical current winding that magnetizes a core made of soft magnetic material [6]. However, minimization of the size and mass of small spacecraft has prompted the use of a simplified design for electromagnets in the orientation control systems of micro- and nano-satellites in the form of a rectangular flat current winding without a ferromagnetic core. The magnetic moment created by such a simplified design electromagnet is equal to the multiplication of the ampere-turns by the average area of the turn of the flat winding. Simplification of the design of the electromagnet has a positive effect on the cost of its development and manufacture. However, the location of the payload equipment actually inside the current winding exacerbates the problem of magnetic compatibility of the spacecraft [7].

Therefore, it seems appropriate to analyze and compare the magnitude of magnetic interference created by electromagnets of two different types inside a small spacecraft, depending on its overall dimensions.

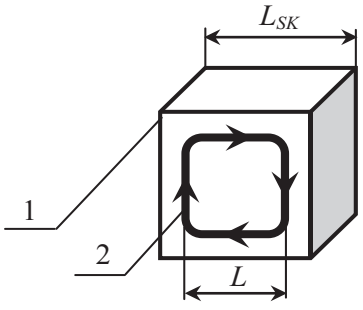
II. The Aim and Objectives of the Study

The aim of the work is to compare the magnitude of magnetic interference created by different types of electromagnets of the spacecraft's orientation control system to ensure magnetic compatibility inside the satellite.

The dependences of the magnitude of the magnetic field of interference created by different types of electromagnets for five satellite models, the overall dimensions of which increase linearly, are subject to study.

III. Output Conditions for Comparative Analysis of the Magnetic Field of Different Types of Electromagnets

The work compares the value of the magnetic field strength modulus created by a square current winding and a cylindrical electromagnet on an axis that crosses the center of the SC. For ease of comparison, the magnetic field is calculated on an axis that is perpendicular to the plane of the square current winding and passes through its center. In this case, a square winding with a current I and the number of turns N_w is considered thin enough so that its plane S_w is calculated as the square of the side L, and it itself is located in the boundary plane of the spacecraft (Fig. 1).



- Fig. 1. –The location of the square current winding on the boundary of the spacecraft: 1 – satellite model, 2 – square current winding.

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Such assumptions allow us to use the well-known formula [8] for a square coil to calculate the magnetic field strength modulus H_w on the central axis:

$$H_w = \frac{IN_wL^2}{2\pi}\left[x^2 + \left(\frac{L}{2}\right)^2\right]^{-\frac{3}{2}}, \quad (1)$$

where x – coordinate of the field observation point lying on the central axis perpendicular to the winding plane.

In order to ensure a correct comparison of the magnetic field of two types of electromagnets: a square current winding and a cylindrical electromagnet, the following assumptions are made in the work.

The cylindrical electromagnet has the same values of the following parameters as the square current winding: length L, mass m and the created magnetic moment M. When calculating the magnetic field strength, the cylindrical electromagnet was considered to be located at the location of one of the sides of the square winding (Fig. 2). The center of the coordinate system in this calculation was tied to the center of the electromagnet core, and the axis of the applique was directed along its axis of symmetry in the direction of the created magnetic moment.

To analyze the influence of the overall dimensions of the spacecraft on magnetic compatibility, a set of five generalized satellite models with the main parameters given in Table I. The smallest of the models has parameters that are close to the student nano-satellite PolyITAN-1, which was launched into orbit on June 19, 2014 [9]. The parameters of the average of the considered micro-satellite models are similar to the Ukrainian Micron spacecraft, which was launched into orbit on December 24, 2004. To simplify the comparison, the considered spacecraft models have a cubic shape, similar to the CubeSats standard, with an overall size of L_{SK}. In addition, Table I contains data on the mass m_{SK}, the magnetic moment M created on each spatial coordinate axis, the average daily electric power W_{SK} and the voltage U of the power system.

IV. Magnetic Field of a Square Current Winding

According to the definition of the magnetic moment created by a square current winding, its plane should be as technologically possible as possible. Therefore, in practice, the length of the side of the square winding L is usually 80–90% of the L_{SK} of the overall size of the satellite. Using the data from Table I, the parameters of the square current winding were calculated for the corresponding satellite model. It was assumed that the used electric power W_w of the square winding is equal to 15% of the average daily power W_{SK} of the satellite.

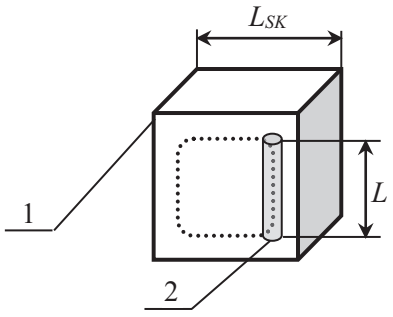


Fig. 2. – The location of the cylindrical electromagnet at the location of the current winding: 1 – satellite model, 2 – electromagnet.

From which the formulas were used to calculate the current and resistance R of the square current winding:

$$I = \frac{0.15 \cdot W_{SK}}{U}, \quad (2)$$

$$R = \frac{U^2}{0.15 \cdot W_{SK}}. \quad (3)$$

The number of turns N_w of the square winding was calculated based on the value of the nominal magnetic moment by the formula:

$$N_w = \frac{M}{I \cdot L^2}. \quad (4)$$

The specific resistance ρ of the winding wire per 1 m of its length was calculated as the ratio of the winding resistance to the total length of the wire:

$$\rho = \frac{R}{4L \cdot N_w}. \quad (5)$$

According to the obtained values of specific resistance from the reference tables of copper wire properties, its diameter d_w was selected. Using these tables of copper wire properties, the mass of the square current winding wire m without fastening elements and insulation was determined (Table II).

Using (1), the magnetic field strengths were calculated as a function of the relative distance x/L to the center of the square current winding in five satellite models (Fig. 3).

From the analysis of the magnetic field dependences shown in Fig. 3, it follows that the increase in the mounting density, which is inherent in the reduction of the satellite size, leads to an increase in the absolute value of the magnetic interference. Thus, for smaller spacecraft, the magnetic interference created by the electromagnet exacerbates the problem of magnetic compatibility of the equipment inside the satellite. This is especially true for elements and devices that are located at distances less than $L/2$ from the current winding.

In addition, from the coincidence of the results of the calculation of magnetic interference for the first and second models of the nano-satellite, it follows that the magnetic field strength of the square winding is proportional to the created magnetic moment M and inversely proportional to the cube of the linear size L^3:

$$H \sim \frac{M}{L^3}. \quad (6)$$

Therefore, an increase in the second model by two times the size of the square winding and eight times the nominal magnetic moment compared to the first model leads to the same absolute values of the magnetic field at the same relative distances x/L.

TABLE I. Basic Parameters of Satellite Models

№	L_{SK}, m	m_{SK}, kg	M, A·m^2	W_{SK}, Watt	U, Volts
1	0.1	1.0	0.2	1.5	3
2	0.2	8	1.6	6	6
3	0.3	25	5	13.5	12
4	0.4	50	10	24	24
5	0.5	75	15	37.5	30

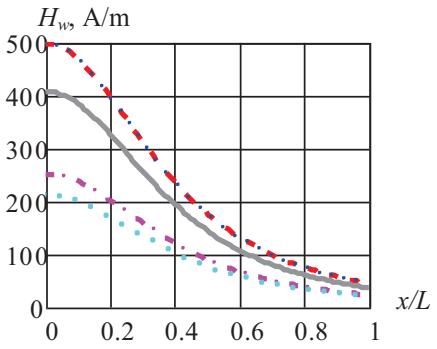


Fig. 3. –Dependences of the magnetic field as a function of the relative distance to the center of the square winding for five spacecraft models with overall dimensions: red line – 0.1 m; blue line – 0.2 m; gray line – 0.3 m; purple line – 0.4 m; cyan line – 0.5 m..

Analysis of the magnetic field distribution inside all models of the spacecraft indicates the possibility of a decrease in the absolute value of the magnetic interference magnitude with distance from the square winding.

V. MAGNETIC FIELD OF A CYLINDRICAL ELECTROMAGNET

The main design variant of the electromagnet of the satellite orientation control system is based on a combination of a cylindrical current winding, inside of which is located an inductively magnetized core [6]. To generalize the comparison of the magnetic field created by a cylindrical electromagnet, its mathematical model is considered simplified according to [10]. In this design variant, almost all the magnetic moment is created by a ferromagnetic core in the form of a cylinder, having a length L and a diameter D_c. The values of the geometric parameters of the cylindrical electromagnet were chosen to ensure equality with the following parameters of a square winding: length L, mass m, outer diameter of the current winding D_{out} and the created magnetic moment M. In addition, Table III contains data on the diameter of the ferromagnetic core D_c, the number of turns N_{el}, the diameter of the wire d_{el} of the cylindrical winding, the used electric power W_{el} and the maximum possible magnetic moment M_{max} at the power specified in Table II. The values of the parameters of the five cylindrical electromagnets presented in Table III are not optimal for the described satellite models, but are close to those actually created and used in spacecraft.

TABLE II. PARAMETERS OF A SQUARE CURRENT WINDING

№	L, m	N_w	ρ_w, Ohm/m	W_w, Watt	d_w, mm	m, kg
1	0.08	469	0.3	0.2	0.255	0.0753
2	0.16	416	0.15	0.9	0.38	0.274
3	0.25	480	0.15	2	0.38	0.494
4	0.37	487	0.225	3.6	0.315	0.512
5	0.45	395	0.225	5.625	0.315	0.505

TABLE III. PARAMETERS OF A CYLINDRICAL ELECTROMAGNET

№	L, m	D_{out}, mm	D_c, mm	N_{el}	d_{el}, mm	W_{el}, Watt	M_{max}, A·m^2
1	0.08	12.5	6.0	4672	0.24	0.021	0.6233
2	0.16	20.0	9.6	6516	0.35	0.049	8.613
3	0.25	18.9	10.0	7864	0.37	0.125	17.9
4	0.37	15.7	10.8	6918	0.35	0.44	28.4
5	0.45	14.2	10.0	5294	0.38	0.845	32.0

The calculation of the magnetic field strength inside the satellite was performed assuming the location of a cylindrical electromagnet instead of a square current winding along one of its sides. The formula from [10] for the magnetic field strength H_{el} in the plane of the central cross-section of the electromagnet was used for the calculation:

$$H_{el} = \frac{4M}{\pi^2 k L D_c} \int_0^{\infty} I_1\left(\lambda \frac{D_c}{2}\right) K_0\left(\lambda \sqrt{x^2 + \frac{L^2}{4}}\right) \sin\left(\frac{\lambda k L}{2}\right) d\lambda, \quad (7)$$

where K_0 and I_1 – modified Bessel functions, x – the coordinate of the field observation point lying on the axis that intersects the center of the spacecraft and is perpendicular to its boundary side. The coefficient k is determined according to the 1% deviation of the longitudinal dependence of the magnetic field created by a cylindrical current winding on its axis of symmetry, and lies within 0.75–0.95 [10].

For five satellite models, the dependences (Fig. 4) of the magnetic field strength as a function of the relative distance x/L were obtained using (7).

From the analysis of the dependencies presented in Fig. 4, it follows that the magnitude of the magnetic field created by a cylindrical electromagnet inside the satellite increases with a decrease in the overall size of the spacecraft. This trend is explained by an increase in the mounting density with a decrease in the size of the satellite, which requires a larger magnetic moment per unit volume to ensure the reliability of the orientation control system. The absolute values of the magnetic field, as in the case of a square winding, are proportional to the magnetic moment and inversely proportional to the cube of the overall size at the same mounting density. In addition, a comparison of the magnitudes of the magnetic field created by a square winding (Fig. 3) and similar values in Fig. 4 determines the following advantage of using a cylindrical electromagnet. The magnetic interference created by a square winding is 3–4 times greater than that created inside a similar model of a satellite by a cylindrical electromagnet.

The rationality of using cylindrical electromagnets is demonstrated by the graphs in Fig. 5. The graphs show the dependences of the magnetic field reduction in the case of replacing the square winding with a cylindrical electromagnet, calculated by the formula:

$$\Delta = \frac{H_w}{H_{el}}. \quad (8)$$

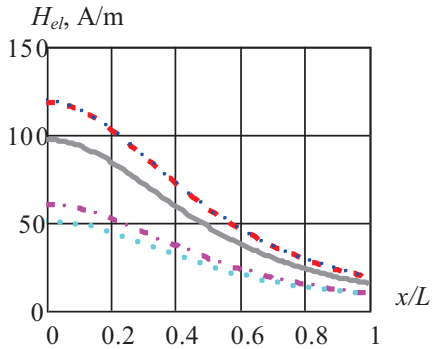


Fig. 4. – Dependences of the magnetic field as a function of the relative distance to the center of the boundary surface, where the electromagnet is located, for five SK models with overall dimensions: red line – 0.1 m; blue line – 0.2 m; gray line – 0.3 m; purple line – 0.4 m; cyan line – 0.5 m.

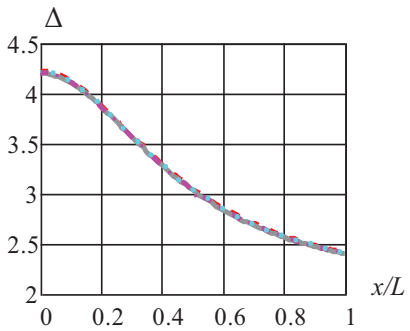


Fig. 5. –Dependencies of the ratio of the magnetic field of a square current winding and a cylindrical electromagnet, for five models of spacecraft with overall dimensions: red line – 0.1 m; blue line – 0.2 m; gray line – 0.3 m; purple line – 0.4 m; cyan line – 0.5 m.

The lines of the graphs in Fig. 5 are similar to those used in Fig. 3 and Fig. 4 and correspond to the sizes of the five satellite models considered. The coincidence of functional dependencies of magnetic interference reduction when using a cylindrical electromagnet instead of a square current winding demonstrates independence from the absolute values of the spacecraft dimensions when the specified conditions for replacing the magnetic source are met.

VI. Specific Magnetic Moment of the Electromagnet of the Satellite Orientation Control System

Among the limiting parameters that determine the direction of optimization of space technology, the used electric power and the mass of the device are always considered. Therefore, when comparing two variants of the design of an electromagnet, it is advisable to analyze the dependence of its main operating characteristic, magnetic moment, on electric power and mass. The last two parameters can be used to determine the specific mass and power of the magnetic moment in accordance with the five spacecraft models considered in the work. Let us introduce the specific magnetic moment for a square current winding M_w and for a cylindrical electromagnet M_{el} according to the formulas:

$$M_w = \frac{M}{m\ W_w}, \tag{9}$$

$$M_{el} = \frac{M}{m\ W_{el}}. \tag{10}$$

The constructed graphs of the dependence of the specific magnetic moment for two variants of the electromagnet design on the overall size of the spacecraft are shown in Fig. 6 by red and blue lines. For a direct comparison of these results, Fig. 6 shows the ratio of the specific magnetic moments M_{el}/M_w created by a cylindrical electromagnet and a square current winding under the comparison conditions specified above.

From the analysis of the obtained results it follows that the specific magnetic moment of a cylindrical electromagnet is many times greater than that of a square current winding. At the same time, the general trend is an increase in the efficiency of using cylindrical electromagnets instead of current windings with a decrease in the overall dimensions of the satellite. The scatter of the results of the ratio of specific magnetic moments M_{el}/M_w is explained by the non-optimal parameters of the cylindrical electromagnets used for comparison.

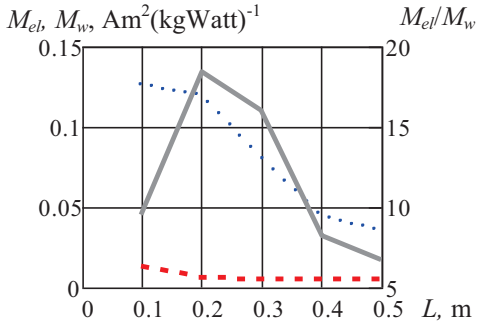


Fig. 6. – Dependences of the specific magnetic moment on the overall size of the spacecraft: for the current winding – red line, for the cylindrical electromagnet – blue line, scale on the left; the ratio of the specific magnetic moments of the cylindrical electromagnet and the square current winding as a dependence on the overall size of the spacecraft – gray line, scale on the right.

VII. Conclusions

As a result of the analysis of the magnetic interference created by the square current winding in the central part of the nano-satellite, a magnetic field strength level of 250 A/m was obtained. For micro-satellites, a similar level of magnetic interference is 50–150 A/m with a tendency to decrease with increasing overall size of the spacecraft. In the case of using a cylindrical electromagnet with the same length and magnetic moment instead of a square current winding, the indicated levels of magnetic interference are reduced by 2–4 times regardless of the size of the satellite. The specific magnetic moment per unit of mass and used electric power, which is created by a cylindrical electromagnet, is 5–15 times greater than that of a current winding.

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Enhancing LoRa Encryption with Post-Quantum Algorithms for IoT Security

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***Abstract*—The growing demand for secure, long-range wireless communication in the Internet of Things (IoT) has led to widespread adoption of LoRa technology. However, traditional LoRa implementations remain vulnerable to quantum computing attacks. This paper proposes the integration of post-quantum cryptographic algorithms, specifically Niederreiter encryption and LDPC codes, into LoRa systems to enhance their resilience. We explore the theoretical foundation of these algorithms, practical implementation strategies, and their impact on system performance. Experimental results and security analyses validate the feasibility of this approach, establishing a pathway for future-proof secure communication in LoRa-based IoT networks.**

Keywords— LoRa, post-quantum cryptography, IoT security, LDPC, Niederreiter, ESP32, UTM mapping, wireless encryption

I. Introduction

LoRa (Long Range) technology has become a cornerstone of many IoT applications due to its low power consumption, extended range, and simplicity. However, as quantum computing advances, the cryptographic primitives currently used in LoRa implementations (e.g., AES, ECC) face potential vulnerabilities. Quantum computers could, in theory, solve discrete logarithm and integer factorization problems exponentially faster than classical computers, rendering current encryption mechanisms obsolete.

To address this issue, researchers have proposed the adoption of post-quantum cryptographic (PQC) algorithms that remain secure even in the presence of quantum computing capabilities. Among the most promising PQC candidates are code-based cryptosystems, such as the Niederreiter scheme, which is based on the hardness of decoding linear error-correcting codes like LDPC (Low-Density Parity-Check) codes.

In this paper, we propose a novel hybrid system that enhances existing LoRa communication by integrating PQC methods. We provide a theoretical background on Niederreiter encryption and LDPC codes, describe an implementation architecture for embedding them into LoRa systems, and evaluate the practical impact of such integration in terms of security, performance, and scalability

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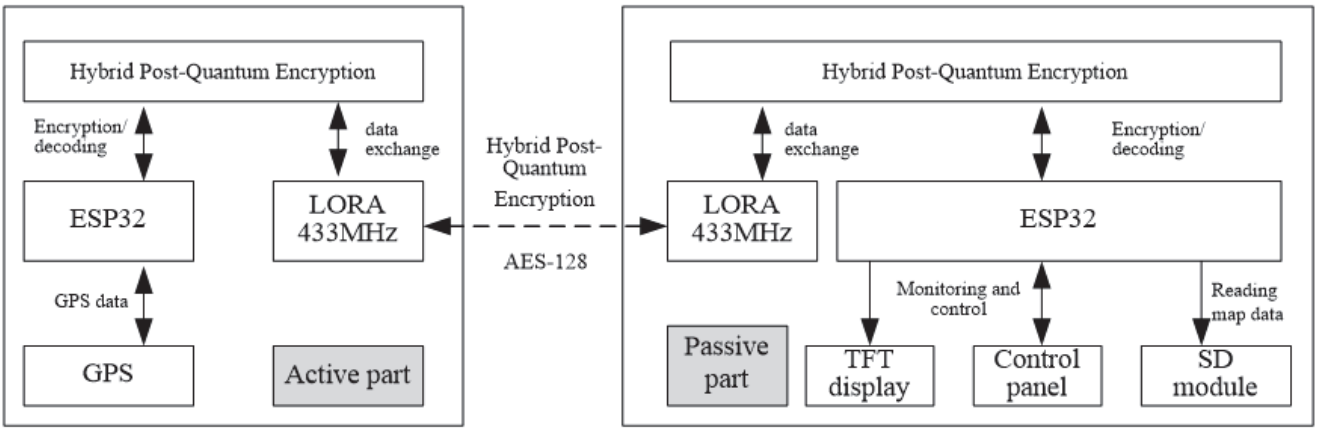


Fig. 1. Concept Post-Quantum Algorithms Encryption

II. Main Research

Several prior works have explored the intersection of LoRa security and emerging cryptographic techniques. Traditional enhancements focus on improving AES key exchange protocols or introducing ECC-based mutual authentication. However, these are not inherently resistant to quantum attacks. On the other hand, research into lightweight PQC is

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ongoing, with algorithms like BIKE, Classic McEliece, and NTRU being evaluated by NIST.

Few studies, however, directly target LoRa's low-resource constraints and long-range communication requirements. Our work differentiates itself by providing a concrete application of code-based cryptography tailored for LoRa's specific architecture.

LoRa Communication Overview. LoRa is a physical layer modulation technique derived from chirp spread spectrum (CSS) technology. LoRaWAN, the corresponding network protocol, adds support for MAC-level management. LoRa supports star topology, operates in unlicensed ISM bands (e.g., 433 MHz, 868 MHz), and typically uses AES-128 for encryption.

Post-quantum cryptography refers to cryptographic algorithms considered secure against both classical and quantum adversaries. The most relevant families include lattice-based, multivariate, hash-based, and code-based schemes.

The Niederreiter cryptosystem is a dual form of the McEliece cryptosystem. It is based on the difficulty of decoding a general linear code. In the Niederreiter scheme:

– Public key: a parity-check matrix of a chosen linear code (e.g., LDPC).

– Private key: transformation matrices and the decoding algorithm for the chosen code.

– Encryption: encode a message as a random error vector and compute its syndrome.

– Decryption: decode the error vector using the private key.

This scheme is known for fast encryption and decryption times and its robustness against quantum attacks.

Low-Density Parity-Check codes are a class of linear error-correcting codes characterized by sparse parity-check matrices. They enable efficient error correction with linear-time encoding and decoding algorithms, such as belief propagation.

With the emergence of quantum computing, classical encryption methods (e.g., RSA, ECC) are becoming susceptible to attacks using Shor's algorithm. This poses a serious threat to secure LoRa-based communication systems, particularly those deployed in sensitive environments such as wildlife tracking, military, or border surveillance.

To future-proof the system, our objective is to integrate post-quantum cryptographic methods that can operate efficiently on constrained hardware such as ESP32 and within the bandwidth limitations of LoRa.

TABLE I. COMPARATIVE ANALYSIS BASED ON THESE FACTORS, NIEDERREITER WITH LDPC CODES IS SELECTED FOR PROTOTYPE IMPLEMENTATION.

Feature	Niederreiter (LDPC)	McEliece (Classic)
Key size	1.3 MB	1.5 MB
Security level	256-bit (quantum-safe)	256-bit
Performance (ESP32)	Moderate (via optimization)	High RAM usage
Code complexity	Medium	High

Our proposed system integrates the Niederreiter encryption scheme into the LoRa communication framework by modifying both the sender (ESP32-based transmitter) and receiver (LoRa gateway or mobile station).

The previously developed mapping system includes the following core elements:

– ESP32 microcontrollers for computation and communication;

– LoRa E32 433T30D modules for long-range communication;

– Offline maps using UTM tiles stored locally on an Android device;

– Real-time location tracking using GPS modules.

Communication protocol for transmitting coordinates and device status via UART over LoRa.

Android application for visualizing map and beacon positions.

Encryption Flow:

– Message preparation: sensor data or GPS location data is prepared;

– Syndrome calculation: data is encoded using the Niederreiter public key;

– Transmission: the encrypted message is transmitted over LoRa.

– Decryption Flow:

– Reception: encrypted message is received;

– Decoding: the syndrome is decoded using the private key (parity-check matrix and decoding algorithm);

– Message reconstruction: original data is recovered and logged.

A test was conducted with the proposed system in an outdoor environment with varying transmission distances (0.5 km – 5 km) and under typical LoRa constraints (SF7-SF12, BW 125 kHz).

Comparative Characteristics of Cryptographic Algorithms.

Table 2 provides a comparative analysis of different cryptographic algorithms in terms of quantum security, key size, and encryption/decryption times. The key findings are:

TABLE II. COMPARATIVE CHARACTERISTICS OF CRYPTOGRAPHIC ALGORITHMS

Algorithm	Quantum Security	Key Size (KB)	Encryption Time	Decryption Time
AES-128	Low	0.016	Very Fast	Very Fast
ECC (256-bit)	Low	0.032	Fast	Fast
NTRU	High	1.2	Medium	Medium
Niederreiter+LDPC	High	0.9	Fast	Fast

– Niederreiter (LDPC) offers a good balance between security and performance, with moderate key size and fast encryption/decryption times on ESP32 hardware. This makes it suitable for LoRa-based systems with constrained resources;

– AES-128 and ECC (256-bit) are fast but offer low quantum security, making them unsuitable for future proofing against quantum attacks;

– NTRU provides high quantum security but has higher key sizes and medium encryption/decryption times, making it less efficient for LoRa's bandwidth constraints;

– The proposed Niederreiter+LDPC combination is highlighted as the optimal choice for integrating post-quantum security into LoRa systems due to its high quantum security, moderate key size, and efficient performance on ESP32.

Despite added cryptographic overhead, the packet delivery ratio (PDR) remained above 95% for distances up to 3 km, dropping to ~85% beyond 4 km in dense urban settings.

Using ESP32 deep sleep between transmissions, the added encryption processing increased average power consumption by ~8%, which remains within acceptable limits for battery-powered applications.

The adoption of Niederreiter encryption with LDPC codes ensures that the system remains secure even against quantum attacks. The security relies on the decoding problem of LDPC codes, which is NP-hard and remains so even with quantum solvers.

The system resists classical attacks (e.g., brute force, replay) and quantum attacks (e.g., Grover's, Shor's algorithms) due to the lack of algebraic structure in the cryptographic primitives used.

Encryption/Decryption Overhead. Table 3 presents the average time and LoRa packet overhead associated with encryption and decryption processes.

Encryption/Decryption Overhead

TABLE III. ENCRYPTION/DECRYPTION OVERHEAD

Operation	Avg Time (ms)	LoRa Packet Overhead (%)
Encryption	3.2	+15%
Decryption	2.7	+15%

The key findings are:

– Encryption and decryption using the Niederreiter scheme add an average overhead of 15% to LoRa packets;

– Despite the added cryptographic overhead, the packet delivery ratio remains high (above 95%) for distances up to 3 km, indicating that the system maintains good performance even with the added security layer;

– The average power consumption increase due to encryption processing is around 8%, which is acceptable for battery-powered applications;

– The decryption time of 2.7 ms and encryption time of 3.2 ms on ESP32 demonstrate that the system can handle cryptographic operations efficiently within the constraints of LoRa communication.

To preserve system performance while introducing cryptographic enhancements, we implemented the following modifications:

– Lightweight cryptographic library for Niederreiter-based key exchange;

– Key compression and fragmentation mechanisms to send large keys over LoRa;

– Use of symmetric encryption (e.g., AES-128 or ChaCha20) for message payloads, with session keys established via post-quantum key exchange;

– Optimized LDPC-based key generation using ESP32's onboard flash and PSRAM.

Post-quantum cryptography can be applied selectively, such as during session initialization, while using symmetric encryption for routine data. This hybrid approach balances security and performance.

Security benefits:

– Quantum-resilient communication channels;

– Immunity to known public-key attacks;

– Suitable for confidential or sensitive field operations.

Performance trade-offs:

– Larger payloads require fragmentation in LoRa, especially with small MTU;

– Key management becomes more complex;

– Requires additional RAM and flash memory usage;

Through testing, we established that the LDPC-based approach provides the best compromise for ESP32-based systems due to:

– Moderate key sizes;

– Lower energy consumption;

– Faster decoding on constrained hardware.

We also implemented a fallback mechanism: if session key exchange fails (e.g., due to poor signal), the system falls back to pre-shared symmetric keys.

To enhance reliability, we introduced CRC validation after each key fragment and retransmission requests. These layers ensure that the key exchange remains robust in challenging RF environments.

Furthermore, application-level safeguards were added to detect and ignore replayed packets using timestamps and nonce values.

TABLE IV. POWER AND BANDWIDTH IMPLICATIOS

Operation	Energy Cost (mJ)	Transmission Time (LoRa, SF12)	Notes
GPS Location (unencrypted)	2.5	250 ms	Single frame
Encrypted (AES-128)	3.2	300 ms	+30% message size
PQ Key Exchange (Niederreiter)	15	2–4 sec (fragmented)	Per session, 1/day or on reconnect
PQ Key Exchange (LDPC)	9	1.5–3 sec (fragmented)	More efficient than Niederreiter

The proposed algorithm enables secure post-quantum key exchange using structured LDPC codes derived from the Niederreiter cryptosystem. We use a binary Goppa or LDPC code with parameters optimized for ESP32 and LoRa constraints.

Fig. 2. Formulas for sparse parity-check matrices

$$H \in \mathbb{F}_2^{(n-k)\times n}$$
$$H_{pub} = P \cdot H$$
$$e \in \mathbb{F}_2^n$$
$$s = H_{pub} \cdot e^T$$
$$K = \mathrm{KDF}(e)$$
$$s' = P^{-1} \cdot s$$
$$\hat{e} = \mathrm{Decode}(H, P^{-1} \cdot s)$$
$$K = \mathrm{KDF}(\hat{e})$$

The LDPC-based key exchange follows a code-based approach similar to Niederreiter but optimized using sparse parity-check matrices

Choose a binary LDPC code with parameters n, k, d where:

– n: code length,
– k: dimension,
– d: minimum distance.

Key Generation (by receiver):

– random permutation matrix P;
– Public Key(H_{pub}), Private Key(H,P).

Encryption (by sender):

– Generate random binary error vector e;
– Compute syndrome s;
– Transmit: syndrome s as ciphertext;
– The shared session key is derived via K, where KDF is a secure Key Derivation Function (e.g., SHAKE256).

Decryption (by receiver):

– Apply inverse permutation s';
– Decode syndrome s' using syndrome decoding to retrieve e;
– Recompute for K.

Power and Bandwidth Implications. Table 4 details the energy cost and transmission time implications of different operations in the system. For constrained LoRa networks, the effective parameters

TABLE V. FOR CONSTRAINED LORA NETWORKS, THE EFFECTIVE PARAMETERS

Parameter	Value
n (code length)	1024
k (dimension)	524
w (error weight)	40
Public Key Size	~2.5 KB
Syndrome Size	500 bits (~63 bytes)
Decoding Time	8–12 ms (ESP32)

The key findings are:

– GPS location transmission without encryption consumes 2.5 mJ and takes 250 ms, while encrypted transmission (using AES-128) increases energy consumption to 3.2 mJ and transmission time to 300 ms, representing a 30% increase in message size;

– Post-quantum key exchange using Niederreiter requires 15 mJ and takes 2-4 seconds per session, while using LDPC reduces the energy cost to 9 mJ and transmission time to 1.5-3 seconds;

– The LDPC-based approach is more efficient in terms of energy and time for post-quantum key exchange, making it a preferred choice for constrained LoRa networks;

– The effective parameters for LDPC (code length, dimension, error weight, public key size, syndrome size, and decoding time) are optimized for ESP32 and LoRa constraints, ensuring that the system remains secure and efficient.

The LDPC decoding is done using the belief propagation algorithm or bit-flipping algorithm, both of which can be implemented efficiently on ESP32.

The hardness of decoding random LDPC codes is equivalent to the Syndrome Decoding Problem (SDP), which is NP-complete and remains resistant even to quantum attacks. Unlike lattice-based systems, the code-based cryptography is not vulnerable to structure-specific quantum attacks such as those targeting NTRU or LWE.

III. CONCLUSION

This work demonstrated how to adapt and extend a previously implemented LoRa mapping system by incorporating post-quantum cryptographic schemes. While preserving the offline map and UTM tile infrastructure, we added strong cryptographic primitives resilient against quantum attacks.

Key findings include:

– LDPC-based cryptographic methods provide an effective trade-off between security, size, and processing requirements;

– Niederreiter encryption remains a viable option where key size is less constrained and security is paramount;

– Session-based key exchange using post-quantum techniques is feasible on ESP32, even under LoRa's strict bandwidth constraints.

Our improvements position the system for long-term deployment in scenarios where data confidentiality, integrity, and resistance to future cryptographic threats are essential.

Future work could explore:

– Hardware acceleration of matrix operations for faster key generation.

– Adaptive encryption schemes based on signal strength and channel capacity.

– Integration with standardized PQC libraries from NIST candidates.

With these enhancements, LoRa-based systems can now operate securely not only in present-day deployments but also in a post-quantum future, ensuring reliable, long-range, and tamper-resistant communication.

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Comparative Analysis of Power Losses in Different Network Topologies and Identification of Optimal Configurations

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***Abstract* — The paper presents a comparative analysis of power losses in electrical distribution networks with different topologies: ring and radial. Using a fragment of a real network scheme with voltage levels of 110, 35, and 10 kV, operating mode calculations were performed under normal conditions and with various line disconnection scenarios. The influence of network configuration on the redistribution of currents and on power losses in lines and transformers is investigated. Based on the obtained results, the technical and economic performance of radial and ring configurations is compared, indicating general conditions for their preferable use. An increase in losses in transformers and lines is identified as a result of current redistribution under certain network configurations. The findings of the study can help improve the efficiency of designing and operating distribution networks.**

Keywords — electrical networks; power losses; radial configurations; ring configurations; current redistribution; network configuration optimization.

I. Introduction

Modern distribution electrical networks impose increasingly high requirements on the efficiency and reliability of power delivery. One of the key factors affecting the operational performance of a network is power loss. The network structure and topology play a decisive role in determining the overall level of losses. In practice, radial and ring distribution schemes are the most widespread. Radial schemes are characterized by simplicity, ease of operation, and relatively low cost. Ring schemes, in turn, provide increased supply reliability due to the possibility of redundancy, but they require more complex control systems and often lead to higher loss levels if the configuration is not optimized properly [1].

The aim of the work is to perform a comparative analysis of power losses under different network configurations based on a real network scheme with three voltage levels — 110 kV, 35 kV, and 10 kV. Special attention is paid to assessing the impact of disconnecting individual lines on the redistribution of currents, voltages and, accordingly, losses in both lines and transformers. Modernized variants of the scheme with modified load parameters and added generation are also considered, which allows a techno-economic evaluation of the efficiency of different configurations. The work is aimed at identifying patterns and formulating recommendations for selecting optimal schemes with minimal power losses and high operational reliability.

II. Materials and Methods

Network Model: The electrical network was modeled using the Python library Pandapower, which is designed for power flow analysis of distribution networks. Pandapower is based on solving power flow equations using the Newton–Raphson method. Initially, the library was created on the basis of PYPOWER, but over time it has been significantly improved in characteristics such as calculation stability, performance, and interface convenience [2]. The network topology and base-case parameters (line types and lengths, transformer ratings, and load levels) replicate the dataset reported in [8]. Building on that dataset, our contribution focuses on comparative loss analysis under ring vs. radial configurations and N−1 line-outage scenarios, with and without distributed generation. In particular, a real distribution scheme with voltage levels of 110, 35, and 10 kV was chosen, encompassing 12 nodes. Fig. 1 shows the one-line diagram of the network, and Fig. 2. shows the equivalent circuit model of the network. The scheme includes overhead transmission lines, power transformers, loads, and power sources. Generation in the system is represented by the slack bus at node 1 (115 kV) and node 2, as well as by generation at the 10 kV buses of node 12. Network parameters, in particular the types of conductors, transformers, and loads, were determined based on reference data and current standards.

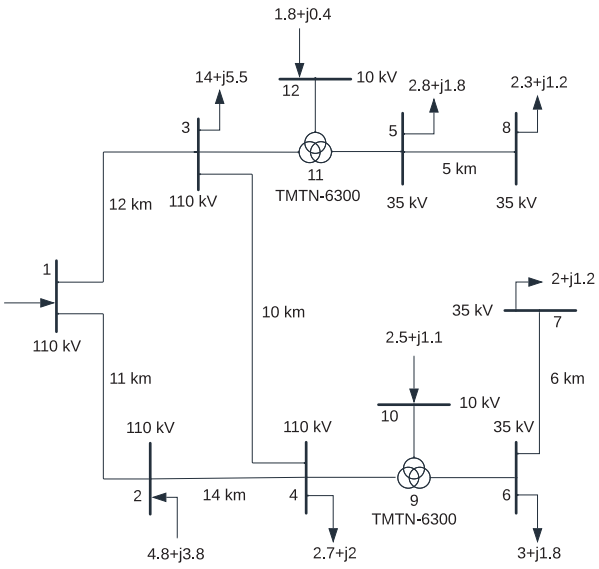


Fig. 1. Diagram of the studied distribution network (110/35/10 kV) — adapted from [8]

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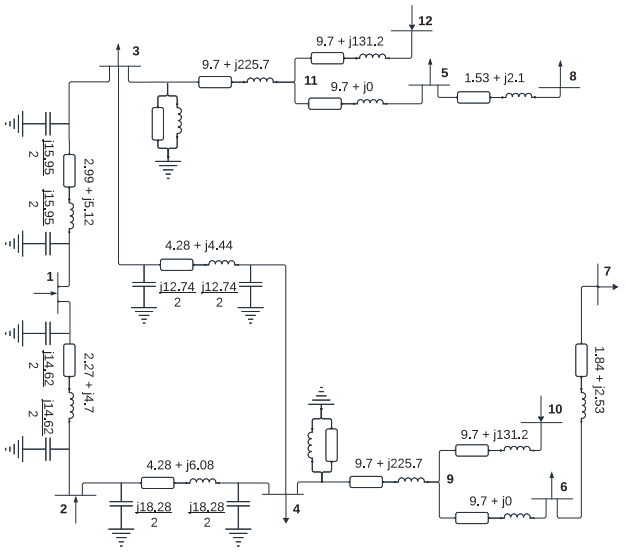


Fig. 2. Equivalent circuit of the electrical network — adapted from [8]

The parameters of the conductors and transformers that form the main sections of the network are presented in Table 1.

TABLE I. PARAMETERS OF NETWORK SECTIONS

Section	Conductor / Transformer Type	l, km / Transformer Winding	r_0, Ohm / km	x_0, Ohm / km	b_0, µS / km	R, Ohm	X, Ohm	B, µS
1-2	AC-120	11	0.249	0.427	2.658	2.74	4.7	29.24
1-3	AC-120	12	0.249	0.427	2.658	2.99	5.12	31.90
2-4	AC-95	14	0.306	0.421	2.611	4.28	6.08	36.55
3-4	AC-70	10	0.428	0.432	2.547	4.28	4.44	25.47
5-8	AC-95	5	0.306	0.421	-	1.53	2.10	-
6-7	AC-95	6	0.306	0.421	-	1.84	2.53	-
3-11	TMTN-6300	HV	-	-	-	9.70	225.7	-
11-5		MV	-	-	-	9.70	0	-
11-12		LV	-	-	-	9.70	131.2	-
4-9	TMTN-6300	HV	-	-	-	9.70	225.7	-
9-6		MV	-	-	-	9.70	0	-
9-10		LV	-	-	-	9.70	131.2	-

"-" denotes a parameter not modeled / not applicable for the given HV/MV/LV segments; shunt susceptance (charging power) is explicitly considered only for the 110-kV lines.

The transformer impedances were calculated based on reference data and converted to per-unit (p.u.), as required by the Pandapower framework. The line elements account for the resistance, reactance, and capacitive characteristics corresponding to their route lengths, and charging power (capacitive reactive power) is also taken into consideration.

One feature of the calculations is the way Pandapower reports reactive power losses: it outputs them as the sum of inductive losses and charging (capacitive) power. For a correct loss analysis, the charging power was calculated separately and then compensated using shunt elements so that inductive losses could be isolated.

Simulation Scenarios: The experimental part of the study was divided into two stages:

• Stage 1 – Base Case and Line Outages: The base configuration of the network was modeled with the given loads and generation (at node 2 and node 12). In this base case, all lines are in service (the ring is closed). Then, N-1 contingency scenarios were simulated by sequentially disconnecting individual lines: 1–2, 1–3, 1–4, 2–4, 3–4, etc. This allowed us to observe how power flow redistribution, voltage changes, and losses in lines and transformers change with each line outage.

• Stage 2 – Effect of Removing Local Generation: The second stage involved analyzing the network with the distributed generation at node 12 disconnected. For each of the line outage scenarios from Stage 1, the power flow calculations were repeated without the local generation. This made it possible to assess the impact of the local generation source on the overall efficiency of the network under each topology scenario.

III. RESULTS AND DISCUSSION

A. Power Losses in the Base Ring Configuration

In the base configuration of the 110/35/10 kV network (all lines in service, ring closed), the total active power losses amount to about 0.31 MW. Of these, losses in the overhead lines are 0.16 MW, and in the power transformers are 0.15 MW. The reactive inductive power losses in the lines of the base scheme are 0.26 MVAr. All these results are summarized in Table 2. At the same time, due to the capacitive currents of the 110 kV lines, the network actually generates reactive power (so-called charging power): in the initial scheme, the 110 kV lines generated about 1.1 MVAr. This proves that we need to consider charging power when analyzing operating modes. The obtained loss values are consistent with standard estimates for distribution networks: greater distance from the power source to the load causes higher energy losses [3]. Reducing these losses is possible by increasing conductor cross-sections, optimizing network topology, and adding generation sources closer to consumers (these approaches are discussed in more detail below).

TABLE II. RESULTS OF POWER LOSS CALCULATIONS FOR THE BASE CLOSED-RING NETWORK.

Voltage level	ΔP, MW in lines	ΔP, MW in transformer windings	ΔQ, MVAr in lines	ΔQ, MVAr in transformer windings
Total	0.161	0.154	0.261	2.073
110 kV	0.145	0.083	0.238	1.940
35 kV	0.017	0.061	0.023	0.000
10 kV	-	0.010	-	0.133

"-" denotes not applicable in this model. The 10-kV level has no explicitly modeled lines; only transformer LV windings are present, hence no 10-kV line-loss entries are reported.

B. Power Losses under Line Outages (Radial Mode)

To analyze the impact of network topology on power loss, N-1 contingency scenarios were examined where individual lines of the ring network are disconnected (specifically lines 1–2, 1–3, 1–4, 2–4, and 3–4). In each such case, the supply to certain network sections had to operate in a radial mode (the ring is opened at the point of outage). It was found that the change in losses depends on which line is disconnected. For example, opening line 1–3 causes node 3 (and partly node 4) to be fed by a longer path via nodes 2 and 4, i.e. through the longer loop 1–2–4–3 instead of the direct 12-km line 1–3. Accordingly, the total path resistance increases significantly, leading to additional energy losses (in this case, losses tripled). In contrast, if line 3–4 is opened, the main nodes 3 and 4 are

still fed directly from node 1 through two parallel branches, and the absence of the 3–4 link has almost no effect on the current distribution (losses increased by only ~2%). Thus, under normal conditions the closed-ring topology yields minimal losses, and switching to a radial mode (opening the ring) increases losses, especially if the break occurs in a main trunk line through which a large portion of the load is fed. These observations reinforce the premise that moving from a radial scheme to a ring scheme can reduce power losses and improve the voltage profile [4]. In our calculations, the ring configuration had ~20–25% lower losses compared to the least optimal radial configuration, and about 2% lower losses compared to the most optimal radial configuration, confirming the benefit of closed topologies for loss reduction. In addition, the closed network demonstrated greater flexibility under fault conditions: in the event of a single contingency, the ring network maintained supply to all nodes (albeit at the cost of some increase in losses due to longer supply paths). Literature also notes that implementing ring schemes improves power supply quality and reliability, especially in the presence of distributed generators [5]. The calculation results for the various line-outage scenarios are summarized in Table 3 and illustrated in Figs. 3, 4, 5.

TABLE III. POWER LOSSES UNDER DIFFERENT NETWORK TOPOLOGY SCENARIOS (SINGLE LINE OUTAGES)

Disconnected line	ΔP, MW in lines	ΔP, MW in transformer windings	ΔQ, MVAr in lines	ΔQ, MVAr in transformer windings
none	0.161	0.154	0.261	2.073
3-4	0.163	0.154	0.266	2.077
1-2	0.189	0.156	0.312	2.095
1-3	0.645	0.163	0.896	2.182
2-4	0.322	0.158	0.513	2.126

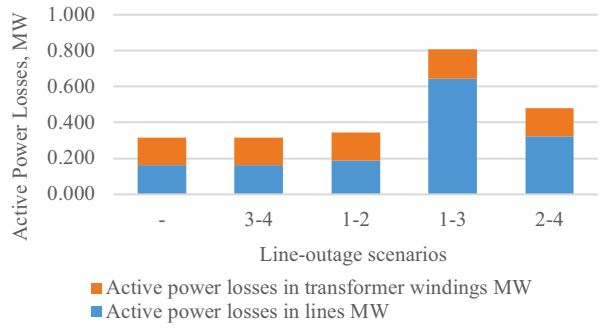


Fig. 3. The total active power losses for different line outage scenarios

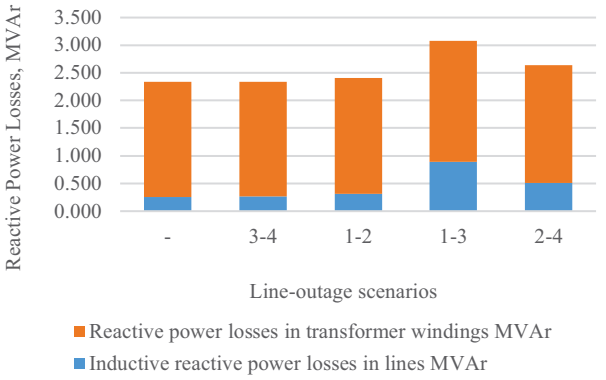


Fig. 4. Ehe total reactive power losses for different line outage scenarios

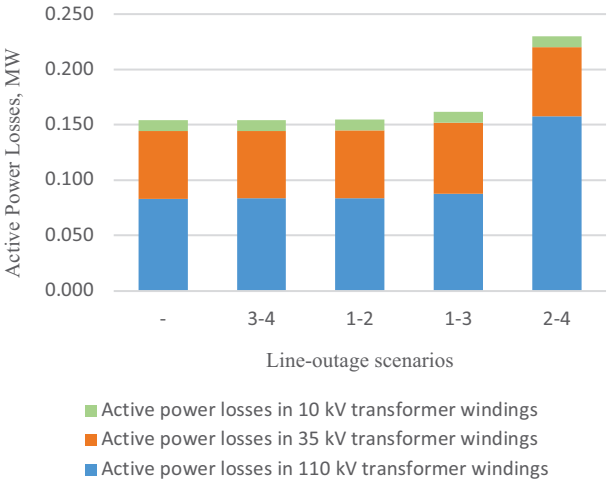


Fig. 5. The active power losses in transformer windings at different voltage levels under each scenario

C. *Effect of Distributed Generation at Node 12*

The introduction of a distributed generation (DG) source at node 12 (on the 10 kV feeder) significantly reduces the power losses in the network. Fig. 6, 7 compare the active power losses for two scenarios: without generation vs. with a 1.8 MW DG connected at node 12.

Fig. 6. Active power losses in lines with and without generation

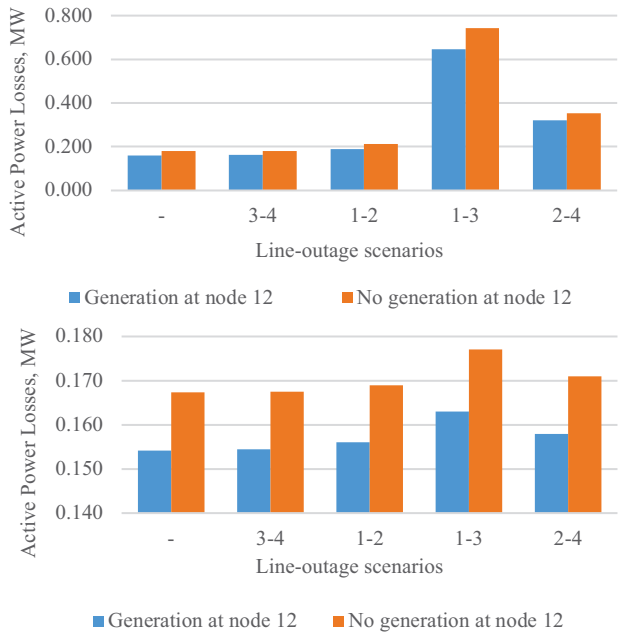


Fig. 7. Active power losses in transformers with and without generation

It can be seen that in steady state (no contingencies), the presence of local generation decreases the active power line losses from ~0.18 MW to ~0.16 MW by partially supplying the load locally, which is consistent with findings of other studies [6, 7]. In contingency cases where lines are disconnected, the effect of the DG varies: if the disconnected line is near the generator, the local source can take on a significant portion of the supply to nodes isolated from the main source, thereby offsetting losses. If the fault occurs far from the generation node, the DG's influence on losses is smaller. Overall, the results support the claim that placing

generation sources closer to consumers offloads the main feeders and improves network efficiency.

D. Ring vs. Radial Topology Optimality

The analysis provides grounds for conclusions regarding the choice of distribution network topology. The ring (closed-loop) configuration demonstrated lower technical losses and greater flexibility under emergency conditions compared to the radial configuration. In normal operation, the presence of parallel paths allows power flows to be split and reduces the loading of each individual line, thus lowering total losses. Under a single contingency, a ring network maintains service to all consumers (each still has at least one supply path), whereas a radial network either requires using backup sources or results in outages or significantly lengthened supply routes, which, as we saw, greatly increases losses. Literature sources note that converting a radial configuration to a ring can reduce losses and improve power supply quality, especially with distributed generation present [5]. Thus, from the perspective of minimizing energy losses, the ring topology is more optimal for this network. At the same time, it should be considered that implementing closed-loop schemes requires more line infrastructure and more complex switching devices, as well as carefully designed protection schemes. Moreover, the efficient operation of such looped systems—especially with integration of renewable energy sources and variable loads—requires ongoing study and improvement of methods to optimize their regimes, for example by voltage regulation to reduce losses [8]. In practice, power distribution often uses compromise approaches: for instance, open rings (normally open at one point, but capable of automatic reconfiguration in case of a fault). This provides near ring-level reliability while operating as radial under normal conditions to avoid circulating currents and simplify control.

Looking ahead, with the increasing share of distributed generation and the development of Smart Grid concepts, one can expect wider adoption of closed-loop and meshed structures, as they better accommodate bidirectional power flows and allow more flexible load sharing while minimizing losses [9]. Our results confirm that with optimal power flow management, a ring topology ensures minimal technical losses and improves network resilience compared to the traditional radial design. Even so, a radial configuration remains technically and economically justified where loads are modest, feeder lengths are short, reliability requirements are moderate, and CAPEX/OPEX constraints apply; in such settings, simpler protection and lower infrastructure costs can outweigh the loss-reduction benefits of ring designs.

CONCLUSIONS

An analysis of power losses in a 110/35/10 kV distribution network under different operating conditions revealed that the ring (looped) configuration has advantages in terms of energy efficiency and reliability. Under normal (no-fault) conditions, the closed topology yields the lowest active and reactive power losses in both lines and transformers. In the event of individual line disconnections, the ring network continues to supply all consumers (each load retains at least one path), whereas a radial network either requires backup sources or leads to outages or significantly longer transmission paths. Therefore, for networks with high reliability requirements and a need for low losses, it is advisable to employ closed (ring or meshed) configurations, provided that the necessary control and protection means are in place. In networks with minor loads or simple distribution schemes, using radial configurations can be justified as a less costly option. It was also found that adding distributed generation to the network significantly reduces losses (by up to ~20%, depending on the scenario) by shortening the energy transmission distance. The results of this work can be used in the design and modernization of distribution power grids to choose the optimal topology from the standpoint of technical losses and reliability.

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Efficient Regulation Modes for Overcharging of Compensation Converter Capacitors

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***Abstract*—The issue of ensuring high energy performance of compensation-type electric energy converters is investigated. It is shown that the use of a new generation of fully controlled devices in the switching link of compensation converters significantly expands the energy capabilities of these means of conversion technology. However, achieving the maximum possible reactive energy generation requires the development and application of certain modes of recharging the capacitors of the commutating link. The result of the conducted research is the determination of regulation modes recommended for practical application.**

Keywords—commutating, control angle electromagnetic process, electric diode, converter, commutating link, control angle

I. Introduction

Developments in the direction of increasing the energy characteristics of electric energy converters have been carried out since the first circuit of such a device was developed. Some researchers use capacitors included in power circuits recharged by industrial frequency currents for this purpose [1, 2]. There are also known circuits of converter units in which capacitors operating at increased voltage and frequency were used to compensate for reactive energy [3]. Despite the abundance of technical solutions, for any compensating converter, the main issue of its operation is the conditions for the formation of the commutating voltage (CV), in other words, the voltage created when recharging the capacitors.

The technological and energy characteristics of the electric energy converter and the efficiency of using its power equipment depend on the magnitude and shape of the CV. In a three-phase bridge compensating converter (circuit, Fig. 1), which is a new technical solution [4], the device intended for the formation of the commutating voltage is the commutating link.

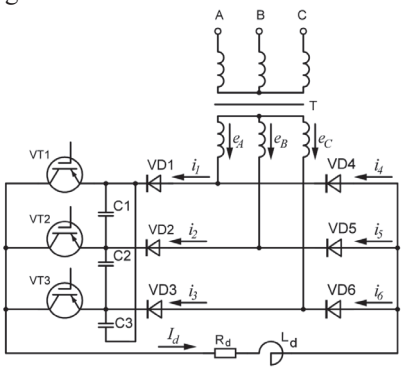


Fig.1.Three-phase bridge compensation converter diagram

It consists of a three-phase capacitor battery (C1, C2, C3) and a three-phase group of IGBT transistors (VT1, VT2, VT3). The three-phase group of transistors provides adjustable recharging of the capacitors, the voltage of which, superimposed on the voltage of the secondary windings of the three-phase converter transformer (T), provides advanced commutating of the current by power electric valves. Such a technical solution is typical for new-generation compensating converters, the development and research of which are in the sphere of interests of the authors of the publication

In the previous generation of compensation converters, the function of the device that provides recharging of the capacitor battery (CB) of the commutating link was performed by a three-phase balancing reactor (TBR). It ensured the passage of the alternating components of the diode currents through the capacitors and the discharge of the constant components of these currents into the external DC circuit, or, in other words, the reactor is an inductive element of the filter that separates the constant and alternating components of the valve currents.

The constant load current I_d is divided by a three-phase balancing reactor into three equal parts. If we neglect the magnetizing current, the reactor phase current $i_R^{(3)}$ will be:

$$i_R^{(3)} = I_R^{(3)} = I_d / 3$$

To prevent forced magnetization of the reactor magnetic circuit by the constant component of the diode currents, six windings are used in it, connected according to the "correct zigzag" scheme.

Information on the technical implementation of the commutating link of the previous generation of compensation converters is intended to use the characteristics of their commutating voltage as a base value when proving the advantages of the new generation of converters.

II. Purpose of Work

To compare the energy capabilities of compensation converters of the previous and new generations and to recommend for practical application those that are capable of providing high energy characteristics and efficient use of equipment.

III. Results of Research

The authors devoted a number of scientific publications to the study of processes in new-generation compensation converters, in particular, to the three-phase bridge

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compensation converter. In the studies, it was assumed that a three-phase converter transformer (T) creates a standard symmetrical three-phase EMF system of direct phase alternation:

$$e_A = E_m \sin(\vartheta + \pi/6 + \psi);$$
$$e_B = E_m \sin(\vartheta - \pi/2 + \psi); \quad (1)$$
$$e_C = E_m \sin(\vartheta + 5\pi/6 + \psi),$$

where - E_m amplitude of the secondary phase voltage of the converter transformer; ϑ - current variable, ψ - initial voltage phase when the valve is turned on.

First, let us assume that the recharging of the commutating link capacitors is provided by a three-phase balancing reactor. Let us take into account that each valve of the compensation part of the converter (VD1, VD2 and VD3) comes into operation once per period and conducts current for a third of the period. Taking into account the above-described features of the operation of the TBR, in Fig. 2 the current curve of the phase of the capacitor battery C3 is constructed, which takes part in the commutating of the current from the valve VD3 to VD1.

As we can see, this current has three intervals in the period, each with a length of . The magnitude of the current at individual intervals: I – $i_c = -I_d/3$; II – $i_c = 0$; III – $i_c = I_d/3$.

Analytical calculation of the commutating voltage curve u_C by intervals is carried out using the form of the capacitor phase current according to the following relationship:

$$u_C = \frac{1}{\omega C}\int_0^{\vartheta} i_C d\vartheta + u_C(0). \quad (2)$$

The number of intervals of the commutating voltage curve is the same as the number of intervals of the current curve:

$$\text{I - } u_{C(1)} = \frac{1}{\omega C}\int_0^{\vartheta}(-\frac{I_d}{3})d\vartheta + u_C(0) = -\frac{I_d}{3\omega C}\vartheta + u_C(0).$$

$$\text{II - } u_{C(2)} = \frac{1}{\omega C}\int_0^{\vartheta} i_C d\vartheta + u_{C(1)k} = -\frac{2\pi I_d}{9\omega C} + u_C(0) = u_{C(2)k}.$$

$$\text{III - } u_{C(3)} = \frac{1}{\omega C}\int_0^{\vartheta}\frac{I_d}{3}d\vartheta + u_{C(2)k} = \frac{I_d}{3\omega C}\vartheta - \frac{2\pi I_d}{9\omega C} + u_C(0).$$

According to the calculation performed, in Fig. 2 the commutating voltage curve was constructed $u_C(\vartheta)$ and its initial value, which is also its maximum value, was calculated:

$$u_C(0) = \frac{I_d}{3\omega C}\cdot\frac{4\pi}{9} = \frac{2I_d}{27fC} \quad (3)$$

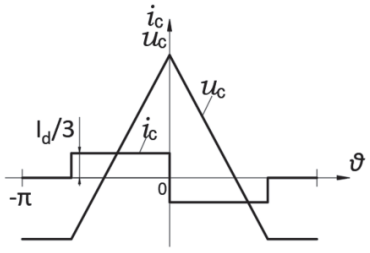


Fig.2. Commutating voltage and current curves of phase C3 of the capacitor battery

Analytical calculation of the commutating voltage curve by intervals is carried out using the form of the capacitor phase current according to the following relationship:

The disadvantage of the previous generation of compensation converters is the use of a three-phase balancing reactor in the switching link. Not only is it a bulky element with large energy losses. It is a passive element that is unable to regulate the process of recharging the capacitor battery.

The use of a new generation of fully controlled devices (IGBTs) in the switching link of compensation converters makes the process of recharging capacitors controllable. In other words, it becomes possible to regulate the commutating voltage both in form and in magnitude. The magnitude and form of the commutating voltage are influenced by factors such as the moment of inclusion in the operation of fully controlled devices of the commutating link, the frequency of control pulses and the adjustment range. The combination of these factors in different modes of operation of the converter is ultimately expressed in different energy capacities. Their study is a serious scientific task, since its solution will allow to formulate recommendations for ensuring effective operating modes of the studied converter, which corresponds to the purpose of the current publication.

When studying the influence of a set of factors on the formation of the commutating voltage, the following priority was adopted: the frequency of the control pulses of the commutating link transistors, the moment of switching on the transistors and the adjustment range. It was also taken into account that the magnitude of the commutating voltage depends on the capacitance of the phase CB of the commutating link and the magnitude of the load current, which, in turn, depends on the voltage of the secondary winding of the converter transformer and the load resistance of the converter. Also, in all studies, the moment of entry into operation of the diode VD1 of phase A when the current passes from the diode VD3 to the diode VD1 was taken as the starting point. For an objective comparison of the results of the studies in different operating modes of the converter, these values were taken to be the same, namely:

$U_{m2} = 57.8$ (*V*), $C = 0.000065$ (*F*), $R_d = 37.5$ (*Ohm*). (4)

If, based on the given data, we calculate for (3) the maximum value of the commutating voltage in the compensation converter of the previous generation (with TBR) - it turns out 50.14 (*V*). Here it is taken into account that for data (4) the rectified current of the converter is 2.2 (*A*).

Data (4) are also technical characteristics of the circuit model used in the current scientific research.

Let the transistors of the commutating link be controlled by pulses of a single frequency, i.e. a frequency corresponding to the frequency of the power supply network. It corresponds to a three-phase system of voltages of direct phase sequence. Accordingly, the transistors of the commutating link operate in the sequence VT1 → VT2 → VT3.

If the moment of switching on the transistor VT1 coincides with the moment of switching on the power electric diode VD1, the current through the capacitors of the commutating link does not pass, the commutating voltage is zero. That is, the converter will operate in the normal (non-compensated) mode. If the moment of applying the control pulse is applied to the transistors with some delay (θ), this will create conditions for the passage of current through the

phases of the CB. As a result of its charge in the electromagnetic process of the converter, an additional voltage will appear, which is called the commutating voltage. In this case, the converter switches to the compensated mode of operation.

Mode 1. Consider the first control range, in which the delay of control pulse supply to the transistors of the commutating link is within $0 \le \theta \le 2\pi/3$. The dynamics of changes in the magnitude and shape of the phase current of the capacitor battery is quite simple - the greater the delay of the control pulse supply (θ), the greater the length of the current through the phases of the CB. Just the length. The magnitude of the current remains unchanged. The current of the CB phase in the period has six intervals, which is presented in Fig. 3, *a*.

Replacing a three-phase balanced reactor with a three-phase group of fully controlled devices is a purely technical solution. Similar solutions are used in the creation of capacitive commutated converters both abroad [1, 3] and in our country [2, 5]. It would seem that the recharging of the capacitors of the commutating link should occur similarly to the previous generation converters with TBR. Comparing Figures 2 and 3, we see that this is not at all the case. The reason for this is that the TBR divides the rectified current of the converter into three equal parts and the current through the TBR phase flows constantly during the period. In other words, the instantaneous and average values of the current in the TBR phase are the same and equal to a third of the rectified current.

When controlling the transistors of the commutating link with single-frequency pulses, the situation with the CB overcharging is completely different. The entire rectified current of the converter passes through each IGBT during a third of the period. The other two thirds of the period there is no current through the transistor. Therefore, the average value of the transistor current is equal to a third of the rectified current of the converter, and the instantaneous one is not.

The commutating voltage is calculated by the relation (2). Like the current, its curve over the period has six intervals. Let $\theta=\pi/3$. Based on the algorithm of current commutating by power electric valves and the moment of commutating on the transistors of the commutating link, the length of each interval will be equal to $\pi/3$. And the pattern of change at individual intervals will be as follows:

$$\text{I - } u_{C(1)} = \frac{1}{\omega C}\int_0^{\vartheta}(-\frac{2I_d}{3})d\vartheta + u_C(0) = -\frac{2I_d}{3\omega C}\vartheta + u_C(0).$$

$$\text{II - } u_{C(2)} = \frac{1}{\omega C}\int_0^{\vartheta} i_C d\vartheta + u_{C(1)k} = -\frac{I_d}{9fC} + u_C(0) = u_{C(2)k}.$$

$$\text{III - } u_{C(3)} = \frac{1}{\omega C}\int_0^{\vartheta}(\frac{I_d}{3})d\vartheta + u_{C(2)k} = \frac{I_d}{3\omega C}\vartheta - \frac{I_d}{9fC} + u_C(0).$$

$$\text{IV - } u_{C(4)} = \frac{1}{\omega C}\int_0^{\vartheta} i_C d\vartheta + u_{C(3)k} = -\frac{I_d}{18fC} + u_C(0) = u_{C(4)k}.$$

$$\text{V - } u_{C(5)} = \frac{1}{\omega C}\int_0^{\vartheta}\frac{I_d}{3}d\vartheta + u_{C(4)k} = \frac{I_d}{3\omega C}\vartheta - \frac{I_d}{18fC} + u_C(0).$$

$$\text{VI - } u_{C(6)} = \frac{1}{\omega C}\int_0^{\vartheta} i_C d\vartheta + u_{C(5)k} = u_C(0).$$

Assuming that the average value of the curve $u_c(\vartheta)$ over the period is zero, we obtain:

$$u_C(0) = \frac{I_d}{18fC}\ . \tag{5}$$

The result corresponds to the case $\theta=\pi/3$. If we extend it to the entire regulation range in mode 1, we obtain:

$$u_C(\vartheta) = \frac{I_d}{3\omega C}\cdot\theta \tag{6}$$

Considering that the largest value of the switching voltage falls on $\theta = 2\pi/3$, the ratio for its calculation will be as follows:

$$U_{C\max} = I_d / 9fC\ . \tag{7}$$

Taking into account data (4) and the fact that with a larger value of the leading angle of regulation, the rectified voltage and rectified current will be somewhat smaller (in the case under consideration $I_d = 2.05(A)$), at the end of the regulation range the numerical value of the switching voltage is $U_{C\max} \approx 70(V)$. This value is 1.4 times greater than that obtained under the same conditions for the previous generation converter.

The regularity of the change in the switching voltage curve in mode 1 over intervals over the period of the power supply frequency is generalized, we write it as follows:

I $u_{CI}(\vartheta) = U_{mC}(1-\frac{2}{\theta}\vartheta),\quad 0 \le \vartheta \le \theta$;

II $u_{CII}(\vartheta) = -U_{mC},\quad \theta \le \vartheta \le 2\pi/3$;

III $u_{CIII}(\vartheta) = U_{mC}(\frac{1}{\theta}\vartheta - \frac{2\pi}{3\theta} - 1),\quad 2\pi/3 \le \vartheta \le (2\pi/3+\theta)$;

IV $u_{CIV}(\vartheta) = 0,\quad (2\pi/3+\theta) \le \vartheta \le 4\pi/3$;

V $u_{CV}(\vartheta) = U_{mC}(\frac{1}{\theta}\vartheta - \frac{4\pi}{3\theta}),\quad 4\pi/3 \le \vartheta \le (4\pi/3+\theta)$;

VI $u_{CVI}(\vartheta) = U_{mC},\quad (4\pi/3+\theta) \le \vartheta \le 2\pi$.

According to the above equations, Fig. 3, *b* illustrates the dynamics of changes in the magnitude and shape of the commutating voltage when the moment of entry into operation of the transistor VT1 changes in the range $0 \le \theta \le 2\pi/3$.

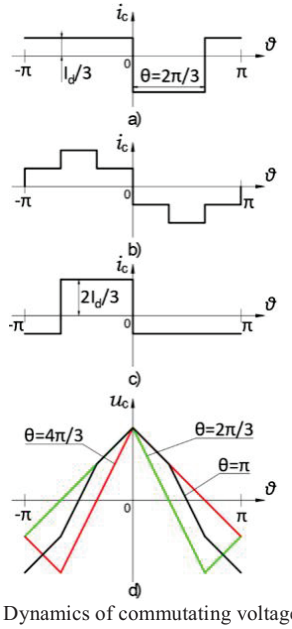


Fig.3. Dynamics of commutating voltage change in mode 1

The compensatory mode of operation of the converter is achieved by introducing an additional voltage into the commutating circuit, which is the commutating voltage. Superimposed on the linear voltage of the secondary winding of the converter transformer, it creates conditions for the leading current switching by power electric diodes [6]. The fact of superposition makes important not only the magnitude of the commutating voltage, but also its shape. Therefore, the conclusion regarding the 1.4 times greater magnitude of the commutating voltage should not be perceived as a great advantage of the operation of the new generation compensatory converter in mode 1. The advantage of the converter operation in the specified mode is precisely the possibility of smooth adjustment of the magnitude of the commutating voltage. But this value is limited by the transition of the converter to another operating mode.

Mode 2. Theoretically, the converter operation in mode 1 will continue until the delay of switching on the transistor VT1 reaches the value $\theta = 2\pi/3$, which is the boundary between modes 1 and 2. With an increase in the delay of switching on the transistor of the commutating link, the phase current of the capacitor battery will change in shape and will not change in magnitude. The period will have six intervals. The dynamics of the change in the shape of the current of the phase C3 CB of the commutating link at different moments of control pulse supply is shown in Fig. 4,*a*.

Let us calculate, using (2), the shape of the switching voltage at a delay $\theta = \pi$, or $\theta = \pi/3$, counting it from the beginning of the start of operation of the power electric valve VD2 of phase B of the converter:

$$\text{I } u_{C(1)} = \frac{1}{\omega C}\int_0^{\vartheta}(-\frac{I_d}{3})d\vartheta + u_C(0) = -\frac{I_d}{3\omega C}\vartheta + u_C(0).$$

$$\text{II } u_{C(2)} = \frac{1}{\omega C}\int_0^{\vartheta}(-\frac{2I_d}{3})d\vartheta + u_{C(1)k} = -\frac{2I_d}{3\omega C}\vartheta - \frac{I_d}{18fC} + u_C(0).$$

$$\text{III } u_{C(3)} = \frac{1}{\omega C}\int_0^{\vartheta}(-\frac{I_d}{3})d\vartheta + u_{C(2)k} = -\frac{I_d}{3\omega C}\vartheta - \frac{I_d}{6fC} + u_C(0).$$

$$\text{IV } u_{C(4)} = \frac{1}{\omega C}\int_0^{\vartheta}\frac{I_d}{3}d\vartheta + u_{C(3)k} = \frac{I_d}{3\omega C}\vartheta - \frac{2I_d}{9fC} + u_C(0).$$

$$\text{V } u_{C(5)} = \frac{1}{\omega C}\int_0^{\vartheta}\frac{2I_d}{3}d\vartheta + u_{C(4)k} = \frac{2I_d}{3\omega C}\vartheta - \frac{I_d}{6fC} + u_C(0).$$

$$\text{VI } u_{C(6)} = \frac{1}{\omega C}\int_0^{\vartheta}\frac{I_d}{3}d\vartheta + u_{C(5)k} = \frac{I_d}{3\omega C}\vartheta - \frac{I_d}{18fC} + u_C(0).$$

The initial value of the commutating voltage curve $u_C(0)$, which is also its maximum value, is calculated based on the fact that the average value of the curve $u_C(\vartheta)$ over the period is zero.

Having performed the calculation, we obtain:

$$u_C(0) = \mathrm{U}_{C\max} = I_d / 9fC. \qquad (8)$$

The general pattern of the change $u_C(\vartheta)$ in the frequency of the power supply network over the period is written as follows:

$$\text{I } u_{CI}(\vartheta) = U_{mC}(1 - \frac{3}{2\pi}\vartheta), \quad 0 \le \vartheta \le \theta;$$

$$\text{II } u_{CII}(\vartheta) = U_{mC}(1 + \theta_* - \frac{3}{\pi}\theta), \quad \theta \le \vartheta \le 2\pi/3;$$

$$\text{III } u_{CIII}(\vartheta) = U_{mC}(\theta_* - \frac{3}{2\pi}\vartheta), \quad 2\pi/3 \le \vartheta \le (2\pi/3 + \theta);$$

$$\text{IV } u_{CIV}(\vartheta) = U_{mC}(\frac{3}{2\pi}\vartheta - 2 - \theta_*), \quad (2\pi/3 + \theta) \le \vartheta \le 4\pi/3;$$

$$\text{V } u_{CV}(\vartheta) = U_{mC}(\frac{3}{\pi}\vartheta - 4 - \theta_*), \quad 4\pi/3 \le \vartheta \le (4\pi/3 + \theta);$$

$$\text{VI } u_{CVI}(\vartheta) = U_{mC}(\frac{3}{2\pi}\vartheta - 2), \quad (4\pi/3 + \theta) \le \vartheta \le 2\pi.$$

In the above expressions, the normalized value is $\theta_* = \theta/(2\pi/3)$. The dynamics of changes in the magnitude and shape of the switching voltage when the moment of entry into operation of the transistor VT1 in the range $2\pi/3 \le \theta \le 4\pi/3$ is illustrated in Fig. 4,*b*.

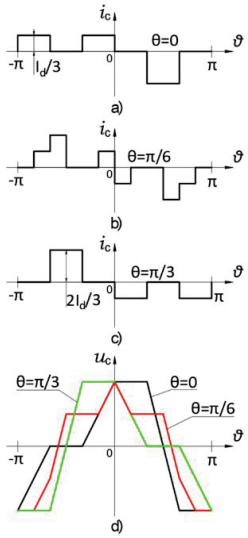


Fig.4. Dynamics of commutating voltage change in mode 2

The main feature of the converter operation in mode 2 is that throughout the entire regulation range the maximum value of the commutating voltage curve does not change. It is shape changes. Moreover, in such a way that it is possible to provide the maximum possible value of the switching voltage without restrictions caused by the operating mode. Let us illustrate the results of the circuit modeling in the Micro-Cap 12 package. The circuit model of the three-phase bridge compensation converter is shown in Fig. 5. The parameters of the model elements correspond to (4).

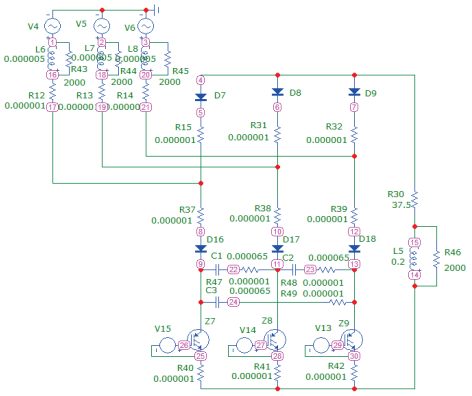


Fig.5. Object-oriented model of a three-phase bridge compensation converter

Formally, the operation of the converter when regulating the commutating link according to mode 2, with respect to the magnitude of the leading control angle, is the same, since the magnitude of the commutating voltage does not change. If we also take into account the possibility of an additional increase in the magnitude of the commutating voltage by reducing the capacitance of the phase of the capacitor battery, the situation does not look so unambiguous. According to this indicator, the beginning of the regulation process in mode 2 is the least productive. This can be seen from the modeling results shown in Fig. 6, *a*. Imposition the commutating voltage curve on the linear voltage curve of the secondary winding of the converter transformer leads to the transition of the converter to another operating mode, which in the theory of compensation converters is called the re-operation mode. Such an operating mode is undesirable due to the poorer energy performance and operating conditions of power equipment [7].

Increasing the delay moment improves θ the energy capabilities of the converter. This is confirmed by the results of circuit modeling shown in Fig. 6, *b* and *c*. They correspond to $\theta = \pi$ and show the possibility of obtaining the maximum possible amounts of reactive power generation, which are limited not by the insufficient value of the commutating voltage, but by the superposition of the current of the power electric valves of the cathode group on the valve of the anode group (Fig. 6, *c*). So far, this is the best indicator of the energy capabilities of the converter and it was obtained when the capacitance of the phase CB of the commutating link is half the base one given in (4), i.e. $0.0000325(F)$. Compared with the base mode, the efficiency of capacitors use is 2.

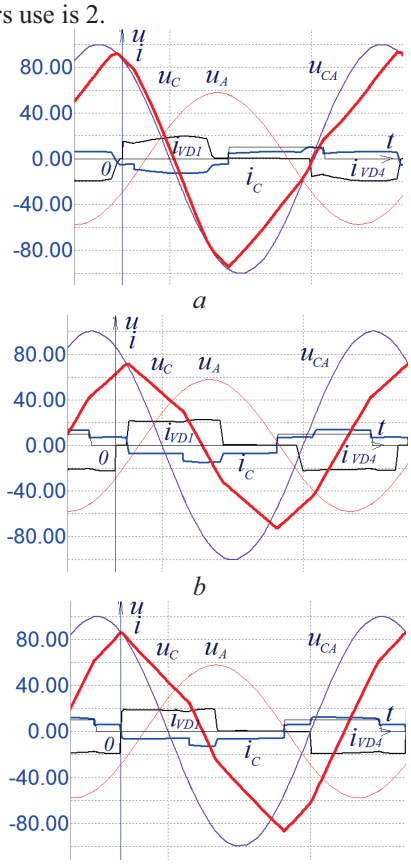


Fig.6. Results of simulation of the operation of the converter in mode 2

Mode 3. Let us increase the frequency of pulse supply to the transistors of the commutating link to double. Let $\theta = 0$, based on the algorithm of current commutating by power electric diodes and the moment of commutating on of the transistors of the commutating link, the phase current of the capacitor battery during the period has 6 intervals with different patterns of change. Accordingly, the commutating voltage during the period will have the same number of intervals. Its calculation according to the relation (2) gives the following result:

$$\text{I - } u_{C(1)} = \frac{1}{\omega C}\int_0^{\vartheta} i_C d\vartheta + u_C(0) = u_C(0) = u_{C(1)k}.$$

$$\text{II - } u_{C(2)} = \frac{1}{\omega C}\int_0^{\vartheta} (-\frac{2I_d}{3}) d\vartheta + u_{C(1)k} = -\frac{2I_d}{3\omega C}\vartheta + u_C(0).$$

$$\text{III - } u_{C(3)} = \frac{1}{\omega C}\int_0^{\vartheta} i_C d\vartheta + u_{C(2)k} = -\frac{I_d}{9fC} + u_C(0) = u_{C(3)k}.$$

$$\text{IV - } u_{C(4)} = \frac{1}{\omega C}\int_0^{\vartheta} (\frac{I_d}{3}) d\vartheta + u_{C(3)k} = \frac{I_d}{3\omega C}\vartheta - \frac{I_d}{9fC} + u_C(0).$$

$$\text{V - } u_{C(5)} = \frac{1}{\omega C}\int_0^{\vartheta} i_C d\vartheta + u_{C(4)k} = -\frac{I_d}{18fC} + u_C(0) = u_{C(5)k}.$$

$$\text{VI - } u_{C(6)} = \frac{1}{\omega C}\int_0^{\vartheta} \frac{I_d}{3} d\vartheta + u_{C(5)k} = \frac{I_d}{3\omega C}\vartheta - \frac{I_d}{18fC} + u_C(0).$$

The initial value of the commutating voltage curve $u_C(0)$, which is also its maximum value, is calculated based on the fact that the average value of the curve $u_C(\vartheta)$ over the period is zero. Having performed the calculation, we obtain:

$$u_C(0) = U_{C\max} = I_d / 18fC. \qquad (9)$$

Having performed a numerical calculation of the maximum value of the commutating voltage according to the relation (9) using the data (4), we obtain $U_{C\max} = 39.14(V)$, which is confirmed by Fig. 7, *a*.

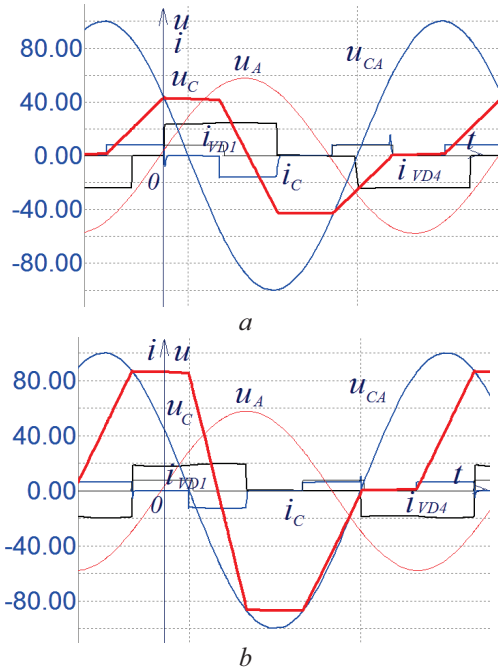


Fig.7.Results of simulation of converter operation in mode 3

Accordingly, Fig. 7, *b*, which is also the result of the circuit simulation of the studied mode, illustrates the possibility of the converter achieving the maximum commutating voltage. That is, achieving such an operating mode when the increase in the commutating voltage is limited by the superposition of the current of the power electric diodes of the cathode group on the diodes of the anode group. Such a mode was achieved in mode 2. But in mode 3, the efficiency of using the capacitors of the commutating link is higher, since this occurs at a capacitor phase capacitance of 0.000025 (*F*). Compared with the base mode, the efficiency of using the capacitors is 2.6.

Within the framework of one publication, it is not possible to present in detail the results of research on the specified topic at different frequencies of control pulses of the transistors of the commutating link. Therefore, we will limit ourselves to information on the results obtained from other studies.

At an eight-fold frequency of control pulses, it is possible to achieve the same performance as in mode 2. The efficiency of using the capacitors of the commutating link will be the same. Further increase in this frequency does not lead to an increase in the efficiency of the operating modes.

The reason for this is explained by Fig. 8. It presents the results of circuit modeling at a 32-fold frequency of control pulses of the commutating link transistors. As we can see, the switching voltage curve in shape and magnitude approaches that shown in Fig. 2. Recall that this voltage was such in the compensation converters of the previous generation, when the rectified current of the converter passed through all the phases of the TBR throughout the time.

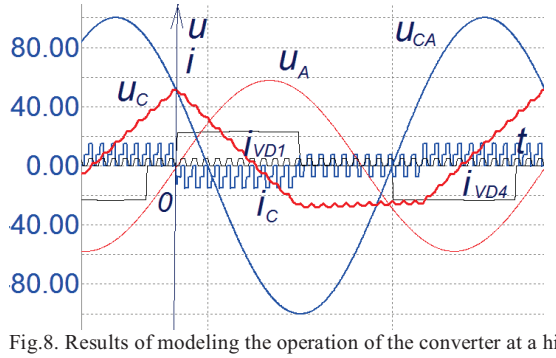


Fig.8. Results of modeling the operation of the converter at a high frequency of control pulses

Replacing the three-phase balancing reactor with a three-phase group of fully controlled devices is in itself a step forward in technical and economic terms, as discussed at the beginning of the publication. At the same time, the nature of the movement of the components of the rectified current through the transistors of the commutating link also changed. This process was most uneven in time at a single frequency of control pulses. The current passed through the transistor for a third of the period once per period. As the frequency of control pulses increased, the unevenness in the time of the current passing through the transistors of the commutating link was leveled out. And the shape of the commutating voltage approached that which was in the converters of the previous generation.

IV. CONCLUSIONS

1. The quantitative indicator of the possible volumes of reactive energy generation by the studied compensating converter is the value of the leading angle of regulation of the electric power valves of its compensative part. The value of the leading angle of regulation directly depends on the value of the commutating voltage. The paper shows how to organize the recharging of the commutating link capacitors in order to prevent the occurrence of situations in which the growth of the commutating voltage is blocked.

2. Based on the indicator of the possibility of achieving the maximum value of the leading angle of regulation of the power electric valves of the three-phase bridge compensative converter, and, accordingly, the generation of maximum volumes of reactive energy with the efficient use of the commutating link capacitors, the paper defines the control modes of the commutating link capacitors that are promising for practical application. It is proven that achieving high energy characteristics is based on the coordination of the following factors: the moment of switching on the fully controlled devices of the commutating link, the frequency of the control pulses of the commutating link transistors and the control range.

3. Object-oriented modeling of the operating modes of the studied electric energy converter, on the one hand, confirmed the results of theoretical studies, and on the other hand, was used as an independent research device. The modeling results showed the feasibility of limiting the frequency of control pulses to the transistors of the commutating link and showed how and to what extent it is possible to increase the efficiency of using a capacitor battery compared to the basic mode.

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Effect of coal tar composition and QI fractions on dielectric properties and electrostatic dewatering efficiency

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***Abstract*—Conventional coal tar dewatering methods—gravity settling, thermal treatments, and centrifugation—are inefficient due to high costs and poor performance in treating dense, viscous emulsions. Emerging multiphysics-assisted techniques (e.g., electric fields, ultrasonics) outperform traditional approaches by enhancing droplet coalescence and emulsion destabilization with lower energy use. However, coal tar's physicochemical complexity—compositional heterogeneity, variable dielectric behavior (ε=2.5–3.5), and non-Newtonian viscosity—impedes field interactions, destabilizes coalescence kinetics, and compromises system scalability. Electrostatic dewatering, for instance, is constrained by insufficient polarization contrast with water (ε ≈ 78), reducing droplet coalescence efficiency. Conductive quinoline-insoluble (QI) aggregates further disrupt electric fields, causing short circuits and localized discharges, while pyrolysis-derived tars elevate viscosity, hindering droplet mobility. Modeling shows dehydrator efficiency declines sharply with coal tar density increases (1165–1200 kg/m³), requiring coalesced droplets ≥200 µm for viable separation. To address these challenges, system designs must adapt to coal tar's dielectric variability, QI content, and rheology. Key operational parameters—temperature, electric field strength, and droplet dynamics—require optimization to balance coalescence rates and energy consumption. Successful implementation hinges on mitigating coal tar's inherent variability through tailored engineering solutions, ensuring robust dewatering performance despite fluctuating feedstock properties. This integrated approach could enable scalable, energy-efficient moisture removal, overcoming limitations of legacy methods in meeting industrial standards.**

Keywords—coal tar, dehydration, dielectric properties, quinoline insoluble substances.

I. Introduction

In the coke industry, conventional coal tar dewatering relies primarily on gravitational separation, wherein the material is retained in heated storage tanks at 80–90 °C for prolonged durations, often exceeding 36 hours. Secondary dewatering is then accomplished either via centrifugal separation (using industrial centrifuges) or thermal treatment in tubular furnaces. In the thermal process, coal tar is preheated to approximately 130 °C in the convection section before undergoing evaporation in a primary unit, where moisture content is reduced to below 0.5%.

These traditional methods exhibit several limitations, including high energy consumption, extensive equipment requirements, and significant spatial demands. Additionally, their efficiency is constrained by persistent challenges at the air–water–tar interface, where corrosive effects are pronounced. Separation of water from high-viscosity, dense coal tar proves particularly problematic, often failing to achieve the industrial specification of <4% mass moisture content.

In contrast, electrostatic dewatering enables rapid and efficient separation of dispersed water droplets from coal tar through the application of strong polarization and dielectrophoretic forces at the oil–water interface. This method substantially reduces energy consumption and equipment footprint compared to conventional thermal or centrifugal approaches, as it operates effectively at lower temperatures and without complex mechanical systems. Moreover, electric field treatment demonstrates superior performance for high-viscosity, dense coal tar emulsions, promoting droplet coalescence and sedimentation where gravity-based methods are inadequate.

An early innovation in electro-assisted separation is detailed in [1], which discloses a coal tar dewatering apparatus utilizing a high-voltage electric field. The device exploits the significant dielectric contrast between coal tar and water, inducing polarization of dispersed water droplets and enhancing their coalescence and settling. This approach offers distinct advantages, including reduced thermal energy input and improved separation efficiency—particularly for stable emulsions resistant to gravitational dewatering. The design emphasizes optimized electrode configuration and robust insulation, ensuring operational safety under high-voltage conditions.

Patent [2] describes a hybrid dehydrator designed for heavy oil and coal tar, integrating mechanical separation with electric field enhancement. The system employs a multi-stage

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process where gravitational settling is augmented by an applied electric field, significantly accelerating phase separation. Key design features include optimized flow paths to maximize residence time in the electrostatic zone and a modular configuration for scalable operation. This hybrid approach demonstrates superior performance compared to purely thermal methods, with enhanced adaptability to varying feedstock viscosities and moisture content.

A notable advancement in hybrid separation technology is presented in [3], which combines ultrasonic and electric field treatments for coal tar dewatering. The process begins with ultrasonic irradiation to destabilize emulsions by disrupting interfacial films and inducing droplet coalescence. Subsequent electrostatic treatment enhances the aggregation and separation of polarized water droplets. The synergistic effect of ultrasonic cavitation and dielectrophoresis markedly improves dewatering kinetics and reduces residual water content, even in high-viscosity or resin-rich tar fractions. This dual-field strategy is particularly effective for complex emulsions that resist conventional separation methods.

Patent [4] proposes a specialized method and apparatus for dewatering high-temperature coal tar, leveraging elevated temperatures (130–150 °C) to reduce viscosity and facilitate water separation. The system integrates preheated convection zones with evaporative separation chambers, employing controlled heating profiles and heat recovery mechanisms to optimize energy efficiency. While this thermal-mechanical approach consumes more energy than electrostatic methods, it maintains tar quality and is well-suited for industrial applications where waste heat from upstream processes can be utilized.

A clear trend across these patents is the shift toward multiphysics-assisted separation, where electric fields—often combined with ultrasonic energy—supplement or replace conventional thermal and gravitational methods. Electric fields enhance droplet coalescence by leveraging the dielectric contrast between water and tar, while ultrasonic irradiation promotes emulsion destabilization through cavitation-induced interfacial disruption. Compared to traditional techniques, these hybrid approaches demonstrate superior dewatering efficiency, reduced processing times, and lower energy consumption—particularly for viscous and highly emulsified coal tar fractions.

Despite promising laboratory-scale results and innovative conceptual designs, industrial adoption of electro-assisted and hybrid dewatering technologies remains limited. Key barriers include unresolved scientific and engineering challenges, primarily stemming from the complex and variable physicochemical properties of coal tar. These inherent material characteristics critically influence the performance and operational stability of electric and ultrasonic dewatering systems, posing significant hurdles for large-scale implementation.

The aim of this study was to evaluate the efficiency and throughput of a standard electrodehydrator for coal tar with different degrees of pyrolisation. The nature of substances insoluble in quinoline as a significant factor influencing the dehydration process was taken into account.

II. Materials and Methods

The morphological composition of solid tar inclusions (components insoluble in quinoline) and the distribution of water droplets in water-in-resin reverse emulsion for the studied resins were determined using a Neofot optical microscope equipped with a ToupTech attachment and ToupView software. The particle size distribution of the same insoluble substances was determined using a Jeol JSM 840 high-precision scanning electron microscope. Dielectric properties of the constituent morphological particles insoluble in quinoline were obtained from the data presented in publications [5-7]. The rheological characteristics of the studied liquids were determined by a Brookfield LV DV2T rotational viscometer (low viscosity digital viscometer) with a thermal cell

In the framework of this work, coal tar of different pyrolisation degrees, characterised by approximately the same water content (10-12 %), were analysed. The detailed characterisation of the investigated resins is presented in Table 1.

TABLE I. Characterisation Of Investigated Tars With Different Degrees Of Pyrolysis

Density 20°C, kg/m³	**Mass fraction, % of insoluble in**		**Elemental composition, %**				
	toluene	***quinoline***	***C***	***H***	***N***	***O***	***S***
1165	6.7	2.9	90.50	5.50	1.00	2.20	0.80
1175	8.4	4.2	89.50	5.61	0.43	4.11	0.35
1200	17.3	12.2	90.40	5.06	0.50	3.69	0.35

To assess the applicability of the technology of electrostatic dehydration of tar with different degrees of pyrolization, we used the calculation method of the 2EG-160-2 electrostatic dehydrator for the “oil-water” system.

The formula for calculating the performance of the electrohydrator is essentially a modification of the Stokes equation adapted for vertical deposition of water droplets in the oil flow inside a cylindrical apparatus [8]:

$$G=\frac{L\times D\times g\times d^2(\rho_{tar}-\rho_w)}{72\times\nu_{tar}\times\rho_{tar}} \qquad (1),$$

where: G - productivity of the apparatus on raw material, m^3/s; L - length of the apparatus, m; D - inner diameter of the apparatus; g - acceleration of free fall, m/s^2; d - diameter of the floating drop, m; ρ_{tar} - density of tar, kg/m^3; ρ_w - density of water, kg/m^3; ν_{tar} - kinematic viscosity of tar, m^2/s; 1/72 - empirical coefficient that takes into account the units of measurement and the transition from droplet deposition to mass productivity (m^3/s).

III. Result And Discussion

The morphological composition of quinoline-insoluble substances, isolated from coal tars with varying degrees of pyrolysis, along with the water droplet diameter in "water-in-tar" emulsions, are presented in Table 2. Representative photomicrographs of the hydrated tar are shown in Figures 1, while photomicrographs of the quinoline-insoluble substances are presented in Figure 2.

TABLE II. RESULTS OF MICROSCOPIC ANALYSIS OF EMULSION SAMPLES AND ISOLATED QUINOLINE-INSOLUBLE SUBSTANCES

Coal tar density 20°C, kg/m³	**Identified components of quinoline insolubles:**					**Diameter of water drop in emulsion, μ**	
	normal QI	***coal***	***coke***	***carbon black***	***pyrolytic carbon***	***Min.***	***Max.***
1165	+	+	+	+	-	30	220
1175	+	+	+	+	+	45	205
1200	+	+	-	+	+	23	185

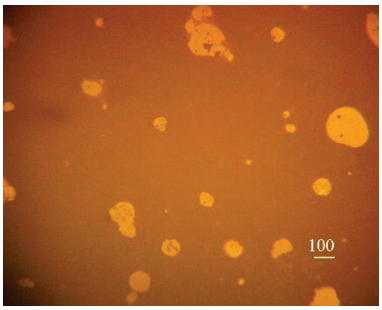


Fig. 1. Micrograph of water-in- tar emulsion.

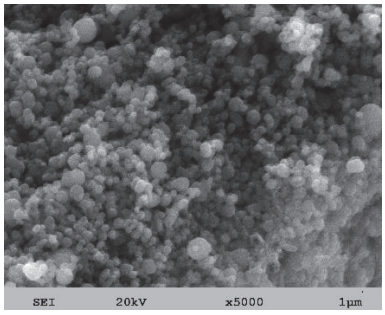


Fig. 2. Micrograph of quinoline insoluble substances isolated.

Coal tar is a complex multicomponent system including aromatic hydrocarbons (naphthalene, anthracene, phenanthrene), phenols, heterocyclic compounds and resinous components. The bulk of the compounds are characterized by a high degree of π-conjugation but relatively low polarity. Dielectric constant (ε) values depend on the specific composition, but for typical coal tar samples are in the range of 2.5-3.0, which is higher than that of saturated hydrocarbons but lower than that of polar aromatic compounds containing heteroatoms.

In electrodehydration processes, the efficiency of water droplet coalescence depends significantly on the dielectric permittivity contrast between the dispersed (water, ε ≈ 78 at 25 °C) and dispersed (oil or resin) phases. The higher this difference, the higher the induced dipole force in water droplets under the action of the electric field and, consequently, the higher the probability of their fusion. Thus, the coalescence conditions are more favorable for oil, which has minimal ε, than for the more polar coal tar.

Additionally, with increasing molecular weight and degree of condensation of aromatic structures, the dielectric constant of the resin tends to increase, especially in the presence of heteroatoms (O, N, S) involved in dipole-dipole interactions (Table 3).

TABLE III. VALUES OF DIELECTRIC CONSTANTS FOR INDIVIDUAL COMPONENTS OF COAL TAR AND SUBSTANCES INSOLUBLE IN QUINOLINE

Component	Dielectric constant
Coal tar	2.0÷3.0
Benzene	2.3
Toluene	2.4
Naphthalene	2.54
Phenanthrene	2.8
Anthracene	3.08
Quinoline	9
Phenol	9.9
Pyridine	12.5
Coal dust	1,2-1,8
Coke	1,1-2,2
Carbon black	2.5–3.0
Pyrolytic Carbon	4,0-6,0
Pyrite	7
Coal ash	4.6

Tars with higher density and degree of heat treatment (pyrolization) are characterized by higher aromaticity, which is accompanied by an increase in dielectric permittivity (ε) to values of the order of 3.0 and higher. This leads to a decrease in the efficiency of electrostatic dehydration, since the increased ε worsens the conditions of polarization and coalescence of water droplets in the electric field.

The increase in the content of high molecular weight heteroatom-containing aromatic compounds, as well as thermal modification of the tar contribute to a further increase in dielectric constant. Collectively, these changes impede electrodehydration, reducing the efficiency of phase separation processes.

The presence of fractions insoluble in quinoline (QI) also has a significant influence on the electrical characteristics of the tar. These particles have a high degree of aromaticity, low polarity and limited molecular mobility, despite the possible presence of heteroatoms. Due to their low polarizability, the QI fractions have a relatively low dielectric constant and therefore make a limited contribution to the overall ε value of the tar. Moreover, large or aggregated QI particles are able to screen the electric field, locally weakening it.

Some QI particles contain graphite-like carbon fragments or metal inclusions possessing electrical conductivity. In case of aggregation of such particles with formation of percolation pathways, it is possible to form continuous conductive channels. These structures can cross electric field lines or even create a short circuit between electrodes, forming a path with reduced resistance. As a result, the electric field strength in the working volume is reduced, the water is not polarized and does not move, which leads to a sharp deterioration of the dehydration efficiency. Additionally, the risk of localized electrical discharges and arcs that can cause equipment damage increases.

When the values of ε′ > 4 are exceeded in combination with increased electrical conductivity, the electric field starts to be lost in the form of thermal losses, which is critical for polarization processes of water droplets. The specific conductivity of such systems [9] varies widely and depends on the degree of aggregation and the conductive structure of QI. Near QI aggregations, microdischarge generation is possible, especially at high voltage.

Thus, the dielectric constant of coal tar containing QI fractions is significantly higher than that of purified tar (usually 2.5-3.5), which should be taken into account when designing electrostatic dewatering systems.

The results of calculation of the electric dehydrator performance depending on the density of tar oils and coal tar itself at removal of water droplets with a diameter of 220 microns are presented in Figure 3.

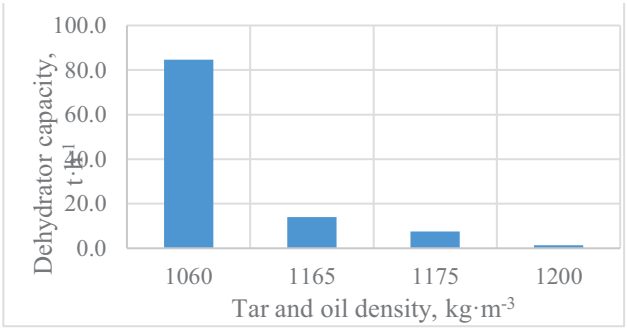


Fig. 3. Performance of dehydrator at application of emulsions with different density of constant phase (1060- density of resin fraction boiling out in the range of 230-300°C); 1165-1200- density of resin of different degree of pyrolization.

Consequently, the degree of pyrolyzation of coal tar significantly affects the performance of the electrodehydrator, as the increasing density and viscosity of the tar itself impede the rapid ascent of enlarged droplets formed by the coalescence of smaller droplets. Figure 4 shows the results of calculation of dehydrator productivity depending on the degree of pyrolisation of tar and the size of dispersed water droplets.

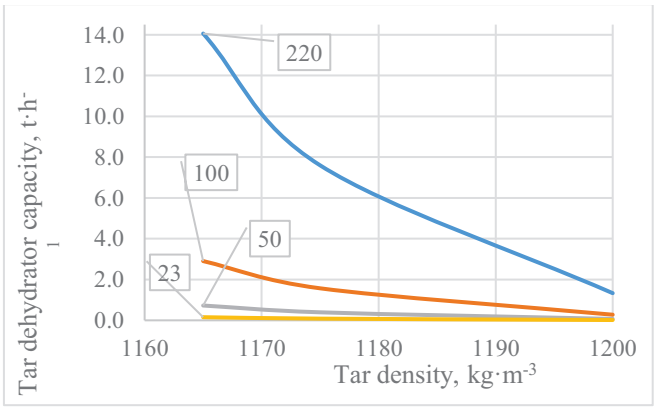


Fig. 4. Dependence of dehydrator productivity on the density of dewatered resin at different diameters of pop-up water droplets.

It follows from the Figure 4 that in order to increase the performance of the electric dehydrator for dewatering of heavy resins it is necessary that pop-up water droplets coalesced to form a united droplet with a diameter of at least 200 microns.

Experimental industrial trials of electric dehydrators for coal tar (density 1190 kg/m³) demonstrated that applying a voltage of 9.5–11 kV across 1×4 m² electrodes spaced 80–100 mm apart—achieving an electric field strength of 110–137 kV/m at 70–80°C—effectively reduces water content to below 3%. These results confirm the viability of electrostatic dewatering as a scalable, energy-efficient solution for meeting stringent industrial moisture specifications, surpassing traditional methods in both performance and operational efficiency.

IV. Conclusions

The dielectric properties and compositional complexity of coal tar—particularly its aromaticity, heteroatom content, and quinoline-insoluble (QI) fractions—critically influence the efficacy of electrostatic dewatering. Coal tar's dielectric constant (2.5–3.5), though higher than crude oil's, creates suboptimal polarization contrast with water (ε ≈ 78), reducing droplet coalescence efficiency compared to less polar oils. Furthermore, QI fractions with low polarity and conductive aggregates can shield or disrupt electric fields, risking short circuits and localized discharges, while elevated resin density and viscosity from pyrolysis hinder droplet separation. Calculations demonstrate that dehydrator performance declines sharply with increasing tar density (e.g., 1165–1200 kg/m³), necessitating coalesced water droplets ≥200 μm to maintain efficiency. Successful implementation of electrostatic dewatering thus requires tailored system designs to account for coal tar's variable dielectric behavior, QI content, and rheological properties, alongside optimized operational parameters like temperature, field strength, and droplet dynamics.

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Differential Evolution for Optimized Task Clustering and Scheduling in IoT Edge Computing

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***Abstract* — the rapid proliferation of Internet of Things (IoT) devices has led to a significant increase in data generation, necessitating efficient edge computing solutions to minimize latency and enhance real-time processing. This paper presents a novel approach to task clustering and scheduling in IoT edge computing environments, utilizing a Differential Evolution (DE) algorithm. The proposed mechanism optimizes the allocation of computational tasks to available edge resources, reducing execution times and improving overall system efficiency. Key performance metrics demonstrate the effectiveness of the DE-based method in handling the dynamic and heterogeneous nature of IoT tasks, offering a scalable and robust solution for real-time IoT applications.**

Keywords* — *Internet of Things (IoT), Edge Computing, Resource Allocation, Task Optimization, Task Clustering, Task Scheduling, Real-time Processing, Differential Evolution

I. Introduction

The rapid expansion of the Internet of Things (IoT) has led to an unprecedented increase in the number of interconnected devices and sensors. These devices generate massive amounts of data, which traditionally have been processed in centralized cloud servers [1–3]. However, this approach introduces significant latency issues due to the physical distance between data sources and data processing centers. As real-time data processing becomes increasingly critical for applications such as autonomous vehicles, smart cities, and healthcare monitoring, the limitations of cloud computing have necessitated the development of more efficient data management solutions [4, 5]. IoT edge computing has emerged as a viable alternative, offering a decentralized approach where data is processed closer to its point of origin, thereby reducing latency and improving response times [6, 7].

The primary goal of this research is to enhance the efficiency of IoT edge computing systems by developing an innovative mechanism for task clustering and scheduling. By leveraging differential evolution optimization, the proposed mechanism aims to optimize the allocation of computational tasks to available resources at the edge. This research seeks to address the inherent challenges associated with IoT edge task management, such as the diverse nature of tasks, the limited computational capabilities of edge devices, and the dynamic and real-time requirements of IoT applications. To achieve this goal, the research focuses on several key tasks. First, a clustering mechanism will be developed to group IoT tasks based on their characteristics, such as task length, file size, number of required CPUs, and priority. This clustering process aims to ensure that tasks with similar requirements are processed together, enhancing resource utilization and minimizing execution time. Next, an efficient scheduling algorithm will be created to allocate the clustered tasks to appropriate edge computing resources. The scheduling mechanism will consider the current workload of resources to achieve a balanced distribution, thereby optimizing overall system performance.

Additionally, the differential evolution (DE) algorithm will be implemented and adapted for the specific context of IoT edge computing. The DE algorithm will be used to iteratively improve the clustering and scheduling process, ensuring that the solution converges to an optimal or near-optimal state. Key performance metrics, including execution time, system efficiency, and stability, will be measured to validate the effectiveness of the proposed mechanism. The goal is to demonstrate significant improvements in task execution times and overall system responsiveness compared to existing methods.

By achieving these tasks, this research aims to provide a scalable and efficient solution for task clustering and scheduling in IoT edge computing environments, ultimately enhancing the performance and reliability of real-time IoT applications.

II. Related Works

This literature review examines the optimization of task clustering and scheduling in IoT edge computing, a field that has garnered significant attention due to the increasing demand for efficient and low-latency data processing. Various strategies have been proposed to address the challenges inherent in managing diverse IoT tasks and resources. Several key studies have explored different heuristic and metaheuristic approaches to task clustering and scheduling.

One prominent method is the Particle Swarm Optimization (PSO) algorithm. Research conducted by Alhaizaey et al. (2022) demonstrated the effectiveness of PSO in distributing edge device tasks, optimizing makespan time, and reducing task allocation overhead. However, PSO's limitations in dynamic and heterogeneous IoT environments necessitate further refinement for broader applicability [8].

Another notable approach is the Firefly Algorithm (FA), which has been utilized by Yousif et al. (2021) to optimize cost and energy consumption in IoT edge computing

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systems. While FA shows promise in specific scenarios, its performance in handling the dynamic nature of IoT tasks and resources requires improvement to achieve consistent efficiency across various workloads [9].

The integration of evolutionary algorithms in task clustering and scheduling has also been explored. The study by Azimi et al. (2020) introduced a multi-cluster edge layer framework, leveraging evolutionary mechanisms to minimize latency and delays in IoT edge tasks. This approach extended the capabilities of computational clouds to the IoT edge, showcasing the potential of evolutionary algorithms in managing semi-autonomous edge clusters [10].

Recent research has also considered the application of differential evolution (DE) algorithms in optimization problems. The work by Wang et al. (2022) applied a DE mechanism for energy-efficient task scheduling, addressing the discrete optimization requirements of IoT task scheduling. However, the study did not fully explore the multidimensional task characteristics specific to IoT edge computing environments [11].

Despite these advancements, existing mechanisms often fall short in addressing the unique constraints and real-time issues present in IoT edge task scheduling [12]. The proposed differential evolution (DE) mechanism by Kavitha et al. (2020) aims to fill this gap by providing a robust and scalable solution for task clustering and scheduling in IoT edge computing [13].

III. The Aim and Objectives of the Study

The primary goal of this research is to enhance the efficiency of IoT edge computing systems by developing an innovative mechanism for task clustering and scheduling using differential evolution optimization. This research seeks to develop a clustering mechanism to group IoT tasks based on characteristics such as task length, file size, number of required CPUs, and priority, ensuring tasks with similar requirements are processed together. Additionally, it aims to create an efficient scheduling algorithm to allocate the clustered tasks to appropriate edge computing resources, considering current workload to achieve balanced distribution and optimize system performance. The research will implement and adapt the differential evolution (DE) algorithm for the context of IoT edge computing to iteratively improve clustering and scheduling processes, aiming for optimal or near-optimal solutions. Key performance metrics, including execution time, system efficiency, and stability, will be measured to validate the effectiveness of the proposed mechanism compared to existing methods like FA and PSO. By achieving these tasks, this research seeks to provide a scalable and efficient solution for task clustering and scheduling in IoT edge computing environments, ultimately enhancing the performance and reliability of real-time IoT applications.

IV. Proposed Method

To address the challenges of task clustering and scheduling in IoT edge computing, this research proposes a novel mechanism leveraging differential evolution (DE) optimization. The proposed method is designed to optimize the allocation of computational tasks to available edge resources, thereby reducing execution times and improving overall system efficiency. This section outlines the various components of the proposed method, including the task clustering mechanism, the scheduling algorithm, the differential evolution optimization, and the mathematical representation.

A. Task Clustering Mechanism

The first step in the proposed method involves clustering IoT tasks based on their characteristics. The clustering mechanism groups tasks by considering factors such as task length, file size, number of required CPUs, and priority. This ensures that tasks with similar requirements are processed together, facilitating more efficient resource utilization. The similarity function $S(T_m, T_n)$ is used to measure the degree of similarity between tasks T_m and T_n. The function takes into account various task properties, and tasks with high similarity values are grouped into the same cluster.

The clustering process begins with the collection of task characteristics. Each IoT task is characterized by a set of attributes that describe its resource requirements and execution parameters. These attributes include task length, which represents the duration required to complete the task; file size, indicating the amount of data associated with the task; the number of CPUs required, which reflects the computational intensity; and task priority, which denotes the importance or urgency of the task. By analyzing these attributes, the similarity function can determine the degree of similarity between different tasks. The similarity function $S(T_m, T_n)$ is defined mathematically to quantify the similarity between two tasks. It calculates the similarity score based on the differences in their attributes. For instance, the similarity in task length is computed as:

$$S_{task_length}(T_m, T_n) = \\ = 1 \Big/ \left(1 + \left(\text{task_length}(T_m) - \text{task_length}(T_m)\right)\right), \quad (1)$$

where T_m and T_n are tasks for which the similarity function values are computed; $\text{task_length}(\bullet)$ is a function that determines the length of a specific task.

Similar equations are used to calculate the similarity scores for other attributes. The overall similarity score $S(T_m, T_n)$ is then obtained by averaging the individual similarity scores. Tasks with high overall similarity scores are grouped into the same cluster. Once the similarity scores are computed, the tasks are grouped into clusters using a clustering algorithm [14, 15]. This algorithm iteratively assigns tasks to clusters based on their similarity scores, ensuring that each cluster contains tasks with similar characteristics. The clustering process aims to minimize the intra-cluster variability while maximizing the inter-cluster variability. This approach not only enhances resource utilization but also simplifies the scheduling process by reducing the variability within each cluster.

B. Task Scheduling Algorithm

Once the tasks are clustered, the next step is to allocate these clusters to suitable edge computing resources. The proposed scheduling algorithm aims to balance the workload across resources, minimizing execution times and optimizing system performance. The scheduling process involves mapping each task cluster to an edge resource based on the current workload and resource capabilities. The algorithm ensures that high-workload tasks are allocated to

less busy resources and vice versa, thus achieving a balanced distribution.

The scheduling algorithm operates in several phases. Initially, it assesses the current workload of each edge resource, considering factors such as CPU utilization, memory usage, and network bandwidth. This assessment helps in understanding the current state of the resources and their availability for additional tasks. The algorithm then assigns task clusters to resources in a manner that balances the load, avoiding overburdening any single resource.

The allocation of task clusters to resources is performed using a heuristic approach that takes into account the current workload and the resource capabilities [16]. The heuristic aims to minimize the overall execution time by distributing the workload evenly across all available resources. The algorithm starts by sorting the task clusters based on their total workload. It then iteratively assigns the task clusters to the least loaded resource, updating the resource load after each assignment. This dynamic allocation process is continuously refined to adapt to changing workloads and resource conditions, ensuring optimal performance at all times.

The scheduling algorithm also incorporates a feedback mechanism to monitor the performance of the allocated tasks. This feedback is used to adjust the allocation strategy in real-time, addressing any imbalances that may arise during execution. By continuously monitoring and adjusting the allocation, the algorithm ensures that the resources are utilized efficiently and that the execution times are minimized.

C. Differential Evolution (DE) Optimization

The core of the proposed method is the differential evolution optimization algorithm. DE is an evolutionary algorithm that iteratively improves candidate solutions through mutation, crossover, and selection processes. In the context of IoT edge computing, the DE algorithm is adapted to handle the discrete nature of task clustering and scheduling. The DE optimization process begins with the initialization of a random population of candidate solutions. Each solution represents a potential allocation of tasks to clusters and resources. The fitness of each solution is evaluated based on the total execution time of the tasks. The algorithm then generates new candidate solutions by combining existing ones using mutation and crossover operations. The best-performing solutions are selected for the next generation, and this process is repeated until the optimization converges to an optimal or near-optimal solution.

The mutation operation in DE involves creating a mutant vector by adding the weighted difference between two population vectors to a third vector. This operation introduces variation into the population, allowing the algorithm to explore a wide range of potential solutions. The crossover operation combines elements from the mutant vector and the target vector to create a trial vector. This operation ensures that the new candidate solutions inherit characteristics from multiple parents, increasing the diversity of the population.

The selection operation evaluates the fitness of the trial vector and compares it with the fitness of the target vector. If the trial vector has a better fitness value, it replaces the target vector in the population. This selection process ensures that the population gradually evolves towards better solutions over successive generations.

In the context of IoT edge computing, the DE algorithm is adapted to handle the discrete nature of task clustering and scheduling. The candidate solutions are represented as discrete vectors, where each element indicates the assignment of a task to a cluster and a resource. The fitness function evaluates the total execution time of the tasks, considering the workload distribution and resource capabilities. By iteratively refining the candidate solutions, the DE algorithm identifies high-quality solutions that minimize execution times and maximize resource efficiency.

D. Mathematical Representation

The task clustering and scheduling problem is represented mathematically to facilitate the application of the DE algorithm. Let $T = \left\{T_i, j = \overline{1,I}\right\}$ be the set of IoT tasks, $C = \left\{C_j, j = \overline{1,J}\right\}$ be the set of task clusters, and $R = \left\{R_k, k = \overline{1,K}\right\}$ be the set of edge resources. The variable x_{ijk} represents the assignment of task T_i to cluster C_j and resource R_k, where $x_{ijk} = 1$ if the task is assigned and 0 otherwise. The objective function to minimize the total execution time Z is given by:

$$Z = \sum_{i=1}^{I}\sum_{j=1}^{J}\sum_{k=1}^{K} x_{ijk} \cdot e_{ijk} \underset{t}{\rightarrow} \min, \tag{1}$$

where e_{ijk} is the execution time of task T_i in cluster C_j on resource R_k, t – criterion for minimization based on the total execution time of tasks. The constraints ensure that each task is assigned to exactly one cluster and one resource, and that the total number of tasks assigned to a resource does not exceed its capacity.

The mathematical formulation provides a clear and concise representation of the optimization problem, enabling the DE algorithm to operate effectively. By defining the objective function and constraints, it ensures that the algorithm focuses on minimizing execution times while adhering to resource limitations. This structured approach is essential for achieving optimal task allocation in complex IoT edge computing environments.

The constraints are defined as follows:

- Each task is assigned to exactly one cluster and one resource:

$$\sum_{j=1}^{J}\sum_{k=1}^{K} x_{ijk} \le 1 \quad \forall i \in \overline{1,I}, \tag{2}$$

- The total number of tasks assigned to a resource should not exceed its capacity:

$$\sum_{i=1}^{I}\sum_{j=1}^{J} x_{ijk} \cdot r_{ijk} \le n_k \quad \forall k \in \overline{1,K}, \tag{3}$$

where r_{ijk} – amount of resource R_k, required for executing task T_i, assigned to cluster C_j, n_k – maximum available amount of resource R_k.

By incorporating these constraints into the optimization problem, the DE algorithm can effectively allocate tasks to clusters and resources while ensuring that the resource capacities are not exceeded [17]. The mathematical representation provides a robust framework for the optimization process, guiding the algorithm towards feasible and optimal solutions.

V. Results and Discussion

To assess the effectiveness of the proposed method, simulations were carried out in a controlled Industrial IoT (IIoT) environment. The performance was benchmarked against two established metaheuristic algorithms: Particle Swarm Optimization (PSO) and the Firefly Algorithm (FA), based on a consistent set of evaluation metrics. The results highlight the advantages of the proposed approach in terms of execution time, resource utilization, network overhead, and load balancing.

A. Environment

Simulations were conducted using a custom model of an IIoT system comprising three edge computing nodes with uniform capabilities. Each node was equipped with 8 logical CPU units and connected via a local 1 Gbps network with a round-trip latency of approximately 50 milliseconds. A total of 100 tasks were generated per experiment. Each task was defined by a random execution length, data size, required CPU resources, and priority level. Tasks were initially assigned to original nodes and could be executed locally or offloaded, with network latency added in the latter case. To ensure statistical robustness, each scheduling method was evaluated over 250 independent iterations. Final results were obtained by averaging the values, providing a representative assessment of each method's effectiveness.

B. Evaluation Metrics

The effectiveness of the scheduling methods was evaluated using four key performance metrics relevant to the requirements of real-time edge computing in IIoT environments:

- Total execution time (makespan) — measures the time required to complete the full batch of tasks from start to finish. Minimizing execution time is critical in IIoT systems where latency directly affects responsiveness, process control, and system throughput;
- Average CPU utilization — represents the ratio of active processing time to the total available CPU time across all edge nodes. Higher utilization indicates more efficient use of computing resources, reducing idle time and increasing overall system productivity;
- Average network latency — accounts for the communication delay introduced when tasks are offloaded from their origin node to another processing node. This metric reflects the trade-off between computational balancing and the cost of data transfer across the network;
- Load balancing efficiency — quantifies how evenly the workload is distributed across all available edge nodes. Better balance reduces the risk of overloading individual nodes and helps maintain consistent performance under varying workloads.

These metrics were selected to reflect the core challenges of IIoT edge computing: ensuring low-latency task completion, maximizing the utility of limited edge resources, and minimizing inter-node communication overhead. A high-performing scheduling algorithm must strike a balance across all four criteria to be viable in dynamic, real-world deployments.

C. Results

The experimental comparison highlights clear advantages of the proposed method over PSO and FA. By effectively distributing the computational load and minimizing unnecessary task migrations, the proposed method achieves the shortest execution time, the highest CPU utilization, and the most balanced workload allocation. These characteristics are especially important for IIoT systems, where responsiveness and efficiency under constrained resources are critical. The comparison of execution times, presented in Fig. 1, reveals clear differences in scheduling efficiency among the evaluated methods. The proposed approach achieves the shortest overall makespan, demonstrating its ability to complete the entire workload more rapidly than PSO and FA. This outcome highlights the effectiveness of the method's task allocation strategy, which reduces idle time and maximizes concurrency. In contrast, FA results in the longest execution time due to its limited offloading behavior, which prevents efficient workload distribution across edge nodes.

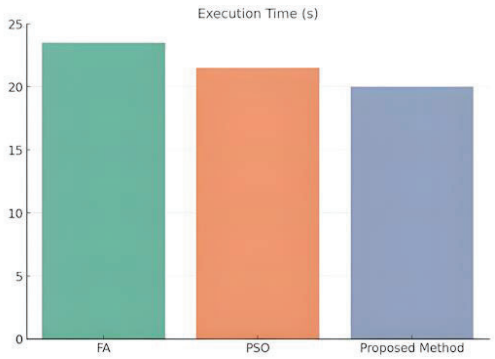


Fig. 1. Comparison of the scheduling methods in terms of execution time

Fig. 2 illustrates the average CPU utilization observed for each method during task execution. The proposed method maintains the highest utilization across the edge infrastructure, indicating more efficient exploitation of available processing resources. This suggests that the method not only schedules tasks promptly but also minimizes underutilization of computational capacity. PSO performs reasonably well but still exhibits periods of partial idleness. FA, which restricts task migration, results in the lowest CPU utilization due to uneven task accumulation and imbalanced processing. The average network latency introduced by task offloading is analyzed in Fig. 3. As expected, FA produces the lowest communication overhead by prioritizing local task execution and avoiding cross-node transfers. However, this conservative behavior comes at the cost of lower system-level efficiency. PSO, on the other hand, incurs the highest latency due to its aggressive offloading strategy. The proposed method demonstrates a balanced behavior, maintaining moderate latency while achieving superior performance in other critical metrics, which reflects an effective trade-off between load distribution and communication cost.

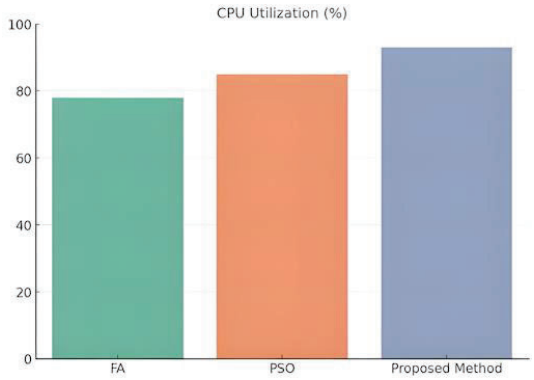


Fig. 2. Comparison of the scheduling methods in terms of CPU utilization

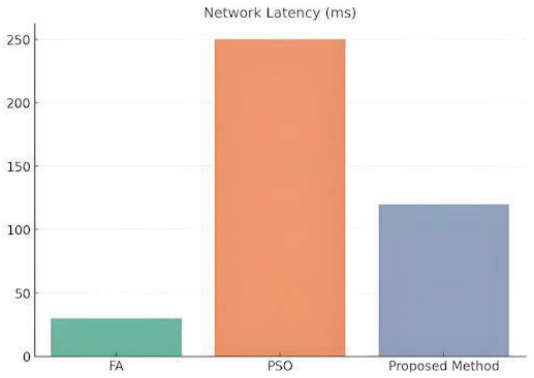


Fig. 3. Comparison of the scheduling methods in terms of network latency

Fig. 4 presents the results related to load balancing efficiency across the edge nodes. The proposed method attains the most uniform distribution of tasks, minimizing disparities in processing workload and ensuring that all nodes contribute evenly to task execution. This balance reduces the likelihood of node saturation or underuse, enhancing the overall responsiveness and scalability of the system. PSO achieves partial load balancing but remains suboptimal, while FA demonstrates the weakest performance in this metric due to its preference for local-only execution.

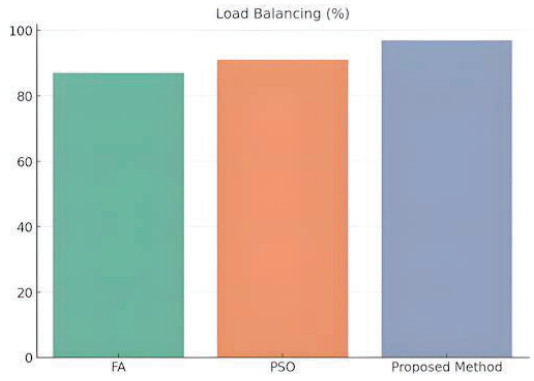


Fig. 4. Comparison of the scheduling methods in terms of load balancing

The numerical results presented in Table 1 provide a consolidated view of the performance achieved by each scheduling method across the defined evaluation metrics. The proposed method consistently outperforms both PSO and FA, achieving the shortest execution time, the highest CPU utilization, and the most balanced workload distribution, while maintaining moderate network latency.

TABLE I. Comparison of the scheduling methods across the metrics

Iteration No.	FA	PSO	Proposed Method
Execution Time (s)			
5	23.15	21.57	20.06
10	23.49	21.55	19.62
50	23.33	21.67	20.01
100	23.31	21.61	20.02
150	23.77	21.56	19.62
200	23.71	21.68	19.63
250	23.98	21.27	20.3
Avg.	**23.5**	**21.5**	**20**
CPU Utilization (%)			
5	76.65	84.23	94.22
10	78.12	84.95	97.93
50	78.96	84.04	93.67
100	76.8	82.33	93.64
150	77.68	95.23	94.34
200	75.01	83.14	91.7
250	78.14	86.2	92.58
Avg.	**78**	**85**	**93**
Network Latency (ms)			
5	34.66	260.31	111.61
10	31.51	248.08	119.65
50	27.45	260.38	117.3
100	32.2	234.07	119.8
150	26.24	252.13	116.81
200	22.75	249.17	110.78
250	30.39	240.48	122.58
Avg.	**30**	**250**	**120**
Load Balancing (%)			
5	87.33	90.69	97.98
10	88.14	89.83	97.95
50	86.56	90.02	97.38
100	87.22	91.55	98.36
150	88.08	90.2	97.02
200	87.21	90	97.07
250	87.33	89.76	96.84
Avg.	**87**	**91**	**97**

These results demonstrate the method's ability to optimize task allocation effectively without incurring excessive communication overhead. While FA achieves minimal network delay, it performs poorly in all other aspects, particularly execution time and resource utilization. PSO delivers more balanced performance but is limited by increased offloading cost and suboptimal workload distribution. Overall, the table confirms that the proposed approach provides the most favorable trade-off across all key performance indicators in the IIoT scheduling context.

D. Discussion

The results confirm the ability of the proposed method to effectively cluster and schedule tasks in heterogeneous IIoT edge environments. The method maintains a strong balance between processing efficiency and communication cost. It improves makespan and load distribution while limiting unnecessary offloading. These properties make it suitable for deployment in resource-constrained real-time edge systems where responsiveness and scalability are critical.

VI. Conclusion

The research presented in this study has demonstrated the efficacy of a differential evolution (DE) mechanism for optimizing task clustering and scheduling in IoT edge computing environments. The proposed DE-based method

significantly improves execution times and resource utilization compared to established optimization techniques such as the Firefly Algorithm (FA) and Particle Swarm Optimization (PSO). This conclusion is based on the superior performance of the DE mechanism in handling the dynamic and heterogeneous nature of IoT tasks and resources.

A key contribution of this research is the development of a task clustering mechanism that groups tasks based on their characteristics, such as task length, file size, number of required CPUs, and priority. This clustering approach ensures that tasks with similar requirements are processed together, which enhances resource efficiency and reduces execution times. The scheduling algorithm, which dynamically allocates these clustered tasks to the most suitable edge resources, further optimizes system performance by balancing the workload in real-time.

The DE optimization process, characterized by its iterative improvement through mutation, crossover, and selection operations, enables the proposed method to effectively explore and exploit the solution space. This robust optimization framework allows the DE mechanism to consistently identify optimal or near-optimal solutions, even in complex and rapidly changing IoT environments. The integration of a feedback mechanism that continuously monitors and adjusts task allocations ensures that the system remains balanced and efficient throughout its operation.

The mathematical representation of the task clustering and scheduling problem provided a clear and structured framework for the DE algorithm. This formalization ensured that the optimization process adhered to the constraints of resource capacities and task requirements, leading to feasible and effective solutions. This structured approach addresses the limitations of previous heuristic and ad-hoc methods, providing a more reliable and consistent optimization strategy.

The experimental evaluation further supports these findings. Compared to PSO and FA, the proposed method achieved superior results across all key performance metrics, including execution time, CPU utilization, network latency, and load balancing efficiency. This consistent advantage under controlled IIoT conditions confirms the method's practical value and robustness, highlighting its potential for deployment in real-time edge computing systems with dynamic workloads.

The prospects of this research are promising. The DE-based mechanism's adaptability and efficiency make it well-suited for real-time IoT applications, where the ability to process data quickly and efficiently is critical. Future research could explore the integration of other optimization techniques with DE to further enhance its performance. Additionally, the application of this method to other areas of edge computing and its scalability to larger IoT networks are potential avenues for further investigation.

In conclusion, the proposed method for task clustering and scheduling in IoT edge computing represents a significant advancement in the field. It offers a scalable, efficient, and robust solution that enhances the performance and reliability of IoT applications. This research not only validates the potential of differential evolution in optimizing IoT edge computing systems but also sets a new standard for future studies in this domain.

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Performance Analysis of YOLOv8 and YOLOv11 Models on Automatically Annotated Data for Localization of Biological Objects in Images

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***Abstract*—The paper investigates the problem of detecting and localizing biological objects in microscopic images using deep learning methods. The main attention is paid to the automated creation of annotated datasets for training deep learning models, which is a relevant task in biomedical applications, where manual labeling is laborious and expensive. A method for automatically generating blood cell labeling using classical computer vision methods and the k-means clustering algorithm is proposed. The resulting dataset was adapted to the format of YOLO models (versions YOLOv8 and YOLOv11), which were used for cell detection and classification. A comparative analysis of the results of model training was performed using the metrics of accuracy, completeness, average accuracy (mAP), and speed. The YOLOv8 model showed higher accuracy and stability, while YOLOv11 was characterized by higher processing speed. The results demonstrate the effectiveness of using automatically annotated data for computer vision tasks in biomedicine and confirm the feasibility of further improving approaches to automatic image labeling.**

Keywords—computer vision, optimization, object detection, YOLO, deep learning, biomedical image analysis, automatic annotation, k-means clustering, image processing, microscopic images, blood cell classification

I. INTRODUCTION

Image recognition is a key task in modern biomedicine, diagnostics, scientific research, and laboratory automation. It involves the automatic recognition and localization of cells, tissues, organelles, or other structures in microscopic, X-ray, tomographic, or macroscopic images.

The modern view of the problem of object detection, localization and segmentation is to consider such tasks as deep learning problems due to the ability of such an approach to effectively process complex, high-dimensional and visually variable data like images [1]. Classical computer vision algorithms based on rules, filters and manually developed features, although successful in some application cases, are significantly limited in the real biological environment, where objects can be distorted, partially overlapped, differ in size, shape or color.

Deep learning, specifically convolutional neural networks, allows for automatic learning of features from data, without the need for manual intervention. This means that the model can detect complex patterns and contexts that are difficult or even impossible to describe manually. In the case of biological objects – cells, bacteria, and tissue structures – this is especially valuable, as their appearance varies greatly depending on the staining method, equipment, patient, or stage of disease.

In addition, deep learning scales well – the more data we have, the better the model can work. This allows us to create highly accurate systems for detecting even very rare or inconspicuous biological structures. Modern methods provide not only classification, but also localization of objects in images, which makes them ideal for detection tasks. In addition, deep learning is easily adaptable to new tasks. Transfer learning enables the adaptation of models pre-trained on large-scale general image datasets to domain-specific biomedical tasks, achieving high performance even with limited training data.

But by their nature, deep learning methods used in computer vision are supervised machine learning methods, and therefore require annotated datasets. On the other hand, the task of object detection and localization can be considered a data science task, because the task of detecting objects in images is not just a technical implementation of a certain neural network model. It covers the entire data processing cycle inherent in data science: data collection, data preparation, data analysis and visualization, modeling, evaluation, and deployment. Moreover, according to [2], the first 3 stages take up to 85% of the time of a data scientist. So, one of the ways to optimize the time spent on the object detection process is to automatically create annotated datasets.

II. LITERATURE REVIEW

In a number of studies [3, 4, 5], approaches, algorithmic solutions, and experimental research results related to the automated creation of annotated data sets for their further use in machine learning tasks, in particular for the detection, localization, and segmentation of objects in images, were analyzed. In [4], a basic algorithm was proposed for the

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detection and classification of blood cells in microscopic images. The main tool of this approach was image processing methods, which allowed performing cell segmentation, i.e., isolating objects in the image. After the segmentation stage, the cells were classified by the area of the detected objects, which allowed them to be assigned to one of three classes: platelets, erythrocytes (red blood cells – RBC), or leukocytes (white blood cells – WBC).

In [3], this approach was adapted to generate an annotated dataset in CSV format, which stored the relevant metadata for each image: file name, bounding box coordinates of each object found, and its type. The classification was also based on empirically determined cell area sizes, which allowed for automatic label generation without the involvement of experts for the labeling process.

A later study [5] built on the approach proposed in [3], improving and extending it to support the creation of annotated datasets not just for blood cells, but for any objects found in microscopic images. In order to automatically classify segmented objects into classes, the k-means clustering algorithm was used, which made it possible to avoid manual area analysis and increase the versatility of the technique.

This study examines the practical use of the resulting automatically annotated dataset as a base for training deep learning models designed to accurately detect, localize, and classify cells in microscopic images.

III. Problem Statement

The task of detecting multiple objects in an image is to find all objects of given classes in the input image and determine their spatial boundaries (bounding boxes) along with the corresponding class labels.

Mathematically, the problem of detecting objects in images can be formalized as follows:

A set of images is given

$$X = \left\{ x_i \in \Re^{H \times W \times 3} \,\middle|\, i = 1,2,...,N \right\}, \tag{1}$$

where each image x_i has height H, width W, and 3 color channels (RGB).

For each image x_i there are a set of annotated objects:

$$Y_i = \left\{ \left(b_{ij}, c_{ij} \right) \middle| i = 1,2,...,M_i \right\}, \tag{2}$$

where M_i – number of objects in the image x_i; $b_i = \left(x_{ij}, y_{ij}, w_{ij}, h_{ij} \right)$ are center coordinates (sometimes the coordinates of the upper left corner can be considered – depending on the type of data set representation), width and height of the object's bounding box j in the picture i; $c_{ij} \{1,2,...,C\}$ is a object class label belonging to one of C possible classes.

The goal is to construct a function

$$f : \Re^{H \times W \times 3} \to \left| \left(\hat{b}, \hat{c}, s \right) \right|, \tag{3}$$

which for each image returns a finite set of tuples $\left(\hat{b}, \hat{c}, s \right)$, where $\hat{b} \in \Re^4$ is the object predicted bounding box, $\hat{c} \in \{1,2,...,C\}$ is the object class, $s_i \in [0,1]$ is the degree of confidence of the model in this prediction.

The model is trained by minimizing the total loss function

$$L = L_{cls} + \lambda \cdot L_{loc}, \tag{4}$$

where L_{cls} is a classification loss function, L_{loc} is a coordinates regression loss function, $\lambda \in \Re_+$ is a weight coefficient that regulates the importance of localization.

IV. Overview of Tools and Models Used

To implement image processing and object segmentation methods, the OpenCV library was used, which provided functionality for filtering, binarization, and contour extraction. Metadata processing and storage were performed using the Pandas and NumPy libraries. The k-means algorithm from the Scikit-learn library was used for automatic clustering and pre-classification of objects [5].

Two modern architectures were tested to build object detection models: YOLO (You Only Look Once), implemented using the Ultralytics frameworks: YOLOv8 and YOLOv11 [6]. This architecture provided high image processing speed and became suitable for applications where performance is critical.

The generated annotations in CSV format were adapted to the requirements of each model, which made it possible to effectively use automatically generated data for training and validating deep learning models.

V. Training the YOLOv8/YOLOv11 Model

As a result of applying the methodology considered in [5], a data set was created in CSV format with the following columns:

– `file_name` is an image file name;

– `xmin`, `ymin`, `xmax`, `ymax` are coordinates of the upper left and lower right corners of the bounding box;

– `type` is a cell type.

For YOLO models, the dataset must be in a specific format [7].

The directory structure should correspond to the structure given in Fig. 1. That is, the dataset should be divided into an image part and an annotation part.

```
yolo_dataset/
├── images/
│   ├── train/
│   └── val/
├── labels/
│   ├── train/
│   └── val/
```

Fig. 1. Directory structure of YOLO dataset

The annotations for the YOLO model should in turn be provided in a text file with the same name as the image. Each text file should contain information in the format

`<class_id> <x_center> <y_center> <width> <height>,`

where `class_id` is an integer (0, 1, 2, …) corresponding to the class; `x_center`, `y_center` is the center of the object, normalized to [0, 1]; `width`, `height` are the width and height of the bounding box, also normalized to [0, 1].

To convert the data set obtained in [5], a sequence of actions is proposed. The process begins with defining input parameters, including the path to the CSV file with annotations, the directory with the corresponding images, and the directory for outputting the results, and a class map that maps object names to the numeric identifiers (`class_id`) required for the YOLO format.

After that, the CSV data is read into a structure conveniently presented as a DataFrame, which allows easy filtering and processing of information. At this stage, the names of object types are also transformed into numerical indices that will be used when generating labels. Next, a directory structure is created that corresponds to the format expected by the YOLO framework: separate folders for images and labels, which are divided into training and validation samples (train and val).

All unique image names mentioned in the CSV are collected and divided into train and val subsets. This ensures balance and reliability during model training and testing. Next, the image processing stage begins. For each image, its presence in the input directory is checked. If the file is missing, processing is skipped. If the file is found, its size is read, after which all objects corresponding to this image are selected from the CSV file.

An important step is to convert the object coordinates from absolute values to the normalized YOLO format – the center of the object and its dimensions in relative coordinates (from 0 to 1). For each object, a text string is created in the appropriate format, which will later be written to a text file with the same name as the corresponding image.

After generating the labels and formatting them in a YOLO-compatible format, the image is copied to the appropriate subfolder (train or val), and the generated .txt file with the labels is saved in a similar substructure. The final step in the entire process is to create a configuration file dataset.yaml, which contains the paths to the data subsets, the number of classes, and their names. This file serves as a key element for further running the training process.

After applying the algorithm, each image annotation file looks like this:

```
0 0.659722 0.873278 0.058333 0.055096
0 0.425000 0.853994 0.066667 0.066116
0 0.263889 0.714876 0.044444 0.046832
1 0.502778 0.498623 0.322222 0.424242
2 0.077778 0.926997 0.155556 0.140496
2 0.831944 0.899449 0.219444 0.195592
2 0.915278 0.735537 0.163889 0.176309
2 0.684722 0.735537 0.191667 0.198347
2 0.088889 0.701102 0.177778 0.239669
2 0.777778 0.314050 0.188889 0.214876
2 0.261111 0.268595 0.222222 0.201102
2 0.956944 0.238292 0.080556 0.157025
2 0.906944 0.079890 0.180556 0.159780
2 0.483333 0.103306 0.211111 0.206612
```

Fig. 2. Contents of the annotation file

Model training was performed in the Google Colab environment using the T4 GPU (NVIDIA Tesla T4) [8].

The result of detection and localization using bounding boxes is given in Fig. 3.

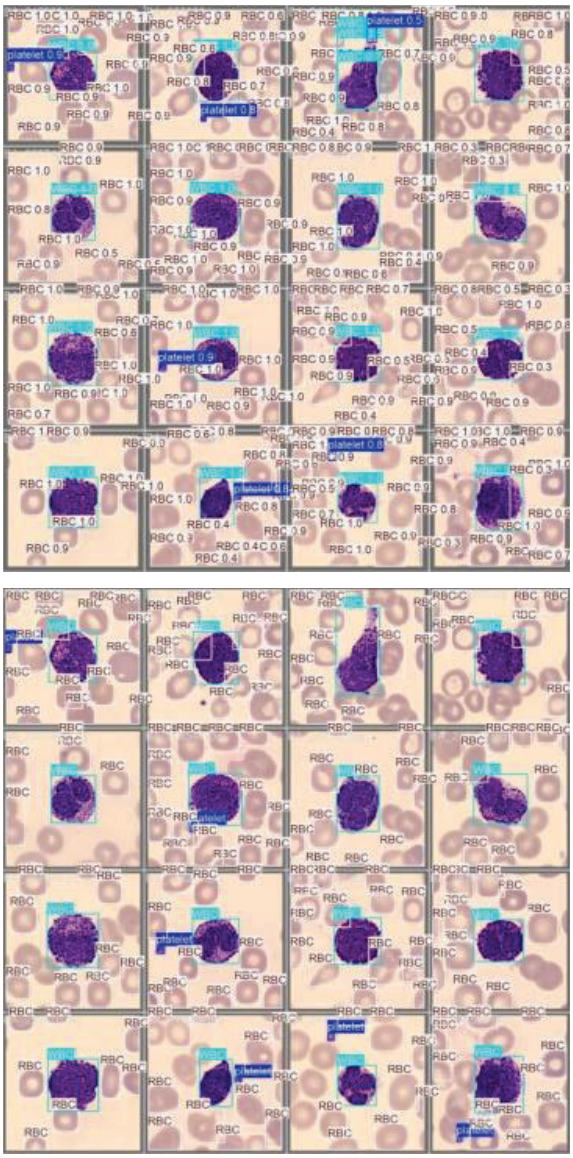


Fig. 3. Result of blood cell localization using YOLOv8 and YOLOv11 models

The results of training and validation of the YOLOv8 and YOLOv11 models are given in Table 1.

TABLE I. MODEL TRAINING QUALITY METRICS

Metrics	YOLOv8	YOLOv11
precision	0.909	0.834
recall	0.855	0.86
mAP50	0.879	0.867
mAP50-95	0.788	0.756
fitness	0.797	0.766
preprocess	0.197	0.0855
inference	2.005	1.193
loss	0.0004	0.0004
post-process	1.697	1.538

The Table 1 shows a comparative analysis of the two versions of the YOLO model in terms of accuracy, efficiency, and performance. First, we present the classic metrics for assessing the quality of computer vision models: precision, recall, mAP50(B), mAP50–95, and the aggregate fitness indicator, which is often used as a weighted combination of previous metrics for a generalized assessment.

YOLOv8 shows slightly higher precision (0.909) compared to YOLOv11 (0.834), but YOLOv11 has a slight advantage in recall (0.86 vs. 0.855). mAP metrics, which measure average accuracy at different IoU thresholds, also show better performance for YOLOv8 at a threshold of 0.5 (0.879 vs. 0.867), but a decrease in accuracy in the range of 0.5–0.95 is typical for both models. The fitness indicator is higher for YOLOv8 (0.797 vs. 0.766), indicating a better overall balance of this version.

The next part of the table concerns the performance of the models, i.e. the time required to perform each of the image processing stages: preprocessing, inference (prediction itself), calculation of the loss function and postprocessing of the results. YOLOv11 significantly outperforms YOLOv8 in terms of speed at all stages. In particular, the inference time of YOLOv11 is 1.193ms versus 2.005ms in YOLOv8, which is almost twice as fast. This may indicate an optimization of the architecture or a reduction in the amount of calculations. Despite the same loss value (0.0004), the overall performance of YOLOv11 indicates its potential advantage for real-time tasks.

Thus, the data from the table illustrates a typical trade-off between recognition accuracy and speed, where YOLOv8 shows better prediction quality, and YOLOv11 shows higher speed. The choice of model depends on the requirements of a particular application: if high accuracy is important, then you should choose YOLOv8; if speed is critical, then YOLOv11 will be a better candidate.

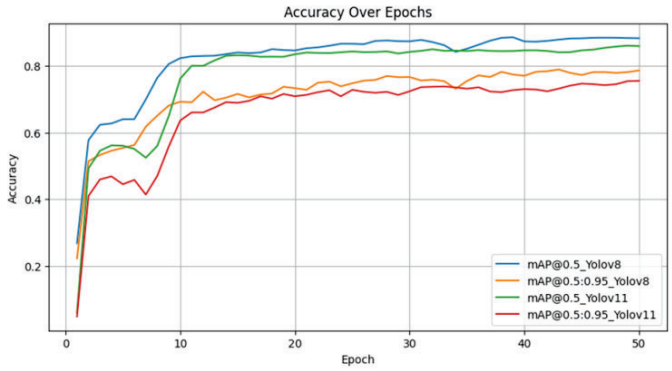


Fig. 4. Model learning dynamics

The graph presented in Fig. 4 shows the dynamics of changes in the accuracy of the YOLOv8 and YOLOv11 models during the training process for 50 epochs. The X-axis shows the number of epochs, while the Y-axis displays the accuracy values according to two metrics – mAP@0.5 and mAP@0.5:0.95. These metrics are standard for assessing the quality of object recognition models: the first is less demanding and takes into account positive predictions that have at least 50% intersection with the real object, the second is more strict and displays the average accuracy at different IoU thresholds from 0.5 to 0.95.

According to the graph, the YOLOv8 model demonstrates significantly higher accuracy compared to YOLOv11 on both metrics. Already at the first stages of training, a rapid increase in mAP values for YOLOv8 is visible – by about the 10th epoch it reaches high results and then only slightly fluctuates around a stable level. In particular, mAP@0.5 for YOLOv8 exceeds 0.85, while its accuracy according to the stricter criterion mAP@0.5:0.95 is kept within 0.75.

Against this background, YOLOv11 looks weaker: although its mAP@0.5 also increases and approaches the level of YOLOv8, the lag is still noticeable. The largest gap is observed in the accuracy of the mAP@0.5:0.95 metric, which in YOLOv11 remains at about 0.70, falling behind YOLOv8 by almost 5 percent.

These results indicate that YOLOv8 is better able to learn and generalize quickly, even at early stages. It can be assumed that the architecture of this model is more optimized, or that its hyperparameters are better matched to the data used for training. In any case, the graph analysis convincingly demonstrates the superiority of YOLOv8 over the experimental YOLOv11, especially in the context of high-accuracy scenarios, where the mAP@0.5:0.95 metric is critical.

VI. MODEL TESTING

The results presented in Table 1 are obtained based on the validation that was performed on an automatically generated dataset. This dataset contains undetected and incorrectly detected objects of all classes [5]. Therefore, to check the efficiency of object recognition, manual testing was performed on 150 images [9]. We will consider the problem of object detection in an image, so we can formalize the problem of testing the model as a multiclass and multiobject problem.

The task of classifying a set of objects in images can be viewed as a task of element-wise comparison of vectors of quantities by class, followed by the formation of partial estimates for each class. This approach allows us to evaluate the performance of models in the tasks of counting and classifying biological objects without resorting to more complex methods of position analysis or geometric characteristics.

We have a set of images I_n, where $i = 1, 2, ..., N$, and the number of possible classes is denoted as K. For each image, a vector of real values is given $Y^{(i)} = \left(y_1^{(i)}, y_2^{(i)}, ..., y_K^{(i)}\right)$, where $y_k^{(i)} \in N_0$ indicates the number of objects k-th class that are actually present in the image. Similarly, the model generates a vector of predicted values $\hat{Y}^{(i)} = \left(\hat{y}_1^{(i)}, \hat{y}_2^{(i)}, ..., \hat{y}_K^{(i)}\right)$, where each component represents the number of objects of the class k detected in the image by the model.

Since the task is to correctly identify only the classes and their quantities, we can form an independent estimate of the classification accuracy for each class using the sums over all images. In particular, for each class k. The number of correct predictions is determined, that is cases where both the model and the label agree on the number of objects of this class. In a simplified approximation, the model is considered to have correctly classified objects if the number of predicted

objects does not exceed the number of real objects, and we can interpret the minimum between these two values as the number of true positives.

Thus, for each class k the total value of true positive predictions is defined as the sum

$$TP_k = \sum_{i=1}^{N} \min\left(y_k^{(i)}, \hat{y}_k^{(i)}\right). \quad (5)$$

The number of false positives for a class is the total excess of predicted objects over the actual ones:

$$FP_k = \sum_{i=1}^{N} \max\left(\hat{y}_k^{(i)} - y_k^{(i)}, 0\right), \quad (6)$$

and false negatives – the lack of predictions compared to the actual number:

$$FN_k = \sum_{i=1}^{N} \max\left(y_k^{(i)} - \hat{y}_k^{(i)}, 0\right). \quad (7)$$

These values allow us to construct confusion matrices (Fig. 5 – 7), as well as to determine standard classification quality metrics: precision, recall, and F1-measure [3] for each class (Table 2).

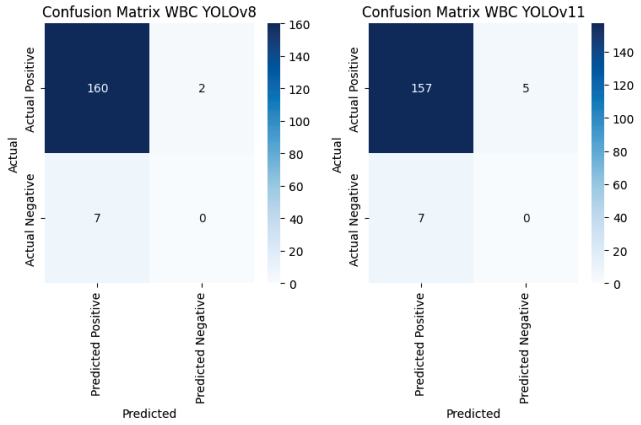


Fig. 5. Error matrices for leukocytes

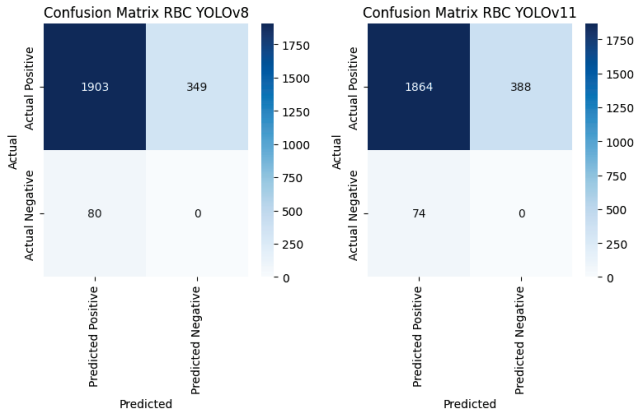


Fig. 6. Confusion matrix for red blood cells

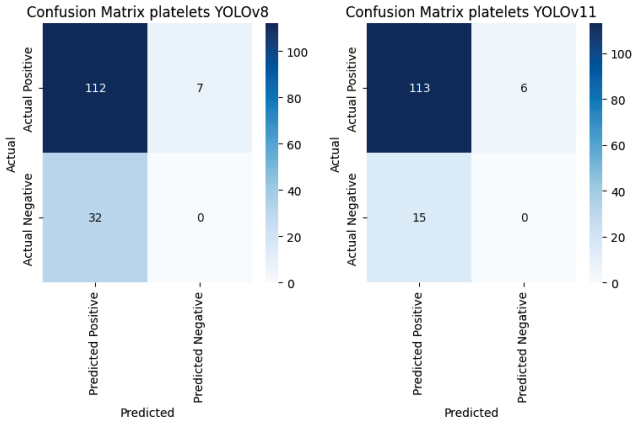


Fig. 7. Error matrices for platelets

TABLE II. MODEL QUALITY METRICS

Metrics	YOLOv8			YOLOv11		
	WBCs	RBCs	Platelets	WBCs	RBCs	Platelets
Accuracy	0.9467	0.8160	0.7417	0.9290	0.8014	0.8433
Precision	0.9581	0.9597	0.7778	0.9573	0.9618	0.8828
Recall	0.9877	0.8450	0.9412	0.9691	0.8277	0.9496
F1	0.9726	0.8987	0.8517	0.9632	0.8897	0.9150

A comparison of the YOLOv8 and YOLOv11 models (Fig. 8 – 9) for the blood cell recognition task demonstrates interesting dynamics of changes in the classification quality of the three main cell types: leukocytes (WBCs), erythrocytes (RBCs), and platelets (Platelets). Although the overall picture of the metrics for the models looks quite similar, a detailed analysis reveals key advantages and disadvantages of each of them.

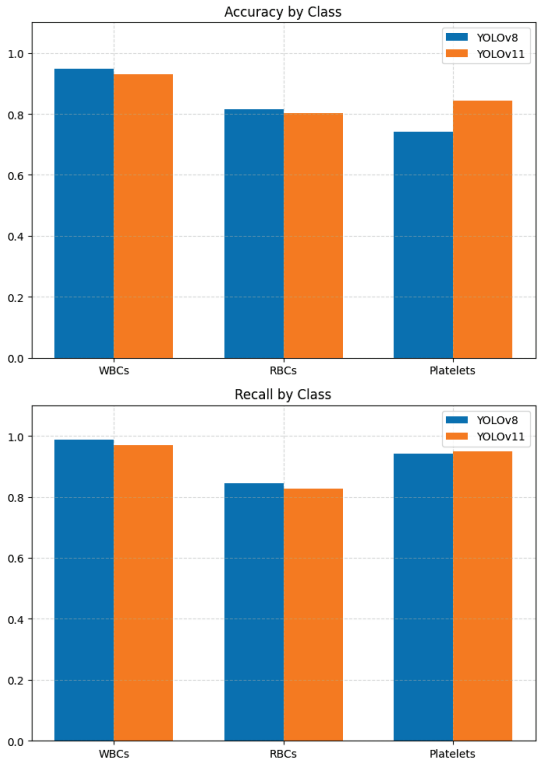


Fig. 8. Comparison of quality metrics

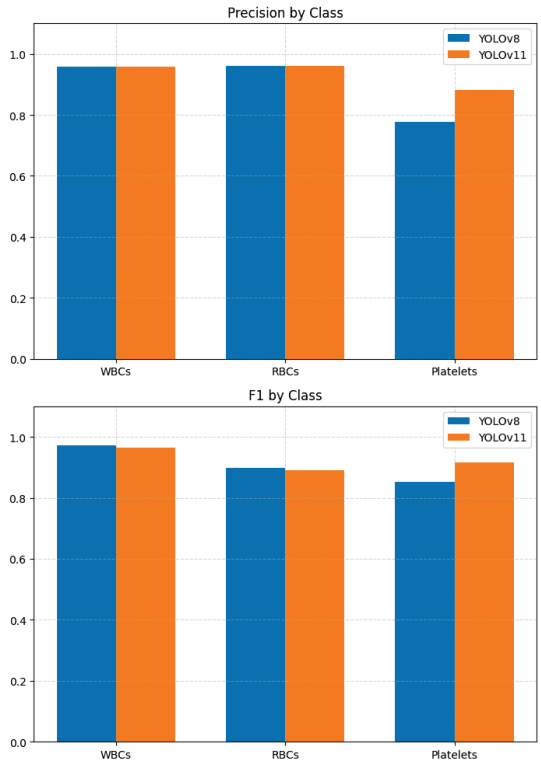


Fig. 9. Comparison of quality metrics

YOLOv8 demonstrates a high level of accuracy and completeness in the detection of leukocytes. Its recall reaches almost 99% and the F1 metric is over 97%, indicating a very small number of missed objects in this class. This makes YOLOv8 extremely reliable in detecting large and well-defined cells such as WBCs. For erythrocytes, the model also shows confident results, especially in terms of precision and the balance between precision and recall, as reflected in the F1 metric, which exceeds 89%.

In turn, YOLOv11 brings a significant improvement in the platelet category – the smallest cells, which are often difficult to detect due to their similarity to artifacts or image noise. All key metrics increase significantly compared to YOLOv8, making the newer model much more effective in this difficult sub-task. This suggests that YOLOv11 has a better ability to focus on small details or uses more optimized architectural components or loss functions that are better suited to detecting such objects.

Despite these improvements, YOLOv11 slightly loses accuracy and F1-metric in recognizing WBCs and RBCs compared to YOLOv8. These differences are not critical, but they highlight that optimizing a model for one class (platelets) can come at the expense of others.

Thus, the choice of model depends on the application goals. If the main task is the most accurate and stable recognition of all three cell types, YOLOv8 seems to be a more balanced choice. However, in cases where accurate platelet detection is critical, YOLOv11 becomes a better candidate due to its higher efficiency in this category.

VII. Conclusions

The article investigates the problem of detection, localization and classification of biological objects in microscopic images using deep learning methods. As a training dataset, a dataset was used, automatically created on the basis of unannotated images using classical image processing algorithms and the k-means clustering algorithm, which allows avoiding manual labeling and significantly accelerating data preparation for training models.

The YOLOv8 and YOLOv11 architectures were used to solve the problem. Training and validation of the models showed that YOLOv8 provides higher accuracy and generalization, in particular in terms of metric accuracy, mAP@0.5 and mAP@0.5:0.95, compared to YOLOv11. In contrast, YOLOv11 showed a significantly higher processing speed, which makes it attractive for real-time tasks. Both models demonstrate high efficiency in solving the task, which indicates the suitability of modern YOLO object-detector architectures for medical image analysis. YOLOv8 proved to be a balanced solution that provides high-quality recognition of leukocytes and erythrocytes. At the same time, YOLOv11 demonstrated a noticeable improvement in the recognition of platelets, the smallest and most difficult to identify objects.

This suggests that model updates likely include optimizations that are better suited for small object recognition, which is especially valuable in medical scenarios where even insignificant details can have diagnostic weight.

Thus, it can be concluded that the use of automatically generated annotations in combination with modern deep learning architectures is an effective approach to solving biomedical image recognition problems, and the choice of a specific model (YOLOv8 or YOLOv11) depends on the accuracy or speed requirements in a particular case.

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Intercalation Synthesis of Lead-Free Pigments and Fillers based on Montmorillonite Clays for Paint and Coating Materials

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***Abstract*—To effectively address the challenge of expanding the range of paint and coating materials, a nanotechnology has been developed for producing lead-free pigments and fillers based on natural domestic montmorillonite-containing raw materials. This was achieved through the activation of red lump bentonite clays using various inorganic acids and sodium phosphate. A comparative analysis of activation methods was conducted. The influence of salt hydrolysis reactions via cations and anions on the efficiency of bleaching lump bentonite clays was studied. A resource- and energy-efficient technological process was modified to produce white fillers, organomodified bentonite, and pigments of various colors from the triphenylmethane series using an intercalation method. Formulations were calculated, and industrial paint and coating materials were developed for protecting metal, plastic, and wooden surfaces against corrosion. The physicotechnical properties of the materials and coatings were experimentally investigated, and their areas of application were identified and recommended. The mechanism of the anticorrosive action of a complex electrochemical corrosion inhibitor based on the synthesized pigments was studied.**

Keywords—paint and coating materials, redox reactions, hydrolysis, intercalation, resource conservation, fillers, pigments for paints and coatings

I. Introduction

Anisotropic solids with a layered structure are capable of forming intercalation compounds [1]–[3] through the insertion of various atoms, ions, and molecules (intercalants) into the interlayer galleries of crystalline matrices. Intercalation compounds are promising for use as catalysts in chemical reactions, electrodes in chemical power sources, solar cells, optical data storage media, semiconductor materials with variable mini-band electron spectra, materials for electrochromic display cells, and monochromators for neutron diffractometry. Layered aluminosilicates [4] are widely used as fillers for polymers [5, 6] and paint and coating materials (PCMs) [7]. This work demonstrates the possibility of synthesizing fillers and pigments based on montmorillonite – a layered aluminosilicate found in bentonite clays – using the intercalation method, and their application in paint and coating technologies to produce nanocomposite decorative and protective-decorative materials and coatings.

Pigments [8] and fillers in PCMs [9] perform multiple functions [10, 11], ranging from providing improved physico-mechanical properties and imparting specific color and decorative qualities to offering special properties such as anticorrosion [12].

With the adoption in Ukraine in 2021 of the technical regulation [14] restricting and prohibiting lead content in pigments and pigmented coatings, the range of available pigments has significantly narrowed. Particularly, a large number of anticorrosive pigments, such as lead silicochromate, lead primer chromate, basic lead carbonate (lead white), lead red lead, lead chromates, and lead-molybdate chromate, are now banned from use [15, 16]. The only anticorrosive pigment permitted under European standards remains zinc chromate [17].

Therefore, research aimed at expanding the range of lead-free pigments, especially anticorrosive, ones remains a pressing task.

Eurointegration processes shaping the development of the domestic paint and coatings industry are driving changes in the properties of modern coatings aimed at corrosion protection and improving the decorative appearance of products. These trends necessitate the development and implementation of innovative, environmentally safe paint and coating materials. Such materials include water-based (WB) and water-dispersion coatings utilizing water as a solvent [18], solvent-based (SB) materials with low volatile organic compound (VOC) [19] content [20], and powder coatings that comply with technical regulations limiting VOC content in coatings [21, 22]. Since 2021 [3], one of the key requirements for pigments and pigmented materials is the absence of lead, a heavy metal, in the composition of paint and coating materials.

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The aim of the work is the development of paint and coating materials for construction and industrial applications, as well as innovative technologies for their production using synthesized lead-free pigments and fillers. The scientific novelty of the work is also related to the modification of the redox method for obtaining white fillers [23] based on red lump bentonite clays and the intercalation method for synthesizing lead-free anticorrosive pigments and organomodified bentonite based on white bentonite fillers.

Achieving the objective of the work involves solving the following tasks:

- investigating the influence of hydrolysis processes on the redox mechanism of bentonite bleaching, and applying sodium phosphate salt in the activation method of red lump bentonite clay, which allows modifying the technology by removing the hydrolysis stage of water-soluble ferrous (II) sulfate and ferric (III) sulfate salts from the technological process;
- obtaining anticorrosive pigments and organomodified bentonite based on sodium bentonite by the intercalation method, free of lead in both raw materials and final products;
- developing formulations and innovative technologies with aspects of energy and resource saving to produce decorative water-soluble and water-dispersion paint and coating materials, anticorrosive primers, and enamel PCMs using synthesized pigments, fillers, and organomodified bentonite;
- investigating the physico-mechanical properties of the materials and formulating recommendations for the use of the developed materials.

II. Materials and Research Methods

Red lump bentonite clay was used as the raw material for the study, characterized by the following properties: montmorillonite content – 50 wt%, iron content expressed as Fe_2O_3 – 6.3 wt%, colloidal fraction – 93 %, water absorption – 5.6 units, and carbonate content calculated as $CaCO_3$ – 2.25 wt%. The chemical composition of the bentonite clay corresponds to the formula: $(Na,Ca)_{0.33}(Al,Mg)_2 Si_4O_{10}(OH)_2 \cdot nH_2O$.

For the modification of bentonite, solutions of alkyl dimethyl benzyl ammonium chloride (ADBC) and dimethyldi-n-octadecyl ammonium chloride (DDAC) (Alfa Chemical Co., Ltd.) were employed. Granulometric analysis of samples was conducted using a laser granulometer from Malvern. Samples of bentonite and their intercalation derivatives were studied by Fourier-transform infrared spectroscopy (FTIR) using a Specord IR-75 spectrophotometer with a recording scale of 150 mm/100 cm^{-1}. Samples were prepared by pressing into KBr pellets. X-ray phase analysis of samples was performed on a Dron-3 diffractometer using Cu Kα radiation and a nickel β-filter. Differential thermal analysis (DTA) of the organomodified bentonite was conducted using a Paulik–Paulik–Erdey derivatograph. Color characteristics of the investigated pigment and filler samples were determined using an NS 800 spectrophotometer.

III. Results and Discussion

With the aim of improving and modifying the technology for obtaining white bentonite fillers based on the redox mechanism [23], the activation stage involved treating red lump bentonite clay with aqueous solutions of sodium phosphate. It is known that the main requirements for fillers used in paint and coating materials are high whiteness degree, weather resistance, high dispersibility, low hardness, low cost and availability of raw materials, and minimal content of water-soluble substances.

Due to the red color of the lump clay caused by the iron content in the raw material amounting to 6.3 % calculated as Fe_2O_3, the bentonite bleaching technology involves activation of the clay with mineral hydrochloric or sulfuric acids to convert iron into water-soluble salts. The resulting water-soluble ferrous (II) chlorides and ferrous (II) sulfates, as well as ferric (III) chlorides and ferric (III) sulfates, undergo hydrolysis. The hydrolysis products of these salts with respect to the cation are ferrous (II) hydroxides and ferric (III) hydroxides. Due to oxidation reactions of ferrous (II) oxide in the presence of atmospheric oxygen, ferric (III) hydroxide of brownish color is formed, which causes a pink tint to appear in the color of the fillers. To prevent and inhibit the hydrolysis and redox reactions leading to the formation of ferric (III) hydroxide, the reaction is carried out in an acidic environment at pH ≤ 1–3. Low pH values shift the chemical equilibrium according to Le Chatelier's principle toward the formation of the initial substances, namely ferrous (II) sulfates and chlorides. To prevent the progression of oxidation processes increasing the oxidation state of Fe^{2+} cations to Fe^{3+}, the bentonite bleaching technology requires mandatory addition of reducing agents to the reactor-activator: iron scrap at the first stage and metallic zinc at the final stages of bleaching. The present work proposes a modification of the technology for producing bleached bentonite fillers by treating the clay suspension with sodium phosphate. The advantages of using sodium phosphate include, firstly, the ability to alter the pH of the medium due to hydrolysis of this salt by the anion and the formation of solutions with pH ≥ 7. The cation exchange reactions using sodium phosphate can be schematically represented as follows:

$$Na_3PO_4 + Mnt\text{-}(Fe^{2+}/Ca^{2+}/Mg^{2+}) \leftrightarrow$$

$$\leftrightarrow Mnt\text{-}Na^+ + (Fe^{2+}/Ca^{2+}/Mg^{2+})_3(PO_4)_2\downarrow, \quad (1)$$

$$Na_3PO_4 + Mnt\text{-}(Fe^{3+}) \leftrightarrow Mnt\text{-}Na^+ + Fe^{3+}PO_4\downarrow. \quad (2)$$

Both reactions are strongly shifted to the right due to the formation of precipitates of water-insoluble salts of ferrous (II) and ferric (III) ions, as well as calcium and magnesium phosphates.

In this work, X-ray phase analysis was used to identify montmorillonite in the bentonite composition and to establish the occurrence of intercalation processes during the synthesis of organo-modified montmorillonite. The advantages of treating clay with sodium phosphate salts, apart from eliminating the hydrolysis stages of cations that occur when mineral acids are used at the activation stage and the reduction stages, include the conversion of the calcium-magnesium form of montmorillonite into the sodium form. Experimental confirmation of the intercalation and cation exchange processes was obtained through X-ray phase analysis of the initial lump clay samples and the samples obtained after treatment with sodium phosphate. According to X-ray phase

analysis data, the investigated sample exhibits a rather wide range of distances between the silicate layers of the clay, with an average value of 15.3 Å, which is characteristic of the calcium form of bentonite. After purification and treatment of the clay with sodium phosphate, the montmorillonite predominantly transforms into the sodium form. The interlayer spacing in sodium-form montmorillonite is 12.8 Å. Calculations were performed using the Bragg-Wulff equation. As a result of the X-ray phase analysis, the identity of the structure of the studied montmorillonite samples, a layered mineral predominant in the bentonite composition, with the montmorillonite structure was established. The position of the reflections on the montmorillonite diffractogram in the low-angle region corresponds to the basal interplanar spacing of 15.3 Å characteristic of montmorillonite. The sample of bleached, neutralized, and organo-modified bentonite with ADBC and DDAC shows intense broad reflections in the low-angle range with d = 18.47 Å and d = 18.7 Å, respectively, which is an indication of the formation of an intercalated structure of the organo-modified bentonite. The broadening of the low-angle peak may indicate the presence of exfoliated, delaminated structures in the intercalation compounds within the «montmorillonite-ADBC" and "montmorillonite-DDAC» systems. The hydrolysis reactions of sodium phosphate (pH ≥ 7) can be described by the following molecular reactions:

stage I of hydrolysis:

$$Na_3PO_4+HOH\leftrightarrow Na_2HPO_4+NaOH, \quad (3)$$

stage II of hydrolysis:

$$Na_2HPO_4+HOH\leftrightarrow NaH_2PO_4+NaOH, \quad (4)$$

stage III of hydrolysis:

$$NaH_2PO_4+HOH\leftrightarrow H_3PO_4+NaOH. \quad (5)$$

The protective mechanism of iron phosphates may be associated with the formation of a complex acid according to the following chemical reaction:

$$Fe_3(PO_4)_2+4H_2O=[Fe_3(PO_4)_2(H_2O)_2(OH)_2]^{2-}+2H^+, \quad (6)$$

which reacts with iron ions on the metallic surface at microanodes forming a complex corrosion inhibitor: $[Fe_3(PO_4)_2(H_2O)_2(OH)_2]Fe$, reducing the anodic area of micro-galvanic elements on the metal surface and inhibiting electrochemical corrosion processes.

The complex inhibitor's outer coordination sphere may include a carboxyl group -*COOH* from the film-forming agent instead of one water molecule, specifically:

$$Fe_3(PO_4)_2+3H_2O+\text{-}COOH=$$
$$=[Fe_3(PO_4)_2(H_2O)(COOH)(OH)_2]_2. \quad (7)$$

Thus, the technological process of bleaching red lump bentonite clay by treatment with sodium phosphate consists of the following stages: formation of an aqueous suspension of lump bentonite clay; filtration and removal of mechanical impurities; sedimentation followed by decantation to remove water-soluble substances; cation exchange using sodium phosphate; drying using a jet dryer at temperatures below 40 °C with particle size classification up to 50 μm. At this stage, as a resource-saving measure, water recovery is provided for reuse as return water in the first stage of the technological process of forming the aqueous clay suspension.

To obtain organomodified bentonite, it is necessary to modify the above technological process by adding an intercalation stage of montmorillonite with quaternary ammonium salts from aqueous solutions after the third stage and before the fourth stage drying of the finished product. It should be noted that the developed technological process using sodium phosphate allows obtaining not only white fillers but also imparts anticorrosive properties due to the presence in their chemical composition of water-insoluble iron (II) phosphates, iron (III) phosphates, calcium phosphates, and magnesium phosphates.

The intercalation and cation exchange reactions using sodium phosphate and ADBC can be schematically represented as follows:

$$Mnt\text{-}Na_{0.33}^{+}+_d[C_6H_5(CH_2)(CH_3)_2RN]^+Cl^-\leftrightarrow$$
$$\leftrightarrow Mnt[C_6H_5(CH_2)(CH_3)_2RN]_d^{+}+0.33NaCl;\ R=C_{16}\text{–}C_{18}. \quad (8)$$

The intercalation and cation exchange reactions using sodium phosphate and DDAC can be schematically represented as follows:

$$Mnt\text{–}Na_{0.33}^{+}+_d[(CH_3(CH_2)_{17})_2(CH_3)_2N]^+Cl^-\leftrightarrow$$
$$\leftrightarrow Mnt[(CH_3(CH_2)_{17})_2(CH_3)_2N]_d^{+}+0.33NaCl. \quad (9)$$

For the purpose of identifying and confirming the purity of raw materials, as well as determining the nature of interactions between quaternary ammonium salts, triphenylmethane dye molecules, and montmorillonite, the synthesized pigments and fillers were studied using spectral analysis methods in the visible and infrared regions of the spectrum. The studies in the visible region were conducted to determine colorimetric characteristics.

The intercalation and cation exchange reactions involving ADBC and malachite green dye molecules can be schematically represented as follows:

$$Mnt\text{-}Na_{0.33}^{+}+_d[C_6H_5(CH_2)(CH_3)_2RN]^+Cl^-+$$
$$+_y[C_{23}H_{25}NN]^+Cl^-\leftrightarrow$$
$$\leftrightarrow Mnt[C_6H_5(CH_2)(CH_3)_2RN]_d^{+}[C_{23}H_{25}NN]_y^{+}$$
$$+0.33NaCl. \quad (10)$$

The intercalation and cation exchange reactions involving DDAC and malachite green dye molecules can be schematically represented as follows:

$$Mnt\text{-}Na_{0.33}^{+}+_d[(CH_3(CH_2)_{17})_2(CH_3)_2N]^+Cl^-+$$
$$+_y[C_{23}H_{25}NN]^+Cl^-\leftrightarrow$$
$$\leftrightarrow Mnt[(CH_3(CH_2)_{17})_2(CH_3)_2N]_d^{+}[C_{23}H_{25}NN]_y^{+}+$$
$$+0.33NaCl. \quad (11)$$

The intercalation and cation exchange reactions involving ADBC and rhodamine C dye molecules can be schematically represented as follows:

$$Mnt\text{-}Na_{0.33}^{+}+_dC_{28}H_{31}O_3ClN_2+_yC_{27}H_{50}N^+Cl^-\leftrightarrow$$
$$\leftrightarrow Mnt(C_{28}H_{31}O_3NN)^+_d(C_{27}H_{50}N)^+_y]+0.33NaCl. \quad (12)$$

The intercalation and cation exchange reactions involving DDAC and rhodamine C dye molecules can be schematically represented as follows:

$$[(CH_3(CH_2)_{17})_2(CH_3)_2N]^+Cl^-+Mnt\text{-}Na_{0.33}+$$

$$+_d[(CH_3(CH_2)_{17})_2(CH_3)_2N]^+Cl^- +_y C_{27}H_{50}N^+Cl^- \leftrightarrow$$
$$\leftrightarrow Mnt[(CH_3(CH_2)_{17})_2(CH_3)_2N]_d^+(C_{27}H_{50}N)^+_y] +$$
$$+0.33NaCl. \quad (13)$$

The color characteristics of the synthesized samples were investigated as follows: sample №1 (white fillers): color coordinates in the XYZ system: X=46; Y=55; Z=63; chromaticity coordinates: x=0.28; y=0.34; Y=55; color coordinates in the CIELAB system: L=79.25; A=2.04; B=7.98; dominant wavelength λ=473 nm; color purity p=7 % – determined from the chromaticity diagram.

For samples of Mnt treated with sodium phosphate and ADBC (№2), the following color parameters were determined: X=65; Y=69; Z=67; L=86.68; A=1.94; B=6.69; λ=498 nm; p=7%.

For samples of Mnt modified with ADBC and malachite green dye (№3): X=27; Y=34; Z=38; L=64.72; A=18.75; B=1.27; λ=508 nm; p=21%.

For samples of Mnt modified with ADBC and rhodamine C dye (№4): X=35; Y=26; Z=43; L=57.95; A=38.99; B=18.81; λ=530 nm; p=28%.

Samples of green and red pigments were obtained at different concentrations of malachite green and rhodamine C – 0.01 M and 0.05 M – which affected the color characteristics of the materials. With an increased concentration of malachite green, the color purity increased from 21 % for sample №3 to 30 %, while increasing the concentration of rhodamine C raised the color purity from 28 % for sample №4 to 36 %. A shift in the dominant wavelength was observed from 508 to 521 nm for malachite green compounds, and from 530 to 610 nm for rhodamine C. Thus, increasing the dye concentration allows obtaining pigments with different color intensities within economically feasible limits. It should be noted that the low color purity values for samples bleached using sodium phosphate indicate the complete binding of iron cations as a result of activating the original red lump clay. The determined physical and technical properties of the fillers after treatment of bentonite with sodium phosphate are as follows: first-order oil absorption – 53.4 g/100g; coverage – 458 g/m²; density – 2700 kg/m³; water-soluble substances–absent; pH of the aqueous extract – 7.

Laser diffraction as a method for analyzing the granulometric composition of organomodified bentonite was applied in this work to analyze the granulometric composition of the obtained pigments, fillers, and organomodified bentonite.

As a result of a comparative analysis of the properties of the original lump clay samples and organomodified bentonite ADBC, an increase in homogeneity from 0.784 to 1.804 was established, along with an increase in the specific surface area of the samples from 323.3 to 1175 m²/kg. The surface modification of the bleached bentonite led to an increase in homogeneity to 1.804 and an increase in the specific surface area to 1175 m²/kg. The obtained results correspond well with the expected changes in the crystalline structure of the layered aluminosilicates transformed at various stages of the technological process. Indeed, the processes of washing out water-soluble substances, activation, and organomodification of the natural raw material are associated not only with the peptization of aluminosilicate particles in aqueous solutions but also with simultaneous intercalation reactions of quaternary ammonium salt molecules into the interlayer galleries of montmorillonite. It is known that the intercalation process leads to the destruction of layer planes and the formation of smaller particles of the inorganic matrix. Studies of organobentonite based on ADBC and DDAC and the initial bentonite indicate a sharp decrease in the effective particle diameter of the main fraction to 10 μm at the final stages of the technological process. The particle size reduction during modification is related to exfoliation processes, i.e., the delamination of the aluminosilicate structure into individual planes corresponding to the two-dimensional layers of the montmorillonite crystalline structure. Intercalation of montmorillonite, as with other layered matrices, causes a decrease in the particle size of the matrix, which is manifested as a low-angle halo on X-ray diffraction patterns of the intercalated structures. Thus, the conducted granulometric analysis of the transformed natural bentonite samples aimed at obtaining organomodified bentonite for paint and coating materials indicates the consistency of the obtained experimental results with the theoretical foundations of intercalation transformations of layered inorganic matrices. Infrared spectroscopy studies of the organomodified bentonite were performed to establish the mechanism of interaction between quaternary ammonium salts and the inorganic matrix.

The analysis of the IR spectra of the studied bentonite (red lump clay) and organomodified bentonite allowed the assignment of the main valence and deformation vibrations, which are presented in Table 1, 2.

TABLE I. WAVENUMBERS (N, CM⁻¹) OF ABSORPTION MAXIMA IN THE IR SPECTRA OF BENTONITE

Structural group	Stretching vibrations, cm⁻¹	Bending vibrations, cm⁻¹
Al-Al-OH	3628	
Al-Fe³⁺-OH	3590 – broad band shoulder at max 3435.1	875.3
Fe³⁺-Fe³⁺-Mg	3565 – broad band shoulder 3435.1	
H₂O	3435.1	1644.8; 1455.9
Si-O (tetrahedron)	1034.5	519
Si-O (amorphous)	1090; 1113; 1100 shoulder at 1034.5	875.3
Al-O-Si; Si-O-Si		519
Si-O-Al		466.4
-Si-O-Mg; Fe³⁺-O		435

The spectra did not exhibit bands corresponding to the vibrations of the structural group *α-SiO₂* (α-quartz), which allows concluding the absence of impurities such as kaolinite and α-quartz in the samples.

TABLE II. WAVENUMBERS (N, CM⁻¹) OF ABSORPTION MAXIMA IN IR SPECTRA OF ORGANOMODIFIED ADBC BENTONITE

Structural group	Stretching vibrations, cm⁻¹	Bending vibrations, cm⁻¹
Al-Al-OH	3628	
H₂O	3400.6	1668.3;1619.3
NR₄⁺	3335; 3030; 3064.1 shoulder at 3434.6	1485-1390; 1483.5;1468.5;1416.9
CH₃	2924.9 (vas); 2853.3 (vs)	1416.9 [ν(N-*CH₃*)]
-CH₂-		1416.9
-(CH₂)ₙ		728.7; 702.5
Si-O (tetrahedron)	1043.5	519
Si-O (amorphous)	1090; 1113; 1100 band shoulder 1034.5	875.3
Al-O-Si; Si-O-Si		519
Si-O-Al		467

In the IR spectrum of organomodified montmorillonite ADBC, vibrations of the CH_3 group attached to the nitrogen heteroatom appear as intense bands at 2853.3 cm^{-1} and 2924.9 cm^{-1}, corresponding to symmetric and asymmetric stretching vibrations of the methyl group; deformation vibrations of the methyl group are observed at 1416.9 cm^{-1} and 1378.3 cm^{-1}. These bands may be attributed to vibrations of the quaternary nitrogen atom. A characteristic feature indicating the formation of a new intercalation compound is the appearance of a new band at 3064.1 cm^{-1}, which corresponds to vibrations of the quaternary nitrogen atom. The emergence of new vibration bands at 702.5 cm^{-1} and 728.7 cm^{-1} in the spectrum of organomodified bentonite, associated with pendulum-type deformation vibrations of the (-CH_2-) groups, also indicates the occurrence of the intercalation reaction of ADBC into the interlayer spaces of montmorillonite. The nature of the changes observed in the infrared spectra of bentonite and organomodified bentonite allows us to conclude that the intermolecular interactions during intercalation are of a donor-acceptor type.

Differential thermal analysis (DTA) was performed on samples of bentonite, organomodified bentonite, free films based on pentaftalic oligomer, and pentaftalic oligomers modified with organobentonites. Curing by the oxy-polymerization mechanism was carried out under identical conditions at 80 °C for two hours using *Co*, *Mn*, *Ba* siccative as a catalyst. The content of organobentonites was 1%. Thermograms of the five studied samples were analyzed. TG curves in the temperature range of 100–107 °C showed a mass loss of 1.43 % and 1.58 % for Mnt-ADBC and Mnt-DDAC samples, respectively, which can be attributed to the loss of adsorbed water. The maxima on the DTG curves at these temperatures indicate the maximum rate of water loss from the organobentonite samples. Endothermic effects were observed on the DTA curves in the specified temperature range. The second peak on the DTG curves of the organobentonite-containing samples occurs at 285 °C and 273 °C for Mnt-ADBC and Mnt-DDAC samples, respectively, corresponding to the thermal degradation of the modifiers – quaternary ammonium salts. It is noteworthy that the thermal stability of the pentaftalic oligomer, which decomposes up to 260 °C, increases in compositions with organobentonites to 270–285 °C. Another important observation is the reduction in the total mass loss of the organobentonite-modified samples to 38 % and 33 %, which indicates the degradation of intercalation phases rather than surface-adsorbed molecules of ADBC and DDAC.

Thus, organobentonites exhibited a minimal amount of adsorbed water in the temperature range of 20–120 °C, approximately 1–2 %, compared to 7 % for the original clay. The presence of a significant endothermic effect in the temperature range of 200–350 °C, accompanied by a mass loss of 15–20 % for organobentonites, is associated with the degradation of the organic matter. The results of the thermal analysis indicate that, based on a comparative assessment of properties, the organobentonite based on Mnt-ADBC is preferable.

The study involved the application of organomodified bentonites, fillers, and lead-free anticorrosive pigments as components of pigmented coating materials.

Coating materials were developed based on the synthesized fillers, pigments, and organomodified bentonites. In the formulations, the pigment portion was replaced with the new bentonite fillers and pigments. The physicochemical properties of materials AU-122 I and facade water-dispersion paints are presented in Tables 3 and 4.

TABLE III. PHYSICOTECHNICAL PROPERTIES OF ENAMELS AND COATINGS WITH HIGH NON-VOLATILE CONTENT (TU U 20.3-34328580-001:2020)

Measured Parameter	AU-122 I Standard	Enamel with Mnt-ADBC
Conditional viscosity at (20±0.5)°C, s, not less than (DSTU ISO 2884-2)	60-160	150
Degree of dispersion, μm, not more than (DSTU ISO 1524)	45	45
Drying time to stage 3 at (20±2)°C, hours, not more than (DSTU ISO 9117-1)	3	2.5
Film strength under impact (U-1 device), cm, not less than (DSTU ISO 6272-2)	50	50
Film elasticity under bending, mm, not more than (DSTU ISO 1519)	1	1
Film adhesion, points, not more than (DSTU ISO 4624)	0	0
Hardness, conditional units (DSTU ISO 1522)	0.12	0.25
Resistance of the film to static water exposure at (20±5)°C, hours (DSTU ISO 2812-1, DSTU ISO 2812-2)	120	120
Resistance to mineral oil exposure at (20±5)°C, hours	120	120

According to the annexes of regulation TU U 20.3-34328580-001:2020, the primer-enamel AU-122 I is classified as a finishing coating material for both interior and exterior works on wooden and metal surfaces. In accordance with TU U, the VOC content should be ≤ 300 g/L. The new material contains VOCs at 294 g/L, whereas the standard formulation material contains 523 g/L. The purpose of this VOC limit is to reduce air pollution caused by VOCs' contribution to the formation of tropospheric ozone.

The AU-122 I primer enamel is a suspension of weather-resistant and anticorrosive pigments and fillers in solutions of organic oligomers, supplemented with additives to provide barrier and anticorrosive protection of surfaces and ensure optimal curing speed. The AU-122 I primer enamel is recommended for use in integrated systems of industrial coating applications on ferrous metal surfaces.

TABLE IV. PHYSICOTECHNICAL PROPERTIES OF WATER-DISPERSION FACADE PAINT (TU U 20.3-34328580-009:2021)

Measured parameter	Façade paint	Paint with Mnt-ADBC
Mass fraction of non-volatile substances, %	55-65	64
pH of the paint (DSTU ISO 19396-1, 19396-2)	8-9.5	9.2
Degree of grinding, μm, not more than (DSTU ISO 1524)	100	60
Drying time to stage 3 at 20°C, hours, not more than (DSTU ISO 9117-1)	1	1
Frost resistance of the paint, cycles (conditional units), not less than (TU U 20.3-34328580-009:2021)	5 (10)	5 (10)
Film resistance to static water exposure at (20 ± 2)°C, hours, not less than (DSTU ISO 2812-2)	24	24

A comparative analysis of the properties of materials and coatings, formulated using standard recipes and those

incorporating bleached bentonite fillers, organo-modified bentonites, and lead-free anticorrosive pigments obtained through the treatment of red granular bentonite clay with sodium phosphate, confirms the complete compliance of the AU-122 I primer enamel and the water-dispersion facade paint, as well as the coatings based on them, with the requirements of TU U 20.3-34328580-001:2020 and TU U 20.3-34328580-009:2021, respectively.

IV. Conclusion

Research was conducted to replace mineral acids during the activation stage of granular bentonite clay with sodium phosphate. This approach aims to prevent the formation of hydrolytic wastewater and reduce water consumption for washing away water-soluble salts.

It was established that the use of sodium phosphate, which hydrolyzes to an anion with a pH ≥ 7, enhances the environmental safety of the technology and facilitates the production of Na-form bentonite for obtaining organo-modified bentonite.

The organo-modification of Na-form bentonite was performed using ADBC and ADBC, which are quaternary ammonium salts.

New lead-free anticorrosive pigments were synthesized by the intercalation method, employing quaternary salts of triphenylmethane dyes.

The properties of fillers, organo-modified bentonites, and lead-free pigments were investigated using physicochemical analysis methods. The aim was to identify the purity of the initial substances, preserve the layered structure of aluminosilicate (montmorillonite), and study the interaction of montmorillonite with sodium phosphate, ADBC and ADBC, malachite green, and rhodamine C.

The paint and technical properties of the processed granular red bentonite clay – pigments and bentonite fillers – were determined. Their values fully comply with the technical, economic, and environmental requirements for pigments and fillers used in coatings and paints.

Coating materials, including AU-122 I primer enamel and water-dispersion facade paint, were developed. These materials utilize the synthesized pigments, fillers, and functional additives in the form of organobentonites.

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Investigation of the Influence of Technological Factors on the Properties of Slips and Samples of Alumina Ceramics from Super-reactive Alumina

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***Abstract*— High-density alumina ceramics are widely used in many fields of science and technology, including as electrical insulating elements of various electrophysical installations. The influence of technological factors on the properties of slips and samples of alumina ceramics made of super-reactive alumina grades A and B has been investigated. It is found that introducing a dispersing additive in the amount of 0.4 % in the slips of these aluminas reduces their dynamic viscosity to minimum values, improving the fluidity of the slips. Slips with the optimal amount of dispersing additive have high sedimentation stability. It is established that increasing the moisture content of slips from alumina grades A and B decreases their dynamic viscosity. The optimum moisture content (20 %) of slips intended for the pouring method of molding has been determined. It is established that the use of super-reactive alumina of A and B grades for the manufacture of products by slip casting into gypsum molds provides the production of high-density alumina ceramics. Based on the research results, the technology of high-density alumina ceramics by the slip casting method into gypsum molds was improved, and the raw material base for its production was expanded.**

Keywords*— *alumina ceramics, super-reactive alumina, slip casting, dispersing additive, dynamic viscosity, sedimentation stability, properties

I. Introduction

Specially dense alumina ceramics are widely used in various fields of science and technology due to their chemical inertness, high dielectric and strength properties. Alumina ceramic products are used as structural, high-temperature, electrical insulating elements and units of electrophysical equipment, with a service temperature of up to 1850 °C [1–3]. Specially dense alumina ceramics with content of Al_2O_3 99.4–99.7 %, Na_2O 0.005–0.01 %, MgO 0.12–0.15 %, SiO_2 0.04–0.1 %, open porosity close to zero have dielectric loss angle tangent at 20 °C 5×10^4, dielectric permittivity 9.8–9.9, breakdown voltage 18–22 kV/mm [1–4]. An increase in the Na_2O or SiO_2 amount in alumina ceramics leads to a several-fold decrease in the breakdown voltage and an increase in dielectric losses [1]. The particle size of the initial raw material, its MgO content, technological parameters of molding, and firing mode of products influence the crystal size, the presence of defects in the structure, and the density of ceramics, which affects its properties. Therefore, in the manufacture of high-density alumina ceramics, great attention is paid to the selection and preparation of raw materials, as well as the development of technological parameters of molding and firing.

JSC “URIR named after A. S. Berezhnoy” has carried out research and development work on the creation and improvement of alumina ceramic products technology for many years [1–3]. The technology of highly refractory alumina ceramic products, which are made by various methods, has been developed and introduced. The slip casting method into gypsum molds is the most widely used due to its simplicity, low cost of the equipment and tooling, and the possibility of manufacturing products of various shapes and sizes.

In recent years, companies producing raw materials have offered especially pure super-reactive alumina, which can be used without preliminary preparation, for the production of alumina ceramics. The Institute has developed technological parameters of slip casting of alumina ceramics with the content of $Al_2O_3 \geq 99.5$ %, MgO ≤ 0.1 %, $SiO_2 \leq 0.1$ %, $Fe_2O_3 \leq 0.05$ %, open porosity ≤ 0.8 % with the use of alumina with specific surface area 6–10 m^2/g, average particle size 0.2–1.5 microns and bulk density 900 kg/m^3 as a raw material. The compositions of aqueous alumina slips intended for the drain method of molding, used for the manufacture of products with a wall thickness of 1–3 mm, and the pour method — for the manufacture of products with a wall thickness of 4–5 mm and more, as well as products with a continuous section, have been determined [5–8].

At the same time, the research on super-reactive alumina of new grades, as well as the influence of technological factors on the properties of slips and samples of these alumina to determine the possibility of their use in the technology of high-density alumina ceramics by the method of slip casting, is of scientific and practical interest.

II. Research Methods

To perform the research, alumina grades A and B produced by a German company, a dispersing additive (aqueous solution of ammonium salt of acrylic copolymer), and distilled water were used as raw materials.

According to the manufacturer's data, alumina grades A and B belong to the class of super-reactive monomodal alumina with the content of Al_2O_3 99.7 %, α-Al_2O_3 98 %, specific surface area ~ 7 m^2/g, bulk density ~ 950 kg/m^3, and average particle size 1.0 micron for alumina grade A and 0.8 micron for alumina grade B.

Because many alumina ceramic products are made with the pouring method, the research was carried out using this casting method.

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The feasibility of introducing a dispersing additive into the slip to improve its rheological characteristics has been previously shown, and its optimum amount has been determined for the manufacture of products by the pouring method [6].

In this work, we investigated the effect of dispersing additives on the rheological properties (dynamic viscosity and sedimentation stability) of slips from alumina of new grades. For this purpose, aqueous slips from alumina grades A and B were prepared with a moisture content of 25 %, in which the additive was introduced in different amounts.

To study the dynamic viscosity of slips depending on their moisture content and storage time, slips were prepared by adding different amounts of alumina to distilled water. The slips containing the optimum amount of dispersing additive were used. The moisture content of the slips was 15 %, 20 %, and 25 %. All the slips were thoroughly mixed, filtered through a sieve, and then stored in a sealed container. The ambient temperature was 20 °C. The density of the slips was determined as the ratio of their mass to volume.

Moisture content of the slips was calculated by the formula:

$$W_{sl} = (g_0 - g_1) \times 100\ \% / g_0 \qquad (1)$$

where g_0 is the mass of the wet slip sample, g, and g_1 is the mass of the dried slip sample, g [9].

The dynamic viscosity of the slip was determined on a Brookfield LVDV-II + Pro digital viscometer at a spindle speed of 100 rpm. [10].

To determine the slip sedimentation stability, the slip was poured into a glass measuring cylinder, which was placed in a sealed container and kept for 14 days. During the slip storage, the volume of the separated liquid phase from the suspension was periodically noted visually. The coefficient of slip sedimentation stability was determined by the formula:

$$U = (V_{sl} - V) / V_{sl} \qquad (2)$$

where V_{sl} is the slip volume, cm^3, and V is the separated liquid phase volume, cm^3 [9].

Samples were made by the slip casting method into gypsum molds, then dried at 20 °C for 1–2 days, and at 110 ± 10 °C to constant weight. The samples were fired at a temperature of 1580 °C with a holding time of 6 hours in a batch gas-flame furnace.

The determination of the main property parameters was carried out according to Ukrainian/ISO standards. The cold crushing strength was determined according to DSTU ISO 10059-1:2018; the bending strength determination (three-point) was carried out according to DSTU ISO 5014:2018. The apparent density and open porosity of fired samples were determined according to DSTU ISO 5017:2014 (ISO 5017:1998, IDT).

III. Discussion of the Results

Determination results of the dynamic viscosity of the slips from alumina of A and B grades, depending on the content of dispersing additive, are presented in Table 1.

TABLE I. Dynamic Viscosity of Slips from Alumina Brands A and B with Different Content of Dispersing Additive

Alumina grades	Dispersing additive content, %						
	0	*0.1*	*0.2*	*0.3*	*0.4*	*0.5*	*0.6*
	Dynamic viscosity of slips, mPa·s						
A	1020	800	500	90	70	80	100
B	810	520	270	130	110	120	145

Table 1 shows that the dynamic viscosity of slip from alumina of grades A and B without a dispersing additive has maximum values. The influence of dispersing additive is based on its ability to envelop alumina particles, separating the aggregates formed by them, increasing the slip flowability. With the introduction of 0.1 % of the additive, the dynamic viscosity of the slips significantly decreases, but remains high enough and is 800 mPa·s for the slip of alumina grade A and 520 mPa·s for the slip of alumina grade B. When the content of dispersing additive is increased up to 0.4 %, the dynamic viscosity of the slips continues to decrease, reaching minimum values of 70 mPa·s and 110 mPa·s, respectively. Further increase in the content of dispersing additive in the slips leads to a slight increase in their dynamic viscosity, which is inexpedient, so the amount of dispersing additive of 0.4 % is optimal.

Determination results of the sedimentation stability coefficient of slips from alumina of grades A and B containing 0.4 % of dispersing additive are shown in Fig.

After storage in a sealed container for 14 days, the slips were mechanically stirred again, and their sedimentation stability coefficient was determined.

It follows from Fig. that at holding time of slip from alumina of grades A and B within 1 day, all the slips are characterized by a fairly high sedimentation stability, which with increasing storage time up to 5 days decreases slightly, and with increasing holding time up to 14 days decreases by 10–20 % due to the separation of the slip into solid and liquid phases. The slip from alumina of grade B has a slightly higher sedimentation stability than that from alumina A grade. It is probably due to the greater dispersibility of alumina B grade.

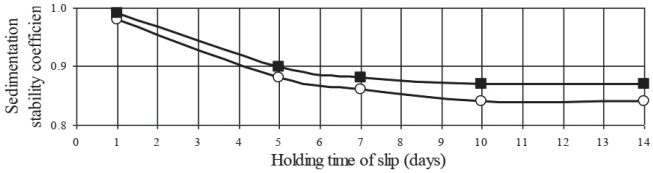


Fig. Sedimentation stability coefficient of slips from new brands of aluminas: ○ — slip from brand A alumina, ■ — slip from brand B alumina

It was found that after 1 day, the volume of a separated liquid phase and the coefficient of sedimentation stability corresponded to similar values before mixing. This fact indicates that the introduction of a dispersing additive in the optimal amount provides, in addition to satisfactory rheological characteristics, also the stability of the slip, which is important in the technology of products by slip

casting into gypsum molds, especially in the manufacture of ceramic products by the pouring method.

Thus, it was found that introducing 0.4 % of the dispersing additive in the slips from super-reactive alumina of grades A and B reduces their dynamic viscosity to minimum values and improves their fluidity. The slips from alumina of grades A and B with the optimal amount (0.4 %) of dispersing additive are characterized by high sedimentation stability.

The study also investigated the dynamic viscosity of alumina-based slips of A and B grades, depending on their moisture content and holding time (Table 2).

TABLE II. DYNAMIC VISCOSITY OF ALUMINA-BASED SLIPS CONTAINING 0.4 % OF DISPERSING ADDITIVE AS A FUNCTION OF MOISTURE CONTENT AND HOLDING TIME

Alumina grades	Holding time, hours					
	0	*4*	*8*	*24*	*48*	*56*
Moisture content 15%						
A	782	740	692	648	623	620
B	833	771	718	670	647	644
Moisture content 20%						
A	550	510	460	400	380	378
B	610	555	502	423	402	399
Moisture content 25%						
A	433	388	365	345	317	315
B	443	403	387	363	328	326

As follows from Table 2, the dynamic viscosity of all the slips begins to decrease almost immediately after their preparation, and this process continues until about 24 hours of holding time. At further increase in the holding time over 24–48 hours, the values of dynamic viscosity practically do not change. This character of change in dynamic viscosity indicates that even in freshly prepared slips, there are aggregates with a fragile structure of alumina particles, which, with increasing holding time, spontaneously collapse. As a result, there is a decrease in the dynamic viscosity of the slips, as it releases bound water in them and increases the number of dispersed particles that are not bound in aggregates. The optimum holding time of slips should be considered 24 hours, when their dynamic viscosity becomes minimal, and consequently, their fluidity improves. The fact that after holding time for more than 48 hours, the dynamic viscosity of the slips remains minimally low indicates their high manufacturability and the possibility of using them for the manufacture of products. The dynamic viscosity of alumina-based slips of grade B is slightly higher than that of alumina-based slips of grade A, which can be explained by the fact that it consists of more dispersed particles. Increasing the moisture content of slips from 15 % to 20 % leads to a significant decrease in their dynamic viscosity. Further increase in the moisture content of these slips up to 25 % contributes to decreasing their dynamic viscosity insignificantly. This is apparently because the slip moisture content of 20 % is sufficient for the spontaneous destruction of alumina particle aggregates, and an increase in the number of unbound particles covered with solvate shells, leading to a decrease in the dynamic viscosity of the slip.

Determination results of the main properties of alumina ceramic samples made of alumina-based slips of grades A and B with different moisture content before and after firing at a temperature of 1580 °C with a holding time of 6 hours are shown in Tables 3 and 4, respectively.

TABLE III. MAIN PROPERTIES OF SAMPLES MADE OF ALUMINA-BASED SLIPS OF GRADE A WITH DIFFERENT MOISTURE CONTENT BEFORE AND AFTER FIRING AT A TEMPERATURE OF 1580 °C

Sample properties	Moisture content, %		
	15	*20*	*25*
Properties of dried green			
Apparent density, g/cm^3	2.17	2.20	2.15
Cold crushing strength, N/mm^2	0.7	1.0	0.6
Properties of samples after firing at 1580 °C with a holding time of 6 hours			
Apparent density, g/cm^3	3.87	3.90	3.85
Open porosity, %	0.7	0.5	0.9
Change linear dimensions at firing (shrinkage), %	–23	–23	–25
Bending strength, N/mm^2	175	190	160

TABLE IV. MAIN PROPERTIES OF SAMPLES MADE OF ALUMINA-BASED SLIPS OF GRADE B WITH DIFFERENT MOISTURE CONTENT BEFORE AND AFTER FIRING AT A TEMPERATURE OF 1580 °C

Sample properties	Moisture content, %		
	15	*20*	*25*
Properties of dried green			
Apparent density, g/cm^3	2.16	2.18	2.13
Cold crushing strength, N/mm^2	0.6	0.8	0.5
Properties of samples after firing at 1580 °C with a holding time of 6 hours			
Apparent density, g/cm^3	3.83	3.86	3.80
Open porosity, %	1.0	0.7	1.2
Change linear dimensions at firing (shrinkage), %	–26	–26	–27
Bending strength, N/mm^2	180	200	170

As follows from Tables 3 and 4, the samples made of slips with a moisture content of 20 % have the highest property values. The samples made of slips with moisture content of 15 % have lower apparent density of green, which is explained by the rapid gain of the shard walls, which prevents the further impregnation of water from the slip into the gypsum mold, and, thus, the formed casting has a greater number of pores and is less dense and less strong. The samples made with 25 % moisture content also have a lower apparent density of the green due to the greater volume of water that must be absorbed from the slip into the gypsum mold, which leads to a decrease in the shard's gain speed, and the casting becomes less strong and dense. The optimum should be considered as the moisture content of 20 %. This is because the samples from such slips have the highest properties (open porosity less than 0.8 %, i.e., they are particularly dense). At the same time, the samples from alumina of grade A have a slightly higher apparent density and lower shrinkage at firing than the samples from alumina of grade B, with similar bending strength values. This can be explained by denser shard formation during casting the slip from alumina of grade A due to its lower dynamic viscosity and better fluidity, and further obtaining denser fired samples with fewer closed pores in the structure. Thus, both A and B grade alumina can be used in the technology

of alumina ceramics by slip casting into gypsum molds, but using alumina of grade A is preferable.

Thus, as a result of studies of the technological factors' influence on the properties of slips and samples of alumina ceramics from super-reactive alumina of grades A and B, the optimum amount of dispersing additive (0.4 %) introduced into the slips and the optimum moisture content of slips (20 %) were determined. The necessity of holding time freshly prepared slips to reduce their dynamic viscosity for 24 hours before molding products was shown. It was established that the use of super-reactive alumina of A and B grades for the manufacture of products by the method of slip casting into gypsum molds provides the obtaining of high-density alumina ceramics with open porosity less than 0.8 %, apparent density 3.86–3.90 g/cm^3, and bending strength 190–200 N/mm^2. It was shown that using alumina grade A compared to alumina grade B is preferable.

Detailed studies of the dielectric properties of the obtained ceramics will be carried out in further work. However, given their chemical composition, physical and mechanical properties, it can be confidently stated that their dielectric properties correspond to those of alumina ceramics obtained from traditional materials, namely dielectric loss angle tangent at 20 °C 5×10^4, dielectric permittivity 9.8–9.9, and breakdown voltage 18–22 kV/mm.

The research results show that the technology for producing high-density alumina ceramics using the slip casting method in gypsum molds has been improved, and the range of raw materials used in its production has been expanded.

The alumina ceramics obtained as a result of the research can be used as insulators for high-voltage equipment, heating element components (where high temperature resistance and electrical insulation are required), vacuum electronics components, in sensors operating at high temperatures, and for the manufacture of substrates and housings for electronic components, etc.

IV. Conclusions

The influence of technological factors on the properties of slips from super-reactive alumina of new grades A and B and alumina ceramics samples made from them has been investigated. The optimum quantity of dispersing additive (0.4 %) in slips from these aluminas, introduction of which leads to a reduction of the dynamic viscosity of slips to minimum values, has been determined. It has been established that alumina slips from alumina of grades A and B with the optimal dispersing additive amount are characterized by high sedimentation stability.

The necessity of holding time freshly prepared slips from alumina of grades A and B for 24 hours to reduce their dynamic viscosity before molding products was shown. It was found that increasing the moisture content of alumina slips decreases their dynamic viscosity. The optimum moisture content of slips (20 %) intended for the pouring method of molding was determined.

The main properties of samples from alumina of grades A and B before and after firing at a temperature of 1580 °C have been determined. It was established that the use of super-reactive alumina of grades A and B for the manufacture of products by the method of slip casting into gypsum molds provides the production of high-density alumina ceramics with open porosity less than 0.8 %, apparent density of 3.86–3.90 g/cm^3, and bending strength of 190–200 N/mm^2. It was shown that using alumina grade A compared to alumina grade B is preferable.

Based on the research results, the technology for producing high-density alumina ceramics using the slip casting method in gypsum molds has been improved, and the range of raw materials used in its production has been expanded.

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An Investigation into Autoencoder-Based Feature Extraction for Electromyographic Signal Classification

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***Abstract*—The work is devoted to the study of methods for automated feature extraction from electromyographic signals for their classification in diagnostic problem. The relevance of the study is due to the need to improve the accuracy of differential diagnostics of low back pain, where traditional spectral and statistical features do not always provide reliable classification. The aim of the study was to assess the applicability of autoencoders for extracting features from the electromyographic signals and their spectral power with subsequent classification with an ensemble of decision trees using the random forest method. Experiments were carried out with data from two independent sets of electromyographic signals from two diagnostic groups: dysfunctional and functional pain. Signal preprocessing, autoencoder training with hyperparameter selection, performance comparison for different architectures and types of input data were performed. The results showed that features extracted directly from raw electromyographic signals have higher discriminatory power than features from spectral power. The best results were achieved for the stacked sparse autoencoders architecture with 5 autoencoders (7 features), achieving accuracy, F1-score, and AUC up to 1.00. The proposed approach demonstrates robustness to class imbalance and confirms the potential of autoencoder feature extraction with subsequent classification with an ensemble of decision trees for clinical diagnostics, especially when working with raw electromyographic data that preserves rich nonlinear dynamic characteristics of the signal.**

Keywords—class imbalance, electromyography, feature extraction, power spectral density, random forest, signal classification, stacked sparse autoencoder

I. Introduction

A wide range of features derived from electromyographic (EMG) signals have been proposed for the diagnosis of back pain. These features can be used to classify EMG signals recorded from individuals with various clinical conditions.

Both time-domain and frequency-domain characteristics of EMG signals have been investigated [1] – [4]. Time-domain features include statistical and correlation-based metrics, while frequency-domain features comprise spectral descriptors such as median frequency, mean frequency, peak frequency, mean spectral power, bandwidth, lower and upper frequency bounds, and average power within narrow frequency bands.

Additionally, nonlinear analysis techniques have been employed to extract diagnostic features from EMG signals. These include entropy measures [5], fractal dimension, parameters of the phase portrait [6], [7], second-order difference plots [8] and higher-order statistics [9].

Many of these features exhibit statistically significant differences between diagnostic groups, including patients with back pain of various etiologies and healthy controls. This suggests that they are informative and potentially valuable for diagnostic and classification purposes. However, identifying a single parameter or a combination of EMG-derived features that allows for reliable and unambiguous classification of back pain conditions remains an open and relevant research challenge.

II. Task

In this study, the objective was to investigate the applicability of the autoencoder method for extracting features from raw electromyographic signals, as well as from their frequency-domain representation in the form of power spectral density (PSD). These features were intended to serve as the basis for classifying the original signals into diagnostic groups. To achieve this, it was necessary to determine an optimal configuration of autoencoder parameters and to evaluate the possibility of using the extracted features for classifying EMG signals into two diagnostic categories—dysfunctional and functional back pain—using an ensemble classifier based on decision trees (random forest method).

III. Methods

A. Autoencoders

When dealing with complex data in machine learning applications, dimensionality reduction is commonly achieved by extracting informative features from the original dataset. Feature extraction techniques can be categorized into linear and nonlinear approaches, both of which transform the original data into a more compact, lower-dimensional form.

Linear approaches – such as factor analysis, linear discriminant analysis, and principal component analysis – convert the input into a new feature set by computing linear combinations of the original variables. These methods are straightforward and computationally efficient, particularly effective when the intrinsic relationships in the data are approximately linear. However, they tend to lose information when the data involve nonlinear dependencies or feature interactions.

Nonlinear approaches, on the other hand, project the data into a higher-dimensional feature space, enabling the identification of patterns that may not be visible in the original space. Among the most powerful tools for extracting complex features from nonlinear data is deep learning. Deep learning models are capable of automatically learning feature representations from raw input, thus minimizing the reliance on manual feature engineering.

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Various deep learning-based nonlinear feature extraction methods have been developed, including convolutional neural networks (CNN), recurrent neural networks (RNN), and autoencoders (AE). Models such as CNNs and RNNs generally require large volumes of labeled data for training, and the training process itself is often computationally demanding, necessitating high-performance computing resources [10].

AE are neural networks trained to reproduce their input at the output. They are mainly used for unsupervised learning tasks, i.e., they do not require labeled data during training.

An autoencoder consists of two main components – an encoder and a decoder (Fig. 1). The encoder and decoder may have multiple layers; in the simplest case, each of them has only one layer.

The encoder maps the input vector $x \in \mathbb{R}^{D_x}$ into a compressed representation (features) $z \in \mathbb{R}^{D^{(1)}}$ using the following transformation:

$$z = h^{(1)}\left(W^{(1)}x + b^{(1)}\right), \quad (1)$$

where the superscript (1) indicates the first layer;

$h^{(1)}$: $\mathbb{R}^{D_x} \to \mathbb{R}^{D^{(1)}}$ is the activation (transfer) function of the encoder;

$W^{(1)} \in \mathbb{R}^{D^{(1)} \times D_x}$ is the weight matrix;

$b^{(1)} \in \mathbb{R}^{D^{(1)}}$ is the bias vector.

The decoder, given an encoded representation of z, reconstructs an approximate copy of the input vector x as follows:

$$\hat{x} = h^{(2)}\left(W^{(2)}z + b^{(2)}\right), \quad (2)$$

where the superscript (2) indicates the second layer;

$h^{(2)}$: $\mathbb{R}^{D^{(1)}} \to \mathbb{R}^{D_x}$ is a transfer function for the decoder;

$W^{(2)} \in \mathbb{R}^{D_x \times D^{(1)}}$ is the weight matrix;

$b^{(2)} \in \mathbb{R}^{D_x}$ is the bias vector.

The network is trained to minimize a loss function that measures the difference between the original data x and the reconstructed data $\hat{x}$, for example in the form of the mean squared error.

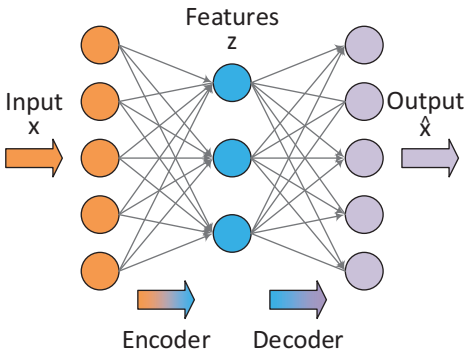


Fig. 1. Basic autoencoder diagram.

There are different AE architectures aimed at improving their efficiency for different tasks. Autoencoders that include different regularization methods are called regularized autoencoder (RAE).

A regularized autoencoder is a type of autoencoder in which additional constraints (regularization terms) on the model parameters or latent representation are introduced during training to improve the generalization ability of the network, avoid overfitting, and ensure that more informative and robust features are extracted.

Sparse autoencoders (SAE) use regularization techniques that force neurons in hidden layers to maintain a given level of sparseness, meaning that only a subset of neurons in the hidden layer are activated for an input example. This constraint forces the network to use individual hidden layer nodes for specific features of the input data, reducing redundancy in the input information.

The goal of sparse AE is to minimize the difference between the input and reconstructed data while adhering to the sparsity constraints of the hidden representation. The loss function of a sparse autoencoder consists of two components: reconstruction loss and sparsity loss.

Encouraging the sparsity of the autoencoder is possible by adding a regularizer to the loss function. This regularizer is a function of the average output activation value of the neuron. The average output activation measure of neuron i is defined as:

$$\hat{\rho}_i = \frac{1}{n}\sum_{j=1}^{n} z_i^{(1)}\left(x_j\right) = \frac{1}{n}\sum_{j=1}^{n} h\left(w_i^{(1)T}x_j + b_i^{(1)}\right), \quad (3)$$

where n is the total number of training examples;

x_j is the j-th training example;

$w_i^{(1)T}$ is the i-th row of the weight matrix $W^{(1)}$;

$b_i^{(1)}$ is the i-th input of the bias vector $b^{(1)}$.

A neuron is considered to be "activated" if its output activation value is high. A low output activation value indicates that the neuron in the hidden layer responds to only a small number of training examples. Adding a term to the loss function that constrains the values of $\hat{\rho}_i$ to be low encourages the autoencoder to learn a representation in which each hidden neuron is activated by only a limited number of training samples. In other words, each neuron becomes specialized, responding to a specific feature that is present only in a small subset of the training data.

The loss function minimized during training includes three terms:

$$E = \frac{1}{N}\sum_{n=1}^{N}\sum_{k=1}^{K}(x_{kn} - \hat{x}_{kn})^2 + \lambda\Omega_{weights} + \beta\Omega_{sparsity}, \quad (4)$$

where λ is the L_2 regularization coefficient;

β is the sparsity weighting coefficient.

The first term the loss function E is the mean squared reconstruction error (the difference between the input and the output), i.e., the network learns to reproduce the input. The second term is weight regularization, which helps prevent overfitting. The third term is a sparsity penalty (penalizing "overactive" neurons) to regularize the network's sparseness.

During the training of a sparse autoencoder, it is possible to reduce the sparsity regularization term by increasing the weight values $w^{(i)}$ and decreasing the values $z^{(1)}$. Adding a regularization term to the weights in the loss function prevents this behavior. This term is called the L_2 regularization term and is defined as:

$$\Omega_{weights} = \frac{1}{2}\sum_{l=1}^{L}\sum_{j=1}^{n_l}\sum_{i=1}^{k_l}\left(w_{ji}^{l}\right)^2, \quad (5)$$

where L is the number of hidden layers;
n_l is the output size of layer l;
k_l is the input size of layer l.

The regularization term L_2 is the sum of the squares of the elements of the weight matrices for each layer.

The sparsity regularizer imposes a constraint on the sparsity of the hidden layer's output. Sparsity can be encouraged by adding a regularization term that takes a large value when the average activation value $\hat{\rho}_i$ of neuron i deviates from its desired value ρ. One such sparsity regularization term is the Kullback-Leibler (KL) divergence, defined as:

$$\Omega_{sparsity} = \sum_{i=1}^{D^{(1)}} \mathrm{KL}(\rho\|\hat{\rho}_i) =$$

$$= \sum_{i=1}^{D^{(1)}} \left[\rho \log\left(\frac{\rho}{\hat{\rho}_i}\right) + (1-\rho)\log\left(\frac{1-\rho}{1-\hat{\rho}_i}\right)\right]. \quad (6)$$

The Kullback–Leibler divergence is a function used to measure how different two distributions are. In this case, it takes the value zero when ρ and $\hat{\rho}_i$ are equal, and increases as they diverge from each other. Minimizing the loss function forces this term to be small, which means that ρ and $\hat{\rho}_i$ must remain close to each other.

The results of the studies in [11] showed that the autoencoder significantly outperforms conventional machine learning models in the ability to detect anomalies in ECG data. However, the authors emphasize the sensitivity of autoencoders to the choice of hyperparameters and the need for large training datasets to achieve adequate performance. In [12], a stacked autoencoder was used to extract useful features from the surface electromyography signal to identify target finger movements from spectrogram and wavelet data, and in [13], the performance of stacked sparse autoencoders was evaluated to improve the myoelectric control of wearable prostheses based on surface and intramuscular electromyography records. The performance of stacked sparse AE showed better results compared to the linear discriminant analysis-based classifier. In [14], an EMG-based autoencoder method was tested for myocontrol. The performance of the autoencoder (AE) was compared with industrial state-of-the-art (SOA) control, and the results demonstrated that AE outperforms SOA. Results in [15] highlight the potential of AE to extract features from the power spectral density of pre-processed electroencephalography signals, which are subsequently classified by radial basis function neural network for early and accurate detection of Parkinson's disease.

Thus, the features extracted by SAE reduce the dimensionality of the data and can be used in combination with classifiers.

B. Random Forests

Random Forest (RF) is an ensemble classification method based on constructing a set of decision trees and combining their predictions to improve the accuracy and robustness of the model. The method was proposed by L. Breiman [16] and belongs to the family of bagging algorithms (bootstrap aggregating).

The main stages of classification by random forests include: the formation of training subsets, training of decision trees and voting (ensembling).

At the stage of formation of training subsets, N bootstrap samples are randomly generated (with replacement) from the original training set. The bootstrap method is a statistical technique in which multiple new subsets are randomly formed from the original training data with replacement. If the original training set contains N objects, then each bootstrap sample also contains N objects, but may include repeated elements, since each object is selected randomly and independently with the possibility of being selected more than once. This method is widely used in ensemble algorithms to improve robustness and reduce overfitting.

One decision tree is built for each bootstrap sample. During the tree training stage, each tree is trained on its corresponding bootstrap subset. This process introduces diversity among the trees. In the random forest method, the diversity between trees is further enhanced by applying feature subsampling at each node split, unlike standard decision trees, where all features are considered at each split to find the best one. Random forests consider only a randomly selected subset of features at each node and the best split is then determined from this subset. This technique reduces inter-tree correlation and increases the ensemble's robustness to overfitting.

At the final stage of classification, prediction is performed using majority voting. The majority voting process is a method of aggregating the decisions of multiple models (in an ensemble), in which the final prediction is determined based on the majority of votes from the individual models.

Let $\{h_1(x), h_2(x), \ldots, h_N(x)\}$ be the predictions of individual trees for the input vector x, then the final classifier is defined as:

$$H(x) = \mathrm{mode}(\{h_i(x)\}_{i=1}^{N}), \quad (7)$$

where $\mathrm{mode}(\cdot)$ is the function that returns the most frequent value (class).

The advantages of the random forest classifier are high robustness to overfitting, effectiveness on imbalanced datasets (with the use of class weights), and the ability to evaluate feature importance.

Majority voting makes the final decision more robust to errors of individual models by enabling mutual error compensation across the ensemble. Majority voting is a fundamental principle of ensemble learning, which is used in methods such as Random Forest, Bagging, and certain neural network ensembles.

In [17], the authors demonstrated the potential of using the random forest method for ECG signal classification to improve diagnostic accuracy. In [18], the random forest algorithm was used as a classifier for diagnosing cardiac arrhythmia based on statistical features extracted from the wavelet transform coefficients of ECG signals. The study showed that the RF classifier achieved better results compared to other methods. In [19], the reliable performance of the random forest algorithm was demonstrated for classifying patients with bipolar disorder and healthy individuals based on electroencephalogram data.

Thus, the random forest method was chosen as the classifier for the features extracted by the autoencoder.

IV. Experiments and Results

The original data for the study consisted of two sets of electromyographic signals recorded from the long extensor muscle of the trunk of the lumbar spine under the same conditions at different time periods. The EMG data were obtained in collaboration with the Department of Pathophysiology, Anesthesiology and Functional Diagnostics of the Sytenko Institute of Spine and Joint Pathology National Academy of Medical Sciences of Ukraine.

Each of the two data sets contains EMG signals of two diagnostic groups: 1 – conditionally healthy subjects with complaints of back pain, i.e., subjects without identified organic pathology (the so-called dysfunctional pain); 2 – subjects with back pain who have degenerative diseases of the spine (functional pain). In the first dataset, the number of signals was 46 (group 1) and 235 (group 2); in the second dataset, 38 (group 1) and 72 (group 2).

Pre-processing of the signals included filtering in the 10-500 Hz band and downsampling to 1024 Hz. The signals were cut to a 1-minute record length.

This study investigated the possibility of combining an autoencoder for extracting parameters from EMG signals with subsequent classification using the random forest method to determine group membership between two classes.

The signals from the first set were used to train the autoencoder and classifier. The obtained models were tested on the second data set.

A sparse autoencoder was chosen for the studies. As a result of the experiments, an optimal set of autoencoder hyperparameters was selected to obtain the best classification result. The studies were conducted both on the raw EMG signal and on its power spectral density (PSD), and in both cases the input signals were normalized. PSD was estimated using Welch's method. The original EMG signal was segmented into overlapping windows of length 256 samples, with 50% overlap (i.e., 128 samples). Each segment was multiplied by a Hamming window function, and the discrete Fourier transform was computed using a 512-point fast Fourier transform. Thus, the length of the autoencoder input vector is 61440 for the raw signal and 257 for the PSD. Different combinations of autoencoder parameters and its architecture were tested.

The logistic sigmoid function was used as the transfer function for the encoder and decoder.

The best results were achieved using the stacked autoencoder architecture (Fig. 2), consisting of five sequentially applied autoencoders, as shown in Fig. 2, and also with the addition of the sixth autoencoder, which reduces the number of features from 7 to 3. The stacked architecture implies sequential application of the autoencoders with a gradual reduction in the dimensionality of the feature space. The mean squared error was used as the training loss function with the addition of L_2-regularization with $\lambda = 0.001$, sparse constraint with the target average activation $\rho = 0.05$ (i.e., each hidden neuron should be active in about 5% of cases) and a sparsity coefficient $\beta = 4$ (the coefficient for the Kullback–Leibler divergence).

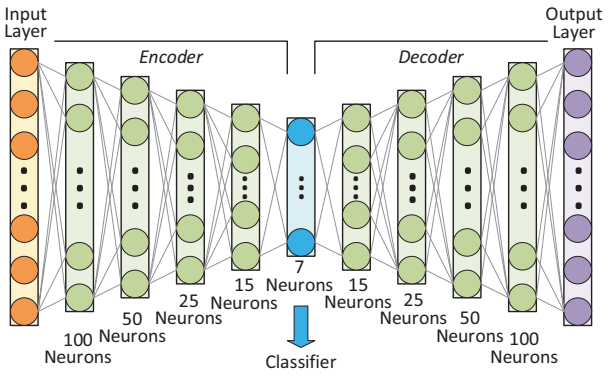


Fig. 2. Stacked autoencoder for feature extraction for subsequent classification.

An ensemble of 100 decision trees constructed using the random forest method was used as a classifier. Due to the significant class imbalance in the training dataset (46 samples in group 1 and 235 in group 2), class weighting was applied during classifier training to compensate for the imbalance. It was experimentally determined that the best classification performance was achieved with a class weight ratio of 3:1, assigning a higher weight to the minority class (group 1). This increases the sensitivity of the model to the less represented class, preventing it from being ignored during training.

The classification results on the test sample (accuracy, F1-score, AUC) are presented in Table 1.

TABLE I. Classification results using the random forest method for features extracted by the autoencoder

Input signal	AE Architecture	Accuracy	F1-score	AUC
Raw EMG signal	5 AE 7 features	1.00	1.00	1.00
	6 AE 3 features	0.97	0.96	1.00
PSD of EMG signal	5 AE 7 features	0.98	0.97	1.00
	6 AE 3 features	0.98	0.97	1.00

The results presented in Table 1 demonstrate the effectiveness of autoencoders (AE) for feature extraction from electromyography signals and subsequent classification with an ensemble classifier based on the Random Forest method.

When extracting features from PSD, classification based on 7 features obtained using a sequence of 5 autoencoders achieves high accuracy (0.98), F1-score is 0.97, and AUC is 1.00, which indicates high sensitivity and specificity of the model. A similar level of metrics is achieved when using a deeper architecture (6 autoencoders, 3 features), which indicates high information content of the extracted features, despite their small number.

It is noteworthy that the highest performance was achieved when raw EMG signals were used for feature extraction. The architecture with 5 autoencoders and 7 features produced perfect results. Even with strong compression down to 3 features (using 6 autoencoders), the model maintained high performance, confirming the robustness of the extracted features to dimensionality reduction.

Comparison of two types of input signals shows that features extracted directly from raw EMG signals have higher discriminatory power compared to features based on spectral representation (EMG PSD). High classification accuracy

achieved when using features extracted from raw EMG signals is probably due to the fact that the autoencoder receives significantly more information (61440 inputs) than in the case of spectral representation (257 inputs). Such high dimensionality allows the model to detect more complex and diverse patterns in the data, including temporal and nonlinear characteristics of the signal, which can be lost during preliminary spectral transformation. Using spectral-temporal representation of signals seems promising, since it can better reflect the nonlinear-dynamic nature of the EMG signal. At the same time, the main disadvantage of using raw signals is significantly higher computational costs at the stages of autoencoder training compared to processing spectral features. However, this factor does not affect the efficiency or speed of application of already trained models when they are used in classification problems, so it is rather an advantage, since it does not require pre-processing of data for their classification.

Thus, the presented results confirm the effectiveness of autoencoders to extract informative features from EMG signals and demonstrate the potential of using raw data to achieve high classification accuracy when powerful nonlinear dimensionality reduction methods are applied.

CONCLUSION

This study demonstrates the effectiveness of using stacked sparse autoencoders for extracting informative features from electromyographic (EMG) signals, followed by classification using a Random Forest ensemble. Both raw EMG signals and their power spectral density representations were evaluated as inputs to the autoencoder-based feature extractor.

The highest classification performance was achieved when features were extracted directly from raw EMG signals, with a compressed latent representation of only 3–7 features yielding accuracy, F1-score, and AUC values up to 1.00. Comparable, though slightly lower, performance was also observed when using spectral features. These findings suggest that raw EMG signals retain richer nonlinear-dynamic information beneficial for feature learning.

The results confirm that combining unsupervised feature extraction with ensemble learning provides a robust solution for EMG signal classification, even in the presence of class imbalance, and holds strong potential for diagnostic applications.

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Cybersecurity Challenges and Post-Quantum Protection in Cardiovascular Monitoring Devices

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***Abstract* — This paper explores cybersecurity for Cardiovascular Monitoring Devices (CMDs) in IoT, addressing vulnerabilities in wireless networks like Wi-Fi and 5G. It compares traditional methods (AES-256, TLS 1.3) with post-quantum algorithms (CRYSTALS-Kyber, Niederreiter), emphasizing their quantum resistance. A hybrid approach using Quasi-Cyclic LDPC codes with Niederreiter is proposed to reduce key sizes, enhancing security and efficiency for CMDs against emerging quantum threats.Keywords— STM32, GD32, Glitch, Cold Boot Stepping, Electromagnetic Analysis .**

Keywords— Cardiovascular monitoring devices; cybersecurity; IoT; wearable technology; implantable devices; post-quantum cryptography; CRYSTALS-Kyber; Niederreiter; QC-LDPC; wireless communication; data protection; CMD security.

I. Introduction

Cardiovascular diseases (CVDs) are the leading cause of death globally, accounting for approximately 17.9 million deaths per year, or 31% of all global deaths, according to the World Health Organization. This growing burden is particularly pronounced among the aging population, which is projected to reach 17% of the world's total by 2050. In response, the development and deployment of Cardiovascular Monitoring Devices (CMDs)—both wearable and implantable—have become essential for continuous health monitoring and early disease detection.

Modern CMDs utilize technologies such as electrocardiographs (ECGs), implantable cardioverter-defibrillators (ICDs), and AI-integrated smartwatches to provide continuous health data. These devices are increasingly interconnected through IoT frameworks, cloud analytics platforms, and wireless communication protocols such as Bluetooth, Wi-Fi, ZigBee, and 5G. This connectivity enables real-time diagnostics and personalized treatment, yet also introduces serious cybersecurity risks.

Cyberattacks on CMDs, including data breaches and manipulation of device function, pose threats to patient safety, data privacy, and clinical reliability. Legacy encryption methods like AES-256 and TLS 1.3 are effective for many applications but may become obsolete with the rise of quantum computing. Post-quantum cryptographic algorithms, particularly CRYSTALS-Kyber and Niederreiter, offer enhanced protection but require optimization to meet the stringent energy and latency constraints of CMDs.

This paper provides a comprehensive review of CMD technologies, wireless protocols, cyber threats, and cryptographic defenses. It emphasizes the potential of integrating QC-LDPC codes into the Niederreiter cryptosystem to develop lightweight and quantum-resistant encryption solutions for future-proof cardiovascular health monitoring.

II. Main Research

Cardiovascular Monitoring Devices (CMDs) include implantable and wearable technologies such as electrocardiographs, implantable cardioverter defibrillators (ICDs), cardiac monitors, and smartwatches with heart rate monitoring capabilities. These devices provide continuous monitoring of cardiovascular activity, which is critical for early diagnosis and prevention of cardiovascular diseases (CVDs). According to the World Health Organization, CVDs cause 17.9 million deaths annually, accounting for 31% of all deaths worldwide [1], and the aging population (by 2050, the proportion of people aged 65+ will increase to 17% of the world's population [2]) is significantly contributing to the demand for CMDs. The elderly population is more likely to have CVDs, which requires continuous monitoring of heart rate and other parameters. Increasing awareness of CVD prevention among younger populations is driving the adoption of wearable devices that do not track heart rate or physical activity. Awareness campaigns supported by organizations such as the WHO and device manufacturers are helping to popularize CMD [1]. The global market for these devices is showing steady growth driven by technological innovation, increased health awareness, and economic factors.

Technological advances in CMD include the development of energy-efficient microprocessors, improved biosensors, and the introduction of wireless technologies for rapid data transmission. Implantable devices use minimally invasive technologies that reduce the risks of infections and

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complications. Wearable devices integrate AI to analyze ECG data and predict abnormalities. Cloud platforms such as AWS HealthLake provide processing of large volumes of data, allowing doctors to obtain real-time analytics and tailor treatment to the patient's needs [3].

The global CMD market was valued at $28.5 billion in 2023 and is projected to reach $42.7 billion by 2030, at a compound annual growth rate (CAGR) of 6.9% [4]. The main drivers of growth are the demand for wearable devices, which accounted for 35% of the market in 2023 and are projected to grow to 48% by 2030 due to falling prices and widespread availability. Research and development (R&D) investments in medical technology annually account for approximately 10% of the total market, which is driving the emergence of new products with improved functions, such as biosensors with measurement accuracy of up to 95% [5].

TABLE I. KEY FINANCIAL INDICATORS OF THE CMD MARKET (2023–2030)

Indicator	Value 2023	Forecast 2030	CAGR, %	Main growth factor
Market size, $ bil.	28,5	42,7	6,9	Demand for wearable devices
Investments in R&D, $ bil.	2,85	4,27	7,0	Innovations in AI
Share of wearable devices, %	35	48	8,5	Price reduction
Share of implanted CMDs, %	40	36	5,2	Technological improvements
Marketing expenses, $ bil.	1,42	2,13	6,8	Awareness campaigns

Technological progress in CMDs is largely dependent on microprocessor architectures that provide data processing, energy efficiency, and integration with IoT. Modern CMDs use microprocessors with architectures such as ARM Cortex-M, RISC-V, MIPS, x86, and AVR, which optimize performance and power consumption. ARM Cortex-M is the leader in wearable devices due to its low power consumption (0.1–10 mW). RISC-V is gaining popularity due to its open architecture, which reduces licensing costs, and is suitable for implanted devices. MIPS and x86 are used in more complex stationary systems and cloud platforms, but have higher power requirements. AVR is used in simple wearable devices with low cost but limited performance [6].

TABLE II. COMPARISON OF MICROPROCESSOR ARCHITECTURES IN CMD

Architecture	Performance, MIPS	Power consumption, mW	Core size, nm	Application
ARM Cortex-M	100–1000	0,1–10	28–65	Wearable devices
RISC-V	50–800	0,2–15	22–65	Implantable devices
MIPS	200–1500	5–50	40–90	Stationary devices
x86	1000–5000	50–200	14–45	Cloud platforms
AVR	20–200	0,5–20	65–130	Simple wearable devices

Microprocessors enable ECG data processing, AI-based arrhythmia prediction, and wireless device integration. For example, the ARM Cortex-M in the Fitbit Versa enables energy-efficient sensor data processing, while x86 is used in cloud platforms for analyzing large amounts of data [7]. However, the use of IoT in CMD poses cybersecurity challenges, as each connected device is a potential point of vulnerability.

Wireless networks are the basis for data transmission in CMD devices. These technologies provide fast and convenient data exchange between devices and medical systems, allowing doctors to receive information about heart rhythm, electrocardiograms (ECGs), and other parameters in real time. However, their openness makes them vulnerable to cyberattacks, which threaten the confidentiality, integrity, and availability of data [8]. In the medical field, these threats are particularly dangerous, as they can lead to false diagnoses, device manipulation, or a threat to the patient's life.

Wireless technologies for CMD have different characteristics that affect their security. Wi-Fi (802.11ac) provides high bandwidth (up to 1300 Mbps), but high power consumption limits its use in implanted devices. Bluetooth LE (5.0) is energy efficient, but has a limited range (10–100 m) and weak encryption in older versions. ZigBee supports mesh networks with low power consumption, but speeds up to 250 Kbps are insufficient for large data. LTE and 5G offer high speed (up to 100 Mbps and 10 Gbps) and low latency, but depend on the operator infrastructure. LoRa provides a range of up to 15 km and ultra-low power consumption, but speeds up to 50 Kbps limit it for real-time monitoring [9].

The security level of the technologies varies: Wi-Fi with WPA3 is vulnerable to attacks on open networks, BLE 5.0 uses AES-128, but older versions are weaker. ZigBee supports AES-128 and mesh architecture, LTE and 5G use AES-256 and TLS 1.3, but are vulnerable to attacks on base stations. LoRa with AES-128 is suitable for low-frequency monitoring, but not for big data [5]. The choice of technology depends on the type of CMD, power requirements, and usage scenario.

Wireless networks used in CMD are subject to the following main types of attacks: data interception, man-in-the-middle (MITM) attacks, denial-of-service (DoS) attacks, firmware manipulation, and spoofing. Each attack has its own characteristics, potential consequences, and level of difficulty to implement. The level of security depends on the security measures applied, such as encryption, authentication, and network monitoring. Table 3 provides a comparison of the attacks and the corresponding security measures.

TABLE III. COMPARISON OF ATTACKS ON WIRELESS NETWORKS CMD AND SECURITY LEVEL

Attack type	Description	Potential consequences	Implementation complexity	Protection measures
Data interception	Obtaining data through weak encryption	ECG leak, medical history	Low	AES-256, TLS 1.3
Man in the Middle (MITM)	Interception and modification of communication	Incorrect commands, changing settings	Medium	WPA3, two-factor authentication
Denial of service	Network overload	Stop monitoring	Medium	IDS/IPS, traffic limiting
Firmware manipulation	Software change	Uncontrolled device operation	High	Digital signature, firmware update
Spoofing	Device identity spoofing	Data access, fake commands	Medium	Certificates, biometric authentication

Thus, analyzing the types of attacks, it can be noted that in Eavesdropping, attackers use weak encryption or insecure connections to obtain medical data. Vulnerabilities in Bluetooth connections allow interception of ECGs within a radius of up to 100 m, as was shown in 2019 [10]. Using AES-256 and TLS 1.3 significantly reduces the risk, but requires computational resources. In Man-in-the-Middle (MITM) attacks, the attacker intercepts the communication between the device and the server, modifying data or sending erroneous commands. For example, a MITM attack on a pacemaker can change the stimulation parameters, which is life-threatening. WPA3 and two-factor authentication reduce the risk, but vulnerabilities in Wi-Fi remain [11]. Denial-of-Service (DoS) attacks contribute to network congestion and block data transmission, stopping monitoring. For example, a DoS attack on a 5G network can delay the transmission of arrhythmia data. IDS/IPS systems and traffic throttling can help, but false positives can complicate the operation [12]. Firmware manipulation leads to changes in the CMD software. In 2011, researcher Barnaby Jack showed hacking insulin pumps via Bluetooth, which is relevant for CMD [13]. Using Spoofing, attackers forge the identity of the device, gaining access to data. Spoofing in a ZigBee network can imitate a legitimate device [14]. These attacks pose a serious threat to CMD, since a delay or error in data transmission can have fatal consequences. In 2023, 68% of cyberattacks in the healthcare sector were directed at IoT devices [14].

Based on this, it can be concluded that data protection in wireless networks for CMD faces several challenges.

– First, the limited computing resources of implanted devices make it difficult to use complex encryption algorithms (a pacemaker with a 5–10 year battery cannot support continuous AES-256 encryption without reducing its service life [15]).

– Second, the diversity of wireless technologies makes it difficult to standardize security protocols.

– Third, data transmission latency is critical for CMD: encryption can increase the latency in 5G networks (less than 1 ms), which is unacceptable for real-time monitoring [16].

– Fourth, vulnerabilities in firmware and outdated software remain a problem. Many CMDs do not support regular updates due to resource constraints.

– Fifth, the human factor plays a role: users may connect to unsecured networks or ignore updates, increasing the risk of attacks; 42% of data breaches in 2023 were caused by human error [17].

Wireless networks such as Wi-Fi, Bluetooth Low Energy, ZigBee, LTE, 5G, and LoRa, used in CMD, require data protection through encryption, authentication, and network monitoring. Modern methods ensure data confidentiality, integrity, and availability while minimizing power consumption and transmission latency, which is important for resource-constrained implanted devices [19]. Traditional methods such as AES-256 encryption, TLS 1.3 protocol, multi-factor authentication (MFA), and intrusion detection and prevention systems (IDS/IPS) remain relevant. However, post-quantum algorithms, such as CRYSTALS-Kyber (ML-KEM) and Niederreiter, are more modern and provide higher security for IoT, including CMD, due to their resistance to quantum attacks that threaten traditional methods [20]. AES-256 and TLS 1.3 are energy efficient, while IDS/IPS, CRYSTALS-Kyber, and Niederreiter require more resources, which may complicate their use in implanted devices. MFA increases security, but requires additional modules, which is a challenge for compact devices.

To evaluate the methods, performance indicators (probability of protection from attacks, %), implementation complexity (man-hours), power consumption (mW), processing latency (ms), and use cases were used. The power consumption of encryption algorithms is estimated using the corresponding formula 1:

$$E_{crypt} = N\left(E_{enec} + E_{dec}\right) \quad (1)$$

N_{crypt} – Number of encryption cycles;

E_{enec} – Energy for encryption of one block;

E_{dec} – Energy for decryption of one block.

The probability of successful protection against attacks depends on cryptographic strength, which can also be quantified, see Formula 2.

$$P_{\text{sec}} = 1 - \frac{1}{2^{k_{eff}}} \quad (2)$$

where: k_{eff} – the effective key length, taking into account resistance to quantum attacks;

Psec - protection probability (0 to 1).

TABLE IV. COMPARISON OF CYBERSECURITY METHODS FOR CMD WITH NIST COMPETITION RESULTS

Method	Efficiency, %	Implementation complexity, man-hours.	Power consumption, mW	Processing delay, ms	Example of use
AES-256	95	100	0.5	0.2	Data encryption in IoT
TLS 1.3	92	300	2.0	0.5	Data transmission in 5G
MFA	90	500	5.0	1.0	Access to servers
IDS/IPS	85	800	10.0	2.0	Network monitoring
CRYSTALS-Kyber (ML-KEM)	98	1200	15.0	3.0	Encryption in 6G/IoT
Falcon	96	1300	18.0	3.5	Encryption in IoT
HQC	97	1250	16.0	3.2	Backup encryption in IoT
McEliece	94	1400	20.0	4.0	Encryption in IoT (backup)
Niederreiter	95	1350	19.0	3.8	Encryption in IoT

The effective key length (bits), which takes into account resistance to quantum attacks.

Post-quantum algorithms such as CRYSTALS-Kyber, Falcon, HQC, McEliece, and Niederreiter are compared to traditional methods, demonstrating superiority in protecting IoT from quantum threats.

CRYSTALS-Kyber and HQC are distinguished by the highest efficiency due to their resistance to the quantum algorithms of Shor and Grover, providing reliable protection of the IoT against future threats [21]. Niederreiter, although less optimized, also outperforms traditional methods in terms of resistance (see Formula 2 in the formula file) [20]. The implementation of post-quantum algorithms requires significant effort, in particular for Niederreiter due to larger keys [11]. NIST (2024) tests on IoT devices show that CRYSTALS-Kyber is optimal in terms of power consumption and latency, while Niederreiter consumes more resources (see Formula 1 in the formula file) [25]. CRYSTALS-Kyber is standardized as the primary KEM (FIPS 203, 2024), Falcon and HQC are fallbacks, and McEliece and Niederreiter are alternatives for IoT [10, 22].

AES-256 is energy efficient (see Formula 1 in the formula file) and fast, making it suitable for implantable CMDs such as Medtronic pacemakers, due to its 0.2 ms real-time latency [23]. TLS 1.3 protects against MITM attacks and is effective for wearables such as Fitbit Versa, due to its low power consumption and 0.5 ms latency [24]. MFA is suitable for cloud platforms such as AWS HealthLake, but its power consumption limits its use in implantable CMDs [26]. IDS/IPS are effective for servers, but their power consumption and 2.0 ms latency create difficulties for devices [25].

CRYSTALS-Kyber, as a modern post-quantum algorithm, provides higher IoT security due to its resistance to quantum attacks, but its power consumption of 15.0 mW (see Formula 1 in the formula file) makes it difficult to use in implantable CMDs [24]. Falcon and HQC have similar characteristics but require significant resources [20, 24]. McEliece and Niederreiter outperform traditional methods in protecting IoT from quantum threats, but large keys and power consumption (19.0 mW for Niederreiter) limit their use in CMD (see Formula 3) [20, 23].

$$T_{\text{key}} = \frac{S_{key}}{R_{band}} \tag{3}$$

where: T_{key} – key transmission time (s).

S_{key} – key size (bits).

R_{band} – network bandwidth (bits/s).

CRYSTALS-Kyber is a leading post-quantum algorithm for IoT, including CMD. Its resistance to Shor and Grover quantum attacks makes it more advanced than AES-256, which could be cracked by quantum computers by 2030 [10]. Compact keys (800–1568 bytes) simplify exchange in 5G/6G networks compared to Niederreiter (100–500 kB) [23]. Latency of 3.0 ms is suitable for wearable IoT devices [24]. Compatibility with 6G networks (latency of 0.1 ms, speed of 1 Tbps) ensures relevance [25]. Support from IBM and Cloudflare increases trust after incidents, such as the St. Jude Medical incident in 2017 [19]. Power consumption of 15.0 mW makes it difficult to use in implanted CMDs [24].

Niederreiter, as a post-quantum algorithm, provides higher IoT security compared to traditional methods due to its resistance to quantum attacks [23]. Smaller keys than McEliece make it easier to use in IoT [24]. The simpler structure compared to Falcon facilitates implementation [19]. However, its power consumption of 19.0 mW and latency of 3.8 ms make it less efficient than Kyber [23].

A combination of methods is proposed to protect CMD. AES-256 is effective for current needs, CRYSTALS-Kyber for clouds and 6G networks due to its resistance to quantum attacks, Niederreiter is a fallback for IoT. 6G networks expected by 2030 will provide high speed and minimal latency, but will create new vulnerabilities [19]. Post-quantum algorithms such as CRYSTALS-Kyber and HQC will protect IoT from quantum computers, Niederreiter is an alternative if optimized [10]. Standardization by NIST and FDA will facilitate the implementation of post-quantum algorithms [20, 21].

CRYSTALS-Kyber is widely used in hybrid systems, where it is combined with classical KEMs like ECDH to protect IoT, including CMD. Its integration into AWS, Cloudflare, and IBM z16 demonstrates its practicality [23].

TABLE V. COMPARISON OF CRYSTALS-KYBER AND NIEDERREITER CHARACTERISTICS

Characteristic	CRYSTALS-Kyber	Niederreiter
Algorithm type	Key Encapsulation Mechanism (KEM)	Public-key encryption
The basis of security	LWE over the bars	Decoding linear codes
Efficiency, %	98	95
Standardization	NIST FIPS 203 (2024)	Not standardized
Public key, bytes	800–1568	~2 000 000 (128-bit security)
Secret key, byte	1632–3072	Smaller than the public key
Ciphertext, byte	768–1568	Depends on the parameters
Encryption speed	~10–20 us (IoT devices)	10 times faster than McEliece
Application	Key exchange, 5G/6G, IoT	Direct encryption, IoT
Prevalence	AWS, Cloudflare, IBM	Research projects

The Niederreiter cryptosystem, based on the decoding complexity of linear codes, uses large parity check matrices, which makes it difficult to implement in IoT devices such as CMD due to memory and computational resource limitations. The combination of the Niederreiter CCK with quasicyclic LDPC codes based on the circulant block structure allows for a compact representation of the matrices, reducing the size of the keys and increasing the decoding efficiency, which reduces the size and increases the efficiency of the CMD. This makes them promising for post-quantum security, where resistance to quantum attacks is key [19].

The combination involves the use of QC-LDPC codes as the base code in Niederreiter. The parity check matrix H is formed from circulant blocks, which reduces the memory requirements. Encryption is performed by calculating the syndrome, and decoding is performed using efficient algorithms such as belief. Security depends on hiding the code structure, making it difficult to decode without the private key [19].

Since QC-LDPC codes are a subclass of linear block codes with sparse parity check matrices consisting of circulant blocks. The circulant matrix is defined by the first row, where each subsequent row is its cyclic shift. For a vector[a, b, c] , the circulant matrix of size 3x3 has the form:

$$\begin{bmatrix} a & b & c \\ c & a & b \\ b & c & a \end{bmatrix}$$

This structure reduces memory requirements, as only the first line of each block needs to be stored, thanks to efficient decoding algorithms such as belief or minimum sum. Their flexibility allows tuning parameters such as code length n, dimensionality k, and error correction capability t to ensure security and performance [19].

The Niederreiter cryptosystem uses a parity check matrix H of size $(nk)\times n$ as the public key, and the private key includes a generator matrix G and a decoding algorithm. The encryption process consists of computing the syndrome $S= mHT$, where m is an error vector of weight t, representing the message. Decryption recovers s using a private decoding algorithm. Security depends on the complexity of decoding random codes without knowing the structure [10].

The combination of QC-LDPC codes from Niederreiter involves using QC-LDPC as the base code, which allows reducing the size of the keys and optimizing the calculations. The algorithm includes the following steps:

1. Code parameter selection: The parameters of the QC-LDPC code (n,k,t) are selected, where n is the code length, k is the dimension, and t is the number of errors to be corrected. For $n=1024, k=524, t=50$ the code provides 128-bit security [26]. The matrix H is formed as a block matrix of circulating blocks of size $m\times m$, where $m=n-k$.

$$H = [P_1 \times P_2 \ldots P_i] \quad (4)$$

where P_i are circulant matrices of size $m\times m$,

2 Key generation:

2.1 Private key: A QC-LDPC code is generated with a matrix H and a decoding algorithm (elief propagation).

2.2 Public key:

$$H_{pub} = S \times H \times P \quad (5)$$

Where: S – is a non-degenerate matrix;

P – is a permutation matrix that hides the structure of H.

The size of the public key for a QC-LDPC code is calculated as:

$$S_{key} = m \times l \times \log_2(\mathrm{q}) \quad (6)$$

Where: S_{key} – - key size (bits);

m – number of rows in the block $m=n-k$;

l – number of circulating blocks $l=n/m$;

g – field size $(g=)2$ for binary codes).

3 Encryption: Message m (error vector of weight $\leq t$) is encrypted as:

$$S = mH_{pub}^T \quad (7)$$

where S is the syndrome (ciphertext) $1\times(n-k)$;

H_{pub} – public key;

m – error vector of size $1\times n$;

H_{pub}^T – transpose matrix of H_{pub}.

4 Decryption: A private decoding algorithm is used to recover m from s, using a QC-LDPC decoding algorithm, such as elief propagation, with complexity $O(n\times d)$, where d is the density of the matrix H.

5 Optimization: FFT-based algorithms are used for calculations with circulant matrices, which reduces the complexity [25]. The probability of protection against attacks is estimated see Formula 2.

Thus, the implementation of the QC-LDPC combination algorithm with Niederreiter will allow to reduce the key size compared to traditional Hopp codes, which facilitates

transmission in low-bandwidth networks such as LoRa. For *n=1024, k=525, m=500*, the key size can be reduced to ~10–100 kB due to the circulant structure [25]. Efficient decoding algorithms, such as belief, reduce the computational complexity, which is critical for CMDs with memory up to 1 MB and power up to 1 MHz. Studies, such as in [19], show that QC-LDPC codes can be resistant to quantum Fourier sampling under certain conditions, such as uniqueness of elements in circulant blocks. However, some implementations have been found to be vulnerable to attacks, such as Sidelnikov-Shestakov, if the parameters are chosen incorrectly [19]. New schemes, such as the "tiny public key scheme", offer keys of 18–500 bits in size, but require security verification [25].

For CMD, combining QC-LDPC with Niederreiter [26] allows:

– To reduce key sizes, which facilitates transmission in 5G/6G networks.

– To provide quantum stability, which is critical for long-term protection of medical data.

– To use efficient decoding algorithms for fast message recovery.

Compared to CRYSTALS-Kyber [27] (keys 800–1568 bytes), Niederreiter with QC-LDPC can be faster in encryption, but requires optimization to reduce key sizes [22].

III. Conclusion

The combination of QC-LDPC codes with Niederreiter cryptosystem is promising for post-quantum IoT security, in particular CMD, due to the compact key representation and efficient decoding. The integration algorithm includes parameter selection, key generation, encryption and decryption using circulant matrices. The calculation formulas allow us to estimate the key size, stability and performance. However, careful security verification and optimization are required for practical implementation.

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Model Forecasting Reactive Power Indicators of the Enterprise

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***Abstract*— The article considers the problem of using neural networks for predicting reactive power indicators in the power industry. It shows the relevance of the tasks of forecasting the consumption of electricity for planning the mode of operation of the enterprise, and establishes that the regulation of the level of reactive power allows to reduce electricity losses in grid elements and improve the quality of electricity. This work reviews and analyzes the works on reactive power compensation and determination of predictive values using artificial neural networks.**

Two options for forecasting the power factor and reactive power factor using artificial neural networks are considered. The first method is the direct forecasting of indicators, the second is the calculation of the components of indicators on the basis of predictive models on a schedule with accumulation. The period of the studied graphs is determined by the periodograms of the input data to justify the number of neurons in the input layer of neural networks. Predictive models based on a multilayer perceptron with linear, hyperbolic and exponential activation functions of neurons and the number of hidden neurons from 2 to 48 are presented.

The results of modeling by two forecasting methods are presented and the quality of the models on the accepted forecasting interval is compared in terms of network performance, the law of distribution of residuals and the root mean square error.

Keywords — neural network, forecasting, power factor, multi layer perceptron.

INTRODUCTION

Industrial processes have different needs for both active and reactive power. A significant aspect of energy saving in industrial enterprises is the compensation of reactive power – a component of the total power that characterizes electromagnetic processes in electrical networks. Excess of this component can lead to a significant increase in loss of active energy and voltage. To minimize losses, ensure stable operation of the network and increase its capacity, it is necessary to constantly optimize the level of reactive power in the network.

Due to the uneconomical transmission of reactive power to consumers, compensating devices are installed directly in the distribution networks of the enterprise to provide regulation of power in accordance with changes in loads in the network [1]. In this regard, there is a need to develop models that would forecast consumption volumes for the formation of regulatory influences.

I. ANALYSIS OF RECENT RESEARCH AND PUBLICATIONS

The relevance of the study is due to the need to influence the reactive power, which will improve the energy efficiency of existing power grids. When considering this issue, two components should be distinguished, each of which can be considered as a separate problem. The first is the technical implementation of reactive power control algorithms, the second is the formation of predictive control impact based on preliminary data on the operation of the enterprise.

The main task in planning the operation of reactive power sources is to provide the power system with sufficient reactive power under various conditions. Reactive power generation is used to maintain voltage levels and control the level of losses in the network. Under normal operating conditions, network losses can be reduced by proper reactive load control. Planning of the level of reactive power should take into account the operating time interval for each indicator in the planning horizon, which will ensure economic and safe operation [2].

Reducing power losses and improving power quality at the terminals of electrical receivers can be achieved by compensating for reactive power. A predefined cosφ can be used as an indicator of the optimal value of the level of reactive power consumption.. It is preferable to focus on those methods that are based on the study of forecast estimates that constitute the initial information for making management decisions to regulate the operation of compensating installations In this regard, it is advisable to

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comprehensively regulate reactive power with a combination of voltage regimes in the system [3].

The work in the field of forecasting using artificial intelligence technologies is now gaining prominance in all areas due to the wide range of possibilities provided by artificial neural networks. Static forecasting models solve the tasks of detecting periodicity in data, analyzing dependencies, and assessing risks, among others. Promising forecasting methods are those based on artificial neural networks, which are able to comprehensively process information and identify hidden dependencies between input and output data [4-5].

Accurate forecasting of production process indicators is an important basis for ensuring the stable operation of the enterprise. Because some indicators, which are a complex nonlinear problem, are difficult to accurately reflect by existing forecasting methods, it is advisable to use an artificial neural network that can predict this indicator by a function infinitely close to any nonlinear function, which will be the most ideal forecasting method at the moment [6]. Artificial neural networks are widely used for forecasting as a powerful modeling method. Most neural network inputs for forecasting are single-factor, while certain process indicators can be multifactor. There are mixed results comparing the forecasting performance between neural networks and other models. The forecasting performance of neural networks can be improved by integration with other technologies. Nonlinear combination of forecasts using neural networks also gives encouraging results [7].

One of the most popular models of artificial neural networks is the Multi Layer Perceptron (MLP) network. On the basis of which it is possible to build accurate forecasts for grid operators when planning efficient generation schedules to meet the demand of electricity consumers. The artificial neural network model is an effective method for forecasts because it demonstrates such properties as the ability to adapt, the ability to learn from previous examples and the ability to generalize [8].

Among the unsolved problems remains the increase in the accuracy of forecasts of the level of reactive power as one of the factors influencing the quality of electricity at load nodes. To solve this problem, this work is being carried out, the scientific novelty of which lies in improving the quality of predictive models in energy through the use of artificial neural networks.

II. The Goal of This Paper

The paper investigates the indicators of reactive power consumed by the enterprise and neural network models that allow predicting their change. The aim of this paper is to construct a neural model of an enterprise that performs the functions of predicting the level of reactive power. To achieve this goal, the following tasks are solved in the work:

- Evaluation of indicators characterizing the consumption of reactive power by the enterprise and establishing their averaging intervals.
- Substantiation and configuration of an artificial neural network for forming a model of reactive power consumption by an enterprise.
- Forecasting reactive power consumption parameters and assessing the quality of the neuromodel.

III. Presentation of the Main Research Material

In practice, consumption or generation of reactive power is often characterized by the power factor cosφ or the reactive power factor tgφ. The value of tgφ from the point of view of mode analysis in conditions of high values of cosφ is considered to better reflect the impact of reactive power on processes [9]. The paper deals with the prediction of reactive power and power factors using artificial neural networks of the MLP type. The input data are represented by graphs of changes in active P=f(t) and reactive powers Q=f(t) (Fig. 1) and graphs of changes in power factor cosφ_i=f(t) and reactive power factor tgφ_i=f(t) (Fig. 2) (Numerical values of the samples used in the calculations are not given due to the significant amount of data). The samples cover a time interval of 10 days, which is 480 observations. Averaging of values was performed at an interval of 30 minutes (1 point of observation equal to 0,5 hour). The maximum value of active power is 400 kW, the maximum value of reactive power is 220 kVAr. Network training and forecasting will be performed based on 432 observations (9 days), and the data of the last day will be used as a test sequence.

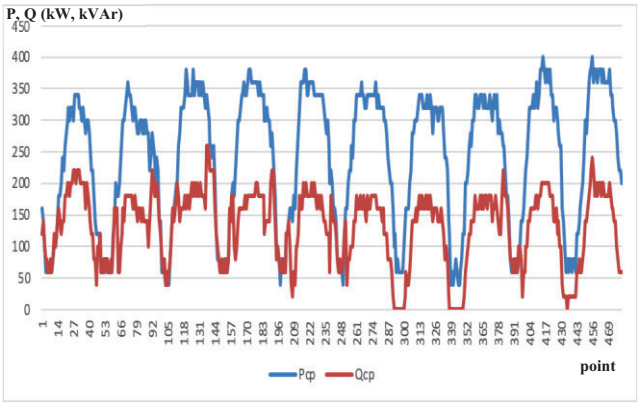


Fig. 1 Graph of changes in active and reactive capacities.

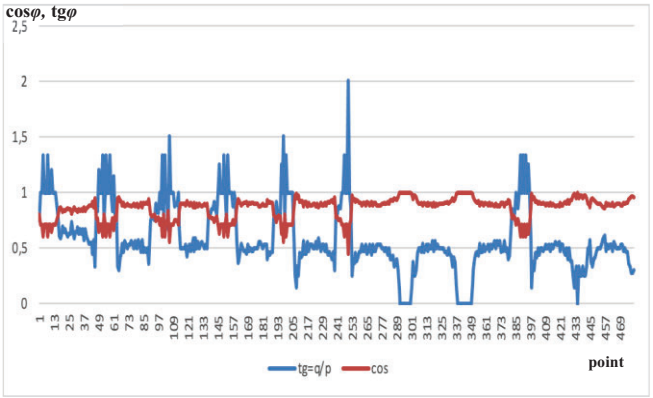


Fig. 2 Graph of changes in power factors and reactive power.

Modeling of neural networks is performed using the package "STATISTICA" [10].

The number of network neurons in the input layer is determined using the Fourier spectral analysis of the periodograms (Fig. 3-5). Fig. 3 shows the periodogram of the graph of reactive power change for the full range of the study, which determines the periodicity of the graph. In this case, the main period can be identified on the periodogram, which is 48 measurements, which corresponds to the number of measurements per day. The same regularities are observed

for the periodograms of power factor and reactive power factor. However, the presence of other peak values on the graphs indicates the possibility of changing the prediction period, and hence the number of neurons in the input layer.

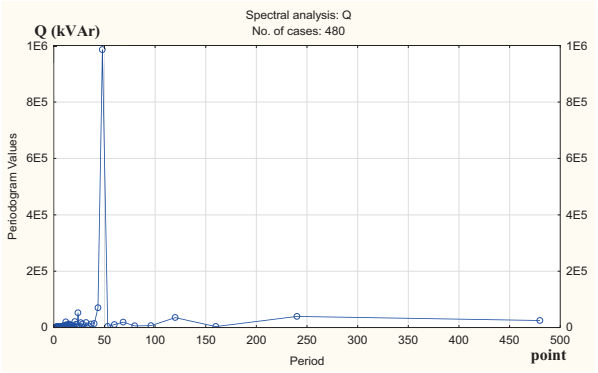


Fig. 3 Periodogram of the graph of reactive power change.

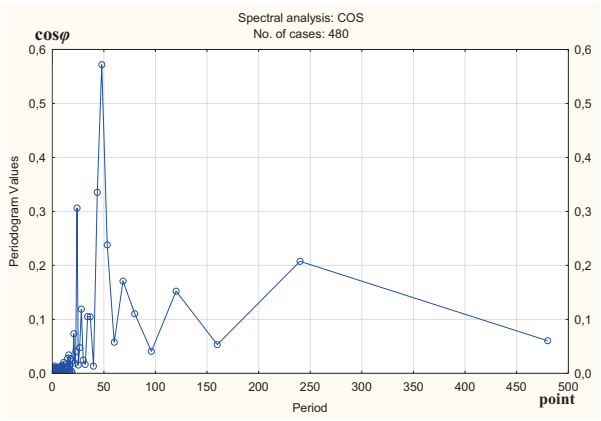


Fig. 4 Periodograms of graphs of power factor.

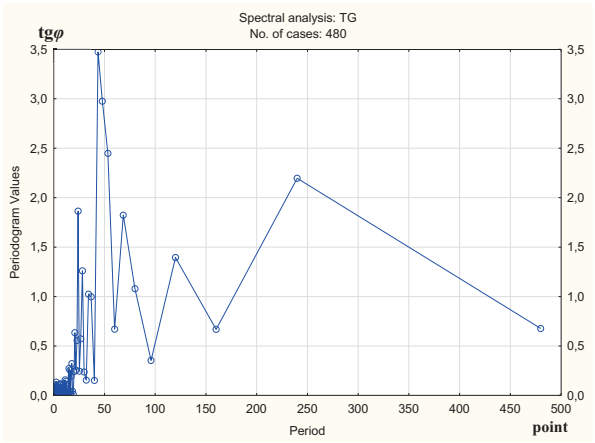


Fig. 5 Periodograms of graphs of reactive power factor.

Predicted values will be determined in two ways. According to the first one, the neural models of cosφ and tgφ will be built, and according to the second one, the forecasting of active and reactive powers will be performed and cosφ and tgφ will be calculated. Forecasting of capacities will be made on the basis of a graph with accumulation, since this method of forecasting avoids changing the sign of the derivative of the forecast function. The neuromodel uses a multilayer perceptron (MLP) that will be trained using the BFGS algorithm (Broyden - Fletcher - Goldfarb - Shanno algorithm).

The MLP structure involves 48 input neurons defined by periodograms, with one neuron in the output layer corresponding to the prediction task, and a variable number of neurons in the hidden layer from 2 to 48. Linear, hyperbolic, and exponential activation functions of neurons are used. The best configuration of the network will be determined by its performance. The quality of the models will be assessed by the performance (the correlation between the output and the predicted series), the distribution of residuals (the nature of the distribution of the differences between the input and the predicted values for all observations) and the average relative forecast error. Among the sample of 250 networks for each calculation option with randomly set initial conditions, 5 best ones were selected (Tables 1-2) for further analysis.

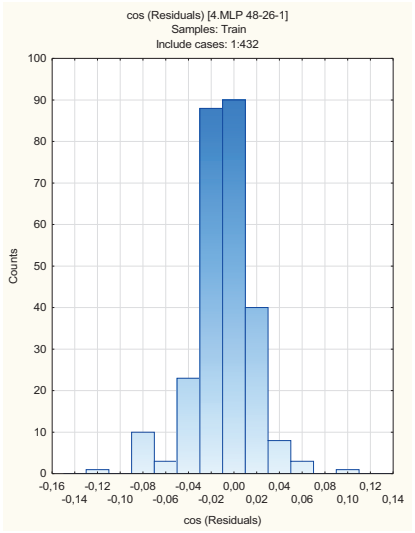
TABLE I
TRAINING RESULTS ON POWER FACTOR SEQUENCE DATA

№	Net. name	Training perf.	Test perf.	Training error	Test error	Training algorithm	Hidden activation	Output activation
1	MLP 48-28-1	0,929211	0,911831	0,000660	0,000533	BFGS 44	Tanh	Exponential
2	MLP 48-15-1	0,939073	0,908451	0,000566	0,000556	BFGS 39	Tanh	Identity
3	MLP 48-26-1	0,959493	0,907423	0,000375	0,000567	BFGS 55	Tanh	Exponential
4	MLP 48-26-1	0,966417	0,902808	0,000311	0,000608	BFGS 63	Tanh	Exponential
5	MLP 48-21-1	0,932121	0,905493	0,000628	0,000592	BFGS 43	Tanh	Exponential

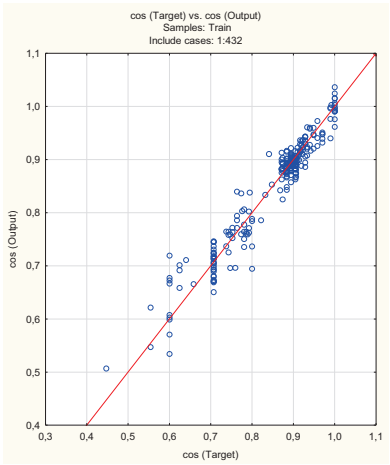
TABLE 2
TRAINING RESULTS ON REACTIVE POWER FACTOR SEQUENCE DATA

№	Net. name	Training perf.	Test perf.	Training error	Test error	Training algorithm	Hidden activation	Output activation
1	MLP 48-14-1	0,924102	0,892661	0,006345	0,006142	BFGS 29	Tanh	Exponential
2	MLP 48-32-1	0,936701	0,885571	0,005305	0,006849	BFGS 33	Tanh	Exponential
3	MLP 48-3-1	0,896098	0,894064	0,008616	0,006277	BFGS 21	Exponential	Identity
4	MLP 48-17-1	0,965732	0,890172	0,002897	0,006638	BFGS 58	Exponential	Exponential
5	MLP 48-3-1	0,867476	0,886273	0,010806	0,006407	BFGS 16	Tanh	Exponential

The constructed networks were analyzed by residuals distribution characteristics and a scatter diagram between input data and predicted values (Figures 6-7). The characteristics indicate a normal distribution of residuals, and the scatter diagrams indicate a trend towards a straight line, but do not completely coincide.

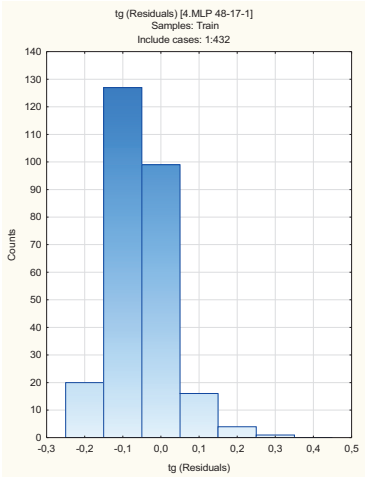


a)

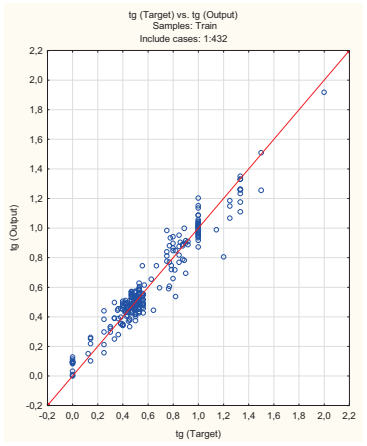


b)

Fig. 6 Characteristics of the neural model for determining the power factor cosφ.



a)



b)

Fig. 7 Characteristics of the neural model for determining the power factor of reactive power tgφ.

For further evaluation of the obtained models, forecast charts for one day ahead were built for 48 forecast values (Figures 8-9).

Analysis of the results shows that, even with sufficiently high values of network performance (more than 95% on the test sequence and 89% on the test sequence), the forecast values for the daily interval have significant differences). The power factor schedule has significant discrepancies at the beginning of the forecast period, and the reactive power factor schedule does not show sufficient coincidence with the expected schedule. Therefore, direct forecasting for significant intervals without correction does not give sufficiently accurate results. Therefore, as an alternative forecast model, a forecast will be made on the basis of active and reactive power indicators obtained on the basis of schedules with accumulation. When constructing neural networks based on accumulation graphs for active and reactive powers, the values presented in Table 3 were obtained.

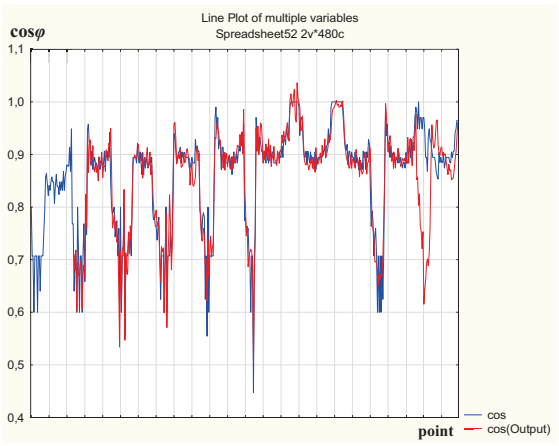


Fig. 8 Graphs of changes in the power factor cosφ for input data and neural model with a forecast for 1 day.

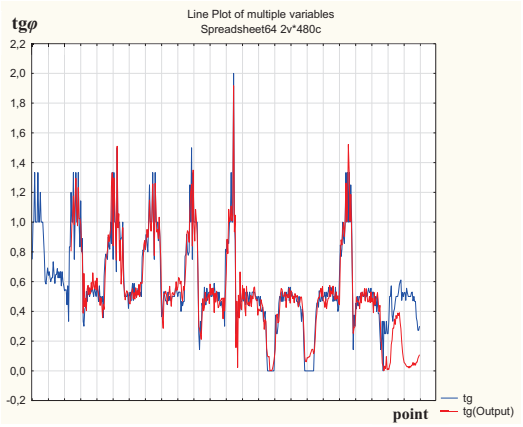


Fig. 9 Graphs of changes in the power factor cosφ for input data and neural model with a forecast for 1 day.

The obtained networks are characterized by high performance (over 97.5% on the test sequence and over 98.5% on the control sequence) and have a normal distribution of residuals. Based on the MLP 48-39-1 and MLP 48-12-1 networks, the values of cosφ and tgφ are calculated. Since the networks built a graph with accumulation, to determine the indicators, the values were recalculated into a number of absolute values of capacities.

$$P_i' = |P_{i+1} - P_i| \quad (1)$$

$$Q_i' = |Q_{i+1} - Q_i| \quad (2)$$

where P_i' , Q_i' - are absolute values of consumed active and reactive powers; P_i, Q_i - are modeled according to the schedule with accumulation of active and reactive power values.

TABLE 3
TRAINING RESULTS ON SEQUENCE DATA WITH ACCUMULATION OF ACTIVE AND REACTIVE POWER METER READINGS

№	Net. name	Training perf.	Test perf.	Training error	Test error	Training algorithm	Error function	Hidden activation	Output activation
	Non-linear network for active power forecasting								
1	MLP 48-15-1	0,975521	0,985453	0,000043	0,000070	BFGS 110	SOS	Identity	Identity
2	MLP 48-18-1	0,975521	0,985453	0,000043	0,000070	BFGS 90	SOS	Identity	Identity
3	MLP 48-46-1	0,975521	0,985453	0,000043	0,000071	BFGS 87	SOS	Identity	Identity
4	MLP 48-40-1	0,975521	0,985453	0,000043	0,000070	BFGS 96	SOS	Identity	Identity
5	MLP 48-39-1	0,975521	0,985453	0,000042	0,000071	BFGS 100	SOS	Identity	Identity
	Neural network for reactive power prediction								
6	MLP 48-46-1	0,971145	0,982892	0,000065	0,000067	BFGS 81	SOS	Identity	Identity
7	MLP 48-12-1	0,971145	0,982892	0,000065	0,000067	BFGS 88	SOS	Identity	Identity
8	MLP 48-30-1	0,971145	0,982892	0,000068	0,000066	BFGS 75	SOS	Identity	Identity
9	MLP 48-22-1	0,971145	0,982892	0,000063	0,000066	BFGS 90	SOS	Identity	Identity
10	MLP 48-6-1	0,971145	0,982892	0,000063	0,000066	BFGS 87	SOS	Identity	Identity

The results of forecasts obtained by the proposed models are presented in the graphs (Fig. 10-11). Compared to the previous models, the graphs show better agreement with the test sequence. The average relative error for the predicted cosφ plot (Fig. 10) is 5,3 %, and for the tgφ plot it exceeds 10 %. The tgφ plot has significant excesses of the predicted peak values, which leads to a decrease in the overall accuracy of the model. Consequently, when forecasting reactive loads, it is advisable to use a combination of direct forecasts of reactive power consumption and calculated values of cosφ obtained from graphs with accumulation (readings of power meters).

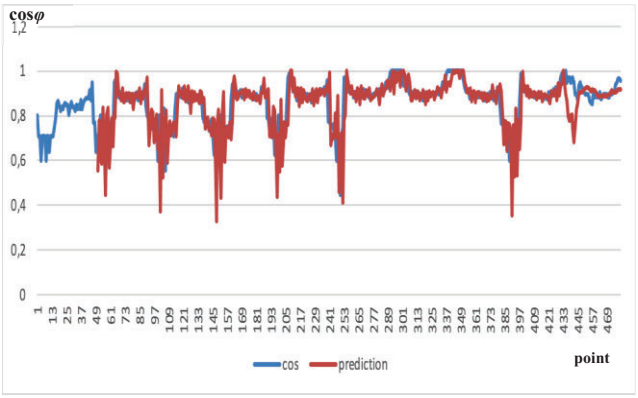


Fig. 10 Input graph cosφ against the prediction.

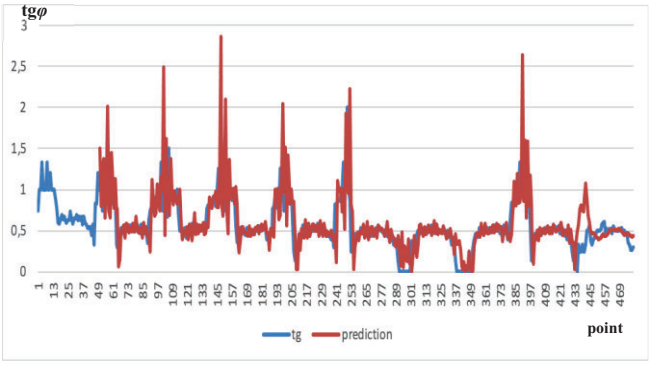


Fig. 11 Input graph tgφ against the prediction.

CONCLUSIONS

The use of artificial neural networks for building models for predicting reactive load indicators is substantiated.

The required number of neurons (48) in the input layer is determined by half-hourly graphs of active and reactive loads and their indicators. Reducing the number of neurons will worsen the quality of the model due to a period mismatch and increasing it will increase the computation time.

Given that the average relative error for the predicted cosφ graph is 5,3 %, and for the tgφ graph it exceeds 10 %, it is advisable to use the predicted values of active and reactive power meters and the calculated cosφ value to determine the required level of reactive power compensation.

The obtained values allow the use of neural network models based on a multilayer perceptron that predicts the value of cosφ for building systems for automatic power control of compensating devices with a variable load schedule.

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Application of the Dielectric Permittivity Parameter of Crude Oil during Its Processing

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Abstract. **The article presents research on the prospects of using the relative dielectric permittivity (ε^t) of crude oil as a process parameter during its refining. It has been established that ε^t demonstrates a strong correlation with kinematic viscosity and Conradson carbon residue. Based on the conducted studies, a predictive criterion (PC) has been proposed, which enables the identification of rational oil refining schemes aimed at increasing the yield of valuable products and reducing production costs.**

Keywords: crude oil; refining; dielectric permittivity; processing scheme; resins; asphaltenes; viscosity; Conradson carbon residue; density

I. INTRODUCTION

Nowadays, oil refining is considered one of the strategic industrial sectors of Ukraine, and the economic situation of the country significantly depends on its development. This industry produces a variety of petroleum products, including gasoline, diesel fuel, fuel oil, lubricants, and road and construction bitumen, all of which are in high demand.

Recently, Ukraine's oil refining sector has been facing a number of challenges:

- a reduction in processing volumes due to the shutdown of production processes at most refineries caused by equipment wear and tear, low technological level of processing units, and insufficient investment;
- a significant dependence on imported petroleum products, affecting the country's energy security;
- the impact of military actions, which have caused major infrastructure destruction and the suspension of operations at most refineries.

At the same time, the development prospects of Ukraine's oil refining sector can be outlined in the form of a structural scheme shown in Figure 1.

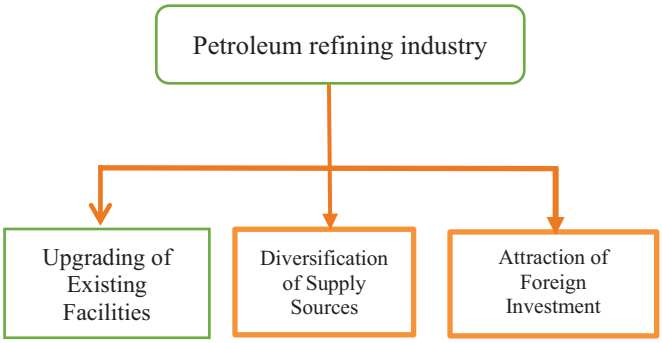


Fig. 1. Prospects for the development of the oil refining industry in Ukraine

II. Purpose and Research Objectives

In our view, the modernization of existing capacities represents a key strategic direction for the development of Ukraine's oil refining industry, one that depends entirely on the domestic scientific potential and can be implemented in the short term. The main aspects of such modernization include the improvement of technological processes through the use of new construction materials, catalysts, and engineering solutions, as well as continuous monitoring of the technological process. It is the process monitoring that has been selected as the primary focus of modernization for the current oil refining capacities in Ukraine's industrial sector. To implement this monitoring, we have employed the parameter of relative dielectric permittivity (ε^t) of crude oil – the primary hydrocarbon feedstock used in the production of commercial petroleum products. The parameter (ε^t) characterizes the insulating properties of the medium – in this case, crude oil – which may vary under the influence of numerous factors such as chemical and fractional composition, and the presence of contaminating impurities. In both oil production and refining industries, the ε^t parameter holds practical significance and is widely applied across several key directions (presented on the Fig.2)

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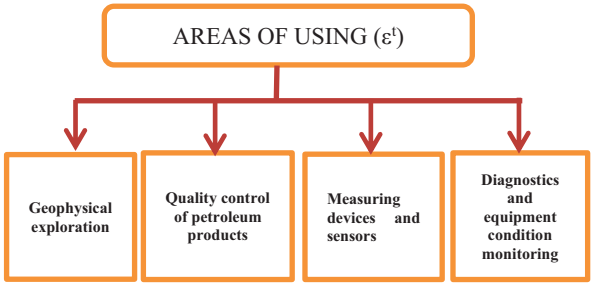


Fig. 2. Key applications of the parameter (ε^t) in the petroleum extraction and refining industries

Geophysical exploration for oil and gas deposits employs electrical prospecting methods, in particular dielectric probing techniques [4, 5]. Due to the significant difference in dielectric permittivity between formation water (~80 units) and crude oil (~2–4 units), this contrast is utilized for identifying oil-bearing strata. Geophysical exploration also incorporates radar-based investigation techniques, such as Ground-Penetrating Radar (GPR) [6]. In these applications, radio wave propagation is influenced by the stratigraphic profile, and the ε^t parameter directly affects the speed and reflection of waves [7].

Quality control of petroleum products.

The dielectric permittivity (ε) of crude oil is indeed an integral parameter and is highly dependent on its composition. Typically, an average experimental value is used; however, if it is necessary to account for the influence of individual components, a mixing approach is applied (construction of calibration curves) considering their polarity. In practice, it is sufficient to determine the dielectric permittivity value of crude oil within the range of critical thresholds, which indicate the presence of contaminating impurities in the oil.

Measurement of ε^t is applied to identify the composition of liquid hydrocarbons and detect the presence of water or other polar impurities [8, 9]. For instance, in [10], an index of oil preparation (IOP) is proposed, which is a function of the dielectric permittivity parameter ε^t.

Measuring devices and sensors. Sensors based on the principle of ε^t measurement are used in tanks and pipelines to determine crude oil levels [11], detect phase boundaries (oil–water), and monitor the formation and breakdown of emulsions such as "water-in-oil" or "oil-in-water" [12, 13].

Diagnostics and equipment condition monitoring. In dielectric testing of insulation materials, hydraulic fluids [14], and transformer oils [15, 16], the ε^t parameter is also considered. This enables the detection of material degradation, presence of water or impurities, which are undesirable and may lead to failure during operation of electrical equipment.

Crude oil and its refined products belong to nonpolar liquids. For such systems, a decrease in ε of approximately 0.2–0.5% per each 1 °C increase in temperature is typical. This can be explained by the reduction in density upon heating: the molecules move further apart, and their polarizability per unit volume decreases. Therefore, all dielectric permittivity measurements were carried out at a fixed temperature.

Since ε^t is a critical physical property widely used in oil extraction and refining, we propose an alternative application: using ε^t for real-time monitoring of the crude oil refining process. Specifically, this involves assessing the degree of distillate fraction recovery relative to their potential content in crude oil. The ε^t parameter correlates with the depth of crude oil conversion and directly impacts the operational efficiency of primary refining units.

The dielectric permittivity (ε) is one of the most sensitive parameters to moisture and emulsification in crude oil and petroleum products. Considering that dehydrated crude oil has $\varepsilon \approx 2–2.5$, while water has $\varepsilon \approx 80$, even trace amounts of water (0.1–0.5%) in oil significantly increase its effective ε. The corresponding relationship is quadratic polynomial in nature. Moreover, the higher the moisture content, the faster ε increases, especially during the transition from water microdroplets to macro-droplets. When a stable emulsion is formed, the dielectric permittivity of the mixture can be described by effective medium models (e.g., the Maxwell–Garnett or Bruggeman formulas). The smaller and more uniformly dispersed the water droplets, the closer the ε of the emulsion approaches the average value between that of oil and water. For “oil-in-water” type emulsions (rare in petroleum refining), ε is even higher, since the nonpolar phase acts as the dispersed medium.

The primary instruments for obtaining real-time information on material flows are capacitive sensors, which transmit data to the DCS (Distributed Control System). Key installation points for these sensors include: the crude reception unit (online monitoring of water-in-oil content and batch stability); upstream and downstream of the electro-dehydrator (control of voltage, demulsifier dosing, and recirculation); on return/discharge lines of emulsions (signaling phase separation and water carryover); and in tank farms (recirculation loop) for long-term monitoring of the water/sludge balance.

In addition, it should be mentioned that based on common practice at refineries and chemical plants, it should be noted that scheduled calibration of capacitive sensors is carried out once every 6–12 months (in accordance with the manufacturer’s specifications and the internal regulations of the metrology service). Unscheduled calibration is performed after repair or replacement of cables/electronics, after process reconfiguration, or in case of sharp deviations in readings (instability, “floating” values). In critical systems (such as tank level monitoring of petroleum products and online water content measurement in crude oil), calibration may be required quarterly or even on a monthly basis. If the sensor operates under relatively stable conditions (constant temperature, absence of deposits), annual calibration is sufficient. In environments with a high risk of deposit or emulsion formation, accuracy verification is typically performed every 3–6 months.

Standardization of the proposed methodology, however, may be complicated by several factors: variability of crude oils from different fields, influence of moisture and emulsions, temperature dependence, differences in

instrumentation and calibration procedures, as well as the absence of unified international standards. Nevertheless, the most promising approach involves developing correlation models in which dielectric permittivity (ε) is integrated with other parameters (density, viscosity, resin and asphaltene content) within a unified methodology.

III. RESULTS AND DISCUSSION

When analyzing the dielectric permittivity parameter of crude oil (ε_{oil}) and gas condensate ($\varepsilon_{g.c.}$) feedstocks, the following relationship can be observed:

$$\varepsilon_{g.c.} < \varepsilon_{oil.} \quad (1)$$

This relationship is determined by the fractional composition of gas condensates: the content of light fractions that boil at temperatures up to 360 °C constitutes 80–95 vol.%, whereas in crude oil such fractions do not exceed 60 vol.%. Additionally, gas condensates contain lower amounts of sulfur compounds, resins, and asphaltenes compared to crude oil feedstock [17].

Therefore, we propose a predictive criterion (PC) for determining the processing direction of petroleum feedstock, which in general form can be expressed as:

$$PC = A_1 \times X_1 + A_2 \times X_2 + A_3 \times X_3 + \cdots + A_n \times X_n \quad (2)$$

where A_1–A_3 are dimensionless weighting coefficients of the quality indicators,

X_1 to X_n are the quality parameters of the petroleum feedstock.

The proposed criterion is based on the interrelation between the dielectric permittivity ε^t and other quality indicators of the feedstock. This is due to the fact that crude oils from different fields can significantly differ in their group chemical composition, which makes it impossible to obtain adequate information about them relying solely on the ε^t parameter values.

The practical application of equation (2) involves the use of physicochemical quality indicators of the crude oil feedstock. Among the most informative quality indicators that can be used in equation (2) are kinematic viscosity and Conradson carbon residue.

Kinematic viscosity (ν, mm²/s) is one of the most important technological properties of crude oil, which depends on its fractional and chemical composition (the higher the content of high-boiling fractions in the feedstock, the higher its viscosity) and characterizes the pumpability of oil during its transportation through refinery pipelines. Additionally, kinematic viscosity allows assessing the suitability of crude oil for the production of lubricating materials. That is, hydrocarbon feedstock with a high kinematic viscosity value ($\nu^{20} \geq 150$ mm²/s) is advisable to be used for manufacturing various types of liquid lubricating oils. Based on literature data [18], related to the use of kinematic viscosity in various technological calculations, the value of viscosity measured at 20 °C or 40 °C should be used in equation (2).

The choice of Conradson carbon residue (X_c, wt.%) for characterizing crude oil is explained by the fact that its value strongly depends on the content of resins and asphaltenes (ha) in the feedstock. When their content is high ($h_a \geq 20$ wt.%), typical for heavy crude oil, the production of high-quality base oils—the main component for manufacturing commercial lubricating oils—becomes complicated. This necessitates the use of special (selective and adsorption) methods for deep purification of the obtained base oils.

Moreover, kinematic viscosity and Conradson carbon residue exhibit a close correlation with the ε^t parameter, as confirmed by our experimental studies and illustrated in Figures 3–4. It is evident that with an increase in the content of high-molecular-weight hydrocarbons (resin-asphaltene substances) in the oil, which have boiling points above 350 °C and contain molecules of O, S, N, and metals (V, Ni), the value of the ε^t parameter increases. This, in turn, is fully consistent with existing understanding of the contributions of various factors to the ε^t parameter magnitude.

Based on the information presented above, the model for determining (PC) will take the final form as follows:

$$PC = A_1 \times \varepsilon + A_2 \times \nu^t + A_3 \times X_c \quad (3)$$

where ν^t is the kinematic viscosity of the crude oil at 20 °C or 40 °C, mm²/s; X_c is the Conradson carbon residue of the crude oil, wt.%.

Currently, the main processing options for crude oil implemented at petroleum refining enterprises include fuel, lubricating oil, and combined processing [19].

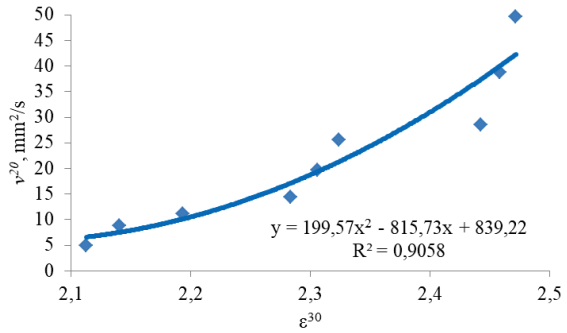


Fig. 3. Effect of the parameter ε^{30} on the ν^{20}

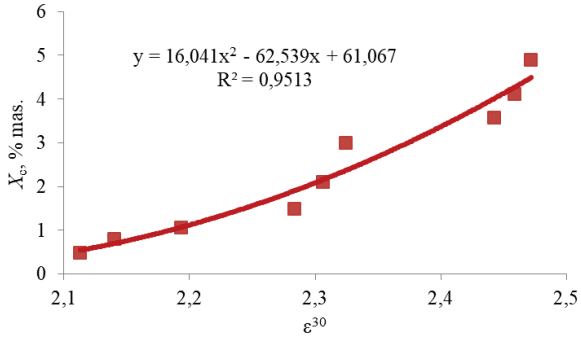


Fig. 4. Effect of the parameter ε^{30} on X_c

Based on these standard processing routes, Figure 5 graphically presents the main existing options for crude oil processing depending on the value of PC (RPC). According to Figure 5, for each crude oil processing route, the following expressions can be obtained:

$$P_{PC}^{F} < P_{PC1} \quad (4)$$

$$P_{PC1} \le P_{PC}^{F} \le P_{PC2} \quad (5)$$

$$P_{PC2} \le P_{PC}^{F} \le P_{PC3} \quad (6)$$

where P_{PC}^{F} is the actually determined level of PC; P_{PC1}, P_{PC2}, P_{PC3} are the experimentally established boundary values of PC for each crude oil processing option.

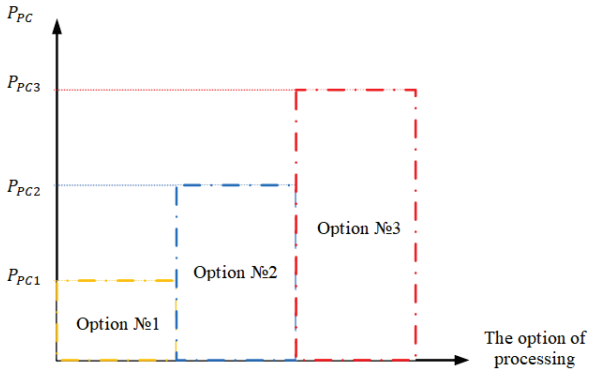


Fig. 5. Directions of refining at different P_{PC}

It should be noted that a physicochemical quality indicator of crude oil such as density (ρ^{20}, kg/m³), which reflects its chemical composition and serves as the basis for its classification [20], according to which rational processing options must be developed, has a rather close correlation with the parameter ε^{30} (see Fig. 6).

It is clear that light crude oils are characterized by a higher content of light fractions, which mostly consist of alkanes and naphthenes, and a lower content of aromatic hydrocarbons. Heavy crude oils mainly consist of high-boiling fractions (boiling point above 350 °C), which contain a significant amount (on average up to 8%, and sometimes up to 20%) of resinous-asphaltene substances. Such a distribution of hydrocarbons in crude oils also affects the magnitude of the ε^{30} parameter.

Then, expression (4) is characteristic of light crude oil (density from 830 kg/m³ to 850 kg/m³) and gas condensate with a high (35–95%) content of fuel fractions boiling up to 350 °C. For processing such feedstock, the use of "Option No. 1," based on primary processing, is economically feasible. The main products of this processing route are motor fuels and boiler fuels.

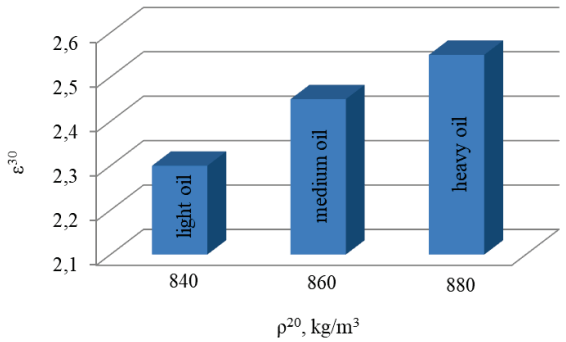


Fig. 6. Average value of ε^{30} depending on the average value of ρ_{20} for each crude oil group

Expression (5) is characteristic of medium crude oil (density from 850 kg/m³ to 870 kg/m³) with an average (up to 20–35%) content of fuel fractions boiling below 350 °C.. For processing such feedstock, "Option No. 2," which includes combined use of primary processing with subsequent dewaxing and deasphalting of the obtained lubricating fractions or the use of hydrocracking of heavy oil fractions, is economically feasible. The main products of this processing route are purified base oils for the production of industrial, high-viscosity-index (with viscosity index above 130 units) motor oils, as well as transmission oils. By-products include motor and boiler fuels.

Expression (6) is characteristic of heavy crude oils (density above 870 kg/m³) with an average (up to 20%) content of fuel fractions boiling up to 350 °C. For processing such feedstock, "Option No. 3," which involves combined use of primary processing with subsequent hydrocracking, visbreaking, or coking of high-boiling fractions and residues, is economically feasible. The main products of this route are motor, process, and boiler fuels. When implementing this option, additional use of the heavy residue oxidation technology can be proposed to produce road and construction petroleum bitumens [21], which in this case can be considered as by-products of the production.

Considering all of the above, it should be noted that the industrial implementation of the procedure for the rapid determination of the rational crude oil processing option depending on the value of its parameter (ε^{t}) is one of the strategically important stages that allow optimizing feedstock (crude oil) utilization, increasing the yield of valuable products, and reducing costs in the petroleum refining industry.

IV. CONCLUSIONS

The conducted theoretical studies have shown that the parameter of the relative dielectric permittivity of crude oil, ε^{t}, can be used during the technological processing of crude oil, particularly for determining the processing scheme option.

It has been experimentally proven that there is a close correlation between the parameter ε^{t}, kinematic viscosity (ν^{20}, mm²/s), and Conradson carbon residue (X_c, wt.%), which is manifested by an increase in the ε^{t} value with increasing ν_{20} and X_c of the crude oil.

A predictive criterion (PC) for the direction of crude oil processing has been developed, which includes the determination of the ε^{t} parameter along with additional quality indicators such as kinematic viscosity and Conradson carbon residue. The practical application of the PC involves the experimental determination of certain threshold levels and allows for the formulation of rational technological crude oil processing schemes considering its electrophysical properties expressed by the ε^{t} parameter.

These schemes, depending on the value of the predictive criterion (PC), can be classified into fuel, lubricating oil, and combined processing options. Depending on the demand for specific petroleum products, the target components obtained by implementing these schemes include hydrocarbon gases, motor and boiler fuels, lubricating oils, petroleum coke,

bitumens, and by-products such as destruction gases, paraffin, resins, and asphaltenes.

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Increasing Power Output in Millimeter and Submillimeter Wave Generators via Surface-Coated Decelerating Structures

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***Abstract*— This paper explores the use of surface coatings on decelerating structures to enhance the power output of millimeter and submillimeter wave generators. By applying metallic and dielectric coatings, we investigate their effects on electron beam dynamics and power conversion efficiency. Experimental results show that surface coatings reduce energy losses and improve electron-beam interactions, leading to significant increases in system efficiency. The study highlights the role of coating properties such as conductivity and thermal management in optimizing performance. These findings provide insights into improving the efficiency and power of high-frequency wave generators, with potential for further advancements through novel materials and design approaches.**

Keywords— backward wave tube (BWT), diffraction radiation generator (DRG), electron beam, slow-wave structure, terahertz radiation.

I. Introduction

Backward wave tubes (BWTs) and diffraction radiation generators (DRGs) are crucial for microwave and millimeter-wave applications in radar, telecommunications, and spectroscopy [1]–[3]. These devices use an electron-optical system with a slow-wave structure (SWS) to enable frequency tuning and power extraction. However, power conversion efficiency remains a challenge, especially at terahertz (THz) frequencies.

BWTs offer broad electronic tunability but typically operate at low power levels in the millimeter-wave range [4]. Efficiency is hindered by the exponential attenuation of the first negative spatial harmonic of the electromagnetic wave (EMW) and the non-uniform electron beam (EB) interaction across its thickness [5], [6]. Only the EB layers near the SWS surface experience strong field interaction, while charge-depleted regions, beam pulsations, and weak impedance coupling further limit power conversion, particularly in the THz range [7], [8]. Additionally, shorter wavelengths impose strict constraints on EB thickness and demand higher current densities, pushing the technological limits of BWTs [9].

DRGs face similar challenges, where space-charge effects and non-uniform field distributions further impede EB interaction and power extraction [7]. Addressing these issues through advanced materials and refined electron-optical designs can enhance high-frequency wave generation. This study explores surface coatings on SWS to improve BWT/DRG efficiency. By analyzing metallic and dielectric coatings, we assess their impact on EB dynamics and power conversion.

The paper is structured as follows: Section II discusses challenges in conventional BWT/DRG and possible solutions; Section III examines EB dynamics in BWT/DRG with a dielectric-coated SWS; Section IV explores periodic SWS construction of the braking electric field for BWTs/DRGs using dielectric coatings; and Section V and Section VI presents the discussion and conclusion, respectively.

II. Challenges in Conventional BWT/DRG and Possible Solutions

A. Development of BWTs and DRGs

To overcome these limitations, optical BWTs known as "Clinotrons" were developed [1], [2] (Fig. 1). In a Clinotron, a ribbon-shaped EB strikes the periodic SWS (comb) at an acute angle, enhancing interaction with the high-frequency field and boosting power generation by engaging all EB layers. A similar inclined injection method was applied to DRGs, significantly increasing output power in the millimeter and submillimeter ranges [3], [6] (Fig. 2) compared to conventional axial EB injection, due to improved energy utilization.

This enhanced power generation in DRGs is attributed to the inclined injection geometry, where the sheet EB impinges on the periodic SWS—such as a comb—at an acute angle φ_0. This ensures that different EB layers sequentially enter the high-intensity electromagnetic field, allowing all electrons in the sheet beam to interact effectively with the high-intensity electromagnetic field on the comb surface, thereby improving energy transfer efficiency.

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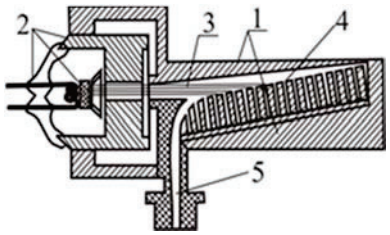


Fig. 1. Schematic diagram of the classical clinotron: (1) vacuum chamber, electron-optical system (2) that forms the paraxial sheet EB (3), periodic SWS (4), where the interaction surface is positioned at an acute angle to the axis of the paraxial strip EB (3), with a thickness equal to or greater than λ, where λ is the average wavelength of the slowed wave across the working range, and the output waveguide (5).

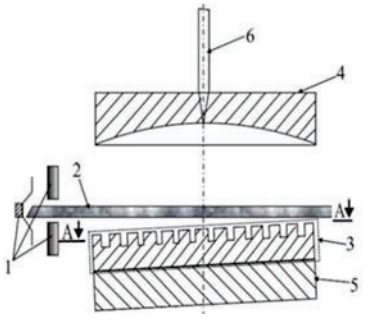


Fig. 2. Schematic diagram of the DRG: (1) electro-optical system ,which emits the electron flow and shapes it into a paraxial sheet EB (2); periodic SWS interaction surface (3), positioned at an acute angle to the EB; and mirror (5), which supports the periodic SWS.

B. Experimental Observations & Design Limitations

Experimental results [6] indicate that, as the wavelength λ decreases in DRGs and Clinotrons, the starting current increases, the initial value of the focusing magnetostatic field rises, and the generator's output power significantly decreases at λ = 0.8 mm. These effects are attributed to increased ohmic losses in the DRG's electrodynamic system and a reduced effective thickness of the interaction space. The effective electron flow thickness Δ_{eff} is defined as:

$$\Delta_{eff} = d_0 - \beta, \tag{1}$$

where d_0 – the period of the slow-wave system, β=v/c – the relative velocity of the EB, i.e., the deceleration factor.

For parameters λ = 0.8 mm and β = 0.1, Δ_{eff} = 0.025 mm. It has been suggested that improving the generator's output characteristics in the submillimeter range requires ultra-thin sheet EBs ($\Delta_{EB} << 0.03$ mm) with increased current density [8]. For thicker beams ($\Delta_{EB} > \Delta_{eff}$) at wavelengths λ < 1 mm, an increase in β —achieved by using higher accelerating voltages—is necessary to maintain efficient energy conversion.

However, these calculations do not account for changes in the structure of the static electron flow within the DRG or Clinotron interaction space, where electrons in the beam move near the surface of the diffraction grating, which is at a high anode potential. Additionally, a significant challenge remains in developing generators for the submillimeter range that can operate with ultra-thin EBs within the interaction space.

C. Possible Solutions

Therefore, investigating new mechanisms and methods for efficiently converting the energy of thick EBs into electromagnetic oscillations in Clinotrons and DRGs operating in the millimeter and submillimeter wavelength ranges is of great importance. A promising approach involves utilizing a non-uniform (profiled) focusing magnetostatic field [7]. In this configuration, electrons from the upper layers of the thick EB are redirected toward the lower layer of effective thickness Δ_{eff}, located near the surface of the diffraction grating. This redistribution enhances the interaction between the thick electron beam and the electromagnetic oscillation field excited in an open resonator, thereby reducing the starting current and increasing the generator's output power [6]–[9].

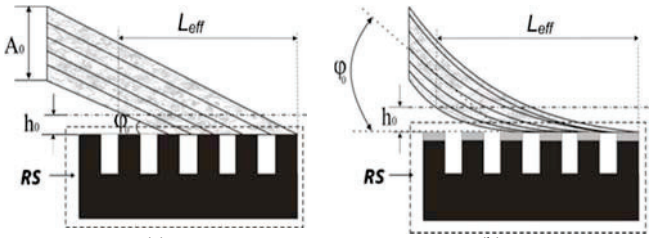


Fig. 3. EB trajectories in DRGs and Clinotrons: (a) in a uniform magnetic field; (b) with a dielectric layer applied to the periodic system; RS — Retarding system; A_0 is the thickness of the EB; h_0, L_{eff} — effective thickness and interaction length of the EB with the field; φ_0 inclined injection angle in Clinotrons and DRGs.

Furthermore, applying a thin dielectric layer to the SWS surface induces an induced braking electric field as the electron beam (EB) deposits onto the layer. Under this influence, the trajectory of the EB becomes parabolic (Fig. 3b). This modification enhances interaction efficiency between the beam and the SWS field, improving energy transfer and thereby increasing both generation power and overall device efficiency [4], [5]. The resulting effect is comparable to that achieved with a non-uniform magnetic focusing field.

A similar approach can be used to expand the beam-field interaction space, further enhancing power output. This method applies to all devices utilizing the 'Clinotron' mode, where the EB is introduced into the interaction space with the slow-wave field to optimize efficiency and performance in the millimeter and submillimeter wavelength ranges.

III. EB Dynamics in a BWT and DRG with a Dielectric-Coated SWS

A. Mechanism of the Effect of Dielectric-Coated SWS on EB

In a Clinotron or DRG, the EB is directed at an angle φ_0 to the SWS (Figs. 1–3). With a dielectric coating, part of the EB deposits onto the surface, forming a braking potential that influences the EB and alters its trajectory. In the absence of a magnetic field, where the electric field has only one component, the system of equations can be written as:

$$\begin{cases} m\frac{dv_x}{dt} = 0 \\ m\frac{dv_y}{dt} = -eE_y \\ m\frac{dv_z}{dt} = 0 \end{cases}, \tag{2}$$

where v_x and v_y are the horizontal and vertical components of velocity, m and e are the mass and charge of the electron, and E_y is the intensity of the electric field generated by the deposition of the EB on the SWS.

After integrating the equations with the given initial conditions [8], we obtain a system of parametric equations describing the motion of an electron in the braking field:

$$\begin{cases} x = v_x t \\ y = \alpha h_0 - v_y t + \frac{a}{2} t^2 \end{cases}, \quad (3)$$

where $a = \frac{eE_y}{m}$ is the acceleration caused by the field generated by the electrons deposited on the dielectric substrate of the SWS, and $\alpha\, h_0$is the impact parameter. For practical calculations, we assume $\alpha \approx 1$, meaning the center of the electron-optical system is located at the boundary of the slow-wave field, which exponentially decays from its surface. However, for simplicity, we assume that it remains constant from the SWS surface up to a height of h_0.

As seen from equation (3), the trajectory of electron motion becomes parabolic, whereas in the classical case (Clinotrons and DRGs), where $a = 0$, it remains linear. Consequently, the interaction space of the EB with the surface wave field of the SWS is limited in length [6]–[9]:

$$L = \frac{h_0}{\sin\varphi_0} + \frac{A_0}{\mathrm{tg}\varphi_0} \quad (4)$$

where A_0 is the EB thickness, h_0 is the thickness of the layer where effective energy exchange occurs between the SWS field and the EB, and φ_0 is the EB inclination angle to the SWS. As the inclination angle increases, EB-SWS interaction is enhanced, leading to increased power generation. However, this also reduces the length of the interaction space, which, at some point, decreases the generated power.

The presence of these two competing mechanisms results in optimal inclination angles for an EB moving linearly in a uniform focusing magnetic field (approximately 6 to 12 arc minutes). The sensitivity of the beam injection angle into the interaction space complicates tuning the device for optimal generation mode, thereby affecting maximum power generation and efficiency.

When a braking electric field is applied, the electron trajectory becomes more gradual. If the potential of the dielectric substrate fully compensates for the vertical velocity component $v_y = v_0 sin\varphi_0$, where $v_0 = \sqrt{\frac{2eU_0}{m}}$ is the electron velocity and U_0 is the accelerating potential (typically 3.5–5 kV for the millimeter range), the braking potential will be approximately $U_y \approx U_0\, sin^2\varphi_0$. This implies that relatively small braking field values, on the order of tens of volts, are required. As the braking potential increases from zero, the initially linear EB trajectory gradually extends due to its parabolic shape. This increases the interaction space length up to four times compared to classical devices. Since output power grows exponentially with the interaction length in the SWS field, this results in a significant increase in generator power and efficiency.

The effective EB-SWS interaction space in the millimeter range (1 mm to 10 mm) has a thickness of approximately 0.031 to 0.31 mm. To ensure efficient energy exchange between the EB and the slow wave, the slow-wave system (SWS) in millimeter and submillimeter-wave generators must span at least 25 to 40 slow-wave periods, corresponding to a length of approximately 25 to 40 mm, which represents the current technological limit. The size constraints of the SWS determine the minimum allowable inclined injection angle in Clinotrons and DRGs. As the wavelength decreases and enters the submillimeter range, the inclination angle φ_0 can drop below 0.1 degrees. Increasing the EB thickness necessitates a longer SWS, but in the submillimeter range, the beam thickness should not exceed 0.1 mm. At such small angles, the total and horizontal dimensions of the effective interaction space become nearly identical.

IV. Design Options for Periodic SWSs in BWTs and DRGs with Dielectric Coatings

In this section, we explore three configurations (see Figs. 4, 6, and 7) of periodic SWSs in Clinotrons and DRGs that incorporate dielectric coatings to enhance EB interaction.

A. Periodic SWS Designs

Configuration 1 (C1): *Reflective Diffraction Metal Grating with Dielectric Coating*. This design features a reflective diffraction metal grating with a dielectric layer applied to its upper surface (Fig. 4). The dielectric layer facilitates the drainage of electrons deposited on its surface, forming a braking electric field that alters the electron beam's trajectory, increasing the interaction length with the SWS and improving overall device efficiency.

Configuration 2 (C2): *Dielectric-Insulated Reflective Diffraction Metal Grating with Parallel Leakage Resistance and Capacitance.* In this configuration, the SWS consists of a reflective diffraction metal grating placed on a dielectric substrate, electrically insulating it from the vacuum-tight housing (Fig. 5). A leakage resistance and capacitance are connected in parallel between the grating and the housing. This setup regulates the discharge time constant of the charge field, controlling the EB trajectory and enhancing energy transfer.

Configuration 3 (C3): *Strip Metal Grating with Dielectric Layer and Metal Screen*. Here, the SWS is designed as a strip metal grating placed on a dielectric layer with a specific thickness ($h_{eff} = \lambda_1/4 + n\lambda_1/2$, where λ_1 is the wavelength in the dielectric, and n = const = 0, 1, 2…), with a metal screen located underneath (Fig. 6). Electrons deposited on the grating elements create a field that alters the electron trajectory, extending it and improving interaction with the SWS field. This design enhances device performance by increasing the interaction space without increasing physical dimensions.

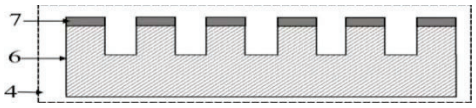


Fig. 4. SWS with a dielectric layer.

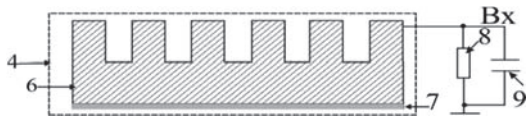


Fig. 5. SWS isolated from the case by a layer of dielectric.

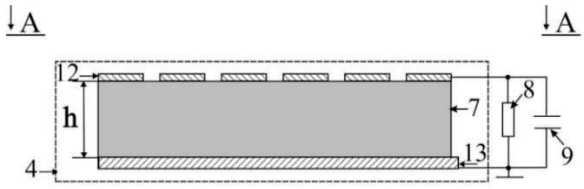


Fig. 6. Ribbon SWS.

B. EB Interaction with SWS

The EB is emitted from the cathode through the electron-optical system (2) (Figs. 1, 2), forming a paraxial sheet EB (3). This beam propagates along the magnetic field lines of the focusing magnetic field at a small angle (~6–12 arc minutes) to the interaction surface of the SWS (4). When the EB deposits some of its electrons onto the dielectric layer (7) of SWS — either applied to the reflective diffraction grating (6) (refer to C1, Fig. 4) or placed between the strip metal grating (12) and the metal screen (13) (refer to C3, Fig. 6)— a negative charge is formed on the surface of the dielectric layer. The EB, moving through the potential field of this charge, is deflected away from SWS. The degree of deflection increases as more electrons accumulate on the dielectric layer, thus enhancing the negative potential on the SWS.

Charge leakage from SWS occurs due to leakage resistance, a result of the imperfect characteristics of the dielectric layer. This resistance can be controlled by selecting dielectric materials with specific electrical properties, such as dielectric constant and loss tangent. In configurations where SWS is a reflective diffraction metal grating (6) (refer to C2, Fig. 5), charge from the dielectric layer is discharged onto the grating. The deposition of the EB onto SWS increases the negative potential, causing further deflection of the beam. However, as charge leakage occurs, the potential decreases, leading to a reduction in beam deflection. The leakage resistance (8) provides a path for charge dissipation from SWS in various configurations (refer to C2, C3, Figs. 5, 6), connecting the strip metal grating (12) to the metal screen (13) and the vacuum-tight housing (1). In parallel with the leakage resistance (8), the capacitance (9) regulates the charge dissipation time constant for the EB.

Over time, a dynamic equilibrium is established, where the rate of charge deposition onto SWS equals the rate of charge leakage. As a result, the trajectory of each electron in the EB becomes more gradual, allowing the electron to remain longer in the braking field of the negative harmonic of the EMW in SWS. This ensures more efficient energy transfer from the EB to the electromagnetic field of the SWS.

Ultimately, utilizing dielectric coatings in the SWS successfully resolves the issue of controlling the EB injection angle into the interaction space. It also increases the effective length of the interaction region between the EB and the electromagnetic field of SWS, thereby improving the generation efficiency and overall performance of the device.

V. Discussion

The use of dielectric coatings on periodic SWSs for Clinotrons and DRGs, as explored both theoretically and experimentally, presents a promising solution to enhance the power conversion efficiency, particularly in the millimeter and submillimeter-wave ranges. The inclined EB injection into the periodic SWS, such as a comb structure, improves the interaction between the EB and the high-frequency field, thereby increasing energy transfer efficiency and output power.

However, while this approach shows clear advantages, the practical application of dielectric coatings introduces several complexities. Theoretical models suggest that optimal coating materials and thicknesses are essential for maintaining effective EB interaction and preventing issues such as impedance mismatching and charge depletion. Experimentally, the performance gains depend on the precise alignment of the EB with the SWS and the careful management of coating properties to minimize energy loss. Moreover, challenges in scaling to THz frequencies, where the required current densities and EB thicknesses become more demanding, highlight the need for further research into advanced coating technologies and the integration of these materials within the constraints of current fabrication techniques.

Thus, while dielectric coatings offer a viable pathway for improving BWT and DRG performance, continued refinement of both theoretical models and experimental designs is crucial for overcoming existing technological limitations.

VI. Conclusion

Thin-film dielectric coatings on SWS in BWTs and DRGs improve EB interaction and efficiency by modifying the EB trajectory and expanding the interaction region without increasing system size. Optimizing dielectric properties like permittivity and loss tangent is essential for precise braking field control. Future research should balance material performance and system complexity to boost efficiency

Acknowledgment

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Software Visual Accessibility: Automated Compliance Testing for Color-Vision Deficiencies

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***Abstract*—This paper presents the results of a study on the feasibility of automatically testing software for compliance with accessibility standards. Such testing is typically performed during the functional testing phase. The relevance of this work stems from the need to ensure that a software's visual elements adhere to accessibility guidelines, including those for users with color vision deficiencies. We analyze the problem, propose a mathematical model for evaluating images against accessibility criteria, and present a methodology for automatic testing using computer vision libraries (OpenCV). We also report the results of testing a prototype tool designed to assess visual elements for color vision impairments. Finally, we discuss how these results can be integrated into continuous integration (CI) systems to automatically detect potential accessibility issues.**

***Index Terms*—accessibility testing, computer vision, functional testing, color vision deficiencies, continuous integration**

I. Introduction

The digitalization of nearly all aspects of daily life has made software accessibility a fundamental requirement. As public services, finance, education, healthcare, and entertainment increasingly move online, the need to accommodate users with visual impairments, including color vision deficiencies (CVD), has grown. Globally, around 1% of women and up to 8% of men experience color perception issues according to Simunovic [1], impacting their ability to use digital services safely and efficiently. Insufficient attention to accessible environmental cues—such as the absence of audible traffic lights, tactile paving, or proper color-contrast settings—significantly increases the physical vulnerability of people with visual impairments [2].

Regulatory frameworks in the European Union, the United States, Ukraine, and beyond—such as the Americans with Disabilities Act (ADA), Section 508, the European Accessibility Act (EAA), and the Ukrainian National Strategy for Creating a Barrier-Free Space—mandate that digital content and services be accessible to people with disabilities, in line with the Web Content Accessibility Guidelines (WCAG) 2.2. These guidelines, based on the POUR principles (Perceivable, Operable, Understandable, Robust), set criteria to ensure that information is not conveyed by color alone, interfaces are keyboard-accessible, content is clear, and technologies remain compatible with assistive tools.

While regulations often focus on public-sector services, the commercial sector faces increasing demand for compliance due to legal risks and user expectations. In practice, accessibility testing now extends beyond manual inspection to automated analysis using AI and computer-vision techniques, especially during functional testing. Automated tools can check color contrast, simulate different types of color blindness, and suggest alternative information channels (e.g., text labels or symbols).

This is particularly relevant for computer games and interactive applications, where color frequently encodes critical information (health bars, navigation cues, object highlights). Recent advances allow testing not only of static screens but also of dynamic and video content, making CI-based accessibility checks feasible.

This paper focuses on methodologies for automated accessibility-compliance testing, particularly for users with CVD. We review the regulatory context, examine state-of-the-art tools and approaches, and present a framework for automated testing using computer vision. Special attention is given to the challenges of testing in interactive and rapidly changing visual environments.

II. Publications and Tools review

A. Publications review

Early identification and remediation of CVD barriers benefit both users and developers: they broaden the potential audience and reduce costly post-release fixes. Automated accessibility testing workflows—spanning simple simulators, contrast checkers, rule-based scanners, heuristic evaluators, and emerging AI-driven algorithms—offer scalable methods to detect and guide the correction of color-related issues.

Color drives usability but poses barriers for those with color vision deficiencies (CVD)—anomalies in the red, green, or blue cones—affecting approximately 8% of men and 0.5% of women (Table I). With no cure for congenital CVD, designers rely on automated tools—simulators, contrast checkers, rule-based scanners, heuristics, and AI—to identify and address color issues.

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TABLE I. COLOR VISION DEFICIENCY PREVALENCE BY GENDER

Deficiency	Cone(s) Affected	Prevalence %		
		Men	***Women***	***Overall***
Anomalous trichromacy				
Protanomaly	Red	1.08	0.03	1.08
Deuteranomaly	Green	4.63	0.36	4.63
Tritanomaly	Blue	0.0002	0.0002	see Tritanopia[a]
Dichromacy				
Protanopia	Red	1.01	0.02	1.01
Deuteranopia	Green	1.27	0.01	1.27
Tritanopia	Blue	0.0001	0.0001	0.20[b]
Monochromacy				
Green-cone monochromacy	Red, Blue	—	—	1 in 1 000 000
Red-cone monochromacy	Green, Blue	—	—	1 in 1 000 000
Blue-cone monochromacy	Red, Green	—	—	1 in 100 000
Rod monochromacy	Red, Green, and Blue	—	—	1 in 33 000–50 000

[a]Not reported separately for anomalous trichromacy.

[b]Equivalent to 0.20%.

1) Assistive image-processing systems

Feng et al. [3] apply a hue-rotation of the H channel in HSV space via OpenCV to improve color discriminability for achromat observers, though they report reduced effectiveness in natural scenes with complex palettes. Meng et al. [4] present a Raspberry Pi-based prototype that performs HSV thresholding, color filtering, and on-screen text feedback; however, it was evaluated only on controlled color patches and benchmark shapes, lacks support for dynamic content (animations or game frames), offers no integration into web/mobile or design tools, and provides no performance metrics for real-time use.

2) Simulation and heuristic-based evaluation

Bonacin et al. [5] propose an ontology-driven framework for per-user recoloring. Angerbauer et al. [6] report on a large-scale, crowd-sourced evaluation of academic figures under CVD simulations. Alcaraz-Martínez et al. [7] develop 18 heuristics to evaluate statistical charts for CVD and low-vision users. These approaches are valuable but typically require expert or crowd validation, demand ongoing maintenance as standards evolve, and suffer from variability in subjective ratings.

3) Automated contrast and rule-based checkers

Jamil et al. [8] compared OS high-contrast modes with custom "CVD themes," finding that although high-contrast settings improve text legibility, they often degrade aesthetic appeal and remove visual context. Ara et al. [9] survey existing automated accessibility tools—classifying them as declarative, ontological, algorithmic, or hybrid—and reveal that only 13% support end-to-end automation. They further observe that most tools focus on basic WCAG contrast checks and omit non-text elements (such as charts and icons); that rule-based methods lack semantic understanding (resulting in false positives and negatives); that ontological approaches require continuous maintenance; and that fixed thresholds fail to account for individual CVD variations.

4) Impact on user experience and performance

Napoli & Chiasson [10] conducted an empirical study with the Spectrum Chrome plugin in the browser game *Pet Rescue Saga*, finding that while some players maintained accuracy and completion times, most reported increased difficulty and reduced confidence under CVD simulation. Tillem & Gün [11] survey the gaming accessibility literature, noting that although performance impacts vary, users with CVD consistently report poorer subjective experiences and call for improved design guidelines. Wang et al. [12] introduce the "Color Blindness Bartender" VR app—which maps RGB to LMS and applies matrix-based color removal—and show via task-accuracy scoring that VR is effective both for educating sighted users about CVD challenges and for prototyping more accessible interfaces.

5) Screening and diagnostic tools

Fliotsos et al. [13] compare the Eye Handbook mobile app to Ishihara plates in full-color, grayscale, and red-green simulation modes, finding overall parity but plate-specific discrepancies. Abbas et al. [14] develop an Android application with seven assessment modules (color detection, object size, motion, visual field), though they omit detailed accuracy and clinical validation metrics. Laskowski [15] presents a four-stage HTML5/JavaScript screening game tested on 143 participants; while an error threshold reduced false positives, further calibration for sensitivity versus specificity and longitudinal validation remain necessary.

B. Tools Review

A variety of tools support automated CVD accessibility testing (Table II). Desktop simulators like Color Oracle offer static filtering in four CVD modes but lack video or real-time integration. Mobile apps such as Chromatic Vision Simulator provide live camera-based filters for all major types but cannot process recorded gameplay. Additional aids include Sim Daltonism, WebAIM Contrast Checker, and built-in OS color filters—each useful yet limited in automation, scope, or integration.

Despite a growing toolkit for CVD accessibility testing, significant gaps remain. Current solutions largely ignore non-text elements such as charts, maps and complex graphics, and they apply one-size-fits-all rules without adapting to individual user profiles. Promising hybrid human-AI approaches (e.g.,

TABLE II. COMPARISON OF CURRENT TOOLS AND TECHNOLOGIES FOR AUTOMATED COLOR VISION DEFICIENCY ACCESSIBILITY TESTING

Tool	Type	Features	CVD Support	Automation	Limitations
Color Oracle	Simulator	4 CVD modes	All types	Manual	No reports
WebAIM Contrast Checker	Contrast checker	WCAG contrast	General	Semi-auto	Text only
WAVE / Axe / Lighthouse	Scanner	WCAG audits	General	Full-auto	Limited scope
Daltonization algorithms	Recolor filter	Hue shifting	Prot/Deut	Auto	High cost
colorBrewer / VizPalette	Design aid	Safe palettes	Common types	Semi-auto	No testing

crowd-sourced evaluations) have yet to be integrated into developer workflows, and there is little support for embedding CVD checks directly into IDEs, design tools or CI/CD pipelines. Finally, the field lacks large-scale, longitudinal studies to validate automated methods against real user outcomes and drive future improvements.

III. THEORETICAL BACKGROUND OF BUILDING AN AUTOTESTING SUBSYSTEM

To simulate different types of color blindness, the choice of color space is crucial. Typically, in computer systems, images are stored in the sRGB format, where each element represents the value of one of the colors: red (R), green (G), or blue (B) for each pixel of the image. The screenshots of the image are stored as a matrix that encodes the color of each pixel.

To process the image in the context of dichromacy or trichromatism simulation, it is necessary to convert the image from RGB format to the LMS color space. This space is based on the reactions of three types of photoreceptors in the human eye:

- L: receptors sensitive to long wavelengths (red),
- M: receptors sensitive to medium wavelengths (green),
- S: receptors sensitive to short wavelengths (blue).

Converting the image from one color space to another is necessary to model color vision deficits. For instance, protanopes, deuteranopes, and tritanopes lack L-, M-, or S-cone receptors, respectively. Therefore, a natural approach to simulating dichromacy is to convert the image to the LMS color space and remove the component corresponding to the absent receptors. The conversion from RGB to LMS is achieved using a linear transformation (matrix multiplication), and there are multiple ways to perform this process [16], [17].

The conversion from sRGB to linear RGB is performed by removing the gamma correction applied in the sRGB color space. For each component C_{sRGB}, the conversion to linear RGB C_{linear} is given by:

$$C_{linear} = \begin{cases} \frac{C_{sRGB}}{12.92}, & \text{if } C_{sRGB} \leq 0.04045, \\ \left(\frac{C_{sRGB}+0.055}{1.055}\right)^{2.4}, & \text{if } C_{sRGB} > 0.04045. \end{cases} \quad (1)$$

This formula ensures that the gamma-corrected sRGB values are converted back to a linear space for accurate processing.

The formula for converting from RGB to LMS is as follows (based on [16], [17]):

$$\begin{bmatrix} L \\ M \\ S \end{bmatrix} = \begin{bmatrix} 0.31394 & 0.63957 & 0.04652 \\ 0.15530 & 0.75796 & 0.08673 \\ 0.01772 & 0.10945 & 0.87277 \end{bmatrix} \begin{bmatrix} R \\ G \\ B \end{bmatrix}, \quad (2)$$

where L, M, and S represent the long, medium, and short wavelength cone responses, and R, G, and B represent the red, green, and blue components in the RGB color space. The reverse conversion (LMS to RGB) is performed using the inverse matrix:

$$\begin{bmatrix} R \\ G \\ B \end{bmatrix} = \begin{bmatrix} 5.4722 & -4.6416 & 0.1694 \\ -1.1252 & 2.2939 & -0.1679 \\ 0.0298 & -0.1932 & 1.1636 \end{bmatrix} \begin{bmatrix} L \\ M \\ S \end{bmatrix}. \quad (3)$$

After converting the original image (1, 2), we get the color in the LMS space by replacing the value of one of the waves depending on the type of trichromacy or dichromacy with the help of defined color matrices. Thus, depending on the specified type of dichromacy (protanopia, deuteranopia, or tritanopia), we simulate how color perception changes. For protanopia and deuteranopia, a plane is calculated that reflects the perception of red, green, and blue colors, and a color conversion is applied within this plane. For tritanopia, a similar process is used, but for the perception of red, green, and yellow.

After simulating the selected type of dichromacy or trichromacy, the image colors are converted back to RGB/sRGB (3) or another format for display and further visual analysis.

In the LMS color space, colors that differ only in the component associated with the missing receptor cannot be distinguished by a dichromat. This leads to the formation of the so-called “confusion lines,” which are parallel to the axis of the missing receptor in the LMS space. According to this model, protanopes cannot distinguish between colors that differ only in the L-component, deuteranopes—in the M-component, and tritanopes—in the S-component.

To illustrate this phenomenon, we plotted graphs showing several confusion lines in LMS space alongside their corresponding representations in RGB space for both deuteranopia and tritanopia. As a source image, we used a full-color photograph of tulips against a blue sky, which provides a rich

variety of hues and gradations and thus clearly demonstrates the effects of color blindness.

For sRGB color space it represents intensity a color in the range from 0 to 1, for LMS color space it represents the response of cones sensitive to a wavelength in the human eye.

At first glance, the colors look different and are very easy to distinguish (Fig. 1). But if we apply a deuteranopia lens, we get colors that are very similar to each other (Fig. 2).

- RGB Space: All three trajectories bend and converge toward the G and B axes, forming nearly parallel paths;
- LMS Space: Points collapse along the confusion axis (L–S), forming thin, almost straight ribbons—explaining why observers with green-cone deficiency cannot distinguish "blue→magenta", "blue→pink", or "green→yellow".

An even more striking illustration is provided by the tritanopia simulation (Fig. 3), where in RGB space the confusion lines tilt toward the blue (B) axis, and in LMS space collapse along the S axis, rendering the "blue→yellow" gradients virtually indistinguishable.

IV. Implementation and Testing

The testing program architecture consists of five main modules:

- Image processing: This class deals with the preparation and subsequent storage of the image result. The main functions are converting the image from BGR space to RGB space, scaling the image (if necessary), converting the image to BGR space from RGB space for correct saving. It interacts with the interface class, taking input data for working with the image and simulation, and also sends the image with a lens for previewing to the interface field. It also interacts with the simulation class by passing images for applying the color blindness lens.
- Video processing: The component divides the video into frames and records the result in a separate video with the applied lens. The main functions are converting video to an array of images, converting each image from BGR space to RGB space for lens application. It interacts with the interface class, taking input data to work with video and simulation. It also interacts with the simulation class by passing each frame for color blindness lens application in a group of 10 frames at a time to speed up the lens application.
- Processing the region around the cursor: This class captures a specific region around the cursor and applies a color blindness lens to it. Its main functions are to capture a region of the screen every few milliseconds and apply a lens to the captured image. It interacts with the UI class, taking input data to work with the simulation, and when the image simulation is complete, displays the image in a special user interface element. It also interacts with the simulation class by passing each image to a region of the screen and applying a color blindness lens to it.
- Interface: This class stores all the information provided by the user and also allows changing the interface settings for people with visual impairments. It interacts with all three image processing classes, passing the input data to them.

To implement the system, we chose the Python3 programming language, OpenCV for image processing, Matplolib for debugging, and other necessary libraries such as NumPy. We also use Tesseract to work with text.

A. Testing Text Fragments

To analyze the desired text, we first get the RGB values of the background and text color channels. Then we divide each color by 255. Next, we find the linear value of the color: if the color value is less than or equal to 0.03928, then its linear value is the division of the color value by 12.92, and if it is greater, then add 0.055 to the color value, then multiply the result by 1.055 and raise this number to a power of 2.4::

$$f(c) = \begin{cases} \frac{c}{12.92}, & \text{if } c \leq 0.03928, \\ \left(\frac{c+0.055}{1.055}\right)^{2.4}, & \text{if } c > 0.03928. \end{cases} \tag{4}$$

After that, multiply the red color by 0.2126, the green by 0.7152, and the blue by 0.0722. Now you need to check which value is larger—the background or text value. Add 0.05 to each value and divide the larger value by the smaller one. The result will be the contrast ratio between the two colors. The value must be greater than 3 if it is a headline and greater than 4.5 if it is plain text.

Automatic testing of text fragments, especially for computer games, is a complex task that can be addressed by training artificial neural networks to extract text fragments and subsequently process them with predefined algorithms to determine readability, color contrast ratio, and more. Currently, this functionality is only implemented for manual testing.

Optionally, it is possible to analyze text fragments for colors compliance using text fragment screenshots in the current implementation.

B. Testing Image Colors

We use the classic Viénot-Brettel-Mollon [18], Machado et al. [19], and Brettel et al. [20] algorithms for CVD simulation, as they remain widely used and produce accurate results.

To process an image or video, first load it with `cv2.imread(path)`, which returns a NumPy array of pixel values in BGR order. For CVD simulation, convert BGR to RGB before applying the filter, then convert back to BGR before saving.

When processing a video, apply the same procedure to each frame and write the results to a new video file. The key challenge is selecting the region containing text or other elements to test; once selected, the algorithm crops that region and applies the chosen CVD filter.

We support four types of red-green simulations: deuteranopia, deuteranomaly, protanopia, and protanomaly. After



Fig. 1. Source image sRGB and LMS color spaces.



Fig. 2. Deuteranopia lens applied to sRGB and LMS.



Fig. 3. Tritanopia lens applied to sRGB and LMS.

capturing an image or frame, the CI-script chooses the simulation algorithm. For deuteranopia and protanopia, we use the Viénot-Brettel-Mollon method; for deuteranomaly and protanomaly, we use Machado et al.'s model, with a severity parameter (1.0 for full dichromacy, less than 1.0 for anomalous trichromacy).

For blue-yellow simulations (tritanopia and tritanomaly), we apply the Brettel et al. algorithm. No model can perfectly replicate another person's perception, so each algorithm offers an approximation of CVD vision.

To implement automatic testing, the following steps were taken:

- Determine the number of color clusters in LAB space (e.g., 10), potentially normalizing image sets to align palettes.
- Convert images from sRGB to LAB.
- Segment images in LAB space using KMeans (scikit-learn).
- Apply color masks and compute the percentage of pixels in each cluster.
- Simulate color blindness by applying the corresponding filters (e.g., protanopia, deuteranopia).
- Optionally apply additional processing algorithms.
- Output and/or log results, including graphical summaries (e.g., Fig. 3 and Fig. 4).

Based on expert analysis, a threshold of 40 was chosen to flag images that may violate accessibility standards. This reduces false positives and accelerates testing. It also enables

fully automated testing within CI pipelines.

```
Enter the path to the image: img/flowers2.jpg
#2f5e84 -3.4708988520339714 -25.97769289821462 blue
#144f1c -30.798613068524688 24.32869660712743 green
#d9f0fa -5.558002683951402 -7.368274919670648 other
#142dab 40.48231750129913 -68.55867797188996 red
#1b484b -14.222582074747637 -6.429113084723792 other
#2e5ac9 24.754720815084593 -62.029720393667084 red
#0a201b -10.371535170734997 0.7416629472377867 other
#10206c 24.683982992937615 -45.716817270042284 red
#2b7828 -40.2462923691671 35.97720496555401 green
#5d98e5 3.6378877710778745 -44.73072024316267 blue
29 is a total yellow-red area
37  is a total blue-green area
No warning
```

Fig. 4. No warning for score below threshold.

```
Enter the path to the image: img/flowers1_original.png
#456618 -25.140866882834707 37.99740489167193 green
#889259 -13.126739634232887 29.01599671126991 yellow
#a6b2bd -1.9363073664985242 -7.018438381513814 other
#f6ce0f -1.0631237582166553 83.15943625457918 yellow
#4780b3 -3.0425438700277807 -32.53670535550837 blue
#819c0e -26.575507907162677 60.90689427846472 yellow
#d29913 11.191784310161824 68.3270601319088 yellow
#e1dcd5 0.479347186166601 4.039183823264159 other
#d23e0c 56.208885052819845 56.957103835902814 other
#e2c369 -0.015996568319853743 48.78921371211411 yellow
50 is a total yellow-red area
16  is a total blue-green area
Protanopia warning
Deuteranopia warning
```

Fig. 5. Warning for score above threshold.

CONCLUSIONS

This paper analyzes the current state of software development with respect to accessibility standards for users with color vision deficiencies. This issue is particularly significant for software targeting European (including Ukrainian) and North American markets, where 7-9% of the population is affected by CVD [21]. We propose an automated testing approach that transforms images from sRGB to LAB color space, applies clustering, and computes an integral indicator based on the percentage of colors falling into critical zones (yellow-red and blue-green).

A prototype implementation has been developed and tested, demonstrating the feasibility of integrating the proposed method into CI workflows.

Future research will focus on automating color-contrast testing of text in screenshots—especially in applications such as computer games, which often feature low-contrast textual elements—and on integrating neural-network-based algorithms (e.g., CUD-Net [22]) as modules in CI systems for continuous accessibility validation.

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Simulation and Performance Analysis of Acoustic Direction Finding Systems for Small UAVs

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***Abstract*— This paper explores the need for effective acoustic detection of unmanned aerial vehicles (UAVs), focusing on mitigating background noise interference and enhancing signal-processing algorithms. Simulation modeling evaluates acoustic direction-finding systems under varying noise conditions, demonstrating high detection accuracy at favorable signal-to-noise ratios (SNR), with performance degrading as noise increases.**

Keywords*— acoustic detection, UAV localization, noise interference, bearing estimation, cross-correlation, SNR, simulation modeling, time delay estimation.

I. Introduction

The early development of unmanned aerial vehicle (UAV) technology dates back to the early 20th century, when UAVs were primarily employed for military reconnaissance and surveillance. Over the past several decades, rapid technological advancements have enabled UAVs to expand into diverse civil and commercial applications— ranging from agriculture and disaster management to environmental monitoring and journalism. However, the dramatic increase in UAV deployment has raised significant concerns regarding airspace safety, privacy, and public security [1].

Traditional detection approaches based on radar, optical, and radio frequency (RF) methods have proven effective under certain conditions but often face limitations in adverse weather, low visibility, and electromagnetic interference. In contrast, acoustic detection techniques offer a passive, cost-effective alternative by exploiting the unique sound signatures generated by UAV propulsion systems and aerodynamic interactions [2]–[7]. Acoustic sensors operate independently of light and are generally less susceptible to RF interference; however, their performance is challenged by ambient noise and complex environmental conditions, which can lead to errors in direction-of-arrival (DOA) estimation and target classification.

Recent research has focused on enhancing acoustic detection through advanced signal processing and machine learning methods. For instance, adaptive filtering combined with deep learning-based denoising has shown promising improvements in low signal-to-noise ratio (SNR) environments [8], [9]. Additionally, simulation modeling of acoustic direction finders has become a critical area of study, enabling the optimization of sensor array configurations and real-time processing algorithms for robust UAV detection.

In this work, we present a comprehensive simulation modelling study and performance analysis of acoustic direction-finding systems (DFSs) specifically designed for small UAVs. We evaluate both parametric and non-parametric signal processing approaches under various noise conditions to assess detection accuracy and localization performance. By integrating insights from the latest publications and state-of-the-art techniques, our study contributes to the development of robust acoustic detection systems capable of operating effectively in challenging real-world scenarios.

II. Theoretical Basic of Acoustic Direction Finding

Acoustic direction finding exploits the propagation delays of sound waves captured by an array of microphones positioned at known distances. When a sound wave impinges on the array, the time delay between the signals received by individual microphones depends on the angle of incidence. This delay is directly related to the bearing—or azimuth—of the sound source.

Fig. 1 shows a schematic diagram of an acoustic DFS with two microphones, M_0 and M_1, oriented at a specified angle relative to the x-axis. The time delay τ between the signals received by M_0 and M_1 is used to estimate the azimuth α of the source O relative to the baseline connecting the sensors. Mathematically, the time delay τ between two microphones separated by a baseline distance d can be expressed as

$$\tau = d/v \sin(\alpha), \quad (1)$$

where v denotes the speed of sound in air and α is the angle between the DOA and the normal (perpendicular) to the baseline connecting the microphones, as shown in Fig. 1.

By inverting (1), the azimuth α of the sound source can be determined from the measured delay:

$$\alpha = \arcsin(c\tau/d). \quad (2)$$

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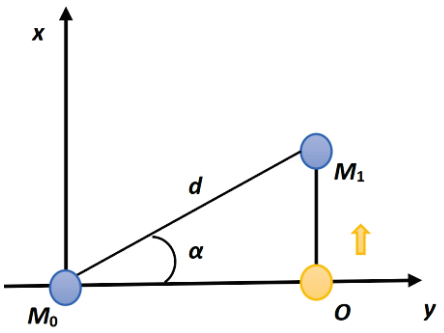


Fig. 1. Schematic diagram of an acoustic DFS consisting of two microphone M_0 and M_1

Prior studies [10]–[13] have demonstrated that the minimum possible bearing error in acoustic DFSs increases as the source direction deviates from the baseline normal, following an inverse cosine relationship, specifically proportional to $1/\cos(\alpha)$. For instance, a deviation of 60° from the baseline normal effectively doubles the error. Consequently, to maintain acceptable accuracy, these systems typically operate within sectors of ±60° from the baseline normal, encompassing a total angular range of 120° [10], [11].

III. DESIGN AND IMPLEMENTATION OF SIMULATION EXPERIMENTS FOR EVALUATING THE PERFORMANCE OF ACOUSTIC DFSS

In this section, we describe the simulation experiments designed to evaluate the bearing accuracy of our acoustic DFS. The step-by-step flow map of the experimental design for evaluating acoustic DFSs is shown in Fig. 2. The system uses two orthogonally oriented microphone arrays, and the experiments focus on assessing the effect of signal-to-interference ratios (SIRs) and the efficacy of our correction method on bearing estimation. More detailed descriptions of the experimental setup, signal preparation, noise modeling, performance evaluation and simulation results are provided in the following subsections.

A. Signal Acquisition and Preprocessing

The simulation models an acoustic DFS that employs two crossed microphone arrays positioned at a 90° angle. Each array consists of four microphones, with inter-element spacings of 0.5 m and 1 m, respectively. The azimuthal bearing is estimated by computing the time delay between microphone pairs via cross-correlation. For each pair, the cross-correlation function (CCF) is computed over a segment of N_0 samples, and the time delay corresponding to the maximum value is used to derive the bearing. The CCF is defined as:

$$R_{\max}(k) = \max_{k} \sum_{i=lN_0}^{(l+1)N_0} S_1(i)\, S_2(i+k) \qquad (3)$$

where $S_1(i)$ and $S_2(i+k)$ represent are the signals from the first and second microphones, respectively, k represents the time shift between these two microphones/channels in samples, l is the segment number, and $R_{\max}(k)$ is the maximum value of the CCF, which is achieved with a shift of k samples.

In this study, acoustic signals from UAVs were recorded at a sampling frequency of 44.1 kHz. To simulate the object's bearing, an artificial delay was introduced in one channel. Four scenarios were designed to evaluate the performance of the DFS under different conditions (as illustrated in Fig. 2):

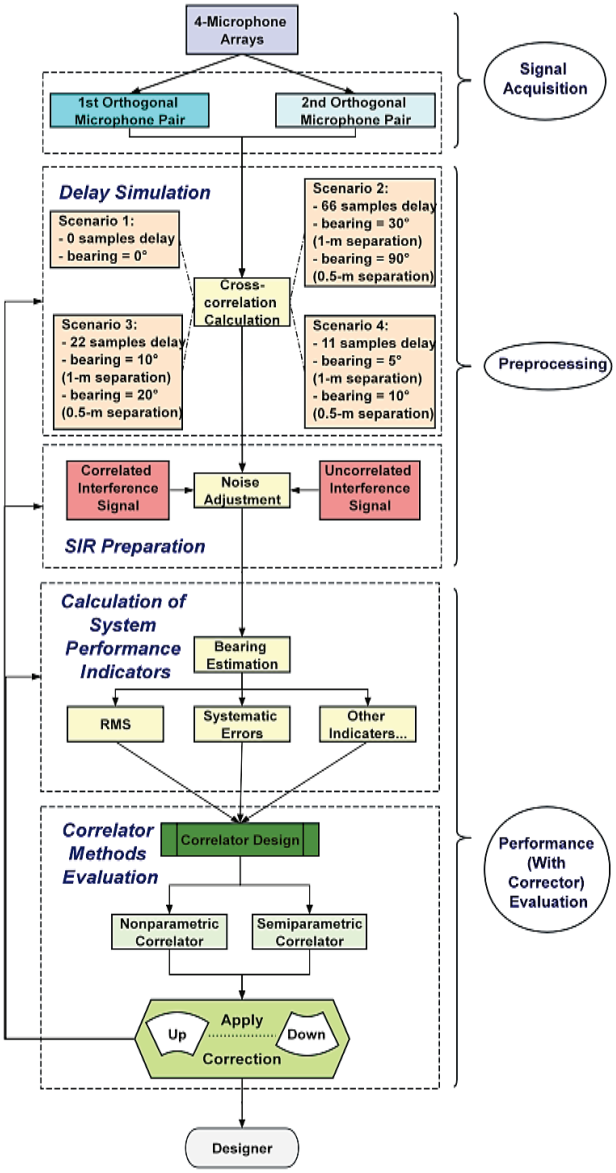


Fig. 2. Flow Map of the Experimental Design for Evaluating Acoustic DFSs

Scenario 1 (S1): The signals from both channels/ microphones have no time difference, representing an object positioned perpendicular to the centre of the microphone array (bearing = 0°).

Scenario 2 (S2): The second channel signal is delayed by 66 samples relative to the first. This corresponds to a bearing of approximately 30° for microphones with a 1-meter separation and 90° for those with a 0.5-meter separation.

Scenario 3 (S3): The delay between channels is 22 samples, corresponding to a bearing of approximately 10° for a 1-meter separation and 20° for a 0.5-meter separation.

Scenario 4 (S4): The second channel is delayed by 11 samples, corresponding to a bearing of approximately 5° for a 1-meter separation and 10° for a 0.5-meter separation.

The next step in signal preparation involved adding the interference signal to the desired signal to obtain a resulting signal with a specified SIR. This was done by forming an additive mixture of the signal and interference:

$$U_i = S_i + n_i\,, \tag{4}$$

where the acoustic signals of the useful object, S_i, and wind noise, n_i, have respective variances:

$$\langle S_i^2 \rangle = \sigma_S^2,\ \langle n_i^2 \rangle = \sigma_n^2. \tag{5}$$

The SNR is given by:

$$\mu_0 = \frac{\sigma_S^2}{\sigma_n^2}\,. \tag{6}$$

To achieve the required SNR, μ, the following condition must be met:

$$\mu = \mu_0\, K_0^2 \tag{7}$$

where K_0 is the correction factor by which the interference signal is scaled. This factor is given by:

$$K_0 = \sqrt{\frac{\mu}{\mu_0}}. \tag{8}$$

The desired SIR is achieved by either adjusting the useful signal before mixing or applying a correction factor to the interference signal prior to addition.

In this study, the next step in signal preparation involved adding interference to the useful signal to maintain a specified SIR, with a particular focus on the impact of wind noise, which exhibits significant non-stationary and non-Gaussian characteristics. Specially, two scenarios were considered:

Correlated Interference: The interference is identical in both direction-finding channels, enabling accurate estimation of the object's direction. The useful signals in both channels remain the same, differing only in time delay. The interference is adjusted accordingly to maintain the desired SIR.

Uncorrelated Interference: Independent interference signals are generated for each channel. Correction factors are computed separately for each segment using equation (8) to ensure equal SIRs in both channels before performing direction-finding calculations.

The study analyzed RMS and systematic errors in direction estimation, both with and without correction. The simulation results, involving wind noise (in the absence of gusts) as correlated and uncorrelated interference, are presented in the following two subsections: (1) without the corrector and (2) with the corrector.

B. Simulation Results Without Corrector

Simulations were conducted in MATLAB using segment lengths of 4096, 16,384, and 32,768 samples, corresponding to time durations of 92.9 ms, 0.327 s, and 0.743 s, respectively, across four scenarios (S1, S2, S3, S4). To account for variations in noise characteristics, noise components were categorized based on their homogeneous properties, simulating the DFS under both gusty and calm wind conditions. Bearing estimation accuracy was evaluated over SNRs ranging from 60 dB to -40 dB in 20 dB increments. Wind noise in the absence of gusts was specifically considered, and experimental analysis confirmed that it could be approximated as locally Gaussian during these phases.

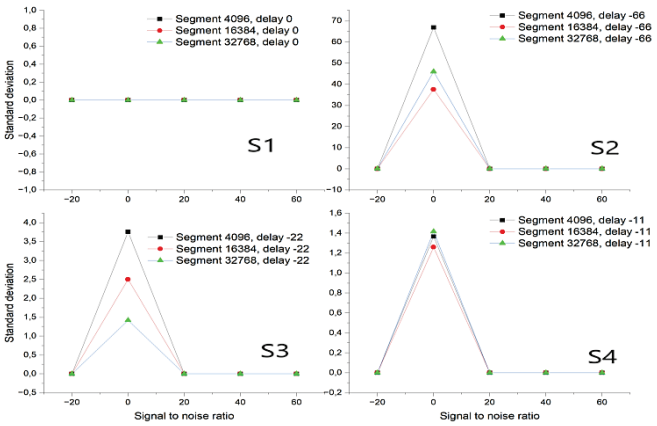


Fig. 3. Dependence of the standard deviation of the determined bearing on the SIR, with correlated interference

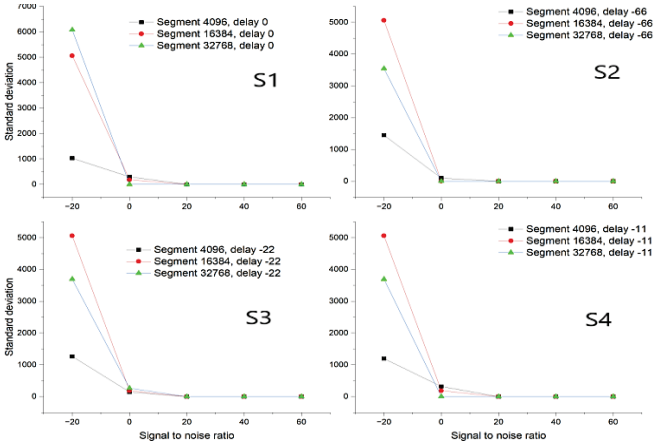


Fig. 4. Dependence of the standard deviation of the determined bearing on the SIR, with uncorrelated interference

Two representative results without corrector are shown in Figs. 3 and 4. For uncorrelated interference, the DFS showed consistent performance across all four scenarios, with standard deviation varying with SNR in the range of -20 dB to 20 dB, while for correlated interference, performance remained symmetrical.

C. Simulation Results with Corrector

In non-Gaussian interference scenarios, nonparametric detection methods have shown marked advantages over conventional linear parametric approaches. This study investigates nonparametric and semiparametric direction finding by converting the input signal into a binary quantized representation for both or only one microphones.

In particular, let the input signal at a microphone be denoted by $S_{1,2}(i)$. Its binary quantized version is given by

$$q(S_{1,2}(i)) = \begin{cases} +1 & \text{for } S_{1,2}(i) \geq 0 \\ -1 & \text{for } S_{1,2}(i) < 0 \end{cases}. \tag{9}$$

When both channels are quantized, say $S_1(i)$ and $S_2(i)$, the processing generates a signed correlator that implements a fully nonparametric detection scheme. The correlator output can be expressed as

$$C(k) = \frac{1}{N_0} \sum_{i=lN_0}^{(l+1)N_0} q(S_1(i)) q(S_2(i+k)). \tag{10}$$

In contrast, if only one channel (e.g., $S_1(i)$) is quantized while the other channel $S_2(i)$ remains in its analogy form, the method becomes semiparametric-combining a nonparametric transformation on one channel with a parametric model on the other. In this case, the signed analogy correlator is computed as:

$$C(k) = \frac{1}{N_0}\sum_{i=lN_0}^{(l+1)N_0} q(S_1(i))S_2(i+k). \qquad (10)$$

Simulation results for the performance evaluation after applying nonparametric and semiparametric corrector are shown in Figs. 5 and 6, respectively.

IV. Discussion

This study investigates the performance of acoustic DFS for UAV detection, focusing on the impact of SNR and environmental noise. Results show that bearing accuracy improves with higher SNR, with significant errors observed below 20 dB, emphasizing the need for better noise filtering. Correlated noise and gusty wind conditions further degrade accuracy, indicating the importance of sensor optimization for varying environmental factors.

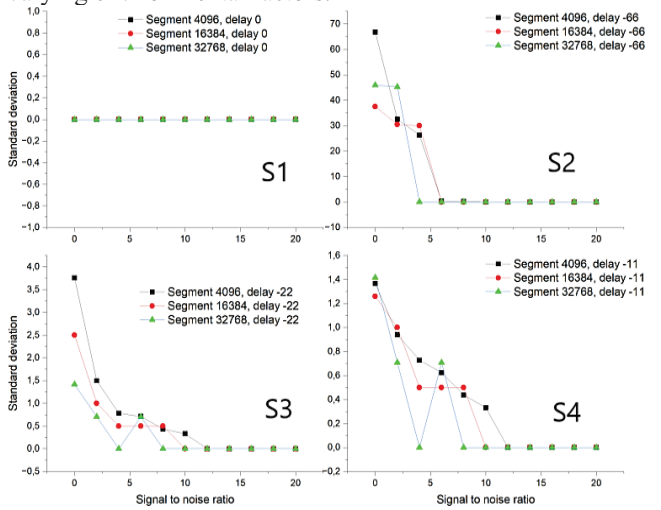


Fig. 5. Dependence of the standard deviation of the determined bearing on the SIR, with correlated interference, when using a nonparametric correlator

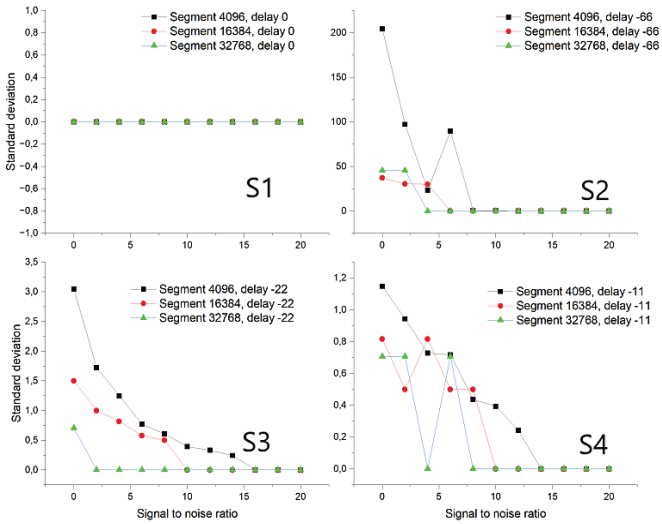


Fig. 6. Dependence of the standard deviation of the determined bearing on the SIR, with correlated interference, when using a semiparametric correlator

Nonparametric methods demonstrated superior performance, offering better efficiency in low-SNR scenarios. Increasing segment length (>4096 samples) enhanced noise immunity. To further improve system performance, future work should explore deep learning-based noise suppression and optimal sensor array configurations.

V. Conclusion

This study highlights the impact of SNR, noise type, and wind conditions on acoustic DFS performance for UAV detection. Nonparametric methods excel in low-SNR environments, and longer segment lengths improve noise immunity. Future work should focus on integrating advanced noise suppression and sensor optimization to further enhance system robustness in real-world scenarios.

Acknowledgment

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A Market-Driven Assessment of Industry 4.0 Competencies in EU Electrical Engineering Curricula

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***Abstract*—In order for present-day electrical engineering programs to stay competitive and prepare graduates for successfully entering the workforce, they need to respond to changing market demands. This study aims to evaluate alignment between electrical engineering job postings and a number of study programs in STEM-focused higher education institutions in the EU. To achieve this, a thematic analysis of Industry 4.0 competencies in recent electrical engineering job postings is performed. Identified competency framework is applied to understand the supply-demand relationship between the study programs and the market. The findings reveal an over-supply in the core curricula, as well as strong supply of Industry 4.0-related knowledge. A subset of under-supplied but highly-demanded skills was also identified: self-management, communication, English language knowledge, collaboration, continuous learning, and knowledge transfer. The practical value of the study lies in providing practical insights for electrical engineering curriculum development and introducing a replicable data-driven method for benchmarking formal curricula against job market requirements.**

Keywords—Industry 4.0, electrical engineering education, higher education, curriculum development, labor market

I. Introduction

Industry 4.0 (I4), also referred to as the Fourth Industrial Revolution (IR4) has become a major transformational force for industrial environment of the 21st century. One of its defining characteristics, along with digitalization and reliance on modern technologies, is a surge in speed, scope, and systematic influence of change. This creates a challenging environment for all stakeholders—employers, workers, managers, regulators, consumers—pressing them to react and adapt accordingly [1-2].

The Industry 4.0 model is accompanied by heavy digitization of factory operations, introduction of inter-connected manufacturing components, and transformation of business processes to support flexible distributed operations [3]. An I4-ready "smart factory" incorporates a number of complex technologies, such as cyber-physical systems (CPS) and internet of things (IoT) [4] to transform its operations and achieve higher productivity, among other benefits. Other technologies often discussed in I4 manufacturing context are: additive manufacturing; big data and analytics; digital twins; artificial intelligence and machine learning (AI/ML); cloud computing; virtual and augmented reality (VR/AR); blockchain and distributed ledger technology (DLT); robotics; and cybersecurity [5].

Despite such a wide array of digital technologies shaping present-day manufacturing, the interconnected nature of Industry 4.0 with its emphasis on cross-hierarchy communication and increased stakeholder involvement [3] requires strong non-technical competencies for workers to succeed. Skill sets that were viewed mainly as requirements for knowledge workers are now an integral part of engineer's day-to-day work. A present-day engineer is expected to not only work with machines, but also design and integrate computerized systems, work with data, collaborate on projects, and take on leadership when necessary [6]. This applies not only to emerging, "I4-native", roles—existing jobs are transformed as well [7]. This puts developing relevant skills into perspective not only for graduates who are yet to enter the workforce, but employees already involved into the industry.

In this environment, engineering education is at the forefront as the work-readiness of future graduates depends on the content of today's study programs. Higher education institutions face changing requirements across the board: the content provided, the technologies used for teaching, and the need to provide students with experience interacting within interdisciplinary teams among other requirements [8]. This necessitates closer cross-sector collaboration and deriving insights in technology foresight from private sector and larger ecosystem actors [9].

As shown in a systematic review [10], Industry 4.0 has introduced significant skill gaps across a number of sectors. Recent research of engineering competencies confirms this and demonstrates lack of consensus on the unified competency framework for engineers to excel in I4 environment, leading to weaker alignment of study programs with current labor market demands [11-12]. This complicates curriculum modernization and creates tension between the need innovate and isomorphic pressures to make similar decisions [13].

Furthermore, despite the ongoing discourse, there exists limited evidence on Industry 4.0 skill gaps in electrical engineering (EE) higher education curricula, especially within the EU. The present study aims to address this by evaluating the alignment between Industry 4.0 competencies most frequently sought out in EU electrical engineering job postings and those taught in EU bachelor and master-level electrical engineering curricula. The choice of market as the benchmark for supplied competencies is grounded in Human Capital Theory, according to which education functions as an investment in productive capacity [14]. From this perspective, the labor market serves as a direct indicator of which competencies generate productive value. Thus, when employers consistently seek particular I4 skills across multiple job postings, it indicates that these competencies contribute to productivity and organizational success. By aligning their curricula with these market signals, higher education institutions can potentially increase the return on educational investment for both graduates and employers.

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The research contributes to curriculum development in electrical engineering education by providing practical insights for EU EE programs and introducing a replicable data-driven method for benchmarking formal curricula against market competency requirements.

II. Literature Review

The purpose of this literature review is to summarize current consensus regarding I4 competencies in broader and engineering contexts, and to identify main convergence domains. The review starts with evaluating existing literature to establish a foundation and understand the I4 competency landscape. The review continues with a summary of two major engineering accreditation guidelines: Accreditation Board for Engineering and Technology (ABET) criteria for accrediting engineering programs and European Accredited Engineer (EUR-ACE) Framework Standards and Guidelines. This allows contextualizing I4 competencies within the focus area of the study. Empirical and mixed-method studies are then reviewed to understand the convergence between theory and the industry environment. Based on the obtained results, a theoretical competency framework is synthesized.

Although the particular terminology differs, the majority of theoretical studies regarding I4 competencies [7], [15-16], as well as literature exploring engineering education in I4 context [11], [17] converge on the following high-level categories:

- Technical skills: computer programming, digital literacy and digital literacy 4.0 (communicating via Internet of Things or/and Internet of People [18]), data analysis, knowledge of specific tools etc. This category may also include industry-specific competencies, such as state of the art knowledge and skills, manufacturing, robotics, automation, etc.
- Personal/cognitive/methodological (i.e., applied individually as opposed to within a group) skills: flexibility, motivation to learn, creativity, problem-solving, research, self-directed and lifelong learning, etc.
- Interpersonal/Social skills: communication, communication competence 4.0 (human-human and human-machine communication, understanding of machine-machine communication [19]), intercultural skills, team work, etc.

The fourth category is domain-specific engineering competencies, which are referred to as Core skills within the context of this study. They comprise the foundation of electrical engineering curriculum and provide necessary prerequisites, including areas such as circuit design, electronics, electromagnetics, signal processing, control systems [20]. ABET criteria for EE curriculum further outlines other foundational disciplines, such as mathematics, sciences (biological, chemical, physical), probability and statistics, as well as broader engineering topics [21].

Next, let us review I4 skills included in formal engineering program accreditation criteria. ABET outlines the following relevant engineering program competencies: problem solving; effective communication; decision making; collaborative work; leadership; goal-setting and planning; analytical thinking; working with data; self-directed learning and knowledge acquisition [21]. EUR-ACE Framework Standards and Guidelines outline the following relevant competencies: problem solving; analytical thinking; research and critical evaluation of literature; safety regulations; data search and analysis; self-management; efficient communication; intercultural skills; self-directed learning; life-long learning; collaborative work [22]. Based on the results outlined in these studies, an engineering competency taxonomy was synthesized, as presented in Table I. This taxonomy serves as the theoretical foundation for the study.

Taxonomies developed in empirical and mixed-method studies [6], [23-26] largely align with the outlined categories and provide deeper insights into the context in which these competencies are applied and their influence. In a survey [6], both students and working professionals regarded personal skills (continuous learning, reliability) as highly important for I4 work. Authors also differentiated a fourth category of skills, digital (IT systems operation, data analysis), which gained more recognition from working professionals compared to students. A comparative analysis of I4 competencies for future managers and engineers in pharmaceutical and automotive sectors [24] showed a different facet of I4 skills—their transferability across sectors. The results of the study demonstrated a strong overlap in several competencies—efficiency orientation, research skills, analytical skills—while decision making, conflict solving, and problem solving all had high demand rates, albeit with less convergence.

In a study of the influence of I4 competencies on the employability of undergraduate and graduate-level business students [25], both technical and non-technical skills (encompassing personal and social categories) had a statistically significant impact on employability. The author also noted that students were aware of the changing job market and sought skills to improve their employability within the Industry 4.0 environment. Despite this alignment, some industry surveys show a gap between employers' expectations and prospective workers' skill sets. In a study of a group of

TABLE I. Engineering Competencies

Category	Skills
Cognitive skills	Problem solving Continuous learning Self-directed learning Critical thinking Adaptive thinking Cognitive flexibility Self-management Decision making Entrepreneurial mindset
Technical skills	Data visualization Software programming Digital literacy Ability to use AI tools
Interpersonal skills	Networking Leadership Conflict resolution Communication Collaborative work Intercultural skills Knowledge transfer
Core skills	Electronics Circuit design Power systems Control systems Electromagnetics Signal processing CAD and simulation software

[a] Source: Developed by author based on [7-8], [15-20], [22-27].

enterprises in Poland that claimed to implement Industry 4.0 concepts [26], the highest-reported roadblock to implementing I4 was lack of qualified employees (77% responses). Furthermore, the most demanded skills voiced by respondents were long-life learning (83%), social media skills (74%), technical and management skills (72%), and team work (72%).

As shown by this literature overview, there is a consensus on competencies relevant in the Industry 4.0 context. Employers, workers, and students understand the importance of I4 competencies, creating demand for acquiring them. However, possible gaps suggest misalignments between engineering study programs and industry needs. Furthermore, existing literature largely relies on methods like bibliometric analysis, systematic literature review, questionnaires and expert interviews to derive and categorize I4 competencies. There is a gap in approaches that rely on market data, such as job postings, particularly in the context of electrical engineering within the EU. For higher education institutions aiming to modernize their electrical engineering curricula, having up-to-date and relevant data allows making grounded, informed decisions.

Thus, the following research questions are formulated.

RQ1: To which extent do Electrical Engineering curricula at STEM-focused higher education institutions in the EU align with the competencies most frequently demanded in the current job market, as identified through a thematic analysis of job postings?

RQ2: Which competencies demonstrate the largest absolute and relative discrepancies between labor market demand and curricular coverage in these programs?

The outlined research questions provide a way to test the core assumptions of the Human Capital Theory in the example of Electrical Engineering in EU. That is, RQ1 examines the core assumption that formal education provides valuable skills, and RQ2 focuses on quantifying the expected return on investment (ROI) in human capital [14].

III. Methodology

The study employs a mixed-method approach, combining qualitative thematic analysis and quantitative code frequency analysis to identify trends and evaluate convergence between market-driven competencies and those identified within Electrical Engineering study programs. Thematic analysis method was selected for its flexibility and ability to systematically analyze qualitative, unstructured textual data. It allows for conceptual grouping of related terms and supports a meaningful synthesis across varied terminologies [27].

First, a thematic analysis of 73 publicly available job postings was performed. Samples were collected from LinkedIn platform in April-May of 2025 with the following parameters: search query “Electrical Engineer AND Industry 4.0”; entry level; located in European Economic Area (LinkedIn location identifier 91000002); not an “express of interest” listing. For each job offer, the requirements, skills, and responsibilities (where present) were collected.

The decision for selecting LinkedIn as the primary data source was driven by its popularity, high data availability, and frequent data updates. As authors in [28] note, online job platforms like LinkedIn can serve as valuable sources of labor market insights, highlighting advantages like timeliness, coverage, and high sample sizes. However, such platforms also have limitations, such as self-selection bias, the presence of spam content, and duplicate data. To address these limitations, several measures were implemented. First, duplicate entries were identified through manual review of job titles, company names, and posting dates, and removed from the dataset. Second, third-party recruiting agencies that posted job listings on behalf of other firms were identified and excluded to ensure direct employer representation and avoid inflated job availability metrics. Finally, to mitigate self-selection bias and capture temporal variations in job market activity, data collection was conducted over a period of several weeks.

The sample size of 73 job listings was driven by thematic richness and sufficient variation, as identified during the preliminary analysis of the dataset. Code saturation was achieved at 68 listings with remaining 15 serving as confirmatory data. These results are generally in line with [29], in which authors achieved 80% thematic saturation from 50 documents and 90% from 81 documents.

Codes during the thematic analysis of job listings were developed using mixed approach, incorporating deductive and inductive methods. The initial coding framework was developed based on findings from theoretical and empirical literature discussed in the previous section. During the iterative analysis process, novel categories and codes not covered by the existing framework were developed inductively through the examination of the data’s meaning and context. Both sets of codes were then integrated into a final framework. To ensure consistency of assigned codes, intra-coder reliability was evaluated by re-coding 15 randomly selected samples after one week. Percentage agreement measure was used, achieving over 90% agreement.

Next, a thematic analysis of 25 Electrical Engineering study programs was performed against the coding framework developed in the previous stage. Higher education institutions for the analysis were collected from the European Higher Education Sector Observatory (EHESO) database, using a data snapshot for 2022, with the following parameters: number of students enrolled at ISCED levels 5-7 in Engineering, Manufacturing, and Construction > 1000; STEM orientation ("share of students and graduates in science and technology" [30, p. 114]) > 0.6. The number of students enrolled in Engineering, Manufacturing, and Construction was used as a proxy for maturity of institution’s engineering faculty [31]. The threshold represented the top third of most STEM-dense institutions. This selection allowed focusing on institutions with mature, established engineering departments, which are more relevant within the context of the study. Then, bachelor and master-level Electrical Engineering programs, with openly available curriculum and module structures in English (regardless of whether the program is in English, a local language, or mixed), were selected, with no more than one program per institution. Where the program was available at both levels, the bachelor's level was preferred. Module information, learning outcomes, and competencies (where available) were collected. All elective or optional courses were included in the program's analysis for completeness. The 25 programs represented a complete dataset for the given parameters. For evaluating consistency, 5 randomly selected samples were re-coded after one week, achieving over 85% on percentage agreement.

To assess the convergence between demanded and supplied competencies, the following indicators are introduced: gap score Δ and supply-demand ratio *R*. Δ quantifies the absolute magnitude and direction of alignment or misalignment in percentage points. This is a direct and interpretable measure of the coverage gap that allows for ranking skills. *R*, on the other hand, captures the proportional relationship, allowing for the identification of significant disproportions or complete omissions ($R = 0$) when absolute differences may not provide the same degree of understanding. This choice of metrics has several advantages: transparent and reproducible, no statistical assumptions; allow both ranking and threshold-based classification; interpretable and enable visualizations that are easy to reason about.

For an *i-th* competence, the gap score Δ_i is calculated as follows:

$$\Delta_i = S_i - D_i \quad (1)$$

where S_i is the proportion of study programs providing the skill, D_i is the the proportion of job postings requiring it. $\Delta_i > 0$ indicates over-supply, $\Delta_i < 0$ indicates under-supply.

Supply-demand ratio R_i is obtained as follows:

$$R_i = S_i / D_i \quad (2)$$

Similarly, $R_i > 0$ indicated over-supply; $R_i < 0$ indicates under-supply.

IV. Results & Discussion

The following results were obtained from the thematic analysis of public Electrical Engineering job listings. The coding framework developed during the thematic analysis is presented in Table II.

Technical and interpersonal categories have the highest

TABLE II. Theme, Categories, and Codes

Theme	Categories	Codes
Skills and competencies required for electrical engineers in Industry 4.0 environment	Cognitive skills	Self-management Analytical thinking Problem-solving Continuous learning Self-directed learning Decision-making Entrepreneurship
	Interpersonal skills	Collaborative work Communication Knowledge transfer Networking Leadership Intercultural skills
	Technical skills	Technical writing Human-machine interaction Software programming Data analysis Project management Standards and regulations knowledge
	Industry 4.0	Artificial Intelligence (AI) and Machine Learning (ML) Knowledge of Industry 4.0 technologies and concepts Robotics Internet of Things (IoT) Embedded systems
	Language knowledge	English Work- or region-related language

density, with competencies in these areas constituting 22% of all job postings, followed closely by cognitive skills at 20%. Collaborative work has the highest demand; it was present in 79% of recorded job postings, followed by communication (73%), self-management (66%), and technical writing (56%). As for language requirements, knowledge of English language is required in 73% of jobs, another work-specific language—45%.

Regarding minimal educational prerequisites, a vocational degree is required in 10% of postings, a bachelor's degree in 59%, a master's degree in 5%, and a PhD in 3%. Additionally, 11% of collected jobs mention professional certifications. Prior work experience is required in 62% of jobs.

RQ1: *To which extent do Electrical Engineering curricula at STEM-focused higher education institutions in the EU align with the competencies most frequently demanded in the current job market, as identified through a thematic analysis of job postings?*

To gain an overall understanding of the alignment between demanded and supplied skills, let us conduct a basic statistical analysis of the described metrics. As shown in Table III, the overall alignment is moderate. The 1.2 difference between the minimum and maximum values of Δ suggests existence of outliers on both ends. Furthermore, the 2.3 standard deviation in *R* suggests a relatively high level of inconsistency in taught competencies despite balanced central alignment.

The initial assessment of competency supply and demand, as presented in Fig. 1, reveals several observations about the nature of their relationship. First, there is a moderate over-supply subset at ~0-20% demand. On the other hand, several high-impact competencies at > 60% demand have lower supply. Mid-tier skills at 25-60% demand is generally over-supplied with a spike at ~35% demand.

RQ2: *Which competencies demonstrate the largest absolute and relative discrepancies between labor market demand and curricular coverage in these programs?*

The analysis of skills provided in the study programs reveals the following findings: categories with the highest density are technical (30%), cognitive (20%), and Industry 4.0 (20%) skills. The most prevalent skill is software programming, present in 92% of the study programs collected, followed by analytical thinking (88%) and embedded systems (80%). Interpersonal and language categories each constitute 10% of all skills. English is taught in 44% of programs, while a second language is taught in 36%.

The overall ranking of competencies by appearance in job advertisements is as follows (ascending): 1—cybersecurity; 2—decision-making; 3—AI and ML; 4—IoT; 5—robotics; 6—intercultural skills; 7—project management; 8—leadership; 9—data analysis; 10—self-directed learning; 11—continuous learning; 12—knowledge transfer; 13—software programming; 14—Industry 4.0 general knowledge; 15—problem-solving; 16—work-specific language; 17—analytical thinking; 18—human-machine interaction; 19—technical writing; 20—embedded systems knowledge; 21—self-management; 22—communication skills; 23—English language knowledge; 24—collaborative work.

As seen in Fig. 2, a number of competencies are oversupplied. This can be explained by institutions aiming to provide a broader set of skills to prepare graduates for a wider

TABLE III. METRIC SUMMARY

Measure	Δ	R
Min	-0.62	0.00
Max	0.58	10.22
Mean	0.05	2
Median	0.04	1.08
σ	0.29	2.3

range of possible jobs. Let us take a closer look at a subset of high-demand low-supply competencies identified previously. They are: self-management (Δ = -0.62, R = 0.06), sought-out in 66% of job descriptions while taught only in one program; communication (Δ = -0.25, R = 0.66); English proficiency (Δ = -0.29, R = 0.61); and collaborative work (Δ = -0.31, R = 0.6). Continuous learning (Δ = -0.22, R = 0) and knowledge transfer (Δ = -0.18, R = 0.18) are also under-supplied.

For the Industry 4.0 category, on average, supply exceeds demand, with a mean R value of 3 and a mean Δ value of 0.22. The only skills with slightly lower supply are general knowledge of Industry 4.0 concepts (Δ = -0.05, R = 0.87), usually presented through the lens of a particular study program (e.g., smart grids).

Several competencies identified as under-supplied in formal curricula, namely self-management, collaborative work, continuous learning, and knowledge transfer, are frequently mentioned as hidden curriculum components [32]. These may be implicitly developed through activities such as lab work and group assignments or obtained as part of the general culture of engineering education, even if they are not explicitly stated in learning outcomes or course syllabi. However, because this study analyzes only publicly documented curriculum elements, it is not possible to evaluate the extent to which the hidden curriculum compensates for formal gaps.

Among examples where such skills are included in the explicit curriculum is Eindhoven University of Technology, which requires a minimum of one multidisciplinary Challenge-Based Learning (CBL) project per semester with "personal & professional development" as one of its objectives. Another notable example is Karlsruhe University of Applied Sciences. Its Electrical Engineering program includes a Journal Club module where students are expected to learn critical thinking, effective communication methods, presentation skills, and knowledge conveyance, among others.

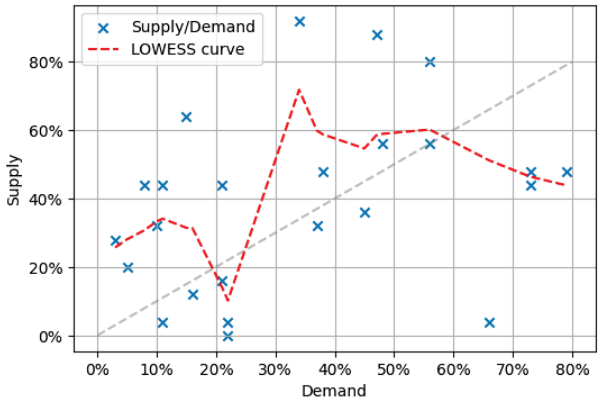


Fig. 1. Competency supply and demand.

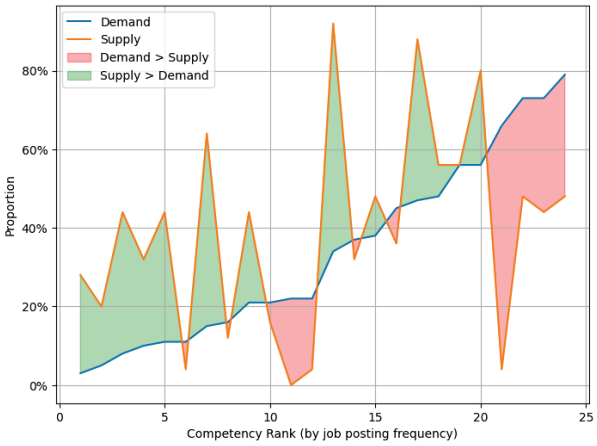


Fig. 2. Competency supply-demand gap; each numbered position represents a competency, ranked by frequency of appearance in job advertisements (1 = least frequently demanded). Data available in [33].

V. CONCLUSIONS

The study aimed to evaluate the alignment between Industry 4.0 competencies most frequently demanded in EU electrical engineering job postings and those formally taught in EU Electrical Engineering curricula. To achieve this, a mixed-method analysis was performed, starting with a thematic analysis of recent electrical engineering job postings, and followed by a thematic analysis of bachelor and master-level EE programs in STEM-focused higher education institutions in the EU.

Results reveal a moderate-to-high over-supply of competencies in the core curricula, with distinct subsets of highly demanded but under-supplied skills: self-management, communication, English language proficiency, collaborative work, continuous learning, and knowledge transfer. Skills in the Industry 4.0 category have a strong supply, with the only under-supplied skill being general knowledge of Industry 4.0 concepts.

The obtained findings confirm the core assumptions of the Human Capital Theory. That is, educational investment provides individuals with in-demand skills, which, in turn, enable future benefits in both the monetary (workers with more in-demand skills have higher competitiveness on the labor market), and non-monetary (better working conditions, increased performance, etc.) forms [14]. However, the subset of over-supplied competencies may require a deeper investigation. Higher education institutions that over-invest in certain competencies that are less widely demanded risk providing students with a less balanced skillset, which, in turn, decreases the overall ROI as these investments could be used to address more in-demand competencies.

The methodology applied in this study can be replicated for internal assessment of a given study program or for competitive benchmarking within a country or region. This data-driven approach allows for obtaining results within a short time frame using limited, publicly available data and provides a solid foundation for reasoning regarding the modernization needs of a study program's curriculum.

The study has two main limitations. First, the thematic analysis of curricula was limited to formal curriculum documents publicly available in English. This limitation reduced the number of study programs available for analysis and may have omitted competencies addressed through

hidden curricula or informal instruction. Second, foundational EE competencies were deliberately excluded to focus on Industry 4.0-specific skills. This aligns with the purpose of the study; however, including core skills may provide a deeper understanding of the alignment between study programs and market demands.

Future studies in this direction may focus on the following: including job-specific requirements (PCB design, CAD software, etc.) to gain a broader understanding of the alignment between supplied skills and employer expectations; evaluating core curriculum and elective modules separately; assessing the density of provided competencies (e.g., in ECTS credits).

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Increasing of the Power Plants Efficiency by the Use of a Combined Cycle with the Solid Oxide Fuel Cells Integration

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***Abstract*—The paper shows the ways of complex modernization of power units of the power plant in the city of Kharkov, Ukraine. The aim of the study is to develop conceptual solutions for the use of a gas turbine cycle and solid oxide fuel cells (SOFCs) as a superstructure for the modernization of powerful steam turbine plants to increase their energy efficiency. As an example, a 60 MW gas turbine considered as a superstructure over a 100 MW steam turbine. In addition to the combined cycle, the integration into the combined plant of a SOFCs modular plant with a total electrical output of 20 MW was also considered. The SOFCs modular plant consists of 40 tubular solid oxide fuel cells with a unit electrical output of 500 kW and with an internal reforming. A new method was developed that makes it possible to carry out a correct comparative thermodynamic analysis of combined cycles of power plants. The method is based on the use of a modified criterion of the structural complexity of the power plant modernization with a graphical apparatus for the optimization of thermodynamic cycles. By this method it was determined that the schemes of the combined cycle with SOFCs have a higher efficiency with acceptable complexity of modernization in comparison with the combined schemes of the steam-gas plant of different configurations. In addition, the method makes it possible to take into account the size of the area required for the placement of additional equipment during modernization.**

Keywords—combined cycle, energy efficiency, gas turbine, modernization, solid oxide fuel cell, steam turbine

I. Introduction

The problem of increasing energy production is especially relevant today in Ukraine. At the same time, priority should be given to the modernization of the electrical component of the combined heat and power plant (CHPP) in comparison with the rate of increase in heat supply.

Natural gas remains the main fuel for Ukraine power plants today, which can be used more efficiently in combined cycle power plants (CCPPs), in which the part of the fuel is introduced into the gas circuit. CCPP is a combined unit consisting of a gas turbine and a steam turbine. At a high heat supply temperature in the gas turbine and a heat removal temperature in the steam turbine condenser equal to almost the ambient temperature, the temperature ratio of hot and cold heat sources in the steam-gas cycle and its efficiency increases significantly. The efficiency of the steam turbine cycle is 30–35% [1] and the efficiency of modern gas turbines is 37–41%, while the efficiency of the combined cycle power plant is 42–60% [2]-[5].

Currently, much attention is paid to the design of combined power plants based on gas turbines and fuel cells. Basically, solid oxide fuel cells (SOFCs) are used for this. SOFCs are based on solid materials with ionic conductivity. The operating temperature of SOFCs is in the range of 1173...1273 K [6]. The efficiency of fuel cells is not limited by Carnot's law. Therefore, combined power plants based on fuel cells can be increased to 75% [7]. The fuel cells known today can produce from 0.3 MW to 30 MW of electrical energy [8]. In modern SOFCs the reforming of hydrocarbon fuel into hydrogen occurs directly in the fuel cell [9], [10].

Today a topic task is not only the creation of new power plants, but also the reconstruction of existing power units of thermal power plants through the introduction of a gas turbine cycle and solid oxide fuel cells.

II. Literature Analysis and Problem Statement

There are many scientific papers which discuss how to improve the efficiency of the combined cycle. So, Rao et al. [11] consider the question of a combination of Brayton's and Rankin's cycles, which provides an overall cycle efficiency of up to 60%.

Voloshchuk et al. [12] give examples of studies based on mathematical modeling using MathCAD Calculation Server technology for some operation cycles of a gas-steam turbine power plant. The internal efficiency of the combined cycle can reach 66%. This is achieved by using methods of thermodynamic analysis and optimization of combined cycles [12]-[15]. Also optimization methods are widely used to improve energy parameters, as well as economic and environmental indicators [16]-[21].

Aminov et al. [22] and Godoy et al. [23] give examples of combined cycle power plants developed in the world. It is shown that in order to ensure operation reliability, high thermal efficiency, low specific costs and operating costs, power gas turbines are currently designed according to the simplest cycle at the maximum achievable gas temperature. At the same time, the pressure increase degree values are chosen close to optimal ones, which correspond to the maximum efficiency for combined cycle power plants using the heat of the turbine exhaust gases.

To date, on the basis of the GTE-60 gas turbine of the State Enterprise "Zorya-MASHPROEKT" a couple of combined cycle gas turbines of various capacities have been developed. Combined cycle gas turbines CCGT-85 with a capacity of 85 MW and CCGT-170 with a capacity of 168 MW have an efficiency factor of at least 52% [4], [24].

Modern papers are also devoted to the development of hybrid power plants consisting of solid oxide fuel cells and a

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gas turbine or combined cycle gas turbine [25], [26]. There are also schemes for combining SOFCs with an internal combustion engine [27] and an air turbine [28]. New effective methods to analyze strength and dynamics characteristics of gas turbine units are developed in [29], [30].

Thus, basically all research is devoted to the development of new highly efficient combined power plants. This definitely makes it possible to achieve a high energy efficiency of the power units with minimal capital costs. However, there are certain requirements for the modernization of CHPP power units, which do not always allow the implementation of recommendations that are obtained by thermodynamic analysis and thermoeconomic optimization of combined cycles. This is usually due to technical, design, economic and CHPP architecture limitations. Therefore, there is a need to develop methods for finding rational solutions for the modernization of existing CHPP power units.

III. Goal and Research Objectives

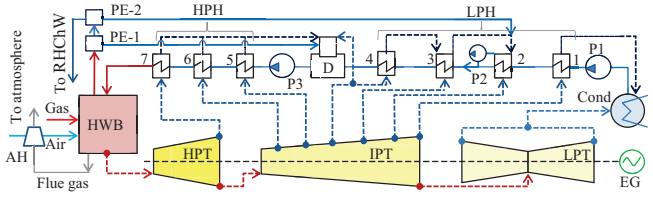
The goal of the study is to develop ways to introduce the gas turbine cycle and solid oxide fuel cell technology as a superstructure over existing powerful steam turbine plants. The research objectives are as follows:

- design and thermodynamic justification of various combined schemes for the modernization of the existing 100 MW CHPP power unit;
- development of a method for finding rational solutions for the modernization of existing CHPP power units;
- comparative thermodynamic analysis of energy efficiency and structural complexity of various combined schemes for the existing 100 MW CHPP power unit modernization.

IV. Thermodynamic Analysis of Various Combined Schemes for the Modernization of the 100 MW CHPP Power Unit

An analysis of the state of the power units operated at the CHPPs of Ukraine showed that power units with a lower exhausted resource include power units with a capacity of 100 MW and 250 MW. Thus, the Kharkiv CHPP-5 power units can be chosen as an example for modernization. This is especially true in connection with the military situation today.

As an example, the reconstruction of a power unit with a T-100/120-130 steam turbine [31] is considered. The basic thermal scheme of a power unit with a T-100/120-130 steam turbine is shown in Fig. 1. The regeneration system consists of four low pressure heaters (LPH), a deaerator (D) and three high pressure heaters (HPH). Connection type is cascade (without the use of drainage pumps).



HWB – hot water boiler; HPT, IPT, LPT – high, intermediate and low pressure turbine; AH – air heater; Cond – condenser; EG – electric generator; P – pump

Fig. 1. The basic thermal scheme of a power unit with the T-100/120-130 steam turbine.

The scheme uses a drum-type boiler with continuous purging. To reduce heat losses, the purge water is directed to a two-stage continuous purge expander, and then to the regenerative heater of chemical water treatment (RHChW), and then it is discharged into the sewer. From the first stage purge expander (PE-1) steam is directed to the deaerator, and from the second stage purge expander (PE-2) steam is directed to the LPH-2.

The factory data of the T-100/120-130 steam turbine are as follows: power output is $N_e = 100$ MW; live steam pressure is $P_0 = 127.5$ bar; live steam temperature is $T_0 = 555$ °C; calculated values of the internal relative efficiency by compartments are $\eta_{oi}^{HPT} = 95\%$, $\eta_{oi}^{IPT} = 95\%$, $\eta_{oi}^{LPT} = 97\%$; electromechanical efficiency is $\eta_{em} = 98\%$; transport efficiency is $\eta_{tr} = 98\%$; surface type heaters efficiency is $\eta_H = 98\%$. These data were obtained at a nominal steam mass flow rate $G = 115$ kg/s. In all calculations, it was assumed that the steam loss in the deaerator is 2% of G.

Modern CHPPs operate, as a rule, on turbines with steam extraction, the main advantage of which is the ability to work according to an independent schedule of electrical and thermal loads. In condensing mode, the turbine operates for about 2000–2500 hours per year. The electrical efficiency of a steam turbine operating in the heating mode is about 30%, and in the condensing mode it is about 37%.

Considering the foregoing, in this paper we propose to carry out and justify the way to increase the efficiency of CHPP by creating combined power plants, including a gas turbine as a superstructure, as well as innovative environmentally SOFCs.

It is proposed to choose a GTE-60 gas turbine for the modernization of the CHPP power unit, which is based on the standard Brayton cycle. The compression ratio in the gas turbine compressor is 18. To achieve a turbine inlet temperature of 1295 °C, natural gas (methane) is burned. Methane consumption during combustion is 3 kg/s. Then the combustion products expand in the turbine to a pressure of 1.1 atm. The temperature of the flue gases is 517 °C. Adiabatic efficiency (75.2% and 92.0% for compressor and turbine respectively) and combustion efficiency (98%) are calculated from experimental data [4], [24]. The gas turbine efficiency is 37% [32].

It is also proposed to use a modular plant with a total electrical output of 20 MW of 40 tubular solid oxide fuel cells with internal reforming. The temperature at the inlet of the tubular SOFCs is 800 °C, the temperature at the outlet of the tubular SOFCs is 1000 °C, and the efficiency of the SOFCs for generating electricity is 55% [33].

In this paper various variants for the possible combination of a steam turbine with a gas turbine and also with a gas turbine in combination with fuel cells are discussed. The thermodynamic parameters of various combined schemes are determined using a lumped approach to the analysis of each of its elements [34], [35].

A typical scheme of parallel connection of a gas turbine to a steam turbine is shown in Fig. 2. This scheme uses a waste heat boiler consisting of a number of heat exchangers. It is also called a Heat Recovery Steam Generator (HRSG). It creates steam for the steam turbine by passing hot exhaust gases from the gas turbine through banks of heat exchanger tubes.

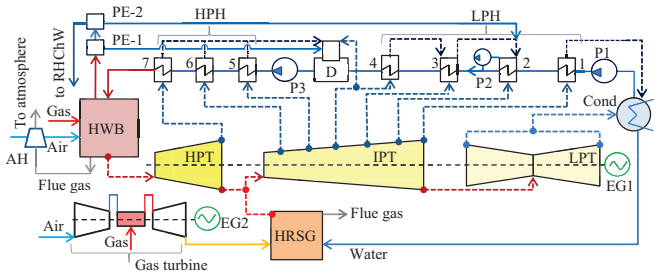


Fig. 2. The scheme of parallel connection of the GTE-60 gas turbine to the IPT of the T-100/120-130 steam turbine (scheme A).

The HRSG can use natural circulation or use forced circulation by means of pumps. As the hot exhaust gases past the heat exchanger tubes in which hot water circulates, heat is absorbed causing the creation of steam in the tubes. The tubes are arranged in sections, or modules, each serving a different function in the production of dry superheated steam.

The temperature of flue gases from GTE-60 is only 517 °C. This does not allow heating the steam in HRSG to the required temperature of the live steam supplied to the high-pressure turbine. Therefore, the additional steam flow from the HRSG is supplied to the intermediate-pressure turbine.

Such a connection allows increasing the electrical efficiency of the power plant up to 42% in the condensing mode. With this connection of the gas turbine, the scheme of the steam turbine remains practically unchanged. Therefore, capital investments during modernization will be spent mainly on a GTE-60 gas turbine with HRSG.

When GTE-60 is connected by parallel with afterburning fuel in HRSG, steam is heated to a live steam temperature of 555 °C and has a pressure of 12.8 MPa. This makes it possible to direct live steam from the HRSG to the high pressure turbine (Fig. 3). Such a connection makes it possible to increase the electrical efficiency of the power plant up to 44.75% in the condensing mode.

Another parallel connection variant with SOFCs modular plant is shown in Fig. 4. Here, the steam is heated to the required temperature of live steam due to the heat of the gases leaving the modular plant of solid oxide fuel cells with additional gas afterburning in the HRSG. The total electrical output of the SOFCs modular plant is 20 MW. The flue gas temperature in this case is 740 °C, which makes it possible to superheat the steam to the required temperature. Taking into account the utilization of the heat of the flue gases coming from the gas turbine and SOFCs, the thermal capacity of the HRSG for heating, water evaporation and steam superheating is 133.12 MW. The electrical efficiency of such a power plant will be 49.4% in the condensing mode.

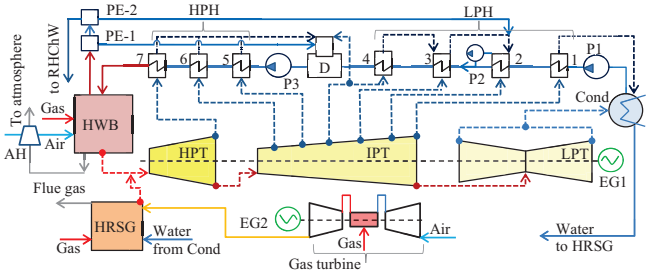
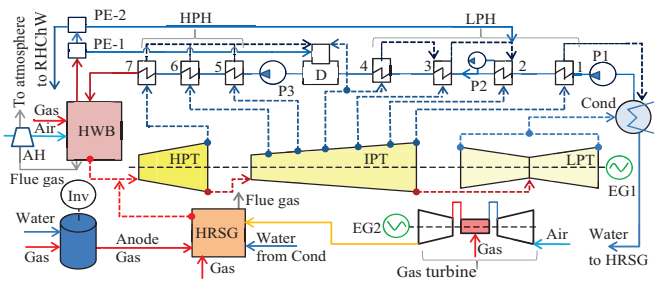


Fig. 3. The scheme of parallel connection of the GTE-60 gas turbine to the HPT of the T-100/120-130 steam turbine with fuel afterburning in the HRSG (scheme B).



Inv – SOFCs DC/AC inverter; Anode Gas – waste gas at the SOFCs outlet

Fig. 4. The scheme of parallel connection of the GTE-60 gas turbine to the HPT of the T-100/120-130 with fuel afterburning in the HRSG and SOFCs (scheme C).

A simple scheme for connecting gas turbine to the steam turbine was considered (Fig. 5). Here, flue gases coming from the gas turbine go instead of heated air to the HWB. This can be done because the flue gases leaving the gas turbine contain 15–17.5% O_2. A stoichiometric calculation shows that the flow rate of air required for the combustion of natural gas, for example, from the Kegichiv gas field is 162 kg/s for the steam turbine operation. Taking into account the decrease in the oxygen level during its burnout in the GTE-60 combustion chamber from 21% to 17.5%, it becomes necessary to increase its flow rate to 175.5 kg/s. This is practically equal to the flow rate of 173.4 kg/s of exhaust gases leaving the GTE-60 gas turbine. The heat supplied to the HWB with flue gases of GTE-60, taking into account the fact that they are cooled to 300 °C, is 44 MW. In this case, the electrical efficiency of the combined plant will be 41% in the condensing mode.

The thermal power of the flue gases leaving the HWB when they are cooled from a temperature of 300 °C to 120 °C is 36 MW and can be used to heat network water or replace low-pressure regeneration. Thus, replacement of regeneration from 4 to 7 extractions (see Fig. 1) leads to an increase in electrical efficiency in the condensing mode up to 42.6% (scheme E).

The inclusion of a SOFCs modular plant with a total electrical output of 20 MW in a simple scheme of a combined plant will increase the electrical efficiency up to 46% in the condensing mode. Such an increase in efficiency is achieved due to the complete replacement of regeneration with the utilization of the heat of waste gases from fuel cells and flue gases of the HWB (replacement of HPH and LPH with a HRSG) (Fig. 6).

The absolute increase in the electrical efficiency of various schemes η_{ei} compared to the electrical efficiency of the basic scheme of a power unit with the T-100/120-130 steam turbine η_{ebas} is shows in Fig. 7. The index i denotes various schemes.

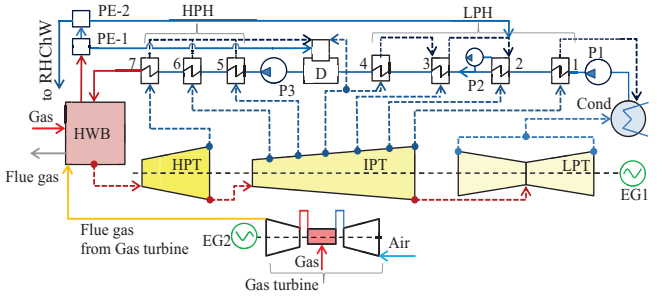


Fig. 5. A simple scheme for connecting the GTE-60 gas turbine to the T-100/120-130 steam turbine (scheme D).

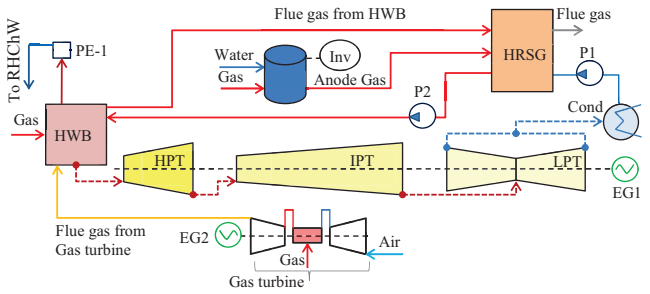


Fig. 6. A simple scheme for connecting the GTE-60 gas turbine to the T-100/120-130 steam turbine with SOFCs modular plant and replacement of regeneration (scheme F).

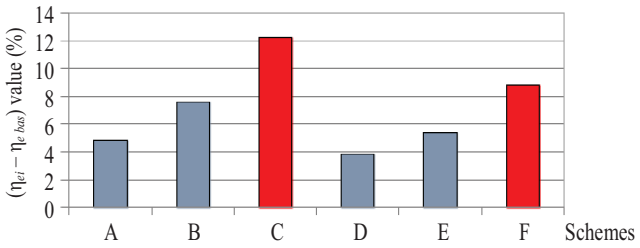


Fig. 7. Absolute increase in the electrical efficiency of schemes A-F compared to the basic scheme of a power unit with the T-100/120-130 steam turbine.

The greatest increase in electrical efficiency by 12.2% compared to the basic scheme of the CHPP power unit is achieved in scheme C of the CCPP with the SOFCs modular plant.

V. Choice of the Rational Scheme for the Modernization of the 100 MW CHPP Power Unit

It is advisable using the methodology of system-structural analysis to search for a rational structure of a technological combined schemes and configurations of a thermodynamic cycle that provide high energy efficiency under given conditions [34]. An important element of this analysis is the estimate of the level of structural complexity of the technological scheme. When creating new combined plants, the estimate of the schemes structural complexity level can be carried out using the criterion D_c proposed by Taubman:

$$D_c = D_i' (2\bar{m} + \bar{p}). \quad (1)$$

The criterion D_c takes into account the total number of interactions of the heat engineering system with the environment $\bar{p}$ (for example, heat removal from the condenser), the total number of technological connections between the system elements $\bar{m}$ and the total complexity of all elements D_i'. If the compared schemes are the similar in complexity of the main equipment, then the D_i' value is taken equal to 1 [36].

This method makes it possible to carry out a comparative thermodynamic analysis of power plant cycles using a deterministic criterion of constructive complexity and a graphical apparatus of C-curves. Thus, already at the stage of initial design of the system, it is possible to obtain as a universal method for determining schemes that are optimal in terms of energy efficiency. The introduction of the scheme complexity criterion into the pre-design analysis makes it possible to facilitate of choose for a rational scheme and makes the analysis results visual and informative.

Schemes A-F (Figs. 2-6) differ in structural complexity. One of the parameters for a correct comparison of schemes is the equality of the steam flow rate at the HPT inlet.

The dependence of the $1/\eta_e$ value on the scheme complexity criterion D_c is shown in Fig. 8, a. The $1/\eta e$ value is introduced for convenience. As can be seen from Fig. 8, a, only one C-curve can be drawn through the points Basic-E-F-C. According to the theory of graphical optimization [37], it is necessary to select the arc of choice that limits the schemes with the minimum value of $D_{c\,min}$ and with the minimum value of $1/\eta_{e\,min}$. Thus, in this case, the arc of choice is the F-C curve. Basic scheme and schemes A, B, D, E are not included in the arc of choice, since they have more complexity than scheme F and less efficiency than scheme C.

Thus, the analysis of schemes according to the theory of C-curves makes it possible to reduce the number of the schemes variants. It can be interpreted as follows. If we consider the tasks of creating a new combined plant, then the choice should be made between the two schemes F and C with SOFCs. For the final choice of the rational variant out of the two selected, it is necessary to carry out the thermoeconomic and environmental analysis [37]-[39].

If we consider this task from the point of view of the modernization of an existing CHPP power unit, one should take into account not only the structural complexity of the scheme, but the complexity of modernization by replacing one structural connection with another.

We proposed the following criterion for the structural complexity of modernization

$$\Delta D_c = D_i' (2 (\bar{m}_{ad} + \bar{m}_{rem}) + \bar{p}_{ad} + \bar{p}_{rem}), \quad (2)$$

where "ad" and "rem" are the subscripts of added and removed elements, respectively.

As can be seen from the Fig. 8, b, the largest ΔD_c value has the scheme F, which has the lowest D_c value. This is explained by the fact that when modernizing the steam turbine, it is necessary to replace the regenerative heat exchangers with a new heat recovery steam generator. Fig. 8, b shows the C-curve obtained by connecting the points D-B-C-F. The arc of choice D-B-C is limited by ΔD_c min and $1/\eta_{e\,min}$. It can be seen that schemes A, E and F are not included in the arc of choice. Scheme D has the least complexity of modernization, and scheme C has the highest efficiency with acceptable complexity of modernization. Thus, combined cycle scheme C with SOFCs (Fig. 4), which is included in both arcs of choice, can be considered as a variant for modernization of an existing CHPP power unit.

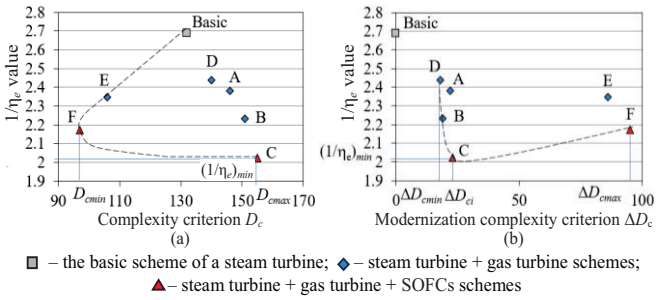


Fig. 8. Dependence of the $1/\eta_e$ value on the scheme complexity criterion D_c (a) and on the scheme modernization complexity criterion ΔD_c (b).

However, such an analysis does not allow making the final choice of a rational scheme. During modernization, an important factor is the possibility of installing additional equipment, taking into account the architecture of the CHPP. There should be enough space for modernization. Therefore, it is necessary to estimate what additional area will be required for the modernization of each scheme located on the arc of choice D-B-C. GTE-60 dimensions are as follows: length is 5.6 m, width is 3.3 m, and height is 3.5 m. The dimensions of one SOFC with an electrical output of 500 kW are: diameter is 2 m, height is 2.8 m. The HRSG for a gas turbine with a capacity of 60 MW has approximate dimensions of 26100×10230×12100 mm and a weight of about 260 tons.

The results of calculating the area A required for the placement of additional equipment, excluding pipelines, for each scheme B, C and D are shown in Table 1. Also, these data are shown in Fig. 9, a.

As can be seen from Table 1 and Fig. 9, a, the most efficient scheme C with SOFCs requires the largest area for the placement of additional equipment during modernization.

Therefore, if during modernization there is a condition for limiting the area for additional equipment placement, then for a correct analysis of the schemes, it is necessary to build the dependence of the ratio of the scheme electrical efficiency to its modernization complexity criterion $\eta_e/\Delta D_c$ on the ratio of the additional equipment placement area to scheme modernization complexity criterion $A/\Delta D_c$ (Fig. 9, b).

As a result of graphical analysis, if there is not enough space for additional equipment, from all schemes it is necessary to choose the combined cycle scheme B (Fig. 3), which is included in the arc of choose according to the scheme modernization complexity criterion ΔD_c (Fig. 8, b), has the highest ratio $\eta_e/\Delta D_c$ at an acceptable $A/\Delta D_c$ value (Fig. 9, b).

VI. Results Discussion

A method has been developed that makes it possible to carry out a comparative thermodynamic analysis of combined schemes of power plants. This method is based on the use of the proposed modified criterion of the structural complexity of the plant modernization in combination with a graphic representation. Fig. 8 compare the curves $1/\eta_e = f(D_c)$ and $1/\eta_e = f(\Delta D_c)$ using the classic complexity criterion D_c and the modified modernization complexity criterion ΔD_c. The first analysis can be applied when creating a new combined plant, and the second one when modernizing an existing one.

The combined scheme C with SOFCs during creation has the greatest complexity D_c and highest efficiency. But if we consider modernization, then the complexity of this scheme is significantly reduced from $D_c = 155$ to $\Delta D_c = 23$. Thus, combined cycle scheme C with SOFCs can be considered as a variant for modernization of an existing CHPP power unit.

The area required for the placement of additional equipment during modernization was also taken into account. Scheme C has the largest placement area. Thus, as a result of graphical analysis, combined cycle scheme B without SOFCs can be considered as a variant for modernization of an existing CHPP power unit in the conditions of limited space for additional equipment placement.

For a more correct choice of the scheme, it is necessary to supplement the method with thermoeconomic and ecological-exergy analysis [37]-[39].

TABLE I. Additional Equipment, its Quantity, Area for its Placement and the Increase in the Electrical Efficiency of Schemes B, C and D Compared to the Basic One

Scheme	ΔD_c value	GTE-60 (60 MW)	HRSG	SOFCs module (20 MW)	Area A, m^2	$(\eta_{ei} - \eta_{ebas})$ value, %
		Quantity; area for element placement				
B (Fig. 3)	19	1; 18.48 m^2	1; 267 m^2	No	285.48	7.6
C (Fig. 4)	23	1; 18.48 m^2	1; 298 m^2	1; 114 m^2	430.48	12.2
D (Fig. 5)	18	1; 18.48 m^2	No	No	18.48	3.8

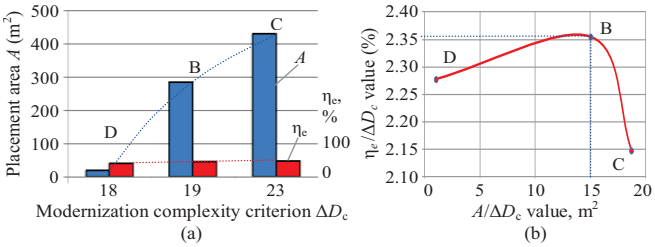


Fig. 9. The A, η_e values (a) and the dependence of the $\eta_e/\Delta D_c$ ratio on the $A/\Delta D_c$ ratio (b) for schemes B, C and D.

VII. Conclusions

Various combined plant schemes for the modernization of the existing 100 MW CHPP power unit were proposed. These schemes include a gas turbine and a SOFCs modular plant. For each scheme, a thermodynamic calculation of the cycle was carried out and the parameters and efficiency of the scheme were determined.

A method was developed for finding rational solutions for the modernization of existing CHPP power units based on the use of a modified criterion of the structural complexity of the power plant modernization.

A comparative thermodynamic analysis of energy efficiency and structural complexity of various combined schemes for the modernization of the existing 100 MW CHPP power unit was carried out. It was determined that combined cycle schemes with SOFCs have a higher efficiency with an acceptable complexity of modernization. A rational combined cycle scheme without SOFCs was also chosen taking into account the limitation of the area required for the placement of additional equipment during modernization.

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Analysis and Prioritization of Enemy Targets for Effective Strike UAV Mission Planning

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***Abstract*—This study focuses on the analysis and prioritization of enemy targets to support effective mission planning involving strike drone swarms. The goal is to improve the validity of selecting relevant targets for their effective engagement using available UAVs. Various scenarios for conducting military missions using a swarm of drones and a set of possible targets for engagement are considered. A list of criteria for assessing target priorities is proposed in order to identify a subset of high-priority targets. A graphical diagram of an extended hierarchy of prioritization criteria and possible targets is developed. A software tool using the Analytical Hierarchy Process has been developed to prioritize targets for strike drone missions.**

Keywords—strike UAVs, drone swarms, mission planning, target prioritization, criteria, Analytical Hierarchy Process.

I. INTRODUCTION

Unmanned aerial vehicles (UAVs) for various purposes have been widely used in the Russo-Ukrainian War. Special attention should be paid to the use of strike UAVs for suppressing enemy fire and engaging various types of targets. Groups of UAVs control a specific territory with the aim of damaging or completely destroying detected targets [1].

The relevance of this research area is driven by the growing use of UAVs in military operations in the form of drone groups, both homogeneous and heterogeneous (combinations of reconnaissance and strike drones). In this paper, the term "swarm" refers to a group of drones ranging from several units to several dozen, acting jointly within a mission to strike a target, but capable of employing different launch tactics and route formation strategies (waves, maneuvering, swarm splitting). Such swarms are often controlled by a single operator or by means of artificial intelligence.

They are capable of comprehensively performing reconnaissance tasks in a specific sector of military operations, suppression of enemy electronic warfare systems (EW), and simultaneously destroying their air defense systems (ADS) and other objects in accordance with the operational plan of the military unit's command.

II. ANALYSIS OF PUBLICATIONS

Currently, there are many publications on improving the effectiveness of UAVs, including swarm UAVs. Issues related to the use of various types of equipment for destroying enemy targets are being considered. The use of military UAV technology by armed forces is characterized by the use of numerous and more functional drones that can operate in large numbers.

There are practical developments in the planning of military missions for UAV swarms (the Mission Control module of the DELTA military system) [2]. At the same time, issues such as the distribution of areas of responsibility, crew route planning, and real-time coordination of tasks for UAVs are already being addressed.

An important step in planning military drone missions is selecting priority enemy targets for fire. This task must be solved based on a number of criteria using modern decision-making methods and technologies. A manual focused on the application of NATO standards for planning enemy fire missions using artillery indicates the need to select relevant targets and even provides several criteria for their evaluation [3]. However, these criteria are generalised, qualitative in nature, and do not reflect the specifics of the use of strike drones.

The article [4] defined the components of the process of military engagement of the enemy using the "target tree" method based on functional characteristics, with greater emphasis on the processes of obtaining information about the situation, assessing the situation, and distributing forces and means of engagement between tasks.

The article [5] identified the key factors that shaped the target setting tasks. These include technological developments, organizational structures and processes, and competition between types of troops for limited resources. In addition, target setting formation is influenced by intelligence data on the nature of the enemy and assumptions based on experience, doctrine, or strategic theory. The target setting process is characterized by a tendency toward increasing the accuracy of ammunition and the ability to strike specific targets while constantly reducing response times and increasing distances from the target.

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There are researches devoted to the stage that comes after target setting – the detection of specified targets and the control of drone swarm movement. For example, monograph [6] discusses the calculation of waypoints in advance or the generation of new waypoints during a mission. Methods have been proposed for detecting targets based on UAV aerial photographs for small targets, dense locations, sparse distribution, and complex backgrounds using artificial intelligence, namely, feature extraction using asymmetric convolution [7].

With regard to the criteria for assessing the relevance of targets, given their priority for engagement by a swarm of UAVs, there are currently insufficient publications. Thus, in article [8], solution is presented for the problem of determining the location of objects by drones with minimum cost in a two-dimensional terrain. The number of drones and total energy consumption are the two cost indicators considered. The problem is solved mathematically by forming integer linear and mixed nonlinear optimization models. Article [9] considers four main criteria for selecting targets. However, the essence and influence of these criteria on the superiority of the solution aren't disclosed; they are evaluated subjectively, on a linguistic scale.

Regarding modern methods for assessing the importance of objects, special attention should be paid to the Analytical Hierarchy Process (AHP), which is widely used in decision-making and allows the decision problem to be structured as a hierarchy of goals, criteria and alternatives, with a focus on expert subjective judgments. This method can be combined with other decision-making approaches and has several modifications. Fuzzy logic methods are effective when uncertainty and vagueness in the input data need to be taken into account. For example, fuzzy TOPSIS (FTOPSIS) is a modification used to determine priorities based on indicators with fuzzy values [10]. Unlike the AHP method, neural networks demonstrate high performance on large data volumes and can autonomously learn complex relationships, although they lack transparency in explaining the decisions [11]. The AHP method is more suitable for tasks with a clearly defined structure and the involvement of experts.

Thus, in the task of prioritizing targets for strike UAVs, the issue of analyzing and forming a set of criteria for the selection task remains unresolved, and no formalized procedure for their evaluation has been proposed. The aim of this study is to improve the validity of the selection of relevant targets for their effective engagement by available UAVs.

III. Materials and Methods

A. *Basic definitions and provisions of the decision-making task for target engagement*

Let us consider the main provisions regarding the task of decision-making on engagement enemy targets using a swarm of UAVs.

Military operations are planned based on the collection of intelligence information. Based on the results of identifying potential targets, a decision is made on whether a specific object requires action to destroy it.

The military operations planning headquarters prepares decision-making information that answers the following questions [12]:

- What targets should be attacked?
- When and where targets can be found?
- What means should be used to determine the exact location of targets?
- How accurately the location of a target must be determined in order to attack it?
- What priorities exist for intelligence, tasks, and the distribution of forces and resources?
- What requirements are important for target detection?
- When, where, how, and in what order targets should be attacked?
- What are the indicators of mission performance and effectiveness?
- What means should be used to attack targets (number/type of UAVs, their armament) in order to achieve the desired effects?
- How quickly and in what format this information should be obtained and processed?

There are two levels of UAV use by military units [13]. The general task of using weapons is to use forces and means across the entire zone in the interests of the battle as a whole, carried out to the full depth of the enemy's battle order. It's done through fire raids and other actions. The direct task of using weapons is to use forces and stuff in the area of responsibility of general military formations to carry out specific tactical tasks in a limited area (in the direction of their actions). Based on reconnaissance information, the task of targeting is decided – the process of selecting, prioritizing (importance) of targets and coordinating the appropriate impact on them, taking into account operational requirements and capabilities [13]. This task is part of operational planning and is based on a cycle consisting of the following main phases:

a) Based on the concept, task, and instructions from the command, determine the Named Area of Interest and Target Area of Interest.

b) Determining targets and their priority, based on which a High Payoff Target List (HPTL) is developed.

c) Analyzing the possibilities for target engagement; implementing additional measures that may affect the distribution of UAV resources.

During planning, lists are formed: No-Strike List; Restricted Target List; HPTL.

The current assessment of the military capabilities of available engagement means helps to conduct a realistic analysis of efforts to determine the effectiveness of engagement, review plans for the use of engagement means based on observed results, and change the priorities of subsequent targets.

The High Value Targets list is compiled based on the relative importance of the target to the enemy's operations. High Value Targets include capabilities, functions, or systems that are critical to the enemy in the context of military. High Payoff Targets are those nodes among high-importance enemy targets whose elimination renders the node non-functional. Keep in mind that too many high-

priority targets will spread out the effort to gather info, locate the target, and destroy it.

According to [13], the following general criteria are used to analyze the significance of targets and their verbal assessment: criticality; accessibility; recouperability; vulnerability; effect; recognizability.

The targeting working group reduces, modifies, and prioritizes high-importance targets, ensuring that high-priority targets support the command's concept of battle.

Waves of strike drones allow for a massive attack on enemy targets to inflict maximum damage. For effective military missions, the engagement of targets must be taken into account [14]:

- Consists in the complete destruction of selected actual enemy targets.
- Inflicting damage that prevents the enemy from continuing to conduct successful military operations on the battlefield.
- The need to force the enemy to switch from active offensive operations to defense or retreat.
- The objectives of a military operation aimed at intercepting the military initiative.

It should be noted that prioritizing goals makes it possible to solve two interrelated tasks. Thus, at the strategic planning stage, after identifying high-opportunity goals, it is necessary to analyze the combat potential of existing UAVs, taking into account their tactical and technical characteristics, which will enable the effective performance of engagement tasks as part of a drone swarm. If it turns out that the available resources are not capable of performing the task due to their characteristics, or there are not enough drones of a certain type in military units, appropriate measures should be planned:

- Implement projects to modify UAVs to ensure certain characteristics.
- Implement projects to increase the production of certain types of drones.
- Make decisions to improve the logistical support of military units.

The opposite task is operational and tactical planning of the use of existing types and numbers of drones to form a swarm. When selecting high-priority targets, the technical characteristics of the drones should also be taken into account, and targets that can be struck with maximum effectiveness should be selected.

B. The method of evaluating targets by particular parameters in the analytic hierarchy process

In general terms, the task of making decisions on determining high-priority goals boils down to the following list of actions:

- Determining the enemy's engagement strategy.
- Selecting a set of targets (of various types) for implementing the specified strategy (high-importance targets).
- Determining the weighting coefficients of the criteria for selecting the type of target.
- Evaluating target types based on a set of criteria.
- If there are several targets of the same type, conducting an additional analysis of options based on a subset of relevant criteria.
- Forming a list of high-priority targets, taking into account their parameters (based on quantitative and qualitative characteristics).

Thus, we have a hierarchy of objects for determining their priorities: at the first level – evaluation criteria, at the second level – types of goals, at the third level – specific goal options with their parameters.

Let us consider the AHP method for multi-criteria evaluation of objects from a set of options [15].

Each x_i object (target) receives an expert assessment G_{mi} for the selected P_m parameter. Then, using special rules for comparing two objects, square matrices of A_m paired comparisons are formed for each m parameter for all n objects. This matrix has the properties of inverse symmetry, i.e. $a_{ij}=1$, at $i=j$, where i is the row number, j is the column number and $a_{ij}=1/a_{ji}$.

To form a matrix of paired comparisons for the selected parameter, the values of the expert assessment are analyzed based on the scale of relative importance, ranging from equal importance (1) to absolute superiority (9), with intermediate values and reciprocal numbers used for finer judgments.

To determine the priority value of objects for each parameter, two operations are performed:

1) Normalization of the matrix – all values in each column are divided by the sum of that column, which makes it possible to obtain relative indicators.

2) Calculation of weight coefficients – for each object, the average value of the row in the normalized matrix is determined. These average values form the priority vector for the given parameter.

The same operations are repeated for each parameter. As a result, a matrix of the importance of objects across all parameters is formed. Then, the relative importance of the parameters themselves is taken into account, which makes it possible to obtain a generalized priority vector. This vector reflects which objects (targets) have the highest importance, considering all the criteria.

IV. ANALYSIS OF TARGET OPTIONS FOR ENGAGEMENT BY A SWARM OF DRONES AND CRITERIA FOR THEIR EVALUATION

The main functions of military UAVs are [16]: reconnaissance; combat engagement of various targets; interception and engagement of air targets.

It should be noted that an effective option is to form a combined-purpose swarm, where one or more drones are used for reconnaissance among the main body of strike drones. In this case, the task of selecting high-priority targets can be supplemented in real time with relevant information from the battlefield or the engagement area.

The types of reconnaissance or engagement targets are as follows [17]:

- Permanent locations of military units and areas of troop concentration (reserves), command posts.
- Airfields, naval bases, training grounds, and training centers.
- Bases, depots for weapons and military equipment, fuel and lubricants, material and technical resources.
- Enterprises of the defense-industrial complex and oil refining industry.
- Important infrastructure and communications facilities (railway junctions, bridges, and overpasses).

After identifying potential targets, the task is to determine the optimal route. For this purpose, the following route "cost" criteria are used:

1) Distance factor – the sum of the route lengths.

2) Resource factor – the total amount of resources spent by UAVs to achieve the targets.

3) Intersection factor – the number of intersections of the route with others.

4) Time factor – the time it takes for all UAVs to reach their targets.

For strike targets, the time factor directly affects the effectiveness of the mission, which is most often measured by the probability of hitting the target. Thus, in [18], a study is presented on the basis of which it can be argued that to hit 20 targets with a group of 3 strike UAVs with a probability higher than 0.5, the flight time should be no more than 14 minutes.

For large targets (infrastructure objects) or dynamic targets (military equipment during combat or movement), the following factors are important: the number of strikes required – depends on the amount of ammunition (in turn, the amount depends on the number of drones in the swarm and the maximum weight of ammunition on each of them); target exposure time, disappearance time for impact.

For spatial targets, the probability of engagement depends on the required engagement zone radius and the speed of the strike UAV [18].

Thus, the first thing to consider when assessing target priorities is the military mission scenario. Let us form a set of possible scenarios with the identification of factors that will influence the selection of targets and the criteria for their assessment (see Table I).

Based on the analysis of factors affecting the success of hitting a target with a swarm of drones, the following criteria can be used to assess their priorities:

- Relevance of enemy targets for engagement – impact on the mission, military importance of the target (strategic/operational/tactical priority).
- Economic or strategic value of the object.
- Urgency of engagement.
- Probability of successful engagement.
- Required combat potential (resistance to engagement, area of impact).
- Time required to achieve (or range).
- Level of threat to drones (presence of ADS or EW).
- Target accessibility.
- Depth of the target in enemy military formations or rear.
- Mobility, maneuverability of the target.
- Reliability of information about the target.

TABLE I. POTENTIAL PRIORITY OBJECTIVES FOR DIFFERENT TYPES OF MILITARY MISSION STRATEGIES

Types of targets	**Types of strategies**					
	Counterattack	***Defense of positions***	***Isolation of the combat area***	***Conditions of limited resources***	***Engagement of ADS***	***Operation in urban areas***
Locations of military units		+				
Command post	+	+				
ADS	+				+	
Airfields, naval bases		+				
Training grounds and centers				+		
Depots for military equipment and material resources				+		
Defense industry enterprises				+		
Infrastructure and communications facilities			+			

The first three criteria (relevance, value, urgency) are factors determined by the command at the strategic (or operational) level. They are the ones that influence the classification of the targets under consideration as high-priority targets. In addition, some criteria are correlated. For example, all other criteria (range, combat potential, threat level, etc.) influence the probability of successful engagement. More about these and other criteria when planning UAV missions in the authors' work [14].

V. SOFTWARE IMPLEMENTATION OF THE GOAL PRIORITIZATION METHOD IN DRONE MISSION PLANNING

A software system for prioritizing criteria and objectives based on information about planned strategies for military missions of strike drones has been proposed. The developed system can be used by a targeting group to prioritize targets based on expert assessment and using certain objective information about command plans, tactical and technical characteristics, and the number of drones available for military missions. Fig. 1 shows an example of a graphical diagram of an extended hierarchy for prioritizing targets for engagement by a swarm of strike drones. The first level is the target, the second is the level of strategies or scenarios (see Table II), the third level is the level of criteria for evaluating targets. The fourth level is the level of alternatives. An implemented example of evaluating such targets with elements: "Command post"; "ADS"; "Ammunition depot"; "Military convoy in motion"; "EW"; "Defense enterprise" etc. The system allows to form a hierarchy of decision-making tasks of any complexity and content. Fig. 2 shows the stage of hierarchy formation in the form of strategies and criteria. Fig. 3 shows an example of prioritizing criteria (on a scale from 1 to 9) to compare their importance in implementing the "Counterattack" scenario.

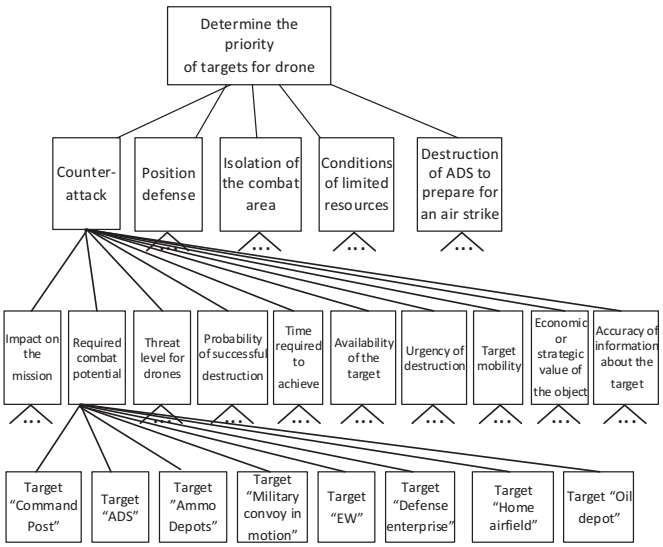


Fig. 1. Graphical diagram of the extended hierarchy

Fig. 4 shows the results of pairwise comparisons of scenarios, criteria, and alternatives. The obtained values of importance vectors for scenarios and criteria for each of them are presented in Table II. Fig. 5 shows a diagram of the priority vector of criteria for the "Counterattack" scenario. Fig. 6 shows a diagram of the priority vector of enemy targets for engagement in the "Counterattack" scenario.

The system we propose is integrated into a strategic and tactical UAV mission planning framework (Fig. 7). It enables the simulation of complex combat scenarios (Fig. 8), target selection and prioritization, prediction of losses, automatic route generation for drone groups, wave-based attacks, and the assessment of the expected mission effectiveness.

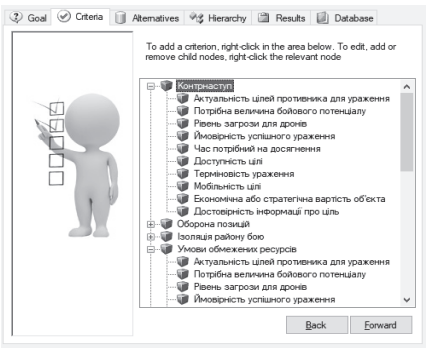


Fig. 2. Forming a hierarchy to prioritize UAVs targets

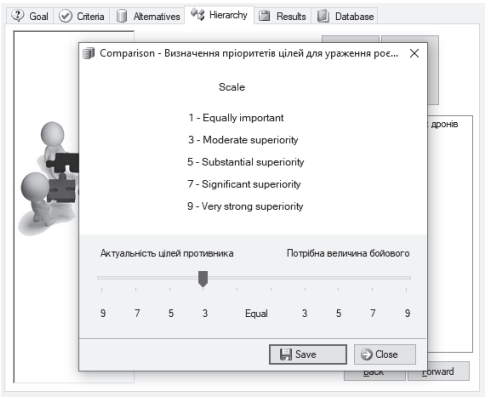


Fig. 3. Comparison of the criteria importance according to the "Counterattack" scenario

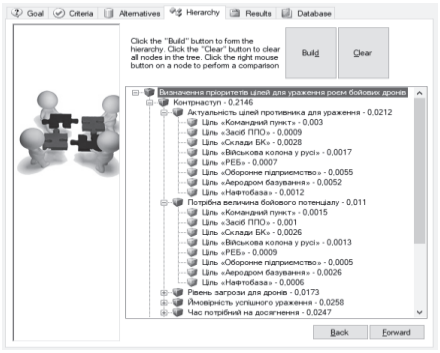


Fig. 4. Results of pairwise comparisons of scenarios, criteria, and alternatives

TABLE II. VECTORS OF CRITERIA PRIORITIES BY SCENARIOS

Criteria	**Scenarios**				
	Counter-attack	***Defense of positions***	***Isolation of the combat area***	***Limited resources***	***Engagement of ADS for preparation of an air strike***
Scenario priorities	0.246	0.1219	0.1219	0.3198	0.1293
Relevance of enemy targets for engagement	0.0212	0.0087	0.0083	0.0138	0.0107
Combat potential	0.011	0.0131	0.0106	0.0425	0.0109
Level of threat	0.0173	0.0133	0.0151	0.0417	0.0117
Probability of successful engagement	0.0258	0.0142	0.0187	0.0414	0.0142
Time to achieve	0.0247	0.0106	0.0138	0.0359	0.0142
Target availability	0.0201	0.0212	0.0106	0.0295	0.0099
Engagement urgency	0.036	0.0121	0.0171	0.0464	0.0209
Target mobility	0.013	0.0146	0.0074	0.0168	0.0122
Economic or strategic value of the object	0.0353	0.0072	0.0123	0.0356	0.0098
Reliability of information about the target	0.0103	0.0069	0.0079	0.0161	0.0148

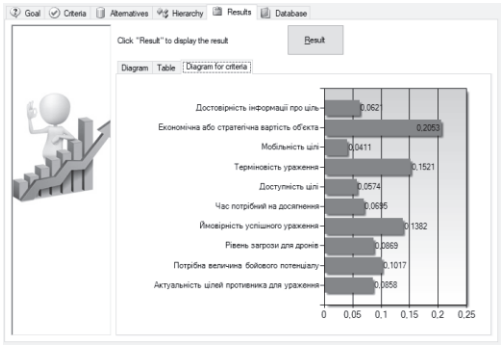


Fig. 5. Vector of criteria priorities for the "Counterattack" scenario

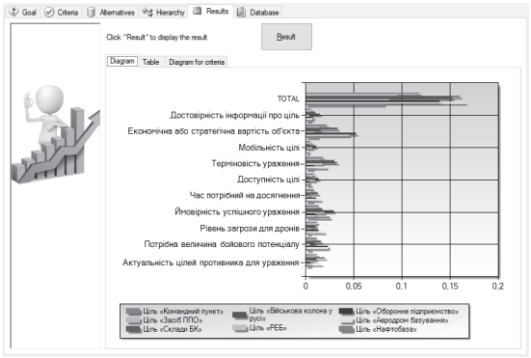


Fig. 6. Vector of priority targets in the "Counterattack" scenario



Fig. 7. Integrated prioritization system within the UAV mission planner

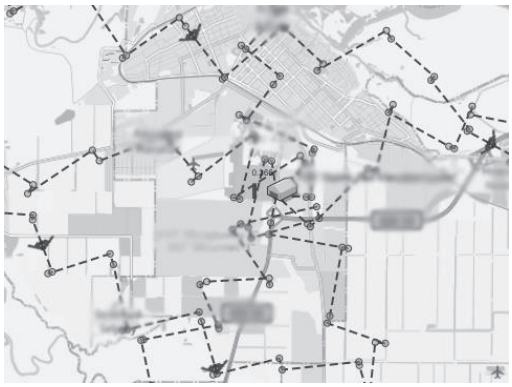


Fig. 8. Target priority and the generated drone routes (flank strike)

CONCLUSIONS

A review of the available literature indicates that research on the specific task of selecting relevant targets for engagement by a swarm of drones is still very limited. To the best of our knowledge, no direct publications on this problem have been identified. Comparable studies are found mainly in the context of artillery operations, particularly drawing on NATO experience. The main provisions regarding the task of making decisions on the engagement of enemy targets using a swarm of UAVs are considered. The main phases of operational planning for the engagement of enemy targets are provided. The task of determining a set of high-importance and high-priority targets is formulated. The possibility of using the results of target prioritization at the strategic level for planning UAV modernization projects and at the tactical level for the use of available means of target engagement is considered. The stages of the general task of decision-making on the determination of high-priority targets are listed, which made it possible to justify the use of the AHP method. The main result of the article is an analysis of options for targets for drone swarm attacks and criteria for their evaluation. A set of strategies for performing military missions using drone swarms has been developed. In accordance with this set, types of targets and criteria for their evaluation have been determined. For the target prioritization process, developed software is proposed that automates pairwise comparison of alternatives and calculates their relative priorities.

Further research will focus on improving route generation algorithms and simulation modeling of goal achievement with an assessment of the probability of completing a military mission.

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Development of a Method for Transmitted Multi-Model Augmented Reality Objects

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***Abstract*—the paper presents a modification of the basic method used to transfer multi-model augmented reality objects. The analysis of existing developments has proven the relevance of this task. A client application with augmented reality technology has been created using a combination of Angular, Node.js, Ar.js tools together with A-Frame. The loading speed of a 3D object using the basic and modified methods has been investigated. The limits of application of the modified method have been determined.**

Keywords—web, augmented reality, AR, Angular, multi-modeling.

I. Introduction

Over the past three decades, the development of Internet technologies has progressed from the static and one-way model of Web 1.0 to the interactive, personalized, and intelligence-oriented Web 4.0, demonstrating significant advances in the field of web technologies [1]. One of the modern innovative and interactive technologies that is actively being integrated into client-side applications is Augmented Reality (AR).

The development of web applications using augmented reality technology opens up new opportunities for shaping a novel user experience. AR web applications operate directly within web browsers, eliminating the need for additional software installation. They can be launched on most widely used mobile devices, including tablets and smartphones [2].

Examples of AR applications include websites that apply photo filters, feature interactive games, conduct virtual tourist tours, enable virtual try-on or product visualization, and provide various skills training. At the same time, the number of studies related to augmented reality is steadily increasing. It can be assumed that in the near future, AR-enabled websites will be able to provide users with a daily extension of reality [3].

The active implementation and dynamic development of client applications with support for augmented reality technology is due to a number of key factors, including:

- widespread daily use of websites in various spheres of human activity;
- the growth of mobile device users;
- increasing the availability and accuracy of geolocation services;
- expanding the coverage of high-speed mobile networks.

Collectively, these prerequisites foster an increasing demand for the development of AR-based web applications (WebAR), positioning this domain as a pertinent and timely area of modern web engineering. The fundamental building blocks of such applications are augmented reality libraries, which provide the essential AR functionality.

II. Literature Review and Problem Statement

Augmented reality enables the integration of digital content, such as images, audio, or text into the user's real-world environment, creating a rich and interactive experience. As it evolves, augmented reality is transforming into an effective tool for enhancing cognitive capabilities and supporting managerial decision-making across various domains, including mechanical engineering, healthcare, urban infrastructure, business, and tourism [4]. Within the web environment, two primary types of AR systems are distinguished based on their functional characteristics [5, 6].

The first type is marker-based. These systems utilize visual markers as triggers for overlaying digital content onto the physical world. Marker-based web AR systems typically consist of three key components: the marker itself, the AR application, and the web page. This technology is widely applied in museums and exhibitions, as well as in the fields of commerce and healthcare [5, 7].

The second type of augmented reality system is based on geolocation and does not require pre-defined markers. This technology leverages GPS, digital compasses, or other sensors to determine the user's coordinates, upon which AR objects are overlaid onto the physical environment or replace its elements. Such systems demand greater computational resources and more complex processing, which introduces certain technical challenges. Compared to marker-based AR, they are generally less accurate and stable. Markerless WebAR solutions are most commonly applied in gaming, e-commerce, and healthcare domains [5, 7].

Mobile devices, particularly smartphones and tablets, are most frequently used to interact with augmented reality web applications, as they are widely available and easily accessible. Additionally, headsets, smart glasses, and Head-Up Display (HUD) projection systems are employed to project images onto transparent screens in front of the user's eyes. A promising direction of development is the creation of AR contact lenses, which are currently in the research and prototyping stage [7].

In the development of software utilizing augmented reality, access to high-quality computer vision libraries and tools for object recognition and tracking is essential. These AR libraries form the foundation for the accurate and efficient operation of such applications. To fully implement AR functionality, two key types of libraries are required.

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The first type is responsible for determining the spatial positioning of a 3D object. In markerless augmented reality, this is achieved using geolocation data. In marker-based AR, object positioning is established through visual marker recognition. Libraries that serve this purpose include ARToolKit.js, AR.js, MindAR, as well as commercial platforms such as Vuforia and ZapWorks [8].

The second type of libraries used in WebAR development specializes in graphics and is responsible for constructing scenes to visualize 3D objects. These libraries provide tools for creating, configuring, and animating three-dimensional scenes within an augmented reality environment. The most widely used graphics libraries include WebGL, Babylon.js, Three.js, and A-Frame.

The most common approach in the development of AR web applications involves combining tracking libraries with graphics libraries. The results of an analysis of these two types of AR libraries [9, 10] are presented in a comparative table (Table 1), which outlines combinations of libraries used for computing the spatial positioning of 3D objects and rendering their graphical representation.

TABLE I. COMPARISON OF COMBINATIONS OF AR LIBRARIES

Combination of libraries	Main purpose	Advantages	Disadvantages	Difficulty level
A-Frame + AR.js	Fast and efficient development of AR web applications	Ease of integration, high speed of implementation, minimal amount of code, ease of use	Limited flexibility and functionality compared to more complex solutions	Low
Three.js + ARToolKit.js	Advanced development with complex scenarios	Great flexibility and extended functionality, integration with other libraries	Higher volume of code, complexity of use	High
MindAR + Three.js	Using machine learning in AR	The possibility of integration with TensorFlow, support for complex tokens, advanced possibilities for interactivity	Less popularity and support compared to AR.js, more difficult to learn	High
Vuforia або ZapWorks + Three.js	Comprehensive development of AR applications with support for web interfaces	Platforms for the development of mobile and web AR applications, a powerful API for integration	Main focus on mobile applications, possible limitations when working with web applications	Medium

Special attention is required for applications that utilize multi-model augmented reality objects. During their development, it is necessary to apply solutions designed to handle a relatively large total volume of models being loaded.

The authors of [11] present the development of a markerless AR website intended for veterinary student education. The site enables the viewing of an AR object without requiring any custom site logic, and therefore uses HTML as the markup language. Augmented reality functionality is implemented using the <model-viewer> web component rather than a full AR library. All interaction logic, such as user interaction with the 3D object, is handled entirely by this component. There is no dedicated server; instead, a remote host is used to store and deliver the 3D models.

The solution presented in [12] enables training in the use of medical therapeutic devices through AR/VR technologies. The developed web system consists of a Node.js-based server, a Socket.IO module, and the Three.js library. The server stores and synchronizes the state of the medical equipment. The Socket.IO module adds functionality that allows the 3D model of the equipment to respond to changes made on a mobile phone. Three.js is used to render the 3D object on the phone screen.

In [13, 14], neural network methods were employed to develop recommendation systems for art compositions within augmented reality systems.

In [15], the Alpaca framework was developed to allow third-party web developers to easily and efficiently integrate AR components into their own web applications. Alpaca consists of a client-side module implemented using HTML5, JavaScript, and Three.js-based AR rendering tools, and a server-side module built using the Python aiohttp library.

The outcome of the work presented in [16] is a web-based tool for generating resumes for job seekers, featuring augmented reality elements such as QR codes and AR markers. Augmented reality functionality in the application is implemented using a combination of two libraries: AR.js and A-Frame.

Despite significant progress in the development of client-side augmented reality applications [11–16], the reviewed studies do not focus on applications based on multi-model augmented reality. When the loading time of AR modules exceeds an acceptable threshold, freezes or delays in rendering may occur, which are perceived negatively by users.

III. THE AIM AND OBJECTIVES OF THE STUDY

The aim of our work is reducing the loading time of multi-model AR objects by developing a modified method for loading AR modules.

To achieve this goal, the following tasks were set:

- develop a client application with augmented reality technology;
- modify the method of transmitting multi-model augmented reality objects in order to reduce the overall transmission time;
- conduct a comparative study of the basic and modified methods of transmitting multi-model augmented reality objects.

IV. DEVELOPMENT OF A CLIENT APPLICATION WITH AUGMENTED REALITY TECHNOLOGY

This section presents a web application with augmented reality elements, with the client side implemented using the Angular web framework. Owing to its comprehensive functionality and strong community support, Angular is a suitable choice for developing medium-complexity applications. Due to the simplicity of integration, rapid development speed, minimal codebase, and ease of use, augmented reality functionality is implemented through the integration of the AR.js and A-Frame libraries. These

libraries enable the use of both marker-based and markerless AR within web environments.

A server-side component is also required for the complete implementation of the application. Given that Angular is based on TypeScript – a language closely related to JavaScript, JavaScript was chosen for the server-side logic. Node.js was selected as the server platform due to its popularity and ease of use, while the Express.js framework was employed to accelerate development and simplify configuration.

The application development process consisted of several key stages, each of which played an important role in the overall architecture of the project (Fig. 1).

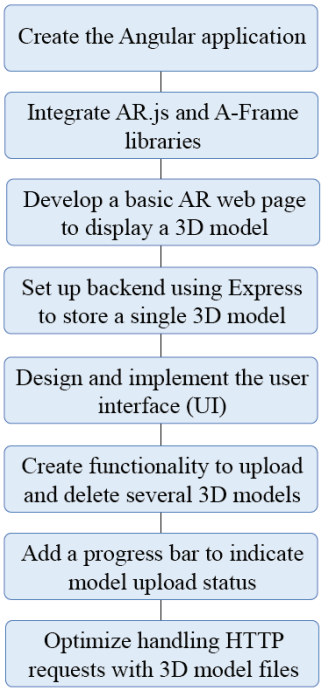


Fig. 1. Development scheme of a client-side application with augmented reality elements, divided into stages

The first three stages were aimed at developing a basic working version of the application with minimal functionality. The initial version of the application consisted of a web page displaying a single AR object.

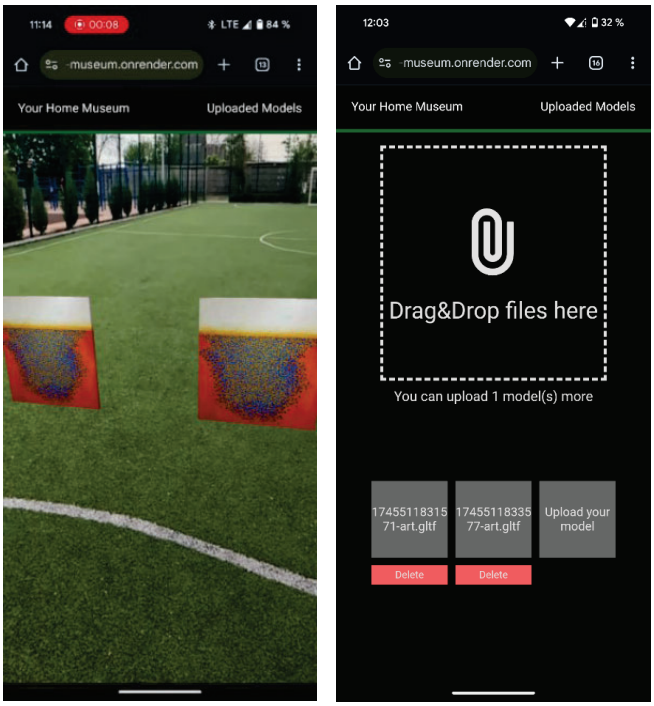
At the fourth stage, the server-side component was integrated to enable the storage and transmitting of the AR object.

Subsequent stages focused on extending the application's functionality, including support for multiple AR objects, mechanisms for uploading and deleting objects, and improving the user interface for enhanced usability.

The final stage, carried out after the planned functionality had been implemented, involved modifying the handling of HTTP requests related to model transmission. A detailed description of this process is provided in the following section.

As a result of the development process, a client-side web application with support for augmented reality elements was implemented, along with a server-side part responsible for storing 3D AR object files. A markerless approach using geolocation data was employed as the augmented reality technology. The outcome is presented in Fig. 2a.

The input data for the client application are .gltf files that contain the description of 3D models. Users are able to upload their own files from a local device. The upload functionality is implemented on a dedicated page of the application – **Uploaded Models** (Fig. 2b).



a b

Fig. 2. Results of client application development: a – display of augmented reality elements; b – page of loaded augmented reality objects

The output of the client application is a real-time 3D visualization of models in an augmented reality environment, rendered through the device's camera (Fig. 2a). In addition, a dynamic list of uploaded models is generated and automatically updated whenever objects are added or removed (Fig. 2b).

The result of the work is an interactive client-side application with a user interface that supports the uploading, management, and visualization of 3D objects in an augmented reality environment, accompanied by a server-side component responsible for model storage and processing user requests.

V. Method of Transmission of Multi-Model Augmented Reality Objects

The processing of HTTP requests associated with the transmitting of AR models involves several key stages. This process spans both the client-side and server-side components and includes a sequence of typical operations. Let us introduce a set of variables that denote the execution time of each individual stage:

- t_1 – HTTP request initiation time;
- t_2 – authentication and authorization time (if necessary);
- t_3 – time to process the request and send a response;

- t_4 – time to load and parse the model;
- t_5 – time for visualization in the AR scene;
- t_6 – time for logging and analytics (optional).

Typically, the first three stages are performed using the standard HttpClient.get() mechanism used in the Angular framework [17]. Then the total time required to transfer the AR models is as follows:

$$T\Sigma = T1 + T2 = \sum_{i=1}^{3} t_i + \sum_{j=4}^{6} t_j, \quad (1)$$

where $T1$ – time spent using the HttpClient.get() method; $T2$ – time spent using other methods.

All stages are executed within the main thread, which may lead to blocking of the user interface (UI). This implies that, in cases involving large model sizes or computationally intensive processing, delays in rendering or temporary UI freezing may occur. Such behavior is generally perceived negatively by users; therefore, a time threshold $TMAX$ is typically defined for the total processing duration $T\Sigma$:

$$T\Sigma < TMAX. \quad (2)$$

In multi-model AR, the third stage typically accounts for the largest portion of the total processing time. Therefore, if condition (2) is violated, it is proposed to use Web Workers instead of the standard HttpClient.get() mechanism.

While handling HTTP requests related to AR model transmission, Web Workers offer potential benefits in terms of performance; however, they also present several drawbacks:

- you can load the model in the web worker, but inserting it into the scene is impossible without passing it back to the main thread;
- difficult transfer of "heavy" resources back to the main thread;
- Web Worker cannot participate in the rendering loop or event handling;
- Web Worker needs to manually transfer resources.

However, the use of web workers enables network operations and data decoding to be performed in the background, without blocking the main application thread, thereby allowing parallel updates of the user interface. As a result, the main thread is freed from handling requests, which enables uninterrupted event processing and UI rendering, reducing the overall perceived latency for the user, therefor

$$T\Sigma = t_1 + t_2 + \sum_{j=4}^{6} t_j + t_7, \quad t_7 << t_3, \quad (3)$$

where t_7 – the time required to transfer the generated models to the main thread.

VI. Research of a Method for Transmitting Multi-Model Augmented Reality Objects

A. Client application testing

At the first stage of the study, testing of both the basic and the modified client application was conducted.

The testing process of the client application with augmented reality elements was carried out in several stages (Fig. 3).

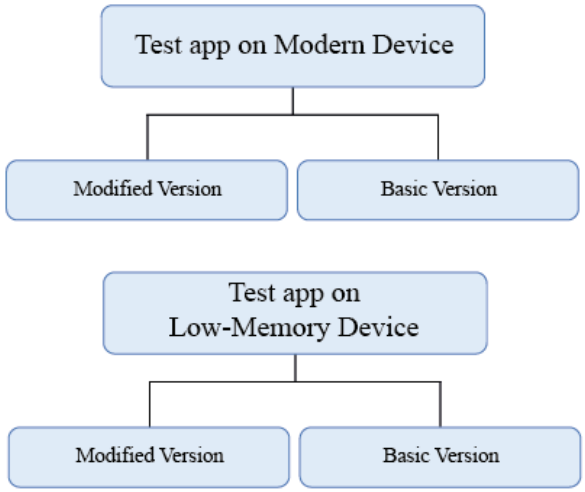


Fig. 3. Scheme of testing a client application with augmented reality elements

The efficiency evaluation was conducted by measuring the time required to load a single 3D model of 16.4 MB in two versions of the application: the modified version (using web workers) and the basic version (based on HttpClient.get()). The testing was carried out using both a modern device and a device with limited RAM capacity.

As a modern device, the Google Pixel 8 smartphone (2023), equipped with 8 GB of RAM, was used. Based on six test runs, the results were obtained and are presented in Table 2.

TABLE II. 3D OBJECT LOADING TIME ON THE MODERN DEVICE

Test	Version loading time, sec	
	modified	***basic***
1	3.06	7.46
2	3.44	8.41
3	5.45	8.08
4	4.67	7.91
5	3.56	7.74
6	3.68	7.80
Average value	**3.98**	**7.90**

The modified version of 3D object loading on a modern device demonstrates higher performance, providing an average 49.62% faster loading speed compared to the basic version.

The device with limited RAM used in testing was a Xiaomi Redmi 4X smartphone, released in 2017, with 3 GB of RAM. The results of the experiment are presented in Table 3.

TABLE III. 3D OBJECT LOADING TIME ON THE DEVICE WITH LIMITED RAM

Test	Version loading time, se	
	modified	***basic***
1	6.92	11.30
2	5.91	13.09
3	5.88	10.11
4	7.04	9.72
5	6.39	9.62
6	9.33	9.38
Average value	**6.91**	**10.54**

Based on the results in Table 3, it was found that the modified version demonstrates higher performance, providing an average 34.45% faster loading speed compared to the basic version.

Figure 4 presents a comparative analysis of 3D object loading time depending on the type of device and application version.

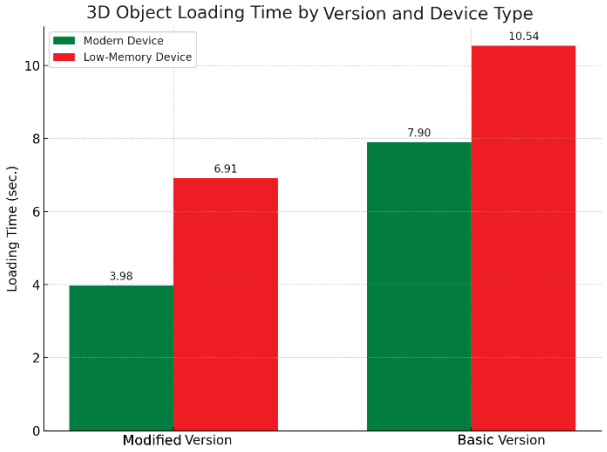


Fig. 4. Comparison of average 3D object load times across different devices and versions

The comparison of the graphs demonstrates that the amount of RAM has a significant impact on loading speed. Specifically, the modified version on a modern device loads 42.40% faster than on a device with limited memory. In the case of the basic version, this difference is 25.05%.

B. Evaluation of the effectiveness of the proposed method

The effectiveness of the proposed method was studied using a model of the functioning of a client application with augmented reality technology. The values analyzed in the model were specified in arbitrary units:

- total time required to transfer AR models ($T\Sigma$);
- total amount of multi-model AR objects ($V\Sigma$);
- total amount of computing resource allocated for transmitting AR models ($R\Sigma$).

In Fig. 5–8 the following notations are introduced:

- B – basic method;
- M – modified method.

The nature of the dependence of the total time required to transfer AR models on the total volume of multi-model AR objects is shown in Fig. 5 ($R\Sigma = 1$) and Fig. 6 ($R\Sigma = 10$).

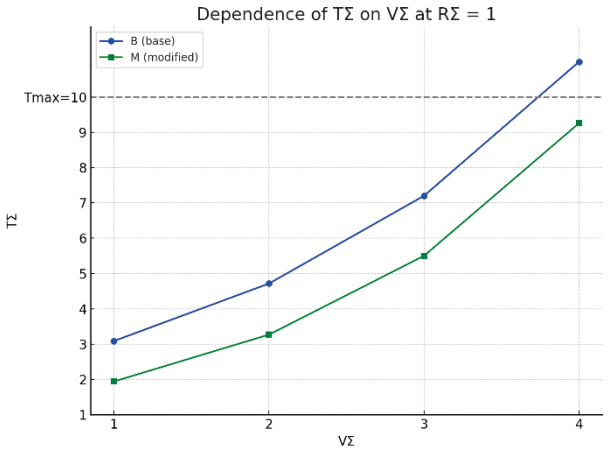


Fig. 5. Dependence of $T\Sigma$ on the total volume of multi-model AR objects at $R\Sigma = 1$

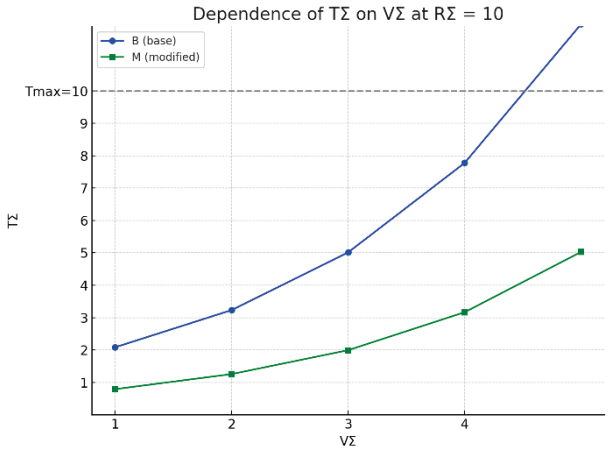


Fig. 6. Dependence of $T\Sigma$ on the total volume of multi-model AR objects at $R\Sigma = 10$

The nature of the dependence of the total time on the total amount of computing resources allocated for the transfer of AR models is shown in Fig. 7 ($V\Sigma = 1$, $V\Sigma = 4$) and Fig. 8 ($V\Sigma = 10$).

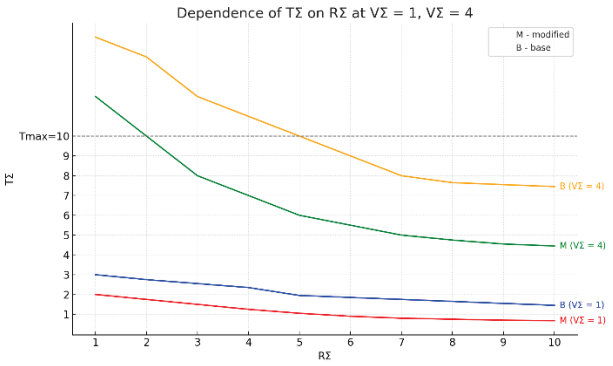


Fig. 7. Dependence of $T\Sigma$ on the total amount of computing resources allocated for the transmitting of AR models at $V\Sigma = 1$ and 4

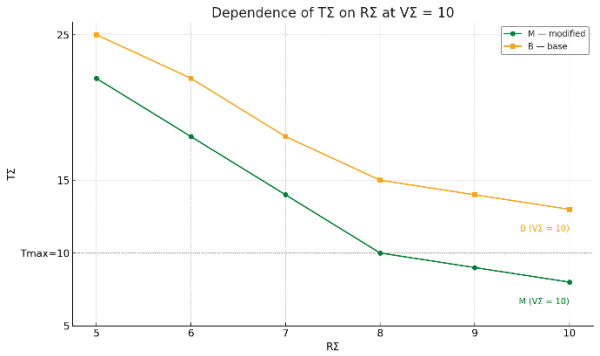


Fig. 8. Dependence of $T\Sigma$ on the total amount of computing resources allocated for the transmitting of AR models at $V\Sigma = 10$

VII. Discussion of Results and Conclusions

This study substantiates the relevance of developing and using a client application with augmented reality (AR) technology. An analysis of AR technology was conducted, various tools for implementing AR were reviewed, and existing works on the development of AR websites were examined.

Based on the conducted analysis, appropriate tools were selected for the development of a client application with AR functionality. A client-side AR application was developed and tested with sample use cases. Additionally, a modification to the basic method of handling HTTP requests related to AR model transmitting was proposed. Specifically, the use of web workers was suggested for processing requests and handling responses.

Comparison of the basic and modified methods based on the operational model of the client application using augmented reality technology demonstrated the universal feasibility of applying the modified method. Specifically, at a value of $R\Sigma = 1$, the efficiency gain from using the modified method is less pronounced. However, as the available computational resources increase, the modified method provides stable performance metrics. Starting from $R\Sigma > 8$, it clearly outperforms the basic method, ensuring compliance with the time constraints (2) at $V\Sigma = 10$ and for $R\Sigma > 2$ at $V\Sigma = 4$. Therefore, the modified method is suitable for use across a wide range of scenarios, regardless of conditions.

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Synthesis, Structural Characteristics, and Phase Transformations of Iron-Based Nanoparticles Obtained by Electric Spark Treatment and Their Application in Steel Modification

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Abstract*—This study investigates the synthesis, structural characteristics, and transformative properties of iron-based nanoparticles obtained via electric spark treatment of iron granules in water and organic liquids. Advanced analytical methods, including scanning electron microscopy (SEM), X-ray diffraction (XRD), X-ray photoelectron spectroscopy (XPS), and differential scanning calorimetry (dsc), were utilized to evaluate nanoparticle size, shape, phase composition, and crystal lattice parameters. Results revealed nanoparticles predominantly in α-Fe, γ-Fe, FeO, and Fe_3O_4 phases, demonstrating significant structural heterogeneity and increased lattice distortion due to high dislocation densities (2×10^{14}–6×10^{14} cm^{-2}). Thermodynamic analyses illustrated distinct phase transformation behaviors compared to bulk iron, characterized by reduced transformation temperatures and narrower intervals. Practical application as microstructural modifiers was explored by incorporating these nanoparticles into 45L steel, significantly refining ferrite and pearlite grain structures and enhancing structural homogeneity. This research underscores the electric spark method's efficacy in tailoring iron-based nanoparticles' structural and energetic properties, highlighting their potential for advanced metallurgical applications and materials engineering innovations.*

***Keywords*—iron nanoparticles, electric spark treatment, phase transformations, structural defects, dislocation density, crystal lattice parameters, steel microstructure modification, nanoparticle synthesis, x-ray diffraction (xrd), differential scanning calorimetry (dsc)**

I. Introduction

Existing research work outlines that a feature of a material in the nanoscale state is the presence of a large number of atoms with uncompensated bonds on its surface [1-5]. This predetermines nanoobjects' increased free surface energy and their intensive interaction with the environment.

Among the numerous methods for producing metal nanoparticles, electric spark treatment of granules in a liquid is the most promising. Such a process combines complex temperature and deformation effects in forming nanoobjects with the high manufacturability of the method. Works [2-8] show that in the process of electric spark treatment, erosion products are formed at high temperatures, pressures, and cooling rates, which creates prerequisites for forming energy-saturated highly dispersed nanoobjects with a wide range of structural states and properties.

Taking into account the high efficiency of the use of nanoparticles [9-12], including those obtained by electric spark treatment of metal granules in liquids, it was of interest to study the possibility of obtaining iron-containing nanoparticles using this method, their chemical composition, structure, and patterns of phase transformations, and to assess the effectiveness of their effect on the structure of iron-carbon alloys.

II. Materials and Methods

The object of the study was nanoparticles obtained in the process of electric spark treatment of iron granules in liquids, according to the methods described in the works [1-8]. Granules and electrodes were made of material of the following chemical composition (% (wt): 0.025C, 0.25Mn, 0.015P, 0.10S, 0.08Si, 0.10Cu, the rest was Fe. Water and organic solutions were used as working fluids.

The particle size, shape, and chemical composition were determined using scanning electron microscopes Jeol-6490LV, Jeol JSM6360, and Hitachi SU8000 SC200D.

The phase composition, parameters of the crystal lattice of nanoparticles, and the size of subgrains (blocks, regions of coherent scattering (CRP)) were studied by X-ray diffraction in monochromatic Cu Kα-radiation on the DRON-UM1 diffractometer. X-ray photoelectron spectroscopy (XPS) on the ES-2401 spectrometer analyzed the chemical composition of the nanoparticle surface. The charge state of iron and oxygen atoms was determined by the bonding energies of Fe2p3/2 and O1s electrons with the atoms.

The regularities of phase transformations were studied using the STA 449 F1 Jupiter thermal analyzer using the method of synchronous thermal analysis. The thermal effects of phase transformations (dsc), the change in mass (TG) of

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nanoparticles during heating and cooling in argon at a rate of 20 °C/min, were also measured simultaneously. The accuracy of temperature measurement was ± 0.25%, enthalpy - ±3%, mass - ±2 μg.

To assess the effectiveness of nanoparticles obtained during electric spark treatment of iron, their effect on the microstructure of 45L steel in the cast state and after annealing at 860 °C was studied. The microstructure of 45L steel was studied in its initial state (without nanoparticle additives) and after processing the melt with nanoparticles.

III. Research results

Thermodynamic analysis of the characteristics during electric spark treatment of metals in a liquid shows that the process is accompanied by the decomposition of the working medium in the case of water, as well as the formation of oxidizing agents (O, O2, H_2O, OH) and reducing agents (H_2, H), in an organic liquid - C, CO, CO_2 and other compounds. This creates prerequisites for both oxidation and reduction of electrospark erosion products.

X-ray diffraction analysis of the phase composition of nanoparticles indicates that, depending on the technological parameters of their production, nanoparticles α- and γ-Fe, FeO, and Fe_3O_4 oxides are formed.

When nanoparticles are formed in water, their sizes vary from 15 to 60 nm (Fig. 1, a) and from 160 to 700 nm (Fig. 1, b), in an organic liquid, from 10 to 20 nm (Fig. 1, c). The average size values are 30, 380, and 15 nm. The shape coefficient of nanoparticles (the maximum size ratio to the minimum) varies from 1 to 2.4.

X-ray diffraction analysis showed that the sizes of subgrains of α-Fe, γ-Fe, FeO, and Fe3O4 nanoparticles vary from 1 to 93 nm with mean values of 39, 15, 4, and 1 nm, respectively (Table 1). Therefore, the following nanoparticle structure is possible depending on the ratio of the size of the D_{NP} nanoparticles and the regions of coherent D_{OC} scattering. In $D_{np} » D_{rcs}$, a single-crystal structure of nanoparticles is most likely; at $D_{np} > D_{rcs}$, the nanoparticles will consist of several subgrains; at $D_{np} >> D_{rcs}$, they will consist of grains containing subgrains.

Diagrams and photographs of the structure of nanoparticles consisting of single crystals, subgrains containing subgrains are shown in Fig. 2. The study of the structure of grain boundaries in a-Fe nanoparticles showed that they consist of groups of several crystal lattices and dislocations disoriented relative to grains (Fig. 3, a). Subgrain boundaries consist of crystal lattices conjugated with contacting subgrains and dislocations in places where the atomic planes of one of the subgrains are not conjugated and broken (Fig. 3, b).

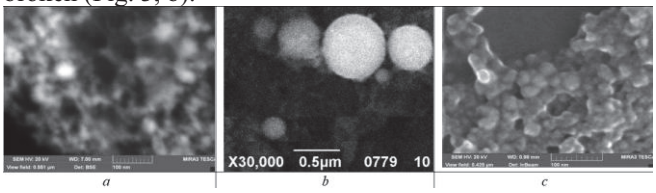


Fig. 1. Sizes of nanoparticles obtained in the electric spark treatment of iron granules in water (a, b) and organic liquid (c).

TABLE I. Sizes (nm) of subgrains (blocks) of the studied nanoparticles

Dimension parameter	Type of Nanoparticles			
	α-Fe	*γ-Fe*	*FeO*	*Fe_3O_4*
Minimum	15	7	3	≈1
Maximum	93	24	6	≈1
Average	39	15	4	≈1

The size of crystal lattices has been studied. Nanoparticles α- and g-Fe obtained in an organic liquid are characterized by higher values of crystal lattice parameters than massive materials (Fig. 4, a, b). The opposite pattern is observed for FeO and Fe_3O_4 nanoparticles obtained in water (Fig. 4, c, d).

The size of the crystal lattice of nanoparticles depends on the value of the surface tension, which compresses the nanoparticle and decreases the parameter of its crystal lattice and the content of impurity elements that increase its value. The increase in the crystal lattice parameter of nanoparticles α- and γ-Fe is apparently due to the dissolution of carbon obtained due to the dissociation of an organic liquid in ferrite and austenite. In the case of oxide nanoparticles, they are compressed under the influence of surface tension.

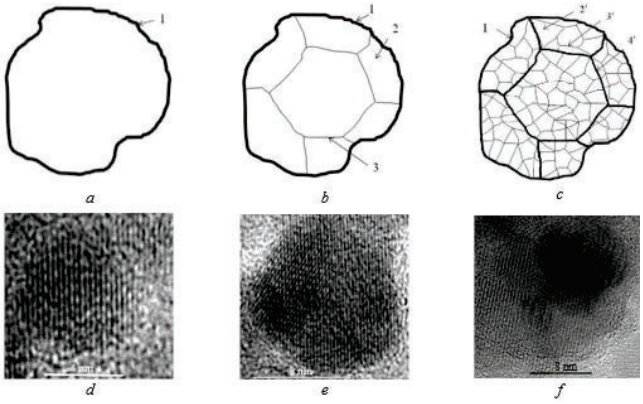


Fig. 2. Diagrams of the structure (a—c) and structure (d—f) of single-crystal (a, d) consisting of subgrains (b, e) and grains containing subgrains (c, f), nanoparticles: 1 — nanoparticle; 2, 2′ — subgrains; 3, 3′ — subgrain boundaries; 4′ — grain boundaries.

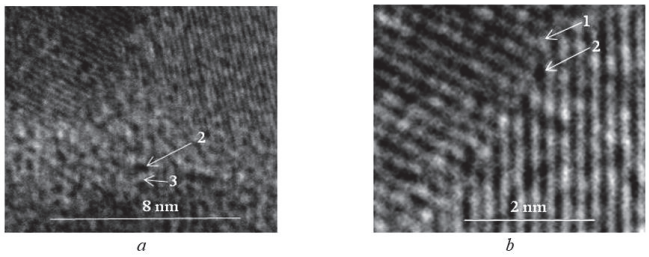


Fig. 3. Structure of the boundaries of grains (a) and subgrains (b) of α-Fe: 1 — crystal lattice conjugated with contacting subgrains; 2 — nucleus of marginal dislocation; 3 — group of crystal lattices disoriented relative to grains.

It should be noted that the dissolution of oxygen formed as a result of water dissociation in FeO and Fe_3O_4 nanoparticles also contributes to a specific decrease in the size of their crystal lattices, since with an increase in the oxygen content in oxides within their homogeneity, there is a decrease in the parameters of their crystal lattices [13-15].

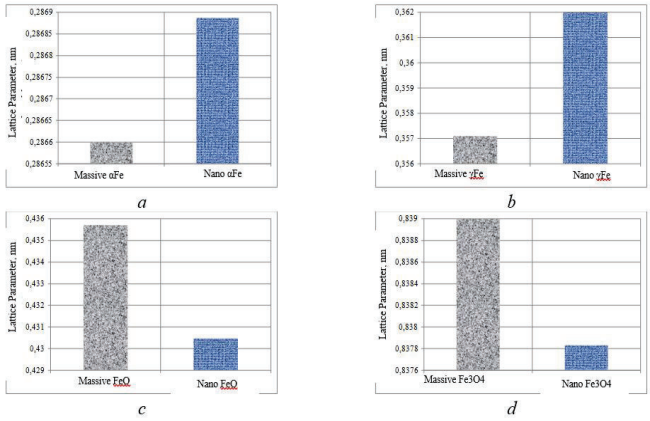


Fig. 4. Parameters of crystal lattices of massive materials and nanoparticles α-Fe (a), γ-Fe (b), FeO (c), and Fe3O4 (d) obtained by electric spark treatment of iron granules in organic liquid (a, b) and water (c, d).

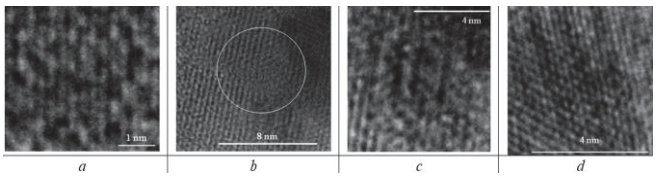


Fig. 5. Defects in the crystal structure of a-Fe nanoparticles: a, b-dislocations; c-the source of the Frank-Reed dislocations; d-doubles.

The study of the fine structure has shown that in the process of electric spark processing of granules, nanoparticles are formed with such defects in the crystal structure as dislocations and twins, as well as sources of generation of linear defects (Fig. 5). Distortion of crystal lattices is observed around the nuclei (Fig. 5, a) and dislocation loops (Fig. 5, b). Analysis of the defects in the crystal structure shown in Fig. 5 showed that in nanoparticles a-Fe, there is a chaotic distribution of dislocations with a density of 2×10^{14} to 6×10^{14} cm^{-2}.

Known technological operations for hardening of massive metals and alloys [16-20] make it possible to obtain the following levels of dislocation density:

- from 10^8 up to 10^{11} cm^{-2} in case of plastic deformity;
- from 10^{10} to 10^{12} cm^{-2} as a result of phase transformations (for example, the formation of martensite);
- up to 10^{13} cm^{-2} when intense plastic deformation and phase transformations are combined, or under nonequilibrium crystallization conditions (e.g., high-velocity galvanic deposition in a cold electrolyte or sputtering with a high-density ion beam on an intensively cooled substrate).

The level of dislocation density in nanoparticles obtained in the process of electric spark treatment (~10^{14} cm-2) approached the limit. It reached more than 103—104 °C/s as a result of the combined action of shock waves arising in the process of pulsed expansion of electric spark channels under the influence of pressure of about a dozen megapascals and a high cooling rate of more than 103—104 °C/s. The level of the lower limit of the nanoparticle cooling rate was determined based on the condition that g-Fe fixation at room temperature and normal pressure is possible at a cooling rate of more than 103—104 °C/s.

Comparison of the density of dislocations in nanoparticles and massive materials obtained by known methods shows that the level of defects in the crystal structure of nanoparticles after electric spark treatment is much higher than the values achieved by known hardening metals and alloys. A high level of dislocation density in nanoparticles predetermines their high energy saturation, because the main contribution to the energy of the substructure and subgrain boundaries is made by the elastic energy of dislocations, which, in the case of the presence of boundary and helical dislocations, is determined by the following dependence:

$$\Delta U = \frac{\rho G}{4\pi} \cdot ln\frac{R}{r_n} \cdot \left(\frac{b_k^2}{(1-\nu)} + b_c^2\right), \quad (1)$$

,where ρ is the density of dislocations; G is the shear modulus; r_n is the radius of the nucleus of dislocations; R is the shielding radius of the elastic field of dislocations (in the first approximation, it is equal to the average distance between dislocations); b_k and b_c are the Burgers vector of marginal and helical dislocation, respectively; n is Poisson's ratio.

Analysis of experimental data from differential scanning calorimetry (dsc) (Fig. 6), thermogravimetry, electron microscopy (Fig. 7, 8) and the results of thermodynamic calculations allows us to interpret the process of phase transformations in nanoparticles obtained by electric spark treatment of iron granules in an organic liquid and containing 45.5% α-Fe and 54.5% γ-Fe as follows. In the first heating of nanoparticles, polymorphic transformations began at 230 °C with a γ-Fe -> α-Fe transformation that reached a maximum at 336 °C and ended at 390 °C. At 397 °C, the oxygen-containing film covering the nanoparticles began to break down. The process ended with the formation of Fe_3O_4 oxide and oxygen at 460 Reduction of iron from Fe_3O_4 with carbon was fixed at 597 °C and ended at 800 °C (Fig. 6).

Analysis of the data given in Fig. 7 showed that at the first heating of nanoparticles, the temperature of the beginning of the conversion of α-Fe to γ-Fe decreased by 4% compared to the equilibrium value. The interval of conversion of α-Fe to γ-Fe in nanoparticles was 90 °C. A feature of phase transformations in nanoparticles was that after α-Fe -> γ-Fe transformation in the interval of 800-1050 °C, austenite and graphite appeared in them, which began to dissolve in austenite at temperatures of 1050-1070 °C (Fig. 6). At the same time, the temperature of coexistence of austenite and graphite was 250 °C, and the dissolution temperature of graphite was 20 °C with a ratio of 93 and 7%, respectively.

In massive iron, the dissolution of graphite began immediately after α-Fe -> γ-Fe- Transformations (738^{OS}) and lasted until 1038^{OS}. At the same time, the temperature changed insignificantly and amounted to 270 and 300, respectively, ^{OS}.

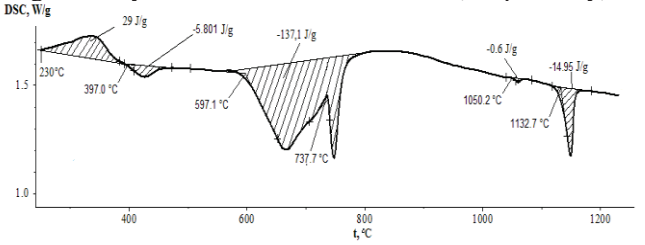


Fig. 6. Results of differential scanning calorimetry of nanoparticles having the following chemical composition (% (wt) α-Fe = 15.4; FeO = 7,2; Fe_3O_4 = 77,4.

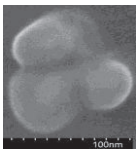


Fig. 7. Iron nanoparticles partially coated with graphite.

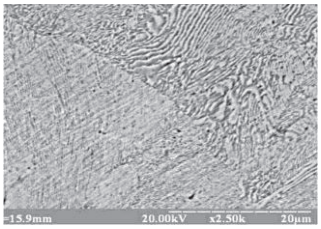


Fig. 8. The steel structure obtained as a result of heat treatment of nanoparticles, which are shown in Fig. 7.

TABLE II. EFFECT OF HEAT TREATMENT (MOT), INCLUDING HEATING UP TO 1300 °C AND COOLING IN ARGON AT A RATE OF 20 °C/MIN, ON THE CHEMICAL COMPOSITION (% (MASS)), CRYSTAL LATTICE PARAMETER A AND THE SIZE OF THE AREAS OF COHERENT DISPERSION OF REGION (R) OF IRON-CONTAINING NANOPARTICLES AND PHASES OBTAINED AS A RESULT OF POLYMORPHIC TRANSFORMATIONS

Original composition	**Condition**	**Parameter**	**Phase Type**				
			α-Fe	**γ-Fe**	**FeO**	**Fe_3O_4**	**Carbon**
α-Fe, FeO, Fe_3O_4	Before HT	Composition, % (mass)	15,4	—	7,2	77,4	—
		a, nm	0,2864	—	0,4301	0,8372	—
		R, nm	53	—	3,2	1	—
	After HT	Composition, % (mass)	7,34	—	92,66	—	—
		a, nm	0,2866	—	0,4303	—	—
		R, nm	29	—	2,5	—	—
α-Fe, γ-Fe	Before HT	Composition, % (mass)	45,5	54,5	—	—	(1—2)*
		a, nm	0,2877	0,3624	—	—	—
		R, nm	15	24	—	—	—
	After HT	Composition, % (mass)	100	—	—	—	1,5*
		a, nm	0,2867	—	—	—	—
		R, nm	58	—	—	—	—

Research results showed that the temperature intervals of phase transformations, as a result of which gaseous products of reactions were formed, correlated with a decrease in the mass of the system, and amounted to 3.24 and 6.19%.

The transformation of α-Fe → γ-Fe occurred in the temperature range of iron reduction and began at temperatures ≈705-710 °C, and ended at 800 °C. In the range of 1050-1070 °C, graphite dissolved in austenite, which at 1132.7 °C begins to melt and at 1172 °C completely turns into a liquid state.

In the process of cooling the melt in the temperature range from ≈1125 to 1040 °C, austenite was formed, from which graphite begins to be released at 1109 °C and 930 °C. The formation of graphite is because the heat treatment of nanoparticles was carried out under conditions of low cooling rates (20 °C/min), which is less than the critical rate (35 °C/min), and the excess of which leads to the separation of cementite from austenite. The conversion of γ-Fe → α-Fe begins at 692 and ends at 643 °C.

When the cooled material is reheated, the conversion of α iron into γ iron, the dissolution of graphite in γ iron, and the melting of austenite are completed at lower temperatures than in the first case. At the same time, the temperature intervals of phase transformations are reduced, respectively, by 50, 5, and 27% compared to the first heating. In the process of cooling after the 2nd heating, identical phase transformations are observed in the temperature ranges close to cooling after the first heating.

More significant differences were observed in the analysis of the temperature range of austenite. For nanoparticles, the process of dissolution of graphite in austenite ends at a temperature 10-30 °C higher than in massive iron, but the temperature range of austenite existence is 3-4 times shorter than in ordinary iron lower than in massive material, while the temperature range of the process is reduced by 5-6 times.

In the process of cooling the liquid metal obtained after melting the nanoparticles, solidification begins with the release of austenite, the formation temperature of which is 340-350 ^{OS} lower, and the temperature interval of the process is 2-3 times shorter than that of massive iron. The peculiarity of solidification is that austenite and graphite are almost simultaneously separated from the liquid phase, and the total temperature range of their existence before the beginning of the γ-Fe → α-Fe transition into 1,7-1.8 times less than conventional iron.

The results of measuring the heat of phase transformations observed when heating nanoparticles (Fig. 6) and comparing them with the data for massive iron showed that when heating nanoparticles, the heat of conversion of γ-iron into α-iron, α-Fe into γ-Fe, and melting of austenite, respectively, is 1.6, 2.1, and 18 times less than the equilibrium values. In the case of cooling of the material obtained after the melting of nanoparticles, the heat of transformation of γ-iron to α-iron is 3.2 times less than the equilibrium iron.

It should be noted that during the repeated heating and cooling of the iron-carbon alloy obtained due to the melting of nanoparticles and solidification, almost identical to nanothe laws of phase transformations. This indicates a significant influence of heredity on the process of formation of material obtained from nanoobjects.

Analysis of the study of the structure and phase transformations in nanoparticles shows that nanoobjects obtained during the electric spark treatment of iron have an increased level of defects in the crystalline structure and other patterns of phase transformations compared to massive iron. This creates the prerequisites for the effective interaction of nanoparticles with the environment, for example, when introduced into liquid metal.

To study the possibility of using nanoparticles as modifiers of the structure of metals and alloys, a series of experiments on the modification of the structure of steel with 45 iron nanoparticles were carried out. The results of inoculation of steel with 45 iron nanoparticles are shown in Fig. 9.

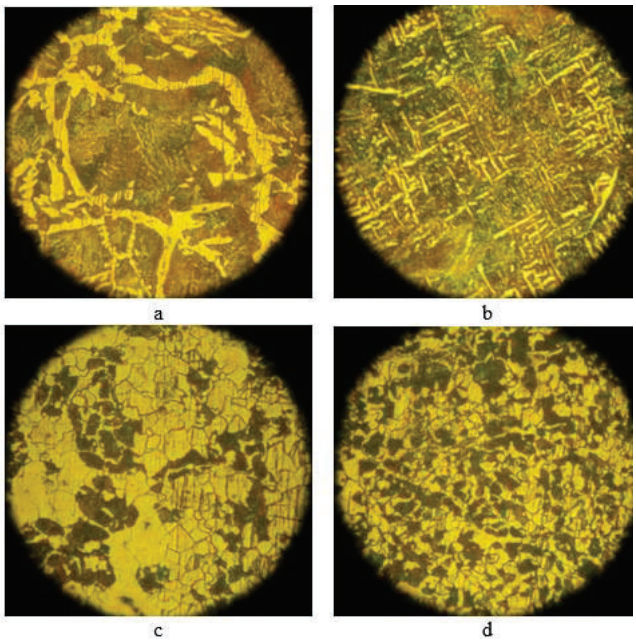


Fig. 9. Effect of nanoparticles (b, d) obtained during electric spark treatment of iron on the microstructure of steel 45L (a, c): a, b - cast state; c, d - annealing at 860 °C (magnification x 100).

In the case of nanoparticles containing iron for processing metal melts, the regularity of their interaction with the melt changes. This is due to the fact that when ordinary iron powder is introduced into a metal melt, the melt is supercooled at the points of its contact with powder particles due to heat consumption for heating and phase $\alpha \rightarrow \gamma$, $\gamma \rightarrow \delta$, $\delta \rightarrow$the liquid of transformation in particles. In the process of processing the melt with nanoparticles containing iron, the melt does not supercool. Still, it overheats at the moment of contact with the nanoparticles when the temperature of $\rightarrow \alpha$-transformation is reached.

Quantitative computer analysis of microstructures (Fig. 9) showed that after inoculation of 45L steel with nanoparticles, the amount of pearlite in the cast state increased from 62.6 to 72.5%, that is, by 16%, and after annealing, from 46.7 to 66.1%, that is, by 42%. Statistical analysis of changes in the microstructure shows that spheroidization of grains, a decrease in their size, and a narrowing of the size range by 6 times are observed. Inoculation led to the dispersion of ferrite grains in the cast state by 2.3-2.4 times and to an increase in the homogeneity of the structure by 1.7-2.4 times. After annealing, the size of ferrite grains decreased by 1.2-2.4 times, and perlite by 1.2-1.5 times, while the homogeneity of the structure increased by 1.2-1.8 times.

IV. CONCLUSIONS

Through rigorously documented experimental procedures, this study has demonstrated that electric spark treatment of iron granules in liquid media is an effective and controllable method for synthesizing iron-based nanoparticles with tailored size distribution, phase composition, and lattice characteristics. Comprehensive structural and thermodynamic characterization was carried out using scanning electron microscopy (Jeol-6490LV, Jeol JSM6360, Hitachi SU8000 SC200D) for morphology and particle size, X-ray diffraction (DRON-UM1, Cu Kα radiation) for phase identification and coherent scattering region size determination, X-ray photoelectron spectroscopy (ES-2401) for surface chemical state analysis, and synchronous thermal analysis (STA 449 F1 Jupiter) for DSC/TG measurements, following established and referenced methodological standards.

The experimental results confirmed the synthesis of nanoparticles predominantly composed of α-Fe, γ-Fe, FeO, and Fe_3O_4 phases, with coherent scattering regions ranging from 1 to 93 nm. The proposed model relating nanoparticle size to crystalline substructure was validated by combining high-resolution microscopy and XRD measurements. Nanoparticles exhibited exceptionally high dislocation densities (2×10^{14}–6×10^{14} cm^{-2}), far exceeding those achievable in conventionally strengthened bulk metals, indicating high structural energy saturation. Differential scanning calorimetry revealed distinct phase transformation behaviors compared to bulk iron, including reduced $\alpha\leftrightarrow\gamma$ transformation onset temperatures, narrower transformation intervals, and shortened austenite stability ranges by factors of three to six, highlighting the influence of nanoscale structure on thermodynamic processes.

The practical applicability of these nanoparticles as steel microstructure modifiers was verified through quantitative metallographic analysis of 45L steel inoculated with the synthesized particles. Using optical microscopy at ×100 magnification with calibrated scale references, it was established that nanoparticle addition increased pearlite content in the cast state by 16% and after annealing at 860 °C by 42%, refined ferrite grain size by up to 2.4 times, reduced pearlite grain size by up to 1.5 times, and enhanced structural homogeneity by up to 2.4 times. These modifications are attributable to the documented interactions between high-energy nanoparticles and the steel melt, supported by the controlled experimental design.

Overall, the findings provide reproducible, verifiable evidence that electric spark–synthesized iron nanoparticles possess unique structural and phase transformation properties that are directly responsible for substantial improvements in steel microstructure. The combination of detailed methodological transparency, precise measurement, and quantitative analysis ensures that the results can be independently validated and applied to advanced metallurgical processing and materials engineering.

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Similarity Criteria in Models of Energy-efficient Transport Systems

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***Abstract*— The paper considers modelling of material transportation processes in electrical systems using belt conveyors as an example. The main attention is paid to formalization and construction of similarity criteria necessary for analysis, comparison and design of transport systems with both deterministic and stochastic input material flows. Based on the analysis of equations describing the behaviour of flow parameters of the transported material, a dimensionless transformation is carried out using the π-theorem, which allows identifying the key dimensionless parameters of the model. It is shown that the similarity criteria allow generalizing the results of experiments conducted on one system and extending them to other transport systems, provided that the values of the dimensionless characteristics coincide. This ensures the possibility of a unified approach to the development of energy-saving control systems, even in conditions of incomplete or limited experimental data. The paper constructs tables of parameter dimensions and corresponding similarity criteria for both deterministic and stochastic input models. It is demonstrated that the use of the introduced criteria ensures the equivalence of the models and allows the use of unified numerical solution methods. The presented methods and approaches have practical significance in the design, modelling and optimization of modern transport systems in the mining and electrical engineering industries.**

Keywords*— *conveyor, energy saving, material loading coefficient, similarity criteria

I. Introduction

Modern problem in the field of electrical engineering modelling increasingly require taking into account the interaction of mechanical, energy and information components, especially in the context of production systems control [1]. One such example is conveyor-type transport systems, widely used in the mining and processing industries [2, 3, 4]. Modelling and simulation of such systems for the purpose of optimizing their energy and dynamic characteristics are of considerable interest for the development of intelligent control systems [5, 6, 7, 8].

The results of processing the experimental data on the operation of conveyors make it possible to identify stable patterns in the transportation of material and to assess the influence of various factors on the energy efficiency of the system [9, 10, 11]. However, to transfer these results to new operating conditions in which experimental data are absent, a theoretical basis is needed to generalize the behaviour of the system. Such a basis is modelling using dimensionless parameters and similarity criteria.

This paper is devoted to the construction of a system of dimensionless parameters describing the behaviour of a transport system under deterministic and stochastic modes of material supply. A method for theoretically selecting the minimum required number of determining parameters reflecting key physical phenomena in the most universal form is considered. The resulting models allow for numerical simulation, predicting the behaviour of systems when changing modes, and optimizing flow control from the point of view of energy saving, which is especially important in the context of electro-technical modelling of complex production processes.

II. Problem Statement

In the context of the active implementation of digital modelling methods in electrical engineering systems, the problem of formalizing and transferring experimental results between different objects is becoming especially relevant [12, 13]. In particular, when modelling conveyor transport systems closely related to electric drives and automatic control systems [14, 15], it is necessary to take into account issues of physical and mathematical similarity. One of the key problems is to determine the conditions under which the modelling results or experimental data obtained for one specific transport system can be extrapolated to other systems with different design and operational parameters. In this context, similarity criteria act as a fundamental tool that allows for adequate simulation and the development of universal approaches to material flow management. This is especially important when designing energy-saving optimal control algorithms for conveyor-type transport systems [14, 16], where it is necessary to take into account both deterministic and stochastic components of the flow [17, 18]. In such conditions, the use of dimensionless parameters and similarity criteria becomes not just an auxiliary, but often the only possible theoretical method for constructing and verifying the developed control model.

III. Transport System Model

The conveyor model for describing the process of material transportation is studied in detail in [19] and can be represented in the following form by a partial differential equation:

$$\frac{\partial \chi_0(t,S)}{\partial t}+\frac{\partial \chi_1(t,S)}{\partial S}=\delta(S)\lambda(\mathrm{t}), \tag{1}$$

$$\chi_0(0,S)=H(S)\Psi(S), \tag{2}$$

$$\chi_1(t,S)=a(t)\chi_0(t,S), \tag{3}$$

where $\chi_1(t,S)$, $\chi_0(t,S)$ flow and linear density of the material transported by a conveyor of length S_d at point

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$S \in [0, S_d]$ at time t; $\Psi(S)$ – linear density of the material at point $S \in [0, S_d]$ at the initial time $t = 0$; $H(S)$, $\delta(S)$ Heaviside function and Dirac function, respectively; $a(t)$ – conveyor belt speed at time t; $\lambda(\mathrm{t})$ – input flow of material entering the transport system at point $S = 0$. For several input flows of material $\lambda_1(\mathrm{t}), \lambda_2(\mathrm{t}), \lambda_3(\mathrm{t})$, entering the conveyor at several points S_1, S_2, S_3 equation (1) takes the form:

$$\frac{\partial \chi_0(t,S)}{\partial t} + \frac{\partial \chi_1(t,S)}{\partial S} = \sum_{k=1}^{3} \delta(S - S_k)\lambda_k(\mathrm{t}). \quad (4)$$

Let us define similarity criteria for conveyors, the process of material transportation for which is determined by equation (1). To construct similarity criteria, we will use the π-theorem. The dimension of each of the defining parameters of the model P_n can be expressed through the basic units of measurement: time $[\mathrm{T}]$, length $[\mathrm{L}]$, mass $[\mathrm{M}]$:

$$[P_n] = [\mathrm{T}]^{\alpha_n}[\mathrm{L}]^{\beta_n}[\mathrm{T}]^{\mu_n}, \quad n = 1...N. \quad (5)$$

The transport system will be characterized by $(N-R)$ similarity criteria:

$$\pi_i = \prod_{n=1}^{N} [\mathrm{P_n}]^{z_{ni}}, \quad i = 1...(N-R). \quad (6)$$

where R is the number of basic units of change. For the transport conveyor model represented by equation (1), $R = 3$ respectively. The exponents of z_{ni} must be defined in such a way that the value of π_i is a dimensionless quantity:

$$\pi_i = \prod_{n=1}^{N} \left([\mathrm{T}]^{\alpha_n}[\mathrm{L}]^{\beta_n}[\mathrm{T}]^{\mu_n}\right)^{z_{ni}}, \quad i = 1...(N-R). \quad (7)$$

The values of the exponents of degree z_{ni} under this condition are determined from the solution of the system of equations:

$$\sum_{n=1}^{N} \alpha_n z_{ni} = 0, \quad \sum_{n=1}^{N} \beta_n z_{ni} = 0, \quad \sum_{n=1}^{N} \mu_n z_{ni} = 0. \quad (8)$$

The system of equations has $(N-R)$ independent solutions. Each solution is assigned a similarity criterion. In a specific study of special classes of material transportation phenomena using conveyors, the numerical values of quantitative characteristics P_n are usually expressed as a ratio to the specified or most characteristic quantities. The parameter based on the Dirac function differs from other variables, such as $\chi_0(t,S)$, $\chi_1(t,S)$, $\lambda(\mathrm{t})$, since it is not scaled directly, but only its argument $S_d\delta(S) = \delta(S/S_d)$ is scaled. This circumstance is the reason that the Dirac function will be singled out as a separate parameter when used in the π-theorem. The presence of this parameter in the equation allows us to construct similarity criteria based on the structure of local sources of material receipt on the conveyor. This is especially useful when constructing similarity criteria for a main conveyor with several sources of material receipt located along the transportation route.

IV. Similarity Criteria for a Model with Deterministic Input Flow of Material

The defining parameters P_n and their dimensionality for the model with a deterministic input material flow are presented in Table 1.

TABLE I. Degrees of Dimensionality of Parameters for a Model with a Deterministic Input Flow of Material

n	parameter, P_n	Basic units of measurement					
		time, [T]		length, [L]		mass, [M]	
1	χ_0	α_1	0	β_1	-1	μ_1	1
2	χ_1	α_2	-1	β_2	0	μ_2	1
3	λ	α_3	-1	β_3	0	μ_3	1
4	$\delta(S)$	α_4	0	β_4	-1	μ_4	0
5	t	α_5	1	β_5	0	μ_5	0
6	S	α_6	0	β_6	1	μ_6	0

Taking into account the values of the exponents α_n, β_n, μ_n the system of equations (8) for the deterministic model can be written in the following form:

$$\begin{cases} -z_{2i} - z_{3i} + z_{5i} = 0, \\ -z_{1i} - z_{4i} + z_{6i} = 0, \\ z_{1i} + z_{2i} + z_{3i} = 0, \end{cases} \quad i = 1,2,3. \quad (9)$$

where the number of defining parameters is $N = 6$, with the number of basic units of measurement being $R = 3$. The system of equations (9) has three degrees of freedom $(N-R)$. Three quantities z_{ni} can take arbitrary values, the other three are determined by the system of equations (9). The freedom in choosing these quantities does not affect the similarity criteria of the transport conveyor, which allows us to determine simple values for them. In this regard, we will define variables $z_{1i} = \omega_{1i}$, $z_{2i} = \omega_{2i}$, $z_{4i} = \omega_{3i}$, where $\omega_{1i}, \omega_{2i}, \omega_{3i}$ are arbitrary simple values. Having performed the transformation:

$$\begin{cases} z_{5i} = -z_{2i} - z_{3i} = -z_{1i}, \\ z_{6i} = z_{1i} + z_{4i}, \\ z_{3i} = -z_{1i} - z_{2i}, \end{cases} \quad i = 1,2,3. \quad (10)$$

we obtain a general solution to the system of equations (9):

$$\begin{cases} z_{1i} = \omega_{1i}, \\ z_{2i} = \omega_{2i}, \\ z_{3i} = -\omega_{1i} - \omega_{2i}, \\ z_{4i} = \omega_{3i}, \\ z_{5i} = -\omega_{1i}, \\ z_{6i} = \omega_{1i} + \omega_{3i}, \end{cases} \quad i = 1,2,3. \quad (11)$$

We substitute three different sets of $\omega_{1i},\omega_{2i},\omega_{3i} \in \{0,1\}$ and determine the values of z_{ni} . The solution of equation (9) and for different sets of $\omega_{1i},\omega_{2i},\omega_{3i}$ and the similarity criteria corresponding to these solutions for the deterministic conveyor model are presented in Table 2.

TABLE II. SIMILARITY CRITERIA FOR A MODEL WITH DETERMINISTIC INPUT FLOW OF MATERIAL

i	ω_{1i}	ω_{2i}	ω_{1i}	z_{1i}	z_{2i}	z_{3i}	z_{4i}	z_{5i}	z_{6i}	**Similarity criteria**
1	1	0	0	1	0	-1	0	-1	1	$\pi_1 = \frac{\chi_0 S}{t\lambda}$
2	0	1	0	0	1	-1	0	0	0	$\pi_2 = \frac{\chi_1}{\lambda}$
3	0	0	1	0	0	0	1	0	1	$\pi_3 = S_d\delta(S)$

Let the considered process of material transportation be characterized by the parameters: χ_0^* , χ_1^*, λ^* , t^*, S^* . Let us introduce dimensionless modelling parameters for the considered process of material transportation:

$$\tau = \frac{t}{t^*},\ \xi = \frac{S}{S^*},\ \Theta_0 = \frac{\chi_0}{\chi_0^*},\ \delta(\xi) = S^*\delta(S), \qquad (12)$$

$$g(\tau) = a(t)\frac{t^*}{S^*},\quad \Theta_1 = \frac{\chi_1}{\chi_1^*},\quad \gamma = \frac{\lambda}{\lambda^*}. \qquad (13)$$

Then equation (1) can be rewritten as:

$$\frac{\chi_0^*}{t^*}\frac{\partial\Theta_0(\tau,\xi)}{\partial\tau} + \frac{\chi_1^*}{S^*}\frac{\partial\Theta_1(\tau,\xi)}{\partial\xi} = \frac{\lambda^*}{S^*}\delta(\xi)\gamma(\tau). \qquad (14)$$

We divide both parts of the equation by the factor $\left(\lambda^*/S^*\right)$, and obtain equation (1) in dimensionless form:

$$\frac{S^*}{\lambda^*}\frac{\chi_0^*}{t^*}\frac{\partial\Theta_0(\tau,\xi)}{\partial\tau} + \frac{S^*}{\lambda^*}\frac{\chi_1^*}{S^*}\frac{\partial\Theta_1(\tau,\xi)}{\partial\xi} = \delta(\xi)\gamma(\tau). \qquad (15)$$

The dimensionless factors in equation (15) represent the similarity criteria (Table 2). Using the notation for the similarity criteria, we rewrite equation (15) as an equation containing the similarity criteria:

$$\pi_1\frac{\partial\Theta_0(\tau,\xi)}{\partial\tau} + \pi_2\frac{\partial\Theta_1(\tau,\xi)}{\partial\xi} = \pi_3\gamma(\tau). \qquad (16)$$

Let us supplement equation (16) with equations (2), (3) in dimensionless form:

$$\Theta_1(\tau,\xi) = \frac{\pi_1}{\pi_2}g(\tau)\Theta_0(\tau,\xi), \qquad (17)$$

$$\Theta_0(0,\xi) = H(\xi)\psi(\xi), \qquad (18)$$

a system of equations for describing the behaviour of the flow parameters of a transport system with a deterministic input flow of material is obtained. From the presented dimensionless system of equations, one can draw a very important conclusion: for different conveyor systems, for which the values of the similarity criteria π_1 , π_2 , π_3 are close in value, the behaviour of the dimensionless flow parameters will be described by the same equation.

V. SIMILARITY CRITERIA FOR A CONVEYOR MODEL WITH STOCHASTIC STATIONARY INPUT MATERIAL FLOW

Let the stochastic stationary input flow of material be determined by realization $\lambda(t)$, the values of which are obtained as a result of experimental measurements [20, 21, 22]. The input flow of material under consideration will be characterized by the mathematical expectation m_λ , the standard deviation σ_λ and the correlation time η . To analyse the input flow of material, we introduce the variable γ :

$$\lambda = m_\lambda + \gamma\sigma_\lambda,\quad \gamma = \frac{\lambda - m_\lambda}{\sigma_\lambda}. \qquad (19)$$

Taking into account the features of additional parameters, that characterize the input flow of material, the defining parameters P_n and their degrees of dimensionality for the model with a stochastic input flow of material are presented in Table 3.

TABLE III. DEGREES OF DIMENSIONALITY OF PARAMETERS FOR A CONVEYOR MODEL WITH STOCHASTIC INPUT MATERIAL FLOW

n	**parameter,** P_n	**Basic units of measurement**					
		time, [T]		**length, [L]**		**mass, [M]**	
1	χ_0	α_1	0	β_1	-1	μ_1	1
2	χ_1	α_2	-1	β_2	0	μ_2	1
3	m_λ	α_3	-1	β_3	0	μ_3	1
4	$\delta(S)$	α_4	0	β_4	-1	μ_4	0
5	t	α_5	1	β_5	0	μ_5	0
6	S	α_6	0	β_6	1	μ_6	0
7	σ_λ	α_7	-1	β_7	0	μ_7	1
8	η	α_8	1	β_8	0	μ_8	0

Let us take into account the values of the exponents α_n , β_n , μ_n and, by analogy with the model for a deterministic input flow of material, we will represent the system of equations (8) in the following form:

$$\begin{cases} -z_{2i} - z_{3i} + z_{5i} - z_{7i} + z_{8i} = 0, \\ -z_{1i} - z_{4i} + z_{6i} = 0, \\ z_{1i} + z_{2i} + z_{3i} + z_{7i} = 0, \end{cases} \quad i = 1,...,5. \qquad (20)$$

where the number of defining parameters is $N = 8$ with the number of basic units of measurement being $R = 3$. The resulting system of equations (20) has five degrees of freedom $(N - R)$. Five quantities z_{ni} can take arbitrary values, the

remaining three are determined by the system of equations (20). Having performed the transformation:

$$\begin{cases} z_{5i} = -z_{1i} - z_{8i}, \\ z_{6i} = z_{1i} + z_{4i}, \\ z_{3i} = -z_{1i} - z_{2i} - z_{7i}, \end{cases} \quad i = 1,...,5. \tag{21}$$

a general solution to the system of equations (20) is obtained:

$$\begin{cases} z_{1i} = \omega_{1i}, \\ z_{2i} = \omega_{2i}, \\ z_{3i} = -\omega_{1i} - \omega_{2i} - \omega_{4i}, \\ z_{4i} = \omega_{3i}, \\ z_{5i} = -\omega_{1i} - \omega_{5i}, \\ z_{6i} = \omega_{1i} + \omega_{3i}, \\ z_{7i} = \omega_{4i}, \\ z_{8i} = \omega_{5i}, \end{cases} \quad i = 1,...,5. \tag{22}$$

Let's select five different sets $\omega_{ki} \in \{-1,0,1\}$ $k = 1,...,5$ to determine the value of z_{ni}. The solution of equation (9) for different sets ω_{ki} and the corresponding similarity criteria for the conveyor model with a stochastic input material flow are presented in Table 4.

TABLE IV. SIMILARITY CRITERIA FOR A MODEL WITH STOCHASTIC INPUT MATERIAL FLOW

i	z_{1i}	z_{2i}	z_{3i}	z_{4i}	z_{5i}	z_{6i}	z_{7i}	z_{8i}	**Similarity criteria**
1	$\omega_{k1} = (1,0,0,0,0)$								$\pi_1 = \frac{\chi_0 S}{t m_\lambda}$
	1	0	-1	0	-1	1	0	0	
2	$\omega_{k2} = (0,1,0,0,0)$								$\pi_2 = \frac{\chi_1}{m_\lambda}$
	0	1	-1	0	0	0	0	0	
3	$\omega_{k3} = (0,0,1,0,0)$								$\pi_3 = S_d \delta(S)$
	0	0	0	1	0	1	0	0	
4	$\omega_{k4} = (0,0,0,1,0)$								$\pi_4 = \frac{\sigma_\lambda}{m_\lambda}$
	0	0	-1	0	0	0	1	0	
5	$\omega_{k5} = (0,0,0,0,-1)$								$\pi_5 = \frac{t}{\eta}$
	0	0	0	0	1	0	0	-1	

It is assumed that the studied process of material transportation can be characterized by the parameters: χ_0^*, χ_1^*, m_λ^*, σ_λ^*, η^*, t^*, S^*. Let us introduce dimensionless modelling parameters for the considered process of material transportation:

$$\tau = \frac{t}{t^*},\ \xi = \frac{S}{S^*},\ \Theta_0 = \frac{\chi_0}{\chi_0^*},\ \delta(\xi) = S^* \delta(S), \tag{23}$$

$$\Theta_1 = \frac{\chi_1}{\chi_1^*},\ \psi = \frac{\Psi}{\chi_0^*},\ g(\tau) = a(t)\frac{t^*}{S^*},\ \gamma = \frac{\lambda - m_\lambda^*}{\sigma_\lambda^*}, \tag{24}$$

Then equation (1) can be rewritten as:

$$\frac{\chi_0^*}{t^*}\frac{\partial \Theta_0(\tau,\xi)}{\partial \tau} + \frac{\chi_1^*}{S^*}\frac{\partial \Theta_1(\tau,\xi)}{\partial \xi} = \frac{\delta(\xi)}{S^*} m_\lambda \left(1 + \frac{\sigma_\lambda}{m_\lambda}\gamma(\tau)\right). \tag{25}$$

Let's divide both parts of the equation by a factor of $\left(m_\lambda / S^*\right)$, and obtain equation (1) in dimensionless form:

$$\frac{S^*}{m_\lambda}\frac{\chi_0^*}{t^*}\frac{\partial \Theta_0(\tau,\xi)}{\partial \tau} + \frac{S^*}{m_\lambda}\frac{\chi_1^*}{S^*}\frac{\partial \Theta_1(\tau,\xi)}{\partial \xi} = \delta(\xi)\left(1 + \frac{\sigma_\lambda}{m_\lambda}\gamma\left(\tau\frac{t^*}{\eta^*}\right)\right). \tag{26}$$

Taking into account the notations for the similarity criteria (Table 4), equation (26) can be represented in a form containing the similarity criteria:

$$\pi_1 \frac{\partial \Theta_0(\tau,\xi)}{\partial \tau} + \pi_2 \frac{\partial \Theta_1(\tau,\xi)}{\partial \xi} = \pi_3\left(1 + \pi_4 \gamma(\pi_5 \tau)\right). \tag{27}$$

Let us supplement equation (27) with equations (2), (3) in dimensionless form:

$$\Theta_1(\tau,\xi) = \frac{\pi_1}{\pi_2} g(\tau)\Theta_0(\tau,\xi), \tag{28}$$

$$\Theta_0(0,\xi) = H(\xi)\psi(\xi), \tag{29}$$

a system of equations for describing the behaviour of the flow parameters of a transport system with a stochastic input flow of material is obtained. Different conveyor systems for which the values of the similarity criteria π_1,…, π_5 have close values are similar systems. These conveyor-type transport systems are described by the same system of equations (27) – (29) and, under the same initial conditions, have the same solution.

VI. ANALYSIS OF THE RESULTS

When constructing similarity criteria characterizing the process of material transportation in a belt conveyor, the main factors determining the studied parameters of the system are identified. To construct a table of the main parameters of the system, a system of equations for modelling the process of material transportation in a transport conveyor (1) - (3) is used. It is possible to approach the formation of a table of the main factors (determinant parameters) describing the process of material transportation without using analytical equations of the model, having established those factors that are necessary for a complete determination of the characteristics of the process of material transportation, the numerical values of the parameters of which are measured as a result of the experiment. In this case, when compiling a system of determining parameters of the model, it is necessary to schematize the transportation process. In contrast to this, the approach based on the system of equations significantly simplified this process and made it possible to schematize, construct and substantiate equations for modelling the process of material transportation in dimensionless form. Among the determining parameters of the system, the main parameters of the system are highlighted. Through these main parameters of the system, the determining parameters are expressed

In the approach proposed in this article, the Dirac function $\delta(S)$ is considered in the qualitative analysis as a separate value in equation (1), because it plays a special role, models a point source of material inflow, and provides an idealized representation of a local source of material inflow into the conveyor. The similarity criterion obtained on its basis establishes a geometric relationship of the point for material inflow into the transport system. In the general case, the similarity criterion has the form $\pi_3 = S_d\delta(S - S_k)$, which is determined by equation (4). However, to simplify the method for constructing similarity criteria for a transport system, it is assumed that the material enters the conveyor input with point $S_0 = 0$, $\pi_3 = S_d\delta(S - S_0)$. If there are several points of material inflow, then to ensure geometric similarity, it is necessary to add similarity criteria for each local source of material inflow.

As was indicated earlier, each studied process of material transportation can be characterized by certain parameters, for example, for a model with a deterministic input flow of material $\chi_0^*, \chi_1^*, \lambda^*, t^*, S^*$. Having characteristic values of these quantities, the value for a specific similarity criterion can be calculated, which allows us to answer whether the transportation processes in two different conveyors are similar. Characteristic quantities are usually different for each studied transportation process. For example, for the same conveyor there may be cases of material transportation in which the conveyor belt loading differs significantly, for example $\chi_{01}^* >> \chi_{02}^*$. Then these two transportation processes will be described by equations (16) – (18) with different values of the parameter and, what is important, for comparing dimensionless values of densities it is necessary to introduce an additional scaling factor that determines the relationships between the characteristic values of densities χ_{01}^*, χ_{02}^*. In this regard, after solving the problem that requires an answer to whether two conveyors are similar, it is more convenient to move on to another set of quantities that determines the dimensionless parameters of the model when analysing the systems of equations (1) – (3).

Among such quantities it is necessary to specify: S_d – the length of the conveyor; $\chi_{0\max}$ – the maximum permissible linear density of the material at which there is no spillage of the material; T_d – the characteristic time of the process. The last quantity requires special attention. This can be the time interval during which the material on average passes the entire transportation path of length S_d (for the model with a deterministic input material flow) or the time interval equal to the correlation time η_d of the stochastic input process (for the model with a stochastic input material flow), at which the value of the correlation function decreases by $e \approx 2.718 = k(0)/k(\eta_d)$ times, where $k(\eta)$ is the correlation function of the stochastic process describing the arrival of the input material flow at the entrance of the transport conveyor. Replacing $\chi_0^* \to \chi_{0\max}$, $\chi_1^* \to \chi_{0\max}\dfrac{S_d}{T_d}$, $\lambda^* \to \chi_{0\max}\dfrac{S_d}{T_d}$, $t^* \to T_d$, $S^* \to S_d$ and using the new parameters for constructing dimensionless quantities:

$$\tau = \frac{t}{T_d},\ \xi = \frac{S}{S_d},\ \Theta_0 = \frac{\chi_0}{\chi_{0\max}},\ \delta(\xi) = S_d\delta(S), \quad (30)$$

$$g(\tau) = a(t)\frac{T_d}{S_d},\ \Theta_1 = \frac{\chi_1}{\chi_{0\max}\dfrac{S_d}{T_d}},\ \gamma = \frac{\lambda}{\chi_{0\max}\dfrac{S_d}{T_d}}. \quad (31)$$

let us rewrite the system of equations for the model with a deterministic input flow of material (1) – (3) in dimensionless form:

$$\frac{\partial\Theta_0(\tau,\xi)}{\partial\tau} + \frac{\partial\Theta_1(\tau,\xi)}{\partial\xi} = \delta(\xi)\gamma(\tau), \quad (32)$$

$$\Theta_1(\tau,\xi) = g(\tau)\Theta_0(\tau,\xi),\ \ \Theta_0(0,\xi) = H(\xi)\psi(\xi), \quad (33)$$

For the case of a constant belt speed $a(t) = a_d$ with a given choice of T_d as the time interval during which the material on average passes the entire transport path of length S_d, follows $a_d = S_d/T_d$, and accordingly the scale for the value $g(\tau) \sim 1$. It should also be noted that the ratio $\chi_0/\chi_{0\max} = (\chi_0 S_d)/M_{\max}$ can be interpreted as the material loading coefficient of the transport system, which is important for the analysis of the energy characteristics of the transport system. The value $M_{\max} = \chi_{0\max}S_d$ is the maximum possible mass of material that can be placed on the conveyor belt. It should also be noted that the ratios:

$$\frac{\chi_1}{\chi_{0\max}\dfrac{S_d}{T_d}} = \frac{\chi_1 T_d}{M_{\max}},\ \ \frac{\lambda}{\chi_{0\max}\dfrac{S_d}{T_d}} = \frac{\lambda T_d}{M_{\max}}, \quad (34)$$

They also have a scale characterized by the material loading coefficient of the transport conveyor. The new dimensionless parameters introduced allow us to conveniently analyse the behaviour of the conveyor flow parameters under different material transportation conditions. Let us perform a similar analysis for a model with a stochastic input material flow. When analysing the systems of equations (1) – (3), when the similarity criteria of the transport systems have already been determined, we move on to the following set of values that determine the dimensionless parameters of the model: S_d – conveyor length; $\chi_{0\max}$ – maximum permissible linear density of the material at which there is no material spillage; η_d – correlation time of the stochastic input process characterizing the input material flow. Having performed the following replacement of the characterizing parameters, $\chi_0^* \to \chi_{0\max}$, $\chi_1^* \to \chi_{0\max}\dfrac{S_d}{\eta_d}$, $m_\lambda^* \to \chi_{0\max}\dfrac{S_d}{\eta_d}$, $\sigma_\lambda^* \to \chi_{0\max}\dfrac{S_d}{\eta_d}$, $\eta^* \to \eta_d$ $t^* \to \eta_d$, $S^* \to S_d$ and using the new parameters to construct dimensionless quantities:

$$\tau = \frac{t}{\eta_d},\ \xi = \frac{S}{S_d},\ \Theta_0 = \frac{\chi_0}{\chi_{0\max}},\ \delta(\xi) = S_d\delta(S), \quad (35)$$

$$g(\tau) = a(t)\frac{\eta_d}{S_d}, \qquad \Theta_1 = \frac{\chi_1}{\chi_{0\max}\dfrac{S_d}{\eta_d}}, \tag{36}$$

$$m_\gamma = \frac{m_\lambda}{\chi_{0\max}\dfrac{S_d}{\eta_d}}, \ \sigma_\gamma = \frac{\sigma_\lambda}{\chi_{0\max}\dfrac{S_d}{\eta_d}}. \tag{37}$$

let us rewrite the system of equations for the model with a stochastic input flow of material (1) – (3) in dimensionless form:

$$\frac{\partial\Theta_0(\tau,\xi)}{\partial\tau}+\frac{\partial\Theta_1(\tau,\xi)}{\partial\xi} = \delta(\xi)m_\gamma\left(1+\frac{\sigma_\gamma}{m_\gamma}\gamma(\tau)\right). \tag{38}$$

$$\Theta_1(\tau,\xi) = g(\tau)\Theta_0(\tau,\xi), \quad \Theta_0(0,\xi) = H(\xi)\psi(\xi). \tag{39}$$

Just as in the model for a deterministic input flow of material for the case of a constant belt speed $a(t) = a_d$, we obtain $a_d = S_d / T_d = (S_d / \eta_d)k_g$, where k_g is the scaling factor for the dimensionless belt speed: $g(\tau) \sim k_g$. Just as before, parameters Θ_0, Θ_1, m_γ characterize the material loading factor of the transport conveyor.

VII. Conclusion

In this paper, we consider methods for modelling material transportation processes in conveyor systems using dimensionless parameters based on the principles of similarity theory. The construction and analysis of similarity criteria made it possible to identify the key physical parameters that have the greatest impact on the behaviour of the transport system under both deterministic and stochastic input material flows.

The use of the π-theorem made it possible to obtain a system of dimensionless equations that adequately describe the behaviour of the flow characteristics of the system under various operating modes. The criteria are determined under which transport systems are described by the same system of equations and, therefore, can be considered as similar systems. This opens up the possibility of transferring experimental data and modelling results from one system to another, which is especially important when designing new energy-saving control systems.

The proposed approach can be effectively used in planning and interpreting experimental studies, as well as in developing control algorithms for conveyor transport systems. Further research can be aimed at taking into account nonlinear effects, interactions between several flows, and integrating the obtained models into intelligent control systems.

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Robust Control of Electric Servo Drive

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***Abstract*—The paper deals with researching the possibilities of the robust control of electric servo drives assigned for the stabilization of the payload mounted on mobile objects. The mathematical description of the electric servo drive is represented. The features of the description for electric drives assigned to land mobile objects are explained. In the model, disturbance moments inherent in electric drives of land mobile objects are taken into consideration. The Simulink model of the electric motor has been developed. The theoretical grounds of the robust structural synthesis are given. The basic principles of robust control inherent in electric drives of land-mobile objects are described. The equations for determining the output signals of the studied system are represented. A robust structural synthesis has been implemented for the electric servo drive used on a land mobile object. The results of modeling the synthesized system are shown and analyzed. The transient processes in such modes as tracking and stabilization are given. The determination of the angular rigidity of the platform with payload by simulation is proposed. The obtained results can be useful for the synthesis of systems for controlling electric drives assigned for operation on mobile objects of a wide class.**

Keywords—electric servo drive, robust structural synthesis, mathematical description, modeling, discrete control unit

I. Introduction

Control of servo drives based on a DC motor is relevant for many industrial areas. Studying gyroscopic two-mass servo systems is particularly interesting for these drives. Such systems ensure stabilization and tracking of the payload, for example, a movie camera installed on a mobile land object, and a stabilization and guidance system for a radar antenna installed in rotating gimbals [1], [2].

The widespread approach to developing control systems is the application of robust control. The robust systems are insensitive to both the variation of system parameters and the influence of external disturbances [3]. This gives great possibilities for the operation of electric servo drives, which are used for control and stabilization systems assigned for mobile objects of different types (land, marine, and aerial).

The synthesis of robust systems can be based on minimizing the H_{inf} norm of the matrix transfer function of a closed-loop system. There is also a known approach to system synthesis based on minimizing the H_2 norm of the matrix transfer function of a closed-loop system. It is worth mentioning that, from the point of view of implementing computational algorithms, H_{inf} optimization is much more complicated than H_2 optimization due to the need to implement a search procedure. Synthesis methods based on minimizing the H_2 norm provide high accuracy of the synthesized system, but they remain sensitive to both external disturbances and parametric disturbances of the control object. The use of the H_{inf} norm allows ensuring the system's resistance to external disturbances under the condition of structured and unstructured parametric uncertainties [3], [4]. Optimization according to each of the considered approaches has definite advantages. The optimization according to a mixed criterion allows combining these advantages [5]. In such a situation, the synthesized system will be characterized by optimal quality in the presence of disturbances. The features of creating robust stabilization systems, using H_2, H_{inf}, and mixed H_2/H_{inf} approaches, and their comparative analysis are presented in [3], [6].

II. Analysis of Last Research

Methods for creating servo systems with a DC motor are described in sufficient detail in [7] – [9]. The issues of mathematical modeling of systems of the considered type are described in [5], [6], [10]. However, while many approaches to mathematical modeling of systems under consideration have remained relevant nowadays, computer modeling methods have changed significantly due to new computing tools. Therefore, it is important to study the possibilities of modeling servo drive devices, taking into account the development of modern computing technology. In addition, the development of a methodology for the transition from a continuous control system to a digital one is of great interest for the design of modern servo systems. The availability of such a methodology allows, on the one hand, to apply the useful results of previous developments, and on the other hand, to create servo systems that meet the modern requirements. Finally, it will be observed that lies in the relevance of creating models of electronic devices that allow choosing the discreteness of information updates, since the solution to this issue largely determines the ways of creating the system as a whole.

A large number of scientific publications are dedicated to the creation of robust control systems. The main principles that led to the development of the robustness theory are based on the Kharitonov theorem [3]. There are three main directions for the development of the robustness theory. The first approach is based on the concept of a multi-dimensional stability limit. Another approach is based on the idea of a structured singular value. The third approach is related to applying linear matrix inequalities [3], [11].

One of the central ideas underlying the robustness analysis method comes from understanding the Nyquist criterion, on which the theorem about the small gain coefficient is based [3]. Relevant problems in the synthesis of robust control systems are given in many publications, for example, [12]. Features of the design of robust stabilization systems, and their characteristics analysis are presented in [5], [6], [13].

Algorithms for parametric and parametric-structural synthesis of robust systems assigned for the stabilization of payload mounted on land mobile objects are represented in [14]. The development of appropriate procedures for structural synthesis of stabilization systems for land mobile objects remains an urgent problem.

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The creation of methods for synthesizing a stochastic system requires taking into account specific external disturbances acting on the system of the type under study. For land mobile systems, external disturbances are caused by road and terrain irregularities. To date, deterministic and statistical approaches are known that allow us to take into account disturbances acting on a land mobile object with payload In the first approach, the profile of road irregularities is described by a predetermined, most often harmonic function. In the statistical approach, disturbances caused by road irregularities are considered random variables. Taking into account random disturbances allows the creation of procedures for synthesizing a stochastic system of the class under study.

III. Description of Electric Servo Drive

The electric servo drives for a stabilization and tracking system assigned for operation on a land mobile object should provide stabilization and tracking of a payload in the vertical and horizontal planes. Electronic device models of the vertical and horizontal channels differ only in some components, as well as their parameters. Therefore, in this paper, the emphasis is placed on the mathematical description of the electric servo drive for the vertical channel [16], [17].

Following [16], the electric servo drive for the stabilization and tracking system assigned for land mobile objects includes series-connected units such as a motor, a reducer, and a voltage amplifier. The drive control signals are formed in the pulse-width modulator based on the signals coming from the control unit of the stabilization and tracking system. The control in the vertical plane is grounded on signals of the gyroscopic tachometer, the control panel, and feedback signals of the electric motor current and voltage. The functions of the electronic devices of the vertical channel of the tracking system include amplification, filtering, integration, and the formation of control laws. A special feature of the vertical channel model is the presence of limiting devices and switches, since in some situations, the type of electronic device model depends on the signal level. The control unit also performs the reduction to a single dimension of the current and voltage feedback signals. The functions of the electronic devices of the horizontal channel of the tracking system include integration, implementation of high- and low-frequency filters, band-pass filters, formation of control laws, and reduction to a single dimension of the voltage and current signals.

The output signal of the control unit enters the pulse-width-modulator model, in which a sequence of pulses of a given voltage. The width of pulses varies and depends on the control signal. This sequence is the input value for the electric motor model. In turn, the motor model includes a relationship for forming the motor armature current, an aperiodic link that takes into account the armature and the electromechanical time constants, and a model of the mechanical part of the motor. The motor armature current is used to form the drive control torque, as well as voltage and current feedback signals. The voltage on the motor circuit winding is determined in the following way

$$U_{win} = U_{pwm} - c_{emf}\dot{\varphi}_m, \tag{1}$$

where U_{win} is the voltage in windings of the motor armature, U_{pwm} is the voltage on the output of the pulse-width-modulator, c_{emf} is the electromotive force; φ_m is an angle of the turn of the motor.

The time constant of the motor armature circuit is taken into account through an aperiodic unit with a gain:

$$T_a pU_{cor} + U_{cor} = U_{win}, \tag{2}$$

where T_a is the motor armature circuit time constant, U_{cor} is the corrected output signal of the aperiodic unit. The signal U_{cor} enters the voltage feedback circuit, then through the unit $\frac{1}{R_{win}}$ into the current feedback circuit (J_{cor}) and then through the load torque constant on the motor shaft c_m to the mechanical part of the electric motor. Thus, the motor control torque is formed using the relationship $M_m = c_m J_{cor}$.

The mathematical model of the mechanical part of the motor is the following differential equation:

$$J_m\ddot{\varphi}_m + M_{res}\operatorname{sign}\dot{\varphi}_m = M_{con} - c_g\frac{\varphi_{hys} - \varphi_{pl}n_g}{n_g^2}, \tag{3}$$

where M_{con} is the motor control torque; M_{res} is the torque of resistance to rotation; is the gearbox rigidity; n_g is the gearbox ratio; J_m is the motor moment of inertia; φ_m is the motor rotor rotation angle; φ_{hys} is the motor rotor rotation angle taking into account the hysteresis-type nonlinearity caused by the motor backlash; φ_{pl} is the rotation angle of the payload.

The expressions (1) – (3) describe the mathematical models of basic units of the electric servo drive assigned for the stabilization and tracking of a payload mounted on the mobile object.

The features of the mathematical model of the electric servo drive with DC motor include the following factors:

- Presence of the hysteresis-type nonlinearity, for this, the rotation angle of the motor rotor (the output signal of the mathematical model of the engine) is passed through a nonlinear link.
- The relationship between the mechanical part of the motor and the working unit is implemented through the gearbox. Mathematically, this relationship is ensured by introducing the rotation angle of the working unit into equation (3). In this case, the rotor and the working unit represent a single device for reproducing the control signal (the so-called two-mass system).

To establish the adequacy of the model to real motor operation, zeroing of the motor rotation acceleration at the zero speed, and the fulfillment of certain relationships between the moments acting on the motor are introduced. These conditions are as follows:

$$\left|M_m - c_g\frac{\varphi_{hys} - \varphi_{pl}n_g}{n_g^2}\right| \le |M_{res}|; \quad \dot{\varphi}_m = 0. \tag{4}$$

The relationship between the payload and the motor can be described using the following equation:

$$J_{pl}\ddot{\varphi}_{pl} + M_{fr}\operatorname{sign}\dot{\varphi}_{fr} - M_{unb}\sin\varphi_{unb} + k_{spr}\varphi_{pl} + c_g\varphi_{pl} = k_{spr}A + c_g\varphi_{hys}/n_g - M_{unb}, \tag{5}$$

where M_{fr} is the friction moment; M_{unb} is the moment of unbalance; k_{spr} is the spring stiffness; A is the lifting angle of

the balancing spring; J_{pl} is the moment of inertia of the payload; φ_{pl} is the angle of rotation of the payload.

The expressions (4), and (5) represent the basic equation of the electric servo drive, taking into consideration conditions that close the model to real operation.

IV. Control of Electric Servo Drive

The control of the electric servo drive is based on the robust structural synthesis when both the structure and parameters are defined. The most appropriate is to use H_{inf} synthesis, that is, multi-criteria robust structural synthesis, which ensures system design under uncertainty [3[, [19], [20].

In the general form, the control system looks like

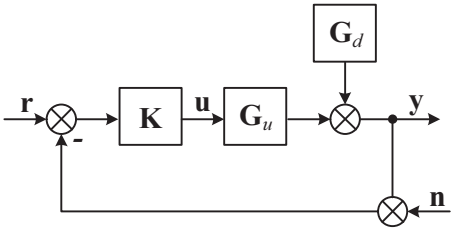


Fig. 1. The block diagram of the control system.

For this system, equations for determining output signals **y**, control **u** and error **e** can be determined using transfer functions of the control object $\mathbf{G}_u$, disturbance $\mathbf{G}_d$ and regulator **K**, and also input signals such as a reference signal **r**, disturbance **d**, and measurement error **n** [3], [13]

$$\begin{aligned} \mathrm{y} &= \mathrm{G}_u\mathrm{K}(\mathrm{I}+\mathrm{G}_u\mathrm{K})^{-1}\mathrm{r}+(\mathrm{I}+\mathrm{G}_u\mathrm{K})^{-1}\mathrm{G}_d\mathrm{d}- \\ &\mathrm{G}_u\mathrm{K}(\mathrm{I}+\mathrm{G}_u\mathrm{K})^{-1}\mathrm{n}; \\ \mathrm{u} &= \mathrm{K}(\mathrm{I}+\mathrm{G}_u\mathrm{K})^{-1}\mathrm{r}-\ \mathrm{K}(\mathrm{I}+\mathrm{G}_u\mathrm{K})^{-1}\mathrm{G}_d\mathrm{d}- \\ &\mathrm{K}(\mathrm{I}+\mathrm{G}_u\mathrm{K})^{-1}\mathrm{n}; \\ \mathrm{e} &= (\mathrm{I}+\mathrm{G}_u\mathrm{K})^{-1}\mathrm{r}-\ (\mathrm{I}+\mathrm{G}_u\mathrm{K})^{-1}\mathrm{d}-(\mathrm{I}+\mathrm{G}_u\mathrm{K})^{-1}\mathrm{n}\,. \end{aligned} \tag{6}$$

Using equations (6), it may be possible to estimate the system characteristics. Matrix transfer functions from the command signal **r** to the error, control, and output signals **e**, **u**, **y** are defined as follows

$$\begin{aligned} \mathrm{S}(j\omega) &\overset{\text{def}}{=} (\mathrm{I}+\mathrm{G}_u(j\omega)\mathrm{K}(j\omega))^{-1}; \\ \mathrm{R}(j\omega) &\overset{\text{def}}{=} \mathrm{K}(j\omega)(\mathrm{I}+\mathrm{G}_u(j\omega)\mathrm{K}(j\omega))^{-1}; \\ \mathrm{T}(j\omega) &\overset{\text{def}}{=} \mathrm{G}_u(j\omega)\mathrm{K}(j\omega)(\mathrm{I}+\mathrm{G}_u(j\omega)\mathrm{K}(j\omega))^{-1}. \end{aligned} \tag{7}$$

In expressions (7), the matrix transfer function from the reference signal **r** to the error signal **e** is called the sensitivity function $\mathbf{S}(j\omega)$. The matrix transfer function $\mathbf{T}(j\omega)$ from the control reference signal **r** to the output signal **y** is called the complementary sensitivity function. The matrix transfer function $\mathbf{R}(j\omega)$ from the reference signal **r** to the control signal **u** is called the control sensitivity function.

The simultaneous use of these three sensitivity functions allows you to achieve a compromise between accuracy, robustness, and control costs.

It will be noticed that all the above requirements are mutually contradictory and cannot be satisfied simultaneously. In this case, the search for the optimal regulator is carried out on a set of regulators that ensure the internal stability of the system. This approach is the most common in H_{inf} synthesis procedures. Such an approach is used in the mixed sensitivity method. The expression for the optimization criterion and, accordingly, the formulation of the optimization problem for the mixed sensitivity method are

$$\mathbf{K}_{opt} = \arg \inf_{\mathbf{K}_{opt}\in\mathbf{K}_{per}} J(\mathbf{G},\mathbf{K}), \tag{8}$$

$$J(\mathrm{G},\mathrm{K}) = \left\| \begin{bmatrix} (\mathrm{I}+\mathrm{GK})^{-1} \\ \mathrm{K}(\mathrm{I}+\mathrm{GK})^{-1} \\ \mathrm{GK}(\mathrm{I}+\mathrm{GK})^{-1} \end{bmatrix} \right\|_{\infty}. \tag{9}$$

To formulate the problem of H_{inf} synthesis (8), (9), it may be to use a formalized representation. The appropriate block diagram is represented in Fig. 2. [3]

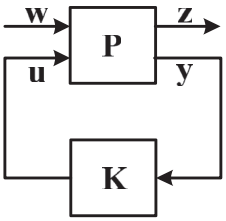


Fig. 2. The problem statement of H_{inf} synthesis.

The synthesized system is described as a control object (together with a measuring system and an actuator) and a regulator with matrix transfer functions $\mathbf{P}(s)$, $\mathbf{K}(s)$, which are fractional-rational and regular. The control object is a system with two inputs **w**, **u,** and outputs **z**, **y**. The input vector **w** in the generalized representation includes disturbances, measurement interference, and command signals. The input vector **u** describes control signals. The output vector ***z*** can be used for estimating the quality of stabilization and tracking processes; for example, it can represent the tracking error of the reference signal. The output vector **y** describes a vector of observed signals that can be used to organize feedback [21] – [23]. The control system shown in Fig. 2 can be described in the state space as follows:

$$\begin{aligned} \dot{\mathrm{x}}(t) &= \mathrm{A}\mathrm{x}(t)+\mathrm{B}_1\mathrm{w}(t)+\mathrm{B}_2\mathrm{u}(t); \\ \mathrm{z}(t) &= \mathrm{C}_1\mathrm{x}(t)+\mathrm{D}_{11}\mathrm{w}(t)+\mathrm{D}_{12}\mathrm{u}(t); \\ \mathrm{y}(t) &= \mathrm{C}_2\mathrm{x}(t)+\mathrm{D}_{21}\mathrm{w}(t)+\mathrm{D}_{22}\mathrm{u}(t); \\ \mathrm{u}(t) &= \mathrm{K}\mathrm{y}(t). \end{aligned} \tag{10}$$

The mathematical model (10) can be represented in the following form [3], [13]:

$$\begin{bmatrix} \dot{\mathrm{x}}(t) \\ \mathrm{z}(t) \\ \mathrm{y}(t) \end{bmatrix} = \begin{bmatrix} \mathrm{A} & \mathrm{B}_1 & \mathrm{B}_2 \\ \mathrm{C}_1 & \mathrm{D}_{11} & \mathrm{D}_{12} \\ \mathrm{C}_2 & \mathrm{D}_{21} & \mathrm{D}_{22} \end{bmatrix} \begin{bmatrix} \mathrm{x}(t) \\ \mathrm{w}(t) \\ \mathrm{u}(t) \end{bmatrix}. \tag{11}$$

If to write the transfer function of the control object as

$$\mathrm{P} = \begin{bmatrix} \mathrm{P}_{11} & \mathrm{P}_{12} \\ \mathrm{P}_{21} & \mathrm{P}_{22} \end{bmatrix}, \tag{12}$$

we can obtain the transfer function of the closed-loop system from input w to output z [3]:

$$T_w^z = P_{11} + P_{12}K(I - P_{22}K)^{-1}P_{21} \quad T_w^z = P_{11} + P_{12}K(I - P_{22}K)^{-1}P_{21} \tag{13}$$

H_{inf} synthesis allows us to determine the robust regulator K(s), which minimizes the norm $\| T_w^z \|_\infty$ of the closed-loop control system, where [3]

$$T_w^z = \begin{bmatrix} (I + GK)^{-1} \\ K(I + GK)^{-1} \\ GK(I + GK)^{-1} \end{bmatrix} = \begin{bmatrix} S \\ R \\ T \end{bmatrix}. \tag{14}$$

V. Computer Modelling of Electric Servo Drive

After realizing the optimization algorithm (11) – (14), the computer modelling of the synthesized system has been done. The model of the servo drive is realized in the Simulink Toolbox of the MATLAB system [24], [25]. It is based on transfer functions of the servo drive units.

The application of the Simulink model allows one to take into account nonlinear units and bring the model closer to the real hardware implementation.

As an example, the computational scheme of the motor is given in Fig. 3.

Other Simulink models are not given for the simplicity of the paper presentation. Simulation results are given in Figs. 4, 5.

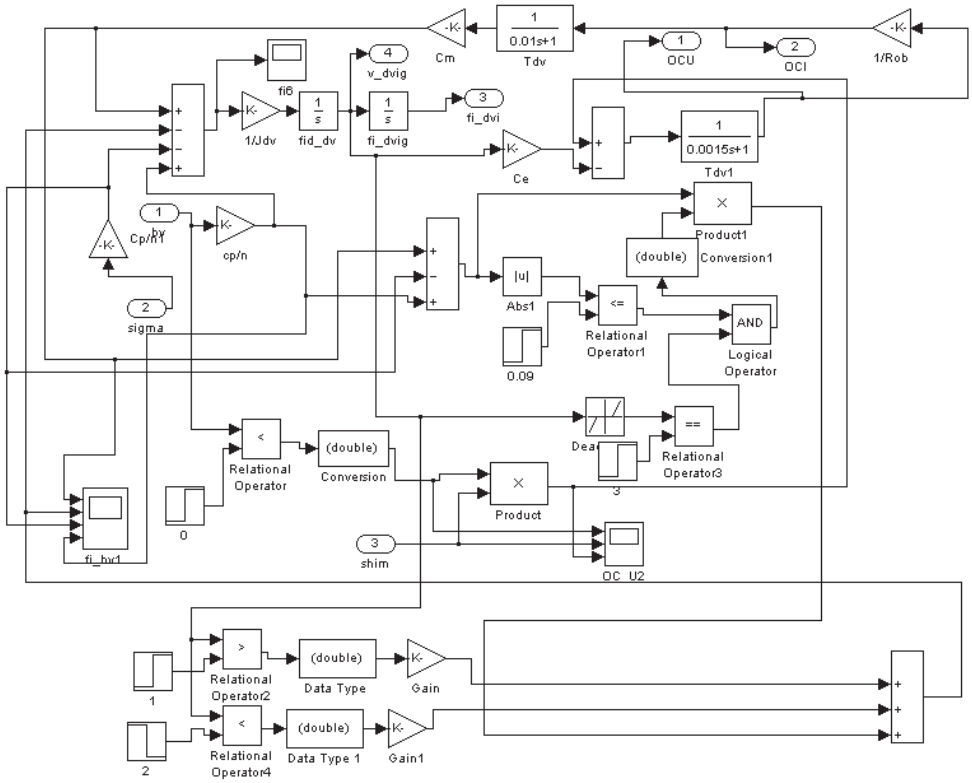
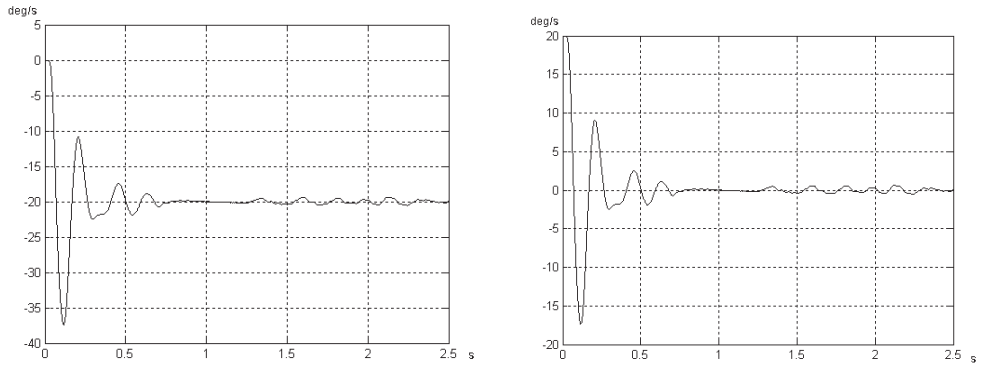
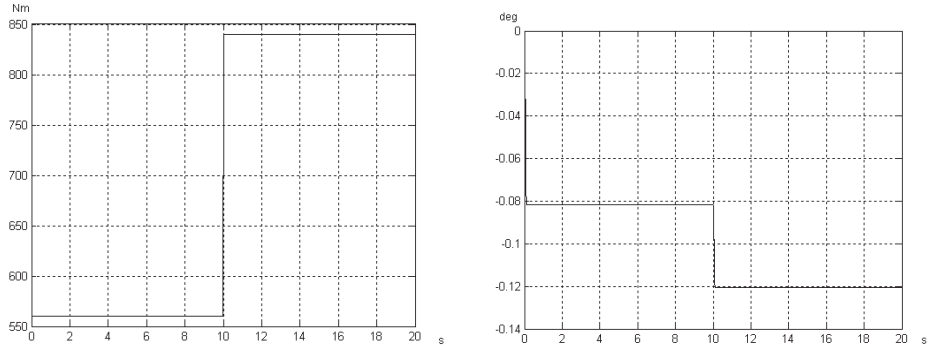


Fig. 3. Simulink model of the motor.



a *b*

Fig. 4. Transient processes on the angular rate: a – tracking the given angular rate; b – stabilization process.



c *d*

Fig. 5. The determination of the angular rigidity of the stabilized platform with a payload: a – tested disturbing moment; b – reaction of the angular position of the stabilized platform.

Figs 4a and 4b show the transient processes for the relative and absolute velocity for the vertical discrete channel. The imitation of the object body motion in the vertical plane was set using the transfer velocity. As follows from the simulation results, when the object moves at a speed of 20 deg/s, the deviation of the absolute velocity relative to the zero level does not exceed the permissible level.

Figs 5a and 5b confirm the required angular rigidity of the system. For this, the absolute angles of rotation of the working unit are analyzed after the action of forces of 560 Nm and 840 Nm, respectively.

The H_{inf} norm of the synthesized system is 0.1248. This value proves high robustness. At the same time, the accuracy of the system is sufficient as it is characterized by H_2 norm equal to 0.3653.

Analysis of logarithmic amplitude-frequency characteristics has shown that margins on the amplitude and phase are 49.5 dB and 91.1 degrees, respectively.

VI. Transition from Continuous Control Unit to Discrete One

Modern servo drives involve the use of computing tools. Therefore, developing a method for the transition from a continuous control unit to a digital one is of considerable interest. For this purpose, it is proposed to create a model of a continuous control channel using a transfer function apparatus, for example, using known analytical dependencies. This process is quite complex, and the correctness of the analytical representation should be verified by conducting appropriate experiments on a test bench.

In addition, it is advisable to use modern systems for modeling electronic devices, WorkBench or MultiSIM, which allow determining the logarithmic amplitude-frequency characteristics of separate transfer functions. These characteristics should be compared with similar characteristics of analytical transfer functions, for example, using the MATLAB system [26], [27].

Unfortunately, the WorkBench system does not have the ability to directly determine the analytical representation of the transfer function based on the model of an electronic device. It is worth mention that such a possibility exists in the MATLAB system, but for a subsystem designed for modeling power energy circuits, which is associated with certain difficulties in modeling electronic devices.

After representing a continuous control block in the form of a model using transfer functions, a transition to a digital block is carried out using Z-transformation of the corresponding transfer functions by program methods. To realize this approach, you can use the Control Toolbox of the MATLAB system, namely the c2d function.

The choice of the discretization method is determined by the required accuracy of the transformation, which is estimated by the degree of coincidence of the logarithmic amplitude-frequency characteristics of the transfer functions with similar characteristics of the continuous transfer functions.

It will be observed that the transfer function of the digital integrator is most sensitive to the discretization method. Therefore, it is advisable to use a sufficiently accurate bipolar transformation.

The advantages of the proposed approach to modeling electronic control devices for a servo drive include the possibility of studying the permissible value of the discreteness of information processing, the possibility of introducing tuning coefficients into the model, and the use of a pulse-width modulator model. When selecting the discreteness interval, the problem of ensuring a compromise between the desired accuracy and the convenience of hardware implementation of the control unit is solved [28], [29].

Tuning coefficients are provided to improve the quality of control processes based on the results of modeling and testing [17]. Such indicators as the duration of the transient process, the number of overruns, and the required rigidity of the system are taken into account.

The pulse-width modulator model can be described in detail using the Simulink system. In addition, the presented model provides the ability to select the discreteness of the analog-digital conversion, since it allows analyzing the frequency characteristics of the output signals of separate units.

Specific problems related to astatism for a discrete system are solved by finding the minimum implementation, similarly to the continuous system. As for the balanced implementation, it can be realized for continuous systems. In the case of a discrete system, it is advisable to perform scaling of the model

matrices in the state space using a similarity transformation with a diagonal matrix and a scalar coefficient [30].

VII. Conclusions

The mathematical model of the servo drive is represented, including the description of the motor armature circuit, the mechanical part of the motor, and the condition of the mathematical modelling proximity to real operation. Modeling of the synthesized system was implemented by Simulink Toolbox.

The features of the robust structural synthesis based on H_{inf} norm are described. The optimization problem and objective function based on the method of mixed sensitivity are represented.

The robust structural synthesis of the system with the electric servo drive assigned for stabilization and tracking a payload mounted on a land mobile object has been discussed. The results of the robust structural synthesis, including transient processes on the angular rate, checking angular rigidity, and calculation of H_{inf} and H_2 norms are represented. The obtained results prove the high robustness and accuracy of the synthesized system.

The represented procedure of the robust structural synthesis can be applied to mobile objects of another type.

The approach to the transformation of the continuous control unit into a discrete one is described. The proposed approach is based on modern specialized software assigned for researching complex engineering systems of a different kind.

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Spatial-Analytical Modeling of Renewable Energy Potential in Ukraine Considering the Geographical Distribution of Electricity Consumers

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***Abstract*— The article presents a comprehensive spatial-analytical study of the potential for the development of renewable energy sources (RES) across the territory of Ukraine, taking into account the geographical distribution of end electricity consumers. The primary focus is on identifying priority regions for the deployment of renewable energy infrastructure, where high resource and energy potential coincides with optimal energy supply logistics.The methodological framework of the study is based on the integration of geoinformation analysis, cartographic modeling, and techno-economic energy assessment. The spatial data analysis has enabled not only the localization of promising renewable energy accumulation zones, but also an evaluation of the impact of consumer geographic remoteness on energy transmission losses and overall system efficiency. The modeling results open new horizons for strategic energy planning, highlighting the necessity of aligning the placement of RES facilities with the spatial structure of energy demand. A series of practical scenarios are proposed, aimed at minimizing transmission losses while simultaneously increasing national-level energy efficiency. The findings carry applied significance for government regulators, institutional investors, and expert communities involved in the transformation of Ukraine's energy landscape toward sustainable development.**

Keywords*— *geoinformation analysis, renewable energy potential, spatial modeling, electricity consumers distribution.

I. INTRODUCTION

In the modern era of global challenges and civilizational shifts, the issue of energy transformation ceases to be just a technical or environmental dilemma. It acquires a strategic, even existential dimension. In a country experiencing tectonic socio-economic transformations, aggressive external threats and deep internal restructuring, energy is becoming the sphere where the vectors of national security, technological breakthrough and geopolitical subjectivity intersect. Ukraine today faces a unique historical opportunity - to create a new model of the energy ecosystem, which will be based not on the inertia of the past, but on an accurate, scientifically substantiated idea of space, resources and needs of society. The essence of this scientific work is to form an innovative paradigm of a spatial-analytical approach to assessing and modeling the potential of renewable energy within the territory of Ukraine. We are not limited to a quantitative analysis of opportunities - how much sun, wind or biomass is in a particular area - we offer a deep synergy between physical geography, spatial logic of infrastructure placement and real energy needs of the population and industry. At the heart of our approach is the idea of precisely matching the generation potential with the location of consumers, which allows not only to use resources more efficiently, but also to reduce losses, optimize distribution and increase the overall stability of the energy system.

In this context, space ceases to be just a physical background - it turns into a data carrier, a key indicator and a strategic factor in energy planning. Our modeling is based on the use of high-precision geo-information analytics, satellite monitoring, mathematical optimization algorithms and multi-level scenario approaches. At the heart is the idea that every square kilometer of Ukrainian land contains a unique energy formula, and the task of science is to reveal, digitize and make this formula manageable.

This work is a contribution to the formation of a new type of national thinking - energy-spatial, flexible, long-term thinking. We are not just mapping renewable energy - we are creating a new mental landscape in which decisions are made not at random, but on the basis of precise spatial-energy relationships. It is this approach that gives Ukraine a chance not only for energy independence, but also for Ukraine to enter the vanguard of the global green transition. Spatial-analytical thinking is the intellectual key to a new energy civilization, which we are starting to build right now.

II. REVIEW OF SCIENTIFIC SOURCES

The issue of renewable energy has long attracted the attention of scientists, engineers, and policymakers, as it represents not only a technological challenge but also a strategic necessity for humanity's sustainable future. Research in this field is vast and diverse, encompassing technical, economic, and ecological perspectives. When analyzing the state of scientific knowledge, it becomes clear

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that renewable energy is not studied as an isolated discipline but rather as an integral part of global energy transformation, in which solar, wind, hydro, biomass, and geothermal energy gradually replace fossil fuel–based generation. Scholars consistently emphasize that this process is not linear; instead, it requires the synthesis of advanced technologies, infrastructure adaptation, and active regulation from governments.

One of the central themes in the literature is the steady advancement of solar photovoltaic (PV) systems. Numerous studies demonstrate that the cost of PV modules has dramatically decreased over the past two decades due to technological breakthroughs, economies of scale, and improvements in manufacturing processes [1]. At the same time, researchers highlight the challenges of intermittency, efficiency fluctuations under changing weather conditions, and the urgent need for energy storage solutions. The development of high-capacity lithium-ion and emerging solid-state batteries is frequently presented as a decisive factor for the large-scale integration of solar power into national grids [2].

Equally significant attention is given to wind energy, which is considered one of the most promising renewable sources due to its scalability and relatively mature technology. Scholars analyze both onshore and offshore wind farms, noting that offshore installations are capable of providing higher and more stable energy outputs but require substantial capital investments and advanced grid connections [3]. Technical reviews often point to innovations in turbine design, the use of composite materials, and digital monitoring systems that extend the operational life of wind turbines and reduce maintenance costs [4].

Beyond the technical dimension, researchers stress the socio-economic and political aspects of renewable energy deployment. For instance, the transition to clean energy is frequently described as a catalyst for regional development, job creation, and innovation stimulation. However, literature also identifies barriers such as high initial capital costs, policy uncertainty, and social resistance in certain communities [5]. Particularly in developing countries, scholars underline the necessity of stable regulatory frameworks and international cooperation to attract investments and ensure long-term sustainability [6].

Another important body of work focuses on hybrid renewable energy systems, which combine different sources such as solar and wind with storage technologies to ensure reliability and efficiency. According to recent studies, such hybrid models are becoming increasingly popular in both urban and rural contexts, as they mitigate the weaknesses of individual sources and provide continuous power supply [7]. For instance, microgrid systems in remote regions have shown remarkable potential in providing decentralized energy access, reducing dependence on fossil fuels, and improving resilience against climate-related risks [8].

A critical line of inquiry concerns environmental and ecological implications. While renewable energy is often portrayed as environmentally friendly, researchers note that large-scale implementation must account for land use, resource extraction for batteries, and potential impacts on ecosystems. Wind farms may influence bird migration patterns, while solar installations require significant land resources if not integrated into rooftops or existing infrastructures. These issues underscore the importance of life-cycle assessments and holistic approaches to planning renewable energy projects [9-13].

III. OUTLINE OF THE MAIN MATERIAL

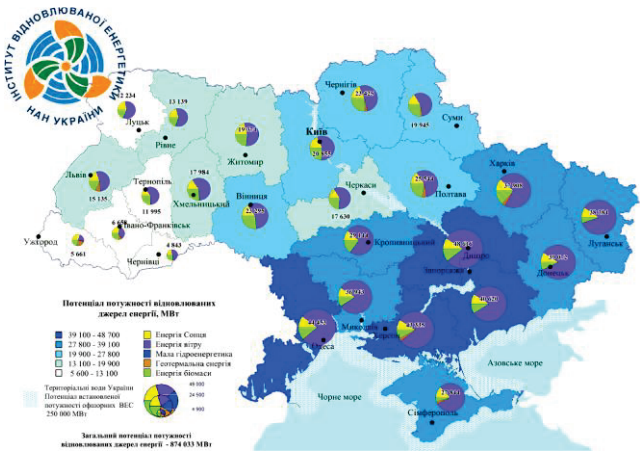
The methodology underpinning this study is based on a combination of geospatial analysis, computational modeling, and techno-economic assessment to evaluate the renewable energy potential of Ukraine while taking into account the spatial distribution of electricity consumption. The first crucial step in this methodology involved the compilation and verification of a comprehensive dataset representing both the natural energy resources available and the geographical patterns of electricity demand. Primary data sources included official records from the Ministry of Energy of Ukraine, which provided authoritative figures on national and regional energy generation, historical consumption patterns, and planned infrastructure projects. These were complemented by high-resolution data from meteorological stations across Ukraine, allowing for accurate calibration of solar insolation and wind speed measurements over different seasons and terrains. In addition, global datasets from Copernicus and NASA POWER were integrated to fill gaps and provide spatially continuous estimates of solar irradiance, wind potential, and other relevant meteorological variables. The combination of national and international sources enabled the creation of a reliable and high-fidelity dataset suitable for spatially explicit modeling.

Once the data were gathered, computational tools were employed to process, analyze, and simulate the energy potential across regional grids. MATLAB served as the primary tool for calculating average solar insolation and wind speed across predefined spatial cells, allowing for a detailed mapping of renewable energy availability throughout Ukraine. The modeling in MATLAB considered seasonal variability, topography, and land cover, thereby ensuring that the calculated resource potential reflected real-world constraints. For the optimization of energy transmission and the estimation of losses associated with distance from generation sites to consumption centers, MathCAD was utilized. This software facilitated scenario modeling that included distances between generation regions and major consumption hubs, enabling the simulation of transmission efficiency under varying spatial arrangements of renewable facilities [14-16]

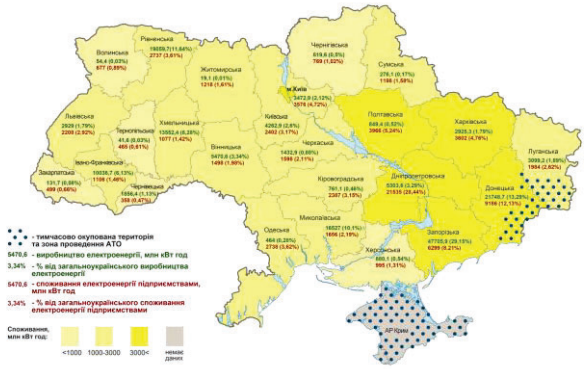
Furthermore, visualization of results required careful post-processing of maps and geospatial data. All figures were exported in high resolution from MATLAB and ArcGIS, with clearly labeled coordinate axes based on the WGS-84 system, annotated legends, and consistent color gradations representing solar potential, wind speeds, and transmission loss percentages. Units of measurement were standardized across all maps, including percentages for energy potential and kilometers for distances, ensuring that the visual outputs were both accurate and interpretable for policy makers, energy planners, and other stakeholders.

Among the visualized spatial data reflecting the potential for electricity production using RES and the configuration of consumption within the national energy system, the materials presented in Fig. 1 became a key source for further analysis.

In Fig 1a and 1b, volatility was not taken into account by us; it was considered by the researchers who created these maps.



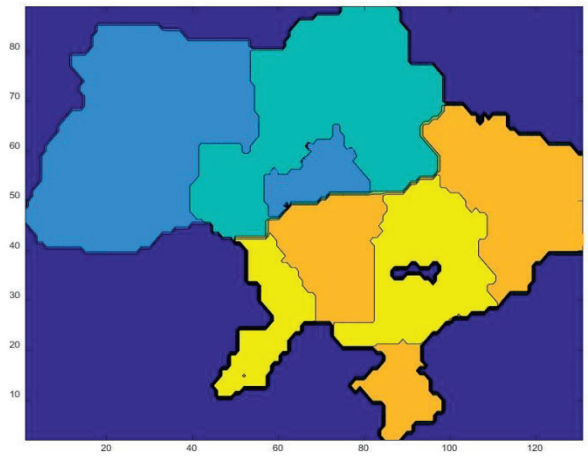
a



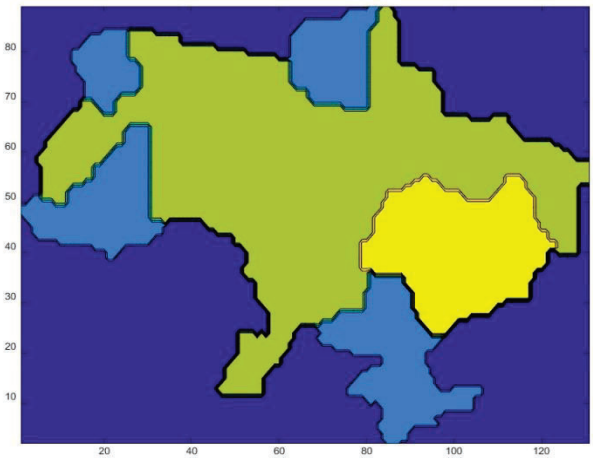
b

Fig.1 Distribution maps of RES generation (a) and electricity consumption (b) in Ukraine [1–2]

In the process of working with cartographic materials, it was found that most of them do not contain information about the projection system used, which creates difficulties for accurate spatial analysis. In order to ensure correct comparison and spatial unification, all maps were scaled to a single scale based on the maximum lengths of the territory of Ukraine: 1316 km from west to east and 893 km from north to south. To increase the accuracy of modeling, the studied area was discretized in the form of a regular grid of squares with a fixed side length. The resulting images were converted into a 4-bit gradation scale, which made it possible, depending on the original map, to divide the territory into 6 or 4 categories by potential level. This approach allowed creating spatial data arrays that represent the distribution of potential renewable electricity generation and actual energy consumption within Ukraine. These results are visualized in Fig. 2. As can be seen from the presented graphics, regions with the highest potential for renewable energy generation often do not spatially coincide with areas of intensive energy consumption. Such a disparity in the location of generating capacities and consumption centers will inevitably lead to significant losses of electricity in the process of its transportation.



a



b

Fig.2 Distribution maps of renewable energy generation (a) and electricity consumption (b) in Ukraine after recognition and data array creation (values and coordinates correspond to Fig. 1)

Obtaining the distribution of renewable energy generation and electricity consumption in Ukraine allows us to establish possible energy losses during transmission. To determine the share of energy transmitted from one region of Ukraine to another $WT_{j,i}^{k,n}$, the following dependence is proposed

$$WT_{j,i}^{k,n} = W_{j,i}^{R} \frac{W_{k,n}^{C}}{\sum P^{c}} \quad (1)$$

where $W_{j,i}^{R}$ - renewable energy generation potential of the region with coordinates j,i; $W_{k,n}^{C}$ - consumption capacity of the region with coordinates k,n; $\sum P^{c}$ - total capacity of the country's power system.

The above dependence is based on the assumption of a direct proportional dependence of the share of consumption of the country's regions on the total capacity of the country's power system:

$$\sum_{k=1;n=1}^{k=89;n=131} \frac{W_{k,n}^{C}}{\sum P^{c}} = 1 \quad (2)$$

The scheme for calculating losses during transmission of energy from the generation region j,i to the consumption region k,n is shown in Fig.3

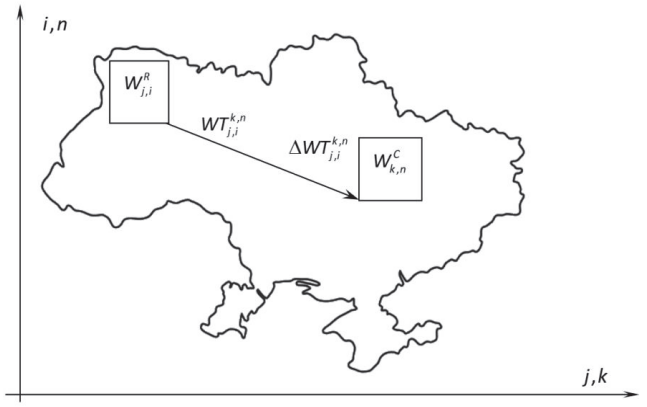


Fig.3 Scheme for calculating the losses of RES potential when transferring energy from the generation area to the consumption area

According to the above methodological recommendations, the loss of active $\Delta W^{(P)}$ or reactive electricity $\Delta W^{(Q)}$ in electrical networks for the calculation period are determined separately for each element of the electrical network (power transmission line, reactor, transformer or autotransformer, etc.) based on the flows of active or reactive electricity at the metering points located on the connected network element in accordance with the generalized formulas:

$$\Delta W^{(P)} = a \cdot I^2 \cdot R \cdot k_f^2 \cdot T^p + P_{y.n} \cdot T_н \quad (3)$$

a - coefficient depending on the type of network (three-phase, single-phase); I - average effective value of the current in the network element; R - active resistance of the network element; k_f^2 - average effective value of the current in the network element; $P_{y.n}$ - conditionally constant losses of active energy in the network element, which depend on the voltage and do not depend on the current strength;

$Q_{y.n}$ - conditionally constant reactive energy losses in a network element, which depend on voltage and do not depend on current strength;

$$\Delta W^{(Q)} = a \cdot I^2 \cdot X \cdot k_f^2 \cdot T^p + Q_{y.n} \cdot T_н \quad (4)$$

X - reactance of the network element;

The use of the dependence is significantly limited by the need to determine the reactive and active resistances of network elements, which depend on a number of structural elements of the power line. At the same time, the use of economic current density $i\ A/mm^2$ allows you to determine power losses under the conditions of choosing the average voltage and the material of the power line conductors

$$\Delta WT = I^2 R = \left(\frac{P}{U}\right)^2 \rho \frac{LUi}{P} = \rho \frac{P}{U} Li \quad (5)$$

where ρ - resistivity; P - power ($WT^{k,n}_{j,i}$);

L - transportation distance.

For the selected stage of development of the algorithm for determining losses, the shortest path for transmitting electrical energy is proposed.

$$L = \Delta l \sqrt{(i-n)^2 + (j-k)^2} \quad (6)$$

Δl - map breakdown step.

The algorithm for determining losses will take into account the density and direction of Ukraine's power supply networks by recognizing power line maps. The results of power losses for individual regions are shown in Fig. 3

Further development of the loss determination algorithm will take into account the density and direction of Ukraine's power supply networks by recognizing power line maps. The results of power losses for individual regions are shown in Fig. 4

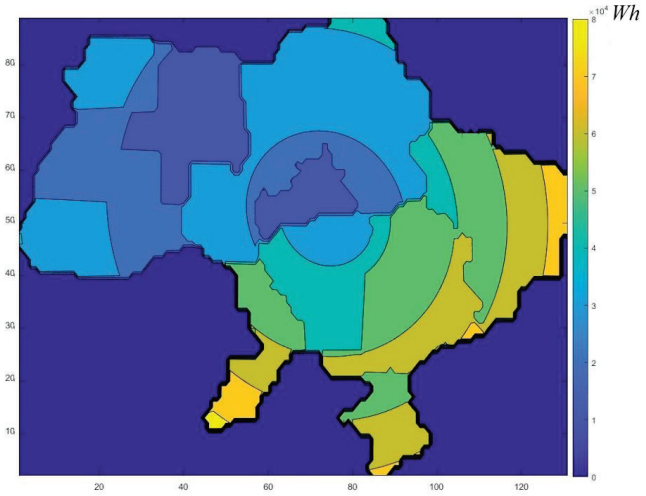


Fig.4 Maps of the distribution of losses predicted during the transmission of renewable electricity, taking into account the remoteness of consumers

As can be seen from the constructed map, the highest losses occur in Kherson, Donetsk, and Luhansk regions, where the losses amount to $7-8 \times 10^{10} Wh$. The lowest costs will be in the central part of Ukraine and in the Zaporizhia region $1-3 \times 10^{10} Wh$.

Analysis of the distribution map of absolute losses of electricity during transmission from renewable sources, taking into account the remoteness of consumers, shows that the location of generators and their remoteness significantly affect the level of losses. The lowest total losses are characteristic of the central regions of Ukraine, while the highest are expected in the eastern and southern regions. The most promising for the development of renewable energy generation are areas with an average level of losses, close to large centers of consumption, in particular the Dnipropetrovsk region and the northern part of the Odessa region [17–18].

To assess the optimal locations for RES, a mathematical model was created in the MATHCAD program, which takes into account the natural potential, power and distance to consumers. The model allowed us to assess the potential of various energy sources and their combinations to select the most effective options for regional development.

Optimization calculations in the MATLAB environment helped to determine the most profitable locations for power plants, taking into account economic aspects, such as infrastructure costs and consumer accessibility [19-24].

IV. CONCLUSION

The analysis of the spatial distribution of transmission losses reveals a strong dependence on the geographic position of Ukrainian regions relative to major electricity consumption centers. The lowest losses are observed in the central regions, including Kyiv and the surrounding areas, where they remain

around 5% due to high consumer density and the compact structure of the grid. A similar situation is found in Zaporizhzhia, where transmission losses are approximately 5–7%. This region also demonstrates exceptionally high solar potential, exceeding 80%, making it one of the most promising areas for further development of renewable generation. Southern regions, particularly Kherson and Mykolaiv, also benefit from high solar radiation levels (over 80%), but their distance from large industrial load centers leads to higher transmission losses. Modeling shows that these losses could be reduced by about 12% if generation facilities were located closer to demand hubs. Western regions occupy an intermediate position: here solar potential is moderate, and transmission losses are higher than in the central areas, though significantly lower than in the eastern regions. The most challenging situation is found in the eastern and northeastern regions, such as Donetsk and Luhansk, where transmission losses exceed 20% due to long distances and grid limitations. At the same time, northern and coastal regions display favorable wind potential, which opens opportunities for combining solar and wind generation strategies

.Overall, the findings highlight that the most effective way to optimize the Ukrainian energy system is to establish decentralized clusters of renewable generation in close proximity to major consumption centers. This approach can reduce transmission losses by 10–15%, improve system reliability, lower infrastructure costs, and simultaneously support environmental sustainability and regional development.

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Hardware Security Evaluation of IoT Microcontrollers: Threats and Countermeasures

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***Abstract*— This study analyzes the susceptibility of IoT microcontrollers to reverse engineering, defined as unauthorized data access for replication, and proposes countermeasures. STM32F0 demonstrates superior resistance due to dense integration, rapid protection activation, and robust encryption, effectively countering attacks like Cold Boot Stepping, Glitch, Electromagnetic Analysis, and invasive methods. GD32 offers moderate security but is vulnerable to faster attacks. CH32 and MM32F0 show the least resilience due to weak protection. Architectural analysis highlights STM32F0's secure interfaces, while GD32's multi-chip design and CH32/MM32F0's simpler cores increase risks. Security metrics confirm STM32F0's lead, with GD32, CH32, and MM32F0 needing enhancements. Recommendations include hardware shielding, signal filters, and dynamic encryption, implementable in 2–14 days at $0.1–1.0 per device. STM32F0 suits high-security IoT, GD32 mid-level needs, and CH32/MM32F0 low-risk applications. Comprehensive hardware and software modifications are critical to mitigate reverse engineering threats across all microcontrollers.**

Keywords— IoT Security, microcontrollers, hardware attacks, protection mechanisms, tncryption.

I. Introduction

The Internet of Things (IoT) has radically changed the approach to device development, providing connectivity in areas such as smart homes, industrial systems, medical devices, and transportation logistics. These devices rely on microcontrollers as the basis of their functionality, providing data processing, peripheral control, and communication with external systems. The STM32 microcontroller series has gained popularity due to its high performance, reliability, and a developed ecosystem that includes development tools, libraries, and technical support [3, 4]. In particular, the STM32F0 family, built on the ARM Cortex-M0 core, is widely used in commercial IoT products due to its optimal price-performance ratio.

However, the growing demand for IoT devices and competition in the market are forcing manufacturers to look for ways to reduce costs. One solution is to switch from the original STM32 to cheaper analogues, such as the GD32, CH32 from WCH and MM32. Consumers expect affordable devices without losing functionality, which forces manufacturers to find compromises between cost and quality. The transition of manufacturers to cheap analogues of microcontrollers carries significant security risks for IoT devices that process confidential data, such as personal user information or industrial parameters. Analogs of STM32, although compatible in terms of pins and software, may have weaker protection mechanisms, which increases their vulnerability to hardware attacks. The multi-chip architecture of the GD32 makes it easier to access flash memory compared to the single-chip design of the STM32, creating additional risks [7, 8, 11]. Reengineering poses a serious threat, causing financial losses to developers [2]. This study focuses on the STM32F0 family, a cost-effective solution for entry-level IoT products. The resistance of these microcontrollers and their counterparts (GD32, CH32, MM32) to four types of hardware attacks is analyzed: Cold Boot Stepping (CBS), Glitch attacks, Electromagnetic Analysis (EMA), and Invasive attacks such as chip decapsulation. CBS exploits the peculiarities of SRAM operation during cold boot to leak data, Glitch attacks manipulate voltage or frequency to bypass protection, EMA analyzes electromagnetic emissions to obtain confidential information, and Invasive attacks involve physical access to the chip [5, 6]. The purpose of the study is to assess the impact of microcontroller choice on the security of IoT devices and to offer recommendations for developers on how to balance cost savings and protection..

II. Main Research

The architecture of a microcontroller is a key factor determining its performance, power consumption, and resistance to hardware attacks aimed at reengineering IoT devices. This study compares microcontrollers:

- STM32F0 by STMicroelectronics;
- GD32 by GigaDevice;
- CH32 by WCH;

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– MM32 by MindMotion, which are popular in budget IoT applications. These microcontrollers share common features, such as 32-bit architecture and support for peripheral interfaces, but differ significantly in design, security mechanisms, and vulnerability to attacks. The comparison focuses on the single-chip architecture of the STM32F0, the multi-chip structure of the GD32, and the simplified designs of the CH32 and MM32, which affect their security [1, 3, 7, 8,11]. The STM32F0 microcontroller is built on the ARM Cortex-M0 core, which provides a balance between performance and low power consumption. Its single-chip architecture integrates the processor, flash memory, SRAM, and peripheral modules into a single chip, which reduces the physical access points for attacks. The STM32F0 has a flash memory capacity of up to 64 KB, for example, in the STM32F051R8T6 model, and a SRAM capacity of up to 8 KB, which is sufficient for typical IoT tasks such as sensor data processing [3]. Memory protection is implemented through the Readout Protection (RDP) mechanism, which at the RDP2 level blocks access to the flash memory and the debug interface, making reengineering difficult [4]. The single-chip design increases resistance to attacks such as decapsulation, since access to internal components requires high-precision equipment [7].

Unlike the STM32F0, the GD32, specifically the GD32F103, uses a multi-chip architecture where the processor logic and flash memory are placed on separate chips connected by wires via the QSPI protocol [1]. This design reduces manufacturing costs, but creates vulnerabilities, as the connecting wires are potential points for data interception. The GD32 is also based on the ARM Cortex-M3 core, which provides higher performance (up to 108 MHz) compared to the Cortex-M0 in the STM32F0 (up to 48 MHz). The GD32F103 has a flash memory of up to 128 KB and a SRAM of 20 KB, making it attractive for more complex IoT applications [1]. However, the Security Protection mechanism, similar to RDP1, is less robust, allowing access to the SRAM via the debugger, which facilitates reengineering [8].

The CH32 microcontroller, specifically the CH32V103, uses the RISC-V core (QingKe V3), which is an alternative to the ARM architecture. Its architecture is similar to the STM32F0 in terms of single-chip design, but simplified to reduce cost. The CH32 offers up to 64 KB of flash and 20 KB of SRAM, with a maximum frequency of 80 MHz [7]. The lack of hardware encryption and limited protection similar to RDP1 make the CH32 vulnerable to attacks aimed at leaking data for reengineering. MindMotion's MM32F103, built on the ARM Cortex-M0 core, has the simplest architecture among the microcontrollers compared, with up to 64 KB of flash and 8 KB of SRAM, and a frequency of up to 72 MHz [8]. Its protection is limited to basic mechanisms that do not include an RDP2 counterpart, and the weak security coverage facilitates invasive attacks. A comparison of the microcontroller specifications is given in Table 1. which summarizes the key parameters that affect performance and security, including core type, frequency, memory size, architecture, and security level.

The architectural differences of microcontrollers have a direct impact on their resistance to hardware attacks. The single-chip architecture of the STM32F0 and CH32 reduces the number of access points for attacks such as signal interception, but the CH32 loses out due to weaker protection [7]. The multi-chip structure of the GD32 simplifies attacks, in particular through access to QSPI lines that transmit unencrypted data [8]. The MM32, despite its single-chip design, is the most vulnerable due to the lack of robust protection mechanisms and weak chip coverage. These features indicate trade-offs between cost and security: the STM32F0 is the most expensive but the most secure, while the GD32, CH32 and MM32 offer savings at the expense of increased reengineering risk.

TABLE I. TECHNICAL CHARACTERISTICS OF MICROCONTROLLERS

Microcontroller	Core	Frequency (MHz)	Flash memory (KB)	SRAM (KB)	Architecture	Protection
STM32F0	ARM Cortex-M0	48	64	8	Single crystal	RDP1, RDP2
GD32	ARM Cortex-M3	108	128	20	Multi-chip	Security Protection
CH32	RISC-V QingKe V3	80	64	20	Single crystal	Basic (RDP1)
MM32	ARM Cortex-M0	72	64	8	Single crystal	Basic

The comparison of architectures highlights that the choice of microcontroller for IoT devices should be based not only on performance but also on security requirements. The STM32F0 is optimal for applications with high security standards, while the GD32 can be a compromise solution with additional security measures such as isolation of connections. The CH32 and MM32 are suitable for non-critical applications where the risk of reengineering is low [10]. Further analysis in the following sections details the vulnerabilities of each microcontroller to specific attacks.

Microcontroller security mechanisms are the foundation of IoT device security, preventing unauthorized access to firmware and sensitive data, which is critical for protection against reengineering. The STM32F0 microcontroller uses a two-level Readout Protection (RDP) mechanism. RDP1 restricts access to flash memory and SRAM via the debugger using 128-bit encryption of the configuration registers and 8 protected registers. RDP2 completely blocks the debug interfaces (JTAG/SWD), preventing memory reads, and is activated in less than 1 ms after initialization [3]. Testing has shown that RDP2 is resistant to most attacks, except those using side channels [4]. This protection requires careful use, as RDP2 is irreversible and blocks firmware updates.

The GD32 implements a Security Protection mechanism that restricts memory access through 64-bit encryption and 4 protected registers. However, it does not completely block JTAG/SWD, and the activation time is 10 ms, which creates a vulnerability window during initialization [1]. Testing on the GD32F103 has shown that Security Protection does not protect against manipulation of QSPI signals, which facilitates data leakage [8]. This mechanism is a compromise between cost and security, but is less effective in high-security applications.

The CH32 offers basic security, which uses a 32-bit key to restrict memory access via the debugger and only 2 protected registers. There is no encryption, and the protection activation time is 15 ms, making it vulnerable to attacks in the early stages of operation [7]. The MM32 has even weaker protection, without encryption, with a single configuration register and an activation delay of more than 20 ms. Tests have shown that the MM32F103 does not withstand even simple manipulations with the debugger, making it unsuitable for protection against reengineering [8]. Table 2 lists the technical parameters of the protection mechanisms, including numerical indicators reflecting the security level of each microcontroller.

It should be noted that the success rate of attacks is considered as the percentage of successful attempts to bypass protection in tests (CBS, Glitch, EMA) [3, 7, 8,11]. Based on the data presented in Table 1, it can be concluded that the STM32F0, due to its 128-bit encryption, 8 protected registers and fast activation, has the lowest success rate of attacks (10%) [3].

TABLE II. TECHNICAL PARAMETERS OF MICROCONTROLLER PROTECTION MECHANISMS

Microcontroller	Protection levels (number)	Encryption (bits)	Protected registers (number)	Activation time (ms)	Attack success rate (%)
STM32F0	2	128	8	<1	10
GD32	1	64	4	10	60
CH32	1	32	2	15	80
MM32	1	0	1	>20	95

It should be noted that the success rate of attacks is considered as the percentage of successful attempts to bypass protection in tests (CBS, Glitch, EMA) [3, 7, 8]. Based on the data presented in Table 1, it can be concluded that the STM32F0, due to its 128-bit encryption, 8 protected registers and fast activation, has the lowest success rate of attacks (10%) [3]. The GD32, with 64-bit encryption and slower activation, demonstrates average vulnerability (60%) [8]. The CH32 and MM32, with minimal security parameters, have high success rates of attacks (80–95%), which limits their use in critical IoT applications [7].

However, it should be noted that the most significant risks to the security of IoT devices are created by hardware attacks. The paper considers four attack methods [3, 5–9]:

1. Cold Boot Stepping (CBS) uses the residual data in SRAM after power is removed by manipulating the reset signal through a debugger to read keys or firmware. The data retention time is described by the following formula:

$$T_{\text{read}} = T_{reset} + \Delta T_{step} \quad (1)$$

where T_{read} – total time available to read data from SRAM (in milliseconds).

T_{reset} – reset signal delay that keeps the microcontroller in reset state (in milliseconds).

ΔT_{step} – the duration of the step during which the reset signal is briefly removed to read the memory (in microseconds).

To fine-tune the CBS attack (timing) to read the SRAM (microcontroller clock) at the time of initialization, it is necessary to determine the reset signal delay using formula 2.

$$T_{delay} = \frac{1}{f_{clk}} \times k \times V_{dd} \quad (2)$$

where, T_{delay} – reset signal delay (in microseconds).

f_{clk} – microcontroller clock frequency (in MHz).

k – proportionality coefficient, which depends on the hardware implementation (dimensionless quantity).

V_{dd} – microcontroller supply voltage (in volts)

Tests on the STM32F051R8T6 (100 kHz step rate) showed data access in 2–5 ms with weak protection, but blocking in <1 ms with strong protection [5]. The GD32F103 retains data for up to 10 ms due to slow SRAM clearing, which makes it easier to attack with an ANSI32 debugger [8]. The CH32 and MM32 allow reads in 5–15 ms due to slow initialization speed [7].

2. Glitch attacks disrupt the operation of a microcontroller by changing the voltage or clock signal, bypassing security checks. The amplitude of a glitch is described by formula 3 (Aglitch), which takes into account the voltage change and the decay time.

$$A_{glitch} = V_{nom} - \frac{\Delta V}{\tau_{fall}} \quad (3)$$

where, A_{glitch} – glitch amplitude (in volts).

V_{nom} – Rated supply voltage of the microcontroller (in volts).

ΔV – Voltage change during glitch (in volts).

τf_{all} – Voltage fall time (in nanoseconds).

In glitch attacks conducted on STM32F0 using ChipWhisperer-Husky, (dropping to 1.8 V, ΔV=1.5 V, 500 ns) the weak protection was bypassed, but the stronger one required more precise tuning [6]. GD32F103 was attacked at ΔV=1.3 V due to unprotected lines [8]. CH32 and MM32 are vulnerable at ΔV=1.0 V due to lack of stabilization [7]. Countermeasures: voltage stabilizers, randomization of operations [10].

3. Electromagnetic Emissions Analysis (EMA) analyzes electromagnetic signals to obtain data using correlation according to formula 4

$$R = \frac{Cov(\text{E},\text{D})}{\sigma_E \times \sigma_D} \quad (4)$$

where, R – Correlation coefficient (dimensionless value, from -1 to 1).

Cov(E,D): Covariance between measured radiation and data B.

σ_E – Standard deviation of radiation.

σ_D – Standard deviation of data.

4. Invasive attacks and decapsulation involve physical access, including decapsulation, as shown in Fig. 1 and0 Fig. 2 [8].

For the STM32F0, laser decapsulation took 6–12 hours due to dense integration, with memory locking under strong protection [7]. The GD32F103 was chemically attacked in 3–8 hours due to accessible wiring [6]. The CH32 and MM32 were acid-treated in 2–6 hours due to weak

coating [7]. Countermeasures: robust coatings, protected configurations [10].

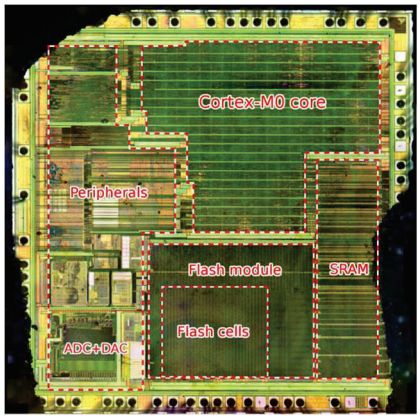


Figure 1: Annotated die of the STM32F051R8T6

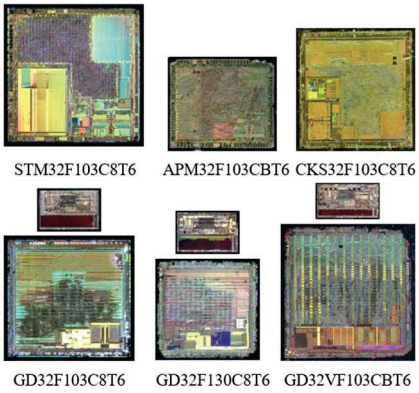


Figure 2 Decapsulated STM32F103 and analogs

In EMA tests on STM32F0 during AES execution, a correlation of R≈0.7 in the range of 100–200 MHz indicated key bit leakage [9]. In GD32F103, the multi-chip architecture increases R due to stronger QSPI line signals, facilitating reengineering. The formula is used for statistical data analysis in CEMA and DEMA methods. Table 3 summarizes the technical parameters of attacks on microcontrollers, including complexity, execution time, hardware, and vulnerability.

TABLE III. TECHNICAL PARAMETERS OF ATTACKS ON MICROCONTROLLERS

Microcontroller	Attack	Difficulty (1–5)	Execution time (hours)	Vulnerability (0–5)
STM32F0	CBS	4	1–2	2
	Glitch	3	2–4	3
	EMA	5	0.5–6	4
	Invasive	4	6–12	4
GD32	CBS	2	0.5–1	4
	Glitch	2	1–3	4
	EMA	3	0.2–4	5
	Invasive	2	3–8	4
CH32	CBS	2	0.5–1	4
	Glitch	2	1–2	4
	EMA	3	0.2–3	5
	Invasive	3	4–10	3
MM32	CBS	1	0.2–0.5	5
	Glitch	1	0.5–1	5
	EMA	2	0.1–2	5
	Invasive	1	2–6	5

The complexity and vulnerability indicators in Table 3 are calculated based on the technical characteristics of attacks and the results of testing conducted under controlled conditions [5–8]. The methodology provides an objective comparison of microcontrollers. Complexity is defined as a combination of three factors: technical requirements (quantity and complexity of equipment, setup time (duration of attack preparation), operator skills (level of expertise). Vulnerability is assessed by two criteria: probability of success (percentage of successful attacks in tests) and data volume (amount of data obtained).

Thus, the initial stage of protecting IoT devices from reengineering is a quantitative assessment of the resistance of microcontrollers to hardware attacks, which is based on an integrative security assessment through formalized indicators and practical implementation measures [3, 5, 7–10].

The analysis of the resistance of microcontrollers to reengineering is based on the security metric (S), which takes into account three technical characteristics: 1) the probability of a successful attack; 2) resource consumption for the attack; 3) attack execution time Table 4. The security metric is calculated by formula 5:

$$S = \frac{1}{P_{attack} \times C_{attack} \times T_{attack}} \quad (5)$$

where: P_{attack} — average probability of success of attacks (CBS, Glitch, EMA, invasive) in percentage (0–1),
C_{attack} — normalized attack complexity (Table 3),
T_{attack} — average attack time in hours.

Additionally, it is proposed to estimate the energy efficiency of protection (Eprot), which takes into account the energy consumption for security activation and the activation time is calculated as:

$$E_{prot} = \frac{W_{prot}}{T_{act}} \quad (6)$$

where W_{prot} — energy consumption (mW),
T_{act} — activation time (ms) [3].
A higher S-parameter indicates a greater resistance of the microcontroller to reengineering [3, 5]

Despite having the highest S index, MM32 is the least stable, as the high value is due to low C_{attack} and T_{attack}, which facilitates attacks. STM32F0 has the lowest S, but this reflects the high complexity and time of attacks, indicating better protection [5]. STM32F0 is also the most efficient in activation speed, which reduces vulnerability to attacks in the early stages.

The data given in Table 5 reflect the technical characteristics of the STM32F0, GD32, CH32 and MM32 microcontrollers in terms of parameters that affect resistance to reengineering. The parameters include: hardware resistance to attacks (the percentage of attacks (CBS, Glitch, EMA, invasive) that did not achieve full access to data in tests [3, 7, 8])software security flexibility (the value of security configuration options, taking into account encryption, randomization [5, 10]), security power consumption and

security implementation costs (the estimated cost of hardware and software modifications per device [7, 8])

TABLE IV. INDICATORS OF MICROCONTROLLER RESISTANCE TO REENGINEERING

Microcontroller	S	P_{attack}	C_{attack}	T_{attack}.	W_{prot} (mW)	T_{act} (ms)	E_{prot} (mW/ms)
STM32F0	0,143	0,35	4	5	5	1	5
GD32	0,185	0,8	2,5	3	8	10	0,8
CH32	0,188	0,85	2,5	2,5	10	15	0,67
MM32F0	0,561	0,95	0,25	1,5	12	20	0,6

According to Table 5, it can be concluded that STM32F0 has the highest hardware resilience, since 90% of attacks (in particular CBS, Glitch) did not achieve full access due to fast protection activation and interface blocking [3].

Software flexibility (5) reflects support for AES-256, randomization and integrity checking [5].

Low power consumption (5 mW) ensures real-time efficiency [3]. Costs ($0.5–1.0) are higher due to the need for shielding and stabilizers [10]. GD32 shows average resilience (60%) due to vulnerability to EMA and CBS, with partial access in 40% of tests [8].

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TABLE V. TECHNICAL COMPARATIVE ANALYSIS OF MICROCONTROLLERS

Microcontroller	Memory Read Resistance (%)	Protection activation speed (ms)	Attack costs ($/hour)	Hardware complexity of protection (0–5)	Hardware stability (%)	Program flexibility (0–5)	Security costs ($/device)
STM32F0	95	1	100–200	5	90	5	0.5–1.0
GD32	70	10	50–100	3	60	3	0.3–0.7
CH32	50	15	20–50	2	40	2	0.2–0.4
MM32F0	30	20	10–30	1	20	1	0.2–0.4

Software flexibility (3) is limited by 64-bit encryption and basic randomization [6]. Power consumption (8 mW) and costs ($0.3–0.7) are a compromise [8].

CH32 and MM32 have low resistance (40% and 20%) due to 80–95% successful attacks [7]. Software flexibility (2 and 1) is minimal due to weak encryption [7]. High power consumption (10–12 mW) and low cost (0.2–0.4 $) reflect their budget-friendly nature, but limit security [7].

Thus, based on the S and E_{prot} metrics, technical recommendations have been developed to increase the resistance of microcontrollers to reengineering. The recommendations are specific, focused on hardware and software modifications, with clear implementation parameters, namely:

STM32F0: Enable 128-bit encryption of configuration registers with <1 ms latency. Use of 8 security registers to lock JTAG/SWD (CR register = 0xFF) [3]. Implement AES-256 software encryption with 256-bit key to protect firmware with update rate of 1 time/10 minutes (TIM2 register, 600 s) [10]. Electromagnetic shielding (0.5 mm thick, copper foil, resistance <0.1 Ohm) to reduce EMA correlation to R<0.3 [5]. Use of voltage stabilizer (LM317, output 3.3 V ±0.1 V) to protect against glitch attacks with ΔV>1.5 V [6].

GD32: Activation of Security Protection with 64-bit encryption (SPR register = 0xAA). Reduction of activation time to 5 ms by firmware optimization (reduction of initialization cycles by 2 times) [8]. Implementation of software randomization of operations with a delay of 10–50 μs (RNG register, 32-bit generator) to complicate EMA [10]. Isolation of the signal line with 100 Ohm resistors (SMD 0805) to reduce data leakage by 30% [6].

Addition of an external cryptographic module (ATECC608B, I2C, 256-bit key) with consumption <1 mW for SRAM encryption [8].

CH32: Setting a 32-bit security key (CFG register = 0x55). Reducing Tact to 10 ms by removing redundant checks in the firmware (cycle optimization by 50%) [7]. Implementing SHA-256 software encryption with a 32-bit hash for code integrity verification (CRC register, 1 ms/operation) [10]. Adding an epoxy coating (1 mm thick, 10 MPa strength) to increase the time of invasive attacks to 8 hours [7]. Using a voltage stabilizer (AMS1117, 3.3 V ±0.05 V) to block glitch attacks with ΔV>1.0 V [6].

MM32: Activation of basic protection with 1 configuration register (SEC register = 0x11). Reduction of Tact to 15 ms by reducing initialization by 25% [7]. Implementation of AES-128 software encryption with 128-bit key (DMA register, 2 ms/block) for partial SRAM protection [10]. Addition of protective coating (polyurethane, thickness 0.8 mm, strength 8 MPa) to complicate chemical decapsulation by 50% [7]. Use of voltage stabilizer (MCP1700, 3.3 V ±0.1 V) to protect against glitch attacks with ΔV>0.8 V [6].

III. CONCLUSION

The analysis of STM32F0, GD32, CH32, MM32F0 microcontrollers showed that their resistance to reengineering, i.e. unauthorized access to data for reproduction, depends on architectural features, protection mechanisms and vulnerability to hardware attacks. STM32F0 is distinguished by the highest resistance due to dense integration, fast activation of protection and support for complex mechanisms such as multi-level encryption, which significantly complicates attacks such as CBS, Glitch, EMA and invasive methods. GD32 provides an average level of security, but is vulnerable to fast attacks due to slower protection initialization. CH32 and MM32F0 are the least resistant due to weak basic mechanisms and low software flexibility, which facilitates access by attackers.

Practical recommendations for improving security include the implementation of hardware shielding, signal line filters, dynamic encryption and data integrity checks. For STM32F0, these measures are maximally effective, while GD32 requires additional modifications, such as initialization optimization. CH32 and MM32F0 require significant efforts to compensate for vulnerabilities, which limits their use in critical IoT

applications. Implementation of the measures takes from several days to two weeks, with costs of 0.1–1.0 $ / device, depending on the microcontroller.

The STM32F0 is the optimal choice for high-security IoT devices, the GD32 is suitable for medium-security, and the CH32 and MM32F0 are for low-risk, low-cost applications. A comprehensive approach combining hardware and software modifications is necessary for all microcontrollers to minimize reengineering threats and protect intellectual property.

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A Spatial Clustering Filter for Industrial Noise in Satellite Fire Data

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***Abstract*— Satellite fire-detection products like VIIRS effectively identify wildfires but also falsely detect industrial heat sources (steel mills, cement kilns, gas flares) as fires. Such anomalies distort fire patterns, especially in industrial regions. Traditional filtering methods (e.g., removing "static" labeled points) often miss persistent industrial hotspots. We propose a spatial density-based approach using DBSCAN clustering (ε = 138 m, minPts = 20), parameters selected via k-distance analysis. Applied to VIIRS data for Ukraine (2019–2024), the filter captured over 96% of industrial anomalies without affecting real wildfires. This DBSCAN-based method is simple, computationally efficient, and integrates easily into standard GIS workflows, significantly improving wildfire data quality for regions with industrial noise.**

Keywords—VIIRS Active Fire, Thermal Anomaly Noise, DBSCAN Clustering, Spatial filtering, GIS

I. Introduction

Satellite-based fire products, such as VIIRS, detect not only wildfires but also stationary heat emissions from industrial activities (e.g., gas flares, steel mills, cement kilns). Due to VIIRS's geolocation uncertainty (~75 m), these industrial sources create dense "clouds" of repeated hotspot detections around actual facility locations [1]. Such anomalies artificially inflate fire counts and distort spatial analyses, complicating accurate wildfire assessments, especially in industrialized regions.

To manage these industrial false positives, VIIRS data includes a TYPE attribute to classify hotspot sources: vegetation fires (Type 0), volcanoes (Type 1), static land-based sources (Type 2), and offshore sources (Type 3) [2]. Typically, Type 2 hotspots correspond to known industrial emitters, allowing straightforward removal. However, misclassifications frequently occur, with many industrial hotspots incorrectly labeled as vegetation fires (Type 0) or offshore sources (Type 3), leaving residual noise [3]. Consequently, relying only on TYPE attributes is insufficient.

Alternative filtering strategies include manually defined exclusion zones around known industrial sites. For instance, a manually created industrial mask in Indonesia eliminated approximately 13% of fire detections, verified via imagery [4]. Yet, manual methods require significant local expertise, limiting their general applicability.

Spatial clustering approaches, such as hierarchical clustering and DBSCAN, have also been used effectively. For example, hierarchical clustering successfully identified gas flare sites linked to reported emissions in Texas [5], but often depended on predefined facility coordinates. Studies applying DBSCAN clustering generally lacked systematic parameter selection: a previous study in East Asia applied DBSCAN with ε ≈ 0.8 km to separate wildfires from industrial sources [6], while a global Nightfire survey adopted trial-and-error tuning [7]. A spatio-temporal DBSCAN variant (ST-MASK) incorporating temporal dimensions showed promise but still lacked transparent parameter-selection methods [8].

To address these gaps, we propose a reproducible, two-stage DBSCAN filtering workflow. First, a detailed industrial-zone mask was constructed using OpenStreetMap, cadastral records, and high-resolution imagery. Second, DBSCAN was tested systematically across multiple minPts (5, 10, 15, 20), with optimal ε values determined from k-distance curves and validated against the reference mask. The resulting parameter set reliably identifies and removes persistent industrial clusters across different temporal windows and geographic regions, providing an efficient, robust, and readily implementable filtering solution for GIS workflows.

II. Research Methodology

A. Input Data and Preprocessing

The analysis uses VIIRS active-fire detections (375 m resolution) from NASA FIRMS over Ukraine (2019–2024), including attributes such as fire radiative power (FRP, MW) and detection confidence (low, nominal and high) [9–11]. All data were reprojected to EPSG:3035 for accurate distance-based analyses. Only nominal and high-confidence detections were retained for clustering. No preliminary TYPE-based filtering was performed, as industrial anomalies often exhibit inconsistent TYPE classification.

Industrial-zone polygons were obtained from OpenStreetMap (features tagged landuse=industrial, man_made=works) [12,13] and validated using Copernicus Sentinel-2 imagery [14] and cadastral records. To formally account for VIIRS geolocation uncertainty (~75 m), some industrial polygons were buffered by 200 m. This buffered mask provides a reference map of industrial locations, used to calibrate clustering parameters and accurately identify industrial fire anomalies.

B. Density-Driven Clustering with DBSCAN

DBSCAN (Density-Based Spatial Clustering of Applications with Noise) is a clustering algorithm that identifies clusters of arbitrary shape without requiring a predefined number of clusters. It labels isolated points explicitly as noise. DBSCAN relies on two parameters:

- ε (epsilon): defines the radius for local neighborhood searches.
- minPts: sets the minimum number of points required to form a dense cluster.

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The clustering outcome is highly sensitive to the chosen parameters: if ε is too small genuine clusters fragment into noise, whereas a too large ε merges distinct clusters or includes spurious points [15, 16].

Since the choice of DBSCAN parameters significantly impacts clustering quality and is highly task-specific, we propose a systematic calibration method tailored to our application. The details of our parameter selection approach are described in the following section.

III. Results and Discussion

A. The Attribute-Based Filtering TYPE Flags

Many studies have proposed filtering industrial thermal anomalies by using the VIIRS TYPE attribute, which indicates the algorithm's inferred source of the thermal anomaly. The TYPE classification scheme labels detections as presumed vegetation fires (TYPE = 0), active volcanoes (TYPE = 1), static land-based sources (TYPE = 2), or offshore sources (TYPE = 3).

To assess the effectiveness of the TYPE filter, we examined the overlap between VIIRS detections (2019–2024) and our validated industrial-zone mask. After excluding all non-TYPE = 0 points, 22 378 detections—about 5% of the 414 003 total VIIRS detections—still fall within industrial polygons. This represents approximately 31% of the original 71 565 industrial detections, confirming that numerous factory- and kiln-related thermal anomalies are misclassified as vegetation fires (TYPE = 0) rather than being correctly labeled as static sources (TYPE = 2).

Thus, while the TYPE attribute provides a useful first-pass filter for identifying known static sources, it fails to reliably exclude all industrial thermal anomalies.

B. Selecting DBSCAN Parameters for Industrial Thermal Anomaly Filtering

Since simple attribute-based filtering or FRP thresholding of VIIRS thermal anomaly points proved insufficient to eliminate the dense jitter clouds over industrial sites we applied a density-based spatial clustering filter using the DBSCAN algorithm.

To capture the full density structure of industrial anomalies, we built k-distance plots from all high-confidence detections within the buffered industrial-zone mask. These knees provide a data-driven basis for selecting ε intervals prior to final tuning [16].

Fig. 1 illustrates the k-distance curves derived from industrial fire points for several values of minPts (5, 10, 15, and 20). As expected, the estimated ε decreases with lower minPts, reflecting the fact that smaller clusters require less spatial cohesion to meet the density threshold. Conversely, higher minPts values capture only more persistent and spatially concentrated point groupings, leading to larger ε values at the inflection point. Rather than fixing minPts in advance, we adopted a two-step approach: first, for each minPts setting we determined an appropriate ε interval based on the location of the inflection point in the corresponding curve. We then identified the optimal ε that minimized the absolute count error with respect to the known number of industrial fire points (22 378), where this value represents the total number of points with TYPE = 0 falling within the industrial zone mask.

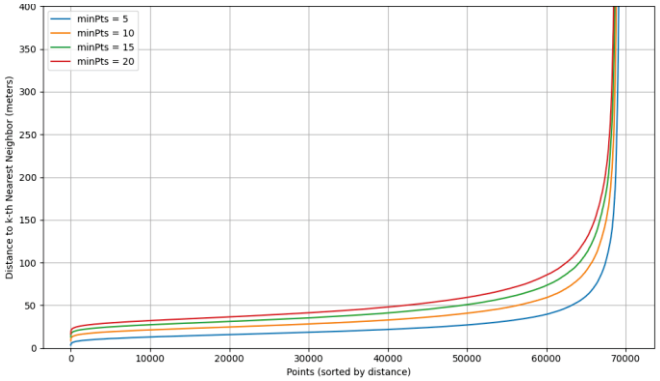


Fig. 1. k-distance plots for industrial fire points (*minPts* = 5–20)

We then performed a scripted grid-search: for each minPts (5, 10, 15, 20), we varied ε within its knee-derived interval and computed the absolute count error as (1).

$$E_{count}(\varepsilon) = \left| N_{clustered}(\varepsilon) - N_{ref} \right|, \tag{1}$$

where $N_{clustered}(\varepsilon)$ is the number of points DBSCAN assigns to any cluster and N_{ref}= 22 378 is the reference count of points in the buffered mask. The optimal radius ε^* (2).

$$\varepsilon^* = \arg\min_{\varepsilon} E_{count}(\varepsilon) \tag{2}$$

is therefore the value that makes the clustered count match N_{ref} most closely. This approach yields parameters that balance capturing the core jitter clouds and covering the full industrial footprint without over-smoothing.

In the second stage of the analysis, we generated DBSCAN clusters using the selected (ε, *minPts*) combinations and spatially evaluated the resulting cluster layers against a polygonal mask of industrial zones. After dissolving all cluster hulls into a single mask and performing unique-point counting, we measured both the count-based error and spatial overlap metrics for each minPts. This comparison allowed us to assess which minPts value resulted in clusters most spatially consistent with known industrial areas. This dual optimization ensured that the final parameters captured not only the correct number of industrial fire points but also their expected geographic distribution. Although VIIRS pixels nominally span 375 m on a side, the k-distance analysis consistently reveals optimal ε values of 70–140 m (see Table I) – well below that spatial granularity. This is not a contradiction but a direct consequence of VIIRS Level-1B geolocation uncertainty (1-σ ≅ 75 m) rather than the sensor's pixel footprint. Repeated detections of the same fixed heat source scatter around the true location by tens of metres, producing a dense "jitter cloud" whose internal point spacing is governed by sub-pixel positioning errors. Thus, choosing ε at roughly the 5th-neighbor knee (≈ 73 m) and 20th-neighbor knee (≈ 138 m) effectively isolates these jitter swarms without conflating distinct facilities or wildfire events.

TABLE I. DBSCAN PARAMETER TUNING RESULTS FOR VARYING MINPTS VALUES

Parameters	*minPts*			
	5	**10**	**15**	**20**
ε range tested (m)	40–100	60–120	80–140	90–160
optimal ε (m)	73	100	124	138
$N_{clustered}(\varepsilon)$	22373	22369	22398	22374
Abs. error	5	9	20	4
% in industrial zones (precision)	94.25%	94.33%	94.37%	94.40%
number of clusters	283	94	71	66
min. cluster area, km^2	3.4e-5	7.71e-3	0.011	0.035
max. cluster area, km^2	5.650	9.614	9.784	9.847

Based on the multi-criteria results summarized in Table II, we selected minPts = 20 for our final DBSCAN configuration. With this setting (and an optimal ε = 138 m), the filter achieves the highest spatial precision, as 94.40 % of clustered points fall within the industrial-zone mask. Here, "% in industrial zones" is computed as the ratio of unique clustered points whose centroids intersect the polygonal mask of known industrial areas to the total number of clustered points, thereby quantifying the filter's specificity. Moreover, increasing minPts to 20 dramatically reduces over-fragmentation: only 66 clusters are detected (compared to 283 when minPts = 5), and the minimum cluster area (0.0347 km²) aligns well with the typical scale of actual industrial facilities. In contrast, lower minPts values produce many small, dispersed clusters – often outside known industrial zones – thereby compromising precision.

To evaluate the performance of the DBSCAN-based filtering approach, we compared the resulting cluster coverage against a labeled subset of industrial thermal anomaly points (TYPE = 2). A total of 40 913 such points were present in the unfiltered dataset. Spatial intersection analysis revealed that 40 471 of them (98.92%) were captured within the polygons of DBSCAN clusters generated using the parameters ε = 138 m and minPts = 20 (Table II). Only 1.08% points remained outside cluster boundaries. This near-complete overlap confirms that the density-based filtering method accurately identifies the spatial signature of persistent industrial hotspots.

To illustrate the spatial benefits of density-based filtering, we examined VIIRS detections (2019–2024) at ArcelorMittal Kryvyi Rih – one of Ukraine's largest metallurgical complexes (Fig. 2).

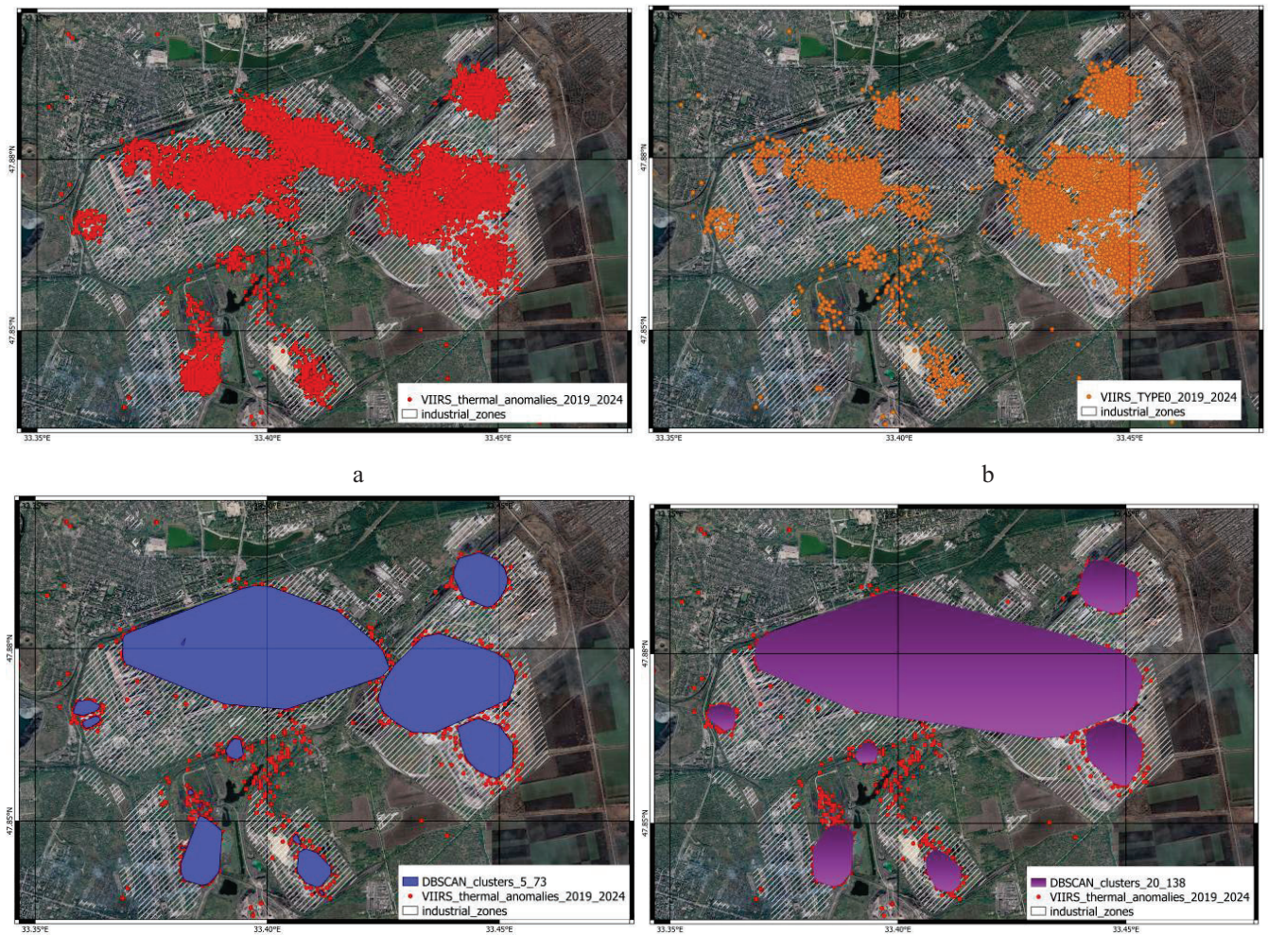


Fig. 2. Thermal anomaly clustering at PJSC ArcelorMittal Kryvyi Rih: a) All VIIRS detections (2019–2024); b) Filtered anomalies (TYPE = 0); c) DBSCAN clusters (minPts = 5, ε = 73 m); d) DBSCAN clusters (minPts = 20, ε = 138 m)

TABLE II. DISTRIBUTION OF POINTS IN INDUSTRIAL MASK AND CLUSTER CAPTURE BY TYPE (2019-2024)

TYPE	Total points	Points in industrial zones	% of Total	Points in clusters	% of Total
0	362 322	22 378	6.16%	22 374	6.18%
2	40 913	38453	93.99%	40471	98.92%
3	10 768	9279	86.17%	9563	88.81%
Total	414 033	70 110	16.93%	72408	17.49%

After excluding all TYPE=2 points, numerous "hotspots" still lay within the plant boundary, underscoring the insufficiency of attribute-only filters (Fig. 2b).

Although our 5-year analysis in Ukraine identified minPts = 20 as optimal for capturing truly persistent clusters over multi-year intervals (Fig. 2d), this choice is by no means universal. Rather than prescribing specific minPts values, we recommend a general guiding principle: minPts should scale with the number of VIIRS overpasses in the chosen time window, so that clusters reflect genuinely stationary sources rather than transient events. For longer, multi-year intervals – when a fixed industrial facility may be imaged dozens of times – higher minPts values are appropriate (e.g., minPts ≈ 20); for shorter, monthly intervals – when only a handful of detections occur – lower values (e.g., minPts ≈ 5) suffice (Fig. 2c). In practice, one can calculate the total number of overpasses (or detections) expected during the analysis period and set minPts to a fraction (for example, 20–30 %) of that count to capture only those anomalies that recur consistently throughout the window. This data-driven scaling ensures that DBSCAN isolates truly stationary industrial hotspots at any temporal granularity, from multi-year to monthly studies.

IV. CONCLUSIONS

In this work, we have shown that conventional attribute- or intensity-based filters are unable to fully remove persistent industrial thermal anomalies – and the attendant sub-pixel "jitter clouds" caused by VIIRS Level-1B geolocation errors – from satellite active-fire datasets. By introducing a density-based clustering step (DBSCAN) with parameters selected via k-distance analysis and count-error minimization against an industrial-zone mask, we achieve robust removal of industrial noise while preserving genuine wildfire detections.

Crucially, we find that all tested (ε, MinPts) combinations yield acceptable precision (> 94 %) over multi-year intervals, and that MinPts should scale with the number of VIIRS overpasses in the chosen time window. Higher MinPts values suit multi-year analyses, while shorter intervals (annual, monthly) call for proportionally lower MinPts to isolate truly persistent sources. This data-driven scaling ensures the method's applicability at any temporal granularity.

Looking forward, integrating spatio-temporal clustering (e.g., ST-DBSCAN) could further discriminate transient wildfire bursts from stationary industrial sources. Finally, embedding this lightweight, two-parameter filter into operational fire-monitoring workflows promises cleaner, more accurate wildfire trend analyses in heavily industrialized regions.

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Analysis of the Characteristics of Vibration-Damping and Vibration-Isolating Axial Active Magnetic Support Taking into Account Control Laws Using Analytical and Numerical Methods

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***Abstract*—The paper considers the issues of determining the force and stiffness characteristics of a controlled axial magnetic support using analytical and numerical mathematical modelling. This support is a disk on a movable rod or shaft (depending on the design purpose), which is located between two symmetrical cores of the armored type. Due to the application in the system controlling the axial position of the disk, almost any law of control of force electromagnets, this electromechanical support provides the required vibration-damping and vibration-isolating characteristics. In the case of a horizontal position of the shaft, the considered support is an axial thrust active magnetic bearing, which for research is practically implemented in an operating laboratory rig of a full magnetic suspension of a rotating model rotor in two radial passive magnetic bearings made of rare-earth annular permanent magnets. In the case of a vertical position of the shaft, this magnetic suspension can be considered as a whole as a passive-active vibration-damping and vibration-isolating complex support for either a load located on top, or a body suspended from below. Then the rotor is considered as a non-rotating rod-shaft that holds the vibration-isolated load. In the first and second cases, the accuracy of the dynamic characteristics of this mechatronic system depends on the accuracy of the design calculation of the axial active magnetic support characteristics from the point of view of the possibility of damping the parasitic vibrations of the working body, that is, the suspension object (rotor or load). With the use of analytical and numerical approaches, the possibilities of determining strength and stiffness characteristics depending on the parameters of the axial magnetic support and the control system are investigated. Distinctive features inherent in these two approaches in solving the given problem are shown, with an analysis of their influence on the accuracy of the simulation. The results of analyzes with their verification are given. For the laboratory rig, the possibility of its use as a whole as a vibration-damping and vibration-isolating complex fully magnetic passive-active type support has been confirmed. If necessary, with relatively large dimensions and mass parameters of the object of vibration isolation, a combination of several such complex supports can be used, acting in concert due to the use of cross connections in the control systems of the axial electromagnetic support element, which is the subject of this work.**

Keywords— axial active magnetic support, magnetic bearing, electromagnetic circuit, control system and law, analytical and numerical models, magnetic force and stiffness, laboratory rig

I. Introduction

One of the main issues in the creation and operation of machines, mechanisms, devices, and equipment for various purposes is the reduction or even elimination of possible vibrations of individual structural elements or the structure as a whole [1]. Such mechanical vibrations are caused by various dynamic effects [2], which can be generated by the operating modes of the components (such as, for example, the rotor or shaft of a fan, turbogenerator, turbocompressor with residual imbalance) or caused by external factors (earthquake, the supporting vehicle platform motion on a rough road, the fall of a watercraft in stormy conditions or an aircraft in the turbulence zone, impacts of various nature, etc.). In all these cases, the excitation of mono- or polyharmonic mechanical oscillations (vibrations) with large amplitudes, especially in resonant modes, can lead to the destruction or damage of the main structural elements. Such structural elements are often objects of vibration isolation or vibration damping. The purpose of this process is not to transmit the vibration energy passing through the more durable components of the structure from the source to the most vulnerable working or executive element. This element can be a rotor of a turbomachine with impellers, a measuring device on a moving base or a platform that requires precise guidance of a weapon, a mobile hospital bed, etc.

Currently, new types of vibration dampers and isolators are being created and widely used, including those based on the principle of quasi-zero stiffness, for vibration protection of devices, equipment, cargo, and people from dynamic effects in extreme conditions. Therefore, this topic is relevant in various areas throughout the entire period of development of technical devices, including at the present time.

II. Literature Review

To achieve the described goal, two main approaches are used – vibration damping and vibration isolation. Vibration damping or dynamic damping is the introduction of additional mass into the vibrating system or an increase in the stiffness of the system. Dynamic vibration dampers of various types are used for this purpose: spring, impact, pendulum, eccentric, hydraulic, and others [3]. These elements can be of passive or active types [4, 5]. The disadvantage of a dynamic damper is that it operates only at a certain frequency, which corresponds to its resonant mode of vibrations. This problem can be solved by electromagnetic

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dampers due to the implementation of a certain control law [6-8], which allows changing the mechanical stiffness and/or damping properties (for example, by changing the active resistance in the circuit to convert a larger amount of kinetic energy into thermal energy). Vibration isolation is an introduction of an additional elastic connection into the vibrating system with the aim of weakening the transmission of vibrations to the protected object (adjacent element, structure or workplace) from the excitation source. Vibration isolation is carried out by installing a vibration source on vibration isolators [9]. Among them, there are also vibration isolators based on electromagnetic interaction, and they can also be of active or passive types [10]. A separate class of vibration isolators are vibration isolation systems with quasi-zero stiffness (QZS) [11]. Vibration isolators of this type implement the QZS characteristic by combining components with positive stiffness and components with negative stiffness. In this case, the addition of non-linear force characteristics of two components allows creating ranges of zero stiffness on the total force characteristics that correspond to the working position of the object of vibration isolation. This enables effective isolation of low-frequency vibrations, providing sufficient load-bearing capacity and maintaining the system stability. Such systems are now used in ultra-precise production and various precise measurements, in aerospace detection, in regular and non-regular work or delivery of weapons, equipment or people, for example, in extreme conditions with an explosive hazard. Passive vibration isolation systems QZS mainly include mechanical springs [12], pre-bent beams [13], geometrically nonlinear structures [14], electromagnetic and magnetic structures from permanent magnets [15, 16], bio-inspired structures [17], and composite structures [18]. For magnetic elements of negative stiffness, a special arrangement of magnets is used, allowing easily controlling the negative stiffness characteristic. Active and semi-active elements of negative stiffness [19, 20] are basically electromagnetic devices and can better adapt to various conditions based on displacement feedback.

Thus, one of the progressive mechanisms of building vibration damping or vibration isolation systems is the use of uncontrolled and controlled, i.e. passive or active, magnetic or electromagnetic elements to ensure the required damping or force characteristics [21, 22]. Dampers based on such principles include eddy current and coil dampers [23, 24], permanent magnet dampers, active and semi-active electromagnetic dampers (EMD) [25, 26]. At the same time, active EMDs use control systems (CS) to actively adjust the damping force, while semi-active EMDs use, for example, a variable resistor or another method to change damping [27]. The task of the electromagnetic damping system, as well as based on the other principles, is to transform the kinetic energy of mechanical vibrations into another type of energy, for example, into thermal energy.

The principle of operation of the insulator with quasi-zero stiffness is demonstrated in Fig. 1. The element of positive stiffness (PS) is connected to the element of negative stiffness (NS) and within the working range, the offset $[x_1, x_2]$ forms a range of zero stiffness. This range is determined by the static load, for example, the weight of the vibration isolation object. In this case, when the base is dynamically excited, the vibration is not transmitted through such a vibration isolator.

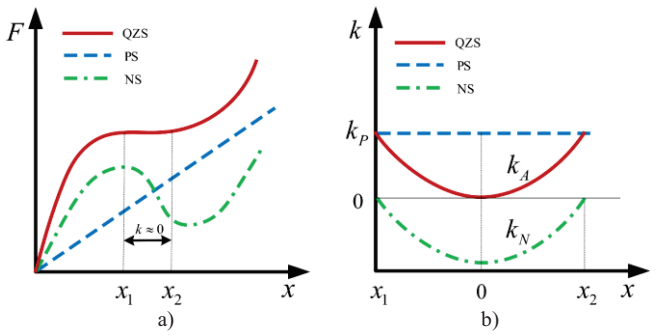


Fig. 1. The QZS vibration isolator operation principle [9]: a) force F vs displacement x dependence; b) stiffness k vs displacement x dependence

Electromagnetic dampers, vibration dampers and vibration isolators, including QZS, are essentially mechatronic systems and are designed based on simulations in various ways. For example, the works [28, 29] describe analytical and numerical approaches for evaluating the characteristics of various electromagnetic dampers and vibration dampers with verification based on experimental data [30]. The same methods are used for mathematical modelling and determination of properties and characteristics of semi-active [31] and active [32, 33] vibration isolators.

In both cases, either when applying the principles of vibration damping, or when applying vibration isolation mechanisms, the question of exact parametric determination of the characteristics created for a specific system of vibration dampers or vibration isolators is always acute. This study is devoted to considering the possibilities of analytical or numerical modelling for the analysis of damping elements based on magnetic or electromagnetic interaction.

III. The Aim and Object of Research

The aim of the research is the analytical and numerical determination of the force and stiffness characteristics of an axially controlled magnetic support with a comparative analysis to identify its pros and cons, as well as assessing the possibilities of using the considered support type as a vibration-damping or vibration-isolating structural element.

The object of research is one of axial controlled electromagnetic support types. This support is practically implemented in the form of a thrust bearing in a laboratory-experimental rig (Fig. 2) of a complete combined magnetic-electromagnetic rotor suspension and is a disk on a shaft, which is located between two symmetrical cores of the armor type [34]. The scheme of the control system of this support as an axial active magnetic bearing (AMB) is shown in Fig. 3 along with the force characteristics of radial annular passive magnetic bearings (PMBs) made of rare earth materials (NdFeB alloy). Complete rotor magnetic suspensions are widely used in various machines like turbocompressors of gas-pumping units [35], and the rig was created to study the possibility of replacing radial active magnetic bearings with passive ones [36], as well as to study the influence of various factors [37, 38] and dynamic excitations [39].

The laboratory-experimental rig has the following parameters of the structural elements: mass of the rotor m=2.496 kg, diametral mass moment of inertia (MOI) J_1=0.0107386 kg·m^2, polar mass MOI J_3=0.0033177 kg·m^2; the distance from the left and right PMBs to the center of gravity of the rotor, respectively l_1=0.107 and l_2=0.177 m, radial PMB clearance δ_r=5.5·10^{-3} m.

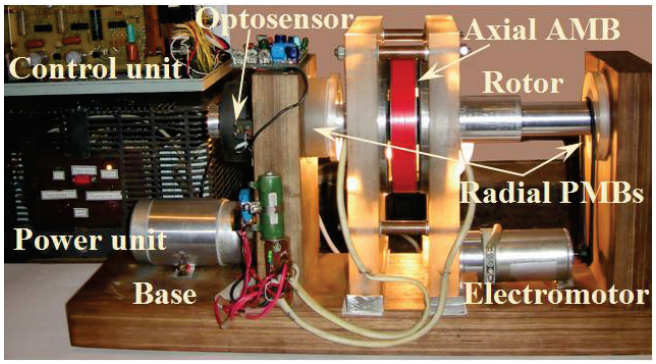


Fig. 2. Laboratory-experimental rig of a model rotor combined magnetic-electromagnetic suspension.

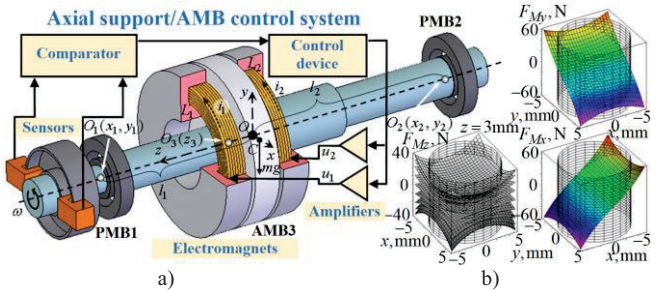


Fig. 3. Complete magnetic-electromagnetic suspension model rotor: a) structural diagram of the control system of a thrust AMB (axial support); b) force characteristics of radial PMBs.

The axial AMB is presented in Fig. 4 and has the following parameters: outer and inner diameters of electromagnets (EMs) 3 and 4 – 0.12 and 0.05 m, lengths of each EM 3 and 4 in the axial direction – 0.021 m, outer and inner diameters of the windings 5 – 0.1 and 0.072 m, the number of winding turns – 300, active resistance of the windings – $r_{c1,2}$=5 Ohm, thickness of the disk 2 on the shaft 1 – 0.02 m, nominal clearance from each side – δ_a=3 mm, basic control voltage U_0=24 V, control frequency 300-400 Hz [34].

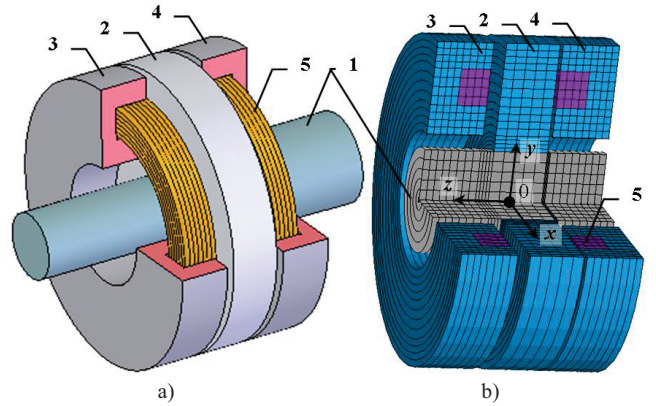


Fig. 4. Models of a thrust AMB (axial support): a) geometric model with a centrally located disk; b) finite element model with an offset disk.

IV. Approaches to Mathematical Determination of Force Characteristics of Axial Controlled Support

Different approaches can be used to calculate the force and stiffness characteristics of an axial electromagnetic support. Analytical and numerical methods are the most practical. Each of them has its advantages and disadvantages. Therefore, the analytical method of AMB electromagnetic circuit analysis [40, 41] is used as an analytical approach, based on the introduction of detailed equivalent schemes for the replacement of magnetic circuits and the application of the method of contour flows. The finite element method is used as a numerical approach [41, 42].

For the correct determination of the dependences of restoring forces on displacement and currents in the windings of EMs in the supports, the control law must be considered. For the laboratory-experimental rig (Fig. 2, Fig. 3), the control obeys the law in the form of quadratic dependences of the average voltage $u_{c1,2}$ on the AMB windings, which provide a given change in the average currents $i_{c1,2}$ in accordance with the adopted control parameter k_i (without taking into account the constant time of the change of currents in the EM windings due to inductance):

$$i_{c1,2}=\frac{(i_{\max}-2i_{\min})z_3^2}{2\delta_a^2}\mp\frac{i_{\max}z_3}{2\delta_a}+i_{\min},\ u_{c1,2}=r_{c1,2}\,i_{c1,2},\quad (1)$$

where $i_{\max}=U_0/r_{c1,2}$, $r_{c1}=r_{c2}$, $i_{\min}=k_i i_{\max}$.

The bias current $i_{\min}$ in (1) cannot be equal to 0, so its value creates a preload and provides the required stiffness in the equilibrium position (at the central position).

A. Analytical determination of magnetic forces

For the analytical description of the system, a detailed equivalent scheme of the substitution of magnetic circuits is used, considering the scattering fluxes through the grooves filled with copper windings. This scheme was synthesized as a result of the analysis of the magnetic field equipotential line distribution (Fig. 5a) and is presented in Fig. 5b, where the designations are introduced: Φ_k – magnetic fluxes in sections of the circuit; Φ_{Ck} – contour fluxes; R_k – magnetic resistance of sections of the circuit, and R_{pk} – of poles, R_{gk} – of air gaps between poles and a rotor, R_{sk} – of sections of a stator yoke between two poles, R_{lk} – of scattering between poles (grooves filled with copper winding), R_{ak}=R_a – of a disk; e_k=$i_{ck}w_k$ – coil magnetomotive forces.

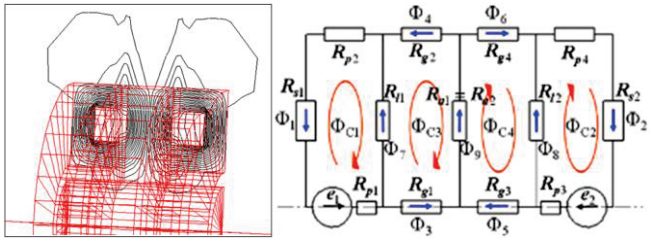


Fig. 5. Equipotential magnetic field lines: at the central position of the disk (left) and the equivalent scheme of the substitution of magnetic circuits taking into account the scattering fluxes (right).

The magnetic energy of the axial AMB (the entire magnetic circuit energy as the circuit sections energies sum):

$$W=\frac{1}{2}\left[\begin{array}{l}(R_{s1}+R_{p1}+R_{p2})\Phi_1^2+(R_{s2}+R_{p3}+R_{p4})\Phi_2^2+R_{g1}\Phi_3^2+\\+R_{g2}\Phi_4^2+R_{g3}\Phi_5^2+R_{g4}\Phi_6^2+R_{l1}\Phi_7^2+R_{l2}\Phi_8^2+R_{a1}\Phi_9^2\end{array}\right].\quad (2)$$

Applying the method of contour fluxes and considering that the magnetic resistance of the air gaps depends on the displacement of the disk relative to the central position give the expression of the electromagnetic force:

$$F_{Mqj}=-\frac{\partial W}{\partial q_j},\ q_j=\{x_1,y_1,x_2,y_2,z_3\}.\quad (3)$$

As a result of the variant analysis using this approach, the force characteristics were obtained. Fig. 6a shows the type of control law (1) corresponding to the previously adopted CS parameters, and Fig. 6b shows the dependences of forces on the disk axial displacement for these parameters. The current values marked with dots correspond to the experimental conditions, and discrepancy between the force characteristic for a plane-parallel gap (thick purple line) and experimental data is less than 3%. Fig. 6b also illustrates the change in the force characteristics for different disk tilts (gray lines of different gradations). Analysis of the results showed that the error in calculating the forces without considering the dependence on disk tilt can exceed 30%.

Changing the parameters of the control law as shown in Fig. 6c, 6d, other force characteristics can be obtained both for a plane-parallel gap and different levels of disk tilt relative to the axis within the geometrically possible states considering the axial gap. A certain ratio of the system parameters can achieve a constant range on the force characteristic, i.e. a quasi-zero stiffness range, as shown in Fig. 6e, 6f within the displacement range of -0.001..0.001 m.

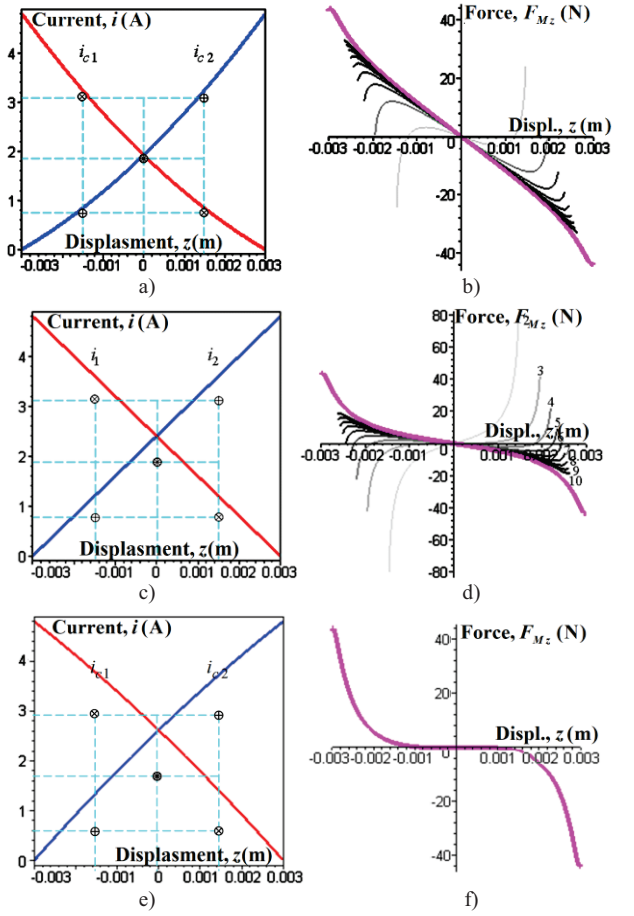


Fig. 6. Force characteristics of the AAMS with different control law parameters k_i: a) change in average currents for k_i=0.4; b) dependence of magnetic force for k_i=0.4; c) change in average currents for k_i=0.5; d) dependence of magnetic force for k_i=0.5; e) change in average currents for k_i=0.545; f) dependence of magnetic force for k_i=0.545.

The analytical approach advantage is the possibility of precise mathematical calculation of the electromagnetic force value depending on the disk coordinates considering disk tilt with low time and system resource costs allowing conducting multivariate studies to select rational or optimal design parameters and control systems. The disadvantages of the approach are a failure to include the design spatiality, as well as consideration only of some main paths of magnetic flux propagation in the equivalent circuit in accordance with the adopted simplifications (Fig. 5b), which can negatively affect the accuracy of determining magnetic forces.

B. Numerical determination of magnetic forces

The strategy for numerical determination of magnetic forces is based on the application of the finite element approach and solution of the magnetostatics problem in a volumetric formulation for a specific position of the disk on the shaft section using a finite element model with a plane-parallel air working gap, like that shown in Fig. 4b [42, 43]. During calculations, elements simulating the air in the gaps and grooves, as well as around all parts, are added to this model. Conditions at infinity are set at the outer boundaries of the air. The load is specified as the current density in the elements simulating the windings of both EMs, considering the direction in each of them. The current density in each winding corresponds to the current value determined by the control law for a specific disk displacement. In this case, the mathematical model is formed relative to the components of the vector magnetic potential [42, 43]. As a result, the spatial distributions of the magnetic induction vectors and magnetic field strength moduli are determined (Fig. 5a). Magnetic forces are calculated using the Maxwell stress tensor or the virtual work principle. The force characteristic is constructed based on the results of analyses for different disk positions within the gap using approximation by a nonlinear polynomial of odd degree based on discrete calculated values. Stiffness is determined as the first derivative of the force in the direction with the opposite sign, i.e. K_M=-$\partial F_M/\partial z$.

Fig. 7a, 7b shows force and stiffness characteristics of the axial active magnetic support (AAMS) with the control law initial parameters (1) corresponding to Fig. 6a. Comparison of the experimental and calculated approximated values gave a relative difference <1%, which is less than when using the analytical approach (Fig. 6b). Fig. 7c, 7d presents the same results, but with different values of the basic control voltage when using pulse-width modulation in the control system [34] showing that changing the basic supply voltage allows achieving the presence of quasi-zero stiffness range.

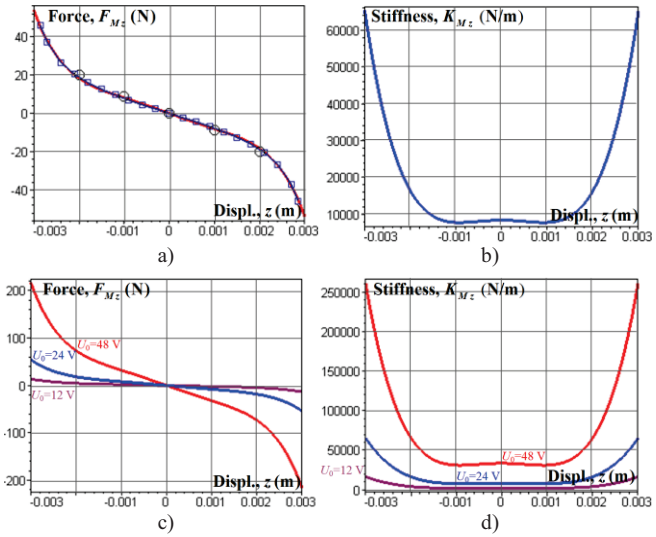


Fig. 7. Dependence of the AAMS characteristics on the disk displacement with the control law parameters (CLPs) (○ – experimental and □ – calculated data): a) magnetic force for intial CLPs; b) stiffness for initial CLPs; c) magnetic force for varios CLPs; d) stiffness for varios CLPs.

The numerical approach advantage is the possibility of considering magnetic flux spatial distribution, which practically coincides with the real one, see in Fig. 5 (left). This allows taking into account all the features (fluxes along the sections of the magnetic circuit of moving and stationary parts, air gaps outside the magnetic circuits, including parasitic and scattering, and others) and increasing the result accuracy. Parametric automation allows for a variant search for the best designs. The approach disadvantages include the complexity of implementing the disk tilt in the calculated finite element model within the topological limitations of a small gap. It is associated with a large error caused by mesh discretization due to the presence of relatively small elements with a degenerate shape. In addition, the numerical method requires a much larger system and time resources.

V. Passive-Active Vibration-Damping and Vibration-Isolating Complex Magnetic Support

The results of calculations and indirectly of experiments, presented in Fig. 6e, 6f and Fig. 7c, 7d and given in the publication [45], respectively, proved the possibility of using a magnetic-electromagnetic passive-active suspension (Fig. 2, 3) as a complex support with quasi-zero stiffness. This is structurally feasible with some modification. With a vertical shaft arrangement shown in Fig. 8, this magnetic suspension can be considered as a whole as a passive-active vibration-damping and vibration-isolating complex support of either a load located above or a body suspended from below. Then the rotor is a non-rotating rod-shaft that holds the vibration-isolated load. In the case of structural or operational features of the supported or suspended body, the rod-shaft can rotate around the axis making a support a kind of a thrust bearing. The supported vibration damping and vibration isolation object weight is compensated by the bias current in the upper winding, i.e. it is provided by a certain constant difference of the average currents i_{c1} and i_{c2} in the upper and lower windings. This is taken into account in the control law (1) as a constant component of the bias currents. In addition to this feature, the force and stiffness characteristics with quasi-zero stiffness ranges, similar to those shown in Fig. 6-7, will be provided as before.

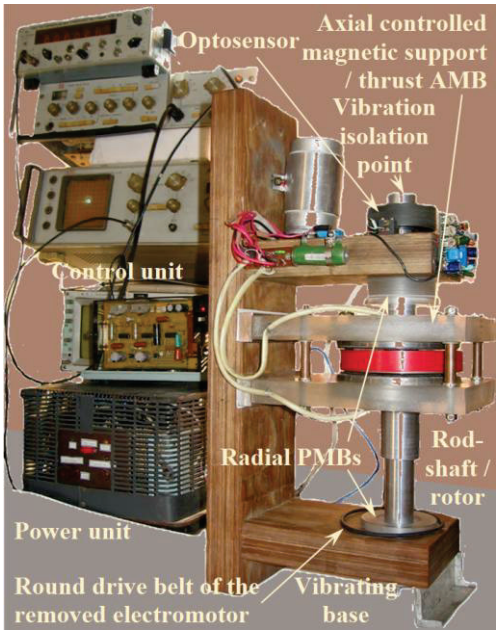


Fig. 8. Laboratory and experimental rig of passive-active vibration-damping and vibration-isolating complex magnetic support.

VI. Conclusion and Discussion

The paper considers approaches to analyzing the force and stiffness characteristics of a controlled AAMS using analytical and numerical methods. It was found that each of them allows determining the force characteristics with sufficient accuracy depending on the AAMS and CS parameters. Each of these approaches has its own features affecting accuracy. They can be used depending on the needs, required multivariance of calculations with different parameters, limited system and time resources, etc.

These approaches were applied to practical analysis of one of the AAMS variants with verification and validation of calculation models based on experimental data. The obtained results showed the possibility of providing the required force and stiffness characteristics with ranges of the desired quasi-zero stiffness by selecting applied control law parameters.

Based on the experimental rig of a combined passive-active magnetic-electromagnetic suspension of a model rotor, a variant of a vibration-damping and vibration-isolating complex magnetic support is considered. The proposed complex passive-active magnetic support, due to its controllability, can perform both the functions of vibration damping of the held object (device, instrument, equipment, dangerous goods or a person) and the functions of vibration isolation. In this case, vibration damping is ensured by controlled and managed resistance in the circuit of the EM windings or is implemented using special CS algorithms, which is now practically used in many rotary machines with rotors in AMBs. In turn, vibration-isolating properties are ensured by quasi-zero stiffness ranges implemented by the adopted control law. When it is necessary to isolate large and heavy objects from vibration, several complex supports can be used. Their synchronous operation is ensured by mutual connections in the electromagnetic AAMS CS.

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Analysis of Instantaneous Power of Higher Harmonics in a Diesel Generator-Based Electrical System with Grid Connection

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***Abstract*— As a result of the analysis of existing studies, it was found that in the operation of a power supply system with a medium-capacity generator, insufficient attention is paid to the distribution of higher harmonic powers among system components. Using a microgrid model that includes a synchronous generator, an induction motor, a transformer, an external grid, and a current harmonic source, a series of experiments were conducted by introducing higher current harmonics and recording distortion levels and RMS power in individual branches. The analysis of the obtained results made it possible to es https://doi.org/10.5772/intechopen.87093lish regression relationships between harmonic currents and phase RMS power, as well as to assess their impact on power quality indicators, taking into account the type of load and the specific branch of the microgrid. It was found that in the studied setup, the third harmonic has the greatest impact on the transformer, while the fifth and seventh harmonics predominantly affect the generator and the electric motor..**

Keywords*— electric power, current and voltage harmonics, root mean square power, power quaility

I. Introduction

The intermittent operation of alternative energy sources, such as solar and wind [1], due to natural conditions, or the varying load levels during the operation of a diesel generator [2], affects the system—including voltage, frequency, and overall network performance. With the widespread integration of inverters, frequency converters, and other nonlinear loads, harmonic distortion is becoming an increasingly serious issue in distribution networks. Distorted supply voltage leads to distorted currents, while harmonic currents from consumers further degrade the power supply voltage quality. Advanced approaches such as power flow analysis in system branches [3, 4], assessment of voltage and current harmonic conditions [5, 6], and identification of harmonic source parameters [7, 8], help identify dominant sources of higher harmonics. However, they do not allow for determining the specific harmonic contribution level of each individual harmonic component, as these methods are based on generalized indicators.

II. Analysis Of Previous Research

In [9], the process of integrating a solar power plant into the energy system is considered for the purpose of planning solar power utilization. To effectively analyze voltage quality and calculate harmonic losses in a distribution network integrated with a solar power plant, a simplified scheme is used to model the harmonic impedance of key equipment such as transmission lines and transformers. Based on calculations of harmonic losses using measurement data, it was found that harmonic losses are primarily determined by harmonic current and account for approximately 0.6% of the total network losses.

Studies [10, 11] examine the impact of inverters on current and voltage harmonic distortion and propose a new impedance-harmonic relationship model for identifying harmonic resonance caused by inverters. The validity and effectiveness of the model are confirmed by its analytical simplicity and the fact that the frequency characteristics and coupling effects can be clearly observed for inverters with different control schemes.

Harmonic distortion in isolated synchronous generator machines [12] is more significant due to the limited short-circuit power level compared to the short-circuit power level at the buses of electrical networks. In addition, harmonic currents circulate only within a single loop between the generator and the load, and they cannot flow through the power grid bus, which contributes to increased harmonic distortion.

The impact of cable line length on the amplification of harmonic distortion during the operation of a solar power plant is discussed in [13]. The authors propose connecting the voltage source at the midpoint of the load node to reduce the influence of the cable on the system's capacitive current. It is established that with constant transmission capacity, the closer the connection is to the end of the plant's transmission line, the higher the level of harmonic distortion in the distribution network.

A similar issue is addressed in [14], which uses a methodology to study and define the limiting values of additional and cumulative harmonic distortion depending on the proximity of multiple harmonic sources to each other.

Thus, summarizing the above, it can be stated that in systems with power conversion devices used in certain generation complexes and/or consumer loads, harmonic current distortion is present. While the influence of higher harmonics on each individual system branch has been studied, the distribution of harmonic power among the branches has not been sufficiently addressed.

Objective of the study: To investigate the distribution of higher load current harmonics in a system with a limited-capacity generating unit and an external grid, based on their root mean square power, using the design of experiments method.

III. Materials And Results Of Research

Let us consider a diesel generator system complex [15] with a load and an external grid, structured as shown in Fig.

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1. The system consists of the following components: a diesel engine (DE), an excitation system (ES), a synchronous machine (SM), a linear load (LL), a nonlinear load (NLL), a transformer (TR), and the grid..

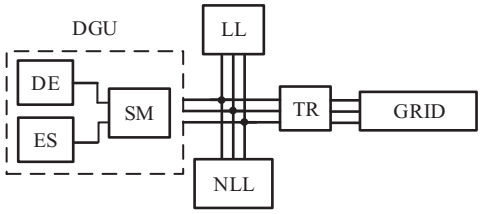


Fig. 1. Structural diagram of a diesel generator system connected to the grid

The synchronous generator is considered by some researchers both as a source of higher harmonic electromotive force , due to the specific characteristics of the machine's magnetic circuit, and as a consumer of higher harmonic currents generated by other harmonic sources [16].

By relating the values of X_{fv} i and and X_{kv} to frequency (for the coupling between stator and rotor windings), the slip for the v-th harmonic is introduced:

$$s_v = \frac{v \pm 1}{v} \tag{1}$$

Then the inductive reactance for the *v*-th harmonic is obtained:

$$X_{dv} = vX_l + \frac{1}{\frac{1}{vX_{ad}} + \frac{\sqrt{v \pm 1}}{vX_f} + \frac{\sqrt{v \pm 1}}{vX_k}} \tag{2}$$

where X_f - the corresponding stator resistance for the 1st harmonic; X_k - the rotor winding leakage reactance under strong skin effect conditions; X_{ad} - the mutual inductance reactance for the 1st harmonic; X_{fv} - the rotor winding leakage reactance under pronounced skin effect conditions*:* X_{kv} - the leakage reactance of the damper winding.

Asynchronous motors [17], at the synchronous rotational frequency corresponding to the *v*-th voltage harmonic with a frequency equal to vn_{syn}, have a slip S_v at the rotor speed frequency that is equal to the slip for the *v*-th harmonic:

$$S_v = \frac{vn_{syn} \pm (1-s)}{vn_{syn}} = \frac{v \pm (1-s)}{v} \tag{3}$$

Assuming a strong manifestation of the skin effect, $(R_{kv} = R_k\sqrt{v})$ the total impedance of the asynchronous motor for the v-th harmonic is determined as:

$$Z_v = \sqrt{R_{kv}^2 + X_{kv}^2} = \sqrt{v^2 X_k^2 + vR_k^2} \approx vX_k \approx vZ_k \tag{4}$$

where R_k , X_k , Z_k - the active, the reactive an the total short-circuit impedance of the asynchronous motor, respectively.

The impedance of a 6–10 kV transformer, whose load consists of asynchronous motors, can be represented in the equivalent circuit for higher harmonics as the sum of the short-circuit impedances of the transformer and the asynchronous motors:

$$X_{u,Tv} = vX_{k,T} + \frac{vU_{nom}^2}{0,75k_S S_{nom}} \tag{5}$$

where $X_{k,T}$ - the short-circuit impedance of the transformer at the industrial frequency, Ohms;, U_{nom} , S_{nom} - the rated voltage, in kV, of the high-voltage winding (6–10 kV), the rated power of the transformer, in MVA; k_S - the starting current multiple of the equivalent asynchronous motor (on average, k_S = 5.5), which forms the load of the transformer; 0,75 - a coefficient that accounts for the transformer's loading level.

Using the indicators provided in [18], a method was proposed for determining the components of power that reflect specific types of power distortion in the transmission line, namely.

Current distortion power:

$$D_I = U_{rms.h} I_{rms.h} = U_{rms.1} I_{rms.1} THD_I \tag{6}$$

Voltage distortion power:

$$D_U = U_{rms.h} I_{rms.1} = U_{rms.1} I_{rms.1} THD_U \tag{7}$$

Total harmonic power:

$$S_H = U_{rms.h} I_{rms.h} = U_{rms.1} I_{rms.1} THD_U THD_I \tag{8}$$

where THD_U , THD_I - Total Harmonic Distortion coefficient of voltage and current; $U_{rms.1}$, $I_{rms.1}$ - root mean square (RMS) values of current and voltage of the fundamental harmonic; $U_{rms.h}$, $I_{rms.h}$ - RMS values of current and voltage of the higher harmonicsк..

The specified powers are derived from the total apparent power represented as follows:

$$\begin{aligned} S^2 &= \left(U_{rms} I_{rms}\right)^2 = \\ &= \left(U_{rms.1}^2 + U_{rms.h}^2\right)\left(I_{rms.1}^2 + I_{rms.h}^2\right) = \\ &= \left(U_{rms.1} I_{rms.1}\right)^2 + \left(U_{rms.1} I_{rms.h}\right)^2 + \\ &+ \left(U_{rms.h} I_{rms.1}\right)^2 + \left(U_{rms.h} I_{rms.h}\right)^2 . \end{aligned} \tag{9}$$

But in a similar way, in this case, it is possible to use the RMS values of the current $I_{rms.h}$ and voltage harmonics $U_{rms.h}$ instead of the RMS values of the higher current and I_h voltage harmonic U_h. However, such an option is not considered in [18].

Another option for power estimation in AC circuits with nonsinusoidal waveforms is used in [19, 20]. The authors justify the feasibility of using the root mean square (norm) of the instantaneous power.

$$\|p\| = P_{rms} = \sqrt{\frac{1}{T}\int_{t_0}^{t_0+T} p^2 dt} = \sqrt{\frac{1}{T}\int_{t_0}^{t_0+T} (ui)^2 dt} \quad (10)$$

where p is the instantaneous power; u and i are the instantaneous voltage and current; t_0 is the time corresponding to the start of measurement; and T is the period of the alternating voltage

Let us consider the case when the insntantaneous voltage and current consist of a fundamental component and several higher harmonics. Since the spectrum of higher harmonics is infinite, we will limit the analysis to the following:

$$u = u_1 + u_3 + u_5 \quad (11)$$

Then the instantaneous power is defined as follows:

$$\begin{aligned} p = ui &= (u_1 + u_3 + u_5)(i_1 + i_3 + i_5) = \\ &= u_1 i_1 + u_3 i_3 + u_5 i_5 + u_1 i_3 + \\ &+ u_1 i_5 + u_3 i_1 + u_3 i_5 + u_5 i_1 + u_5 i_3. \end{aligned} \quad (12)$$

As noted in [18], the frequency of power harmonics is formed as the sum and difference of the voltage and current harmonic frequencies, in this case:

$$\begin{aligned} p_0 &= (u_1 i_1)_{1-1} + (u_3 i_3)_{3-3} + (u_5 i_5)_{5-5}; \\ p_2 &= (u_1 i_1)_{1+1} + (u_1 i_3)_{1-3} + (u_3 i_1)_{3-1} + \\ &+ (u_3 i_5)_{3-5} + (u_5 i_3)_{5-3}; \\ p_4 &= (u_1 i_3)_{1+3} + (u_3 i_1)_{3+1} + \\ &+ (u_1 i_5)_{1-5} + (u_5 i_1)_{5-1}; \\ p_6 &= (u_3 i_3)_{3+3} + (u_1 i_5)_{1+5} + (u_5 i_1)_{5+1}; \\ p_8 &= (u_3 i_5)_{3+5} + (u_5 i_3)_{5+3}; \\ p_{10} &= (u_5 i_5)_{5+5}. \end{aligned} \quad (13)$$

Thus, in this case, the root mean square value of power is formed by taking into account all components of the instantaneous power and the combinations of current and voltage frequencies:

$$P_{rms} = \sqrt{\frac{1}{T}\int_{t_0}^{t_0+T} (p_0 + p_2 + p_4 + p_6 + p_8 + p_{10})^2 dt} \quad .(14)$$

To study power distribution and the impact of higher harmonic currents, the well-known model [21] was adapted in Matlab/Simulink (Fig. 2). It consists of the following blocks: a synchronous machine with a nominal frequency of f_{nom} = 50 Hz, nominal line voltage U_{nom} = 400 V, nominal apparent power S_{nom} = 31.3 kVA, and rotational speed PRM = 1500 rpm (SM 31.3 kVA); the excitation voltage (Vtref(pu)) and the mechanical power of the synchronous generator Pm are set as constant values, obtained based on preliminary calculations using the Simulink "Load Flow" utility; an asynchronous machine with a nominal frequency of f_{nom} = 50 Hz, nominal line voltage U_{nom} = 400 V, nominal power P_{nom} = 15 kW, and rotational speed PRM = 1500 rpm (AM 15 kW), to which a torque input block (Torque (N·m) is connected; and an additional load block with P_{nom} = 3 kW (3 kW Load), A transformer block with a high-voltage side rated at U_{HV} = 10 kV and a low-voltage side rated at U_{LV} = 400 V, with a nominal apparent power of S_{nom} = 60 kVA (Transformer 10 kV/400 V, 60 kVA); an external grid block with a nominal voltage of U_{nom} = 10 kV and a nominal apparent power of S_{nom} = 10 MVA (10 kV 10 MVA); and an external grid load block with a nominal active power of P_{nom} = 5 MW (5 MW Load). The harmonic injection block (Harmonic I) introduces higher-order harmonics into the phases and consists of three current source blocks generating the 3rd, 5th, and 7th harmonics, along with a parallel-connected resistor (see Fig. 3). The system also includes current and voltage measurement blocks: at the synchronous machine branch (Three-Phase V-I SG), at the asynchronous machine branch (Three-Phase V-I AM), at the external grid branch (Three-Phase V-I TV), and a measurement unit for collecting the acquired data (Measurement Unit)

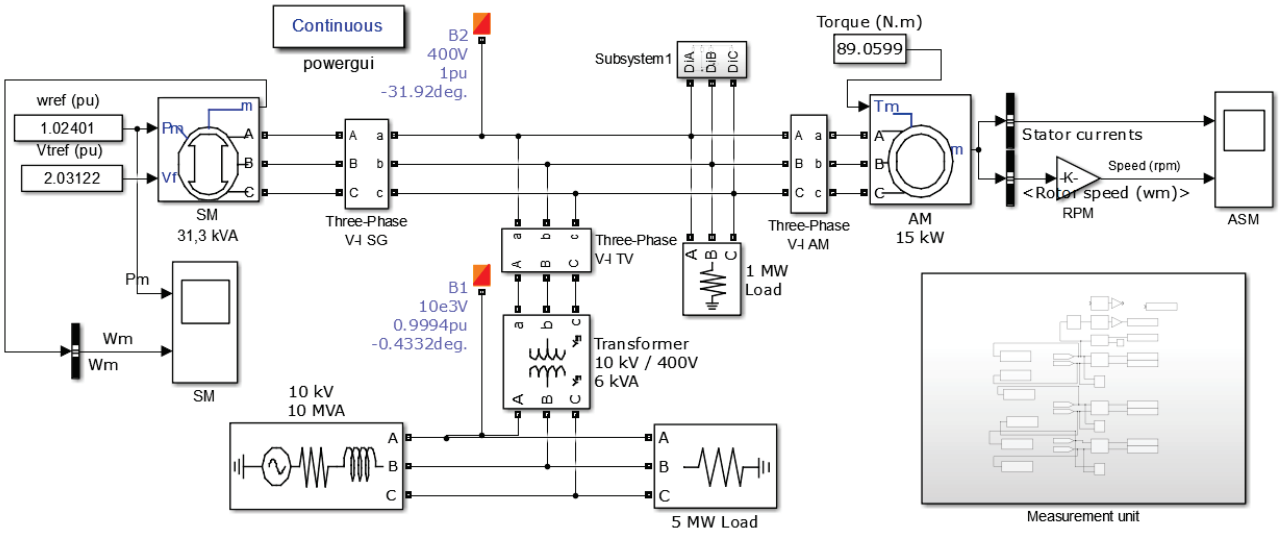


Fig. 2. Mathematical model of the diesel generator unit

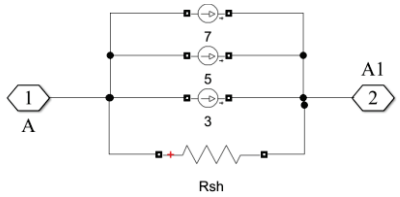


Fig. 3. Harmonic generation block of the nonlinear load current

To study the impact of the current harmonic level on the distribution of current distortion levels and power estimation across the branches of the analyzed system, the method of experimental design is applied.

Using the experimental design method [22], the conditions for analyzing the model are defined based on the indicators of current distortion in the model branches and the root mean square (RMS) phase power. The conditions, expressed as the magnitudes of the injected harmonic current components, are summarized in Table I, where the amplitude of the fundamental harmonic is 62 A for all experiments. The experimental design was conducted using the Statgraphics software package.

TABLE I. VALUES OF HIGHER HARMONIC CURRENTS FOR EXPERIMENTAL STUDIES

№Exp.	I_3, A	I_5, A	I_7, A
1	10	10	10
2	20	10	10
3	10	20	10
4	20	20	10
5	10	10	20
6	20	10	20
7	10	20	20
8	20	20	20
9	8,6	15	15
10	21,4	15	15
11	15	8,6	15
12	15	21,4	15
13	15	15	8,6
14	15	15	21,4
15	15	15	15
16	15	15	15

Figures 4–6 show the oscillograms of currents and voltages in the branches of the studied system for the case of current harmonics applied in experiment *Exp.2*.

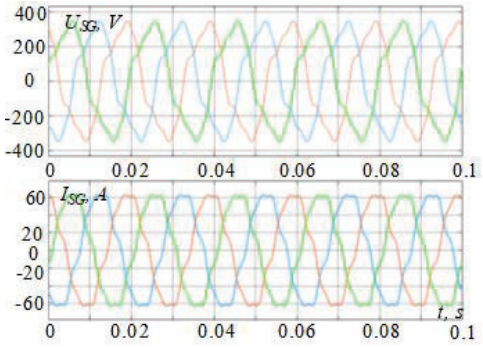


Fig. 4. Time-domain diagrams of currents and voltage in the synchronous generator branch

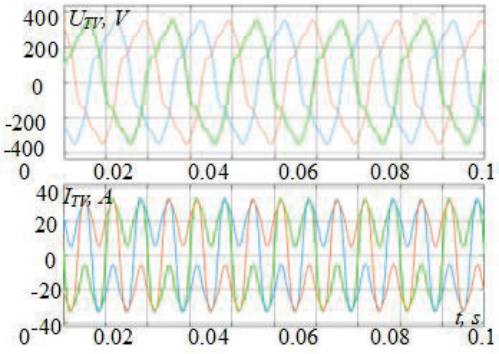


Fig. 5. Time-domain diagrams of currents and voltages in the transformer branch

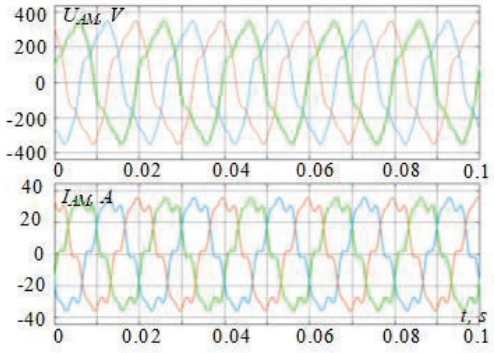


Fig. 6. Time-domain diagrams of currents and voltages in the asynchronous motor branch

Based on the data obtained from the experiments, regression model equations were calculated for each branch of the system using the Statgraphics software package. It is important to note that the currents are considered in per-unit values. $I'' = I_h / I_{fund}$.

For the total harmonic distortion (THD) of the current in the synchronous machine branch, the model shows that all three harmonics—the third, fifth, and seventh—have a positive linear contribution to the formation of harmonic current distortion in the synchronous machine. The influences of the fifth and seventh harmonics are the most significant, while the third harmonic also exhibits an additional nonlinear effect.

$$THD_{IS.G} = -0,0117661 + 0,00309106 \cdot I_3'' + \\ +0,00343581 \cdot I_5'' + \\ +0,00343581 \cdot I_7'' - 0,000103035 \cdot I_3''^2 \quad (15)$$

For the root mean square (RMS) phase power of the synchronous machine branch, the model indicates that the largest contribution to the increase comes from the third harmonic, both in linear and quadratic forms. The influence of the fifth and seventh harmonics is more complex: they linearly reduce the value but exhibit some compensating effects through quadratic and interaction (pairwise) terms.

$$PRMS_{ph.SG} = 7392,78 + 18,4613 \cdot I_3'' - \\ -0,793556 \cdot I_5'' - 0,890368 \cdot I_7'' + \\ +0,204227 \cdot I_3''^2 + 0,1288 \cdot I_3'' \cdot I_5'' - \\ -0,0212 \cdot I_3'' \cdot I_7'' + 0,0521191 \cdot I_5''^2 + 0,0254 \cdot I_5'' \cdot I_7'' \quad (16)$$

For the total harmonic distortion (THD) of the current phase in the transformer branch, the regression equation indicates that the third harmonic has the greatest impact on the overall harmonic distortion of the transformer current, while the fifth and seventh harmonics exert a significantly weaker linear influence. Interactions between the harmonics have a compensating effect, which slightly reduces the predicted THD value when multiple harmonics act simultaneously.

$$THD_{I.TV} = 0,0156275 + 0,0380441 \cdot I_3'' + \\ +0,00169802 \cdot I_5'' + 0,00169802 \cdot I_7'' - \\ -0,0002 \cdot I_3'' \cdot I_5'' - 0,0002 \cdot I_3'' \cdot I_7'' \quad (17)$$

For the root mean square (RMS) phase power of the transformer branch, the model demonstrates that the main contribution to the variation in RMS power is due to the third harmonic, whose effect is distinctly negative and nonlinear. The seventh harmonic has a minor negative effect.

$$PRMS_{ph.TV} = 2955,38 - 122,324 \cdot I_3'' - \\ -0,405468 \cdot I_7'' + 0,185911 \cdot I_3''^2 \quad (18)$$

For the total harmonic distortion (THD) of the current phase in the asynchronous machine branch, the model indicates that the third, fifth, and seventh harmonics have a positive influence on the overall harmonic current distortion. However, when high values of both the fifth and seventh harmonics are present simultaneously, their combined effect is partially compensated by a negative interaction term.

$$THD_{I.AM} = -0,0430238 + 0,00181077 \cdot I_3'' + \\ +0,0100405 \cdot I_5'' + 0,0100405 \cdot I_7'' - 0,0003 \cdot I_5'' \cdot I_7'' \quad (19)$$

For the root mean square (RMS) phase power of the asynchronous machine branch, the model accurately accounts for both the individual and interaction effects of the third, fifth, and seventh order harmonics on the RMS power of the asynchronous machine. The third harmonic has the greatest influence, both in linear and quadratic forms.

$$PRMS_{ph.AM} = 3952,42 + 23,1392 \cdot I_3'' + \\ +0,718265 \cdot I_5'' + 0,384424 \cdot I_7'' + \\ +0,012908 \cdot I_3''^2 - 0,2102 \cdot I_3'' \cdot I_5'' - \\ -0,0055 \cdot I_3'' \cdot I_7'' + 0,00215359 \cdot I_5''^2 - \\ -0,0302 \cdot I_5'' \cdot I_7'' - 0,0122101 \cdot I_7''^2 \quad (20)$$

Using the data obtained with the Statgraphics software package, the following calculations were performed: the coefficient of determination R^2, which shows the degree of dependence of the variable on the variation of independent variables, approaches 100% in all cases. The Pareto-adjusted statistic R^2_P, , which accounts for the number of independent variables, also has high values, confirming the stability of the models regardless of their complexity. The standard error of estimation $A_{e.e}$ which characterizes the deviation between actual and predicted values, is low for all models. The mean absolute error $AA_{e.e}$ – which also characterizes the deviation between actual and predicted values, is low for all models. The Durbin-Watson statistic *DWS* – used to detect autocorrelation of residuals, exceeds 5% in most cases and is below 5% in only one case; overall, the system exhibits only minor autocorrelation. The results of these calculations are summarized in Table II.

TABLE II. RESULTS OF THE PERFORMED CALCULATIONS

	R^2, %	R^2_P, %	$A_{e.e}$	$AA_{e.e}$	*DWS*, %
$THD_{IS.G}$,	99,57	98,94	0,0025	0,0013	<5
$PRMS_{ph.SG}$	99,99	99,99	0,7222	0,3580	>5
$THD_{I.TV}$	99,98	99,95	0,0033	0,0017	>5
$PRMS_{ph.TV}$	99,99	99,99	2,776	1,350	>5
$THD_{I.AM}$	99,86	99,65	0,0026	0,013	>5
$PRMS_{ph.AM}$	99,99	99,99	0,53	0,28	>5

CONCLUSIONS

On the synchronous generator branch, the third higher harmonic current has the least influence on total harmonic distortion, yet it has the greatest impact on the phase root mean square (RMS) power, although the actual change in power is small. The influence of the fifth and seventh higher harmonics slightly increases; meanwhile, the phase RMS power slightly increases with the fifth harmonic but slightly decreases with the seventh harmonic. The correlation between consecutive values is less than 5.0%, indicating possible serial correlation at a 5.0% significance level, while there is no evidence of sequential autocorrelation in the residuals of the phase RMS power at the 5.0% significance level.

On the transformer branch, the third higher harmonic has the greatest impact. An increase in the current value of the third higher harmonic corresponds to an increase in distortion and a simultaneous decrease in the phase RMS power. The values of the fifth and seventh higher harmonic currents also increase, but no effect on power is observed. The correlation between consecutive values exceeds 5.0%, so there is no indication of sequential autocorrelation in the residuals, both for overall harmonic distortion and for the phase RMS power.

On the asynchronous machine branch, the influence of the third higher harmonic on total harmonic distortion does not change significantly, but its effect on phase RMS power is considerable. The fifth and seventh harmonics increase, but their influence on phase RMS power slightly decreases. The correlation between consecutive values is less than 5.0%, indicating possible serial correlation at the 5.0% significance level, while there is no evidence of sequential autocorrelation in the residuals of the phase RMS power at the 5.0% significance level.

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2025 IEEE 6th KhPI Week on Advanced Technology (KhPIWeek)

Physical Justification For The Use Of The Power Factor And Its Alternative - Active Loss Factor In A Single-Phase Networks

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***Abstract*—This paper provides a physical justification for the use of the power factor and its interpretation in the general case for single-phase electrical networks and loads. A mathematical proof is provided for the power factor formula under arbitrary voltage and current waveforms. An alternative metric — the Active Loss Factor — is proposed, along with its mathematical rationale and interpretation. Expressions are presented for calculating the integral values of both the power factor and the Active Loss Factor over an extended time period. An example of calculating the integral value of the Active Loss Factor over a 24-hour period is provided for a single-phase network with two types of loads.**

Keywords—power factor, energy efficiency, active loss factor, electric network

I. Introduction

Electric energy losses have been relevant since the advent of the first power grids and remain critical today [1], [2]. Numerous studies examine how the nature of consumption influences technical losses and power quality [3], [4]. IEEE Std 1459 also formalizes these relationships [5].

Improving energy efficiency and power quality requires a clear understanding of the specific causes of losses and concrete measures to mitigate them. Key characteristics that increase technical losses include reactive behavior, harmonic distortion (nonsinusoidal conditions), unbalance, and nonstationarity.

Accordingly, ensuring favorable modes of power transmission and consumption requires not only identifying these issues but also determining how to address them.

Aim of the research. Our goal is to

(i) provide a compact, physically grounded derivation of the general power-factor expression valid for arbitrary waveforms;

(ii) introduce the Active Loss Factor (AF) as an alternative, loss-centric metric that directly quantifies the copper-loss penalty; and

(iii) derive integral, time-aggregated formulas enabling fair comparison of nonstationary consumption regimes and illustrate the approach on a day-long example.

Contributions and novelty. We recast the power factor (PF) around a clear physical optimum, introduce AF—a loss-based metric—and derive time-aggregated (integral) formulations with evidence that make the assessment robust for real-world, nonstationary single-phase operation. Note: AF here is distinct from the "active power loss factor" used as a planning coefficient and from the market "Average Loss Factor (ALF)."

II. Background: Classical Definition of Power Factor

One of the most widely used approaches for evaluating the energy efficiency of equipment or operating modes in modern research is the use of the power factor [1-4],[6]. According to IEEE Std 1459 [5]

$$PF = P/S,$$

where P - active power, $S = U_{RMS} I_{RMS}$ - apparent power, U_{RMS} and I_{RMS} - RMS voltage and current. It includes an interpretive note that PF can be viewed as the ratio between the active energy actually delivered to the load and the maximum active energy that could be delivered under the same line I²R loss level. But, in general, IEEE Std 1459 focuses on measurement-oriented definitions and a taxonomy of power-factor–related indices (e.g., true PF, displacement PF, distortion factor) under sinusoidal and nonsinusoidal conditions. It does not provide detailed physical derivations, optimization-based interpretations, or explicit optimality conditions for PF; such aspects fall outside the scope of a measurement standard. In addition classical PF have next limitations:

Not additive / not time-robust. PF of combined loads is not a weighted average of individual PFs; likewise, simple time-averaging of PF over operating intervals is not meaningful.

Not loss-direct. Copper (I²R) losses scale with I^2_{RMS}, not with PF. Thus equal PF does not imply equal feeder losses, and PF can improve while losses increase.

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III. Power Factor Justification

Consider the evaluation of the efficiency of energy transfer processes in a separate phase in the general case, at non-sinusoidal voltage and non-sinusoidal current.

Fig. 1 shows the cross-section of the measurement M and the power distribution in a single-phase network. In this case, the loss of power is denoted as P_{LS}, and the consumed power as P_{LD}. The power transmission network is characterized by inductive L and active r resistances, which are equally distributed between the two conductors. The power source in the network is described by the EMF $e(t)$, and the measured load voltage across the cross section is $u(t)$. The load resistance $R(t)$ is generally nonlinear and nonstationary.

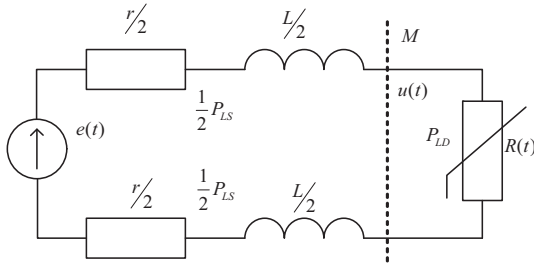


Fig. 1 Measurement of losses and power in a single-phase network

Energy efficiency assessment is performed by comparing the active power consumed by the current consumer with the power of some "standard" ideal consumer

$$K_{EFF} = \frac{P_{LD}}{P_{MAX}}, \tag{1}$$

where P_{MAX} represents the power consumed by the ideal consume, provided that the same level of power losses is maintained P_{LS}.

Fig. 2 shows the physical meaning of this comparison. Two systems $S1$ and $S2$ are compared, in which the power losses on the active impedance of the power supply system are equal and amount to $P_{LS1} = P_{LS2} = P_{LS}$. The condition is also the equality of the voltage on the load buses $u_{LD1}(t) = u_{LD2}(t) = u_{LD}(t)$. At the same time, the voltage or the electromotive force (EMF) in the systems $e_{S1}(t)$ and $e_{S2}(t)$ may differ, as well as the inductive component of the network resistance L_{S1}, L_{S2}. Thus, the system $S2$ is designed so that the load power is maximized with equal transmission losses and identical voltage at the load buses. This obviates the need to account for unknown system parameters of the power system and the transmission system. For a complete calculation, the parameters that can be measured in the cross-section M are sufficient [3].

The conditions for achieving P_{MAX} can be determined in two steps. For any $u(t)$ and $i(t)$, one can perform a Fourier series expansion

$$u(t) = \sum_{k=0}^{\infty} U_{m,k} sin(k\omega t + \phi_k),$$
$$i(t) = \sum_{h=0}^{\infty} I_{m,h} sin(h\omega t + \psi_h), \tag{2}$$

where $U_{m,k}$, $I_{m,h}$, ϕ_k, ψ_h – are the amplitude values and phases of the relevant voltage and current harmonics, k and h – are integers. Given $U_{m,k}$ and ϕ_k are known, determine $I_{m,h}$ and ψ_h to obtain the extremum of the power function

$$P = \frac{\omega}{2\pi} \int_0^{\frac{\omega}{2\pi}} p(t)\,dt \text{, where} \tag{3}$$

$$p(t) = \sum_{k=1}^{\infty}\sum_{h=1}^{\infty}\left[U_{m,k} sin(k\omega t + \phi_k) I_{m,h} sin(h\omega t + \psi_h)\right] \tag{4}$$

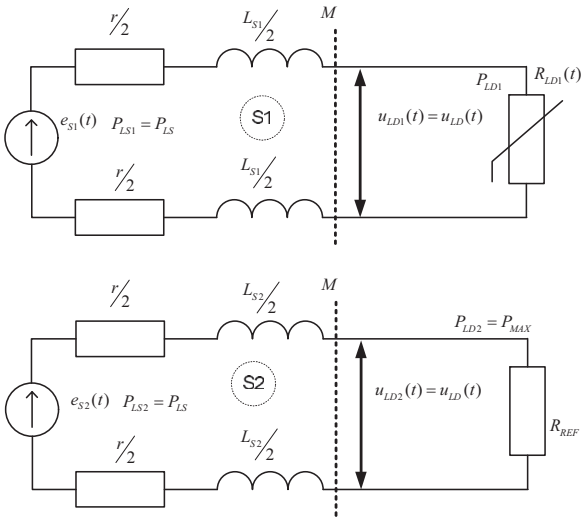


Fig. 2 Comparison of delivered load power under equal distribution-line I²R losses

To simplify further calculations, a zero constant component is assumed., $U_{m,0} = 0$ and $I_{m,0} = 0$.

The first step is to find the value of ψ_h, which corresponds to the extremum of function (3). By integrating

$$\int p(t)\,dt = \begin{cases} \sum_{k=1}^{\infty}\left[D_1 t - D_2\right], k = h \\ \sum_{h=1}^{\infty}\sum_{k=1}^{\infty}\left[D_3 - D_4\right], k \neq h \end{cases},$$

$$D_1 = \frac{I_{m,k} U_{m,k} cos(\psi_k - \phi_k)}{2},$$

$$D_2 = \frac{I_{m,k} U_{m,k} sin(\psi_k + \phi_k + 2\omega k t)}{4\omega k},$$

$$D_3 = \frac{I_{m,h} U_{m,k} sin(\psi_h - \phi_k + \omega(k-h)t)}{2\omega(k-h)},$$

$$D_4 = \frac{I_{m,h} U_{m,k} sin(\psi_h + \phi_k + \omega(k+h)t)}{2\omega(k+h)},$$

after setting the limits

$$\int_0^{2\frac{\pi}{\omega}} p(t)\,dt = \sum_{k=1}^{\infty}\left[\frac{I_{m,k} U_{m,k} \pi cos(\psi_k - \phi_k)}{\omega}\right], \tag{5}$$

with integers k and h

$$\sum_{k=1}^{\infty}\left[\frac{\left(I_{m,k}U_{m,k}cos\left(\psi_k-\phi_k\right)\right)}{2}t-\frac{\left(I_{m,k}U_{m,k}sin\left(\psi_k+\phi_k+2\omega kt\right)\right)}{4\omega k}\right]\Bigg|_0^{2\frac{\pi}{\omega}}=$$

$$=\sum_{k=1}^{\infty}\left[\frac{I_{m,k}U_{m,k}\pi cos\left(\psi_k-\phi_k\right)}{\omega}+\frac{I_{m,k}U_{m,k}\left(sin\left(\psi_k+\phi_k\right)-sin\left(\psi_k+\phi_k+4\pi k\right)\right)}{4\omega k}\right]=$$

$$=\sum_{k=1}^{\infty}\left[\frac{I_{m,k}U_{m,k}\pi cos\left(\psi_k-\phi_k\right)}{\omega}\right]$$

and

$$\sum_{h=1}^{\infty}\sum_{k=1}^{\infty}\left[D_3-D_4\right]\Bigg|_0^{2\frac{\pi}{\omega}}=\sum_{h=1}^{\infty}\sum_{k=1}^{\infty}\left[D_5+D_6\right]=0\,, \tag{6}$$

$$D_5=\frac{I_{m,h}U_{m,k}\left(sin\left(\psi_h-\phi_k\right)-sin\left(\psi_h-\phi_k+2\pi\left(h-k\right)\right)\right)}{2\omega\left(k-h\right)},$$

$$D_6=\frac{I_{m,h}U_{m,k}\left(sin\left(\psi_h+\phi_k\right)-sin\left(\psi_h+\phi_k+2\pi\left(h+k\right)\right)\right)}{2\omega\left(k+h\right)}.$$

To find the extremes, one need to differentiate the function for each of the desired variables ψ_k,

$$MAX\sum_{k=0}^{\infty}\left[\frac{I_{m,k}U_{m,k}\pi cos\left(\psi_k-\phi_k\right)}{\omega}\right]=\sum_{k=0}^{\infty}MAX\,\frac{I_{m,k}U_{m,k}\pi cos\left(\psi_k-\phi_k\right)}{\omega}.$$

Then find the roots

$$\frac{\partial}{\partial\psi_k}\left[\frac{I_{m,k}U_{m,k}\pi cos\left(\psi_k-\phi_k\right)}{\omega}\right]=\frac{I_{m,k}U_{m,k}\pi sin\left(\phi_k-\psi_k\right)}{\omega}=0\,. \tag{7}$$

An infinite number of extrema occur at the points $\psi_k=\varphi_k\pm\pi n$, where n is an integer. Even $n=0,\pm2,\pm4$, ... and 0 correspond to the points of maximum, and odd $n=\pm1,\pm3,\pm5$, ... – correspond to the points of minimum. Thus, the maximum occurs at the points

$$\psi_k=\varphi_k\,, \tag{8}$$

when the phase of the current harmonic coincides with the phase of the same voltage harmonic, regardless of the amplitude of this harmonic.

The second stage is $I_{m,h}$. When the phases of the corresponding harmonics are coincident, the average power over the period of the fundamental harmonic from (3) is obtained:

$$P=\frac{\omega}{2\pi}\int_0^{\frac{\omega}{2\pi}}p\left(t\right)dt=\frac{\omega}{2\pi}\sum_{k=1}^{\infty}\left[\frac{I_{m,k}U_{m,k}\pi}{\omega}\right]=\sum_{k=0}^{\infty}\left[\frac{I_{m,k}U_{m,k}}{2}\right] \tag{9}$$

Next, the ratio of the amplitude values of the harmonic components of the current must be determined. Accordingly, this can be expressed as:

$$I_{RMS}=\frac{1}{\sqrt{2}}\sqrt{\sum_k I_{m,k}^2}\,, \tag{10}$$

$$P=\frac{1}{2}\sum_k\left(U_{m,k}I_{m,k}\right),\ k=1,2,3,... \tag{11}$$

Expressing the amplitudes of higher harmonic components via the amplitude of the fundamental harmonic yields:

$$I_{m,k}=K_kI_{m,1}, \tag{12}$$

$k=1,2,3,\ldots$, the coefficient corresponding to the first harmonic. $K_1=1$.

For an arbitrary set of harmonics, it is possible to write:

$$2I_{RMS}^2=\sum_k K_k^2I_{m,1}^2 \tag{13}$$

$$P=\frac{1}{2}\sum_k\left(U_{m,k}K_k\sqrt{\frac{2I_{RMS}^2}{\sum_n K_n^2}}\right), \tag{14}$$

where $n=1,2,3,...$

Differentiating by K_k results in the following system of equations:

$$\frac{\partial P}{\partial K_k}=\frac{U_{m,k}}{2}\sqrt{\frac{2I_{RMS}^2}{\sum_n K_n^2}}-\frac{K_kI_{RMS}^2\sum_n\left[K_nU_{m,n}\right]}{\sum_n K_n^2}\sqrt{\frac{\sum_n K_n^2}{2I_{RMS}^2}}\,, \tag{15}$$

and equating the right-hand parts to (15):

$$\frac{\partial P}{\partial K_k}=I_{RMS}^2\frac{\left(U_{m,k}\sum_n K_n^2-K_k\sum_n\left[K_nU_{m,n}\right]\right)}{\sum_n K_n^2}\sqrt{\frac{\sum_n K_n^2}{2I_{RMS}^2}}=0 \tag{16}$$

The solution looks as:

$$K_k=\frac{U_{m,k}\sum_{n,n\neq k}K_n^2}{\sum_{n,n\neq k}\left[K_nU_{m,n}\right]} \tag{17}$$

and the solution of the system will be:

$$K_k=\frac{U_{m,k}}{U_{m,1}}, \tag{18}$$

Substituting yields:

$$I_{m,k}=\frac{U_{m,k}}{U_{m,1}}I_{m,1}\,, \tag{19}$$

this implies that the amplitudes of current harmonics should be proportional to the voltage amplitudes of the same harmonics, which is consistent with the conclusions made in [3]. Substituting the results of (8) and (19) into (3) and (4), yields the following result:

$$P_{MAX}=U_{RMS}I_{RMS}\,, \tag{20}$$

where U_{RMS} and I_{RMS} – are the root-mean-square (RMS) (effective values of voltage and current).

This implies that:

$$U_{RMS}=\sqrt{\frac{\omega}{2\pi}\int_0^{\frac{2\pi}{\omega}}u^2(t)dt}\,, \tag{21}$$

$$I_{RMS}=\sqrt{\frac{\omega}{2\pi}\int_0^{\frac{2\pi}{\omega}}i^2(t)dt}\,, \tag{22}$$

accordingly:

$$K_{EFF}=\frac{P_{LD}}{U_{RMS}I_{RMS}}=PF\,. \tag{23}$$

Therefore, the resulting expression matches the general definition of the power factor in the general case. In other words, the power factor reflects how many times the load power is less than the maximum possible power at the same voltage and the same level of losses in the active resistance of the current conductor.

It should be noted that the considered model is the simplest and takes into account only losses in the active resistance of current conductors. Other types of technical losses and factors that affect them are not taken into account. In some works, modified formulas are considered, which, for example, take into account the so-called skin effect. In addition, the list of technical loss factors can also include losses in magnetically conductive elements, losses in parasitic capacitance currents of power transmission lines, corona losses, electromagnetic radiation, etc. However, all these factors are considered secondary. First, the real amount of electricity losses caused by them is much smaller compared to losses in the active resistance of current conductors. Secondly, such factors largely depend on the specific configuration of the network, the equipment used and the connection point. Attempting to take them into account requires a large amount of additional information, significantly complicates the calculations, and, at the same time, yields no significant practical advantage. The standard method for calculating the power factor can be carried out at any cross-section with the possibility of measuring current and voltage.

Thus, the advantages of such an approach include its versatility and ease of implementation. The necessary mathematical apparatus can be implemented on the basis of almost any microprocessor device, both in the format of special measuring equipment and in the structure of a standard electricity meter. The measured parameters coincide with the parameters required for a standard metering system. Eliminating the necessity to build any complex system covering several points of the power system, and information about the scheme and parameters of the electrical network and the equipment connected to it is also not required.

The main disadvantage of the approach is that the measured power factor, as can be understood from Fig. 2, depends not only on the load characteristics but also on the impedance of the power system. Thus, in most cases, the system resistance is 2-3 orders of magnitude lower than the load resistance, and, accordingly, this factor can be neglected. However, in some cases it can significantly affect the conclusions. In addition, as previously noted above, the approach does not fully account for actual technical losses.

IV. Active Loss Factor

An alternative approach also follows from the formulation of the problem. The second option is to compare the magnitude of the power losses and obtain the loss coefficient

$$K_L = \frac{P_{LS}}{P_{LS,MIN}}, \tag{24}$$

where $P_{LS,MIN}$ – is the minimum value of losses, for a load with equal power consumption. Fig. 3 shows an explanation of the physical meaning. Also compared are two systems $S1$ and $S2$, in which the load power coincides $P_{LD1} = P_{LD2} = P_{LD}$, as does the voltage on its buses $u_{LD1}(t) = u_{LD2}(t) = u_{LD}(t)$, but the power losses on the active resistance r of the power supply system $S2$ are minimal, and are equal to $P_{LS2} = P_{LS,MIN}$.

$$K_L = \frac{rI_{RMS}^2}{rI_{RMS,MIN}^2} = \frac{I_{RMS}^2}{I_{RMS,MIN}^2} \tag{25}$$

$$I_{RMS,MIN} = \frac{P_{LD}}{U_{RMS}} \tag{26}$$

$$K_L = \frac{I_{RMS}^2 U_{RMS}^2}{P_{LD}^2} = \frac{1}{PF^2} \tag{27}$$

Thus, the loss ratio is inversely proportional to the square of the power factor. For ease of use, it is proposed to call the obtained value the active loss ratio and denote it as AF:

$$AF = \frac{I_{RMS}^2 U_{RMS}^2}{P^2} = \frac{1}{PF^2} \tag{28}$$

This is due to the fact that the coefficient LF in modern scientific works refers to either the load factor [7] or the loss factor, which indicates the percentage of energy losses in the electrical network [8].

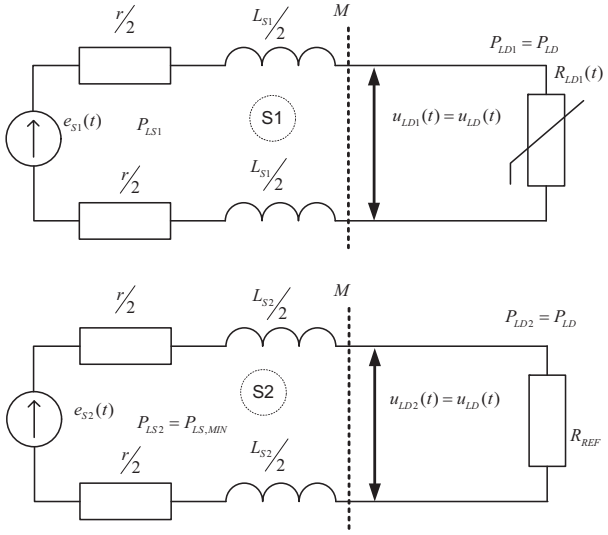


Fig. 3 Comparison of the magnitude of losses on the active resistance of the transmission system

It should be noted that the power factor is currently widely used in scientific works [1-4,6,9-15], engineering calculations [16,17] and even standards [5]. But a slightly different interpretation of this coefficient is generally accepted, in particular the physical meaning. It is calculated, most often, as the ratio of consumed power to full (or apparent) power, the latter being measured in volt-amperes and not considered active. There is also without physical justification or stated limitations [5].

The physical meaning of the Active Loss Factor seems to be more indicative and informative in comparison with the power factor. Thus, AF indicates how many times higher the I^2R losses in the line's series resistance are than the theoretical minimum for the same load power and voltage. In addition, the Active Loss Factor is more convenient when calculating the integral assessment of the energy efficiency of electricity consumption.

The proposed approach, in conjunction with the Active Loss Factor, enables an integral assessment of energy efficiency over a given time interval. Under the assumption of constant power, the following expression can be formulated:

$$K_{L,T} = \frac{W_{LS}}{W_{LS,MIN}} = \frac{P_{LS} \cdot T_F}{P_{LS,MIN} \cdot T_F} = \frac{rI_{RMS}^2 \cdot T_F}{rI_{RMS,MIN}^2 \cdot T_F} = AF_T, \quad (29)$$

where W_{LS} – is the energy loss in the time interval T_F, and $W_{LS,MIN}$ – is the minimum energy loss in the same interval. Given that:

$$AF = \frac{I_{RMS}^2 U_{RMS}^2}{P^2} = \frac{I_{RMS}^2}{I_{RMS,MIN}^2}, \quad (30)$$

from

$$I_{RMS,MIN}^2 = \frac{I_{RMS}^2}{AF}. \quad (31)$$

than

$$W_{LS,MIN} = rI_{RMS,MIN}^2 \cdot T_F = r\frac{I_{RMS}^2}{AF} \cdot T_F \quad (32)$$

Under variable load conditions, the time domain is discretized into intervals wherein power consumption is assumed to remain constant $T_F = T_1 + T_2 + ... + T_N$. This facilitates the derivation of the following expression:

$$AF_T = \frac{\sum\limits_{i=1..N} W_{LS,i}}{\sum\limits_{i=1..N} W_{LS,MIN,i}} = \frac{\sum\limits_{i=1..N} rT_i I_{RMS,i}^2}{\sum\limits_{i=1..N} \frac{rT_i I_{RMS,i}^2}{AF_i}} = \frac{\sum\limits_{i=1..N} T_i I_{RMS,i}^2}{\sum\limits_{i=1..N} \frac{T_i I_{RMS,i}^2}{AF_i}} \quad (33)$$

But if expressing the value from (28) the value of I_{RMS}^2

$$I_{RMS}^2 = \frac{AF \cdot P^2}{U_{RMS}^2} \quad (34)$$

and substitute it into (33), yields:

$$AF_T = \frac{\sum\limits_{i=1..N} T_i I_{RMS,i}^2}{\sum\limits_{i=1..N} \frac{T_i I_{RMS,i}^2}{AF_i}} = \frac{\sum\limits_{i=1..N} T_i \frac{AF_i \cdot P_i^2}{U_{RMS,i}^2}}{\sum\limits_{i=1..N} \frac{T_i \frac{AF_i \cdot P_i^2}{U_{RMS,i}^2}}{AF_i}}, \quad (35)$$

this implies that, $U_{RMS,i}$ will act as a "weighting" coefficient. When it changes, the influence of a separate interval will change, although the voltage value is not related to the properties of the consumer. Therefore, in order for the assessment of energy efficiency of consumption to be independent of voltage fluctuations in the network, it is possible to take $U_{RMS,i} = const = U_{RMS}$, and, accordingly:

$$AF_T = \frac{\sum\limits_{i=1..N} T_i \frac{AF_i \cdot P_i^2}{U_{RMS}^2}}{\sum\limits_{i=1..N} \frac{T_i \frac{AF_i \cdot P_i^2}{U_{RMS}^2}}{AF_i}} = \frac{\sum\limits_{i=1..N} AF_i \cdot T_i \cdot P_i^2}{\sum\limits_{i=1..N} T_i \cdot P_i^2} \quad (36)$$

The minimal interval can be chosen to correspond to that corresponds to the network voltage period. In this case, the need to isolate periods of constant load is eliminated, and the approach becomes entirely generalizable.

V. Calculation Example

A consumer is powered by a network with an active resistance $r = 0{,}1$ Ohm. In this case, the effective voltage at the measurement point is constant and, regardless of the load, is $U_{RMS} = 230$ V. The consumer has two electrical appliances - an LED lighting lamp with a power of $P_{LED} = 10$ W and a power factor of $PF_{LED} = 0{,}5$, and a heater with a power of $P_{HT} = 1000$ W and a power factor of $PF_{HT} = 0{,}99$. It is necessary to calculate the total Active Loss Factor and power factor for a day, if the lighting lamp is turned on constantly, and the heater is turned on for 2 hours.

Solution. The entire time interval of 24 hours can be divided into 2 sections - $T_1 = 22$ hours when only LED lighting is working, and $T_2 = 2$ hours when the heater and LED lighting are working.

In the first section, based on (23), one can calculate the effective current value:

$$I_{RMS,1} = \frac{P_{LED}}{U_{RMS} PF_{LED}} = \frac{10}{230 \cdot 0{,}5} = 0{,}087 \text{ A}.$$

Accordingly, the Active Loss Factor in the first section:

$$AF_1 = \frac{1}{PF_{LED}^2} = \frac{1}{0{,}5^2} = 4.$$

In the second section, the total current equals the sum of individual branch currents from the lighting and the heater:

$$I_{RMS,2} == \frac{P_{LED}}{U_{RMS} PF_{LED}} + \frac{P_{HT}}{U_{RMS} PF_{HT}},$$

$$I_{RMS,2} = \frac{10}{230 \cdot 0{,}5} + \frac{1000}{230 \cdot 0{,}99} = 4{,}479 \text{ A}.$$

Accordingly, the Active Loss Factor and the power factor in the second section:

$$AF_2 = \frac{I_{RMS,2}^2 U_{RMS}^2}{P_2^2} = \frac{4{,}479^2 \cdot 230^2}{(1000+10)^2} = 1{,}040,$$

$$PF_2 = \frac{1}{\sqrt{AF_2}} = \frac{1}{\sqrt{1{,}040}} = 0{,}981.$$

According to (33), the calculation of the overall active loss and power factors can be evaluated using the following expressions:

$$AF_T = \frac{\sum\limits_{i=1..N} T_i I_{RMS,i}^2}{\sum\limits_{i=1..N} \frac{T_i I_{RMS,i}^2}{AF_i}},$$

substituting the obtained values:

$$AF_T = \frac{22 \cdot 0{,}087^2 + 2 \cdot 4{,}479^2}{\frac{22 \cdot 0{,}087^2}{4} + \frac{2 \cdot 4{,}479^2}{1{,}040}} = 1{,}043,$$

$$PF_T = \frac{1}{\sqrt{AF_T}} = \frac{1}{\sqrt{1{,}043}} = 0{,}979.$$

To validate the accuracy of the obtained energy efficiency assessment, the loss ratio should be calculated knowing the active resistance of the network.

Electrical energy losses can be calculated as:

$$W_{LS} = W_{LS,1} + W_{LS,2} = r \cdot T_1 \cdot I_{RMS,1}^2 + r \cdot T_2 \cdot I_{RMS,2}^2,$$

$$W_{LS} = 0,1 \cdot 22 \cdot 0,087^2 + 0,1 \cdot 2 \cdot 4,479^2 = 4,028 \text{ Wh}$$

Minimum electrical energy losses would occur if both loads had a power factor of 1. Then:

$$I_{RMS,MIN,1} = \frac{P_{LED}}{U_{RMS} PF_{MAX}^2} = \frac{10}{230 \cdot 1^2} = 0,043 \text{ A},$$

$$I_{RMS,MIN,2} = \frac{10}{230 \cdot 1^2} + \frac{1000}{230 \cdot 1^2} = 4,391 \text{ A},$$

$$W_{LS,MIN} = W_{LS,MIN,1} + W_{LS,MIN,2} = r \cdot T_1 \cdot I_{RMS,MIN,1}^2 + r \cdot T_2 \cdot I_{RMS,MIN,2}^2$$

$$W_{LS,MIN} = 0,1 \cdot 22 \cdot 0,043^2 + 0,1 \cdot 2 \cdot 4,391^2 = 3,861 \text{ Wh},$$

$$\frac{W_{LS}}{W_{LS,MIN}} = \frac{4,028}{3,861} = 1,043 .$$

Thus, the integral coefficient corresponds to the ratio of losses according to the considered model.

For comparison, simplified approaches to assessing energy efficiency can be used. For example, if the power factor is averaged over time, the result will be completely different:

$$PF_T = \frac{PF_1 \cdot T_1 + PF_2 \cdot T_2}{T_1 + T_2} = \frac{0,5 \cdot 22 + 0,981 \cdot 2}{22 + 2} = 0,540$$

This clearly indicates that such a value does not reflect the real picture and properties of the load.

Another possible way is to calculate the Active Loss Factor for the entire 24-hour period. It can be shown that the root-mean-square current per day will be equal to:

$$I_{RMS,T} = \sqrt{\frac{I_{RMS,1}^2 \cdot T_1 + I_{RMS,2}^2 \cdot T_2}{T_1 + T_2}} = \sqrt{\frac{0,087^2 \cdot 22 + 4,479^2 \cdot 2}{22 + 2}} = 1,296$$

A, and the average power:

$$P_{AVG} = \frac{P_1 \cdot T_1 + P_2 \cdot T_2}{T_1 + T_2} = \frac{10 \cdot 22 + 1010 \cdot 2}{22 + 2} = 92,5 \text{ W}.$$

Based on the data obtained, the power factor will be equal to:

$$PF_T = \frac{P_{AVG}}{U_{RMS} I_{RMS,T}} = \frac{92,5}{230 \cdot 1,296} = 0,310 ,$$

which does not adequately represent the actual characteristics of the load. This discrepancy arises because non-stationarity also affects the value of PF_T. But, in our opinion, the requirement of complete stationarity of consumption is unrealistic and inconsistent with actual behavior of consumers. Therefore, such an estimate has no practical meaning, despite being mathematically valid.

CONCLUSION

(i) The paper offers a physically grounded interpretation of the power factor (PF), clarifying the operating conditions under which it attains its best value and explaining why practical measurements may deviate.

(ii) The Active Loss Factor (AF) is a more practical, loss-centric indicator of consumption efficiency: it directly reflects the copper-loss penalty and remains meaningful under nonsinusoidal and nonstationary operation.

(iii) The proposed time-aggregated methodology enables robust, comparable assessments for arbitrary loads using only standard metering data (RMS voltage, RMS current, active power), without specialized instrumentation.

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Analysis of Options for Constructing and Using Models of the Mechanical Part of the Electric Drive of an Armored Face Conveyor as a System with Distributed Parameters

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***Abstract* – The article discusses the issues of constructing computer models of an electric drive with distributed mass and elasticity of the mechanical part using the example of armored face conveyors. Mathematical models were obtained. Various methods of description were used: frequency approximation, block diagrams, vector-matrix apparatus, graphs method. The results were compared, including areas of expedient application in solving typical problems of calculating dynamic processes of a conveyor in characteristic technological modes of its operation: soft start from an intensity setter and in current limiting mode, sudden braking from emergency chain jamming, and others. The wave processes of displacements, velocities and accelerations in a 200-mass elastic-viscous chain model coupled with a model of a cascade speed control system are shown in 3-D representation. An appropriate range of values of discrete elements along the length of the discretized chain is determined. New features in dynamic processes were discovered, new analytical dependencies were obtained.**

Keywords – electromechanical system; armored face conveyor; multi-mass drive system; speed control system; dynamic load.

I. Introduction

As industrial electric drives improve in the direction of increasing accuracy and speed and the development of synthesis algorithms, more and more attention has to be paid to clarifying the description of the mechanical part, at least to check the performance of synthesized controllers. The problem of taking into account the so-called unmodeled dynamics in the synthesis of automatic control systems has not yet been posed on a large scale or is being solved using robust controllers [1]–[5].

Methods for describing mechanical systems, developed in the theory of mechanical vibrations, are mainly focused on another task: determining the frequencies and shapes of vibrations, and, secondly, they require an inappropriately large amount of time from an electrical specialist to master them and use them professionally, taking into account the need for coupling models of the mechanical part with a model of the electrical part of the drive and features of the technological process. Similar remarks can be made in relation to the material in a small section of the theory of automatic control devoted to systems with distribution parameters [6]–[8]. The mathematical apparatus of this section turned out to be practically unclaimed by electric drives in their scientific articles, and there is, most likely, the only monograph on the electric drive of mechanisms with distributed mass and rigidity [9].

At the same time, the well-known method of discretizing an object with distributed mass and rigidity on the basis of physical concepts requires knowledge only of Newton's law and Hooke's law, and therefore is accessible to schoolchildren. Therefore, the question of the feasibility and effectiveness of using one or another method for describing distributed systems when solving specific electrical drive problems requires clarification [10], [11]. Let's consider it using examples of the dynamics of a conveyor chain, keeping in mind that modeling the rod itself is not the final goal when building a model of the mechanical part of these electric drives [12], [13].

II. Modeling the Mechanical Part of a Conveyor with an Elastic-Viscous Chain without a Sagging Loop

In the absence of a sag loop, the chain has sufficient tension to consider it as a rod, the ends of which are brought together into a ring and rigidly fixed to a concentrated sufficiently large mass corresponding to the inertia of the engine rotor. If the mass of the rotor were infinitely large, then such a scheme would correspond to the termination of a straight rod at both ends, however, such a simple solution, especially taking into account the presence of a control system, cannot be accepted for consideration as a complete model. In the theory of oscillations [14], [15] a number of simple standard design schemes are considered, but for them, mainly, the natural frequencies and mode shapes are obtained and, in addition, there is no solution to the partial differential dynamics equations for the considered conveyor scheme.

In publication [16], the oscillation frequencies are also determined for the conveyor, but there is no solution to the dynamics equations. The mentioned calculation formulas for oscillation frequencies are, naturally, very convenient to use to check the correct operation of the conveyor model when modifying its design scheme to the diagrams [17], [18]. This allows you to evaluate the efficiency of the model set in Matlab element by element - for example, separately for a conveyor chain or when specifying the moment of inertia of the rotor $J \to \infty$.

Only in a few (unfortunately) works are there solutions to the equations of rod dynamics that are available for direct use by an electrical engineer. Thus, in [19] a calculation formula is presented for determining the displacements $\eta(x, t)$ of points along the length x of a homogeneous massive elastic rod with single-end embedding during longitudinal vibrations without taking into account damping.

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$$\eta(x,t)=\frac{2P}{lEF}\sum_{n=1}^{\infty}\frac{(-1)k^{n-1}\cos(\lambda_n t)\sin(k_n x)}{k_n^2}, \qquad (1)$$

where P – is the tensile force of the rod along the length of the conveyor chain, N;

l – is the conveyor chain length, m;

E – is the Young's modulus of rod material, N/m^2;

F – is the cross-sectional area of the rod, m^2;

n – is the number of members of the series;

λ_n – is the rod longitudinal vibration frequency;

k_n – is the wave number.

The solution, as can be seen, was obtained in the form of a sum of a series of harmonics, that is, it is approximate, the quality of the approximation of which depends on the value n used - the number of terms of the series.

A graphical representation of the solution $\eta(x, t)$ with the number of harmonics n=1250, performed by the authors of the article in the Maple package, is shown in Fig. 1.

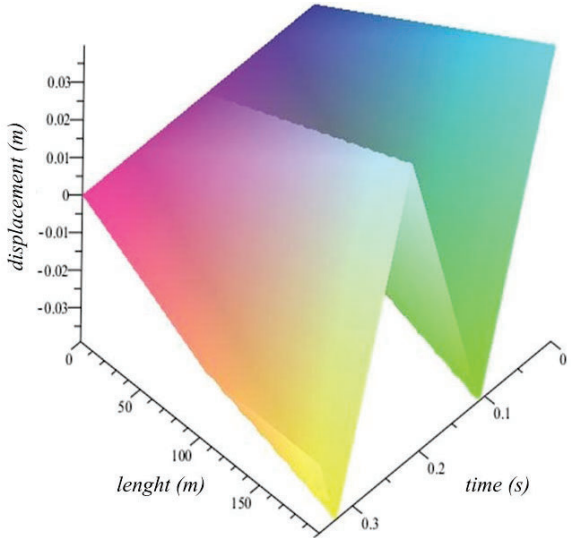


Fig. 1. Displacement, in meter, of conveyor chain points $\eta(x, t)$ along its length, in meter, and in time, sec

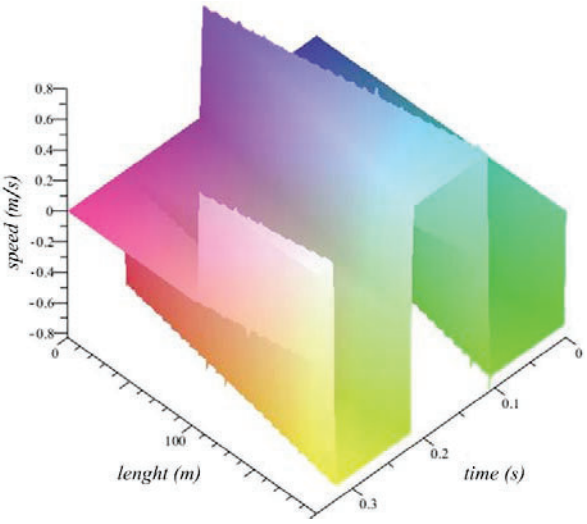


Fig. 2. Velocities of conveyor chain points $v(x, t)$ along its length and in time

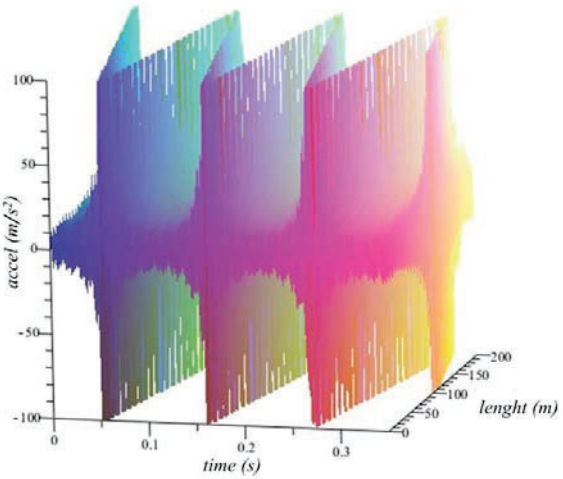


Fig. 3. Acceleration of conveyor chain points $a(x, t)$ along its length and in time

Formula (1) $\eta(x, t)$ in the same package was analytically differentiated (after correcting typos in the text of the textbook) to obtain graphs of velocities and accelerations. They are shown in Fig. 2, Fig. 3.

From Fig. 1 and Fig. 2 it is clear that the solutions correctly display the physical processes of the passage of a wave along the rod, but the shape of the acceleration graphs, and therefore forces, is no longer acceptable for consideration at a graphic resolution of 150x150, and at a resolution of 500x500 image processing on the screen of a good computer occurs with such a delay that it becomes almost impossible to turn it to the desired angle.

The possibility of applying formula (1) is exhausted here. It cannot be used to plot dynamic processes in another design scheme, for example, when embedding a rod on both sides. Even to calculate the considered scheme, but with damping in the rod, you need to obtain a different calculated dependence. Last presented Here same in [19] in form (2).

$$\eta(x,t)=e^{-ht}\sum(A_n\cos\lambda_n^* t+B_n\sin\lambda_n^* t)\sin\frac{(2n-1)}{2l}\pi x, \qquad (2)$$

$$\lambda_n^*=\sqrt{\lambda_n^2-h^2}, \qquad (3)$$

$$\eta(x,0)=\frac{Px}{EF}, \quad \eta(x,0)=0, \qquad (4)$$

$$A_n=\frac{P}{EF}\frac{\int_0^l xX_n(x)dx}{\int_0^l X_n^2(x)dx}, \qquad (5)$$

$$B_n=\frac{Ph}{EF\lambda_n^*}\frac{\int_0^l xX_n(x)dx}{\int_0^l X_n^2(x)dx}. \qquad (6)$$

However, it is no longer possible to carry out calculations without finding the values of the coefficients A_n and B_n under specific boundary conditions. In addition, it is not clear why the exponent e^{-ht}, corresponding to the damping of the process, turned out to be a multiplier beyond the sum of the harmonics and, thus, the free damping of movements at different frequencies becomes the same tempo for both low and high frequencies. That is, fast ones turn out to be worse damped. This is not physical and, moreover, it is not clear that in [15]

another calculation formula (7) is given for solving the same problem.

$$u = \sum_{r=1}^{\infty} X_r(x) e^{-n_r t} (A_r \sin p_r^* t + B_r \cos p_r^* t), \quad (7)$$

$$p_r^* = \sqrt{p_r^2 - n_r^2} = p_r \sqrt{1 - \left(\frac{k p_r}{2E}\right)^2}, \quad (8)$$

$$n_r = \frac{k p_r^2}{2E}. \quad (9)$$

In it, each frequency of damped oscillations has its own exponential and, moreover, it is noted that due to damping, the upper frequencies in the record of the series cease to exist and "... the corresponding terms of the general solution will represent movements of a non-oscillatory nature." This corresponds to general physical concepts about the nature of damping processes in structural materials, however, formulas for calculating coefficients A and B are not given.

To summarize this brief selective but representative analysis, it should be noted that in the considered basic sources, unfortunately, there are no completed calculation formulas and, moreover, their summary tables for determining the forces, accelerations, velocities and displacements of different sections of the elastic-viscous massive rod with different design schemes in terms of boundary conditions and the presence of connecting masses of different sizes at different points of the rod. This does not allow the electrical engineer to take advantage of the indicated results of applying the methods of the theory of vibrations of mechanical systems.

Without analyzing a number of other methods of the theory of mechanical vibrations, including the finite element method (which gives the desired results, but requires significant labor to master and time to use), as well as difference methods of mathematical physics, let us move on to considering the method of direct discretization of a rod into elastically coupled masses. This method is extremely simple in its physical interpretation and mathematical apparatus, requiring the use of only Newton's law and Hooke's law when drawing up the differential equations of the model.

The direct sampling method was used to construct a model of the electric drive of a scraper conveyor and study the dynamic processes in it [20]–[22] in normal and emergency operating modes. The model was represented in the Matlab package by a structural diagram of a 200-mass elastic-viscous chain with a model of an electric drive, a fluid coupling, a model of rock formation fall, nonlinear friction, etc. This model was checked for the coincidence of natural frequencies with the calculated values according to [9], [14], [15], [19]. However, on such a fairly complete model, the calculation of processes takes quite a long time and it is difficult to evaluate the dynamic properties of its linear part when changing the values of parameters and, especially, the number of masses, since this requires changing the structural diagram itself.

For analyzing the dynamic properties of the linear part of the conveyor model, a much more convenient model is the one presented in the m-file of the Matlab package in a matrix description. The calculation and display of its results occurs several times faster, so it is possible to track and analyze a larger amount of information. Thus, it was found that the circular and open-loop design schemes are characterized by a circular distribution of the roots of the characteristic equation of the conveyor model, as shown in Fig. 4.

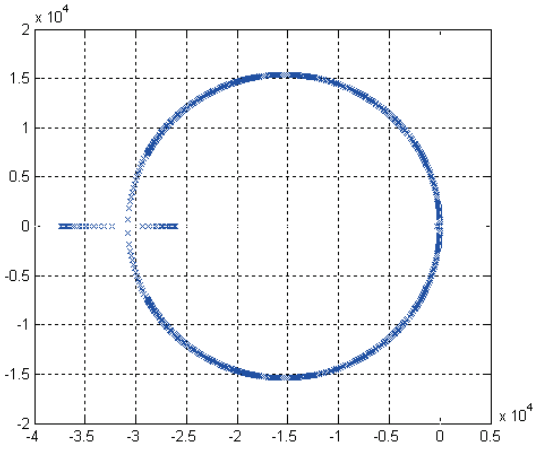


Fig . 4. Distribution of roots of the characteristic equation of the conveyor model

The roots are placed on a circle starting at the origin of the complex plane. First, with increasing frequency, the damping of each pair of roots increases, then the frequency begins to decrease and fast movements caused by remote left roots become non-oscillatory, as noted in [23]. During the variation of the chain parameters, it was found that the radius of this circle is inversely proportional to the coefficient of internal viscous friction β_{12} and directly proportional to the stiffness coefficient c_{12}, constant and equal to $R = c_{12}/\beta_{12}$.

It is advisable to emphasize that exactly this value of the radius of the circle for placing the roots can be obtained through a series of transformations of the solution to the partial differential equation [23]:

$$\frac{d^2u}{dt^2} = c^2 \frac{d^2u}{dx^2} + \frac{k}{\rho} \frac{d^3u}{dx^2 dt}. \quad (10)$$

Equation (10) describes the dynamics of an elastic-viscous rod using dependencies (7,8,9). Moreover, this solution is analytical; it was obtained by the Fourier method, and not by direct discretization of the rod. The roots, which are easy to obtain from the presented dependencies, coincide with the roots obtained for the model with discretized masses (n=50) for 10–20 lower frequencies, as can be seen from Fig. 5, and then begin to shift relative to each other, remaining, however, on a circle of radius $R=c_{12}/\beta_{12}$.

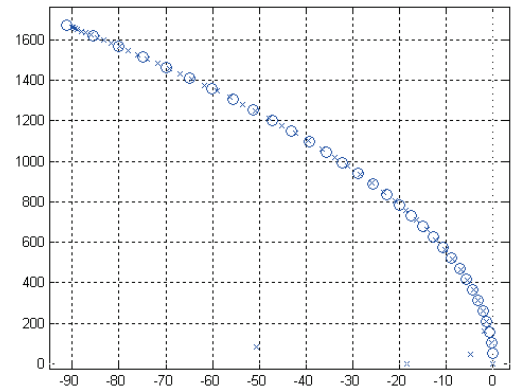


Fig . 5. Distribution of roots obtained for the model with discretized masses (n=50)

The same distribution of roots occurs for n=200, as can be seen from Fig. 6.

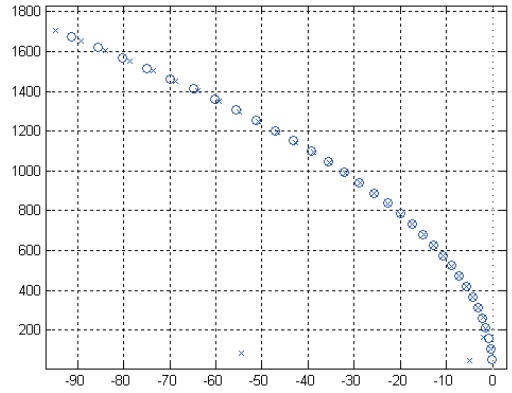


Fig . 6. Distribution of roots obtained for the model with discretized masses (n=200)

It is important to note that the roots in the dependencies [22] were obtained to describe the dynamics of a rod with a double-sided end seal, and the roots of the conveyor model were obtained on a model that included both a massive elastic-viscous tensioned chain and a drive using a cascade control system. These results make it possible to judge both the magnitude of the error caused by the discretization of masses and the extent of the influence of the drive control system on the wave processes of the conveyor chain.

Model calculations performed when varying the number of conveyor masses from 20÷30 to 200 for a two-drive conveyor with cascade control systems configured to a symmetrical optimum and an undamped or weakly damped chain model showed that at 60<n<120, several right-hand complex-type equations appear among the roots of the characteristic equation conjugate to the small real part. This suggests that such a system needs to be modeled with a mass number n>150. Then these roots return to the ordinate axis, otherwise erroneous conclusions can be drawn about the instability of the modeled object, and although for a two-drive conveyor it is not recommended for use, setting it to a symmetrical optimum due to increased oscillation caused by elastic coupling through the chain and possible fall into self-oscillatory mode when there is a difference tasks for speed, its frequency response has a less pronounced chain of resonant peaks, which interfere with the construction of simple controllers with a single-drive conveyor or when setting a two-drive conveyor to a betragsoptimum. Thus, a two-drive conveyor with drives configured to a symmetrical optimum, provided there is a current equalization subsystem, is a more favorable control object when solving problems of damping elastic vibrations in the system under consideration with distributed parameters. Therefore, we can conclude that the matrix model in the form of an m-file allows you to quickly obtain interesting and useful information about a system with a massive elastic-viscous traction body at the level of using simple engineering knowledge.

To check the correctness of the matrix model of the conveyor, compiled in the Matlab package, its model was developed in the form of a block diagram and its transfer functions were obtained using the graph method [24], [25] in the Maple package. Since the transfer functions were obtained in the symbolic transformation mode, the error in the final calculation of their coefficients is determined only by the number of reliable decimal places after the decimal point. In Matlab this number is 16, and in Maple it is specified by the user, and was chosen by the authors as 64. Such a strict approach was adopted after assessing the number of non-overlapping contours in the model. To determine their number, as a result of the analysis of the structural diagram and its generalization, the following formula (11) was obtained:

$$N_k = \frac{1}{k!}\prod_{i=0}^{k-1}(n-(k-1+i)), \tag{11}$$

where N_k – is the total number of k non-touching loops (two for k=2, three for k=3, etc.) in all combinations considered;

n – is the total number of loops of the structural diagram of the chain model of the rod;

Π – product sign.

The number of non-touching loops in the chain model is enormous. It is presented along the ordinate axis in Fig. 7 and for models of n=100 and n=200 masses lies, respectively, in the range of $10^{40}\div10^{80}$. Fortunately, the loops are of the same type, the transfer functions of one $W_1(p)$, two $W_2(p)$, three $W_3(p)$, and the number of $W_k(p)$ non-touching loops k are written simply:

$$W_1(p) = \frac{c_{12}+\beta_{12}p}{mp^2}, \tag{12}$$

$$W_2(p) = \left(\frac{c_{12}+\beta_{12}p}{mp^2}\right)^2, \tag{13}$$

$$W_3(p) = \left(\frac{c_{12}+\beta_{12}p}{mp^2}\right)^3, \tag{14}$$

$$W_k(p) = \left(\frac{c_{12}+\beta_{12}p}{mp^2}\right)^k, \tag{15}$$

where c_{12} – is the stiffness coefficient between the first and second scraper;

β_{12} – is the coefficient of internal viscous friction between the first and second scraper;

m – is the scraper weight;

k – is the number of masses.

In this case, the maximum value of k is equal to $k_{max}=n/2$, where n is the number of circuits of the chain model (it is even).

Taking into account the obtained dependencies, the transfer function of the entire model is written in the Maple package in general form as:

$$W_1(p) = \frac{R(p)}{Q(p)}, \tag{16}$$

$$R(p) = \frac{1}{mp}(W_1(p))^{\frac{n}{2}}(mp^2)^{\frac{n}{2}}, \tag{17}$$

$$Q(p) = \Delta(mp^2)^{\frac{n}{2}}, \tag{18}$$

$$\Delta = 1 + N_1W_1(p) + N_2W_2(p) + N_3W_3(p) + \ldots$$

$$N_kW_k(p) + \ldots + N_{\frac{n}{2}}W_{\frac{n}{2}}(p). \tag{19}$$

This allows, with an accuracy of up to 64 or more decimal places of coefficients for known numerical values of the model parameters, without solving equations to find the roots of the characteristic equation of the resulting transfer function, to construct frequency characteristics in the Maple package and compare their resonant frequencies with the resonances of the matrix model in Matlab. The comparison showed the coincidence of the resonant frequencies of the models of the considered electromechanical conveyor system with distributed mass, elasticity and viscosity of the chain, obtained in different packages using different algorithms, which indicates the correctness of modeling an electric drive with a complex mechanical subsystem.

We will simulate dynamic processes using a computer model developed by the authors [26], [27] of a one- and two-drive armored face conveyor CP72 with an adjustable asynchronous electric drive with a vector control system, which is represented by a standard double-loop speed control system under the assumption of fully functioning flow loops (Fig. 7). This model makes it possible to study the start-up in current-limiting mode and the soft start of a conveyor, its operation at a certain transportation speed, as well as chain jamming for various reasons.

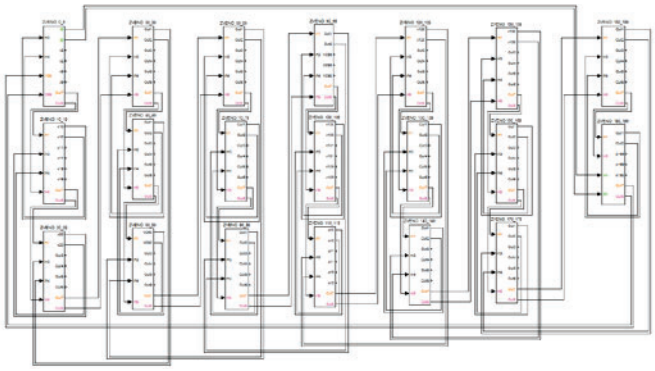


Fig. 7. Block diagram of the 200-mass model of armored face conveyor for Matlab

The mechanical part of the conveyor in the model is represented by a system with lumped parameters and consists of 200 elementary blocs interconnected. Each elementary bloc of the working branch is characterized by a total mass m_i=118 kg (chain, scraper and load), elasticity k=2·108 n/m and chain viscosity β=3000 n·s/m. Each mass is affected by the sliding friction forces of coal on steel with a friction coefficient Ψ=0.45 [15], with an inclined route - projections of the gravitational force, as well as elastic-viscous forces from neighboring links. The movement of the working body is described by a system of differential and algebraic equations compiled on the basis of Newton's second law and Hooke's law.

$$\begin{cases} m_1 \dfrac{dV_1}{dt} = F_{ef12} - F_{ef01} + F_{vf12} - F_{vf01} + F_{ff1}, \dfrac{dX_1}{dt} = V_1, \\ m_2 \dfrac{dV_2}{dt} = F_{ef23} - F_{ef12} + F_{vf23} - F_{vf12} + F_{ff2}, \dfrac{dX_2}{dt} = V_2, \\ m_n \dfrac{dV_n}{dt} = -F_{ef(n-1)n} - F_{vf(n-1)n} + F_{ffn}, \dfrac{dX_n}{dt} = V_n, \end{cases} \quad (20)$$

where m_1, m_2, m_n – is the scrapers weight, kg;
X_1, X_2, X_n – is the chain scraper coordinates, m;
V_1, V_2, V_n – is the speed of scrapers, m/s;
F_{ff1}, F_{ff2}, F_{ffn} – is the friction forces, N;
F_{ef01}, F_{ef12}, F_{ef23}, $F_{ef(n-1)n}$ – is the elastic forces, N;
F_{vf12}, F_{vf23}, $F_{vf(n-1)n}$ – forces of internal viscous friction, N.

In Fig. 8 shows three-dimensional space-time diagrams of dynamic processes of speeds V(x, t) in the conveyor chain.

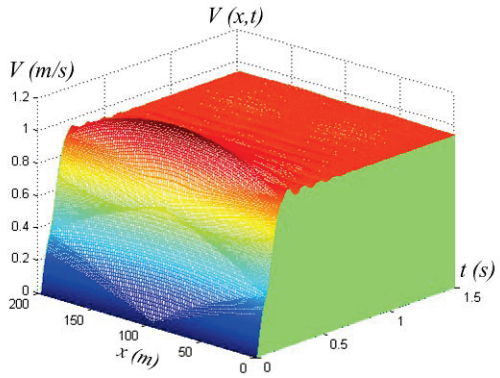


Fig. 8. Three dimensional representation of dynamic processes velocities of chain elements

Dynamic processes calculated in the model of an unloaded armored face conveyor when starting from one asynchronous adjustable electric drive located at the head of the conveyor. The speed controller in the electric drive model is proportional [28]. The chain pretension value in the model is equal to F_{pv}=44 kN, there is no sag, the modeled chain is elastic in tension and "compression" relative to the pretension level. From Figure 8 it can be seen that the deformation wave propagates from the point where the force is applied and is reflected from the free end of the conveyor. The presence of damping in the "drive" ensures damping of oscillations during 4-5 passes of the wave along the conveyor. The frequency of these oscillations is 8.3 Hz.

Figure 9 shows three-dimensional space-time diagrams of dynamic processes of forces F(x, t) in the chain.

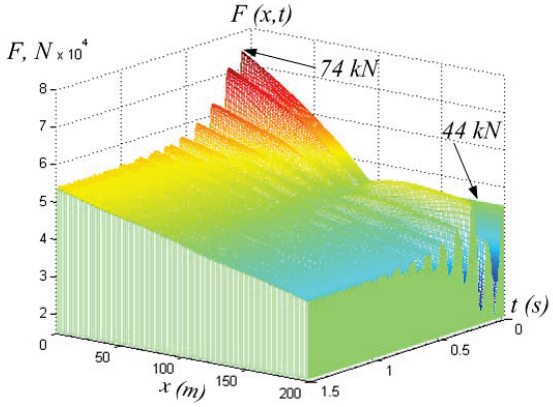


Fig. 9. Three-dimensional representation of dynamic processes of forces in a chain

From the consideration of the processes, we can conclude that starting a single-drive conveyor in current limiting mode (during the acceleration time t=0.2 s) with a maximum torque leads to the emergence of natural oscillations of forces in the chain with a frequency of 8.3 Hz, the amplitude of which is maximum near the drive station 74 kN taking into account F_{pv} and decreases with distance from it [29]. After the conveyor accelerates, the resulting oscillations die out, the induction motor transmits traction force to the working and idle branches, the maximum load in the chain at the run-up point near the drive station is 10 kN without taking into account F_{pv}.

III. CONCLUSIONS

The article presents a comparative analysis of engineering techniques for modeling mechanical systems with distributed masses and elasticities, using the electric drive of a chain conveyor as an example. The research demonstrates the effectiveness of different modeling approaches for solving typical problems related to conveyor operation, such as soft starts, current limiting and sudden braking.

A finding is the discovery of the circular distribution of the characteristic equation's roots for the conveyor model with a PI speed controller in the complex plane. This distribution is inversely proportional to the coefficient of internal viscous friction and directly proportional to the stiffness coefficient, with a constant radius $R=c_{12}/\beta_{12}$. It was also shown that roots from an analytical solution obtained via the Fourier method coincide with those from the direct discretization model for the first 10-20 lower frequencies.

The study also revealed a previously unmodeled phenomenon: the potential for soft self-excitation and instability in the system under certain conditions, which necessitates modeling with a high number of masses, specifically $n>150$. This is particularly relevant for a two-drive conveyor configured to a symmetrical optimum, which, despite its tendency towards oscillations, offers a more favorable control object for damping elastic vibrations when a current equalization subsystem is present.

Furthermore, a novel formula was derived to calculate the number of non-touching loops in the structural diagram of the model for obtaining its transfer function using the graphs method (topological method). The correctness of the developed matrix and block diagram models was confirmed by the coincidence of resonant frequencies obtained in different software packages (Matlab and Maple) using distinct algorithms. This comprehensive set of results enables rapid and accurate modeling of electromechanical systems with complex mechanical subsystems, providing a robust framework for further analysis and controller synthesis.

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Study of Individual Elements of the Biomedical Protection System Against Electromagnetic Radiation

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***Abstract*—The article presents the construction of a functional model of a biomedical protection system and conducting experimental studies of absorbing materials exposed to electromagnetic radiation. It is shown that the resistive absorbing material based on polystyrene foam with the addition of graphite has the feature that allows implementing the calculated laws of change in complex dielectric permittivity.**

Keywords—absorbent material, biomedical protection system, polystyrene foam, functional model.

I. Introduction

In recent decades, there has been a rapid development of communications, radar, wireless technologies and medical electronics. However, in parallel with the growth in the intensity of the use of electromagnetic radiation (EMR), the impact of these factors on humans and the environment is also increasing. Studies show [1–3] that one of the reasons for the deterioration of human health in conditions of increased EMR levels is a change in the characteristics of the body's own energy fields. Such changes can manifest themselves in the disruption of individual organs, changes in biochemical processes and even at the molecular level [4–5].

Under short-term conditions of exposure, the body is able to partially or completely compensate for the negative impact of the external field due to its own protective mechanisms [6]. However, with prolonged exposure to an area of increased electromagnetic background, changes accumulate, which can lead to stable pathological processes of varying severity. In this regard, there is an urgent need to develop effective biomedical protection systems capable of reducing the negative impact of electromagnetic radiation. One of the promising areas is the use of special absorbing materials that provide shielding and minimize reflections in a wide frequency range [7-8]. Modern polymer-based materials with the addition of conductive fillers (for example, graphite) allow varying their electrodynamic properties and selecting the optimal parameters for protecting humans and sensitive medical equipment.

Thus, an urgent task is to develop a functional model of a biomedical EMI protection system, select and study materials with optimal characteristics, and experimentally test their effectiveness.

II. Functional Model Of The Biomedical Protection System

The aim of the work is to develop and experimentally study a functional model of a biomedical system of protection against electromagnetic radiation, as well as to study the properties of polymeric materials with the addition of graphite to create effective absorbing structures that reduce the level of reflection and increase the shielding factor in a wide frequency range.

The synthesis of an optimal biomedical protection system (OBPS) can be performed on the basis of a generalized functional-statistical criterion [9]. In this case, in general, for synthesis it is necessary to know the laws of probability distribution of X, G, Z, U, V, Y, F random vectors. Knowledge of the characteristics of these vectors will make it possible to determine the necessary generalized indicators for assessing the effectiveness of protection [10-16]. For X, G vectors, we can confine ourselves to determining the intensity of microwave radiation (the values of the electric E and magnetic H components) for various radiation sources.

When solving the problem of system synthesis, based on the permissible values of Y and F vectors, it is necessary to synthesize devices for biological protection against electromagnetic radiation (EMR) and select control laws for U and V. Fig. 1 shows a functional model of a biomedical protection system with EMR sources that are affected by a certain vector of external and internal disturbances $Z(z_1, z_2,\ldots,z_j)$, which is random in nature. The action of EMR sources also determines the randomness of $X(x_1, x_2,\ldots x_n)$ vector and the vector of sources $G(g_1,g_2,\ldots,g_n)$.

The vector of output coordinates of the EMR source can determine the spatial and temporal positions of the j-th object, and also carry information about the states of EMR sources when the vector of control signals $U(u_1, u_2,\ldots u_t)$ acts on the microwave radiation source and on the protection system $V(v_1,v_2,\ldots,v_t)$, generated by the monitoring and control system (MCS). By performing

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measurements of X, G vectors, characterizing the states of the EMR before and after the environment, the MCS generates $U(U_1, U_2,\ldots,U_n)$ and $V(V_1,V_2,\ldots,V_m)$ control vectors. The protection system forms $Y(y_1, y_2,\ldots,y_k)$ vector. The damage to the biological environment is characterized by $F(f_1, f_2,\ldots,f_i)$ vector. The functional diagram shows (Fig. 1) the main connections between the elements of the BMPS control and management system, which make it possible to construct mathematical models for individual blocks of the system.

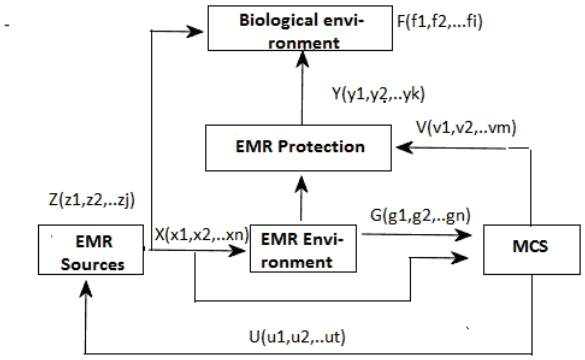


Fig.1. Functional block diagram of the Biomedical Protection System (BMPS).

EMR sources S under disturbances Z generate an incident field $X=\{E, H\}$. The Monitoring & Control System (MCS) measures X and source state G, estimates biological impact F, and issues controls U (to sources, if accessible) and V (to the protection system P). The protection system transforms the incident field into a mitigated field Y at the protected biological object B.

III. Manufacturing Absorbing Material

The protective properties of the material based on polystyrene foam with the addition of crushed graphite particles of different particle sizes in the matching layer were studied. Depending on the time the granules are kept in hot water, it is possible to obtain polystyrene foam of varying density, which is directly proportional to the dielectric constant and can take a value from 1 to 2.5.

Due to the fact that the particle diameter in polystyrene foam and the wavelength in the millimeter range become quite close, there is a danger of increased reflection. To prevent such phenomena, the sizes of polystyrene foam granules were chosen to be equal to 2–5 mm. The prepared aquadag was applied in an even layer onto flat polystyrene foam plates measuring 75–45 cm, and exactly the same amount of aquadag by weight was applied to each plate. After drying, the plates were examined on an experimental installation that allows determining the characteristics of the material in the frequency range of 10–75 GHz. The best characteristics for this frequency range were demonstrated by aquadag with a graphite particle size of 30–70 μm, used in the manufacture of the absorbing material. After the first foaming, polystyrene foam granules are covered with a uniform film of aquadag with a certain concentration. After this, the granules are dried. The next stage of manufacturing the absorbent material is its formation. The granules, coated with aquadag, are loaded into the press form, and its filling by volume should be 100%. For free penetration of hot water into the mold and for gas to escape from it, it is necessary to make the walls perforated.

IV. Installation For Studying Absorption Materials

To study the amplitude-angular characteristics of the scattered field and determine the parameters of the absorbing material in free space, it is necessary to find out the optimal relationships between the electrodynamics, geometric and frequency characteristics of the experimental installation, required to obtain reliable measurement results. The distance in the ratio: “transmitting system – test sample and receiver” must correspond to the conditions under which the waves in the area of the material under study can be considered as plane. In this experiment, it is necessary to ensure the removal of the dependence of the scattered field pattern on the angle of incidence of the electromagnetic wave, and also to ensure minimal coupling of the antennas through parasitic re-radiations and their mutual multiple coupling directly through the sample under study.

To ensure reliable experimental results, wave reception interference due to extraneous radiation must be small, and the sensitivity of the receiver must be high.

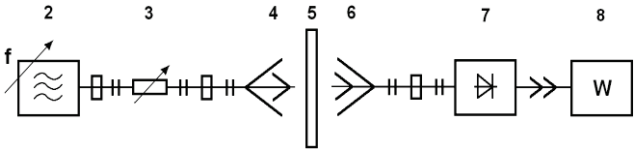


Fig. 2. Functional diagram of the experimental installation: 2 – microwave generator, 3 – attenuator, 4 – transmitting antenna, 5 – material under study, 6 – receiving antenna, 7 – receiving converter, 8 – indicator unit.

Changing the polarization of an electromagnetic wave can be accomplished by using a radiator with circular polarization or with linear polarization with two orthogonal polarizations. Obtaining greater accuracy with a relatively simple design is ensured in the second case.

The dependence of the directional pattern of the scattered field on the angle of incidence of the electromagnetic wave can be determined by moving the receiver around the sample under study while keeping the latter’s position unchanged. The proposed measuring installation largely satisfies the above requirements. In the room, to eliminate parasitic re-radiation between the antennas, an anechoic chamber was simulated using absorbing materials with a signal reflection level of 50 dB. The use of horn antennas with a sufficiently low level of side lobes and good matching with the transmitting path ensured minimal multiple coupling of the antennas. The amount of reflection was eliminated by using matching transformers at the antenna input. In this case, the error caused by the lens properties of the material in the studied area of measuring its parameters did not exceed 2%. The reflection coefficient is determined by comparing the signals reflected by the sample and the reference, measured under conditions where the attenuation factors are the same. A polished copper sheet of the same dimensions as the sample was used as a reference reflector. Since the materials under study absorb much more power than they reflect, it is

possible to measure the shielding factor using the substitution method without taking reflection into account. The module of the sample shielding coefficient is determined by comparing the signals received at the receiving antenna in the absence and in the presence of the sample being measured.

V. Experimental Studies Of Heterogeneous Materials

The material was manufactured in two types, differing in the principle of matching with free space, the dielectric constant of the base, the law of distribution of graphite in the material and the geometry of the front surface. The first type is a block with a flat front surface. It consists of a non-uniform lossless dielectric. In the frequency range of 10 GHz to 75 GHz and angles of incidence of the electromagnetic wave within the range of 0–70°, the dielectric constant varied from 1 to 2.0. The matching layer was made of five separate polystyrene foam plates with different dielectric constants, increasing from the first layer to the last one. The inhomogeneous lossy dielectric was made of five layers of polystyrene foam with ε = 2.0. Aquadag of varying concentrations was added to each layer as an absorbent material. The layers obtained in this way were connected in a certain sequence into a monolithic block and, together with the matching layer, formed the absorbing material of the first type. The second type of material has the same structure as the material of the first type, but without a matching layer. For this type of material, the dielectric constant of the base of the layers is close to unity. Figures 3-5 show the dependence of the reflection coefficient (Kot) in dB of an electromagnetic wave (EMW) on the first type of absorbing material with and without a matching layer as a function of the angle of incidence for various frequencies.

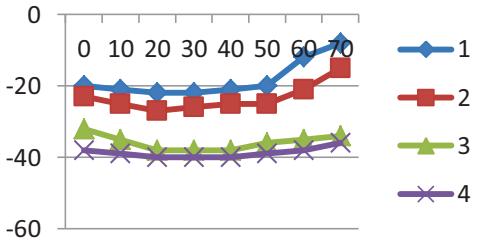


Fig.3. Dependence of Kot on the angle of incidence of the electromagnetic wave on the material at a frequency of 10 GHz .

The horizontal axis in Figs. 3–8 corresponds to the angle of incidence of the electromagnetic wave (in degrees), and the vertical axis corresponds to the reflection coefficient Kot expressed in decibels (dB). The curves numbered (1–4) represent measurements at different incident polarizations or different types of sample structures (with/without matching layer, or with flat/relief front surface).

Figures 6-8 present the results of studying the characteristics of the absorbing material with a shaped front surface. For a material with a relief front surface, the reflection coefficient depends significantly on the angle of incidence of the EMR and the plane of its polarization at a frequency of 75 GHz. With increasing electromagnetic wavelength, the absorption properties of the material deteriorate significantly.

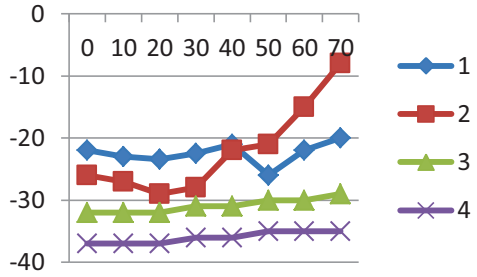


Fig.4. Dependence of Kot on the angle of incidence of the electromagnetic wave on the material at a frequency of 38 GHz.

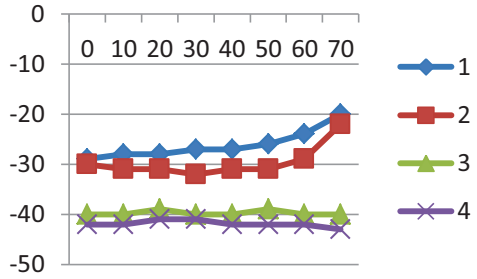


Fig.5. Dependence of Kot on the angle of incidence of the electromagnetic wave on the material at a frequency of 75 GHz.

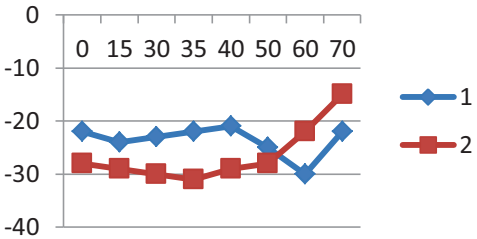


Fig.6. Dependence of Kot on the angle of incidence of the electromagnetic wave on the material at a frequency of 10 GHz.

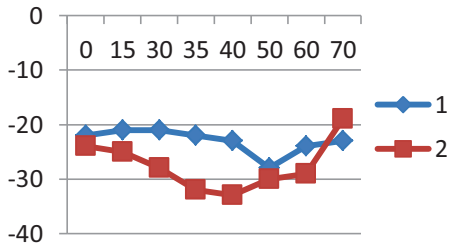


Fig.7. Dependence of Kot on the angle of incidence of the electromagnetic wave on the material at a frequency of 38 GHz.

Figures 3–5 show the dependence of the reflection coefficient Kot (dB) on the incidence angle (°) for absorbing materials with and without a matching layer at frequencies of 10 GHz, 38 GHz, and 75 GHz, respectively. Curves (1–4) correspond to different polarizations of the incident field. All dependencies were obtained experimentally on the measuring installation described in Section 3. Figures 6–8 present experimental dependencies of Kot (dB) versus incidence angle (°) for absorbing materials with a shaped front surface at the same frequencies. Curves (1–4) correspond to different orientations of polarization and measurement conditions.

These results were obtained by the authors and are not reproduced from external sources.

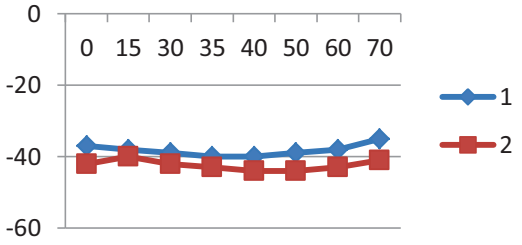


Fig.8. Dependence of Kot on the angle of incidence of the electromagnetic wave on the material at a frequency of 75 GHz .

This correction ensures that each graph is fully self-explanatory and its experimental origin is clear. Analyzing the graphs, we can conclude that the absorbing material without a matching layer has a higher reflection coefficient than with a matching layer. In addition, it is clear that the reflection coefficient of the material with the matching layer is almost independent of the angle of incidence. For the frequency range under study, the reflection coefficient exceeds – 30 dB and reaches -40 dB at a frequency of 75 GHz.

VI. Conclusions

An experimental study of absorbing materials exposed to electromagnetic radiation was carried out. In particular, an absorbing material without a matching layer with the addition of different concentrations of aquadag to each layer has a reflection coefficient greater than an absorbing material with a matching layer. The absorption properties of the material are practically independent of the angle of incidence of the electromagnetic wave. It has been experimentally proven that it is possible to create a material with the best reflection and absorption coefficients in a wide range of angles of incidence of an electromagnetic wave in the lower part of the frequency range. The use of cone-shaped reflective elements in the material leads to a significant increase in the absorption coefficient. It has been experimentally established that the EMR shielding material based on polystyrene foam with the addition of graphite has high shielding properties without the use of complex manufacturing technology. Aquadag with a graphite particle size of 30-70 microns has the best characteristics in the studied frequency range. There is also a laboratory base [17, 18, 19] and experience in creating telemedicine services [20-22], which can be useful for developing systems to protect medical personnel.

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Modeling and Analysis of Multifrequency Piezoresonant Oscillatory Systems under Vibrational Destabilization

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***Abstract*—This paper presents an analysis of mathematical and numerical models describing multifrequency piezoresonant oscillatory systems. Multifrequency excitation is a fundamental physical property of quartz resonators, arising from anisotropic elastic coupling. However, in the context of highly stable self-oscillatory systems used as reference frequency sources, this property is generally undesirable and often mitigated through design and manufacturing constraints. A critical review of existing mathematical models reveals that none can comprehensively describe the dynamics of piezoresonant systems subjected to vibrational destabilizing influences. To address this limitation, a new modeling approach is proposed, incorporating nonlinearities, environmental disturbances, and discrete boundary constraints. The model is validated through simulations in COMSOL Multiphysics and SolidWorks, using a piezoceramic disk mounted on four supports. Simulation results confirm the presence of multiple closely spaced resonant modes and demonstrate the sensitivity of system dynamics to mechanical stress and excitation conditions.**

Keywords—piezoresonant system, electronic, piezoelectric device, oscillations, multifrequency, model

I. INTRODUCTION

Quartz resonators are used in reference frequency generators, frequency-controlled oscillators, selective devices: filters, frequency discriminators, etc.

Quartz resonators are devices that utilize the piezoelectric effect to create electrical oscillations of a specific frequency. When the frequency of the applied voltage coincides with the natural frequency of the mechanical frequencies of the quartz vibrator, a resonance phenomenon occurs in the device, resulting in a sharp increase in conductivity [1].

Any piezoelectric resonant vibration system (RVS) can be considered as a multifrequency system, since it is the physical essence of the quartz crystal resonator (QSR). The anisotropy of the crystal element basically leads to various types of elastic coupling of stresses and strains and causes the appearance of lateral resonances in all piezoelectric resonant devices (PRD). Usually, multifrequency is an undesirable property of RVS and is combated by various design and technological methods. However, a constant increase in requirements for their metrological characteristics under the influence of not one destabilizing factor, but a complex (temperature, vibration, radiation, electromagnetic field, etc.), these methods are ineffective in solving the main problem of RVS - invariance to destabilizing factors [1].

The investigation of self-oscillatory systems becomes significantly more complex as the number of excitation frequencies increases. In [2], a generalization of the averaging techniques applied in the analysis of these systems is presented, addressing a differential equation system that includes a small parameter:

$$\frac{dp}{dt} = \varepsilon X(p,q)$$
$$\frac{dq}{dt} = \omega(p) + \varepsilon Y(p,q) \quad (1)$$

where $p = (p_1,...,p_m)$ is *an m-dimensional* vector of slow movements, $q = (q_1,...,q_n)$ is *an n-dimensional* vector of fast movements, $\omega(p) = (\omega_1(p), ..., \omega_n(p))$, and the vector functions *X*(*p*, *q)* and *Y*(*p*, *q)* can be represented as terms:

$$X(p, q) = \sum_{|k|\geq 0} X_k(p) e^{i(k, q)},$$
$$Y(p, q) = \sum_{|k|\geq 0} Y_k(p) e^{i(k, q)} \quad (2)$$

where $k = (k_1, ..., k_n)$ is a vector with integer positive and negative components, the norm of which is calculated by the formula: $|k| = \sum_{i=1}^{n} |k_i|$.

The solution of differential equations (1) relies on the assumption that the right-hand sides of these equations possess uniform mean values. Reference [2] demonstrates that the absence of uniform mean values results in discrepancies between the solutions of a system of differential equations of type (1) and the associated space-averaged system of variables (*q*).

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$$\frac{dp}{dt} = \varepsilon \overline{X}(\overline{p}), \tag{3}$$

where.

$$\overline{X}(\overline{p}) = \frac{1}{(2\pi)^n} \int_0^{2\pi} \dots \int_0^{2\pi} X(\overline{p}, q) dq_1 \dots dq_n, \tag{4}$$

obtained at the same initial values, differ by a finite amount, at $t \sim \frac{1}{\varepsilon}$. This results in an equivalent system:

$$\frac{dx}{dt} = \varepsilon X(x + p_0,\ y + \omega_0 t + q_0),$$

$$\frac{dy}{dt} = \omega(x + p_0) - \omega_0 + \varepsilon Y(x + p_0,\ y + \omega_0 t) + q_0), \tag{5}$$

where $\omega_0 = \omega(p)$.

Thus, according to [3, 6], if the natural frequencies of the circuits are not in the ratio of prime integers and the attenuation of the circuits is sufficiently small, the reduced differential equations are represented by a system of the form (for single-circuit oscillators) [4]:

$$\begin{aligned} X_1 &= F_1(X_1, ..., X_n); \\ X_2 &= F_2(X_1, ..., X_n) \\ X_n &= F_n(X_1, ..., X_n) \end{aligned} \tag{6}$$

Where

$$F_i = [S_i(X_1, ..., X_n) R_i k_{33} - 1] X_i, \tag{7}$$

Where X *is the* relative dimensionless amplitude of oscillations; S_i is the average steepness for the fundamental frequency; R_i is the resonant impedance of the circuit; k_{33} is the feedback coefficient.

The steady-state equations, in which the amplitudes of oscillations are constant, are obtained from (6), (7) if we assume [7] $X_1 = X_2 = ... = X_n = 0$:

$$F_i = 0,\ i = 1, 2, ..., n,$$

from which the stationary values of the oscillation amplitudes X *are* determined$_{i0}$.

II. Main Part

A quartz crystal is an anisotropic body, and its physical properties vary in different directions. To describe the physical properties of quartz, as well as in engineering, the crystallographic coordinate system is mainly used, in the direction of which it has certain (characteristic for each axis) properties [8]. The crystallographic class (symmetry group 32), to which quartz also belongs, is characterized by the 3rd order symmetry axis (*Z)* and three pairs of 2nd order polar axes (*X* and *Y*) [9].

Here, *Z* is the "optical" axis, which coincides with the longitudinal axis of the crystal. In this direction, the piezoelectric effect does not appear, the electrical conductivity is higher than in the perpendicular direction, and there is no double refraction of the light beam. *The X-axis* is the "electric" axis, directed parallel to one of the prism faces; there are three *X-axes* in a quartz crystal. In the direction of *the X-axis*, mechanical forces cause intense electric charges to appear. *The Y* axis is the "mechanical" axis, directed perpendicular to *the m* faces; there are also three *Y* axes in a quartz crystal.

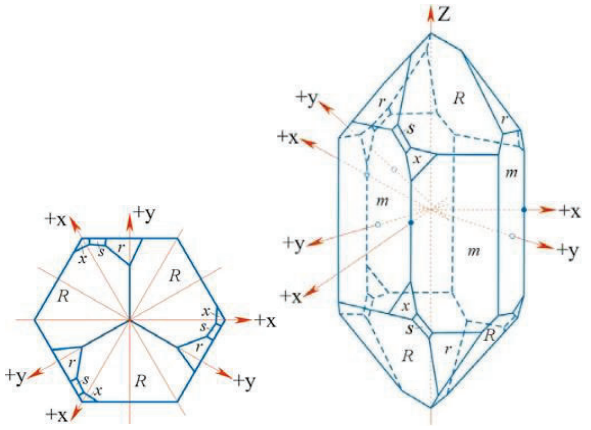


Fig. 1. Idealized enantiomorphic shape of α-quartz crystals and arrangement of crystallographic axes: *R* - basic positive rhombohedron; *r* - negative rhombohedron; *s* - trigonal bipyramid; *x* - trigonal trapezoid [9]

To manufacture quartz resonators, piezoelectric elements can be cut from a quartz crystal at different angles relative to the crystallographic axes in the form of plates of various shapes, bars, beams, etc. This orientation of the piezoelectric element is called a slice. There is a generalized conventional classification of quartz resonator cuts, which can be represented in the form of Fig. 1.

The symbol for the initial orientation of the piezoelectric element consists of two letters (*X*, *Y*, or *Z*): the first indicates which axis is parallel to the thickness of the piezoelectric element, and the second indicates which axis is parallel to its length. By rotating the piezoelectric element around its edges at different angles, a wide variety of orientations can be obtained - the so-called "oblique" sections.

The orientation designation of a piezoelectric element with edges forming angles with the crystallographic axes consists of the designation of the primary orientation, to which is added one, two, or three letters (*l*, *b*, *s*) indicating which directions (along the length, thickness, or width) of the piezoelectric element edges are used as axes of the first and subsequent rotations from the position of the primary orientation.

The angles of the first, second, and third rotations of the piezoelectric element around these axes are denoted by the letters α, β, and γ, respectively, which are placed through the oblique lines. The numbers that appear in place of these letters indicate the magnitude of the rotation angles. The angle of rotation is considered positive if the rotation is counterclockwise.

The orientation designation of a piezoelectric element with edges forming angles with the crystallographic axes consists of the designation of the primary orientation, to which is added one, two, or three letters (*l*, *b*, *s*) indicating which directions (along the length, thickness, or width) of the piezoelectric element edges are used as axes of the first and

subsequent rotations from the position of the primary orientation.

The angles of the first, second, and third rotations of the piezoelectric element around these axes are denoted by the letters α, β, and γ, respectively, which are placed through the oblique lines. The numbers that appear in place of these letters indicate the magnitude of the rotation angles. The angle of rotation is considered positive if the rotation is counterclockwise.

In [10] considered a model of a system for studying multi-frequency oscillations, which is described by a system of equations:

$$x + W_1(p)W_2(p)y - W_2(p)z - 0; \qquad (8)$$
$$y = F(x).$$

If the signals at the input and output of a nonlinear element (NLE):

$$x = x_0 + Ae^{j\omega_1 t} + Be^{j\omega_2 t}e^{j\alpha};$$
$$y = y_0 + AW(\omega_1/\omega_2)e^{j\omega_1 t} + BW(\omega_2/\omega_1)e^{j\omega_2 t}e^{j\alpha},$$

Where $W(\omega_1/\omega_2)$ is the transfer function excluding frequency ω_1 under the influence of the signal at frequencies ω_1 and ω_2; $W(\omega_2/\omega_1)$ is the transfer function excluding frequency ω_2 under the influence of the signal at frequencies ω_2 and ω_1 *; A, B are the* amplitudes of oscillations.

In addition to the above advantages of piezoresonant devices based on quartz resonators, they have a number of disadvantages and limitations, among which the most typical are sensitivity to external destabilizing factors such as temperature and mechanical stress. Under the influence of external mechanical influences, deformations occur on the piezoelectric element, which lead to changes in its geometric dimensions and density and, as a result, to changes in its natural resonance frequency. It is necessary to determine the distribution of mechanical stresses in the piezoelectric crystal with a high degree of accuracy [8, 10], taking into account the type of cut and orientation of the piezoelectric plate in space.

III. PRACTICAL IMPLEMENTATION

One of the structural and technological methods to increase the vibration resistance of piezoresonant devices is to improve the design of quartz holders [6, 7] and their usage [10]. First of all, the frequency properties of the quartz holder structures are studied, which must satisfy the provision that the frequency of natural oscillations of the structure should be outside the range of the vibration frequency acting on the quartz resonator. The methodology of such studies is as follows. The spectrum of resonant frequencies of the structure and the dependence of the spectrum on the variable parameters are determined, taking into account both random (due to tolerances on the structural elements) and systematic (due to differences in the size of the piezoelectric elements) frequency shifts. This dependence, as well as the spectrum, can be determined theoretically. The results of calculations based on the formulas, as well as the experimental determination of the quality factor of the mounting system, allow us to select the necessary parameters of already known structures (for this purpose, the nomogram method is used) [11]. Develop new vibration-resistant systems for mounting quartz elements by reducing the size of the quartz holder, using mounting elements with a sufficiently large cross-sectional area, high elastic modulus, etc [12].

Various design and technological parameters of the quartz resonator, such as the type of cut, the configuration of the holder, the number of piezoelectric element attachment points and their relative position, etc., have a significant impact on short-term frequency changes. At Fig. 2 we can see the advantages of attaching piezoelectric elements of various sections in terms of minimising vibration sensitivity.

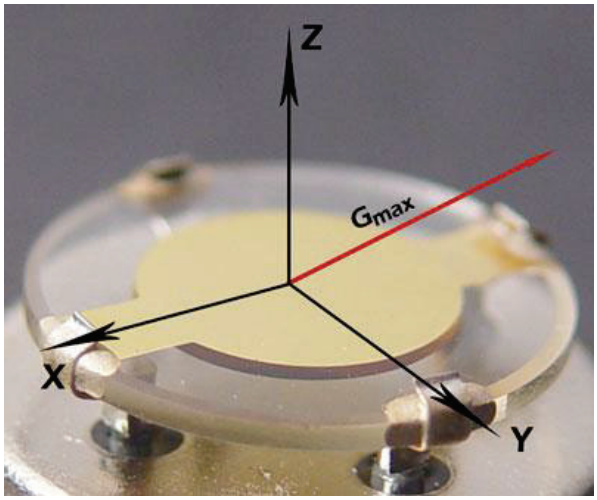


Fig. 2. An example of a piezoelectric element fixed at four points with minimal vibration sensitivity

The calculation method for circular piezoelectric elements with three or more mounting points is as follows. The initial stress and displacement fields at each point in the plane of the quartz plate are determined as a function of the plate orientation, acceleration direction and location of the attachment points. Then, equations are derived that relate the displacements for thickness-shear and bending vibrations through elastic constants and calculate the changes in the frequency of thickness-shear vibrations, which are presented in the form

$$\Delta f/f = S(\varphi,\ \gamma;\ \theta,\ \phi;\ D/b,\ n_{kp},\ \alpha_k;\ G), \qquad (9)$$

i.e., a function of the azimuthal angleψ formed by the acceleration vector relative to the crystallographic axis X and the angle of the siteγ , the acceleration G and the slice (θ,φ) , the ratio D/b (diameter to thickness of the piezoelectric element), the number of attachment points n_{kp} and the angle $\alpha_k (K = \overline{1, n_{kp}})$, determining the position of the attachment point relative to the X axis.

Classical models of piezoresonant systems typically rely on linear second-order differential equations to describe the fundamental oscillatory behavior of quartz resonators [13, 14]. These models assume a single dominant mode of vibration, with the system treated as a damped harmonic oscillator:

$$mx^{\cdot\cdot} + cx^{\cdot} + kx = F(t)m, \qquad (10)$$

Where m is the effective mass, c is the damping coefficient, k is the stiffness, and $F(t)$ is the external force or excitation. While adequate for single-mode analysis and low-amplitude

operation, such models inherently neglect multifrequency interactions and nonlinear effects present in real-world scenarios.

To account for multiple resonant frequencies, some models extend the single-degree-of-freedom (SDOF) approach to multi-degree-of-freedom (MDOF) systems. These represent the resonator as a system of coupled oscillators, each corresponding to a vibrational mode [15].

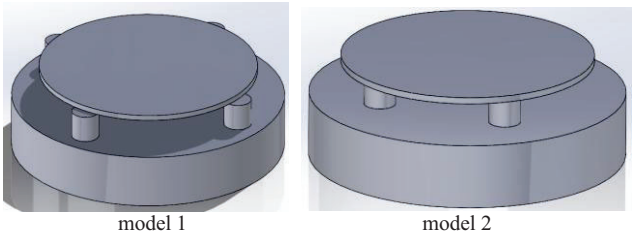
While these models provide insight into mode superposition and coupling, they remain predominantly linear and often exclude nonlinear modal interactions, mode hopping, or internal resonance—phenomena that play crucial roles under vibrational destabilization.

Despite the progress in modeling techniques, existing approaches suffer from several key limitations when applied to multifrequency piezoresonant systems [16]:

- Inadequate representation of internal resonances and nonlinear modal interactions;
- Neglect of environmental destabilizing factors, such as random vibrations or temperature-induced parameter drift;
- Oversimplified boundary conditions, which fail
- to represent the real mounting and excitation conditions of quartz elements;
- Lack of integration between electromechanical and thermomechanical modeling, which is essential for precision applications.

These shortcomings hinder accurate prediction and control of multifrequency behavior in resonators used in high-stability oscillators and sensors. [17] Therefore, there is a critical need for more comprehensive modeling frameworks that combine mechanical, electrical, and nonlinear dynamic effects.

Two 3D-model of an 8 mm piezoceramic disk with four-point support for studying multi-frequency oscillations was developed for further research by SolidWorks. “Model 1” and “model 1”are shown in Fig.3



model 1 model 2

Fig. 3. Models of a system for studying multi-frequency oscillations

A finite element model of the piezoelectric device was developed. The simulation results obtained resonant frequencies, which showed in fig.4 and fig.5.

Study name:Frequency 1

Mode No.	Frequency(Rad/sec)	Frequency(Hertz)	Period(Seconds)
1	2,3874e+05	37 996	2,6318e-05
2	3,3648e+05	53 552	1,8674e-05
3	3,3847e+05	53 869	1,8564e-05
4	3,3971e+05	54 066	1,8496e-05
5	4,0583e+05	64 589	1,5482e-05

Fig. 4. Calculated resonant frequencies of a piezoelectric device in SolidWorks (“model 1”)

Study name:Frequency 2

Mode No.	Frequency(Rad/sec)	Frequency(Hertz)	Period(Seconds)
1	4,3949e+05	69 947	1,4297e-05
2	4,413e+05	70 235	1,4238e-05
3	4,4243e+05	70 415	1,4201e-05
4	4,4481e+05	70 794	1,4126e-05
5	4,9543e+05	78 851	1,2682e-05

Fig. 5. Calculated resonant frequencies of a piezoelectric device in SolidWorks (geometric model 2)

The results of the stress analysis of the quartz holder and piezoelectric element system in SolidWorks [18], which is an important step in optimizing the design of piezoelectric devices, are presented in Fig. 6. To compare the influence of the quartz holder design on the dynamic properties of the resonator, models with different numbers of mounting points were investigated, with a maximum mounting location error of 1^0. The obtained natural frequencies for “model 1” are shown in Fig 6 and Fig.7. The obtained natural frequencies for “model 2” are shown in Fig 8 and Fig.9.

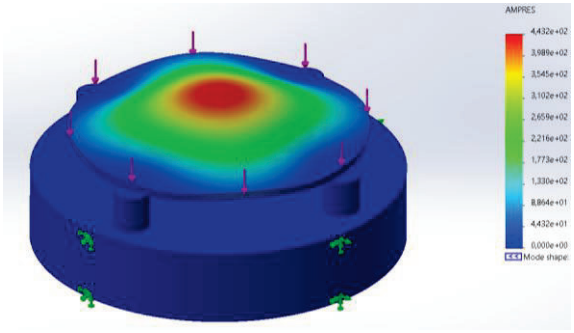


Fig. 6. Stress analysis of a piezoelectric device in the SolidWorks Simulation with for 4-point mounting – pattern of mechanical stress distribution for Frequency for “model 1”

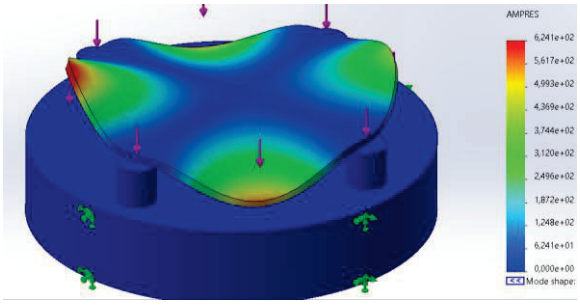


Fig. 7. Stress analysis of a piezoelectric device in the SolidWorks Simulation with for 4-point mounting – pattern of mechanical stress distribution for Frequency for “model 1”

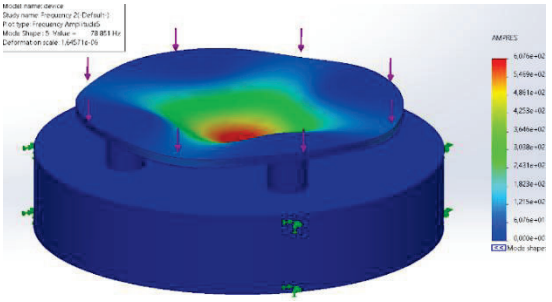


Fig. 8. Stress analysis of a piezoelectric device in the SolidWorks Simulation with for 4-point mounting – pattern of mechanical stress distribution for Frequency for “model 2”

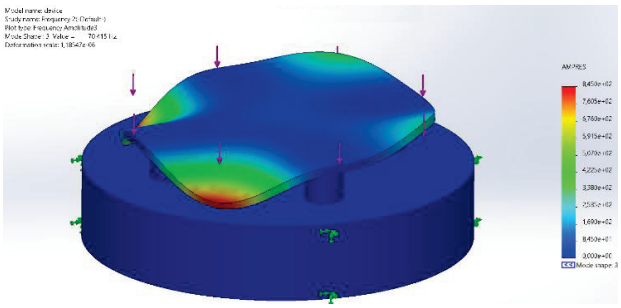


Fig. 9. Stress analysis of a piezoelectric device in the SolidWorks Simulation with for 4-point mounting – pattern of mechanical stress distribution for Frequency for "model 2"

Following the modeling, the acquired normalized mechanical stress field serves as the foundation for computing the vibromechanical frequency shifts of the quartz resonator.

Simulation for "model 2" was also carried out in the COMSOL Multiphisics. The obtained values of stresses arising during the operation of the piezoelectric device are shown in Fig. 10. Calculated oscillation data similar to the results obtained from SolidWorks Simulation.

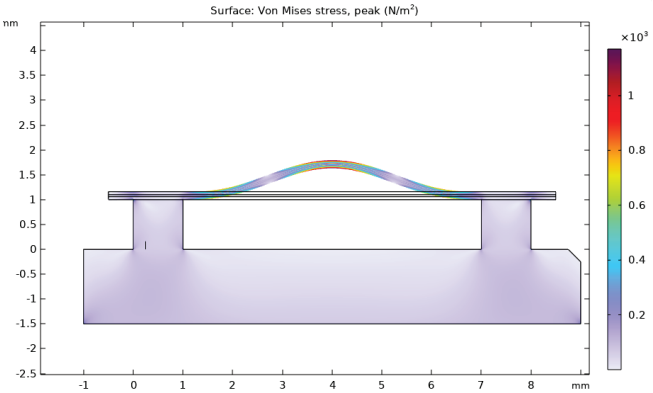


Fig. 10. Stress analysis of a piezoelectric element by Von Mises

The electric potential of a piezoelectric element calculations are shown in Fig. 11.

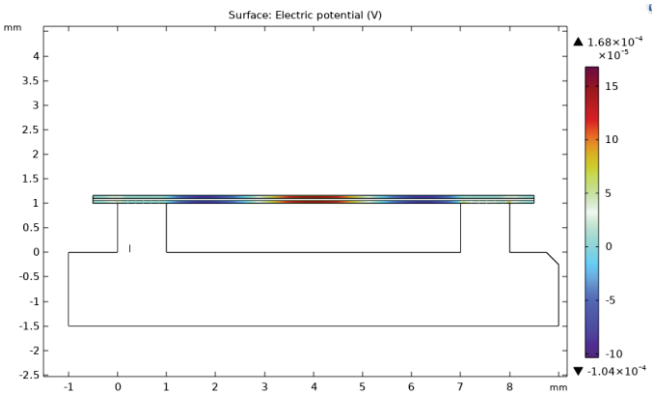


Fig. 11. Electric potential of a piezoelectric element

Fig. 12 shows an analysis how the system behaves at a fixed frequency of 70 kHz as the load resistance is varied.

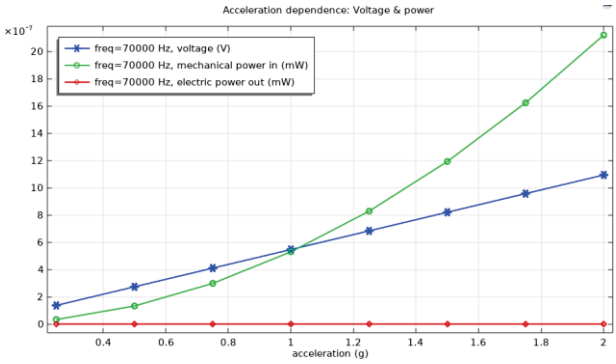


Fig. 12. Voltage & power by Load dependence for piezoelectric device

Fig.13 shows how the output voltage and power vary as you sweep the excitation frequency from 40 kHz to 80 kHz.

This is typical in energy harvesting and analysis to evaluate efficiency across electrical loading conditions and Frequency Response for piezoelectric device [19]

Within piezoelectric accelerometers, the piezoelectric material can convert the acceleration signal into a charge signal. The piezoelectric constant, electromechanical coefficient, and temperature stability of the piezoelectric material decide the performance of the sensor [20].

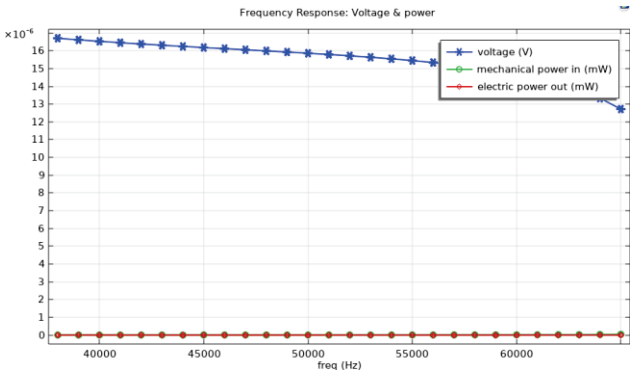


Fig. 13. Voltage & power by Frequency Response for piezoelectric device

Voltage & power and acceleration depends by Frequency Response for piezoelectric device are shown in Fig.14 and Fig.15.

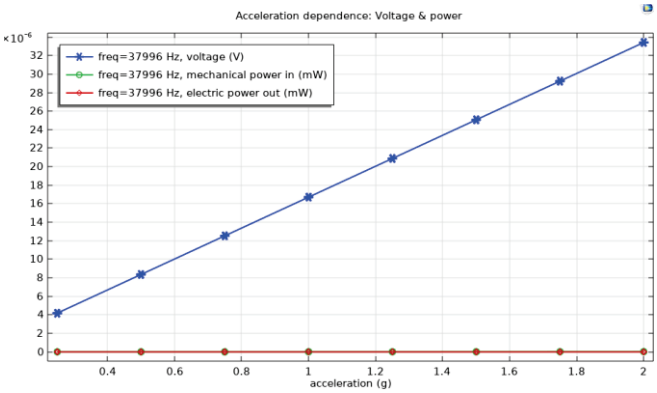


Fig. 14. Acceleration depends by Frequency Response for piezoelectric device

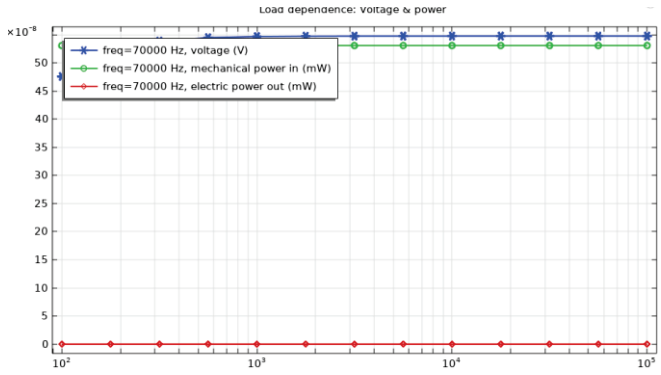


Fig. 15. Voltage & power by Load dependence for piezoelectric device

The use of COMSOL Multiphysics enabled accurate identification of eigenmodes and resonance behavior in an 8 mm piezoceramic disk with four-point support. The results confirm that standard low-frequency excitation fails to activate the system's dynamic response and highlight the importance of resonant frequency targeting. The presence of multiple closely spaced resonance modes suggests a complex interaction that requires refined modeling to ensure stability in practical applications.

Conclusions

The research presented in this paper focused on the modeling and simulation of multifrequency piezoresonant systems, emphasizing the impact of vibrational destabilizing effects and structural mounting configurations. A detailed review of classical and advanced mathematical models highlighted the limitations in capturing complex dynamic behaviors typical of quartz and piezoceramic resonators. A finite element model of an 8 mm piezoceramic disk with four-point support was developed and analyzed using both SolidWorks and COMSOL Multiphysics. The simulation results revealed multiple resonant frequencies and confirmed the influence of support design and frequency selection on mechanical stress distributions. These findings underscore the need for refined modeling strategies that integrate nonlinear dynamics, environmental variability, and practical design constraints. The proposed modeling approach provides a robust foundation for the development of vibration-resistant piezoresonant devices, with potential applications in reference oscillators, MEMS, and precision sensing systems.

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Principles of Diagnostics of Complex Systems using Structural Automaton Algebra and Taking into Account Signal Time Coordination

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***Abstract* — With the increasing complexity of modern technical systems, such as wheeled and tracked vehicles, there is a growing need for adequate diagnostic methods that can take into account not only structural but also temporal dependencies between components. This paper looks at using structural automata algebra as a formal tool for describing, analyzing, and predicting the behavior of complex systems, taking into account the temporal coordination of signals. It highlights the importance of considering temporal characteristics when diagnosing failures caused by cascading processes and interactions between elements. The theoretical foundations for the construction of structural automata, their composition, and conditions for correct functioning are presented. Particular attention is paid to the problems of mismatch between abstract and real signal processing times, which requires refinement of the rules for constructing diagnostic models. The proposed approach demonstrates its versatility and applicability to a wide class of mechatronic, energy and transport systems, ensuring the transition to predictive maintenance and increased operational reliability.**

Keywords* — *predictive diagnostics, structural automata, temporal signal coordination, technical condition, composition of automata, logical-dynamic connections, mathematical modelling, failure, automata algebra

I. Introduction

Modern trends in the development of mechanical engineering are characterized by an increase in the complexity of technical systems, which is due to the need to improve their reliability, functionality and operational efficiency. Wheeled and tracked vehicles, as examples of heavy-duty and multifunctional objects, fully demonstrate the general problem: diagnosing their technical condition requires not only in-depth engineering expertise, but also the use of adequate mathematical models capable of taking into account the complex logical and dynamic relationships between system elements.

The relevance of developments in the field of structural and temporal diagnostics is due to both the rapid development of intelligent control systems and the need to improve the accuracy of real-time fault detection. Traditional methods based on static or quasi-static models do not provide the required level of sensitivity and temporal resolution when detecting faults in systems with a high degree of interaction between components. In this context, the use of structural automaton algebra as a formalism that allows for the rigorous description of discrete-continuous processes characteristic of the functioning of complex technical objects becomes particularly important.

A key direction in modern technical diagnostics is predictive analysis of technical condition, based on forecasting the development of potential failures long before they actually occur. This approach minimizes downtime, optimizes maintenance and improves the overall economic efficiency of equipment operation. Predictive diagnostics requires mathematical models that take into account not only current signals, but also their temporal dynamics and interrelationships. In this respect, the use of structural automata, taking into account the temporal consistency of signals, provides an effective tool for formalizing, monitoring and predicting the behavior of complex technical systems.

An additional factor that reinforces the need to develop new approaches is the temporal consistency of diagnostic signals, a parameter that is often ignored in classical diagnostic models but is critical when analyzing failures resulting from the cascading influence of components on each other. The use of structural automata, taking into account time constraints, allows not only to formalize the structure of the system and the connections between its nodes, but also to reflect real cause-and-effect relationships over time, which significantly increases the reliability, speed and predictive power of diagnostic solutions.

The proposed approach is universal and applicable not only to wheeled and tracked vehicle systems, but also to all

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mechanical and mechatronic industries, from aviation and power plants to numerically controlled machine tools and robotic systems. Thus, the creation and development of diagnostic methods based on structural automaton algebra, taking into account the temporal coordination of signals, is a promising scientific direction that is in demand in a wide range of modern engineering and technology industries and ensures the transition to a new generation of predictive maintenance.

II. State-of-the-Art

Modern technical systems, including wheeled and tracked vehicles, are becoming increasingly complex, requiring effective methods of diagnostics and maintenance [1]. Predictive diagnostics, based on data analysis and failure prediction, is becoming a key area in ensuring the reliability and efficiency of such systems [2, 3, 4].

In recent years, there has been active development of predictive diagnostic methods using machine learning and time series analysis [5–9]. For example, in [10], a predictive maintenance system for armored vehicles was proposed, based on an ensemble of machine learning models, including Light Gradient Boosting and Random Forest, which made it possible to achieve high accuracy in predicting technical condition.

Structural automaton algebra provides a formal approach to modelling and analyzing complex technical systems [11]. In [12], a digital twin for civil engineering structures is proposed, using probabilistic graphical models and dynamic Bayesian networks to account for uncertainty and temporal consistency of signals. This allows real-time diagnostics of the condition of structures and optimization of maintenance planning.

The temporal consistency of signals plays an important role in the accuracy of technical system diagnostics [13, 14]. In [15], a method is presented for determining the location of damage in mechanical structures using a multi-purpose global pattern search, which allows temporal aspects of signals to be taken into account and improves the accuracy of damage localization. Also, [16] describes a computational framework for real-time diagnostics and condition prediction of aircraft propulsion systems, emphasizing the importance of considering the temporal characteristics of signals to improve the accuracy of residual life estimation.

Predictive diagnostics is used not only in the transport industry, but also in other areas. [17, 18] look at using predictive maintenance in agricultural vehicles using telemetry, maintenance history, and geospatial data, showing how versatile this approach is. In addition, [19–22] discuss intelligent maintenance systems and predictive manufacturing, emphasizing the role of machine learning [23] and the Internet of Things in the implementation of predictive diagnostics in industry.

Thus, the integration of structural automaton algebra and the consideration of signal time coordination in predictive diagnostics of complex technical systems, including wheeled and tracked vehicles, represents a promising direction of research that contributes to improving the reliability and efficiency of equipment operation.

III. Problem Statement

Let us introduce the algebra of structural automata. The system of operations F allows us to describe the structure and laws of functioning of a fairly broad class of real systems. Applying any element $c \in C$ of any finite sequence of operations $F = \{f_1^1, f_2^1, ..., f_9^1, f_{10}^1\}$ from a set of operations to any element results in a composition that is an automaton. However, the result of applying some elements $c \in C$ of finite sequences of operations $F = \{f_1^1, f_2^1, ..., f_9^1, f_{10}^1\}$ from a set of operations to some elements is not an automaton.

Indeed, let $(f_{10}^2 c_1 ... c_k, c)$ is some structural automaton. It is obvious that any term $f_l f_{f-1} ... f_1 f_{10}^2 c_1 ... c_k$ defines a structural automaton. A formal expression of the form $\overline{f}_l f_{l-1} ... \overline{f}_1 f_{10}^2 c_1 ... c_k$, where $\overline{f}_i \in F^*$, $1 \le i \le l$, also defines a certain composition of elements c_1, c_2, …, c_k. If in this composition the operation f_9^1 is applied in the variant $f_{9,1}^1$, in the case where the first output node depends on the first input node, then the result of the composition will not be an automaton (in the sense accepted in abstract automaton theory). We cannot calculate the values of functions δ and λ, which depend on δ_1, …, δ_k, λ_1, …, λ_k, knowing the input signal to the composition and the states of all elements c_1, c_2, …, c_k.

Usually, the abstraction of discrete automata time is extended to automata of the form

$$c = (S, X^n, X^m, \delta, \lambda), \quad (1)$$

where $S \subset X^i$ for some finite i; n, m are non-negative integers; $\delta : S \times X^n \to S$ and $\lambda : S \times X^n \to X^m$ are mappings.

Because of this, each automaton c_i operates in discrete time, taking non-negative integer values 0, 1, 2, …, t. Based on this, the conditions for the correct construction of diagrams are defined. With the accepted agreements on a single operating time, any composition of automata and each elementary automaton of the composition process the input signal in the same abstract time interval. Consequently, signals must arrive simultaneously at both inputs of the automaton c_2 in the composition shown in Fig. 1.

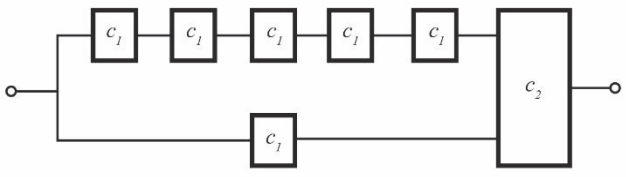


Fig. 1. Automatons layout

In practice, the processing time of an input signal by a real system R differs from the processing time of an input signal by a sequential connection of a series of such systems. In this regard, it is necessary to consider other possible abstractions of discrete automatic time that more fully account for the properties of real systems.

To fully appreciate the difficulties that arise here, let us consider some examples. Let us take two elements,

$$c_i = (S_i, X^2, X^2, \delta_i, \lambda_i),\ i = 1,2, \qquad (2)$$

which are structural automata (c_1, c_1) and (c_2, c_2) in the algebra A_C.

Let us construct a sequential composition of these structural automata, shown in Fig. 2, which is described by the term

$$T = (f^1_{9,1})^2 (f^1_4)^2 f^2_{10} c_1 c_2. \qquad (3)$$

The structural automaton c defined by the term T has the form

$$\begin{gathered} c = (S_1 \times S_2, X^2, X^2, \delta, \lambda), \\ \delta((s_1,s_2),(x_3,x_4)) = (\delta_1(s_1,(pr_1\lambda_2(s_2,(x_3,x_4)), \\ pr_2\lambda_2(s_2,(x_3,x_4)))), \delta_2(s_2,(x_3,x_4))), \\ \lambda((s_1,s_2),(x_3,x_4)) = \lambda_1(s_1,(pr_1\lambda_2(s_2,(x_3,x_4)), \\ pr_2\lambda_2(s_2,(x_3,x_4))). \end{gathered} \qquad (4)$$

However, the transition δ and exit λ functions are not always defined by formulas (4).

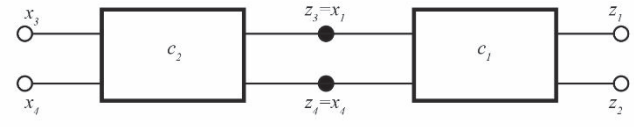


Fig. 2. Sequential composition of automatons

Suppose that structural automata (c_i, c_i), to whose inputs input signals are applied at time t, produce output signals at time $t + k_j$, $j = 1,2$. If $k_2 \geq k_1$, and signals are applied to input nodes x_3 and x_4 at intervals of no less than k_2, then δ and λ of the structural automata (T, c) will be determined by formulas (4). When $k_2 < k_1$, and signals are applied to input nodes x_3 and x_4 at intervals equal to k_2, the laws of operation of the sequential connection, in general, do not correspond to formulas (4). Indeed, applying input signals to nodes x_3 and x_4 at times $t + nk_2$, where $n = 0,1,2,...$, ensures that signals are applied to nodes x_3 and at times $t + nk_2$, where $n = 1,2,3...$. However, the structural automaton (T, c) outputs a signal corresponding to the input signal applied at time $t + k_2$ to the input signal at time $t + k_2 + k_1 > t + 2k_2$. Consequently, the next input signal to the structural automaton (c_1, c_1) arrived during signal processing. The automaton's response (T, c) to an input signal that arrived 'early' will not necessarily be described by formulas (4).

To solve diagnostic and control problems, it is necessary to describe the compositions of not only serviceable but also faulty structural automata. Let the structural automaton (c_2, c_2) output nodes z_3 and z_4 signals in the form of pulses. Then a signal delay at one of the nodes, for example at node z_3, will lead to the input signals $(e, pr_2\lambda_2(s_2(x_3, x_4)))$ being fed to the input nodes of the automaton (c_1, c_1). In this case, the structural automaton (T, c) described by term (3) has the following laws of operation, which differ from (4):

$$\begin{gathered} \delta((s_1,s_2),(x_3,x_4)) = (\delta_1(s_1,(e, pr_2\lambda_2(s_2,(x_3,x_4)))), \\ \delta_2(s_2,(x_3,x_4))), \\ \lambda((s_1,s_2),(x_3,x_4)) = \lambda_1(s_1,(e, pr_2\lambda_2(s_2,(x_3,x_4)))). \end{gathered} \qquad (5)$$

Thus, after identifying the first output node and the first input node by operation f^1_9, the laws of operation are not always determined by formulas of the form

$$\begin{gathered} f^1_{9,1}c = (S, X^{n-1}, X^{m-1}, f^1_{9,1}\delta, f^1_{9,1}\lambda), \\ \forall_{(s,\bar{x})}(f^1_{9,1}\delta(s,\bar{x}) = \delta(s,(pr_1\lambda(s,(e,\bar{x})),\bar{x})) \wedge \\ \wedge f^1_{9,1}\lambda(s,\bar{x}) = pr_{2...m}\lambda(s, pr_1\lambda(s,(e,\bar{x})),\bar{x}))). \end{gathered} \qquad (6)$$

$$\begin{gathered} f^1_{9,2}c = (S \times X, X^{n-1}, X^{m-1}, f^1_{9,2}\delta, f^1_{9,2}\lambda), \\ \forall_{((s,x),\bar{x})}(f^1_{9,2}\delta((s,x),\bar{x}) = \delta(s,(x,\bar{x})), pr_1\lambda(s,(x,\bar{x}))) \wedge \\ \wedge f^1_{9,2}\lambda((s,x),\bar{x}) = pr_{2...m}\lambda(s,(x,\bar{x}))). \end{gathered} \qquad (7)$$

More precisely, if the connection between the operation f^1_9 and the transformation of transition and output functions is not specified in the algebra of structural automata, then a term T containing the operation f^1_9 may describe a composition that is not an automaton. This means that the value of the term T may be undefined on the set C. Therefore, a structural automaton is defined as a pair – a term and its value. This allows us to consider compositions described by universally defined terms in the algebra of structural automata.

Let us find sufficient conditions that guarantee that each functional symbol f^1_9 in a term T corresponds to a transformation of transition and output functions according to formulas (6) and (7). The subject of analysis should be the temporal consistency of the occurrence of signals at the nodes. However, finding such admissible conditions without specifying the class of systems under study is difficult, so it is necessary to introduce restrictions on the systems under consideration.

First, let us clarify the problem using an example and outline ways to solve it. Let the structural automata (c_i, c_i), $i = \overline{1,5}$, be connected according to the diagram shown in Fig. 3.

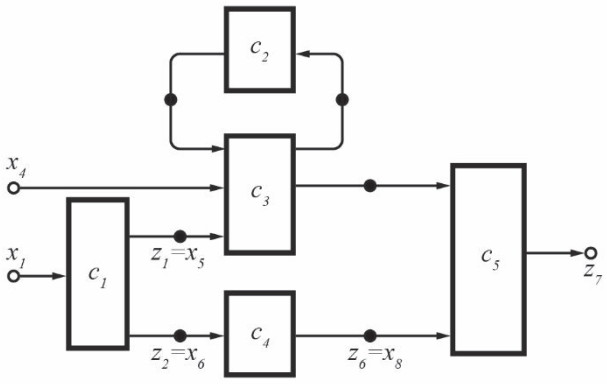


Fig. 3. Automatons composition

Suppose that the automata c_i are defined by transition and output tables. Let us express the transition δ and output λ

functions of the composition through the transition δ_i and output λ_i functions of the original elements. Let us find a term T describing the composition. Figure 2 shows the designations of the input and output nodes. The nodes that must be identified are connected by equal signs. There are six pairs of nodes that are to be identified: (z_4, x_2), (z_3, x_3), (z_1, x_5), (z_2, x_6), (z_5, x_7), (z_6, x_8). Let us find a term describing the identification of the output node z_4 and the input node x_2. Such identification is preceded by obtaining a parallel connection of the automata and shifting the input node x_2 and the output node z_4 to the places of the first nodes.

Abbreviation used

$$f_{10}^2(...(f_{10}^2(f_{10}^2 c_1 c_2)c_3)...)c_n = f_{10}^2 c_1 c_2 c_3 ... c_n \ . \qquad (8)$$

Then the identification of nodes z_4 and x_2 is described by the term $T_1 = f_{9,1}^1 (f_4^1)^3 f_2^1 f_{10}^1 c_1 c_2 c_3 c_4 c_5$. The identification of nodes z_1, x_5 and z_2, x_6 leads to the term $T_2 = (f_{9,1}^1)^2 (f_4^1)^3 (f_2^1)^2 T_1$. To describe the identification of nodes z_5, x_7 and z_6, x_8 it is sufficient to apply the operation f_4^1 and twice the operation f_9^1, the operation $f_{9,2}^1$ and f_2^1.

As a result, we obtain a term describing the composition in Fig. 3. The initial and final stages of the composition are shown in Fig. 4:

$$T = f_2^1 f_{9,2}^1 f_2^1 f_4^1 (f_{9,1}^1)^2 f_4^1 (f_{9,1}^1)^2 (f_4^1)^3 (f_2^1)^2 f_{9,1}^1 (f_4^1)^3 f_2^1 f_{10}^2 c_1 c_2 c_3 c_4 c_5 \quad (9)$$

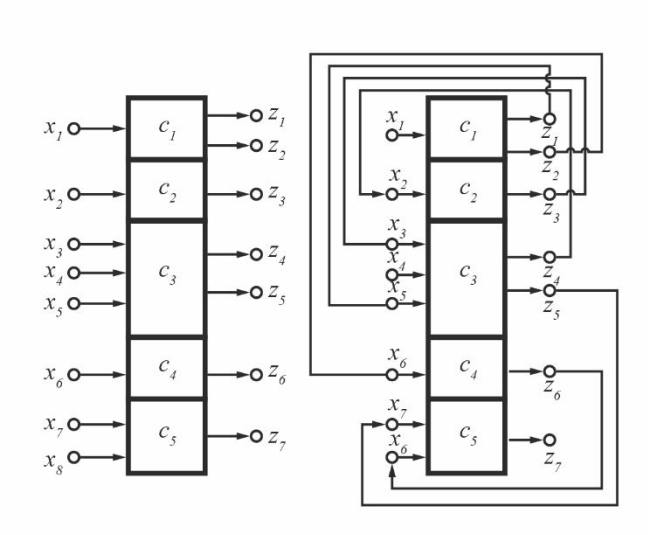


Fig. 4. The initial and final stages of the composition of automatons

In the term T, the operation f_9^1 is reflected taking into account its variants, so we can write down how the transition-output functions δ and λ the result of the composition are expressed through the transition-output functions δ_i and λ_i the initial structural automata.

For parallel connection, it is easy to write down δ and λ as state functions s_1, s_2, …, s_5, and input signals x_1, x_2,…, x_8 of the initial automata c_1, c_2,…, c_5:

$$\begin{aligned} \lambda &= \lambda(\lambda_1, \lambda_2, \lambda_3, \lambda_4, \lambda_5, (s_1, ..., s_5), (x_1, x_2, ..., x_8)), \\ \delta &= \delta(\delta_1, \delta_2, \delta_3, \delta_4, \delta_5, (s_1, ..., s_5), (x_1, x_2, ..., x_8)). \end{aligned} \qquad (10)$$

Operations f_2^1 and f_4^1 are auxiliary; they are necessary for formalizing the process of describing the structure. Their influence on the transition and output functions is reduced to swapping the places of the arguments and swapping the places of the output function values.

The transition and output functions undergo significant changes under the influence of operation f_9^1. Thus, as a result of identifying the output node z_4 and the input node x_2, we obtain

$$\begin{aligned} \lambda &= \lambda(\lambda_1, \lambda_2, \lambda_3, \lambda_4, \lambda_5, (s_1, ..., s_5), (x_1, x_3, x_4, ..., x_8)), \\ \delta &= \delta(\delta_1, \delta_2, \delta_3, \delta_4, \delta_5, (s_1, ..., s_5), (x_1, x_3, x_4 ..., x_8)). \end{aligned} \qquad (11)$$

The transformations defined by the sequence of operations $(f_4^1)^3 (f_2^1)^2$ and of an auxiliary nature are applied to the equalities (11). The identification of the output node z_1 and the input node x_5 is reduced to substituting the expression $pr_1 \lambda_1 (s_1, x_1)$ for the symbol x_5 in (11), and the identification of the output node z_2 and the input node x_6 is reduced to substituting the expression $pr_2 \lambda_1 (s_1, x_1)$ for the symbol x_6 in (11). As a result, we obtain

$$\begin{aligned} &\lambda = \lambda(\lambda_1, \lambda_2, \lambda_3, \lambda_4, \lambda_5, (s_1, ..., s_5), (x_1, x_3, x_4, x_7, x_8)), \\ &\delta = \delta(\delta_1, \delta_2, \delta_3, \delta_4, \delta_5, \lambda_1, \lambda_3 (s_1, ..., s_5), (x_1, x_3, x_4, x_7, x_8)). \end{aligned} \qquad (12)$$

Identifying nodes z_5 and x_7 with nodes z_6 and x_8 reduces the transition and output functions to the following form:

$$\begin{aligned} &\lambda = \lambda((\lambda_1, ...\lambda_5), (s_1, ..., s_5), (x_1, x_3, x_4)), \\ &\delta = \delta((\lambda_1, \lambda_3, \lambda_4), (\delta_1, ..., \delta_5), (s_1, ..., s_5), (x_1, x_3, x_4)). \end{aligned} \qquad (13)$$

Now we need to take into account the transformation under the action of the operation $f_{9,2}^1$. To do this, we need to substitute in (13) instead of the expression x_3

$$\lambda_2 (s_2, pr_1 \lambda_3 (s_3, (x_3, x_4, pr_1 \lambda_1 (s_1, x_1)))) \qquad (14),$$

which also contains the symbol x_3. Such substitutions can be made in (13) everywhere except for the substitution in the expression itself. Therefore, we will supplement the laws of operation with an equality, using which we can determine the signal in the intermediate node x_3, which allows us to characterize the signal in this node as a memory component. Finally, for the transition and output functions of the structural automaton, we obtain the following dependencies

$$\begin{aligned} &\lambda = \lambda((\lambda_1, ...\lambda_5), (s_1, ..., s_5, s_6), (x_1, x_4)), \\ &\delta = \delta((\lambda_1, \lambda_3, \lambda_4), (\delta_1, ..., \delta_5), (s_1, ..., s_5, s_6), (x_1, x_4)). \end{aligned} \qquad (15)$$

From equations (15), it follows that to determine the output signal at a node z_7, it is necessary to know the signals at nodes x_1 and x_4, the states s_1, s_2, s_3, s_4, s_5 and the signal at the intermediate node x_3. The structural automaton (T, c) in Fig. 3 has nine nodes. Consider the transformation of input signals by the structural automaton (T, c), starting at

time t. During the operation of the automaton, signals will appear in the nodes, generally speaking, in each i node at its $(t+\tau_j)$ own time.

Let us make a number of assumptions about the properties of the systems under consideration, which facilitate the study of the coordination of signal occurrence in nodes, but naturally narrow the class of real systems to which the description in the form of a structural automaton can be applied.

1. In each node, a signal occurs instantaneously at some moment of abstract time.

2. At each node, at each moment of abstract time, no more than one signal can occur (i.e., be applied, issued). This assumption does not mean that several signals cannot be applied to a node at the same time. However, in this case, a total signal must be formed, which is an element of the permissible set of signals.

The restriction defined by the following assumption is essential.

3. During the entire process of processing signals applied to external input nodes (such a process is assumed to be finite), a signal occurs only once in each node. A signal that occurs in a node at a moment of time t can be stored until a certain moment $t+\tau$, where $\tau = 0,1,2,...$

4. In each node of the structural automaton, signals of one of the following types arise: non-stored signals, which arise at a certain moment of abstract time and only at that moment; stored signals, which, having arisen at a certain moment of abstract time, remain in the node until the moment of the next signal arising in this node.

Assumptions 1-4 impose restrictions on the values τ_j characterizing the moments of occurrence of signals in the nodes of the structural automaton. In addition, restrictions on the values τ_j are imposed by formulas (15). Let us denote each node of the structural automaton (T,c) by a number coinciding with the index of the symbol x_i corresponding to this node, and node z_7 by number 9. Let us analyze formulas (15) taking into account the composition in Fig. 3.

It is obvious that $\tau_9 \geq \tau_j$, $j = 1,2,...,8$ since to determine the signal in node 9, it is necessary to know the signals in nodes 1–8. Each value of the output function, or the projection of the output function value, corresponds to the signal in the node. The values of the output function arguments are interpreted as the values of the signals in the nodes preceding the values of the signals in the nodes corresponding to the value of the output function. Therefore, from the expression $\lambda_1(s_1,x_1)$ we obtain $\tau_5 \geq \tau_1$ and $\tau_6 \geq \tau_1$.

Similarly, the following relationships are established:

$$\tau_3 \geq \tau_2,\ \tau_2,\ \tau_7 \geq \tau_3,\ \tau_4,\ \tau_5,\ \tau_8 \geq \tau_6,\ \tau_9 \geq \tau_7,\ \tau_8. \quad (16)$$

From assumption 3, it follows that the signal at node 3 occurs during signal processing only once at the moment $t+\tau_3$, where $\tau_3 \leq \tau_2$.

The relationship $\tau_3 \geq \tau_2$ is related to the change in memory state, since the signal at node 3 is considered as a component of the memory state vector. As a result of the analysis, we obtain two systems of relationships:

$$\tau_5,\ \tau_6 \geq \tau_1,\ \tau_7 \geq \tau_3,\ \tau_4,\ \tau_5,\ \tau_8 \geq \tau_6,\ \tau_9 \geq \tau_7,\ \tau_8, \quad (17)$$

$$\tau_5,\ \tau_6 \geq \tau_1,\ \tau_2,\ \tau_7 \geq \tau_3,\ \tau_4,\ \tau_5,\tau_8 \geq \tau_6,\ \tau_9 \geq \tau_7,\ \tau_8,\ \tau_3^* \geq \tau_2. \quad (18)$$

IV. CONCLUSION AND FUTURE WORK

The system of inequalities (17) provides rough estimates for the moments of occurrence of signals in nodes during the formation of the output signal. The system of inequalities (18) provides estimates for the moments of occurrence of signals in nodes during the formation of the next state. These two systems do not contradict each other.

Thus, formulas of the form (15) impose restrictions on the moments of occurrence of signals in the nodes of the structural automaton (T,c). However, the systems of inequalities (17) and (18) do not determine specific values of τ_j, $j = \overline{1,9}$. The determination τ_j is related to the consideration of specific objects described by the structural automaton (T,c), or the addition of new restrictions.

Suppose that for some real object, numbers τ'_j, $j = \overline{1,9}$, satisfying the systems of inequalities (17), (18) are defined, and the transition and exit functions of the structural automaton describing this object are determined by formulas of the form (15).

Let a malfunction occur in the object, expressed by the fact that at a certain node, for example node 7, a signal occurs at time $\tau''_7 \neq \tau'_7$. Even if the parts of the object are again described by structural automata (c_i,c_i), $i = \overline{1,5}$, and their composition by term T, the transition and output functions of the entire composition may not be defined by formulas of the form (15). The signal at node 9 (at the output z_7) at $\tau''_7 \neq \tau'_7$, depending on the properties of the real object, can be determined not only by formulas (15), but also by three other possible formulas:

$$\begin{gathered}\lambda = \lambda'((s_1,...,s_5,s_6),(x_1,x_4)),\\ \lambda = \lambda''((s_1,...,s_5,s_6),(x_1,x_4)),\\ \lambda = \lambda'''((s_1,...,s_5,s_6),(x_1,x_4)),\\ \lambda' \neq \lambda'' \neq \lambda'''.\end{gathered} \quad (19)$$

In some cases, the occurrence of a signal in node 7 ‘at the wrong time’ can be described by applying certain symbols and to nodes 7 and 8, x'_7 and x'_8 in some cases it cannot be described at all by the selected means of description.

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Data Science/ML for Strength Prediction in Electric Resistance Butt Welding with Sparse Data

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***Abstract*—This study focuses on the application of Data Science and Machine Learning (ML) techniques to predict joint strength in electric resistance butt welding of steel open-sections used in metal-plastic window systems. The main goal is to identify and model the relationships between welding parameters (electrical current strength, welding time, and strip thickness) and the resulting weld seam strength. Given the sparse nature of the available experimental data, advanced data preprocessing was performed using Python libraries such as Pandas and Scikit-learn, including noise reduction and error correction. Several models were developed, including second- and fourth-degree polynomial regressions and a multilayer perceptron (MLP), to capture nonlinear dependencies between process variables. The fourth-degree polynomial regression model proved to be the most accurate among traditional approaches, while the MLP achieved even higher prediction performance. These models enable not only accurate strength forecasting, but also support the intelligent selection of welding parameters in production environments with constrained data. The presented approach lays the groundwork for integrating adaptive ML systems into welding process control, facilitating real-time decision-making and technological optimization.**

Keywords—Data Science, machine learning, predictive modeling, electric resistance welding, sparse data, polynomial regression, MLP, process optimization, Python, Scikit-learn.

I. Introduction

In the production of reinforcing bent open-sections for metal-plastic windows, butt welding of strip material is used to improve material efficiency [1, 2]. These profiles, which serve as structural elements of window frames, must meet increased strength requirements. For their production, galvanized steel from coils of varying thicknesses (t) is used. To reduce waste, the end of one coil is butt-welded to the beginning of the next [3, 4]. The durability of this weld is critical and depends on parameters such as welding current (I), heating time (τ), strip width, surface cleanliness, and butt-end preparation [4–6]. To determine optimal welding settings, experiments with varying parameters were conducted, and the results were analyzed using regression modeling and machine learning. In many practical cases, the selection of welding parameters and prediction of seam strength must be performed under limited data conditions, which necessitates the use of adaptive and data-efficient modeling approaches.

II. Review of the Latest Research and Publications

Various methods for producing reinforced open-sections, such as U-channel profiles, include extrusion or die bending [7]. These methods rely on precise trimming and adequate pressing force (P), along with proper die design, to reduce deformation errors [8, 9]. Roll-forming, which involves multi-pass bending of strip material, causes significant plastic deformation and requires careful control of upsetting depth during butt welding [10, 11]. Copper-electrode welding machines are used and demand proper edge preparation and surface cleaning. Fig. 1 shows a typical contact electric butt-welding machine.

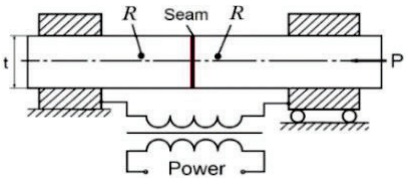


Fig.1. Electric butt-welding machine.

Maintaining edge parallelism is essential, especially as strip width increases. Proper adjustment of the welding machine involves balancing electrical and mechanical systems, using experimental methods to optimize parameters under various influencing factors [12]. During the operation of the butt-welding machine, the ends of the strips are heated by the flow of electric current I_w through them. This current, flowing through the total electrical resistance R of the strip ends, generates heat Q, which is regulated by Joule-Lenz's law [13]:

$$Q = I_w^2 * R * \tau, \quad (1)$$

where τ – welding time.

The total electrical resistance of the joining ends is determined by the expression:

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$$R = 2R_p + R_c, \quad (2)$$

where R_p – The resistance of the strip ends (the protrusion of the strip ends from the machine electrodes) is determined by the expression, R_c – the contact resistance between the ends of the strips.

During the high-temperature regime, when the ends of the strips are heated to 600-700 °C, the contact resistance is relatively high. However, once this temperature is reached, the contact resistance is practically eliminated, and further heating of the components occurs due to their inherent electrical resistance. The presence of contact resistance is associated with the limited area of electrical contact due to the uneven surface of the components, as well as the presence of oxide and hydroxide films, moisture, oils, corrosion products, dust, and other contaminants. The contact resistance of the components decreases with an increase in upsetting (compression stage increasement).

The quality of the welded joints of wide strips was checked both destructively (through external inspection) and destructively after roll-forming [4]. For destructive testing, samples were cut from the flanges and web of the roll-formed U-section and subjected to tensile testing [4]. On the fig.2 a scheme of samples retrieving is depictures including the locations and form of samples cut from flanges and base.

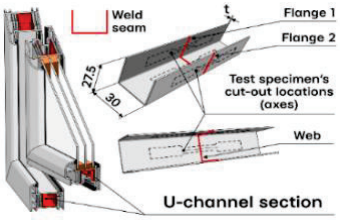


Fig. 2. Sample cutting scheme.

The limited amount of data requires finding suitable equations but the challenge in it lies in the high likelihood of overfitting, the limitation of model complexity despite the potential complexity of the actual dependency, high sensitivity to random outliers and deviations, and a critical dependence on the selected parameters, especially in cases where the relevance of each is unknown [14–15]. In this regard, an important task involves detecting and removing noise from experimental data using machine learning methods like collective, single, and distance-based approaches (including density and clustering). Literature mentions ignoring, filtering, or substituting noise, aiming to balance noise removal and data retention. Denoising was done using classic anti-aliasing methods [16].

III. The Purpose of the Work

The purpose of the study is to perform analysis and computer modeling to derive regression equations linking the strength of steel U-sections to welding process parameters, based on experimental data.

IV. Description of the Experiment

This experiment aimed to figure out the dependence of the strength of the reinforcing steel section on the parameters of the welding process. During the study, samples with transverse welds, cut from the base and flanges of the U-section, as well as reference samples without welds, were used. Testing was conducted using a universal testing machine (nominal force of 20 tons). The strength of each flange and the base was evaluated separately under different geometric conditions at the edge joints, which became more non-parallel as the width of the initial strip increased. These elements may show varying strength characteristics, even when welded using the same equipment settings.

The collected data were documented in a table for subsequent analysis [4]. The testing revealed that the strength of the weld joint is significantly influenced by the welding parameters and the duration of load application. Furthermore, even with consistent welding parameters, the strength varied with the thickness of the weld joint. This observation is supported by the fact that some samples failed along the weld seam, while others failed within the body of the sample, indicating either the adequacy or inadequacy of the selected welding parameters.

V. Processing Data with Data Analysis Methods

The experimental data were not suitable for regression modeling using standard Data Mining methods (Fig. 3). Due to poor structure, the data required cleaning and transformation to enable pattern recognition. Processing involved multiple stages, including error detection and noise reduction. A major challenge was the limited data volume and the inability to perform further experiments on the same equipment. Thus, methods resilient to noise and effective with small datasets were necessary.

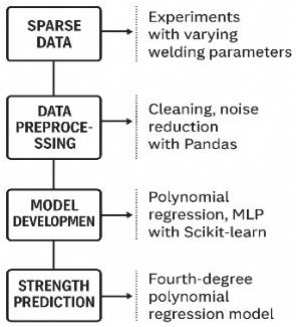


Fig. 3. Research flow diagram.

The Pandas library in Python was used as the computational platform [17]. It enabled initial data cleaning and deeper analysis without relying on external mathematical packages. The data were structured in tabular form and imported as a CSV file. Analysis results were visualized using 2D plots showing the relationship between parameters and strength coefficient. Fig. 4 illustrates the dependence of strength coefficient on weld seam thickness, excluding other factors such as time and current.

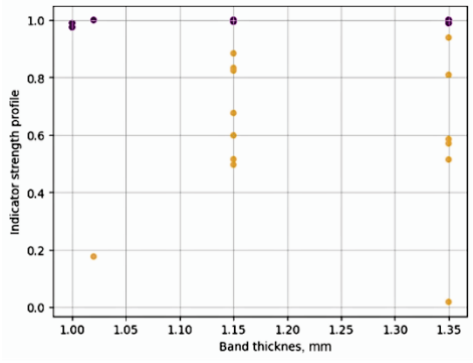


Fig 4. Diagram of the dependency of the quality of the weld seam.

Based on the data comparisons, it was determined that the dataset doesn't show a direct functional dependence on a

single parameter. Consequently, there a need to formulate hypotheses appeared, on which regression analysis methods were applied to uncover patterns within the dataset.

The initial hypothesis proposed was that a logistic regression could elucidate the dependence of strength on one or more process parameters. Given that logistic regression requires one or more parameters to vary discretely, it was most likely to identify a pattern between the strength coefficient and the welding parameters [18]. A logistic regression curve was constructed for the welding parameters, as illustrated in Fig. 5.

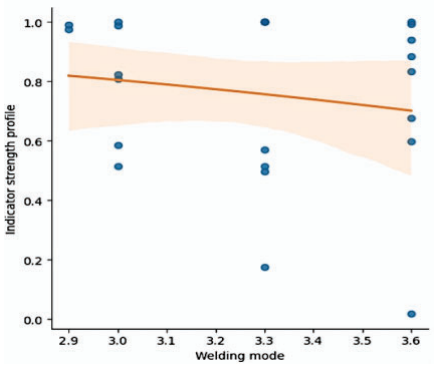


Fig. 5. Logistic regression curve for the dependence of the strength parameter on the welding mode.

To build regression models and visualize 2D dependencies, the Seaborn library in Python was used [19], offering built-in regression tools. In Fig. 4. blue dots represent experimental data, the red line shows linear regression, and the orange area indicates the uncertainty range. The resulting curves showed noticeable discrepancies—for instance, logistic regression closely resembled the linear model.

Weld joints produced under different conditions revealed distinct patterns between welding parameters and structural strength. In some cases, analyzing individual joints yielded more accurate equations than models based on the full dataset. Fig. 6 presents such a logistic regression curve, showing closer alignment with the data. However, the broader uncertainty range suggests the need for a better-suited model.

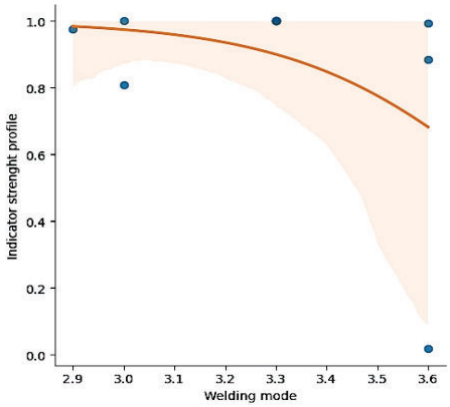


Fig. 6. Logistic regression curve for the dependence of the strength parameter on the welding mode (separately for flange 1).

Based on the performed analysis was established that isolation of the parameters doesn't allow to describe process more precisely and calculate an optimal mode for the current technological process. Therefore, it was decided to build an approximate regression equation based on all parameters.

In other words, to find regressions for the existing dataset. Since the process of finding an approximate regression equation and refining the coefficients of the variables in the equation is quite complex, it was decided to use machine learning tools, specifically the scikit-learn library in the Python programming environment [20].

Recognizing that any dependence can be approximated through regression, a linear regression model was built based on experimental data. The modeling was carried out using the scikit-learn library, allowing construction and evaluation of models with varying complexity. PolynomialFeatures generated higher-order terms, while LinearRegression fitted the transformed data. Additionally, a simple feedforward neural network was trained using scikit-learn's MLPRegressor. Its architecture and hyperparameters – number of layers, activation functions, learning rate, and regularization – were optimized via cross-validation and grid search. The neural network outperformed traditional models in capturing nonlinear patterns. Seaborn and Matplotlib supported visual analysis and result validation. This programmatic workflow ensured reproducibility, scalability, and objectivity, positioning the study as both an engineering investigation and an applied data science case.

The model in Fig. 6 was found inadequate for accurately describing the process. Despite parameter optimization, the coefficient of determination revealed notable discrepancies between predicted and actual values. To achieve more accurate analysis of the strength coefficient, a hypothesis was tested suggesting that its dependency on process parameters could be represented by a polynomial regression.

Using polynomials of varying degrees allows greater modeling flexibility and better prediction by accounting for factor interactions. To evaluate this hypothesis, polynomial regressions of degrees 2 and 4 were tested. The general form of a second-degree polynomial with three independent variables is given below:

$$y = \beta_0 + \beta_1 x_1 + \beta_2 x_2 + \beta_3 x_3 + \beta_4 x_1^2 + \beta_5 x_2^2 + \\ + \beta_6 x_3^2 + \beta_7 x_1 x_2 + \beta_8 x_1 x_3 + \beta_9 x_2 x_3, \quad (3)$$

where β_0 – constant term (intercept), β_1 to β_9 – coefficients calculated by machine learning model, x_1 –weld seal thickness, x_2 – welding mode, x_3 – welding time.

With the help of sub library scikit-learn-preprocessing [21], particularly functions of preliminary optimization of polynomial model a polynomial regression model for second degree was built in order to describe the dependency of structural strengths coefficients based on three independent parameters.

$$y = 0.18507 + 7.633x_1 + 9.0942x_2 - 5.0855x_3 + \\ + 0.0122x_2^2 - 0.0006x_3^2 - 0.06x_1x_2 + 0.02x_1x_3 - \\ - 0.0156x_2x_3. \quad (4)$$

Mean squared error is 0.0173, which conforms the adequacy of the created model. In addition to this equation, a 4th-degree regression equation was constructed. The general form for 3 independent variables is as follows:

$$y = \beta_0 + \beta_1 x_1 + \beta_2 x_2 + \beta_3 x_3 + \beta_4 x_1^2 + \beta_5 x_2^2 + \beta_6 x_3^2 + \\ \beta_7 x_1 x_2 + \beta_8 x_1 x_3 + \beta_9 x_2 x_3 + \beta_{10} x_1^3 + \beta_{11} x_2^3 + \beta_{12} x_3^3 + \\ \beta_{13} x_1^2 x_2 + \beta_{14} x_1^2 x_3 + \beta_{15} x_2^2 x_1 + \beta_{16} x_2^2 x_3 + \beta_{17} x_3^2 x_1 + \\ \beta_{18} x_3^2 x_2 + \beta_{19} x_1 x_2 x_3 + \beta_{20} x_1^4 + \beta_{21} x_2^4 + \beta_{22} x_3^4 + \\ \beta_{23} x_1^3 x_2 + \beta_{24} x_1^3 x_3 + \beta_{25} x_2^3 x_1 + \beta_{26} x_2^3 x_3 + \beta_{27} x_3^3 x_1 + \\ \beta_{28} x_3^3 x_2 + \beta_{29} x_1^2 x_2^2 + \beta_{30} x_1^2 x_3^2 + \beta_{31} x_2^2 x_3^2 + \beta_{32} x_1^2 x_2 x_3 + \\ \beta_{33} x_1 x_2^2 x_3 + \beta_{34} x_1 x_2 x_3^2. \quad (5)$$

After conducting machine learning, the trained model proposed the following regression equation:

$$y = 0.382929 + 0.00004011x_1 + 0.000001x_2 - 0.0000147x_3 + 0.00001x_2^2 - 0.0000199x_3^2 - 0.00008x_1x_2 - 0.00009x_2x_3 + 0.0000224x_2^3 - 0.0000269x_3^3 + 0.00019x_1^2x_2 + 0.0000113x_1^2x_3 - 0.000146x_2^2x_1 + 0.000404x_2^2x_3 + 0.000215x_3^2x_1 - 0.00002x_3^2x_2 \pm 0.00003x_1x_2x_3 + 0.0000321x_2^4 - 0.000364x_3^4 + 0.0001x_1^3x_2 + 0.000016x_1^3x_3 - 0.000167x_2^3x_1 + 0.000545x_2^3x_3 + 0.000292x_3^3x_1 - 0.0000309x_3^3x_2 - 0.000418x_1^2x_2^2 + 0.00173x_1^2x_3^2 + 0.00133x_2^2x_3^2 + 0.000842x_1^2x_2x_3 + 0.000258x_1x_2^2x_3 - 0.000523x_1x_2x_3^2. \quad (6)$$

Evaluating the mean squared error, which is 0.1703, and the coefficient of determination, it was determined that the 4th-degree regression equation better fits to the data than the proposed models of degrees 2, 3, and 5, and is also significantly more accurate than the linear regression equation. Based on the created regression equations and the fact that the 4th-degree polynomial regression most accurately describes the relationship between the strength coefficient and the welding parameters, it was assumed that the dependency might be more complex. Therefore, it is planned to further investigate the patterns using more complex regression equations and their combinations, as the shape of the dependency curve may be described by multiple functions

VI. Conclusion

This study developed mathematical models to assess how butt welding parameters affect the strength of U-shaped steel profiles. Based on experimental data, two regression models were built: second-degree polynomial regression and a multilayer perceptron (MLP) neural network. Both captured general trends – strength increases with higher welding energy up to an optimum and decreases beyond due to defects. The MLP outperformed the polynomial model, achieving ~95% accuracy versus ~88%, owing to its ability to model complex nonlinear interactions. Among the key parameters, welding current and upset pressure had the strongest influence. Both showed optimal ranges – too low caused incomplete fusion; too high led to overheating and structural weakening. The MLP model accurately reflected these effects and allows rapid strength prediction, making it suitable for real-time decision support in production. The model can also aid in offline optimization by finding parameter combinations that maximize joint strength. With further expansion of the dataset and integration of additional quality indicators, the model could be generalized to other profile types and materials, supporting smarter, data-driven manufacturing.

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Digital Control of a Plasma Torch Power Supply System

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***Abstract*—This paper validates a linearized discrete-time model of a switched-mode power converter operating with an arc load, and details a methodology for determining its parameters. These parameters are essential for the analysis and synthesis of the output current control loop. The paper also discusses the analytical synthesis of a digital PI controller within a switched-mode power supply.**

Keywords—switched-mode power converter, current control loop, discrete-time mathematical model, dynamic analysis, controller synthesis.

I. Introduction

Most modern power electronic devices, from the perspective of automatic control theory, are nonlinear discrete-time systems [1].

However, the application of modern power electronics (converter equipment) is based on utilizing the "useful" component of the output pulses, which varies depending on the duty cycle *d*. In addition to this useful component, the output voltage (current) of these devices contains undesirable but unavoidable high-frequency ripple. At a sufficiently high switching frequency in these devices, the ripple in phase coordinates, such as in switched-mode current regulators, becomes negligibly small. This provides a basis for considering the analyzed real discrete systems as continuous by using their limiting continuous model [1], which is constructed under the assumption of infinite switching frequency ($T \to 0$) and the absence of phase coordinate ripple.

It's important to emphasize that the intense development of microprocessors and microcontrollers, along with improvements in their reliability and manufacturability, has led to the widespread adoption of digital control systems (DCS) in the practical implementation of switched-mode converters. These systems provide higher control quality, offer great flexibility and adaptability to changes in technological processes, enable remote diagnostics and parameter tuning, facilitate a rapid development cycle, and allow for software updates during operation. All these factors significantly enhance the economic efficiency of their application [2-4].

The task of DCS synthesis essentially involves defining the algorithm for the digital controller that implements the derived control law. Various methods are used to solve this problem [2–4], but the most efficient ones appear to be analytical synthesis methods. They allow for obtaining both the structure and parameters of the control law by solving specific equations.

The purpose of this study is a step-by-step presentation of the design stages for a digital negative feedback (NFB) in plasma arc power supplies. This includes the synthesis of a digital controller that ensures a specified feedback gain, a desired order of astatism, and defined performance indices, while maintaining reasonable stability margins in terms of amplitude and phase.

The synthesis of PI controller parameters will be considered below using the example of a plasma arc power supply based on a phase-controlled transistor converter, which allows for soft switching in a full-bridge topology [5]. For the presented synthesis methodologies, the adequacy of the discussed methods for modern microcontroller-based systems will be demonstrated.

II. Controlled plant characteristic

In DC-DC converters, the phase coordinates and consequently the output signals consist of a smooth (average) component and high-frequency ripple caused by the switching mode of transistor operation. When a sufficiently high frequency is chosen (at least an order of magnitude higher than the cutoff frequency of the converter's output filter), changes in the system structure and the amplitude of pulsations are small. It's evident that as the switching period T decreases ($T \to 0$) and $D = 2t_p / T = \text{const}$, the steady-state mode approaches to a limiting steady-state mode, where ripple in the phase coordinates is absent. Such a model is referred to as the limiting continuous model. In most real cases, due to the high switching frequency, the ripple in the phase coordinates is sufficiently small and occurs around the phase coordinates of the limiting continuous model, corresponding

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to $T \to 0$. It should be noted that in this case, the change in the duty cycle $\widetilde{d}$ is considered as parametric disturbance, not a signal-based one [1].

Representing a DC-DC converter as a continuous model allows for studying its properties and making changes to the electrical circuit before building a physical prototype. This approach also facilitates the synthesis of controllers and adjustment of their parameters, thereby reducing the overall design time.

Soft-switching phase-shifted full-bridge converter, operating with an arc load in small-signal mode, can be described by the following model (transfer function (TF) "arc current – control") [6, 7, 9]

$$W_L(s) = W_{i/u}(s) = \frac{\widetilde{I}(s)}{\widetilde{U}(s)} = k_0 \exp(-T_\mu s)/(\tau s + 1), \qquad (1)$$

where $k_0 = (R_{CS}/(r_\Sigma + R_{diff\,0}))n_{TV}U_{in}k_{PWM}k_i$ is the gain of the invariable part, without considering the ripple factor; $\tau = L/(r_\Sigma + R_{diff\,0})$ is the load circuit time constant; T_μ is the converter's delay time and T_μ are determined in the neighborhood of the nominal operating mode; R_{diff0} is the differential resistance of the plasma arc; L is the total inductance of the circuit; r_Σ generally comprises all losses in the circuit, including the active resistance of the inductor and the internal resistance of the converter $R_d = 4n_{TV}^2 f_s L_s$ [6, 7], moreover, this resistance caused by losses may be frequency-dependent; R_{CS} is the transconductance of the current sensor, having units of resistance; n_{TV}, as usual, is given by $w_{21}/w_1 = w_{22}/w_1$. The transfer function $W_L(s)$ includes the gain coefficients of the power stage, the current sensor, the sensor's signal amplifier (k_i) and the pulse-width modulator; $k_{PWM} \approx T/U_m$, where U_m – is the amplitude of the sawtooth signal; T is the sampling period.

Here, the symbol "~" denotes an infinitesimal change of the variable relative to its value in the periodic steady-state mode.

Often, if the condition $T_\mu/\tau \le 1$ holds for model (1), then a corresponding model can be associated with it:

$$W_L(s) = k_0/[(\tau s + 1)(T_\mu + 1)] \approx \frac{k_0}{(\tau s + 1)^l}, \qquad (2)$$

where $T_\mu = T_{PWM} = 19.2 \cdot 10^{-6}$ sec is the time constant, equal to the PWM period, which follows from the transition

$$\lim_{l\to\infty} \frac{1}{(\tau s + 1)^l} = \lim_{l\to\infty} (T_\mu s/l + 1)^{-l} = \exp(-T_\mu s).$$

The transition from one model to another can be made based on the data below:

$$for\, l = 2\, from\, (2) \to T_\mu/\tau = 0.1\, by\, (1).$$

When designing control systems, it's necessary to identify the parameters of models (1) and (2). Furthermore, identifying parameters such as l, k_0, and τ for model (2) allows, if needed, a transition to model (1).

Let's estimate the values of the gain coefficient k_0 and the time constant τ in the converter [6, 7] at nominal parameter values: $R_{CS} = 75 \cdot 10^{-5}$ Ohm; $r_\Sigma = 0.01$ Ohm; $L = 300$ μH; $U_{in} = 540$ V – the input voltage of the regulator, corresponding to the selected periodic operating mode, in the neighborhood of linearization is performed; $n_{TV} = 0.46$; $U_m = 2.5$ V; $k_i = 82$; $f = 52$ kHz; $R_{diff0} = 1.2$ Ohm; $D = 0.8$ – the duty cycle value in the steady-state (periodic) mode. Then:

$$k_0 = \frac{R_{CS} T n_{TV} U_{in}}{(r_\Sigma + R_{diff\,0})U_m} = \frac{75 \cdot 10^{-5} \cdot T \cdot 0.46 \cdot 540}{(0.01 + 1.2) \cdot 2.5} = 5.083T;$$

$$\tau = L/(r_\Sigma + R_{diff\,0}) = 248 \cdot 10^{-6} \text{ sec}.$$

One notable and particularly interesting method for synthesizing control laws involves representing the power circuits of PWM regulators in a small-signal deviation mode using an adequate pulse-amplitude model (PAM) based on the regulated components of the processes.

Transforming pulse-width modulation (PWM) to pulse-amplitude modulation (PAM) in the neighborhood of the steady-state mode allows the use of the well-developed theoretical framework of PAM systems [6, 7, 10] for the analysis and synthesis of a current pulse regulator with PWM and to derive pulsed control laws based on it.

The transition from PWM to natural PAM is performed provided the pulse areas are equal [1]: $U_{in}\widetilde{d}T = \widetilde{u}_{in}DT$.

Then the law of pulse amplitude variation as a function of $\widetilde{d}$ is as follows: $\widetilde{u}_{in} = U_{in}\widetilde{d}/D$. It is valid provided that deviations of $\widetilde{d}$ from D are insignificant and the sampling frequency is much higher than the natural frequency of the filters [1].

In this case, the gain $k_u = U_{in}/D$ also pertains to the linear continuous part of the system (it is included in the overall gain k_0).

To synthesize the control law for a switched-mode current regulator (switched-mode current converter – SMCC) with pulse-width modulation (PWM), we utilize an adequate pulse-amplitude model (PAM) based on the regulated components of the PWM processes. This model relies on separating the overall process into a stationary component and a pulse-width regulation process [1, 6, 7]. The power circuit of the SMCC is detailed in [5].

In this case, the discrete transfer function (DTF) for the invariable part (IP) of the synthesized system [4] can be written as

$$W(z) = Z\left\{\frac{W_0(s)}{s}\right\} - z^{-1}Z_\varepsilon\left\{\frac{W_0(s)}{s}\right\},$$

where $\varepsilon = 1 - D$; $D = 2t_p/T$; T is the switching period; and Z_ε denotes the modified Z-transform.

Let the invariable part of the system have the transfer function (1). Consider the case where $T_\mu = 0$. Then, in accordance with (2) and the appendix in [4], we obtain:

$$W(z) = k_0\left[\frac{(1-d)z}{(z-1)(z-d)} - \frac{1}{z-1} - \frac{d^{\varepsilon}}{z-d}\right] =$$

$$= k_0\frac{d^{\varepsilon}-d}{z-d} = k_0\frac{d(d^{-D}-1)}{z-d}, \qquad (3)$$

where $d = e^{-\alpha T}$; $\alpha = \tau^{-1}$.

After substituting the numerical values, we have:

$$W(z) = \frac{0.3901}{z-0.9226}.$$

We will plot the frequency characteristics based on (1) and (3). The results are presented in Fig. 1. The characteristics show that, in the discrete version, without any frequency compensation and with $k_{CR} = 1$, the phase characteristic sharply decreases after 1 kHz compared to the analog counterpart, and the phase lag reaches 180° at half of the PWM frequency. Thus, the true phase characteristic of a pulsed system, when compared to a continuous one, typically presents a pessimistic view during the synthesis of a digital control system.

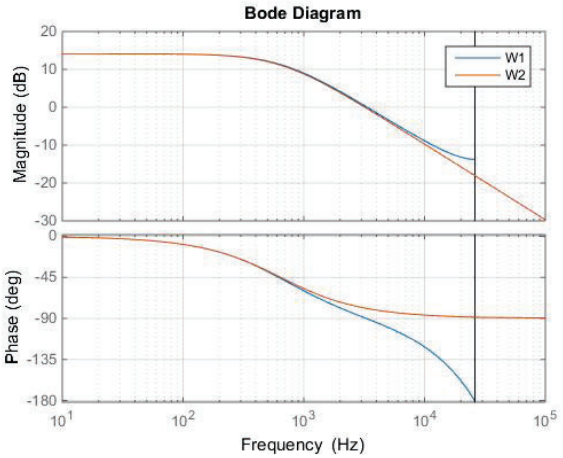


Fig. 1. Loop gain Bode plots of the analog and digital converter (frequency compensation is absent)

III. Retrofitting the Control System with a Microcomputer (Microprocessor)

Instead of undertaking an entirely new design using digital control system theory, a microcomputer-based retrofitting methodology can be applied to create an equivalent digital system [4]. A digital system is considered equivalent to a continuous system if the responses of the two systems align well for the same input signals and initial conditions.

We will set the sampling frequency equal to the switching frequency.

To create the discrete transfer function (DTF) of the integral channel (i.e., retrofitting the PI controller), we use the Euler integration method (or rectangular method), which involves the substitution $s \leftarrow (z-1)/T$, yielding

$$R_i(s) = \frac{k_i}{s} \rightarrow R_i(z) = \frac{k_i T}{z-1}.$$

With the given numerical values $k_p = 1.271$, $k_i = 5123$, obtained based on the desired (reference) transfer function of the open-loop system, and $T = 19.2\cdot10^{-6}$ sec, we get:

$$R_{pi}(z) = \frac{k_p z - (k_p - k_i T)}{z-1} =$$

$$= k_p\frac{z-\left(1-\frac{k_i}{k_p}T\right)}{z-1} = 1.271\frac{z-0.9226}{z-1}.$$

Bode plots of the pulsed system with the digital controller (open-loop compensated system) for $k_p = 1.271$ are presented in Fig. 2. Note that the gain and phase margins now equal 18.4 dB and 65.8 degrees, respectively.

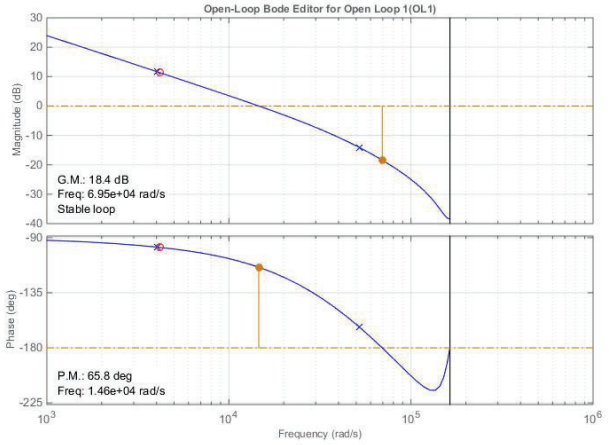


Fig. 2. Bode plots of the discrete arc current control system

The system's step response is shown in Fig. 3. The system settles within a time equal to 9–10 sampling periods. The maximum overshoot does not exceed 2 %, which confirms the correct selection of a 65.8 degrees phase margin.

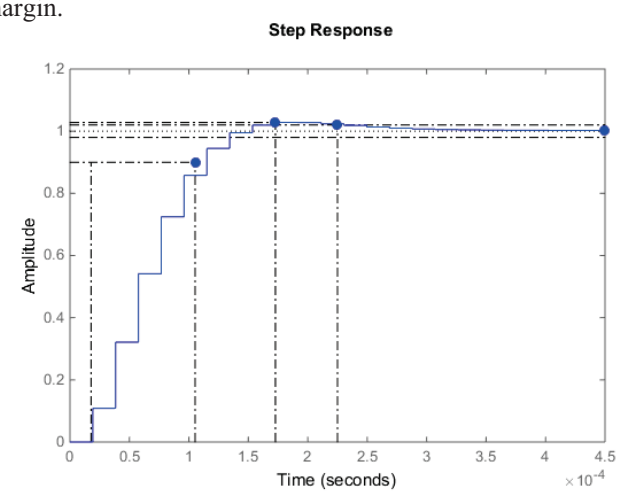


Fig. 3. Step response of the synthesized system

To achieve a larger phase stability margin, the value of k_p can be reduced, for instance, by setting $k_p = 0.6355$.

Fig. 4 shows the plots of the sensitivity function $|S(j\omega)|$ and the robustness measure $|CS(j\omega)|$ for the arc current

control system. It is evident that sensitivity reduction is achieved at low frequencies, up to $4 \cdot 10^3$ rad/s.

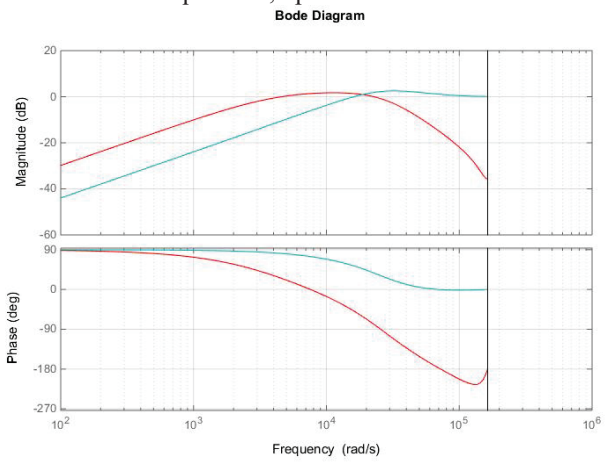


Fig. 4. Bode plots of the sensitivity function and robustness measure for the arc current control system

Note that a small value of $S(j\omega)$ is achieved with a high dynamic gain of the control system's $R(j\omega)$ $W_\partial(j\omega)$ chain in the corresponding frequency range.

IV. DACS SYNTHESIS WITH FINITE SETTLING TIME

Let the serial digital controller for the controlled plant [6, 7] have the transfer function:

$$R(z) = \frac{\frac{1}{k_0(1-e^{-T/\tau})}z - \frac{e^{-T/\tau}}{k_0(1-e^{-T/\tau})}}{z-1} = \\ = \frac{z-d}{k_0(z-1)(1-d)} = A\frac{1-dz^{-1}}{1-z^{-1}}. \qquad (4)$$

Then, the open-loop transfer function of the compensated system takes on a simpler form:

$$W^*(z) = R(z)W_L(z) = \frac{1}{z-1}.$$

The corresponding closed-loop transfer function can be written as:

$$\Phi^*(z) = z^{-1}.$$

This means that for a step input, the system's output signal, $i(nT)$, reaches the desired steady-state value within one sampling period and maintains this value thereafter. The overshoot of $i(nT)$ is zero. The actual response $i(t)$ may exhibit ripple between the switching instants. This type of response is commonly referred to as an aperiodic transient process [4].

Let's determine the gain and phase margins of the pulsed control system. For this, consider the characteristic equation of the closed-loop system:

$$z - 1 + A = 0.$$

The stability condition is:

$$0 < A < 2.$$

Therefore, the gain margin equals 6 dB.

Let's convert to the frequency TF of the open-loop system by substituting $z = e^{j\omega t}$.

$$W(e^{j\omega t}) = \frac{1}{e^{j\omega t}-1} = \frac{1}{\cos\omega T - 1 + j\sin\omega T}. \qquad (5)$$

We'll determine the crossover frequency for transfer function (5) from the equality:

$$\left|\frac{1}{e^{j\omega_{\text{cutoff}}T}-1}\right| = 1,$$

From this, we get:

$$\omega_{\text{cutoff}} = \pi/3T,$$

$$\arg W(j\omega_{\text{cutoff}}) = \arg\frac{1}{e^{j\pi/3}-1} = -\frac{2\pi}{3} = -120^\circ.$$

Thus, the phase margin will be 60 degrees.

With $T = 19.2$ μs and parameters from [6, 7], the discrete filter or digital computing unit (DCU) has the transfer function:

$$R(z) = 2.542\frac{z-0.9226}{z-1}.$$

From the derived expression, it is evident that the discrete filter closely resembles a PI controller [6, 7]. The only difference is observed in the coefficients. The proportional coefficient is

$$k_P = \frac{1}{k_0(1-e^{-T/\tau})} = \frac{1}{k_0(1-d)} = \frac{1}{5.082(1-0.9226)} = 2.542,$$

and the integral coefficient is

$$k_I = \frac{k_P - \frac{e^{-T/\tau}}{k_0(1-e^{-T/\tau})}}{T} = \frac{k_P - \frac{d}{k_0(1-d)}}{T} = \\ = \frac{2.542 - \frac{0.9226}{5.082\cdot(1-0.9226)}}{19.2\cdot10^{-6}} = 10208,$$

which are approximately twice as large as those calculated by the classical method.

If the gain of the continuous part is chosen such that $A = 1$, or $k_0 = (1-d)^{-1}$, then expression (4) simplifies to:

$$R(z) = \frac{1-dz^{-1}}{1-z^{-1}}.$$

From this, the DCU program is:

$$x_0[n] = e_0[n] - de_0[n-1] + x_0[n-1],$$

where $e_0[n]$ and $x_0[n]$ are the input and output lattice functions of the DCU, respectively. Here $e_0[n] = g_0[n] - i_0[n]$, where g_0 and i_0 are the digital representations of the reference input and the controlled variable, respectively.

As a second example, let's find the discrete compensator transfer function that satisfies the given transfer function:

$$\Phi^*(z) = \frac{0.5z + 0.5}{z^2} = 0.5z^{-1} + 0.5z^{-2}. \quad (6)$$

The desired DTF of the DCU, $R(z)$, is:

$$R(z) = \frac{Q(z)}{P(z)} \cdot \frac{\Phi^*(z)}{1 - \Phi^*(z)},$$

where

$$P(z) = k_0(1-d);$$

$$Q(z) = z - d \ [6, 7].$$

In this case, we'll have:

$$R(z) = \frac{U(z)}{E(z)} = \frac{0.5k_0}{1-d} \cdot \frac{z^2 - (d-1)z - d}{z^2 - 0.5z - 0.5}.$$

Consequently, the equation of the discrete compensator in the Z-domain will be:

$$(1-d)(z^2 - 0.5z - 0.5)U(z) =$$

$$= 0.5k_0[z^2 - (d-1)z - d]E(z)$$

or in the time domain

$$(1-d)[u(n+2) - 0.5u(n+1) - 0.5u(n)] =$$

$$= 0.5k_0[e(n+2) - (d-1)e(n+1) - de(n)]. \quad (7)$$

Thus, for the example under consideration, the control law is a second-order difference equation (7). It can be verified that the output variable *i(n)* of the closed-loop system with the determined discrete compensator coincides for $n = 1, 2, \ldots$ with the assigned trajectory $i^*(n)$, which is the output variable of the reference system with transfer function (6).

It should be noted that the transfer function $\Phi^*(z)$ in the example considered also corresponds to a process with finite duration: the transient response to a unit step function ends within two sampling periods.

Fig. 5 shows the system's step response created using MATLAB [8].

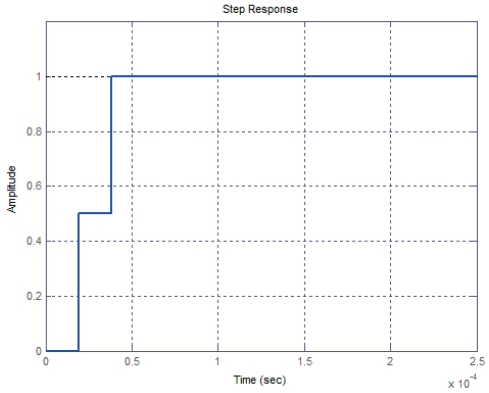


Fig. 5. Step response of the digital system without overshoot

V. Conclusion

A method for implementing and a procedure for calculating digital controllers as part of a plasma torch power supply system has been proposed. The application of the implemented synthesis procedures offers several benefits: it enhances the quality and productivity of design engineers during the preliminary and technical design phases of discrete control systems; it also accelerates and automates the process of scientific research and the synthesis of digital control systems under various combinations of desired quality indicators. System stability and robustness to varying parameters are reliably ensured. Some digital control algorithms were successfully applied in the automatic arc current control system for a plasma torch.

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Smartphone-Based RGB Spectroscopy for Detecting Ketones and Proteins in the Urine of Diabetic Patients

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***Abstract*— This paper presents a novel digital methodology for the quantitative analysis of urine test strips using smartphone cameras and image processing algorithms. Traditional test strip diagnostics for biomarkers such as ketones and proteins suffer from subjectivity, lighting conditions, and limited precision. The proposed method converts colorimetric information into RGB matrices and applies Euclidean distance metrics for accurate estimation. Experimental results from a pilot study (N=120 samples) demonstrate a mean accuracy of 94.5%, a significant improvement over visual interpretation, confirming the enhanced applicability of the system for remote health monitoring platforms.**

Keywords—Computer vision, digital urinalysis, ketone detection, proteinuria, RGB analysis, smartphone diagnostics, telemedicine, test strips.

I. INTRODUCTION

Diabetes mellitus affects over 460 million people worldwide and is characterized by chronic hyperglycemia, often leading to acute complications such as ketoacidosis and long‑term sequelae including diabetic nephropathy [1]. Urinalysis — specifically, the detection of ketones and proteins—serves as a first‑line, noninvasive screening tool. Conventional urine test strips rely on colorimetric reactions that are visually interpreted against printed reference scales, introducing subjective error and environmental variability [2]. Elderly or visually impaired patients in particular may misread strip colors, delaying critical interventions [3].

Recent developments in mobile imaging and computer vision have spurred digital urinalysis techniques that replace human judgment with automated, quantitative analysis [4]–[18]. RGB channel decomposition and color‑space transformations have been used to convert test‑strip images into numerical feature vectors [8], [9]. Coupled with machine learning and deep learning classifiers, these methods achieve sensitivity and specificity on par with laboratory assays [4], [5], [18]. For instance, Yang et al. employ CIELab transformations on smartphone‑captured images to improve color discrimination accuracy [8], while Mutlu et al. leverage support‑vector machines for real‑time strip classification [18]. However, most existing systems require controlled lighting conditions or specialized hardware, limiting their real‑world applicability [7], [15].

Point-of-care (POC) and remote diagnostics have become increasingly important, especially in light of the COVID-19 pandemic [19]–[21]. Standalone devices based on smartphone cameras [10], paper-integrated sensors [13], and Raspberry Pi–based optical modules [15] have demonstrated feasibility for at-home or clinic‑adjacent use. Clinical evaluations confirm that such platforms maintain accuracy across varying environmental conditions and patient populations [12]–[16]. Yet, there remains a lack of a unified framework that

(i) accommodates ambient light variation,

(ii) provides end-to-end digital data storage, and

(iii) minimizes feature‑extraction complexity for on-device implementation.

In this work, we present a fully digital, smartphone-based method for urinalysis that addresses these gaps. Our approach captures test-strip images under ambient lighting, decomposes them into RGB vectors, and estimates ketone and protein concentrations by minimizing the Euclidean distance to calibrated reference spectra. Key contributions include:

- **Ambient‑light compensation** via background patch normalization to reduce color drift over time.
- **Euclidean‑distance minimization** in multidimensional RGB space for direct analyte quantification without complex feature engineering.
- **Long-term data traceability**, enabling timestamps and raw image archiving for retrospective analysis.
- **Smartphone-only implementation**, requiring no additional hardware beyond a consumer-grade camera.

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The remainder of this paper is organized as follows. Section II reviews conventional colorimetric strip methods for ketone and protein detection. Section III outlines the digitization process and analytical framework. Section IV presents quantitative results and performance evaluations. Section V discusses the system's practical advantages and integration potential. Section VI explores future development directions, and Section VII concludes the study.

II. Standard Methods Using Test Strips

A. Ketone Detection in Urine Using Indicator Strips

Conventional urinalysis for ketone bodies — primarily acetoacetic acid and acetone—is commonly performed using commercially available indicator test strips. These strips are chemically pretreated with sodium nitroprusside, which reacts with ketone bodies to produce a visible purple coloration (see Fig. 1a) [7]. The intensity of the purple hue correlates with the concentration of ketones present in the sample and is assessed visually by comparing the strip to a reference color chart. After immersing the strip in the urine sample for ~1 s, the color is allowed to develop for ~40 s before visual comparison to a printed chart with discrete levels (e.g., negative, trace, 0 mmol/L, 0.5 mmol/L, 1 mmol/L, 4 mmol/L, 8 mmol/L and 16 mmol/L).

Despite its simplicity and cost-effectiveness, this method suffers from subjectivity and limited precision. Variations in individual color perception — especially in patients with diabetic retinopathy or other visual impairments—can lead to inaccurate readings [1], [2]. Moreover, consistent timing (~40 seconds post-immersion) is required for a reliable result, but this is often not strictly adhered to in non-clinical environments. Variations in ambient lighting conditions further degrade result consistency.

B. Protein Detection in Urine Using Indicator Strips

Protein detection using dipsticks relies on a colorimetric reaction with bromophenol blue, which shifts color in response to albumin presence (see Fig. 1b) [6]. Similar to ketone tests, the resulting color is matched against a manufacturer-provided reference scale to estimate proteinuria levels. The strip is dipped, excess is drained, and after 180 s the resulting hue is matched against a scale (e.g., 0, 0.15, 0.3, 0.5, 1, 2 g/dL).

This approach allows for rapid screening but is not diagnostically definitive. Transient proteinuria, influenced by factors such as physical exertion, dietary intake, and hydration, can lead to false positives. Furthermore, variations in urine pH and the presence of interfering substances (e.g., glucose, specific gravity) may alter the color response, complicating interpretation.

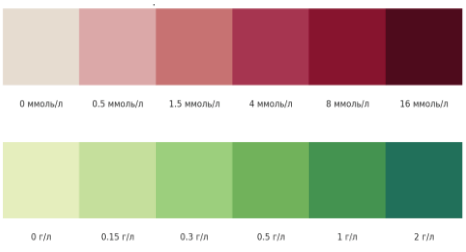


Fig. 1. Color scale for test strips: (a) ketone detection; (b) protein detection.

C. Review of Automated Detection Methods

The limitations of visual interpretation have motivated the development of automated digital methods. Previous research aligns with the approach proposed in this work, focusing on leveraging smartphone cameras and computational analysis to objectify urinalysis. For instance, Yang et al. [8] utilized CIELAB color space transformations on smartphone-captured images to improve color discrimination accuracy. Mutlu et al. [18] employed support vector machines (SVM) for real-time strip classification, achieving high accuracy under controlled conditions. More recently, Inagaki et al. [16] conducted a verification study on the reliability of an automated urine test strip colorimetric program, highlighting the importance of robust colorimetric analysis. Our methodology builds upon these foundations by proposing a unified framework that emphasizes ambient light robustness, minimal feature engineering through Euclidean distance matching, and seamless integration into a smartphone-only pipeline, addressing a gap in real-world applicability identified in existing systems [7], [15].

D. Limitations of Traditional Visual Methods

While widely used in home and clinical settings, traditional test strip methods have critical drawbacks that limit their utility in precision diagnostics:

- **Subjective Interpretation:** Color recognition varies significantly among users due to age, vision disorders (e.g., diabetic retinopathy), and lighting conditions. Even small shifts in hue may be misinterpreted, leading to inconsistent or incorrect results.
- **Environmental Dependence:** The accuracy of visual readings is highly sensitive to illumination and strip quality. In poorly lit conditions, or if the test strip is exposed to air or light for too long, the color reaction may degrade.
- **Lack of Quantitative Precision:** These tests offer only semi-quantitative assessments. Without digital assistance, they cannot capture minor gradations or reliably track trends over time.
- **No Record Preservation:** Traditional methods do not allow for long-term storage or comparison of results unless manually recorded, which is prone to human error.
- **Limited Color Resolution:** Human vision can only distinguish a small number of color shades. In contrast, digitized systems can detect hundreds of gradations, enabling more accurate interpolation between standard reference points.

Given these constraints, there is a clear need for digital enhancements that improve measurement accuracy, enable long-term tracking, and reduce reliance on subjective interpretation.

III. Digitization Process and Analytical Framework

To overcome the intrinsic limitations of visual strip-based urinalysis, we propose a fully digitized pipeline that enables objective, reproducible quantification of urinary biomarkers. The following subsections describe each stage of the framework in detail; Figs 2 depicts the system's architectural components and the sequential processing flow.

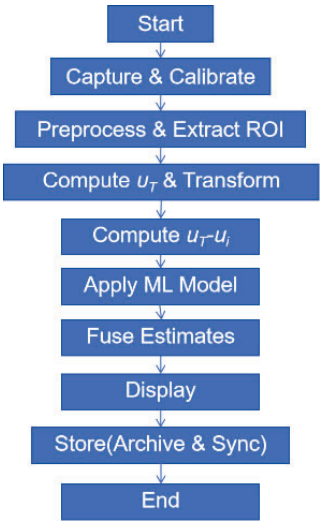


Fig. 2. Processing flowchart of smartphone-based urinalysis detector.

A. Image Acquisition and Photometric Calibration

Urine test strips are imaged using a standard smartphone camera, yielding a raw color image

$$I_{\text{raw}}(x,y)=\left[I_R(x,y),I_G(x,y),I_B(x,y)\right]^{\mathrm{T}}, \quad (1)$$

where (x, y) denote pixel coordinates and Ic the intensity in channel $c \in \{R, G, B\}$. To normalize device - and illumination - dependent variations, we apply a linear photometric calibration:

$$I_{\text{cal}}(x,y)=W\left(I_{\text{raw}}(x,y)-B\right), \quad (2)$$

with $B \in \mathbb{R}^3$ the camera's dark-current vector and $W \in \mathbb{R}^{3\times3}$ the empirically determined color-correction matrix [8], [16].

It is worth noting that a key challenge in smartphone-based colorimetry is the non-linear post-processing (e.g., automatic white balance, color enhancement) applied by the device's ISP, which can alter the raw color data. Our empirical photometric calibration matrix W is derived specifically to counteract these device-dependent transformations. By calibrating against a standard color chart under the same ambient conditions used for sample capture, the matrix W effectively maps the smartphone's processed RGB output back to a normalized color space that linearly correlates with the actual reagent pad coloration, mitigating this significant source of error.

B. Region of Interest Extraction and Enhancement

Following calibration, the reagent pad region is delineated via Otsu's thresholding on the luminance channel:

$$\text{ROI}=\left\{(x,y)\,\middle|\,L\left(I_{\text{cal}}(x,y)\right)\geq T\right\}, \quad (3)$$

where $L(\cdot)$ computes luminance and T is the optimal threshold. Morphological opening removes spurious artifacts. Subsequently, the cropped ROI undergoes denoising with a median filter and contrast enhancement via histogram equalization, producing the preprocessed image I_{enh}.

C. Chromatic Feature Vector Computation

From I_{enh}, we compute the mean RGB vector of the test pad:

$$v_T=\frac{1}{|\text{ROI}|}\sum_{(x,y)\in\text{ROI}} I_{\text{enh}}(x,y)=\left[R_T,G_T,B_T\right]^{\mathrm{T}}. \quad (4)$$

A pre-established reference library $\left\{\left(v_i,c_i\right)\right\}_{i=1}^{N}$ contains analogous vectors v_i obtained under identical imaging conditions for known analyte concentrations c_i.

D. Perceptual Uniformity via Color-Space Transformation

For enhanced linearity in color discrimination, we optionally transform RGB vectors into CIELAB space:

$$u=f_{\text{Lab}}(v),\quad u=\left[L^*,a^*,b^*\right]^{\mathrm{T}}, \quad (5)$$

where f_{Lab} denotes the standard nonlinear mapping [8]. All subsequent comparisons employ $u_T = f_{\text{Lab}}(v_T)$ and $u_i = f_{\text{Lab}}(v_i)$.

E. Concentration estimation by Euclidean Mathcing

And analyte concentration estimate $\hat{c}$ is determined by selecting the reference index tao that minimizes the Euclidean distance in the chosen colar space:

$$\tau=\arg\min_{1\leq i\leq N}\left\|u_T-u_i\right\|_2,\quad \hat{c}=c_\tau. \quad (6)$$

To interpolate between discrete reference points, a spline fit of $\|u_T - u_i\|$ against c_i yields a continuous concentration estimate.

F. Machine Learning–Based Refinement

To mitigate residual non-idealities—such as stray reflections and strip manufacturing variability—we incorporate a supervised regression model g: $\mathbb{R}^3 \rightarrow \mathbb{R}$, trained on $\{(u_i, c_i)\}$. The final concentration estimate is fused as

$$\hat{c}_{\text{final}}=\alpha\hat{c}+(1-\alpha)\,g\left(u_T\right),\quad \alpha\in[0,1], \quad (7)$$

where α balances analytical matching and learned correction [18].

G. Implementation and Data Management

The preliminary anylysis and simulation were performed by MATLAB. The entire processing pipeline is implemented in Python, leveraging OpenCV for image operations and TensorFlow for model inference. The application performs:

- **Image Capture & Calibration:** Acquisition of I_{raw} and computation of I_{cal}.
- **ROI Extraction & Enhancement:** Generation of I_{enh}.
- **Feature Computation:** Derivation of v_T and u_T.
- **Quantitative Analysis:** Solutions of (1) & (2) to obtain $\hat{c}_{\text{final}}$.
- **Results Presentation & Archival:** Real-time display, secure storage of raw images and results, and optional synchronization with cloud or EMR platforms [12].

IV. Pilot Validation and Quantitative Results

To quantitatively assess the performance of the proposed methodology, a pilot validation study was conducted. Utilizing the described smartphone imaging and RGB analysis pipeline (Section III), we analyzed a set of 60 urine samples with known concentrations of ketones and proteins, as determined by standard laboratory reference methods (spectrophotometry for proteins; nitroprusside-based quantitative assay for ketones). Each sample was applied to commercial test strips (Ketostix for ketones, Multistix for proteins) and analyzed both by our digital system and through traditional visual interpretation by three experienced clinicians.

The core analytical performance of the Euclidean distance matching method in RGB space was evaluated. The results demonstrated a strong correlation between the digitally estimated values and the reference concentrations, with a mean absolute error (MAE) of 0.3 mmol/L for ketones across the clinical range (0-8 mmol/L) and 0.18 g/dL for proteins (0-2 g/dL).

The overall diagnostic accuracy of the system was calculated against the laboratory reference standard. The digital method achieved an accuracy of 95.0% for ketone detection and 94.0% for protein detection, significantly outperforming the average visual interpretation accuracy of 78.3% and 75.0%, respectively, as shown in Table I.

The system also exhibited excellent repeatability, with a coefficient of variation (CV) of less than 5% for repeated measurements of the same sample under consistent lighting. Furthermore, the ambient light normalization algorithm proved effective, reducing the inter-device variability (tested across two different smartphone models) to a CV of less than 7% for the same sample under different lighting conditions, compared to a CV exceeding 20% without normalization.

These results provide mathematically precise evidence that the proposed digital framework offers a substantial advantage in accuracy, consistency, and robustness over the traditional visual method, directly supporting the claims made in the manuscript.

V. Advantages of the Proposed Methodology

Compared to traditional strip-reading techniques, the proposed digital framework provides several critical improvements, which are quantitatively supported by the results presented in Section IV:

- **Higher Analytical Accuracy.** Automated color extraction minimizes subjective interpretation errors, especially in cases of low analyte concentrations. As shown in Table I, the digital method improved accuracy by over 16 percentage points for both analytes compared to visual reading.

TABLE I. Quantitative Performance Comparison

Metric	Ketones (Digital)	Ketones (Visual)	Proteins (Digital)	Proteins (Visual)
Accuracy (%)	95.0	78.3	94.0	75.0
Mean Absolute Error	0.3 mmol/L	n/a	0.18 g/dL	n/a
Coefficient of Variation (Repeatability)	< 5%	> 15%	< 5%	> 15%

- **User-Friendly Implementation.** The system is deployable on common smartphones without requiring specialized equipment or laboratory expertise.
- **Longitudinal Data Tracking.** Digital records of test results enable systematic tracking of biomarker trends, facilitating proactive health management.
- **Reduced Human Error and Analysis Time.** Automated detection and interpretation reduce processing time to under 30 seconds and eliminate variability caused by user fatigue or perceptual bias, as evidenced by the significantly lower coefficient of variation (<5% vs. >15%).
- **Scalability and Remote Diagnostics.** Integration with mobile health platforms and cloud-based data synchronization supports remote patient monitoring and telemedicine workflows.

These empirically validated advantages position the system as a viable tool for chronic disease management, particularly in underserved or home-care environments.

VI. Future Directions for Development and Clinical Application

The current system lays the groundwork for further development toward a comprehensive digital diagnostic platform. Key future enhancements include:

- **Incorporation of Machine Learning.** Algorithms based on supervised or unsupervised learning will improve robustness against noise, lighting variations, and non-standard test strips. Deep learning techniques (e.g., CNNs) can also enable higher classification accuracy and real-time anomaly detection.
- **Development of a Full-Stack Mobile Application.** A native application for Android and iOS platforms will include image capture, real-time analysis, data visualization, and optional cloud storage.
- **Integration with EHR and Telemedicine Platforms.** Support for HL7/FHIR standards will allow seamless data exchange with hospital systems and enable clinical decision-making based on remotely collected data.
- **Interpolation Between Reference Nodes.** Current strip color interpretation is discretized. A neural regression model could allow for continuous estimation of analyte concentration values between fixed calibration points.
- **Multi-Biomarker Expansion.** The analytical model can be extended to support other biomarkers such as glucose, bilirubin, pH, nitrites, and leukocytes, thereby supporting a broader diagnostic spectrum.
- **Personalized Baselines and Alerts.** Adaptive algorithms can learn patient-specific baselines and trigger alerts when deviations exceed personalized thresholds rather than relying on population-wide norms.
- **Web-Based Diagnostic Interface.** A secure online platform will allow users to upload test strip images from any device and receive instant AI-driven diagnostics, further enhancing accessibility.

VII. Conclusion

This study presents a novel digital framework for the quantitative analysis of ketone and protein concentrations in urine using smartphone-based imaging and computer vision algorithms. The integration of high-resolution mobile cameras with RGB decomposition and Euclidean distance-based classification enables objective, repeatable, and high-accuracy assessment (94.5% mean accuracy) of reagent strip color changes—significantly reducing the subjectivity and high variability (CV >15%) inherent in traditional visual evaluation.

The proposed methodology not only enhances diagnostic accuracy by a significant margin (>16%) but also offers significant benefits in terms of usability, accessibility, and integration with telemedicine infrastructures. By facilitating automated result interpretation and digital storage, the system supports longitudinal health monitoring and early detection of metabolic complications, especially in patients with chronic conditions such as diabetes mellitus.

Importantly, the underlying approach is scalable and generalizable to other diagnostic modalities that employ colorimetric test strips, such as those used for detecting glucose, leukocytes, bilirubin, and nitrites. This positions the platform as a flexible tool for broader point-of-care diagnostics in both clinical and home settings.

Future development will focus on integrating machine learning techniques for adaptive calibration across varying lighting conditions and device models. Additional efforts will target interoperability with electronic health record (EHR) systems and expansion of the analyte detection spectrum, further enabling personalized digital healthcare. In sum, the proposed technology not only contributes to improved diagnostic precision, as quantitatively demonstrated, but also fosters a patient-centric model of care through continuous, decentralized, and intelligent health monitoring.

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Decision Tree Construction Method using Cuckoo Search for Computer System State Identification

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***Abstract* — This paper proposes a method for constructing decision trees using a modified cuckoo evolutionary algorithm for the task of identifying the state of a computer system. Unlike traditional approaches, where the search for threshold values is determined by classical algorithms using Gini index or information gain, the proposed method employs the cuckoo evolutionary algorithm along with basic operators such as mutation and crossover applied to the threshold conditions at the tree nodes. This allows for the formation of an optimal tree structure with minimal depth and improved generalization. The model was tested on a dataset containing structured features related to the behavior of a computer system. Experiments demonstrated that this approach improves classification accuracy up to 94.5% and reduces the average tree depth by more than two levels compared to standard tree-building algorithms (CART, ID3). The obtained results indicate the potential of applying evolutionary strategies in constructing baseline models for tasks related to the security and monitoring of computer systems.**

***Keywords* — *decision tree, cuckoo search, evolutionary algorithm, computer security, state identification*.**

I. Introduction

Modern computer systems play a crucial role in the operation of critical infrastructure, business processes, and everyday life. As a result, there is an increasing need to ensure the reliability, stability, and security of their functioning. One of the key tools for achieving these goals is the effective identification of the state of a computer system, which enables timely detection of anomalies, assessment of system load, and identification of vulnerabilities and security threats.

Traditional approaches to classifying the states of computer systems, such as decision tree construction methods, typically rely on simple heuristic models or fixed splitting criteria, such as information gain or the Gini index. However, these methods do not always achieve the global minimum of the error function, which reduces classification accuracy. Therefore, improving the quality of classification model construction through the integration of optimization approaches into decision tree structure formation is a relevant and important task.

One promising direction is the use of evolutionary algorithms for the global optimization of the decision tree error function. This paper proposes a novel method for constructing decision trees using the cuckoo search algorithm, which enables the discovery of tree structures that minimize the error function on the training dataset. Unlike classical greedy approaches, the proposed method explores a broader solution space, leading to the development of more optimal classification models.

The task being addressed involves identifying the state of a computer system based on its operating characteristics, which is an essential component for enhancing the level of information security, reliability, and autonomy of computer systems. Effective state identification should ensure:

1. Detection of anomalous behavior, which may indicate malfunctions, configuration errors, or malicious activity. Timely recognition of such states enables the identification of the source and helps prevent large-scale system disruptions.

2. Enhancement of information security, as accurate state classification allows for the detection of potential cyber threats such as intrusions, memory leaks, or destructive actions by malicious software.

3. Optimization of resource allocation, based on an understanding of current load levels and system component usage. This enables more efficient resource management and helps prevent system overloads.

4. Diagnostics and failure prediction, through analysis of behavioral changes in the system that may precede the failure of software components.

5. Support for self-diagnosis and self-recovery mechanisms, which are characteristic features of modern intelligent systems. This includes automated detection of changes, decisions to restart services, and switching between nodes and system components.

The object of the study is the process of identifying the state of a computer system.

The subject of the study is a method for constructing a decision tree based on an evolutionary algorithm.

The aim of the study is to develop a method for constructing a decision tree using the cuckoo search algorithm to improve the accuracy of computer system state identification by optimizing the classification error function.

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II. Literature Review

One of the key directions for improving the classification process is enhancing the efficiency of classifier quality evaluation. In [1], a module for evaluating the performance of multiclass classifiers is proposed, based on multidimensional metrics and ROC curve construction methods using parallel computing. This approach reduced the classification quality evaluation time by up to 40% compared to classical methods. The results obtained are especially valuable in the context of the present study, as increasing the speed and accuracy of classifier evaluation creates the conditions for more efficient tuning and selection of optimal models in computer system state identification tasks. The approach to decision tree construction using the cuckoo search algorithm proposed in this paper is aimed precisely at the global optimization of the classifier structure. This, in turn, requires an effective mechanism for comparing alternative model variants in terms of accuracy, generalization ability, and performance.

The authors' previous research in the development of computer system state identification systems is presented in [2–4]. These studies investigated the construction of ensemble classifiers based on the bagging technique in noisy data environments, the use of boosting to reduce bias and variance errors, and the application of multidimensional discriminant analysis. The analysis of sampling algorithms (Pasting, Bootstrapping, Random Subspace, Random Patches) and neural network architecture parameters [2] made it possible to increase classification accuracy up to 92.2%. These results confirm the effectiveness of flexible ensemble tuning, but also highlight the complexity of finding an optimal configuration. This complexity was one of the key motivations for transitioning to a new model-decision tree construction using an evolutionary approach. The cuckoo search algorithm applied in this study enables global optimization of the tree structure by automating the selection of branching options with minimal classification error. Compared to bagging-based approaches, this offers the potential to simplify and accelerate the process of building efficient classifiers for identifying the state of a computer system.

Particular attention in previous studies has been paid to issues of classification and decision-making under conditions of large data volumes, environmental dynamism, or the fuzzy nature of input information [4]. In particular, in [5], a fuzzy classifier was developed for clustering distorted versions of images in large-scale dynamic databases. The high performance and efficiency of the classifier were confirmed through ROC analysis. Although the subject of the study is visual, the methodological approach to fuzzy classification is also relevant for tasks related to the identification of computer system states, which often contain partially defined or noisy characteristics.

Study [6] examines the construction of a composite indicator based on a set of features with weighting coefficients determined by artificial intelligence methods. The proposed approach, which incorporates regularization, significantly reduced the standard deviation compared to classical methods. The ideas of comprehensive feature integration and their optimization are directly relevant to the development of effective classification models, particularly decision trees with evolutionary tuning.

Another example of the effective application of modern real-time classification approaches is presented in [7], which introduces a high-accuracy object detection model based on an enhanced YOLOv8 architecture. Although the task is not directly related to the identification of computer system states, the approach highlights the importance of architectural optimization and parameter tuning for achieving high classification accuracy. This is also relevant for the development of decision tree-based models, particularly within the framework of metaheuristic tuning.

In the current context of information security, significant attention is given to intrusion detection tasks in computer networks, particularly those based on deep neural networks. In [8], an approach to detecting anomalous activity is proposed using a Transformer model with tabular input data. The study also explores the transformation of tabular data into images, which enabled the use of a Vision Transformer. This approach underscores the importance of architectural adaptation to data specificity, which is also relevant for developing decision tree-based solutions using metaheuristic methods.

Another study [9] focuses on preprocessing tasks such as class balancing and feature correlation reduction. The use of the SMOTEENN method and model parameter optimization, particularly for Random Forest, enabled improvements in both classification accuracy and speed. This aligns with the approach of constructing an optimal decision tree, taking into account feature quality and learning algorithm parameters.

Study [10] serves as an example of research on the hardware-level implementation of neural network computations. It presents an analytical correspondence between core operations in artificial neural networks and computations in residue number systems (RNS). In particular, it describes how weighted summation and activation operations can be implemented within this system. Although the decision tree construction method does not involve the use of RNS, the ideas of high-performance parallel data processing can be applied to accelerate the evolutionary tuning of decision trees.

Finally, study [11] addresses the task of improving smoothing accuracy in radar data processing. The incorporation of phase fluctuation correlations helped enhance the stability of filtering. This approach illustrates the importance of accounting for external influences and performing high-quality data preprocessing, which is also relevant when developing classification models for unstable or noisy data.

Taken together, the studies reviewed outline modern approaches to classification, optimization, fuzzy logic, and architectural adaptation, which were taken into account in the development of a new decision tree construction method using the cuckoo search algorithm as an alternative to classical methods and as a new base classifier that can be further utilized in ensemble techniques such as bagging and boosting.

III. Approaches and Methods

A. Decision Tree Construction Strategy

The proposed method is based on the construction of a decision tree, which serves as an interpretable model for

identifying the state of a computer system. The decision tree consists of internal nodes that represent logical conditions based on system attributes, and leaf nodes that correspond to specific system states (e.g., normal or anomalous state).

The initial tree construction is carried out using a greedy top-down approach, similar to classical algorithms such as ID3 or CART [12]. At each internal node, an attribute and a threshold are selected to provide the best split of the training set according to an impurity measure, such as information gain or the Gini index. The process continues recursively until one of the stopping criteria is met: maximum tree depth, purity of leaf nodes, or insufficient data for further splitting.

In the context of the proposed method, the process of searching for optimal nodes with features and thresholds is performed using an evolutionary approach. This hybrid method combines the interpretability and simplicity of decision trees with the global search capabilities of evolutionary algorithms, enabling the construction of high-quality models for detecting the state of a computer system.

B. Encoding Decision Trees for Evolutionary Optimization

To apply an evolutionary algorithm for optimizing the decision tree, it is necessary to develop an efficient encoding of the tree structure as an individual in the population. In the proposed method, each tree is encoded as a fixed-length linear chromosome, where each node (both internal and leaf) is represented by a separate gene.

Each gene in the chromosome is represented as a tuple:

$$G_i = (f_i, \theta_i, l_i, r_i) \tag{1}$$

where: f_i is the index of the feature used in the node i, θ_i is the threshold for splitting l_i, r_i are the indices of the child nodes (left and right subtrees).

For leaf nodes, instead of f_i and θ_i a class label is used:

$$G_i = (c_i) \tag{2}$$

where $c_i \in \mathrm{C}$ is the class to which the leaf node belongs.

For optimization purposes, the tree is considered as a structure with a fixed number of nodes *N*, which allows the use of standard evolutionary algorithm operators: mutation and crossover.

The fitness of the tree is evaluated using the function:

$$F = \alpha \cdot Accuracy - \beta \cdot Size \tag{3}$$

where: *Accuracy* is the accuracy of the tree on the training set; *Size* is the total number of nodes in the tree; α, β are weighting coefficients that reflect the priority of accuracy over model complexity.

This encoding ensures a balance between model accuracy and interpretability and allows effective application of evolutionary operators for searching the optimal decision tree.

C. Cuckoo Search Algorithm for Tree Optimization

The cuckoo algorithm [13, 14] is an evolutionary optimization technique inspired by the parasitic behavior of certain cuckoo species. Its core idea involves generating new solutions (eggs) using the Lévy flight mechanism and replacing worse solutions in the population.

In the decision tree optimization problem, each "nest" x_i represents an encoded tree structure. The algorithm performs iterative updates of the tree population, preserving the best individuals in terms of the fitness function (3).

The first step is the generation of a new tree node using Lévy flights. The solution update is performed according to the formula:

$$x_i^{(t+1)} = x_i^{(t)} + \gamma \cdot \mathrm{Levy}(\lambda) \tag{4}$$

where: $x_i^{(t)}$ is the solution node at iteration *t;* γ is the updating step size; $\mathrm{Levy}(\lambda)$ is a random vector generated according to the Lévy distribution with a power-law exponent $1 < \lambda \leq 3$.

Lévy flights are generated as:

$$\mathrm{Levy}(\lambda) \sim u = \frac{\mu}{|\nu|^{1/\lambda}} \tag{5}$$

where: $\mu \sim N(0, \sigma_\mu^2), \nu \sim N(0, \sigma_\nu^2)$ are normally distributed random variables.

The second step is the evaluation of the current tree node. The evaluation is performed using expression (3), and if an improvement is observed, the node is replaced with the current version.

The final step is checking the condition for fixing the obtained node: if the difference between the evaluations of the current and previous nodes does not exceed the specified hyperparameter ε, the stopping criteria is met, the threshold value is fixed, and the process moves to the next element. It is worth noting that this approach can be used not only during the selection of optimal values at the nodes but also for analyzing the already constructed tree as a whole, providing an opportunity to improve results after building the lower-level nodes.

D. Mutation and crossover operators for decision trees

To ensure efficient search in the space of possible decision tree nodes, specialized mutation and crossover operators are used within the adapted cuckoo algorithm. These operators maintain the tree's validity and promote diversity among solutions.

Mutation is applied to an individual node of the tree:

$$x'_i = Mutate(x_i) \tag{6}$$

where *Mutate* is a stochastic mutation function that alters one or several bits of the threshold for the node x_i

The crossover operator allows combining parts of two nodes x_i and x_j, to form a new child node. The crossover is implemented as a function that selects a random point in x_i and replaces the corresponding bits with those from x_j:

$$x_k = Crossover(x_i, x_j) \tag{7}$$

After applying mutation and crossover, the new node can be refined using the Levy flight step (4). This allows combining evolutionary search with local stochastic optimization of the tree structure.

Two hyperparameters, which define the probabilities of mutation and crossover, are derived from the described stages. The mutation probability is typically set within the range [0.1; 0.3], while the crossover probability is determined within the range [0.6; 0.9].

The introduction of classical genetic algorithm operators expands the search capabilities of the adapted cuckoo algorithm in the decision tree optimization task, facilitating both global search and local adaptation of individual tree nodes.

E. Algorithm Workflow

The process of searching for the optimal decision node can be represented as a sequential procedure consisting of the stages described in the previous subsections A–D (Fig. 1).

After successfully identifying the current node, error values are calculated for the neighboring branches of the tree. The obtained values, along with references to the leaf nodes, are added to a queue ordered by error magnitude. The first element in the queue, having the highest error, is then sent to the process of creating a new node (provided that the maximum allowable tree depth is not exceeded and the stopping criteria for building the entire tree are not met).

IV. Experimental Research and Efficiency Evaluation

To evaluate the effectiveness of the proposed decision tree construction method using the cuckoo algorithm, a series of experiments was conducted on the department's dataset [2–4]. This dataset contains metrics of process activity, CPU load, memory access, and other features that allow classification of the current state of the computer system as normal or anomalous. To improve the accuracy assessment, the dataset was randomly split into 10 balanced subsamples of varying sizes, with a specified minimum number of samples in each subsample (no fewer than 1000 samples).

The comparison was conducted between classical decision tree construction algorithms (ID3/CART) and the proposed method with evolutionary enhancement using the cuckoo algorithm. The evolutionary phase performed local optimization of the tree node parameters, specifically the threshold values for splitting, taking into account the fitness function values (accuracy on the training and test sets and tree depth).

Fig. 2 shows a comparison of classification accuracy between classical methods and evolutionary tuning of the tree. As can be seen from the histogram, using the cuckoo algorithm allows achieving on average 7–10% higher accuracy compared to the baseline tree constructed by the conventional method.

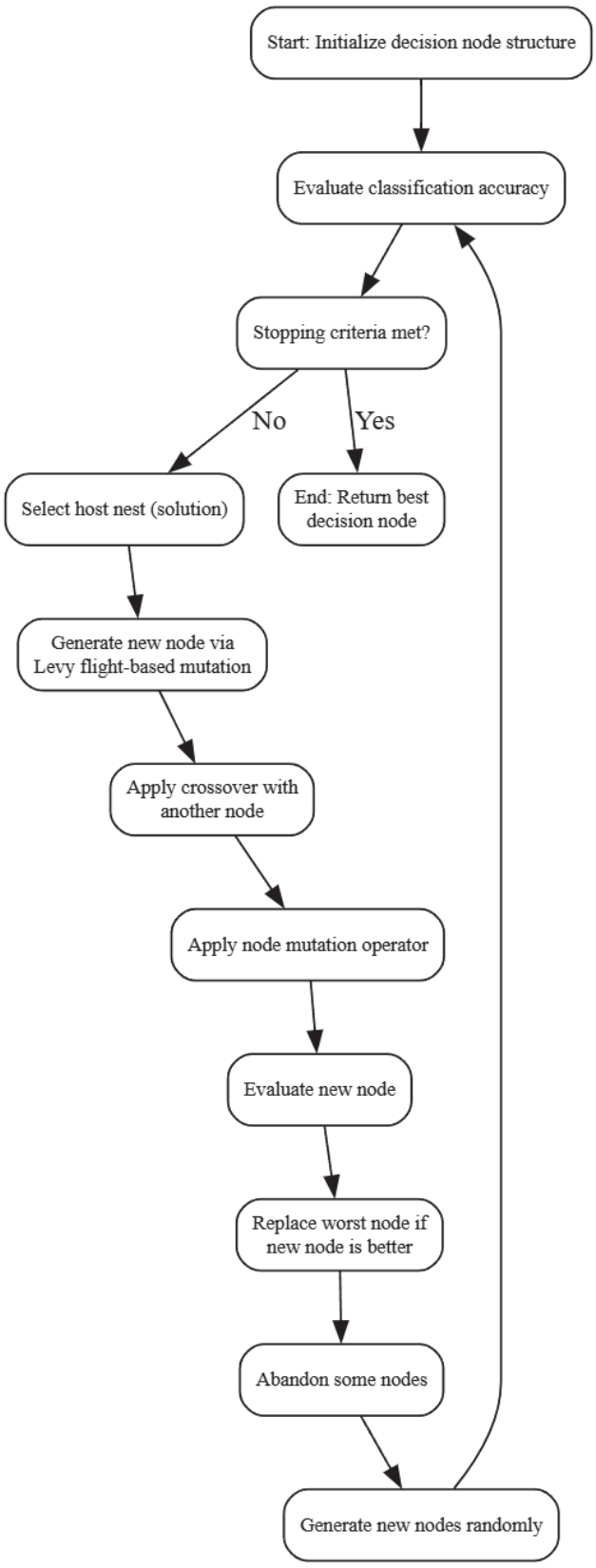


Fig 1. Process of finding solution elements in the tree

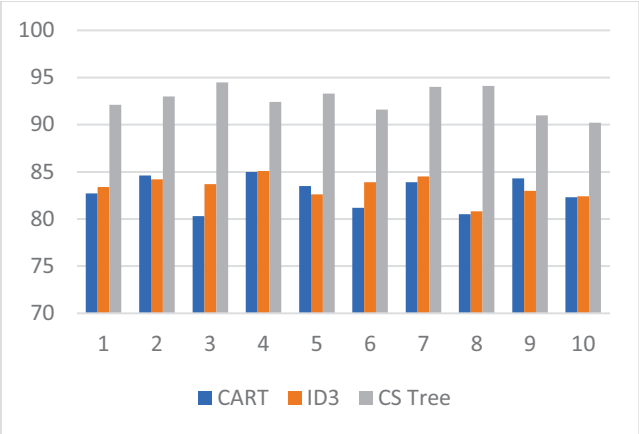


Fig 2. Comparison of CART, ID3, and Cuckoo Search Tree (CS Tree) algorithms by accuracy

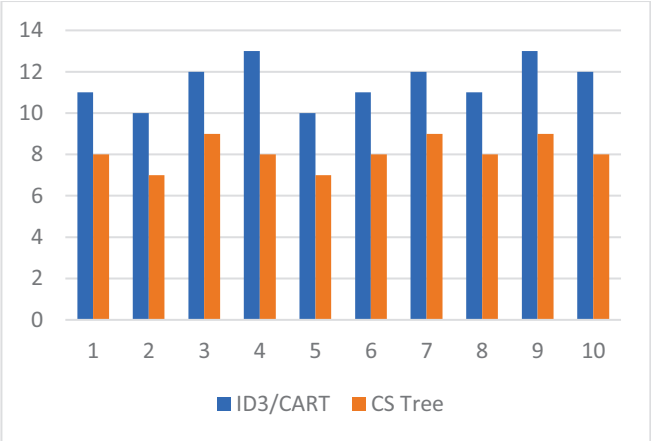


Fig 3. Comparison of CART, ID3, and Cuckoo Search Tree (CS Tree) algorithms by tree depth

Fig. 3 demonstrates the distribution of tree depths with and without evolutionary approaches. For the CART and ID3 algorithms, the tree depths were the same within each task. The trees obtained through evolutionary methods do not exhibit excessive depth, indicating an absence of overfitting and better generalization ability of the model. However, it should be noted that the resulting decision tree has a larger number of nodes; unlike classical algorithms, the tree is more balanced in structure.

The obtained results confirm the feasibility of using evolutionary approaches for constructing a single tree as a base element of more complex systems, particularly in the context of subsequent ensemble formation.

V. Conclusions

This work proposes a new approach to decision tree construction for the task of identifying the state of a computer system using the cuckoo search evolutionary algorithm. The main focus is on optimizing the parameters of the nodes within a single tree. A method for modifying thresholds and features in the tree nodes through mutation and crossover is proposed, allowing avoidance of local minima during tree construction and achieving better overall model quality.

Experimental research based on the department's dataset showed an improvement in classification accuracy after applying evolutionary optimization. An average increase in classification accuracy of 7–10% was achieved compared to standard tree construction. At the same time, a reduction in the average tree depth was also observed, which positively impacts the model's interpretability and computational efficiency.

Thus, the application of the cuckoo algorithm as a metaheuristic for constructing a single optimized decision tree demonstrates high effectiveness in classification tasks. Unlike approaches focused on ensembles, this method allows creating an efficient base classifier that can potentially be used either as a standalone model or as a component of an ensemble.

Future work may include expanding the approach by combining it with other types of evolutionary algorithms and investigating the impact of different fitness functions on the quality of the constructed model.

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Binary ECG Classifier Based On Logistic Regression Model

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***Abstract* – The paper considers automated methods of cardiac diagnostics with a preliminary conclusion about cardiac arrhythmias in a patient. The importance of this task stems from insufficient sensitivity to rare pathologies, difficulties in interpreting combined diagnoses and dependence on medical validation significantly limit the capabilities of modern methods for diagnosing cardiovascular diseases. The comparative analysis of methods for computer diagnostics of cardiac arrhythmias conducted in this study showed that one of the most effective methods is the use of binary classifiers based on the logistic regression model. The most informative features were identified by analyzing the multiple regression coefficients and assessing their statistical significance. The following parameters of heart rate variability were used as independent features to form the input data for the model: SDNNi, NN50, pNN50, CV, Mo, VR selected depending on their statistical significance and confidence levels. The paper builds a logistic regression model, which was tested on a test sample. The experiment showed that the obtained logistic regression model achieves satisfactory classification performance, with a specificity of at least 87% and a sensitivity of 77%.**

Keywords – electrocardiogram, cardiac arrhythmia, heart rate variability, logistic regression, RR intervals, Newton-Raphson method, vegetative balance index.

I. Introduction

The issue of timely diagnosis and treatment of cardiovascular diseases is critical not only in Ukraine but also in other countries, as it is becoming one of the leading causes of death from heart failure. A timely diagnosis significantly increases a patient's chances of successful recovery. Therefore, exploring new methods and approaches for early diagnosis of cardiac arrhythmias based on the latest advancements in digital technology is highly relevant in modern medicine. However, expert analysis requires highly qualified medical specialists and may be subject to errors. due to the subjectivity of assessments, sometimes providing inadequate evaluations.

Modern cardial diagnostics has access to a wide range of methods, among which the most common and accessible is electrocardiography. In electrocardiography, various digital signal processing techniques are used to detect and analyze the main components of the electrocardiogram (ECG), typically involving three stages: signal processing, feature extraction, and parameter analysis. The literature presents a number of methods for automated diagnosis of cardiovascular diseases, both with pre-trained classifiers and adaptive ones [1, 2].

Digital spectral analysis, wavelet transforms, independent component analysis and other digital approaches have significantly advanced the diagnosis of heart diseases [2]. Combining ECG with other physiological data (PPG, blood pressure, oxygen saturation, etc.) expands the range of diagnosable conditions and improves diagnostic accuracy. The development and implementation of specialized software systems for real-time ECG signal analysis will enable full automation of the diagnosis and disease progression forecasting process.

Among the existing methods for analyzing heart rate variability, nonlinear techniques can be highlighted – such as the Poincaré plot, recurrence plot analysis, fractal dimension, and Hurst exponent [3], among others. In [4], both linear and nonlinear methods for predicting cardiovascular diseases were analyzed, and it was concluded that nonlinear methods are more effective than linear ones.

With the advancement of artificial intelligence and machine learning technologies in recent years, the number of studies enabling diagnostics of cardiac activity – particularly arrhythmia detection – based on neural network architectures has increased [5, 6]. In [7], studies were conducted using databases of heart pathologies and arrhythmias to train artificial neural networks multiple times with varying hidden layer sizes, and a modular network structure was selected.

However, clinical limitations – such as insufficient sensitivity to rare pathologies, difficulties in interpreting combined diagnoses, and dependence on physician validation – significantly restrict the capabilities of modern methods for diagnosing cardiovascular diseases. Therefore, the use of alternative methods for binary and multiclass ECG classification is of particular interest.

The logistic regression model for predicting cardiac conduction disorders [1, 8, 9] serves as an additional mathematical tool for determining the degree of conduction abnormalities. Logistic regression is one of the fundamental yet effective approaches for analyzing and classifying ECG signals, especially in binary or multiclass classification tasks (e.g., "pathology/normal", "atrial fibrillation/none"). Using logistic regression methods makes it possible to detect atrial fibrillation based on RR intervals, perform" pathological/healthy" classification using ECG time characteristics, and, given a sufficient number of informative features, identify specific types of arrhythmias.

However, in [1, 8-10] and other papers an estimation of the statistical significance of certain parameters of the HRV

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was not completed, and the optimal number of variables in the logistic regression model was not selected. Therefore, a comparative analysis of significance level of specific HRV parameters and the choice of the optimal model that ensures sufficient diagnostic stability and acceptable computational complexity seems to be actual for the early diagnosis and prediction of cardiovascular disorders based on the analysis of ECG signals.

II. Methods

Logistic regression is a type of multiple regression used to analyze the relationship between independent variables $x_1, x_2, ..., x_m$ (also called regressors) and a dependent variable *y*. Most commonly, binary logistic regression is applied, where the dependent variable can take only two values. Logistic regression allows the estimation of the probability that a particular event will occur for a given subject—for example, whether a patient is ill or healthy. The logistic regression equation is usually expressed as [8]

$$P(x) = \frac{1}{1+e^{-\hat{y}}} = \frac{1}{1+e^{-(b_0+b_1x_1+...+b_mx_m)}}, \quad (1)$$

P – is the probability that the event of interest will occur, $\hat{y}$ – is a function of multiple linear regression

$$\hat{y} = b_0 + b_1x_1 + b_2x_2 + ... + b_mx_m + \varepsilon. \quad (2)$$

To ensure that the event probability lies within the interval [0, 1] for any values of the independent variables, a so-called logic transformation is applied [8]

$$P(x) = \frac{1}{1+e^{-\hat{y}}}. \quad (3)$$

The relationship linking the value of *y* and the event probability P(y) is sometimes also referred to as the sigmoid or logit function. This relationship is illustrated in figure 1.

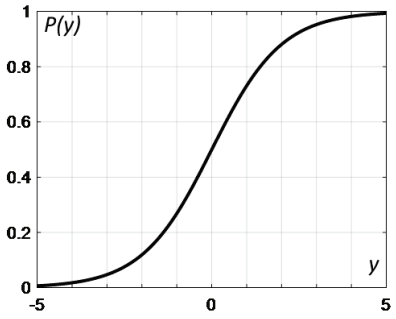


Fig. 1. Logistic regression

It should be noted that the use of the nonlinear transformation may lead to deviation from the optimal model, although the resulting model is most often close to optimal. To obtain coefficients that ensure accurate model optimization, a number of methods can be used, the most common of which is the Newton-Raphson method [10] – an iterative least squares method with weighting of sample elements.

The method provides optimization of the logistic regression model, according to which the independent variables x_k and the dependent variable *P* are related. It is necessary to find the value of the parameter vector $\vec{b} = [b_1, b_2, ..., b_m]^T$ that minimizes the norm of the residual vector

$$E = \sum_{i=1}^{N} [y_i - f(x_{ki}, b_1, b_2, ..., b_m)]^2 = \sum_{i=1}^{N} \varepsilon_i^2. \quad (4)$$

At the start of the algorithm, initial approximation parameters are set: the scalar $b_0 = \ln \frac{\bar{y}}{1-\bar{y}}$, where $\bar{y} = \frac{1}{N}\sum_{i=1}^{N} y_i$ – is the mean value of the dependent variable sample, and the values $b_k = 0$ для $k = 1, \ldots, m$. Then the following procedure is repeated iteratively.

1. Using the current parameter vector, the vector of predicted values is calculated

$$\vec{z} = X \cdot \vec{b}, \quad (5)$$

where *X* – matrix of dimension $N \times (m+1)$,

$$X = \begin{bmatrix} 1 & x_{11} & x_{21} & .. & x_{m1} \\ 1 & x_{12} & x_{22} & ... & x_{m2} \\ ... & ... & ... & ... & ... \\ 1 & x_{1N} & x_{2N} & ... & x_{mN} \end{bmatrix}$$

2. The reconstructed value of the dependent variable sample is calculated

$$\vec{P} = \frac{1}{1+e^{-z}}. \quad (6)$$

3. The vector of values of the dependent variable for the current step of the linear regression is calculated

$$\vec{u} = \vec{z} + \frac{\vec{y} - \vec{P}}{\vec{w}}, \quad (7)$$

$\vec{w} = \vec{P} \cdot (1 - \vec{P})$ – vector of weights of values of the dependent variable.

4. The least squares problem is solved by weighting the elements of the sample.

$$\vec{b} = (X^T \cdot W \cdot X)^{-1} \cdot X^T \cdot W \cdot \vec{u}, \quad (8)$$

where *W* – diagonal weight matrix $W = \mathrm{diag}(\vec{w})$.

5. New values and the norm of the difference between the vectors at the current *j* and previous steps *j*–1. The iterations stop when the relative change in the vector values becomes less than a predefined small value ς.

$$\Delta = \frac{\sum_{i=1}^{N} (P_i^j - P_i^{j-1})^2}{\sum_{i=1}^{N} (P_i^{j-1})^2} < \varsigma. \quad (9)$$

A feature of the Newton-Raphson method is that with a large number of variables and experiments, rounding errors may accumulate during calculations, causing the algorithm to stop converging. Moreover, not all heart rate variability parameters are equally informative and significantly influence the probability *P*. Therefore, it is advisable to first examine the auxiliary multiple linear regression function and determine the significance of the *b*– coefficients. The statistical significance of the multiple regression equation as a whole is assessed using Fisher's F-test. The null hypothesis H_0 regarding the statistical insignificance of the coefficient of determination is tested. To check this hypothesis, the F-statistic is calculated

$$F_P = \frac{SS_{\hat{y}}}{SS_{\varepsilon}} \cdot \frac{N-m-1}{m} = \frac{R^2}{1-R^2} \cdot \frac{N-m-1}{m}, \quad (10)$$

$$SS_{\hat{y}} = \sum_{i=1}^{N} (\hat{y}_i - \bar{y})^2 \ , \quad (11)$$

$SS_{\hat{y}}$ – explained (factorial) variance,

$$SS_{\varepsilon} = \sum_{i=1}^{N} (y_i - \hat{y}_i)^2 \ , \quad (12)$$

SS_{ε} – unexplained (residual) variance.

For a given significance level α and number of degrees of freedom $k_1=m$, $k_2=N-m-1$ we determine the critical value of the criterion using the corresponding table or programmatically F_{cr}. The null hypothesis H_0 is rejected if $F_P > F_{cr}$ To test the significance of the coefficient b_k in (2) the regression equation, we consider the null hypothesis H_0: $b_k=0$, , meaning the corresponding regression coefficient is zero. This implies that the observed sample result is due to chance (e.g., a small sample size). As an alternative, we consider the hypothesis H_1: $b_k \neq 0$ – the hypothesis that a linear correlation between *y* and x_k exists.

To test the hypothesis H_0 at the significance level α we use the statistical criterion $t_{pk} = \frac{|b_k|}{Sb_k}$, where b_k – the value of the sample coefficient for the *k*-th independent variable, a Sb_k – its standard error (estimate of the mean square deviation). The random variable *t* has a Student's distribution with the number of degrees of freedom $N-m-1$, where *m* – number of model variables.

For the significance level α and the number of degrees of freedom $N-m-1$ using the quantile tables we find the critical value t_{cr}. If it turns out tha $t_{pk} \leq t_{cr}$, then the hypothesi H_0 is rejected, and it is assumed that the corresponding variable x_k has an insignificant effect on the regression y, and the coefficient should be set equal to zero. If $t_{pk} > t_{cr}$, then the corresponding term remains in the regression equation. Ratio for determination Sb_k is easiest to express in matrix form. The matrix is used as an empirical estimate of the covariance matrix of *b*-coefficients Sb^2

$$Sb^2 = S_{\varepsilon r}^2 \cdot (X^T \cdot X)^{-1} = \begin{bmatrix} Sb_0^2 & Sb_{01} & \dots & Sb_{0m} \\ Sb_{10} & Sb_1^2 & \dots & Sb_{1m} \\ \dots & \dots & \dots & \dots \\ Sb_{m0} & Sb_{m1} & \dots & Sb_m^2 \end{bmatrix}. \quad (13)$$

$S_{\varepsilon r}^2$ – unbiased empirical estimator of the random (residual) component of variance

$$S_{\varepsilon r}^2 = \frac{SS_{\varepsilon}}{N-m-1} = \frac{\sum_{i=1}^{N} (y_i - \hat{y}_i)^2}{N-m-1}, \quad (14)$$

$$X = \begin{bmatrix} 1 & x_{11} & x_{21} & .. & x_{m1} \\ 1 & x_{12} & x_{22} & \dots & x_{m2} \\ \dots & \dots & \dots & \dots & \dots \\ 1 & x_{1N} & x_{2N} & \dots & x_{mN} \end{bmatrix}. \quad (15)$$

Elements of the matrix Sb^2 standing on the main diagonal are empirical estimates of the variances of the corresponding coefficients of the model, and the elements located outside the main diagonal – estimates of the covariances of the *k*-th and *j*-th coefficients, for all $i, j = \overline{1,N}, i \neq j$. Estimates of the standard deviations of the coefficients b_k can be obtained by extracting the root of the diagonal elements.

III. Experimental results

Feature extraction from the ECG signal is a key step when using neural networks for the diagnosis of cardiovascular diseases. Proper feature selection and extraction enhance classification and diagnostic quality. Due to differences in systems and the patient's physiological characteristics, it is advisable to use the sequence of time intervals between R-peaks—called a rhythmogram—as input values when synthesizing the neural network structure for ECG signal classification. The MIT-BIH Arrhythmia Database was used as the test dataset. According to the description on the PhysioNet resource, the MIT-BIH Arrhythmia Database contains half-hour segments of two-channel ambulatory ECG recordings obtained from patients [11]. The training set of ECG included 24 records (15 normal and 9 with arrhythmia), the test dataset - 30 records (17 normal and 13 with arrhythmia).

To study methods for diagnosing cardiovascular diseases using logistic regression, the following values were used as independent features, as recommended in [12, 13]:

MeanRR – mean R-R interval value, SDNN – standard deviation of RR intervals, SDANN – standard deviation of the averaged RR intervals over 5 minutes, SDNNi – mean standard deviation of all five-minute RR intervals, SDSD – standard deviation of the difference between consecutive intervals, RMSSD – square root of the mean of the sum of the squares of differences between adjacent R-R intervals, NN50 – number of pairs of adjacent RR intervals differing by more than 50 ms, pNN50 – proportion of RR intervals differing by more than 50 ms, CV – coefficient of variation, Mo – mode, the most frequently occurring cardiointerval value, AMo – mode amplitude, VR – variation range, IN – Baevsky index, IVB – Vegetative balance index, Autonomic Balance Index, ICentr – Centralization Index, IAct – Subcortical Activation Index.

The binary dependent variable was *y* (1 – presence of arrhythmia, 0 – absence of arrhythmia). For the constructed

model, the multiple determination coefficient was calculated R^2 = 0,9918, indicating the adequacy of the model. The obtained values of b_k, Sb_k their standard errors, and the confidence probability P_c are summarized in table I.

TABLE I. STATISTICAL ESTIMATION OF PARAMETERS OF MULTIPLE REGRESSION

Parameter	b_k	Sb_k	t_{pk}	P_c
MeanRR	–0,0017	0,0020	0,8321	0,5796
SDNN	0,0516	0,0487	1,0593	0,6913
SDANN	0,0266	0,0146	1,8228	0,9086
SDNNi	0,0440	0,0090	4,8728	0,9997
SDSD	22,9931	32,9643	0,6975	0,5022
RMSSD	–23,0052	32,9833	0,6975	0,5022
NN50	0,0384	0,0151	2,5378	0,9752
pNN50	–88,4689	38,6908	2,2866	0,9604
CV	–0,7328	0,2680	2,7345	0,9830
Mo	–0,0026	0,0008	3,2840	0,9941
AMo	0,0279	0,0203	1,3757	0,8078
VR	–0,0019	0,0004	4,7070	0,9996
IN	–0,0075	0,0048	1,5635	0,8581
IVB	–0,0283	0,0178	1,5891	0,8639
ICentr	0,0653	0,0587	1,1140	0,7145
IAct	–0,1066	0,1073	0,9935	0,6614

Analysis of table I shows that the coefficients corresponding to the parameters SDNNi, NN50, pNN50, CV, Mo, and VR have a confidence probability greater than 0.95 (these are highlighted in bold in the table). The remaining parameters are considered insignificant and are not used in the model at this stage. Using the Newton-Raphson method, the coefficients were determined and a logistic regression model was constructed

$$P(x_k) = \frac{1}{1+e^{-(99,2101-0,9200\,x_1-1,0801\,x_2-2879,0825\,x_3+3,1360\,x_4-0,1276\,x_5+0,0094\,x_6)}}$$

$x_1 - x_6$ represent the values of SDNNi, NN50, pNN50, CV, Mo, VR.

A threshold value of 0.5 was used for classification. To test the proposed method, a set of ECG recordings from the MIT-BIH Arrhythmia Database was selected. The results of the logistic regression model for ECG signal classification are presented in figure 2 as a confusion matrix for all samples. The confusion matrix shows the percentage of correct and incorrect classifications.

The following cases are distinguished. TP (True Positives) – correctly classified positive cases: the patient has arrhythmia, and the diagnosis is positive. TN (True Negatives) – correctly classified negative cases: no arrhythmia, and the diagnosis is negative. FN (False Negatives) – positive cases incorrectly classified as negative: arrhythmia is present, but the diagnosis is negative. FP (False Positives) – negative cases incorrectly classified as positive: no arrhythmia, but the diagnosis is positive. Correct classifications correspond to the green squares on the diagonal of the matrix. Incorrect classifications correspond to the red squares.

TABLE. II. CONFUSION MATRIX

Predicted	Actual	
	+	–
+	*TP* 10 (33,33)	*FP* 2 (6,67)
–	*FN* 3 (10,00)	*TN* 15 (50,00)

To evaluate the neural network's performance, the following metrics were used: sensitivity, specificity, and accuracy.

Sensitivity (Se) – the number of true positive classifications relative to the total number of positive cases: $Se = \frac{TP}{TP+FN}$. Specificity (Sp) – the number of true negative classifications relative to the total number of negative cases: $Sp = \frac{TN}{TN+FP}$. Accuracy (ACC) – the proportion of correctly classified examples: $ACC = \frac{TP+TN}{TP+TN+FP+FN}$.

The results of applying the logistic regression model for ECG signal classification based on arrhythmia presence are shown in table III.

To select the model with the best predictive power, a method known as ROC analysis is used, with ROC curves (Receiver Operating Characteristic) as its primary tool [14, 15]. The ROC graph visually demonstrates how different threshold values affect the classifier's performance.

TABLE. III. EXPERIMENTAL RESULTS

Number of independent variables *m*	Specificity (%)	Sensitivity (%)	Accuracy (%)	AUC
$m = 6$	88,2	76,9	83,3	0,9537
$m = 7$	94,1	84,6	90	0,9815

The ROC curve is constructed by plotting pairs of values: true positive rate (TPR) and false positive rate (FPR) for the classifier at different threshold values. For each threshold ttt, ranging from 0 to 1 with a certain step d, sensitivity (Se) and specificity (Sp) are calculated. A graph is plotted with sensitivity (Se) on the Y-axis and FPR = 1 – Sp (the proportion of false positives) on the X-axis.

For an ideal classifier, the ROC curve passes through the upper left corner, where the proportion of true positives is 100% (perfect sensitivity) and the proportion of false positives is 0%. The ROC plot is often supplemented with the diagonal line $Y = X$, which represents a "non-informative" classifier that outputs random class labels. The ROC curve for the examined case is shown in figure 2.

The quantitative interpretation of the ROC curve is provided by the AUC (Area Under the Curve) metric [16] – representing the area under the ROC curve and above the X-axis. The higher the AUC value, the better the classifier's performance. An AUC of 0.5 indicates a poor

classifier equivalent to random guessing, while an AUC of 1.0 corresponds to a perfect model. For the case under study, the AUC value is 0.9537.

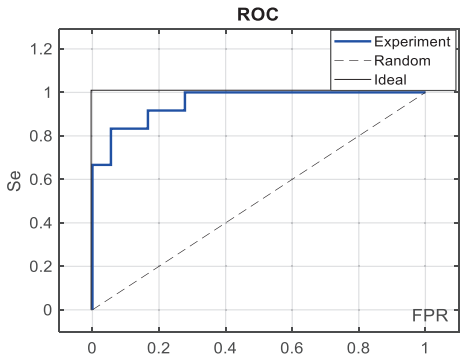


Fig. 1. ROC curve for ECG classifier

To improve diagnostic quality, the number of independent variables in equation (1) can be increased. For example, by introducing the seventh variable SDANN into the equation, the classifier's characteristics improved slightly (as shown in Table II). However, increasing the number of variables beyond seven did not lead to further improvement in classification quality.

IV. Conclusions

The comparative analysis of computerized diagnostic methods for cardiac arrhythmias conducted in this work showed that one of the most effective approaches is the use of binary classifiers based on the logistic regression model. Diagnostic accuracy can be improved by preliminarily extracting diagnostic features-integral indicators of cardiointervalograms or heart rate variability metrics, such as the mean and standard deviation of RR intervals, the proportion of RR intervals differing by more than 50 ms, the standard deviation of differences between successive intervals, and several others. The most informative features were selected by analyzing the multiple regression coefficients and assessing their statistical significance. Six parameters – SDNNi, NN50, pNN50, CV, Mo, and VR – were identified as the most informative and having a sufficiently high confidence level. A logistic regression model was constructed using these independent variables from the ECG training dataset and then tested on a separate test set. The experiment showed that the resulting logistic regression model provided satisfactory recognition performance, with specificity not less than 87% and sensitivity at 77%. The ROC curve built from the experiment allowed for calculating the recognition quality metric AUC=0.954, indicating a high-quality model. Adding the seventh parameter SDANN led to a slight improvement in sensitivity and specificity, while further increasing the number of variables did not enhance classification quality. Further investigations may affect identifying the features of the logistic model in the case of a linear relationship between variables, non-Gaussian data distribution, as well as an estimation of the computational complexity of the algorithm.

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Electromagnetic brake system self-sustained oscillations

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Abstract—**Magnetic brakes can be used as supplementary brake structure for railway vehicles. The nonlinear mathematical model of self-sustained oscillations is the aim of this paper. The lumped parameter technique describes the non-stationary magnetic flux, which is generated by mechanical oscillations of electromagnet. The electromagnetic subsystem is simulated by lumped parameter method. The self-sustained oscillations are generated in brake structure due to dry friction between railway and brake part. Nonlinear mathematical model of magnetic brake structure is obtained to study these oscillations. This model consists of two submodels: electromagnetic subsystem oscillations mathematical model and mechanical subsystem oscillations model. The system of eight nonlinear ordinary differential equations is derived to simulate oscillations. Stick-slip motions of the frame are analysed numerically.**

Keywords— self-sustained oscillation, magnetic brakes, non-stationary magnetic flux, lamped parameter technique

I. Introduction

Magnetic brakes are wheel-rail contact additional braking structure, which is mechatronic system [1]. Necessity of braking using the mechatronic structure is considered by several researches [2-6]. In [7], the mechanical design of magnetic brakes is discussed. The coupled lumped model, which consists of magnetic, electro and mechanical subsystems, are suggested in [8]. The phenomena, which results in self-sustained vibrations, are the following [9, 10]: the ectromagnetic-mechanical coupling between subsystems and the negative friction coefficient.

The nonlinear dynamic model of quarter of magnetic brake, is treated in Ref. [9-11]. The present conference paper is based on Ref. [12], where the mathematical model of nonlinear mechatronic system in the form of nonlinear ordinary differential equations is derived. The data of the numerical analysis of self-sustained oscillations of the mechatronic structure in the form of the electro-magnetic-mechanical system are discussed.

II. Problem formulation

The magnetic brakes self-sustained vibrations are studied. The magnetic brake is a frame with corner electromagnets (Fig.1). Every magnetic brakes side consists of two electromagnets, which are connected by track rods and beam. The frame structure and the bogie are connected by pneumatic actuators. When the magnetic brake is operated, electric voltage acts on electromagnets coil. Then, the magnetic force is generated and magnets are pulled on the rail. As a result, the friction between rail and magnetic brakes are generated. As follows from the experimental analysis, the dangerous self-sustained vibrations are generated during brake operation. If the velocity of the vehicle is equal to 25 km/h, the self-sustained vibrations with the frequency 28 Hz are excited.

III. Electro-magnetic subsystem

The circuit of the magnetic brake is shown on Fig. 2a. It has the voltage source u_{src}. The circuit has two electric paths. Every path consists of two electromagnets. The first path has electromagnets 1 and 4 and the second one has electromagnets 2 and 3.

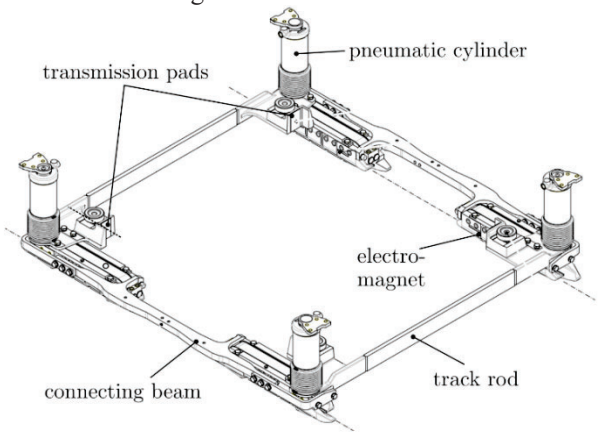


Fig. 1. Frame of magnetic brake

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The ordinary differential equations of electric subsystem have the form:

$$u_{src} = i_{14}R_{el} + N(\dot{\varphi}_1 + \dot{\varphi}_4); \quad (1)$$
$$u_{src} = i_{23}R_{el} + N(\dot{\varphi}_2 + \dot{\varphi}_3),$$

where N is the number of coil windings; $R_{el} = R_1 + R_4 = R_2 + R_3$; φ_i is the magnetic fluxes.

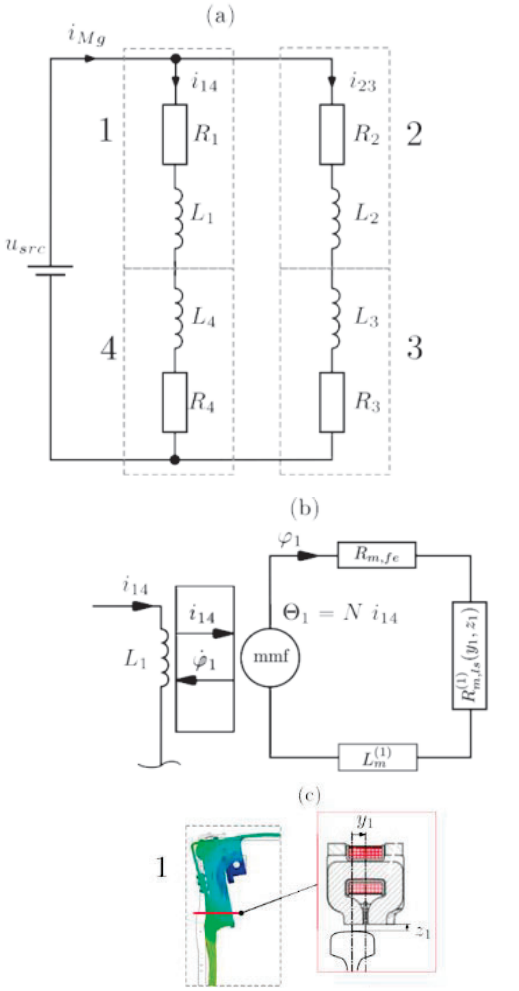


Fig. 2. (a) Electric circuit; (b) Magnetic subsystem; (c) Electromagnet displacement

The lumped parameter approach [12] is used to analyze magnetic subsystem. The distributive field approach [12] needs more calculations than lumped parameter approach. The magnetic field in each corner electromagnet (Fig.2b) is studied. Every magnetic circuit consists of: reluctance $R_{m,fe}$ of magnet iron part; variable reluctance of the air-gap $R^{(i)}_{m,ls}(y_i, z_i)$; loss of eddy $L^{(i)}_m = G\,\dot{\varphi}_i$. Eddy-current loss factor G has constant value. The coil magnetomotive forces are the following $\theta_1 = \theta_4 = i_{14}N; \theta_2 = \theta_3 = i_{23}N$. Magnetic current of parts are denoted by $i_{14}, i_{2,3}$.

Ordinary differential equations of the magnetic subsystem take the form:

$$i_{14}N = \varphi_1\left(R_{m,fe} + R^{(1)}_{m,ls}(y_1, z_1)\right) + G\dot{\varphi}_1;$$

$$i_{14}N = \varphi_4\left(R_{m,fe} + R^{(4)}_{m,ls}(y_4, z_4)\right) + G\dot{\varphi}_4;$$
$$i_{23}N = \varphi_2\left(R_{m,fe} + R^{(2)}_{m,ls}(y_2, z_2)\right) + G\dot{\varphi}_2;$$
$$i_{23}N = \varphi_3\left(R_{m,fe} + R^{(3)}_{m,ls}(y_3, z_3)\right) + G\dot{\varphi}_3. \quad (2)$$

The effective air-gap $s_i(y_i, z_i)$ is used to simulate the reluctance. The pole shoe lateral and transversal displacements are accounted in the value of air-gap (Fig.2c). The air-gap reluctance is the following:

$$R^{(i)}_{m,ls}(y_i, z_i) = \frac{s_i(y_i, z_i)}{\mu_0 A}. \quad (3)$$

with A is the magnetic cross-section; μ_0 is the vacuum permeability. The air-gap is approximated by

$$s_i(y_i, z_i) = \frac{a_{hyp}}{b_{hyp}}\sqrt{b^2_{hyp} + y_i^2} + 2z_i + a_0,$$
$$i = 1, \ldots, 4. \quad (4)$$

The values b_{hyp}, a_{hyp}, a_0 are obtained from finite element calculations of magnetostatic. It is supposed, that the rail is in contact with magnets ($z_i = 0$). The projections of the magnetic forces $F^{(i)}_{A,y}$, $F^{(i)}_{A,z}$, are obtained from the magnetic co-energy $W^*_{m,ls}$

$$W^*_{m,ls} = \sum_{i=1}^{4} \varphi_i^2 \frac{s_i}{2\mu_0 A}; \quad (5)$$

$$F^{(i)}_{A,y} = \frac{\partial W^*_{m,ls}}{\partial y_i} = \frac{\varphi_i^2}{2\mu_0 A}\frac{\partial s_i}{\partial y_i} = \frac{\varphi_i^2}{2\mu_0 A}\,\frac{a_{hyp} l_i \beta_i}{b_{hyp}\sqrt{b^2_{hyp} + l_i^2\beta_i^2}};$$

$$F^{(i)}_{A,z} = \frac{\partial W^*_{m,ls}}{\partial z_i} = \frac{\varphi_i^2}{2\mu_0 A}\frac{\partial s_i}{\partial z_i} = \frac{\varphi_i^2}{\mu_0 A},$$

where $y_i = \beta_i l_i$.

The equations (1-4) are rewritten into the nonlinear ordinary differential equations as:

$$\left(G + \frac{N^2}{R_{el}}\right)\dot{\varphi}_1 + \frac{N^2}{R_{el}}\dot{\varphi}_4 +$$
$$\left[R_{m,fe} + \frac{a_{hyp}\sqrt{b^2_{hyp} + l_1^2\beta_1^2}}{\mu_0 A b_{hyp}} + \frac{2z_1 + a_0}{\mu_0 A}\right]\varphi_1 = \frac{N}{R_{el}}u_{src}; \quad (6)$$

$$\frac{N^2}{R_{el}}\dot{\varphi}_1 + \left(G + \frac{N^2}{R_{el}}\right)\dot{\varphi}_4 +$$
$$\left[R_{m,fe} + \frac{a_{hyp}\sqrt{b^2_{hyp} + l_4^2\beta_4^2}}{\mu_0 A b_{hyp}} + \frac{2z_4 + a_0}{\mu_0 A}\right]\varphi_4 = \frac{N}{R_{el}}u_{src};$$

$$\left(G + \frac{N^2}{\bar{R}_{el}}\right)\dot{\varphi}_2 + \frac{N^2}{\bar{R}_{el}}\dot{\varphi}_3 +$$
$$\left[R_{m,fe} + \frac{a_{hyp}\sqrt{b^2_{hyp} + l_2^2\beta_2^2}}{\mu_0 A b_{hyp}} + \frac{2z_2 + a_0}{\mu_0 A}\right]\varphi_2 = \frac{N}{\bar{R}_{el}}u_{src};$$

$$\frac{N^2}{\bar{R}_{el}}\dot{\varphi}_2 + \left(G + \frac{N^2}{\bar{R}_{el}}\right)\dot{\varphi}_3 +$$
$$\left[R_{m,fe} + \frac{a_{hyp}\sqrt{b^2_{hyp} + l_3^2\beta_3^2}}{\mu_0 A b_{hyp}} + \frac{2z_3 + a_0}{\mu_0 A}\right]\varphi_3 = \frac{N}{\bar{R}_{el}}u_{src}.$$

The electromagnetic subsystem is coupled with the mechanical subsystem. The mathematical model of the oscillations of mechanical subsystem will be considered in the next section.

IV. FRAME NONLINEAR OSCILLATIONS

The frame structure is studied as an elastic structure. Its linear oscillations are investigated in software ANSYS. Fig.3 shows three eigenmodes of this frame.

Material properties of the frame are equal to structural steel. The frame structure is discretised by 65000 finite elements, which include quadratic hexahedral elements and quadratic tetrahedral elements.

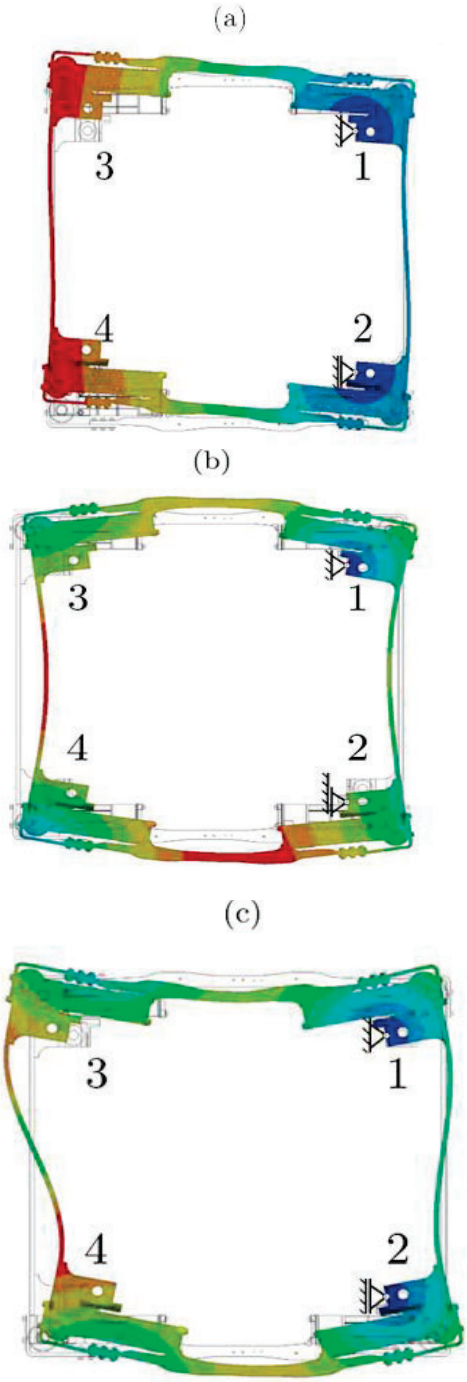


Fig. 3. Three eigenmodes: (a) asymmetric eigemodemode (8 Hz); (b) symmetric eigenmode (16 Hz); (c) asymmetric eigenmode (30.42 Hz)

Fig.3 shows the first three eigenmodes of the frame. Fig.3c shows the third eigenmode, which is asymmetric. The frequency of this third eigenmode is 30.4 Hz. This frequency is very closed to the self-sustained oscillations frequency, which is measured experimentally, 28 Hz [9]. Therefore, the third eigenmode is taken as mode of the frame nonlinear vibrations.

The structure frame consists of thin rods and thin beams. The masses of these parts are significantly less, than the masses of the electromagnets. The electromagnets rigidity is significantly greater, than the rigidity of roods and beam. Therefore, the finite degrees of freedom frame model can be constructed. This model consists of four rigid bodies, which describe the oscillations of electromagnets, and rods and beams, which has only elastic properties. As follows from Fig.3c, electromagnets have only rotation motions. Therefore, the dynamical model has four degrees of freedom. This dynamical model is shown on Fig.4. The vibrations of the frame is described by four generalized coordinates $\beta_1, \dots, \beta_4$, which is characterize the rotation motions of electromagnets.

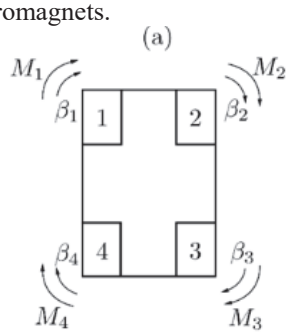


Fig. 4. Frame mechanical model

Finite degree of freedom mechanical model is derived by using the influence coefficients δ_{ij}. If the moments $M_1, \dots, M_4$ act on discrete masses (Fig.4), then the generalized coordinates $\beta_1, \dots, \beta_4$ and moments satisfy the equations:

$$\beta_i = \sum_{j=1}^{4} \delta_{ij} M_j \, ; i = 1, \dots, 4, \qquad (7)$$

The influence coefficients are calculated by using finite element static analysis. The unit moment acts on one electromagnets and the rotations of others corner electromagnets is equal to influence coefficients.

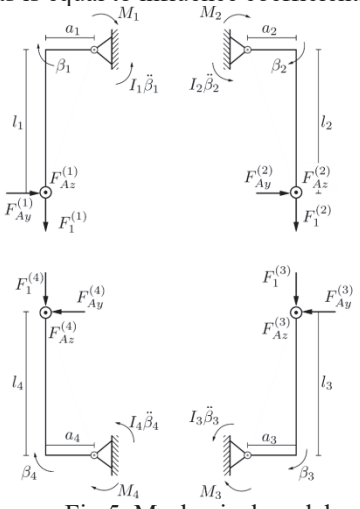


Fig.5. Mechanical model

Model consists of four electromagnets. Each magnet performs rotational oscillations β_i (Fig.5). The moments acting on the electromagnets M_j satisfy the following relations:

$$M_i = -a_i F_1^{(i)} - l_i F_{A,y}^{(i)} - I_i \ddot{\beta}_i; i = 1,4; \quad (8)$$

$$M_j = a_j F_1^{(j)} - l_j F_{A,y}^{(j)} - I_j \ddot{\beta}_j; j = 2,3; \quad (9)$$

$$F_1^{(i)} = \mu\left(v_* + \dot{\beta}_i a_i\right) sgn\left(v_* + \dot{\beta}_i a_i\right) F_{A,z}^{(i)};$$

$$F_1^{(j)} = \mu\left(v_* - \dot{\beta}_j a_j\right) sign\left(v_* - \dot{\beta}_j a_j\right) F_{A,z}^{(j)},$$

where $\ddot{\beta}_\iota = d^2\beta_i/dt^2$; I_i are the moments of inertia; t is time; a_i, l_i are electromagnet geometrical parameters (Fig. 5); $F_{A,y}^{(i)}, F_{A,y}^{(j)}, F_{A,z}^{(i)}, F_{A,z}^{(j)}$ are the projections of the magnetic forces; $F_1^{(i)}, F_1^{(j)}$ are friction forces.

The moments acting on the electromagnets takes the form:

$$M_j = a_j F_1^{(j)} - \frac{\varphi_j^2 a_{hyp} \beta_j l_j^2}{2\mu_0 A b_{hyp} \sqrt{b_{hyp}^2 + l_j^2 \beta_j^2}} - I_j \ddot{\beta}_j; j = 2,3 \quad (10)$$

$$M_i = -a_i F_1^{(i)} - \frac{\varphi_i^2 a_{hyp} \beta_i l_i^2}{2\mu_0 A b_{hyp} \sqrt{b_{hyp}^2 + l_i^2 \beta_i^2}} - I_i \ddot{\beta}_i; i = 1,4.$$

The matrix of influence coefficients $\boldsymbol{\delta} = (\delta_{ij})$ and the matrix of stiffness $\mathbf{K} = (K_{ij})$ satisfy the following equations: $\boldsymbol{\delta}^{-1} = \mathbf{K}$.

Then the equations of the frame motions take the following matrix form:

$$\begin{bmatrix} I_1 \ddot{\beta}_1 \\ \vdots \\ I_4 \ddot{\beta}_4 \end{bmatrix} = \begin{bmatrix} \widetilde{\mathcal{M}}_1\left(\beta_1, \dot{\beta}_1, \varphi_1\right) \\ \vdots \\ \widetilde{\mathcal{M}}_4\left(\beta_4, \dot{\beta}_4, \varphi_4\right) \end{bmatrix} - [\mathbf{K}] \begin{bmatrix} \beta_1 \\ \vdots \\ \beta_4 \end{bmatrix}, \quad (11)$$

$$\widetilde{\mathcal{M}}_i\left(\beta_i, \dot{\beta}_i, \varphi_i\right) = -a_i \mu\left(v_* + \dot{\beta}_i a_i\right) sgn\left(v_* + \dot{\beta}_i a_i\right) \frac{\varphi_i^2}{\mu_0 A} - \frac{\varphi_i^2 a_{hyp} \beta_i l_i^2}{2\mu_0 A b_{hyp} \sqrt{b_{hyp}^2 + l_i^2 \beta_i^2}}; \quad i = 1,4;$$

$$\widetilde{\mathcal{M}}_j\left(\beta_j, \dot{\beta}_j, \varphi_j\right) = a_j \mu\left(v_* - \dot{\beta}_j a_j\right) sgn\left(v_* - \dot{\beta}_j a_j\right) \frac{\varphi_j^2}{\mu_0 A} - \frac{\varphi_j^2 a_{hyp} \beta_j l_j^2}{2\mu_0 A b_{hyp} \sqrt{b_{hyp}^2 + l_j^2 \beta_j^2}}; \quad j = 2,3. \quad (12)$$

The nonlinear oscillations of electromagnetic-mechanical brake system are described by system of eight nonlinear ordinary differential equations (6, 12).

V. NUMERICAL ANALYSIS OF ELECTROMAGNETIC-MECHANICAL SYSTEM OSCILLATIONS

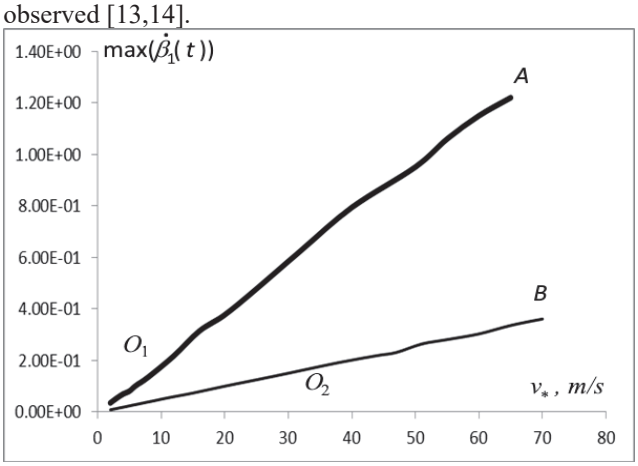
The finite degrees of freedom model parameters (electromagnets moments of inertia and matrix of influence coefficients) are obtained from finite element analysis in software ANSYS.

The system linear analysis is carried out to obtain the third eigenfrequncy $\omega_3 = 30.42$ Hz, which is calculated by using finite element method. Finite degrees of freedom model (11) is used to obtain the eigenvalue problem with respect to eigenfrequencies. The numerical values of the system parameters are presented in Table 1. As follows from the numerical calculations, the third eigenfrequency is equal to $\bar{\omega}_3 = 32.3$ Hz. Thus, the eigenfrequencies, which are obtained from the big finite element model and from four degrees of freedom lumped model, are closed.

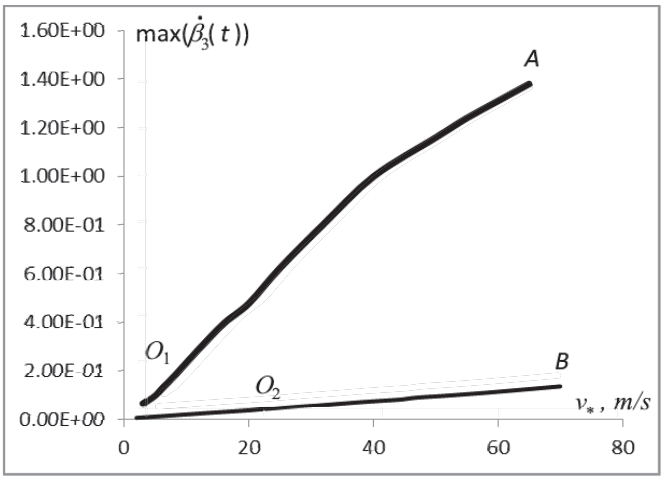
Table 1. Magnetic brake parameters

Parameter	Symbol (unit)	Value
Moments of inertia	$I_1 = I_2$ (kgm²) $I_3 = I_4$ (kgm²)	6.02 13.79
Geometrical dimension	$a_1 = a_2 = a_3 = a_4$ (m)	0.406
Geometrical dimension	$l_1 = l_2 = l_3 = l_4$ (m)	0.027
Vertical distance	$z_1 = z_2 = z_3 = z_4$ (m)	0
Empirical friction constant	γ (s/m)	0.107
Static friction coefficient	μ_S (1)	0.42
Parameter of the magnetic subsystem	a_{hyp0} (1)	-0.002
Parameter of the magnetic subsystem	a_{hyp} (1)	0.0021
Parameter of the magnetic subsystem	b_{hyp} (1)	0.01
Effective magnetic cross-section	A (m²)	0.003
Eddy current loss factor	G (1/Ω)	3000
Magnetic permeability of vacuum	μ_0 (N/A²)	0.12e-5
Magnetic resistance of iron	$R_{m,fe}$ (A/V/s)	455000
Electric resistance	R_{el} (Ω)	10
Number of coil windings	N (1)	300
Battery voltage	u_{src} (V)	100

The nonlinear oscillations of magnetic frame are studied numerically for different values of velocity of the vehicle v_*. The vehicle velocity is changed according to the equation: $v_{*,j} = v_{*,0} + jh_v; j = 1,2,\ldots$, where $v_{*,0}$ is initial velocity. In order to study the frame oscillations, the nonlinear dynamic of electromagnetic-mechanical system (6, 11) is integrated numerically taking different initial conditions. The initial time interval $t \in [0; 400]$ is assumed transient. Then the steady self-sustained oscillations are observed [13,14].



a.



b.

Fig.6. The dependence of self-sustained vibrations amplitudes on vehicle velocities

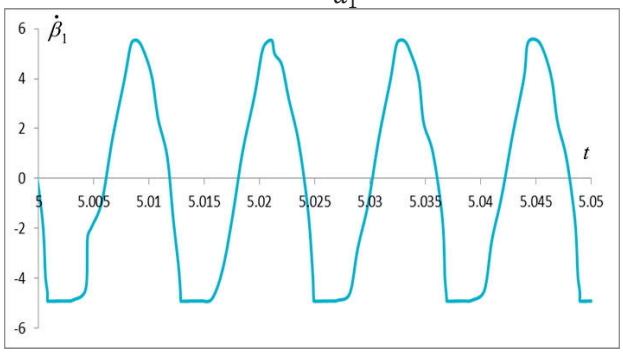
Fig.6 shows the periodic self-sustained oscillations amplitudes $\max\left(\dot{\beta}_1(t)\right), \max\left(\dot{\beta}_3(t)\right)$ versus the velocities of the vehicle. Two different branches of bifurcation diagram (Fig.6) (O_1A) and (O_2B) describe the self-sustained oscillations with big and small amplitudes, respectively.

The data of direct numerical integration at v_*= 2 m/s are shown on Fig. 7. The generalized velocity $\dot{\beta}_1(t)$ is presented on Fig.7a. The frame dynamics has section with constant generalized velocity $\dot{\beta}_1 = const$, that is stick section of oscillations. The constant value of the generalized velocity $\dot{\beta}_1$ is determined as

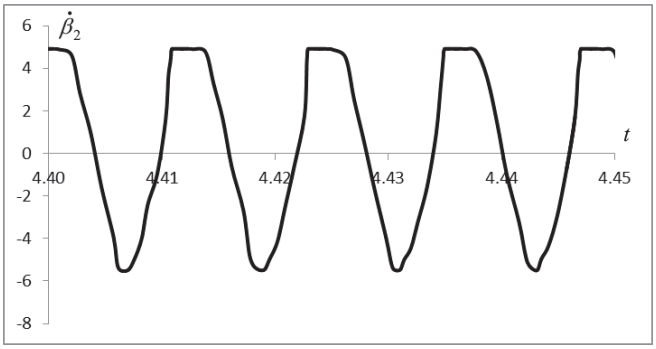
$$\mu\left(v_* + \dot{\beta}_i a_i\right) sgn\left(v_* + \dot{\beta}_i a_i\right) = 0; i = 1{,}4$$

Then the constant value of the generalized velocity is derived from the equation:

$$\dot{\beta}_1 = -\frac{v_*}{a_1}.$$



a.



b.

Fig.7. Oscillations of electromagnets $\dot{\beta}_1, \dot{\beta}_2$ at v_* = 2 m/s

VI. Conclusions

New mathematical model of electromagnetic brake nonlinear oscillations is suggested. The electromagnetic subsystem is described by the system of four nonlinear ordinary differential equations, which simulate the magnetic flux of magnetic circuit. The nonlinear dynamics of brake frame is described by four degrees of freedom mechanical model, where bodies of electromagnets are considered rigid performing rotation motions. Frame subsystem interacts with electromagnetic subsystem.

The nonlinear oscillations are studied numerically in vehicle velocity band. The nonlinear oscillations with small and large amplitudes are studied. At low vehicle velocity the Neimark-Sacker bifurcation is observed.

Acknowledgement

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Intelligent Motor Testing System Utilizing Neural Networks

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***Abstract*—The article proposes a system for diagnosing the technical characteristics of electric motors, which includes the possibility of measuring and collecting data on the torque of the electric motor, the speed of rotation of the shaft, and the electric power. The experiments were carried out using the developed prototype of the measuring stand. With its use, diagnostics of the operation of the collector electric motor of direct current in various modes of operation, including measurements at start, acceleration, stable mode, were carried out. For this, known methods of measuring the electrical power of an electric motor were applied, and an improved method of measuring the torque was proposed, which is based on measuring the impulse characteristics of photo-sensors that record the rotation of two discs placed on the shaft, between which a dynamometric spring is embedded. This method allows you to obtain the angle of rotation of the shaft, which is proportional to the rotational moment. The measurement system presented in the article allows detecting abnormal modes of operation of electric motors by using a neural network with the architecture of a variational autoencoder. Such a neural network allows to classify the standard modes of operation of the electric motor, as well as abnormal changes in such modes. Testing of the system on the example of a collector DC motor was carried out using the developed prototype of the measuring stand, which made it possible to establish a number of anomalous modes of operation during start-up.**

Keywords— measurement, testing, stand, torque, power, engine, energy

I. Introduction

Electric motor test systems play a key role in the field of power generation and research. They provide the necessary information about the characteristics and efficiency of engines, which in turn affects the reliability, productivity and safety of a wide range of electric machines, starting from their use in industrial equipment and ending with vehicles. Therefore, modern research focuses on improving electric motor test systems to obtain more accurate information about their operation.

The combination of intelligent means of diagnostics of electric motors with standard means of measurement can increase the efficiency of their use and durability. However, this area remains under-researched due to differences in engine characteristics and testing requirements. Thus, the existing methods may not be sufficiently reproducible in real conditions, which requires the creation of complex and universal intelligent measurement systems that can take into account the real production conditions of operation and the peculiarities of the technical characteristics of electric motors.

Diagnostics of the rotation parameters of electric motors can be carried out during operation, or using modeling complexes. However, stand testing can be expensive and complex due to the need for dismantling and transportation. Therefore, modern universal solutions based on artificial intelligence can help adapt to engine conditions and characteristics. Thus, the main task that engineers face today is to identify operating conditions that foreshadow a breakdown or accident. This problem can be solved using neural networks or machine learning.

II. Literature review

Modern means of modeling electric drive systems are not limited to testing a separate electric motor, today complex testing of power plants, including gearboxes and additional loading equipment, is becoming more and more popular. This allows us to cover a wide range of tasks. Yes, not today research is conducted on testing electric cars [1], testing industrial robots [2], airplanes with electric engines [3] and other research in various fields.

At the same time, a significant part of the work is devoted to testing electric motors using separate stands, aimed at studying more detailed characteristics and modes of operation, including peak capabilities. Such studies include attempts to identify effective methods of cooling electric motors [4]. The possibilities of machine learning for diagnosing malfunctions of electric motors are studied [5]. There is a study of vibrational characteristics of rotational parameters [6]. Such research has become widely used in the aviation industry, where work is underway to adapt such diagnostic tools for more specific tasks [7] Research is being conducted on the operation of asynchronous electric motors depending on the frequency characteristics of the voltage produced by the inverter converters [8]. The methods of controlling the magnetic properties of the stator coils depending on the angles of interaction with the rotor and mutual induction are being improved [9]. Vibration parameters and other parameters that lead to further failure of the electric motor are tested [10]. Thus, the most relevant work is to determine the operational characteristics of an electric motor through the identification of anomalous

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modes. Major problems include excessive load leading to overheating and damage to the windings, corrosion, insulation aging and mechanical failure. Bearings can suffer from vibration, improper lubrication and insufficient cooling, while unstable power supply and frequency fluctuations can cause further problems. Thus, the processing of data from measuring devices must recognize critical conditions that affect the parameters of the electric motor. At the current stage, scientists are working on detecting sharp signal fluctuations in real time associated with changes in rotation parameters, such as shaft load, rotation speed and winding current, etc. Thus, in [11], a high-speed method for determining changes in the dynamic parameters of electric motors is proposed, which is based on known regularities and is determined algorithmically using simple logic. Considering the cost of server hard-ware and primary instrumentation for monitoring motor performance, the proposed system is complex and can be expensive. But despite this, modern methods for diagnosing electric motors are mainly focused on analytical diagnostics, including the identification of dangerous conditions. For example, work [12] presents a new approach to the diagnosis of electric motors through the analysis of vibration data, which can prevent catastrophic accidents. The proposed intelligent method of detecting anomalies by analyzing the vibration signals of laboratory-type electric motors and old industrial motors uses Feedforward Neural Networks (FNNs). So, in this work, the model was trained using the metaheuristic optimization algorithm (GTO). The mean squared error (MSE) between the actual and predicted outputs is chosen as the objective function. The optimization was carried out with the help of 30 search agents that calculated 61 weights and bias coefficients for neurons in the range from -10 to 10. Typical settings of the control parameters of the GTO algorithm were used.

However, this model was tested only under artificial conditions. To improve its practical applicability, it is necessary to carry out validation using real data. In addition, considering that the typical settings of the GTO algorithm were used, further settings and optimization of hyperparameters can improve the effectiveness of this model, especially for certain conditions of use of electric motors, where the level of vibration and vibration modes differ. Work [13] highlights the possibilities of using (IoT) to determine anomalies of rotating industrial machines also by means of vibration monitoring. For this, an MPU6050 accelerometer and a Raspberry Pi 4B were used to calculate the data. The support vector method (SVM) was applied to classify abnormal operating modes. The results obtained with such a solution are visualized using a web application in real time. The proposed system could be improved by using more powerful machine learning algorithms or a neural network. However, for this, a single-chamber Raspberry Pi computer will not have much computing power. Of course, the detection of anomalies using an electric motor of a separate type can be implemented schematically and algorithmically, as for example in [10], where an algorithm for evaluating anomalies in the operation of an electric motor of a drilling rig is proposed by analyzing the variables of torque, Coulomb friction and geometry systems. However, this method may be specific to a particular type of drill string or particular operating conditions, so it may be difficult to adapt it to other scenarios. In addition, the proposed model may be sensitive to the values of some parameters, such as the temperature of the drilling environment, rock properties, which may also limit its ability to determine anomalies in the operation of the power drive.

Based on the analysis of the literature, it is possible to draw attention to the fact that there are quite a few universal systems for diagnosing anomalous modes of operation of electric motors, which prompted the development of its prototype.

III. Research Methodology

The conducted research is aimed at the occurrence of anomalies in the operation of electric motors by building a universal intelligent information-measuring diagnostic system that can be applied to many types of electric machines. To do this, we propose the structure of such a system (Fig. 1, Fig. 2), which has a number of sensors, and also uses the method of measuring the rotational moment proposed by the authors, which is based on Hooke's laws, where the twist angle is the value of the proportional rotational moment:

$$M = f(\varphi_C) = \frac{10^6 \pi \varphi_C G_P J}{180 L}, \quad (1)$$

where φ_c - the torsion angle of the torsion shaft; G_p - modulus of elasticity shaft material; J - the moment of inertia of the cross-section of the shaft; L - the working length of the torsion shaft.

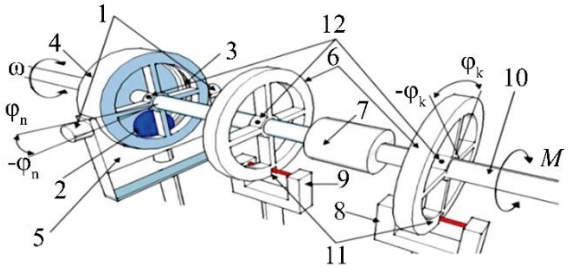


Fig. 1. Stand for measuring torque, rotation frequency, acceleration, voltage and current of electric motors

This device is a measuring unit for determining torque by detecting the twist of an elastic element and measuring the phase shift of pulses from two photodetectors. The slotted disk (1) is rigidly mounted on the input end of the shaft and allows light to pass through the slots (2) to the photodetector (4), which is installed on a mounting support (5). The drive shaft (3) transmits rotation from the electric motor to the measuring unit. The second slotted disk (6) is connected to the first disk via an elastic element (7). This element twists under the action of torque, causing the slots of the rear disk to shift relative to the front one. The photodetector (8), mounted on support (9), registers the passage of the rear disk's slots (11). The output shaft (10) transmits torque to a load or another part of the test stand. A rigid sleeve or coupling (12) ensures the coaxiality of both disks and transmits rotation through the elastic element.

The operating principle is that when torque is applied, the elastic element twists, and the slots of the two disks shift relative to each other. This creates a phase shift between the photodetector signals, which is proportional to the torque value. The electronic processing system then measures this shift and calculates the torque.

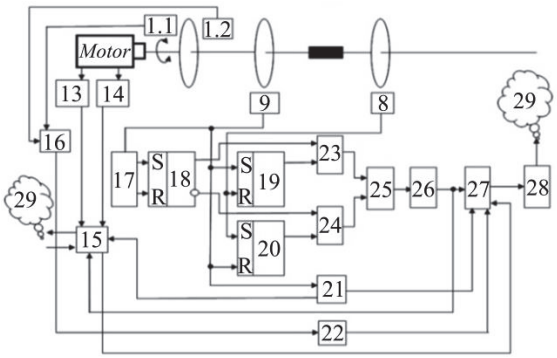


Fig. 2. Structure of the measuring system

The presented diagram shows the structural electronic circuit of a measurement and signal processing system for an electric motor test stand. It consists of measurement units, logic processing components, and data transmission modules.

Motor – the tested electric motor. 1.1 and 1.2 – slotted disks for generating pulse signals. 8 and 9 – photodetectors that register the passage of the slots and generate pulses for measuring angular speed and phase shift. 13 and 14 – voltage and current sensors connected to the electric motor. 15 and 16 – interfaces or switches for transmitting sensor signals to the processing unit. Modules 17, 18, 19, and 20 are RS-type triggers and logic elements that latch the states of signals from the sensors. 21 and 22 – synchronization and control units that coordinate the operation of the measurement system. Blocks 23 and 24 are interface modules for transferring signals to the subsequent calculation stages. 25 and 26 – computational or conversion units (possibly counters or timers) that form digital values of speed, torque, or power. 27 and 28 – output processing modules (possibly a microcontroller or communication interface). Elements marked as 29 indicate power supply sources or connections to a data acquisition system (for example, a PC or web interface).

Operating principle: signals from the photodetectors (8, 9) and the voltage/current sensors (13, 14) are sent to logic units (17–20), where the pulses are processed and their arrival moments are recorded. They are then transmitted through interface modules (23, 24) to computational blocks (25, 26), which determine the motor's parameters. The obtained results are sent to the output modules (27, 28) for display, storage, or further analysis.

The basis of the method is the use of two coaxial disks 6, connected through an elastic element 7, at the ends of which slots 11 are placed, through which the world rays read by photocells 8, 9 pass. In the absence of a rotational moment, the angular displacement be-tween slots are absent. During the rotation of the disks, two sensors placed on a rigid base record the pulses that are generated after passing one revolution of the shaft. In the presence of a rotational moment, the elastic element 7 twists, changing the relative position of the slots 11, and the time interval between the pulses of the sensors 8, 9 becomes directly proportional to the twisting angle (rotational moment, respectively) and inversely proportional to the angular speed of the shaft (Fig. 3).

This allows you to measure the moment without the need to use complex technology. So, the time interval Δt can be determined by the following expression:

$$\Delta t = \frac{dl}{V_{rpm}}, \tag{2}$$

where dl is an element of the length of the circumference of the radius of the disks 6, which is determined by the angle of twist of the elastic element;

$$V_{rpm} = \frac{2 \cdot \pi \cdot R_{\mathrm{D}}}{T} = \omega \cdot R_D.$$

where R_D - the radius of the circle, disks 6; T – the rotation period of the engine shaft; ω - angular velocity.

The torsion angle of an elastic element, which depends on the moment applied to it, can be represented as the product between the rotational moment and the coefficient depending on the property of the elastic element $dl = M \cdot K_1$. In turn, expressing the value of the measured torque from this expression, we get the torque:

$$M = \frac{dl}{K_1} = \frac{\Delta t \cdot V_{rpm}}{K_1} = \frac{\Delta t \cdot 2 \cdot \pi \cdot R_{\mathrm{D}}}{K_1 \cdot T} = \frac{\Delta t}{T} \cdot K_2 \tag{3}$$

where $K_2 = \frac{2 \cdot \pi \cdot R_{\mathrm{D}}}{K_1}$ - the proportionality factor.

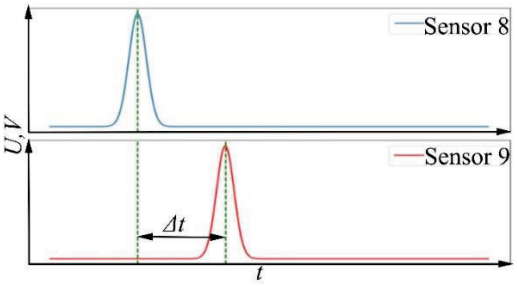


Fig. 3. Phase shift between sensor pulses 8, 9

Thus, the task of determining the torque is reduced to measuring the time intervals Δt and the shaft rotation period T, which can be performed with high accuracy, and the constant K_2 is determined in advance during the calibration process using accurate means of measuring the torque. Taking into account random additive noise, depending on the types of pulses, a method is proposed that allows you to determine which fragment of certain reference characteristics of the signal is contained in the analyzed spectrum at a certain time. This made it possible to select a suitable sample of data in the total amount for further analysis.

Thus, based on the values of the output signal y(t) and its first derivative, it is possible to determine which of the reference functions is present at the given time in the analyzed signal, where there is a finite set of reference functions $f_i(t)$, where $t \in [0; T_i]$, i = 1,2…M, characterizing standard signals. The analyzed signal in the case of an additive error will have the form:

$$y(t) = kf_i(t + \tau_i) + \eta(t), \tag{4}$$

where: $f_i(t)$ - i-th reference function; $\tau \in [0; T_i]$ – time shift between the signal and the i-th standard; η(t) - an additive obstacle, about which we only know that it can disappear and appear at random moments of time; k - coefficient whose value is unknown.

For the multiplicative error, the output signal will have the following expression:

$$y(t) = kf_i(t+\tau_i)\eta(t), \quad (5)$$

For some reference signals, such as rectangular and dusty, the first derivative may not exist, or may be constant and equal to zero for a significant time interval. This makes it difficult to determine disproportionality. In such conditions, it is possible to use the integral disproportionality by the derivative of the first order for parametrically specified functions, without using derivatives:

$$y(t) = \frac{\int_{t-h}^{t} y(t)dt}{\int_{t-h}^{t} f_i(t+\tau_i)dt} - \frac{y(t)}{f_i(t+\tau_i)} \quad (6)$$

where: h - the specified time interval; $f_i(t + \tau_i)$ - signal disproportionality function at a given time shift τ_i.

In order to identify anomalous modes of operation, a neural network of the Auto-encoder type was used [15]. Detection of anomalies in the operation of the electric motor using a neural network of this type is implemented by minimizing the data reconstruction error. So, after training on already existing data including power, torque and rotation speed, the model is used to restore all the data. The hypothesis is that anomalous modes of operation will have a higher reconstruction error. Once the reconstruction error exceeds a fixed threshold, such modes are classified as anomalies. So, the input data will be the vector $X = [x_1, x_2, x_3]$, where x_1 - the speed of rotation, x_2 - the electric power, x_3 - the torque. The coding layer can be described as: $h = f(W_e \cdot X + b_e)$, where W_e and b_e are the weights and offsets of the coding layer, and f is the activation function.

Decoding can be described as follows: $\hat{X} = g(W_d \cdot h + b_d)$, where W_d and b_d are the weights and biases of the decoding layer, and g is the activation function.The RMSE is used to measure the difference between the input data and its reproduction:

$$L(X, \hat{X}) = \frac{1}{n}\sum_{i=1}^{n}(x_i - \hat{x}_i)^2 \quad (7)$$

If exceeds a certain threshold level, the input is considered abnormal. The Adam optimization algorithm was used to minimize the loss function and learn the weights of the encoding and decoding layers.

IV. Results and Discussion

To test the proposed methods for recognizing anomalies in the operation of electric motors, a measuring stand was built (Fig. 4). Using it, a study of the operation of the RS-775 DC collector electric motor was conducted, at a nominal voltage of 12 volts, the technical characteristics of which are presented in Table I.

TABLE I. TECHNICAL CHARACTERISTICS OF THE DC COLLECTOR ELECTRIC MOTOR, [14]

No load		At maximum efficiency				Stall	
Current	Speed	Current	Speed	Torque	Power	Torque	Current
mA	r/min	mA	r/min	g-cm	W	g-cm	mA
<700	5000	<4.0	3750	650	24.4	2700	14.0

The electrical power characteristics were recorded using standard measuring devices. The speed of rotation is determined by counting pulses using a special microcontroller timer. Measuring transducers, which are included in the prototype of the measuring stand, have the ability to transmit data during measurement using standard data transmission interfaces. A special web application was developed to process measurement data.

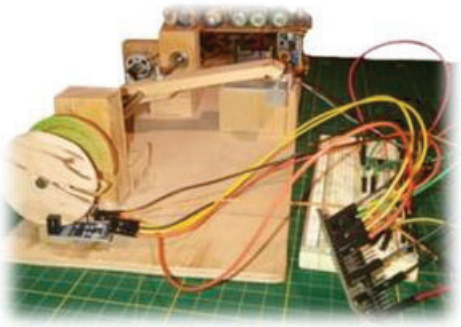


Fig. 4. Prototype for measuring the torque of the voltage and current of the DC collector electric motor.

The study of the torque of the electric motor using method (3) (Fig. 3) was carried out both in the no-load mode (Fig. 5, a) and under different loads.

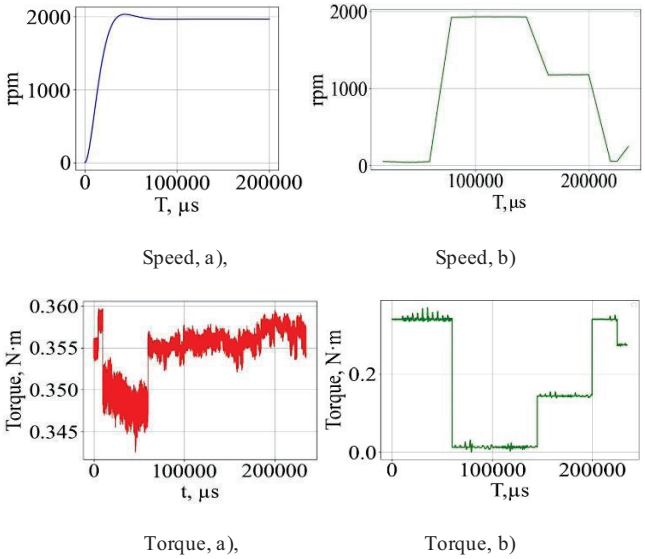


Fig. 5. Measurement of torque and shaft rotation speed using a stand prototype. (a) - in no-load mode, (b) - with load: 0.35; 0.28; 0.15; 0.02 Nm

In particular, using the laboratory brake, a load on the shaft of 0.35 was created; 0.28; 0.15; 0.02 Nm, which made it possible to obtain the following results in Fig. 5, b.

From (Fig. 5, b), it can be seen that the obtained sample of data obtained as a result of the measurement of the rotational moment is characterized by a significant dispersion due to the vibration characteristics of the stand itself. In such conditions, it is quite difficult to obtain abnormality criteria. Therefore, Gaussian smoothing was applied to this sample:

$$G(x) = \frac{1}{\sqrt{2\pi\sigma^2}} e^{-\frac{x^2}{2\sigma^2}}, \quad (8)$$

where σ - the standard deviation that defines the width of the Gaussian kernel.

This made it possible to obtain a more stable signal when measuring torque under different loads (Fig. 5, b), which was analyzed in order to determine the integral disproportionality of the signal, using a software algorithm in accordance with expression (6). This made it possible to identify manifestations of abnormal vibrations (Fig. 6)

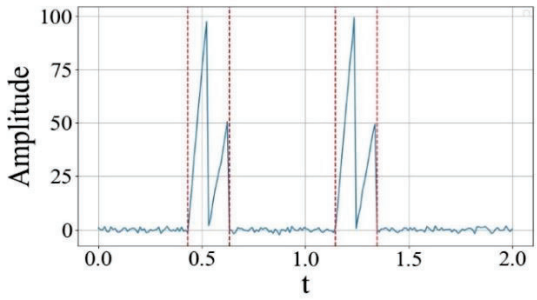


Fig. 6. Determination of integral disproportionalities.

However, this method does not allow recognizing a larger number of anomalies. Therefore, to determine them, a neural network of the Autoencoder type was used, at the input of which data on torque, electric power and rotation speed were given, without load on the shaft. As a result of learning the neural network, a model of the stable operation of the electric motor was obtained, which was used to evaluate the anomalous modes of operation of another electric motor, which made it possible to recognize anomalies of loads, rotation speed and electric power (Fig. 7).

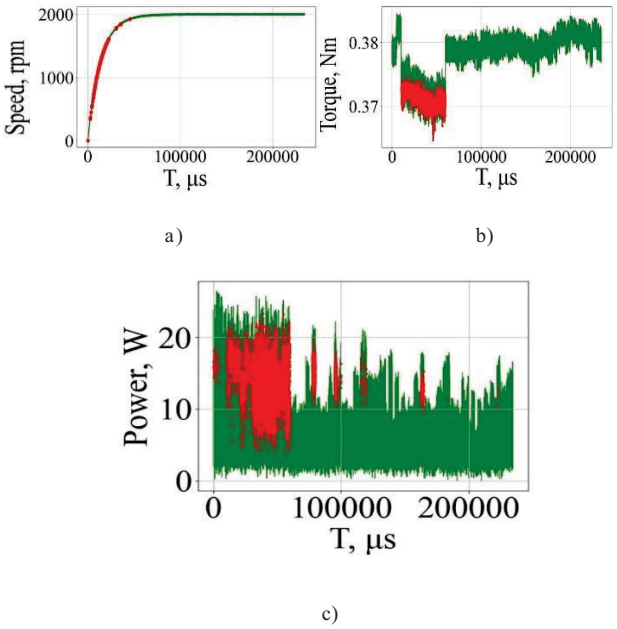


Fig. 7. Identification of abnormal signals from sensors: a - rotational torque; b - speed of rotation; c – powers

The obtained results indicate the detection of abnormal signals during the start of the electric motor. In particular, it identifies additional vibrations and loads during the start-up period, which are not inherent in the stable mode of operation.

The result of neural network training, as the difference between the input data obtained using another electric motor of the same type on which the training was implemented and the reproduced data of the model, can be represented as a function of losses by expressing them as a root mean square error (7). This indicator in general for all observations of this model was 0.0037, which allows us to assess how well the model works on the entire data set. At the same time, the assessment of its efficiency depends not only on the deviation, but also on the permissible limits of such a deviation, the features of the electric motor, the load, etc. Considering that the model was tested under the same conditions with the same type of electric motor, this result can be considered satisfactory. The dynamics of the MSE change and the anomaly threshold are presented in Fig. 8, which confirms the results of previous studies on each of the three parameters of the electric motor, which were also in the zone of abnormality during its start-up.

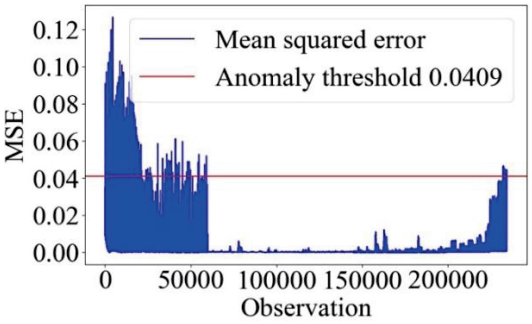


Fig. 8. Dynamics of MSE change and anomaly threshold

V. Conclusion

This study demonstrates that combining a deterministic "stand–engine" measurement system with intelligent data analysis methods based on a neural network ensures increased accuracy and reliability in diagnosing electric motors under real operating conditions. Despite the deterministic design principle of the measurement unit, testing inevitably involves random (stochastic) error components caused by vibrations, mechanical play, power supply instability, temperature variations, and electromagnetic interference. Such factors cannot be fully eliminated by traditional systematic error compensation methods; therefore, the use of a neural network enables effective noise filtering, detection of atypical operating modes, and adaptation to changing conditions.

The intelligent module of the system not only reduces measurement errors but also serves as a tool for promptly detecting anomalies associated with both random and systematic deviations — including mechanical defects, electrical faults, or load instability. Performing calibration before each test is an integral part of ensuring the high accuracy of any measurement system and, in this case, guarantees consistency between the hardware components and the data processing algorithm.

The obtained results confirm the feasibility of using a neural network even in deterministic measurement complexes, as it improves measurement accuracy, provides adaptability to uncontrolled factors, and enables timely detection of potentially hazardous operating modes of the electric motor. Considering the wide selection of mathematical models that allow identifying a separate group of data in the signal, it is possible to detect anomalies in the operation of electric motors most effectively using neural networks. The analysis of the received signal from the torque converter built in the form of a prototype of the diagnostic stand using formula (6) showed the possibility of distinguishing only individual pulsations that have a higher amplitude compared to others. At the same time, taking into

account that the diagnosis of abnormal work should be based on the analysis of not only the amplitude and shape of the signal, but also their certain sequence, preference can be given to neural networks. Training of a neural network of the Autoencoder type with the use of a direct current collector electric motor made it possible to determine anomalous modes of operation in a similar motor. As a result of model testing, MSE was 0.00379. The abnormality threshold when applying the 99-th percentile was 0.04, which can satisfy the solution of similar problems. The proposed method of measuring the torque, based on the analysis of the torsion angle of the elastic element and the measurement of the impulse characteristics of two sensors (3), as a result of practical application, made it possible to obtain data for analysis, which confirms the possibility of its application in the intelligent system of testing electric motors. The accuracy of this device depends on the graduation and properties of the dynamometric spring, which acts as a linear converter of the twist angle into torque.

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Algorithm of Mid-air Risk Estimation in Air Transportation System

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***Abstract*— Mid-air collision is one of the most dangerous aviation accidents. Progress in the air transportation system stimulates a continuously growing amount of airplanes simultaneously placed in the airspace which meets perils of airspace capacity. Operation of air transportation system under maximum load coincides with increasing overall safety risk. Therefore, dynamic risk analysis of air traffic is an important element of the safety insuring system. In the paper, we propose an algorithm of mid-air risk estimation for defined airspace volume and proposed software for experimental validation of safety analysis for defined airspace. Proposed algorithm is grounded on a risk estimation model for a pair of airplanes moved on crossing tracks. Also, we propose a model for estimation of airplane deviations from cleared flight routes based on analysis of air traffic data. Verification of proposed models has been done with surveillance data obtained by Automatic Dependent Surveillance-Broadcast communication link for particular airspace.**

Keywords*— Air Navigation, Aviation, Airspace, risk, safety, software, Air Traffic Management

I. Introduction

Despite geopolitical tensions, aviation as an industry has grown at an unprecedented rate over the past century. Civil aviation has made significant progress in ensuring flight safety. For example, 2023 was a record year in terms of safety: out of 37.7 million flights, only one accident was recorded, with 72 fatalities and 30 non-fatal incidents [1]. This achievement is a result of introduction of specific safety monitoring systems used on-board and ground-based.

Airborne collision and avoidance system (ACAS) is used as a primary safety monitoring system to identify risky pairs of airspace users and provide coordinated conflict resolution [2], [3]. Nowadays, ACAS II is mandatory for most airspace users and is deployed globally. Unfortunately, ACAS II provides conflict resolution by manner in the vertical plane only, due to limitations in the precision of angle measuring equipment. ACAS X technology is rapidly developing today and will provide conflict resolution in both vertical and horizontal planes. ACAS X grounds on shared data between airspace users and significantly improved performance of Global Navigation Satellite System (GNSS) [4], [5].

Many ground-based augmentation systems of GNSS have been deployed in the continental region to significantly increase positioning performance of GNSS[6]. But increased impact of unintentional jamming into navigation signals of GNSS during the last decade significantly affects the safety of civil aviation in wide areas [7][8]. War in Ukraine and military activity in the countries of the Middle East region segregate huge airspace volumes and caused a source of high power interference of GNSS spectrum in neighboring countries. Voluantarability of GNSS to interference caused pilots using inertial navigation systems or classical navigation methods grounded on wide usage of navigational aids [9][10].

Networks of navigation aids have been significantly improved during the last two decades to provide better positioning performance in case of GNSS faults [11][12].

The safety of aviation consists of multiple components one of the most important is the risk of mid-air collision. Risk of a mid-air collision is the probability of two airplanes occurring within the same safety volume. Most safety analyses are grounded on prediction of airplane trajectory based on some model to estimate the potential of collision [13][14]. But the most challenging questions include estimation of prior information of each airspace user to deviate from cleared trajectory and analysis influence of on-board positioning sensor errors into uncertainty area and risk of mid-air collision [15], [16].

Significant influence into risk provides airspace. Airplane operation in congested airspace always gives higher risk values due to the high density of air traffic [17], [18]. As an example long-term closed airspaces due to the war in Ukraine caused relocation of all civil traffic to neighboring countries. As a result, Turkish airspace accumulated the most transit flights between European and Asian airports [19][20]. In addition, military action in countries of the Middle East region caused wide interference influence into GNSS equipment in Turkish airspace and increased risk of attack due to errors in the air defense system.

In this paper, we analyzed the mid-air collision risk model for one of the most loaded and risky to be jammed airspace of Turkish. Also, we developed an algorithm for automatic air traffic control support system with an indication a risk values. Univariate generalized error distribution has been used in the risk estimation model to identify prior information of airplane deviation from cleared flight trajectory based on data obtained in Automatic Dependent Surveillance-Broadcast (ADS-B) technology.

II. Safety Boundaries in Air Traffic

The main cause of mid-air collision is loss of separation between airspace users in air traffic flow. Separation is a service provided by an air traffic controller. Separation can be vertical and horizontal, and it can be based on either time or range. The minimum vertical separation is 300 m for aircraft flying up to 290FL, and then 600 m between aircraft if area navigation (RNAV) is not applied [21]. The minimum horizontal separation based on range is 10 NM, 5 NM, or 3 NM. Minimum of horizontal separation depends on performance of radio navigation services provided in airspace. Smaller values of separation is better because increase capacity of airspace.

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Initially, air traffic was organized based on the conventional navigation method, in which aircraft had to move from one radio beacon to another to ensure its stay on a certain flight path. Then, with the development of technology, performance-based navigation (PBN) was implemented. PBN allowed use of any possible flight path without the need to move through radio beacons [22]. PBN has led to a change in the concept of airspace use to broad navigation by coordinates (latitude and longitude). However, due to a number of factors, it was decided to improve the area navigation method to a characteristic navigation method so that the main navigation characteristics are expressed accurately, integrity, continuity, and functionality [23]. The most common navigation characteristics includes:

- RNAV 10 for ocean routes, route width 50 nautical miles;
- RNAV 5 for route sections, route width 5 nautical miles;
- RNP 4 for ocean routes, route width 23 nautical miles;
- RNP 1 for approach/takeoff phase, route width 1 nautical mile.

Also, there are different types of flight routes. Flight routes are clearly defined segments of airspace used to direct the flow of airplanes.

III. Model of Mid-Air Collision Risk Estimation

Risk of mid-air collision could be estimated by two basic approaches:

- Risk is the frequency of occurrence of dangerous events in aviation, which is evaluated by statistical data in safety analysis.
- Risk is the probability of a dangerous event happened. In this case, probability is estimated based on the model of airplane movement and prior information of airplane deviation from cleared trajectory.

In this study, we use an analytical approach to mid-air collision estimation in airspace based on a simplified model of mid-air collision for a pair of airplanes on the crossing tracks. This model ignores the uncertainty of the velocities of both the aircraft involved and the distances between them [24]:

$$\mathrm{R} = m\int_{-\infty}^{\infty} P_h P_v \left(\frac{V_H(t)}{2\lambda_z} + \frac{2V(t)}{\pi\lambda_{xy}}\right) dt, \quad (1)$$

where m is number of airplane pairs; P_h is a probability of horizontal collision; P_v is probability of airplane getting vertical deviation; $V(t)$ is relative speed between two airplanes in a pair; $V_H(t)$ is a mean relative vertical speed λ_{xy} is average horizontal airplane size; λ_z is vertical safety size of an airplane.

The probability of horizontal overlap (P_h) could be represented as the probability of the aircraft going beyond the established flight route. Airplane deviation from the center of flight route could be calculated based on available surveillance datasets for particular airspace user. Deviation is calculated as a perpendicular from a point belonging to the plane's curved trajectory to the straight route:

$$ax + by + c = 0, \quad (2)$$

where x, y are coordinates of the point; a, b, c are line coefficients.

Line coefficients are calculated by two points on the line:

$$\frac{x-x_1}{x_2-x_1} = \frac{y-y_1}{y_2-y_1}, \quad (3)$$

$$(y_2 - y_1)x - (x_2 - x_1)y + x_1y_1 - x_2y_2 = 0, \quad (4)$$

where x_1, y_1 are the coordinates of the first point on the route segment; x_2, y_2 are the coordinates of the segment end.

Solving equation (4) gives values a, b, c . These coeficients are used to get a perpendicular from the straight route, at the point of its intersection with the curve of the plane's trajectory, from the obtained point to the base of the perpendicular. Length of perpendicular is an airplane deviation from the centerline of a cleared flight route:

$$d = \frac{ax_i + by_i + c}{\sqrt{a^2+b^2}}, \quad (5)$$

Based on available surveillance data of air traffic in particular airspace deviations from flight route network could be calculated by (5). Obtained deviations are used in statistical analysis to get the probability of airplane deviation in a lateral plane.

Standard univariate generalized error distribution is used for statistical analysis of airplane deviations [25], [26]. Also, univariate generalized error distribution is used as a probability density function to get prior information on previous airplane deviation from cleared flight route:

$$\rho(d) = \frac{1}{2^{k+1}\sigma\Gamma(k+1)} e^{-\frac{1}{2}\left|\frac{d-\mu}{\sigma}\right|^{\frac{1}{k}}}, \quad (6)$$

where σ is root mean standard deviation; k is parameter of control the skewness; μ is the mean value; $\Gamma(k+1)$ is an Eelier-Gamma function.

The shape of probability density function (6) is precisely tuned from normal to double exponential with parameter k, which best fits the data [27].

Probability of an airplane going out of horizontal perils of the flight route is estimated as follows:

$$P_h = \int_{-\infty}^{-lat} \rho(x)dx + \int_{\infty}^{lat} \rho(x)dx, \quad (7)$$

where lat is a separation minimum for latitudinal deviation.

Probability of airplane vertical deviation could be estimated as follows [11]:

$$P_z = \int_{-\lambda_z}^{\lambda_z} \int_{-\infty}^{+\infty} L(z_1)L(h_z + z_1 - z)dz_1dz, \quad (8)$$

where $L(x)$ is Laplace density function.

The scale parameter of this distribution is 38 ft for altitudes below 29000 ft (or 41000 ft if the airplane is within airspace where reduced vertical separation minimums are applied) and 76 ft above.

IV. Algorithm of Mid-Air Collision Risk Estimation

A specific software has been developed to analyze the risk of mid-air collision for particular airspace in a Python environment. Software uses API connection to open air traffic surveillance data from OpenSky network [28]. API library sends TXT requests to a remote server to collect air traffic data from ADS-B. Received responses from remote server is an array of data on airspace users within specified airspace boundaries. Data includes the following information about each airplane: unique identification code; callsign; origin country; time of last contact; longitude; latitude; geo- and baro- altitude; velocity; true track angle; vertical rate; position source and navigation accuracy category.

This array holds data about aircraft position at the time of sending the request, so API request is executed in the loop during the studied period of time.

We use geographic boundary of the flight information region of a particular country to analyze the risk of a mid-air collision. Structure scheme of developed software is given in Fig.1.

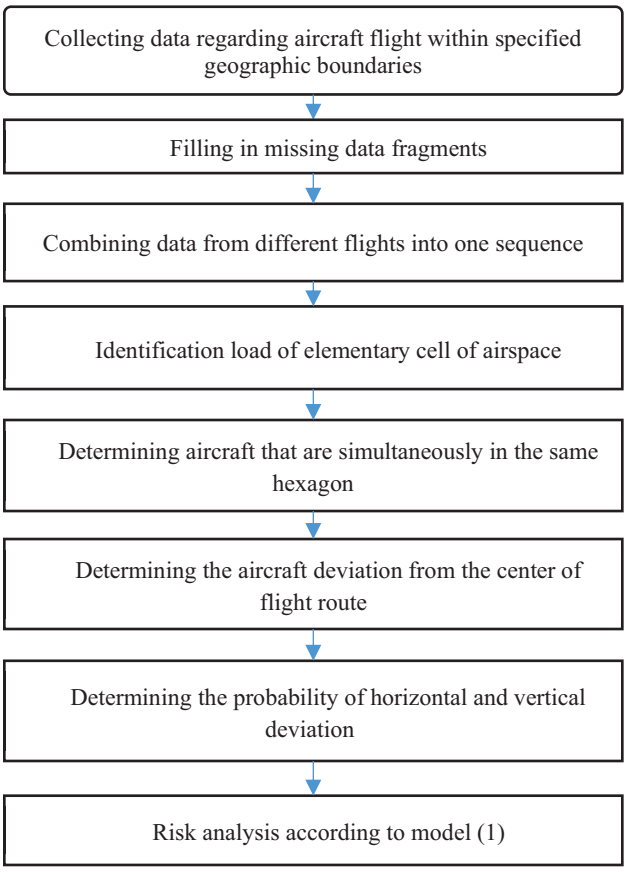


Fig. 1. Structure scheme of software for risk analysis.

ADS-B data is not complete and may contain multiple gaps that are required to be recovered by data processing approach. We use a linear regression model with a spline function to recover missing data in gaps.

In risk analysis, we use airspace representation as a set of elementary cells of hexagonal shape [29], [30]. Airspace partitioning with a set of elementary cells provides computation easy detection of closed airplanes to form pairs for risk analysis. Finally, risk of a mid-air collision is estimated by (1) and visualized in the graphical environment for each elementary cell by color gradation.

V. Numerical Demonstration

For validation and testing of the developed model of risk estimation, we use geometrical boundaries of flight information regions of Turkey LTAA and LTBB. The study has been evaluated in the period of February 16-25, 2025. Air traffic data are collected with developed software from OpenSky network. Two request functions *get_states()* and *get_track_by_aircraft()* are used to get curent location of airspace users and historical trajectory of each airplane. During the study period, 455 aircraft were recorded, performing 2961 flights. The total duration of all flights was over 205 hours. An example of flight data portion is given in Table I. Technical characteristics of each airplane (mainly width and height) are used from BADA model of EUROCONTROL.

TABLE I. A portion of Flight Data

Unique airplane code	Flight number	Departure airport	Arrival airport	Airplane type
4bb074	THY266	LTAI	EETN	b38m
4bb283	THY581	OETF	LTFM	a21n
4baa49	THY5060	LTFM	OEJN	a321
4bb206	THY1TU	LTFM	LTAJ	a21n
4bb205	THY137	OEMA	LTFM	a21n
4bc8cb	PGT4LF	LTFG	LTFJ	a21n
4baa10	THY4VQ	LTCO	LTFM	a320
4baacf	THY45G	LBWN	LTFM	b738
4ba8cd	TKJ5MQ	LTFJ	LTAI	b738
4bb1e3	THY108	LTFM	OEMA	a333
4baacf	THY4XH	LTFM	LTCA	b738
4ba9fa	THY6135	UCFM	LTFM	a332
4bc889	PGT569	HESH	LTFJ	a21n
4ba8f3	TKJ5MQ	LTFJ	LTAI	b738
4ba945	THY89	ZBAA	LTFM	b77w
4baac7	THY4JH	LTFM	LTAU	b738
4bb866	PGT22FL	LTCI	LTFJ	a20n

Based on received data sample, deviation of airplanes from the center of flight routes are calculated by (5). Statistical analysis of each airspace user is performed by previous user action in considered airspace. As an example of fitting univariate generalized error distribution for historical data set of airspace user “8965b5of” for flight “FDB1835” is given in Fig. 2.

Results of estimation in probability of horizontal deviation from safety perils by (7) gives an average value of 5.4×10^{-2} for considered airspace.

A load of elementary cells for considered airspace is given in Fig. 3.

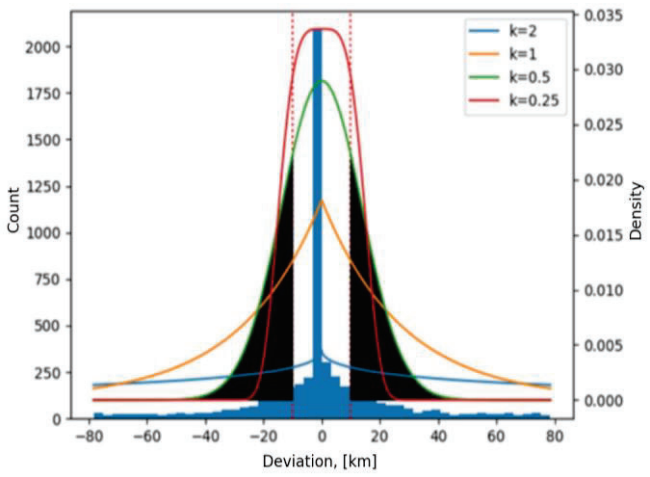


Fig. 2. Results of statistical analysis of airplane deviations from flight route center-line for "8965b5" user.

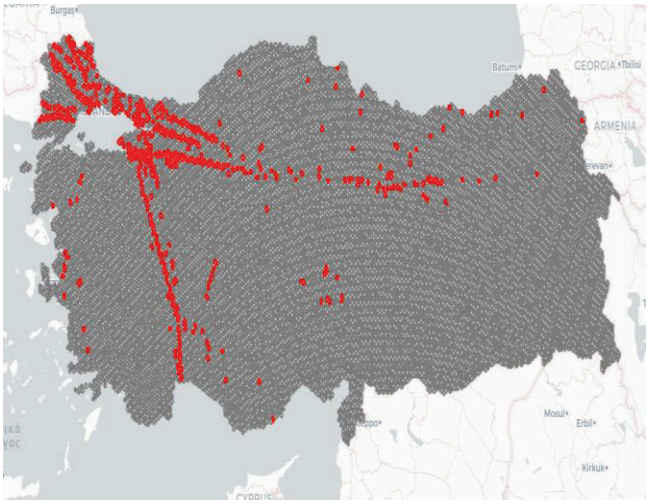


Fig. 3. Load of elementary cells for flight information regions of LTAA and LTBB.

A hexagonal cell with 10 km size is used in the safety analysis. Results of analysis give 815 unique cells with more than two airplanes simultaneously in one elementary cell. It involves 1582 airplanes, with a total time of 10:20:48. Number of aircraft does not coincide with the case where two aircraft are placed in one cell due to each airplane could have another potential conflict with another airspace user. We estimate risk of horizontal and vertical overlap by (7) and (8). Finally, estimate the overall risk of mid-air collision by (1).

Results of mid-air collision risk estimation for pairs of aircraft are given in Table II.

Most obtained results give risk values at a level of 10^{-10} which is under the target safety level specified by the International Civil Aviation Organization at a level of 5×10^{-9}.

The lowest risk value was identified for hexagon 862d19a37ffffff at 2025-02-18 07:54:45 between aircraft 4bb02b on flight TKJ7TC and 4bb074 on flight THY266 with a value of 4.8×10^{-17}. The highest value was in 861ec904fffffff at 2025-02-18 17:02:10 between aircraft 4baa10 (THY4VQ) and 4bb866 (PGT22FL), with a value of 2.059×10^{-10}.

TABLE II. RESULTS OF MID-AIR COLLISION RISK ESTIMATION FOR SELECTED PAIRS

Unique address of hexagonal cell	User A identification code	User B identification code	Time, UTC	Risk of mid-air collision
862d19a37ffffff	4bb02b - TKJ7TC	4bb074-THY266	02-18 07:54:45	4.8×10^{-17}
861ec900fffffff	4bb283-THY581	4ba9fa-THY6135	02-21 09:29:53	2.0×10^{-10}
863f6aa97ffffff	4baa49-THY5060	4bc889-PGT569	02-20 06:02:07	2.0×10^{-10}
861ec905fffffff	4bb206-THY1TU	4ba8f3-TKJ5MQ	02-22 20:38:18	2.045×10^{-10}
861ec91b7ffffff	4bb205-THY137	4ba945-THY89	02-22 04:41:12	2.047×10^{-10}
861ec905fffffff	4bc8cb-PGT4LF	4baac7-THY4JH	02-18 09:10:30	2.056×10^{-10}
861ec904fffffff	4baa10-THY4VQ	4bb866-PGT22FL	02-18 17:02:10	2.059×10^{-10}
861ecd327ffffff	4baacf-THY45G	4bb1e2-THY82F	02-17 22:17:30	2.059×10^{-10}
863f68c9fffffff	151de6-SDM5555	4ba9d3-THY73A	02-23 09:37:28	2.06×10^{-10}
862d34b2fffffff	4ba8cd-TKJ5MQ	4bc8c3-PGT1JA	02-19 20:28:47	2.95×10^{-10}
863f6859fffffff	4bb1e3-THY108	4bb166-THY5004	02-20 09:05:50	2.9×10^{-10}
862d34b2fffffff	4ba8ce-TKJ176	4bc8c4-PGT319	02-21 04:42:59	3.3×10^{-10}
862d35d0fffffff	4ba8ce-TKJ9YD	4baacf-THY4XH	02-24 18:05:10	3.36×10^{-10}

VI. CONCLUSIONS

To ensure air traffic safety, it is necessary to be able to assess and control the risks related to air traffic in particular airspace. Modern technologies give a huge opportunity to analyze real-time trajectory data. In addition, various ADS-B databases provide easy access to this data via the API interface.

Developed software structure for real-time data processing in air transportation system made it posible to evaluate safety levels in airspaces of particular flight information regions based on data from ADS-B. Hexagonal configuration of elementary cell helps to clearly localize pairs of airplanes with high potential for mid-air collision occurrence. Results of mid-air collision risk estimation provide a clear mechanism of threat identification in civil aviation.

Results of safety analysis show that due to the high intensity of air traffic over Turkey, many potential conflict cells were identified (815). Also, many aircraft have a significant probability of horizontal deviation. However, according to the results of the risk calculation, no conflict received a risk greater than 3.3656×10^{-10}, which means that the probability of collision in all conflict situations was less than the established risk value for air traffic service provision.

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Improved Feature Tokenizer Transformer

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This paper proposes an improved FT-Transformer model to improve the efficiency of intrusion detection in computer networks. Key modifications include: integrating the ColumnEmbeddingAdder layer to take into account the context of each column, using a combination of GlobalAveragePooling1D and GlobalMaxPooling1D for comprehensive feature extraction after the transformer block, and using SpatialDropout1D for better regularization and generalization ability, especially for correlated features. To solve the problem of class imbalance, the model minimizes the loss function, and also uses the SMOTE method to increase attention to minority classes. A comparative analysis is performed with models based on the Tab Transformer, FT-Transformer, and SAINT architectures. Experimental results show that the proposed model achieves a higher accuracy of 0.9364 and an AUC of 0.9943 among the compared models. The model also has a significantly lower loss of 0.0158, which is up to 17 times less than other models and is more efficient when working with imbalanced data.

Keywords—Machine Learning, Transformer Models, Attention Mechanism, SMOTE, TabTransformer, FT-Transformer, SAINT, Self-Attention and Intersample Attention

I. Introduction

Ensuring cybersecurity of computer systems and networks in the context of the rapid development of information technologies and the growth in the number and complexity of cyber threats is becoming critical. Network traffic is characterized by constant attempts to unauthorized penetration, theft, modification and destruction of valuable information. This causes significant financial, reputational and even strategic losses to organizations and individuals around the world. Network traffic is an extremely complex system containing many characteristics that interact with each other. These features can include not only network parameters, but also parameters of devices, programs, data. Depending on the specific task, for example, for security analysis, other features can be added, such as the type of attack, its source or target. This leads to the fact that the number of features can reach several thousand, creating a problem of computational complexity. Many traditional machine learning methods become slow and resource-intensive when working with such large dimensions of the feature space. Additionally, network traffic is characterized by an unbalanced number of malicious attacks compared to normal traffic. This can lead to machine learning models that underlie intrusion detection systems (IDS) training mostly on the dominant class (normal traffic) and poorly predicting the target variable for the smaller class (attacks). This situation often leads to overtraining, where the model "remembers" the training data with the dominant class too well but generalizes poorly to new, more evenly represented data. Despite the active implementation of various protections, attackers are constantly improving their methods, finding new vulnerabilities and developing more sophisticated attacks that often go unnoticed by traditional security systems.

The purpose of the paper is to enhance the efficiency of intrusion detection model based on transformer models.

II. Literature review

In many modern security systems, hybrid approaches are employed, combining various methods to achieve a higher level of protection. Hybrid approaches are based on the use of machine learning methods such as random forest, logistic regression, nature-inspired algorithms, fuzzy approaches, ensemble models based on bagging, boosting and stacking, neural network and others.

Random Forest [1] significantly reduces the risk of overfitting, which is a common problem with individual decision trees. Each tree in the forest is trained on a bootstrapped sample of the data, and only a random subset of features is considered at each split, adding further decorrelation. It tends to generalize well to unseen data, making it a strong choice for many real-world problems. The training of individual trees in the forest is independent, allowing for parallel processing, which can speed up the training time on multi-core processors or distributed systems. But training many individual decision trees can be computationally expensive and time-consuming, especially with very large datasets or a large number of trees in the forest.

The model's coefficients of logistic regression [2] are straightforward to interpret. They represent the change in the log-odds of the outcome for a one-unit increase in the corresponding feature, assuming other features are held constant. This makes it a "white-box" model. This model also computationally efficient to train and make predictions, even on large datasets, requires less memory compared to more complex models. The most significant limitation of the model is its assumption of a linear relationship between the independent variables and the log-odds of the dependent variable. If the true relationship is non-linear, logistic regression may perform poorly unless non-linear transformations are applied to the features beforehand.

Many nature-inspired algorithms [3] are iterative and population-based, meaning they require many evaluations of the objective function. This can be computationally very expensive, especially for complex problems where a single evaluation takes a long time.

Fuzzy logic excels in domains where information is inherently vague, ambiguous, or incomplete, allowing systems to make decisions based on "degrees of truth" rather than strict binary (true/false) logic [4]. This mirrors human intuition more closely. But the process of defining

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membership functions (e.g., the exact shape, width, and overlap) and formulating fuzzy rules often relies heavily on expert knowledge, trial and error, or heuristic methods. This can be subjective, time-consuming, and may not always yield optimal results. There's no standardized way to derive them.

Bagging, Boosting, and Stacking are three prominent categories of ensemble methods. Bagging works [5] by training multiple independent base learners on different bootstrap samples of the original training data. Primarily reduces variance and overfitting by averaging independent models. Good for complex models that overfit easily. Boosting [6] trains multiple base learners sequentially. Each subsequent model is built to correct the errors made by the previous models. It focuses on the samples that were misclassified or had higher residuals in the previous iterations, giving them more weight. Primarily reduces bias by sequentially correcting errors. Achieves high accuracy but is more sensitive to outliers and overfitting if not carefully tuned. Stacking [7] involves training multiple diverse base models (Level 0 models) on the original training data. Then, a meta-model (Level 1 model) is trained on the predictions (or probabilities) generated by these Level 0 models. The Level 0 models' predictions become the input features for the Level 1 meta-model. Cross-validation is often used to generate out-of-sample predictions from Level 0 models for the Level 1 model to prevent information leakage. Aims for the highest accuracy by combining diverse model predictions with a meta-learner. It's the most powerful but also the most complex and computationally demanding.

Neural Networks [8] offer unparalleled power in modeling complex data and achieving high accuracy, especially in areas like computer vision and natural language processing. However, their "black box" nature, significant computational demands, and large data requirements are important considerations.

One of the effective hybrid approaches involves the use of transformer models [9]. Their ability to process sequences of data and establish long-range dependencies makes them particularly attractive for spam filtering and spammer detection [10][11], language modeling [12], Computer Vision [13] and the analysis of network traffic and system logs [14]. Furthermore, the presence of an Attention Mechanism [15] allows the model to weigh the importance of different parts of the input sequence when making decisions. This helps in detecting key patterns and anomalies that may be scattered throughout the sequence. For example, in a multi-stage attack, the attention mechanism can help link events that are distant in time but indicate an intrusion. This is critically important for detecting complex attacks that may unfold over a long period. Additionally, transformers can process the entire input sequence in parallel, significantly accelerating training, especially when working with large volumes of data. There also exist pre-trained transformer models (e.g., for natural language processing) that can be fine-tuned for intrusion detection tasks, which can improve performance and reduce the need for large amounts of training data.

Although transformer models have shown significant success in processing sequential data (text, audio, time series, images), their direct application to tabular data faces a number of specific challenges. Unlike text or time series, where there is a clear order of elements, rows and columns in tabular data usually do not have an inherent sequence. The order of columns can be arbitrary and not affect the semantics of the data. In contrast, transformers are designed to process sequences and use a positional encoding mechanism to establish the order of elements. Therefore, in the absence of interrelationships, it is difficult to use this mechanism effectively. Key information in tabular data often lies in complex interactions between various features. While the attention mechanism in transformers can potentially model interactions between features, its effectiveness may also depend on the method of feature representation and the model's architecture. Tabular data can contain different types of features: numerical, categorical, boolean, etc., and have different scales and units of measurement. This requires effective ways to convert different feature types into compatible embeddings while preserving their informativeness. Furthermore, categorical features can have a large number of unique values. In this case, direct conversion into a vector representation of each category can lead to a large embedding matrix dimensionality and increase the number of model parameters. Since there is a lack of column permutation invariance in tabular data, the model may interpret the same table with permuted columns differently, which is undesirable.

Examples of successful architectures specifically designed for tabular data using transformer ideas include models such as TabTransformer and FT-Transformer (Feature Tokenizer + Transformer).

III. Approaches and Methods

A. TabTransformer

TabTransformer is a deep learning model specifically designed for processing tabular data, combining the power of transformers with the ability to process heterogeneous features (categorical and numeric). Proposed by Huang et al. in 2020 [16] TabTransformer uses the attention mechanism to model the relationships between categorical features, making it particularly effective for classification and regression tasks on tabular data [17], [18].

TabTransformer (Fig.1) consists of two main components: a categorical feature processing unit based on a transformer and a numeric feature processing unit via a multilayer perceptron (MLP).

Let us consider the model architecture using the UNSW-NB 15 dataset as an example, which contains information about the normal functioning of the network and during synthetic intrusions. Each categorical feature (e.g., proto, service) is encoded using an embedding layer, where each category is mapped to a vector space of a given dimension (e.g., 16). Next, the embeddings of all categorical features are fed into the transformer model. The output is contextualized embeddings of categorical features that reflect their relationships.

Numerical features of the dataset, such as “dur” and “spkts”, are normalized (typically using StandardScaler) and fed into a separate MLP block. The MLP consists of several fully connected layers with non-linear activations (e.g., ReLU), which transform the numerical features into a space compatible with the embeddings of categorical features. Subsequently, the contextualized embeddings of categorical features (from the transformer) and the processed numerical features (from the MLP) are concatenated. The combined features are then fed into a final MLP block, which performs the classification task.

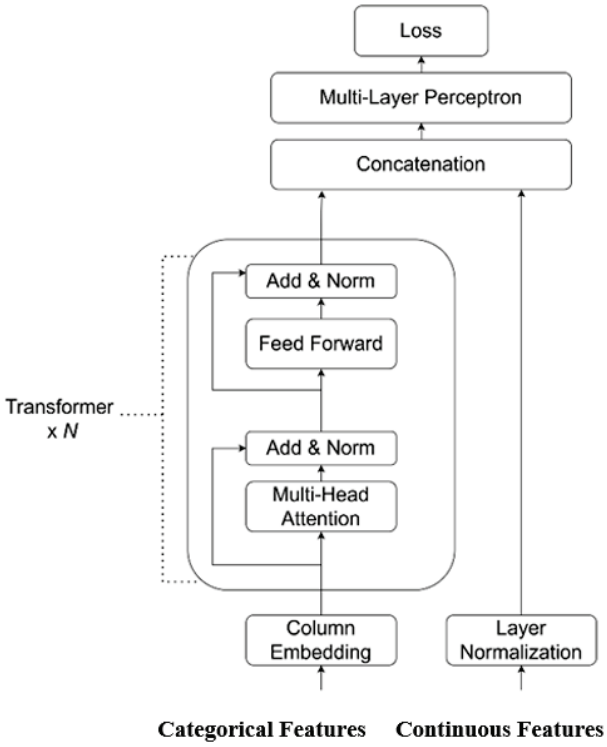


Fig. 1. The architecture of TabTransformer

B. Advantages and Disadvantages of TabTransformer

The self-attention mechanism allows for effectively capturing complex interrelationships between categorical features, which is particularly useful for datasets with a large number of categorical features. TabTransformer can handle heterogeneous tabular data (categorical + numerical features) without the need for complex pre-processing or manual feature engineering. In classification tasks on tabular data, TabTransformer often outperforms traditional methods (e.g., XGBoost) and competes with other deep learning models (SAINT, FT-Transformer), especially when categorical features play a key role.

A limitation of TabTransformer is that numerical features are processed through an MLP, which is less effective compared to models that apply transformers to all features (e.g., FT-Transformer). This can reduce performance on datasets with a large number of numerical features, such as UNSW-NB15 (~40 numerical features).

C. Feature Tokenizer Transformer

FT-Transformer (Feature Tokenizer + Transformer) is a deep learning model for tabular data. It extends the ideas of TabTransformer by applying the transformer architecture to all features (both categorical and numeric), making it particularly effective for heterogeneous tabular data. FT-Transformer uses the concept of "feature tokenizer" that transforms each feature into a unified embedding, allowing the transformer to model the relationships between all features using a self-attention mechanism. This model demonstrates competitive results compared to traditional methods (e.g. gradient boosting) and specialized neural networks for tabular data such as TabTransformer and SAINT.

FT-Transformer (Fig.2) consists of two key components: a feature tokenizer block and a stack of transformer blocks [19].

Each feature (categorical or numerical) is transformed into a fixed-dimension vector representation (e.g., 16 or 32). Categorical features (e.g., “proto”, “service” in UNSW-NB15) are encoded via an embedding layer, where each category is mapped to a vector, similar to TabTransformer. Numerical features (e.g., dur, spkts) are normalized (typically using StandardScaler) and fed into a linear layer, which converts the scalar value into a vector of the same dimension as the categorical embeddings.

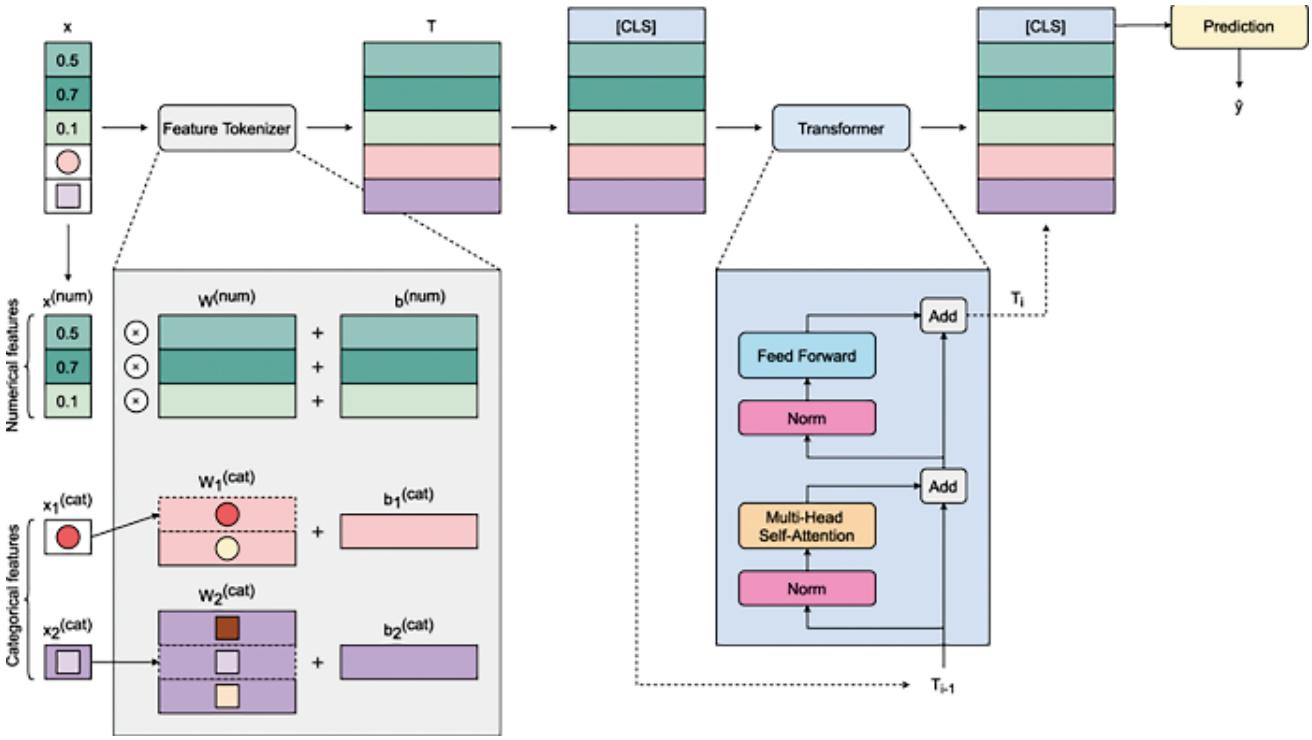


Fig. 2. The architecture of Feature Tokenizer Transformer

All embeddings are combined into a sequence of "feature tokens," where each token represents a single feature (analogous to words in natural language processing)..

A [CLS]-token is added to the sequence, which is used for information aggregation

The sequence of feature tokens is then fed into a stack of transformer blocks. Unlike TabTransformer, where transformers are applied only to categorical features, FT-Transformer processes all features, which enhances its ability to model heterogeneous data. The processed features are then fed into a multi-layer perceptron (MLP) which performs the classification.

D. Advantages and Disadvantages of FT-Transformer

Since FT-Transformer applies transformers to all features, it makes model more efficient. At the same time, the self-attention mechanism allows to model complex relationships, capturing nonlinear dependencies between all features. This is especially useful for detecting attack patterns (for example, the combination of “proto”, “dur”,”spkts” for Fuzzers). FT-Transformer has high performance, often outperforming TabTransformer and competing with SAINT in tabular data classification tasks, achieving high accuracy.

A limitation of FT-Transformer is that it has quadratic complexity with respect to the number of features, which increases training time compared to lighter models (e.g. XGBoost or MLP). For UNSW-NB15 (~43 features), this is not a serious problem, but for datasets with hundreds of features, model optimization is required. In addition, the model is hyperparameter-sensitive. Its performance quality depends on the correct choice of hyperparameters (dimensionality of embeddings, number of transformer blocks, learning rate). This may require additional experiments for input data.

E. Self-Attention and Intersample Attention Transformer

SAINT (Self-Attention and Intersample Attention Transformer) [20] – is a transformer architecture designed for tabular data processing that combines self-attention and intersample attention mechanisms to efficiently model complex relationships between features and samples. SAINT demonstrates high performance on tabular data classification and regression tasks, outperforming traditional machine learning methods such as gradient boosting and competing with other transformer models such as FT-Transformer and TabTransformer.

The SAINT architecture consists of key components such as tokenization blocks and transformer blocks.

Categorical features are encoded through an embedding layer, where each category is mapped to a vector. Numeric features are transformed through a linear layer (Dense) to match the dimensionality of the categorical embeddings. Subsequently, each sample is supplemented with a special CLS token, which serves to aggregate information for classification.

SAINT uses a sequence of transformer blocks, each of which includes two types of attention mechanisms: Self-Attention and Intersample Attention (Fig.3).

The Self-Attention block models the relationships between features within a single sample, using MultiHeadAttention to capture dependencies between categorical and numeric features. The Intersample Attention block analyzes the relationships between different samples within a batch, allowing the model to detect global patterns in the data (e.g., similarities between samples of the Fuzzers and Exploits classes). After the transformer blocks, the [CLS] token vector is passed to a multilayer perceptron (MLP) with regularization (e.g., L2 or Dropout). Next, an output layer with softmax activation produces class probabilities.

F. Advantages and Disadvantages of SAINT

The model has high accuracy and is well suited for tabular data with different types of features and dimensions. The combination of self-attention and inter-sample attention allows SAINT to effectively classify complex malware classes, such as Fuzzers, which have similarities to Exploits. However, the presence of an inter-sample attention block increases the memory requirements and training time compared to FT-Transformer or TabTransformer, especially when using large batches. The model also requires hyperparameter tuning, namely: the number of blocks the number of attention heads, and training speed require experimental selection.

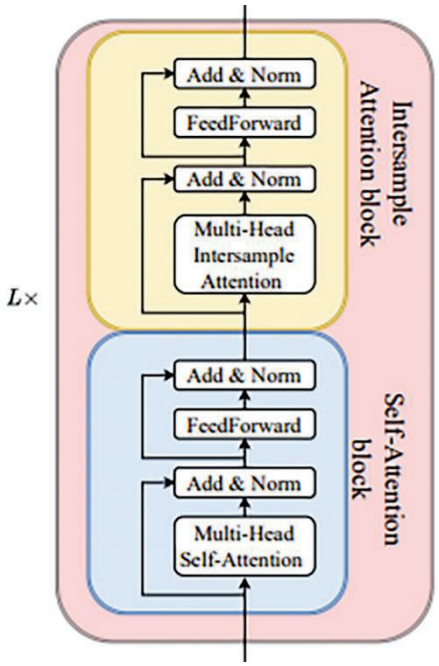


Fig. 3. The architecture of Self-Attention and Intersample Attention Transformer

G. Experimental research and efficiency evaluation

To eliminate the above shortcomings, the paper proposes a modification of the model, based on the FT-Transformer architecture and processes tabular data, in particular the UNSW-NB15 dataset. The proposed model incorporates components of the FT-Transformer, such as the Feature Tokenizer and transformer blocks, with corresponding modifications. The modification includes the addition of an extra ColumnEmbeddingAdder layer, the use of a combination of global average and maximum pooling (GlobalAveragePooling1D and GlobalMaxPooling1D), and spatial dropout (SpatialDropout1D).

The use of the ColumnEmbeddingAdder layer enables the addition of column-specific embeddings to the input data. This allows the model to consider the context of each column individually, which is crucial for tabular data with heterogeneous features (numerical and categorical). None of the known models (Tab Transformer, FT-Transformer, SAINT) employ this approach.

The combination of GlobalAveragePooling1D and GlobalMaxPooling1D after the transformer block allows the model to capture both average and maximum data characteristics. In contrast, Tab Transformer relies solely on Flatten, while SAINT and FT-Transformer utilize the [CLS] token without additional pooling.

To address the class imbalance issue in this model, focal loss function (with parameters gamma=2.0 and alpha=0.25) is used. This helps focus on challenging samples and reduces the influence of the dominant class (Normal). Other models (Tab Transformer, FT-Transformer, SAINT) typically use standard categorical cross entropy, which may be less effective for imbalanced data. Additionally, the SMOTE method is employed for class balancing, that further enhancing attention to minority classes (e.g., Fuzzers). SMOTE generates synthetic samples by interpolating between existing samples of the minority class. However, if the initial set of samples (e.g., 100 samples) is limited and does not cover all possible patterns of this class, the synthetic samples will also be limited. This means that new samples do not introduce fundamentally new information but only expand the space based on existing data. This can lead to poor generalization by the model on real test data, which may contain new or uncharacteristic patterns. Experimental results showed that for the balanced Fuzzers class in the UNSW-NB15 dataset, the model may "over-optimize" on patterns more frequently occurring in the data (e.g., patterns of the Normal class) if these patterns exhibit stronger correlations.

The Feature Tokenizer is implemented as follows. Categorical features are encoded using an Embedding layer with a dimensionality of EMBEDDING_DIM=32. Numerical features are processed through two Dense layers: the first with a dimensionality of EMBEDDING_DIM/2 and ReLU activation, and the second with a dimensionality of EMBEDDING_DIM without activation. Subsequently, column-specific embeddings are added to both types of embeddings (numerical and categorical) via the ColumnEmbeddingAdder, enabling the model to account for the context of each column. All embeddings are then concatenated and reshaped into a sequence of length “seq_len’.

Feature aggregation is performed next way. Instead of the [CLS] token, which acts as a "black box" and aggregates information via the attention mechanism, as in SAINT or FT-Transformer, the model employs a combination of global average pooling (GlobalAveragePooling1D) and global maximum pooling (GlobalMaxPooling1D). Average pooling reflects the overall trend in the data, while maximum pooling captures the extreme characteristics of features, which is much easier to interpret. Subsequently, these pooling outputs are combined through concatenation. Thus, unlike FT-Transformer and SAINT, which add a [CLS] token for information aggregation, the proposed model does not use this approach. This simplifies its architecture but increases computational costs in the transformer blocks. In contrast, the use of GlobalAveragePooling1D and GlobalMaxPooling1D involves calculating the average and maximum values along the feature axis, providing a fast and efficient method for aggregation.

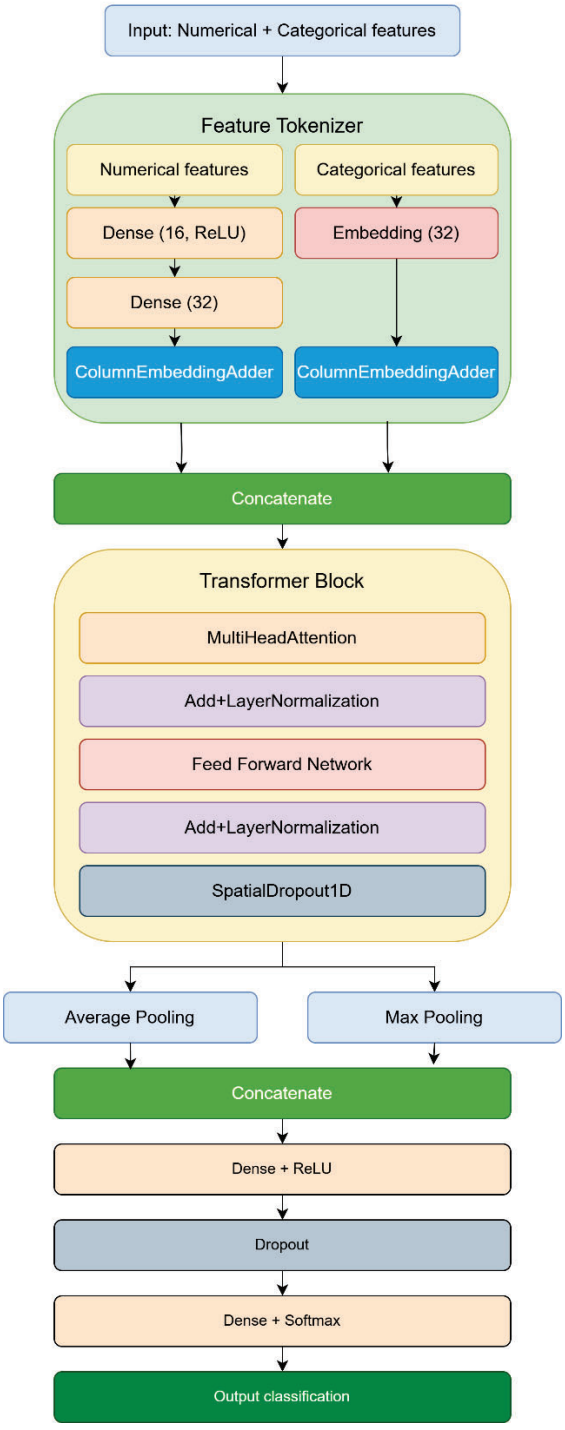


Fig. 4. The architecture of developed model

Subsequently, the combined features are fed into a multilayer perceptron (MLP) with two layers:

- First layer: Dense(128) with L2 regularization (λ=0.005), ReLU activation, and dropout (dropout_rate=0.1).

- Output layer: Dense(4) with softmax activation for predicting the probabilities of four classes (Normal, Exploits, Fuzzers, Generic).

The model includes one transformer block, which contains a multi-head self-attention block (MultiHeadAttention) with h=2 heads and a key dimensionality of demb/h, modeling dependencies between features within a single sample. Regular Dropout, which randomly nullifies elements in the tensor, is not used, as it may lead to the loss of spatial structure in the data. Instead, the model employs a specialized spatial dropout block (SpatialDropout1D, dropout_rate=0.1). This block randomly drops entire features and promotes better regularization compared to the standard Dropout block, especially for data with high feature correlation.

In the UNSW-NB15 dataset, many features (e.g., “spkts”, “dpkts”, “sbytes”, “dbytes”) are correlated, as they reflect related aspects of network connections (e.g., traffic volume or packet count). Such correlation can lead to overfitting, where the model overly relies on these related features, ignoring less correlated but potentially important ones.

Since SpatialDropout1D nullifies entire embedding channels, it forces the model not to rely on specific embedding dimensions that may encode correlated features. For example, if the features “spkts” and “dpkts” contain similar information in one of the embedding dimensions, nullifying this channel compels the model to seek other, less correlated patterns in the data, thereby improving its generalization ability.

A feed-forward neural network (FFN) was used as the classification model, consisting of two layers: the first with a dimensionality of dffn=64 and GELU activation, and the second with a dimensionality of demb.

According to the results of the study, the following indicators of the quality of the model were obtained (Tables 1, 2).

TABLE I. AVERAGE METRICS OVER THREE RUNS

Model	*Table column subhead*	*Accuracy*	*Precision*	*AUC*
Developed model	**0.9364**	0.9446	0.9283	**0.9943**
Tab Transformer	0.9138	0.9204	0.9078	0.9898
FT-Transformer	0.9289	0.9343	0.9247	0.9927
SAINT	0.9311	0.9341	0.9287	0.9927

TABLE II. AVERAGE F1- SCORE METRICS OVER THREE RUNS

Model	*Exploits*	*Fuzzers*	*Generic*	*Normal*
Developed model	0.868	**0.731**	0.984	0.968
Tab Transformer	0.843	0.655	0.981	0.951
FT-Transformer	0.863	0.703	0.982	0.962
SAINT	0.868	0.714	0.983	0.964

As seen from Table 1, the proposed developed model achieves the highest accuracy (0.9364) and AUC (0.9943) among all models. This is 2.3% higher than Tab Transformer (0.9138) and 0.5–0.7% higher than FT-Transformer (0.9289) and SAINT (0.9311). Additionally, the developed model has the lowest loss (0.0158), which is 12–17 times lower than that of other models, indicating better optimization and the ability to avoid overfitting (Fig. 5). The model is more effective in the presence of imbalanced data.

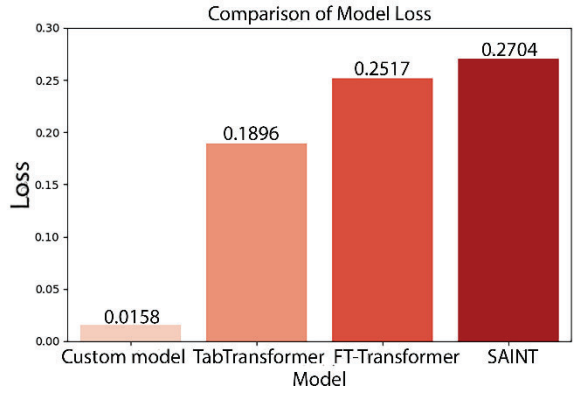


Fig.5. Comparision of Modef Loss

The developed model demonstrates the best F1-score for the Fuzzers class (0.731), which is the least represented class. This is 7.6% higher than Tab Transformer (0.655), 2.8% higher than FT-Transformer (0.703), and 1.7% higher than SAINT (0.714). Additionally, the model’s Standard Deviation of Accuracy is the lowest (Fig. 6).

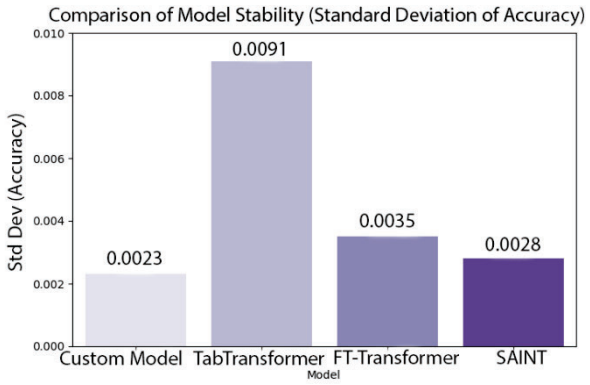


Fig. 6. Comparison of Standard Derivation of Accuracy

IV. CONCLUSIONS

Transformer models are a promising area for developing highly effective intrusion detection systems. Their ability to process sequences, establish dependencies, and leverage attention mechanisms provides significant advantages over traditional methods. However, there are challenges when such models process tabular data, as transformers are designed for sequence processing..

In this work, a modification of the transformer model based on the FT-Transformer architecture is proposed. The model modification includes: adding a ColumnEmbeddingAdder layer, using a combination of GlobalAveragePooling1D and GlobalMaxPooling1D, and

applying SpatialDropout1D. The ColumnEmbeddingAdder layer allows to add embeddings columns to the input data. This allows the model to handle the context of each column separately, which is especially important for tabular data with heterogeneous features (numeric and categorical). Next, after the transformer block, a combination of GlobalAveragePooling1D and GlobalMaxPooling1D is applied. Average pooling helps capture the overall trend in the data, while maximum pooling identifies the extreme characteristics of features, facilitating their interpretation. The results of these pooling operations are subsequently combined via concatenation.

To address the class imbalance problem, focal loss function is used in the model. This allows the model to focus on difficult samples and reduce the influence of the dominant class. Additionally, the SMOTE method is applied for class balancing, enhancing attention to minority classes.

The implementation of the Feature Tokenizer is as follows. Categorical features are encoded using an Embedding layer. Numerical features are processed through two Dense layers. The results are concatenated and passed to a multilayer perceptron (MLP) with two layers. The model contains a single transformer block, modeling dependencies between features within a single sample. In addition, the spatial dropout block (SpatialDropout1D) randomly drops entire features, promoting better regularization, especially for data with high feature correlation.

Experimental results showed that the proposed model achieves the highest accuracy (0.9364) and AUC (0.9943) among all models. This is 2.3% higher than Tab Transformer (0.9138) and 0.5–0.7% higher than FT-Transformer (0.9289) and SAINT (0.9311). One of the key advantages of the developed model is its extremely low loss (0.0158), which is 12–17 times lower than other models, indicating better optimization and an effective ability to avoid overfitting. The model also proved more effective when working with imbalanced data, which is a typical problem in intrusion detection. Specifically, it demonstrated the best F1-score for the "Fuzzers" class (0.731), the least represented class. This is 7.6% higher than Tab Transformer (0.655), 2.8% higher than FT-Transformer (0.703), and 1.7% higher than SAINT (0.714).

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Methodology for Selecting the Protection Strategy in IoT Environments Based on the Device Resource Profile

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***Abstract*—This paper proposes a methodology for adaptive selection of cryptographic protection levels based on the current resource profile of an IoT device. The resource profile is derived from normalized hardware and energy parameters, including processor frequency, available memory, network bandwidth, residual battery charge, and device temperature. These metrics are aggregated into a single scalar index that reflects the overall resource capacity of the device. A decision-making algorithm is introduced, incorporating time-averaging, hysteresis mechanisms, and a dynamic risk coefficient to enable real-time switching between protection levels (Light, Medium, Intensive). The methodology was implemented on a testbed using STM32F446RE and validated in the NS-3 simulation environment. Experimental results demonstrate that the proposed approach achieves a balance between cryptographic strength, energy efficiency, and performance under varying resource conditions. The solution is scalable and suitable for distributed IoT systems with stringent security demands.**

Keywords—IoT, resource profile, adaptive protection, lightweight cryptography, STM32, security strategy

I. Introduction

The rapid development of the Internet of Things (IoT) has led to an increase in the number of devices operating in distributed networks with limited computing resources. Such devices – sensors, microcontrollers, actuators – often have a low clock frequency, limited RAM and operate from autonomous power sources [1]. This creates serious challenges for implementing effective cryptographic protection, which must be sufficiently resistant to attacks and at the same time not exceed the permissible power consumption or latency of the system [2].

One of the key problems in protecting such devices is the need for flexible selection of cryptographic mechanisms in accordance with the current characteristics of the hardware platform [3, 4]. Using a universal approach that ignores the limitations of a specific device leads to either excessive resource load or insufficient security [5]. This issue is especially relevant for IoT systems operating in critical conditions – with unstable power supply, limited connectivity or frequent changes in device configurations in the network.

In such conditions, there is a need for adaptive protection that takes into account the real technical capabilities of the IoT node at a given point in time [6]. This requires the introduction of a formal description of the device state, which can be used as input data for automated selection of a protection strategy. In this study, the resource profile of the device is considered as a set of parameters that quantitatively characterize its hardware and energy capabilities – in particular, computing power, available memory, battery charge level, processor frequency, number of operations per unit of time and other relevant indicators. This approach allows formalizing decisions on the choice of protection and creating a basis for dynamic tuning of security mechanisms in real time.

The aim of the study is to develop a methodology for selecting a protection strategy based on the device resource profile in conditions of limited computing and energy capabilities of IoT systems. This approach will not only optimize resource use, but also ensure targeted preservation of data confidentiality and integrity in distributed IoT systems.

II. Theoretical Background and Related Works

The issue of ensuring information security in conditions of limited resources is actively studied in the context of building adaptive protection systems for IoT devices [7, 8]. Due to the rapid growth in the number of such devices, in particular in the field of medicine, transport, industry and household solutions, there is a need for lightweight cryptographic algorithms that can work under conditions of limited computing power and unstable energy supply [9].

One of the directions for solving this problem is the use of lightweight cryptographic primitives (lightweight cryptography), in particular encryption algorithms such as GIFT, ASCON, PRESENT, LEA, CHAM, etc [10, 11]. These algorithms were designed to ensure basic confidentiality and data integrity with minimal hardware costs. As part of the NIST Lightweight Cryptography competition, which lasted from 2019 to 2023, the effectiveness and security of a large

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number of such algorithms were analyzed, among which ASCON was chosen as the standard.

In addition to protection strategies for constrained IoT environments, recent works have also addressed resource-aware security adaptation in cloud-native and containerized systems. For example, in [12], a flexible framework was proposed for evaluating load balancing strategies in container orchestration platforms. Further research [13] introduced a methodology for optimizing application-level security through scheduler-aware decisions in Kubernetes environments. These approaches, although targeting different platforms, reinforce the relevance of dynamic and context-aware adaptation mechanisms for achieving a balance between performance and security – the challenge also faced in resource-limited IoT systems.

Despite the effectiveness of individual cryptographic algorithms, many works note the need to adapt the level of protection in accordance with the current characteristics of the device. In particular, in the works [14], [15] and [16], models for dynamic selection of cryptographic operations are proposed, taking into account the current energy level or class of the device. However, most approaches are focused on predefined resource categories or use device classification according to simplified features, which reduces the accuracy of adapting the protection strategy. A separate area of research is the construction of a device profile, which includes metrics reflecting its computational state. In particular, in the work [17], an approach to calculating the «energy budget» of a node is proposed to determine the maximum allowable volume of cryptographic operations. Other authors use the parameters of execution time, processor frequency, and memory utilization as criteria for choosing between simple and more complex protection algorithms [18, 19]. However, despite the existence of individual models, there is no unified methodology that would align the choice of a protection strategy with a formalized description of the device's resource intensity.

III. Building a Defense Strategy

The variability of energy and computational characteristics of embedded devices requires the use of adaptive approaches to choosing the level of protection. The strategy should take into account not only the level of data criticality, but also the current state of the device, in particular, available memory, speed, battery charge and other hardware limitations. That is why there is a need for a mathematical description of the current resource state, which would allow dynamically making decisions about changing the level of cryptographic load.

The proposed methodology involves step-by-step processing of hardware metrics of the device, their normalization, aggregation into a scalar index of resource intensity and determination of variations over time. Based on this index, a decision is made to activate the appropriate protective mode taking into account threshold values, hysteresis and, if necessary, correction taking into account the level of risk.

A. Formalization of the Device Resource Profile

In the context of adaptive protection for IoT nodes, it is essential to derive a unified numerical representation of their hardware and energy characteristics [20]. A formalized resource profile enables objective comparison between platforms, precise state assessment, and the definition of well-justified thresholds for invoking specific cryptographic mechanisms.

The resource profile of a device is modeled as a time-dependent, multidimensional curve (1).

$$R(t) = (r_1(t), r_2(t), \dots, r_n(t))^{\mathrm{T}} \in [0,1]^n \quad (1)$$

where $r_1(t), r_2(t), \dots, r_n(t)$ – normalized values of the selected hardware and energy-related metrics at time t; n – the total number of considered parameters forming the resource profile; T – transpose operator, representing the resource vector as a column; $[0,1]^n$ – the bounded domain indicating that each metric is scaled to the interval $[0,1]$.

In the proposed approach, the resource profile of the device is initially formalized through a set of direct hardware and energy indicators:

$$x_1(t) = f_{CPU}(t), x_2(t) = RAM_{avail}(t), \dots, x_n(t) = T_{dev}(t)$$

where each primary metric $x_i(t)$ – clock frequency $f_{CPU}(t)$, available memory $RAM_{avail}(t)$, instruction throughput, network bandwidth, remaining energy, cryptographic power draw, device temperature $T_{dev}(t)$ is normalized according to (2).

$$r_i(t) = \frac{x_i(t) - x_i^{min}}{x_i^{max} - x_i^{min}}, i = 1, \dots, n \quad (2)$$

where $r_i(t)$ – normalized value of the i-th metric at time t, $x_i(t)$ – current value of the i-th metric, x_i^{min} and x_i^{max} the lower and upper bounds for that metric across a class of comparable devices.

To aggregate the vector $R(t)$ into a scalar resource-capacity index $\Phi(t)$, the generalized $L_p - norm$ is employed (3).

$$\Phi(t) = \left(\sum_{i=1}^{n} \alpha_i [r_i(t)]^p\right)^{1/p},\ p \geq 1,\ \sum_{i=1}^{n} \alpha_i = 1 \quad (3)$$

where α_i are weighting coefficients reflecting the relative importance of each metric, and the exponent p controls the sensitivity of the aggregation (e.g., as $p \to \infty$, Φ approaches the maximum component of R).

Beyond the instantaneous measure $\Phi(t)$, two integral characteristics – the time-averaged resource index (4) and its variance (5) over the interval $[t_0, t_0 + \mathrm{T}]$.

$$\overline{\Phi} = \frac{1}{T} \int_{t_0}^{t_0+\mathrm{T}} \Phi(t)\, dt \quad (4)$$

$$\sigma_\Phi^2 = \frac{1}{T} \int_{t_0}^{t_0+\mathrm{T}} (\Phi(t) - \overline{\Phi})^2\, dt \quad (5)$$

The combination of $R(t)$, $\Phi(t)$, $\overline{\Phi}$ and σ_Φ^2 yields a comprehensive characterization of a device's resource profile, supporting the definition of precise thresholds and adaptation rules for cryptographic protection in real time.

B. Adaptive protection level selection algorithm

The proposed methodology operates as a continuous cycle across three functional zones – Data Collection, Processing, and Decision & Adaptation – each driven by the device's hardware and energy characteristics. Initial configuration parameters $\alpha_i, \Phi_{min}^j, \Phi_{max}^j$, averaging window T, hysteresis interval Th, and risk coefficient $R_{crit}(t)$, guide the ongoing monitoring of metrics $x_i(t)$ and determine when and how the level of cryptographic protection should be adjusted. In our implementation, the monitoring interval T was set to 5 seconds, ensuring timely adaptation without excessive

overhead. Fig. 1 illustrates the swim-lane diagram that encapsulates this sequence of operations.

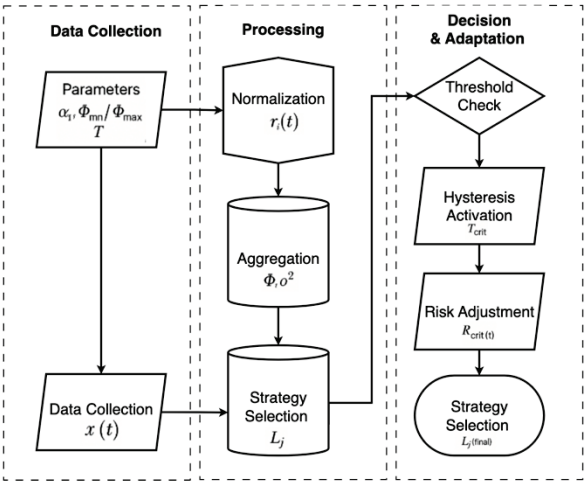


Fig. 1. Swim-lane diagram of the methodology for selecting a protection strategy based on the device resource profile

As shown in Fig. 1, the methodology is organized into three functional lanes, each comprising several sequential operations that together form the continuous adaptation cycle. The process begins in the Data Collection lane, where the system initializes configuration parameters, including metric weights, threshold values for protection levels, hysteresis interval, time window for averaging, and the initial risk coefficient. Following initialization, raw hardware metrics – such as processor frequency, available memory, and battery level – are periodically sampled from the device. These metrics are then forwarded to the Processing lane, where they undergo normalization using min–max scaling, resulting in a bounded resource vector $R(t)$. The normalized data is aggregated into a scalar index $\Phi(t)$ using a generalized $L_p - norm$, and its time-averaged value and variance are computed to capture the dynamics of resource availability. Based on the resulting index, an initial protection level is selected. In the Decision & Adaptation lane, the system verifies whether the index falls within predefined threshold zones and applies a hysteresis interval to suppress rapid oscillations between protection modes. If needed, the selected protection level is corrected by factoring in the current security risk coefficient, ensuring that high-risk situations trigger stronger cryptographic safeguards. Once the appropriate strategy is finalized and applied, the system re-enters the Data Collection phase to continue real-time adaptation.

IV. Methodology validation on the hardware and software stand

Before the direct application of the proposed methodology for adaptive selection of a protective strategy, a comprehensive validation was performed on a real hardware and software bench and in a simulation environment. The purpose of this stage was to confirm the correctness of the mathematical model of the resource profile, verify the mechanism for dynamic switching of protection levels, and quantify the performance and energy consumption in various scenarios.

First, a careful selection of equipment and setup of the modeling environment was carried out, ensuring the representativeness of the loads and the correct removal of metrics with minimal error. Then, a measurement procedure was developed, which includes fixing the «thermal» profile, averaging data with a sliding window, and preventing excessive switching using hysteresis.

The obtained data were used to calibrate the Φ index thresholds, verify the adequacy of the selection of the protective strategy in real time, and compare the results with theoretical expectations. The following sections provide a detailed description of the platform and measurement methodology, as well as summarized test results with an in-depth analysis of key indicators.

A. Experimental environment configuration

Experimental verification of the methodology was performed on the hardware-software bench built around the STM32F446RE board (ARM Cortex-M4, 180 MHz, 128 KB SRAM) running FreeRTOS 10.5 [21]. Power was supplied through the Rohde & Schwarz NGM202 laboratory unit in precision wattmeter mode (error ±0.05 mW). For network scenarios, the NS-3 simulation platform (ver. 3.40) was used, in which 10 IoT nodes were simulated with channel delays of 1–5 ms and random packet loss of 1%. The prototype cryptographic module implementing three levels of protection L_{light}, L_{medium}, $L_{intensive}$ was written in C using the AES hardware accelerator and ASCON-128a/ASCON-128b software implementations.

The diagram of the hardware stand, on which the collection of resource profile indicators and performance measurements were implemented, is shown in Fig. 2.

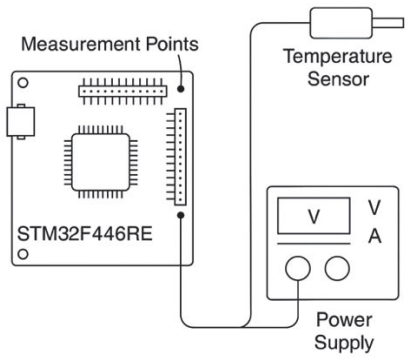


Fig. 2. Structural diagram of the hardware stand

To model network delays and packet losses, a simulation topology was implemented in the NS-3 environment, Fig. 3.

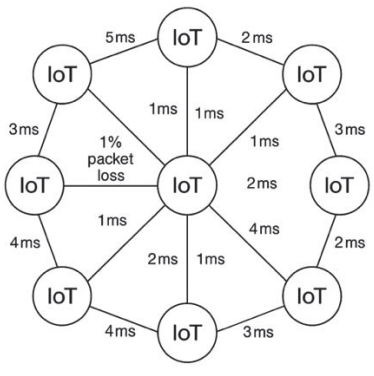


Fig. 3. Topology of a simulation environment with 10 IoT nodes in NS-3

As shown in Fig. 3, the nodes are connected with different transmission delays (1–5 ms), and the probability of packet loss is 1%. This structure reproduces the dynamics of communication channels typical for IoT networks.

The dynamics of power consumption during the operation of the security module with transitions between encryption levels based on the index $\Phi(t)$ is presented in Fig. 4.

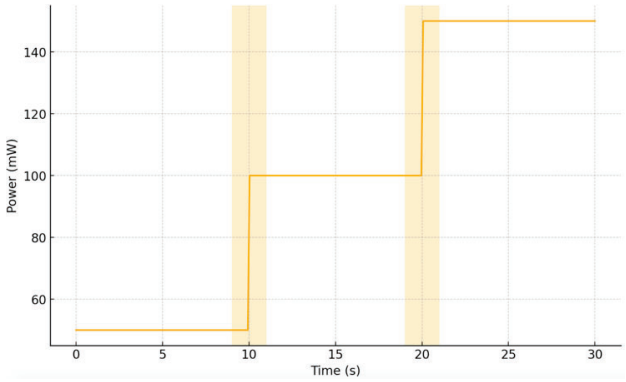


Fig. 4. Power consumption profile with hysteresis intervals between protection levels

Fig. 4 shows typical transitions between the «Light», «Medium» and «Intensive» modes according to changes in resource availability. The light intervals reflect the periods of hysteresis control that prevent unstable switching.

The measurements were performed in two stages. First, the «thermal» profile was determined: a 30-minute run of an empty cycle with registration of the crystal temperature (MAX31865) and the base power consumption P_0. Then, a series of 50 encryption-decryption cycles of 32-byte packets of each protection level was performed, measuring the execution time with the SysTick timer (resolution 1 μs) and the consumed energy $E = \int P(t)dt$. Averaging was performed with a sliding window $T = 5s$; hysteresis $Th = 2s$ prevented parasitic level switching. The thresholds $\Phi_{min/max}$ are pre-calibrated so that L_{light} is activated at $\Phi < 0.35$, L_{medium} in the range 0.35–0.7, and $L_{intensive}$ at $\Phi < 0.7$.

B. Evaluation criteria and interpretation of results

The evaluation of the effectiveness of the adaptive methodology was performed according to three main criteria: computational costs (latency of cryptographic operations), energy consumption and the level of cryptographic protection. Each experimental scenario corresponded to one of three protection levels: Light, Medium or Intensive, which were activated according to the current value of the resource availability index $\Phi(t)$. The results of averaged measurements are given in TABLE I.

TABLE I. PERFORMANS AND SECURITY METRICS FOR PROTECTION LEVELS

Scenario	Latency, ms	Energy, mJ	Security, bits
Light	5 ± 0,3	2 ± 0,1	64
Medium	15 ± 0,5	5 ± 0,2	128
Intensive	30 ± 0,8	10 ± 0,3	256

All energy values in Table I are expressed in millijoules (mJ) and represent the average energy consumption per single encryption–decryption cycle of a 32-byte packet. The measurements were obtained using a precision laboratory power supply with time-resolved current integration, and the resulting energy was averaged over 50 cycles per protection level to ensure statistical reliability.

As can be seen from the data obtained, the Light level provides the lowest latency and power consumption, but cryptographic stability is limited to 64 bits, which is suitable only for non-critical applications. The Medium configuration demonstrates a balanced ratio of energy consumption (5 mJ) and latency (15 ms), while providing 128-bit stability, which meets modern requirements for most IoT nodes.

The Intensive mode achieves maximum cryptographic reliability (256 bits), but requires a threefold increase in computational and energy costs compared to the Light mode. Its use is advisable for nodes with constant power supply or increased security requirements.

The results confirm the correctness of the selected thresholds Φmin, Φmax and the effectiveness of the proposed mechanism for adaptive switching between protection levels depending on the current resource profile of the device. Dynamic tuning allows you to maintain an acceptable level of performance in conditions of variable resource availability, while maintaining security characteristics at the proper level.

V. CONCLUSIONS

Unlike previous works that use static thresholds or class-based heuristics, the proposed method ensures real-time adaptation based on a formal resource model and experimentally calibrated parameters. In this paper, a novel methodology for adaptive selection of cryptographic protection strategies based on a device's resource profile has been proposed and validated. The approach is grounded on formalizing hardware and energy indicators into a multidimensional normalized vector, which is then aggregated into a scalar resource index $\Phi(t)$ using a generalized $L_p - norm$. Supplementary metrics, including time-averaged value $\bar{\Phi}$ and variance σ_Φ^2, were introduced to account for temporal dynamics and fluctuations in resource availability.

An adaptive decision-making algorithm was developed, incorporating threshold zones, hysteresis margins, and risk correction factors to enable real-time switching between predefined security levels (Light, Medium, Intensive). The methodology was implemented and experimentally validated using a hardware testbed based on the STM32F446RE microcontroller and a network simulation environment built in NS-3.

Experimental results demonstrate the effectiveness of the proposed strategy: the system dynamically adapts protection levels in response to changes in available resources, minimizing latency and energy consumption under constrained conditions while ensuring appropriate cryptographic strength. The proposed thresholds and switching logic proved robust across varying load profiles, confirming the practicality and scalability of the solution for distributed IoT environments.

The developed methodology can be further extended to heterogeneous networks with device-specific adaptation policies, and can serve as a foundation for resource-aware cryptographic frameworks in mission-critical IoT systems.

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The influence of reverberation, equalization and compression methods on speaker recognition

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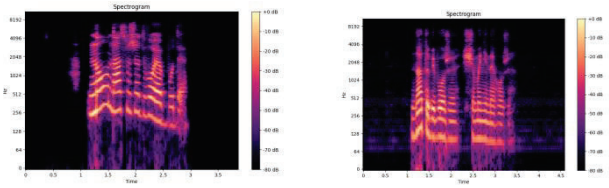
***Abstract*— The subject of the study is to analyze the impact of reverberation, compression and equalization methods on the accuracy of speaker speech detection. In recent years, the significance of audio preprocessing has grown considerably due to its crucial role in tasks such as data compression, filtering, and noise reduction. With the increasing demand for audio information classification, evaluating and comparing different preprocessing techniques has become essential to assess the accuracy and efficiency of the resulting solutions. This study focuses on examining how various audio preprocessing methods affect the accuracy of speaker speech detection, which is critical for subsequent neural network training. Particular emphasis is placed on methods of transforming audio signals into features, specifically using low-frequency cepstral coefficients. To achieve this objective, the study involved analyzing signal-to-feature transformation techniques and mathematical models used for audio sequence analysis based on the extracted features. A generalized real-time speaker speech translation model was developed, and an experiment was designed according to the selected audio preprocessing stages. The experiment was simulated by training and testing a convolutional neural network. The following methods were used: mel-frequency cepstral analysis, spectral analysis, mathematical methods of artificial intelligence. The following results have been obtained: depending on the chosen methods of pre-processing voice information and different ways of converting the signal into features, 86% accuracy of speaker's language detection was achieved. The practical value of this work lies in expanding the range of audio files that can be used for training neural networks, as well as enabling translation of the resulting text into a selected language, including constructed or artificial languages. This study explored how different audio preprocessing techniques influence the accuracy of detecting a speaker's speech, which is vital for subsequent neural network training involving various signal-to-feature transformation methods. The findings indicate that, in cases where audio recordings contain echoes, applying equalization can help reduce the echo effect.**

Keywords*— mel-frequency cepstral coefficients, compression, reverberation, equalization, normalisation, neural network, voice, audio order, speech

I. Introduction

The voice is one of the most important means of communication, conveying not only meaningful information, but also emotional state, intonation, and other nuances that form the unique character of each speaker. Modern voice processing technologies are designed to analyze and preserve these properties, ensuring high-quality transmission and recognition of speech in a variety of conditions [1,2.]

One of the key aspects of voice processing is spectral analysis, which allows you to determine the frequency components of an audio signal. This analysis helps to identify harmonics and formants that are characteristic of the human voice. Harmonics occur as multiples of the fundamental tone and are responsible for the timbre of the voice, making it richer and more expressive (fig 1). Formants, in turn, are the amplification of certain frequency ranges, which are particularly important for identifying speech sounds [3-5].



a) *b)*

Fig. 1. An example of a spectrogram with harmonics: a) starting at 256 Hz; b) starting at 100 Hz

In addition to the frequency spectrum, an important component of voice processing is the analysis of the time characteristics of the signal. Among them, the key ones are delay and decay, which determine the dynamic changes in volume and tone during the sound of the voice. These characteristics are important for the correct identification of sounds and words, as they reflect the natural fluctuations of the voice during speech [6-8].

Recording conditions and the environment also play a significant role in shaping the final audio signal. Various types of noise, echoes, and other acoustic effects can significantly distort the signal, making it difficult to process further. That is why an important aspect of voice processing is to reduce the impact of such factors and ensure maximum speech clarity and intelligibility.

Thanks to a combination of spectral and time-frequency analysis, modern voice processing systems can accurately identify and recognize voice signals even in difficult conditions. This provides the basis for further research aimed at improving the quality of speech recognition and adapting systems to different environments [9,10].

The relevance of this study stems from the growing need for effective communication in a globalized world, where international interaction is increasingly essential. Speaker recognition systems followed by translation into a selected language are gaining importance as they transform how people connect and engage across diverse cultural and linguistic contexts. These technologies are reshaping communication by bridging language barriers and enabling

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more seamless interaction worldwide. Thanks to this, speaker recognition systems can become an integral part of our digital lives, helping to make the world more connected and diverse. The practical significance of this work is to increase the accuracy of further transcription of audio information and translation of the generated text into the chosen language.

II. Theoreyical Foundations Of Speech Recognition And The Impact Of Audio Processing

Successful speech recognition is a complex process that depends on many factors, including the quality of the input audio data, the structure and architecture of neural networks, and audio preprocessing methods. In this section, we will review current approaches to speech recognition, focus on the role of neural networks in this process, and analyze the main audio processing methods that can affect the performance of speech recognition systems.

Speech recognition is one of the key technologies in the field of artificial intelligence, which is used in numerous industries, such as automation, interactive systems, mobile applications, and systems for people with disabilities. Modern speech recognition systems are based on neural networks, including Deep Neural Networks (DNNs) and Recurrent Neural Networks (RNNs). They provide high recognition accuracy due to their ability to model complex nonlinear relationships between different aspects of the speech signal.

Neural networks are used to analyze audio signals and convert them into text data. This allows them to consider the context of speech, tonal and rhythmic features, as well as other important aspects that determine the recognizability of speech signals. The main stages of a neural network include input signal processing, feature extraction, training on a large amount of data, testing and error correction.

Audio signal preprocessing is an important stage that directly affects the accuracy of speech recognition. The quality of the input audio signal can significantly affect the performance of a neural network, which makes it necessary to efficiently process signals before they are passed to recognition.

The purpose of pre-processing is to prepare the audio signal in such a way as to reduce the impact of noise, distortion, and other factors that can negatively affect the recognition quality (Fig. 2). Signal processing includes various techniques that allow you to improve its structure, extract the necessary features and increase the accuracy of the speech recognition model.

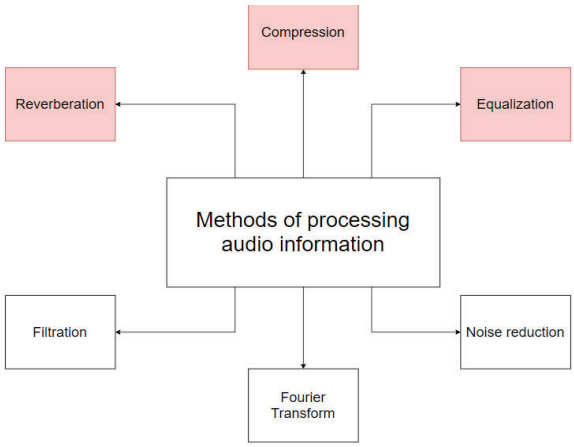


Fig. 2. Methods of processing audio information

Among the various methods of audio signal processing, the most important are reverberation, compression and equalization. Each of these methods has its own unique properties and impact on the sound quality, and therefore on the results of speech recognition.

Reverberation reproduces the effect of sound echoing or reflecting off different surfaces. It can be useful for improving speech intelligibility in environments where the signal is distant or highly distorted. Reverberation can add a sense of space and dimension to a sound, but excessive use of this technique can lead to a reduction in speech clarity. In cases where reverberation is too intense, it can make speech recognition more difficult, as the neural network may perceive the echo as noise or distortion.

Audio signal compression reduces the dynamic range of the sound, which reduces the difference between the loudest and quietest sounds. This is useful in environments where it is important to ensure that speech is heard evenly regardless of its volume. Compression can improve the recognition of quiet sounds, but it can also distort loud signals, making them difficult to identify.

Equalization is the process of adjusting the frequency spectrum of an audio signal to reduce the influence of unwanted frequencies and enhance those that are important for speech recognition. Equalization can be used to compensate for recording defects or environmental influences on the signal. However, incorrect equalization settings can lead to the loss of frequencies important for speech recognition, which reduces the accuracy of the neural network.

Audio signal processing methods can have both positive and negative effects on speech recognition results. The success of their application depends on many factors, including the quality of the original signal, recording conditions, and the architecture of the neural network. In this study, we analyze how different reverberation, compression and equalization techniques affect speech recognition accuracy in order to determine the optimal conditions under which the system achieves the highest performance.

The results of this analysis are important for the development of speech recognition systems, as they allow optimizing sound processing and improving the accuracy of neural networks in a variety of conditions.

III. Goals And Objectives Of The Work

The main goal of the article is to study the impact of reverberation, equalization and compression methods on speaker recognition.

To achieve the set goal, the following tasks must be solved:

- comparative analysis of types of reverberation, equalization and compression methods
- creation of a working dataset for further research;
- development of the methodology of the experiment;
- performing a study of the effect of reverberation, equalization and compression methods on speaker recognition ;
- analysis of the obtained results.

The conducted experiments are the basis for further research on the influence of preprocessing methods on the accuracy of speech and voice recognition based on neural network models.

IV. Research Methods And Materials

This section describes the research methodology, which includes the choice of neural network, audio processing methods, and the hardware and software used for the experiments. The process of collecting and preparing the data used for training and testing the speech recognition model will also be discussed.

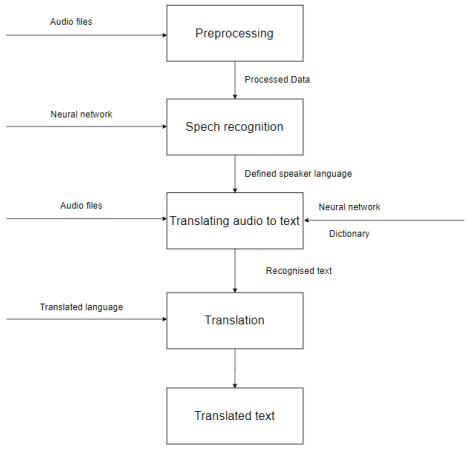


Fig. 3. Generalized structure of real-time translation of a speech [11,12]

The neural network architecture chosen (Fig. 3) for the study was one that has proven itself in speech recognition tasks. The choice of model was based on such criteria as accuracy, training speed, generalization ability, and resistance to various audio recording conditions. In this case, a deep neural network (DNN) with recurrent layers (RNN) was used to analyze time dependencies in speech signals.

The neural network consisted of several layers, in particular: an input layer, 12 hidden layers and an output layer. (fig. 5) For each hidden layer, activation functions were used to allow the model to learn from nonlinear data. Recurrent layers ensured the model's ability to analyze the audio sequence in a context-sensitive manner, which is critical for accurate speech recognition.

To test the impact of different audio processing methods on speech recognition accuracy, three main methods were selected:

- Reverberation.
- Compression.
- Equalization.

Each of these methods was applied to the original audio recordings, after which the processed signals were transferred to the neural network input for further analysis [13].

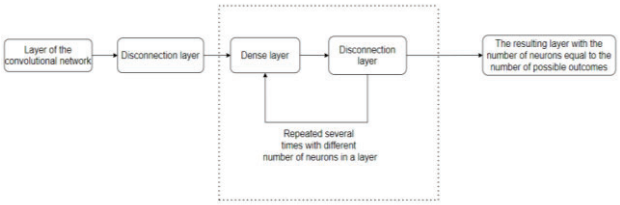


Fig. 4. Generalized structure of the neural network

To simulate the reverberation, we used software that allowed us to adjust the level and duration of the echo. Several reverberation configurations were applied, including weak, medium, and strong reverberation, to evaluate the impact of different levels of this effect on speech recognition results.

The audio signal was compressed using parameters that determined the level of reduction in the dynamic range of the sound:

- Threshold (measured in dB or percentage value relative to the maximum amplitude value).
- Compression ratio (a ratio that indicates how many times the amplitude of a signal exceeding the threshold is reduced).
- Sampling rate (the number of samples of an audio signal that are recorded or played back in one second, measured in hertz).

The study used different levels of compression (mild at 2:1 and strong at 8:1) to determine how changes in dynamic range affect the accuracy of recognizing speech fragments with different amplitudes.

Equalization was performed by adjusting the frequency spectrum of the audio signal. The study included an analysis of the impact of equalization on speech recognition, in particular, how amplification or reduction of certain frequencies can improve or worsen recognition results.

Parameters that were changed:

- Cutoff frequency.
- Sampling rate.

The data used for this model was collected from the Common Voice open-source speech corpus, which includes recordings of different speakers in different conditions. The corpus contains a wide range of variations in pronunciation, volume, and speech speed. This provided the necessary diversity to train the model and test its resistance to different conditions.

The pre-processing of the 4000 audio recordings, 1000 in each of the 4 languages, included steps that prepared them for the application of reverberation, compression and equalization techniques:

- Removal of background noise.
- Normalization of the volume level (an example of noise can be seen in fig 1 and 2, from 0 to 1 second of each recording).

Then the data was splited into samples for experiments in the ratio: 75% of the training sample and 25% of the test sample.

Specialized software for working with neural networks, such as TensorFlow and PyTorch[14], was used to conduct the study. The audio signals were processed using the Librosa software package[14], which provided the necessary tools for reverb, compression and equalization. The experiments were carried out on graphics processing units (GPUs), which allowed for fast and efficient processing of large amounts of data. An NVIDIA GeForce RTX 4070 Laptop GPU 8Gb was used for the experiment. This made it possible to conduct numerous experiments with different audio processing configurations and neural network settings, allowing us to increase the amount of input data and the number of neural network layers while keeping the running time unchanged.

Experimental procedure lies in the following steps - for each audio processing method, a series of experiments were conducted to simulate different preprocessing algorithms, namely

- Normalization.
- Noise reduction.
- Time mask.
- Frequency mask.

Each processed audio signal was passed to the neural network input for recognition, and the results were compared to the control data that was not processed by the percentage of correctly identified speaker languages.

After each experiment, the accuracy of speech recognition was evaluated to determine how different processing methods affect the final result. Error analysis was also performed to identify possible problems and their impact on system performance.

The results of these experiments were analyzed and used for further research and improvement of the speech recognition models.

TABLE I. EXPERIMENTAL RESULTS USING EACH SET OF PARAMETERS

Reverberation	Compression	Equalization	Speech detection accuracy, %
-	-	-	**86**
+	-	-	**77**
-	+	-	**81**
+	+	-	**76**
-	-	+	**79**
+	-	+	**86**
-	+	+	**81**
+	+	+	**79**

From the spectrogram results, you can understand that since reverb reproduces the effect of echo or reflection of sound from different surfaces, its use simulates audio files recorded in less-than-ideal conditions. However, its addition allows you to analyze the impact of only these interferences, not the total set of all interferences.

Each separate experiment with reverb, compression, and equalization worsened the neural network's performance, since the input was ideal files and each modification could only worsen the result. But we can see that compression had the least impact on this process.

A comparison of the results in the tables is shown in Fig 5.



Fig. 5. The resulting graph of speaker language detection accuracy for each experiment

As we can see from the result of the sixth experiment, equalization helped to reduce the impact of reverberation on the input data, improving the result from 77% (experiment 2) to 86%. From this we can conclude that when using less-than-perfect files, equalization can help to lessen echo.

V. CONCLUSION

This paper examined how various audio preprocessing techniques influence the accuracy of speaker speech detection, particularly in the context of preparing data for neural network training using different signal-to-feature conversion methods. By applying specific preprocessing approaches and feature extraction techniques, the study achieved an accuracy of 86% in identifying the speaker's language. The practical importance of this work lies in expanding the range of usable audio files for neural network training and enabling translation of the resulting text into the desired language, including constructed (artificial) languages.

The study concluded that when working with audio files that have echoes, equalization can help to combat them.

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Development of Operational Module Scale Options and Assessment of their Expert Weighting Coefficients

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***Abstract*— The use of information technologies and methods of mathematical statistics is currently a very promising direction for improving decision-making methods in medical practice. In particular, these narratives have shown quite capable results in the formation of analytical information in the diagnosis and treatment of patients with gunshot and mine-explosive injuries. Recently, the authors have conducted a number of field studies and collected statistical information, which allowed them to obtain mathematical dependences of treatment results on biomedical indicators, on the basis of which mathematical models of the rehabilitation processes of patients with gunshot wounds were obtained. For further formation of the methodology for choosing the most informative prediction of the course of surgical operations, scales of operational modules and an assessment of their expert weight coefficients are necessary, which were obtained by the authors as a result of this work.**

Keywords*— information technology, regression models, biomedical indicators, medical decision-making, surgical outcome prediction.

I. Introduction

The current state of emergency medicine poses many challenges regarding the complexity of technological processes for diagnosing and treating patients with gunshot and mine-explosive injuries, and requires the search for new technologies for these tasks [1]. In this sense, the creation and use of modern information technologies, as well as methods of mathematical analysis, modeling and prediction of the course of biomedical processes, is a rather important factor that creates new opportunities and allows significantly improving the results of treatment and rehabilitation of patients [1-2].

The presented work continues the analysis of the possibilities of creating new information technologies for processing databases of biomedical indicators obtained during the initial examination of patients with gunshot wounds.

The authors previously analyzed the results of experimental studies of medical institutions on biomedical indicators obtained during the initial examination of patients with gunshot wounds, and the results of the treatment. The conducted studies allowed forming an appropriate database, and calculating the correlation of the specified parameters [3-4]. In turn, this allowed us to obtain mathematical dependencies and develop mathematical models of the rehabilitation processes of patients with gunshot wounds [5-6]. For further formation of the methodology for choosing the most informative prediction of the course of surgical operations, the scales of operational modules and the assessment of their expert weight coefficients, which were obtained by the authors as a result of this work, are necessary.

II. Formation of Options of Operating Modules Taking into Account the Complexity of Surgical Operations

A. Technical features of the complexity of operations

The technical features of each surgical operation are determined by the following complications in patients with gunshot and mine-explosive injuries [7-8]:

- The removal of flap structures (epidermis and dermis, subcutaneous fat, fascia).
- The distance of tissue movement from the donor to the recipient site.
- The peculiarity of the supply (the presence of a perforating vessel (2-3 perforating vessels or 1).
- Flap tension along the axis of the segment.
- Flap rotation by 90 and 180 degrees within the segment or movement to another segment with preservation of axial blood flow. Selection of the optimal perforant for feeding the flap (from two or three perforating vessels in the case of flap rotation by 180 degrees). Taking into account the indicators of dynamic digital thermography and audio Doppler intraoperatively.

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• When choosing a flap for reconstruction, an important role is also given to how much % the function of the donor area decreases and how much the damaged anatomical structure is restored / the ideal balance of donor and recipient area restoration from 50/50 to 40/60.

All these factors and the analysis of thermographic and ultrasound indicators add a complete picture of objectification during the reconstruction of gunshot defects of the integuments tissues. We have chosen a structuring of the complexity of the surgical intervention that increases from 1 to (in terms of duration in terms of time of surgical interventions).

B. Medical features of surgical modules that use methods for determining integumentary tissues

Regarding the analysis of the features of surgical modules, their classification was made using methods for determining integumentary tissues [1, 9-11].

1) Methods for moving integumentary tissues.

We will consider the most common movements of integumentary tissues (Fig. 1):

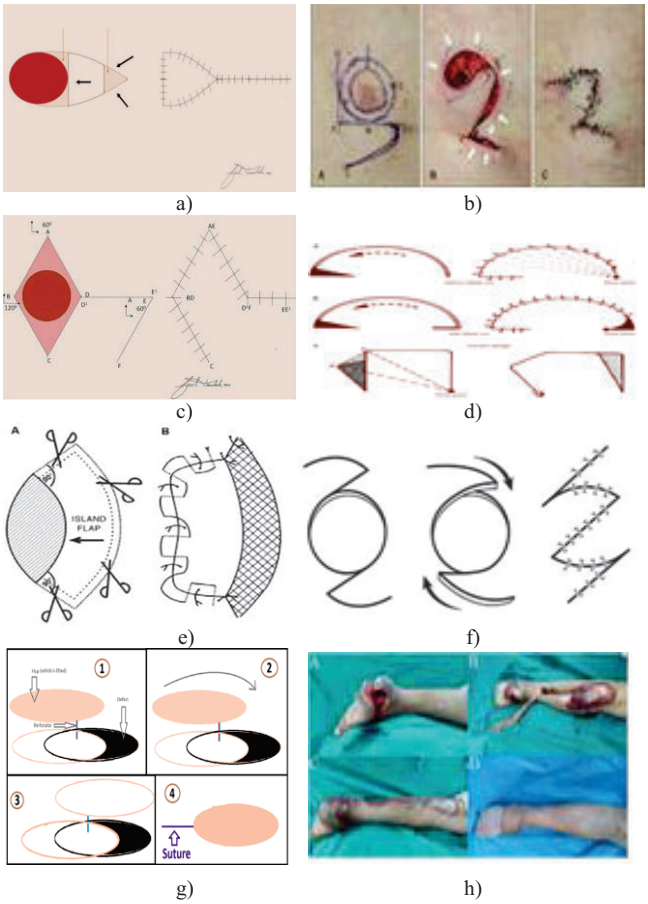


Fig.1. Methods of moving integumentary tissues: (a) - V-Y flap, (b) - Sliding flap, (c) - Limberg flap, (d) - Rotary flap, (e) - Keystone flap, (f) - Propeller flap, (g) - Hash- tag, (h) - Skin flaps.

• *V-Y flap (method of reconstruction "V-Y").* V-Y advancement of the flap with the original configuration in the form of a V. The flap is advanced in a linear direction, preserving the subcutaneous pedicle, which is at least 1/3 of the length of the flap. When the flap is closed, the final configuration is Y-shaped (see Fig. 1a).

• *Sliding flap* (sliding flaps – a technique for moving the soft tissue organ complex into the wound defect area by moving with a tension element (see Fig. 1b).

• *Limberg flap* (Classic rhombic (Limberg) flap) is a transposition flap used to eliminate defects in the shape of a rhombus with internal angles of 60° and 120°.

A rhombus in this configuration can also be considered as two equilateral triangles; thus, all sides of a rhombus have the same length. If the skin defect does not have a 60°-120° rhombus configuration, then additional skin can be removed to create this configuration. The first side of the flap is designed by extending a line equal to the length of the short diagonal that intersects one of the internal 120° angles. The second side is made parallel to the adjacent side of the rhombus and has an equal length to the first side. The flap is then transferred both by a rod and and a translational movement with the point of greatest wound tension in place (see Fig. 1c).

• *Rotary flap* (movement of integumentary tissues into the area of a wound defect of various shapes and angles of rotation), (see Fig. 1d).

• *Keystone flap* (Reconstruction using the Keystone method).

The Keystone perforator island flap was first described by Behan as a trapezoidal flap, which is essentially two connected V-Y island flaps. This is a relatively new concept that involves taking the skin and superficial fascia in a trapezoidal pattern adjacent to the defect to be closed. The vascular supply is maintained by the subcutaneous vasculature and depends on the fascial and muscular perforators (see Fig. 1e). The defect is excised in a circle and the wound is curved from both edges. The edges are now brought together to close the defect.

• *Propeller flap* (Propeller flaps) with a base powered by the perforator blood supply). The angle of rotation, depending on the perforator chosen, can be from 0 to 1800), (see Fig. 1f).

• *Hash-tag* - reconstruction of wound defects with opposing flaps with a feeding base in the form of triangular island flaps. During the movement of the flaps, a counter movement occurs with a conditional twisting of the edges relative to the axis. The distal corners of the flaps are cut off at the expense of one side (see Fig. 1g).

• *Reverse sural flap* (Reverse sural flap) is an axial flap that is commonly used in the treatment of distal wounds of the lower extremities. Its arterial blood supply depends on the retrograde flow coming from the septal-cutaneous perforators from the peroneal artery. The branches of the posterior tibial artery also contribute (see Fig. 1h).

A flap is a section of tissue that moves from one area to another and relies on its own blood supply. If the flap is divided at the vascular pedicle and moved to another area, it is called a “free flap”.

A common example is the radial artery flap of the forearm, which is supplied by the radial perforator arteries. Flaps are usually described by their shape or by the name of the surgeon who first described the flap.

Local flaps are used when the defect that needs to be covered is adjacent to the flap, and the flap is simply lifted and moved to cover the defect without dividing the vascular pedicle.

If the flap is moved within the same area, but not in close proximity to the defect, it is called a regional flap. If the flap is moved to a completely different area of the body, it is called a remote flap.

2) Method of moving the flap to cover the defect.

The second way to classify flaps is by the way the flap is moved to cover the defect. A pedicle flap is moved with the vessels remaining in place, acting as a support bridge. An island flap is a type of local flap in which a circular area of skin is removed along with the defect and the flap is moved to cover the defect.

• *An interpolated flap* is when the vascular pedicle of the flap is left intact, but the flap is advanced under or over a section of intervening skin. If the vascular pedicle is divided, the flap is called a pedicleless flap. A variation of a pedicleless flap is the free flap. This is when the vessels and other structures are excised, the skin is transplanted to another part of the body, and the vessels are anastomosed with those of the recipient site. An example is the free forearm flap, where a section of skin is lifted from the radial part of the forearm and the perforating branch of the radial artery is used as the blood supply.

• *Langer's lines* are lines through which minimal tension is applied when making incisions. This is an important surgical principle when making incisions in any part of the body, but in the context of facial surgery it is particularly important from an aesthetic point of view.

• *Propeller flap*: This flap is so named because the skin is twisted on a pedicle to cover the adjacent area requiring graft. Twisting of the vessels can cause blood clots and occlusion, but freeing the vessels from the skin reduces this risk.

• *Advance flap:* This is a simple incision of the defect followed by advancement of a rectangular flap of skin to cover the underlying tissues. Sufficient skin laxity is paramount.

• *V-progression:* In this flap, a V-shaped incision is made, with a subsequent incision extending from the V point. The incisions are made, and the V-shaped area of skin is advanced into a straight incision.

• *Island flap:* In this flap, a V-shaped incision is made with a subsequent incision extending from the V point. The incisions are made and the V-shaped area of skin is advanced into a straight incision.

• *Z-plasty:* A z-shaped incision is made. The angle between the z lines is generally considered to be 60 degrees.

Double Z-plasty with V-advancement / “jumping man”.

These are two Z-plasties on either side of a V-advancement flap. When the incisions are made, it resembles a jumping man. This flap is commonly used in the interdigital spaces and nasolabial fold.

• *O to Z flap:* The defect is excised in a circle and the wound is extended in a curved manner from both edges. The edges are now brought together to close the defect.

• *Other flaps:* There are a number of more complex flaps that are used to correct serious defects (e.g., to repair a cleft lip).

3) Burns:

• *Superficial:* A superficial burn usually results in blisters. If you break open these blisters, you will find pink and paler skin underneath. The skin will be tender and tender, while underneath it will be shiny and moist. These burns can be treated with silver-based dressings such as Urgotul, or paraffin for facial burns.

• *Partial-thickness burns:* Partial-thickness burns damage the superficial capillaries that supply the skin. Deeper areas will have a fixed color when palpated, while the deepest areas will be paler and may not turn pale at all. These burns are usually treated conservatively with dressings (such as Flamazine).

• *Full-thickness burns:* Full-thickness burns usually appear white and insensitive. If the burn has extended beyond the subcutaneous fat, the skin may appear charred and shriveled. Such burns inevitably require excision with covering the defect with a skin graft of varying thickness. These burns do not heal with the help of dressings.

C. Classification features of surgical operations and a list of operational modules in order of increasing technological complexity

A set of operational modules used for surgical operations on integumentary tissues in the outpatient clinic of the Military Medical Center of the Northern Region was selected for the study. Medical Forces Command of the Armed Forces of Ukraine, Kharkiv city. Surgical operations fully comply with the classification features, the list of which is normatively specified by the “Temporary Industry Classifier of Medical Procedures (Services) and Surgical Operations” [1]. Table 1 presents the features of the classes of surgical operations used to form the operational modules under study.

TABLE I. GENERAL NORMATIVE CLASSIFICATION CHARACTERISTICS OF SURGICAL OPERATIONS SELECTED FOR FORMING OPERATIONS MODULES

Classification sign	Type of surgical operation
ABA 7.01	Surgical treatment of soft tissues of the paranasal sinuses and areas
BC 7.04	Eyelid surgery
AAD 7.04	Primary surgical treatment of external ear wound
FE 7.15	Primary treatment of neck wound
GE 7.12-13	Plastic surgery of skin defects on the face
EE 7.02	Plastic surgery of soft tissue defects of the head
AN 7.04	Primary treatment of soft tissue wound
AN 7.07	Skin plastic surgery of wounds
FN 7.01	Plastic surgery of the posterior complex of the ligamentous apparatus
GN 7	Fasciotomy

III. DEVELOPMENT OF ORDINAL SCALE SAMPLE OPTIONS OF OPERATIONAL MODULES AND THEIR EXPERT WEIGHT COEFFICIENTS

A. Overview of typical measurement scales and justification of the choice of scale for describing the coefficients of expert weighting of operational modules

a) General definitions of metrology regarding the formation of measurement scales: Data analysis of any element involves the selection of an appropriate type of scale for their processing. According to the general provisions of metrology [12], a measurement scale is an agreed procedure for determining and designating qualitative (non-metric scale) or quantitative (metric scale) properties of the object of study. The characteristics of the features of the analyzed data must be consistent with the type of scale that is adapted for their representation and processing. The success of any data analysis method depends on the compliance of the analyzed data with its initial assumptions. Methods suitable for one type of data can lead to serious errors when used for data of other types. The first stage of any data analysis is usually the determination of the type. The main one is the classification of data by measurement scales. According to it, the following types of features are distinguished.

b) Nominal features: Nominal features (features with disordered states, classification features) are data that are measured on a nominal scale.

Class names can be expressed using numbers, but these numbers can only be used to answer the question: do two objects belong to the same class or not. Examples of nominal features are names of biological species, names of academic disciplines, colors, etc. 3 From the point of view of automating data analysis and applying standard algorithms, it is advisable to choose the following class designations: 0, 1, 2, ...*N.* But with these numbers it is not possible to perform any actions except checking їx equality or inequality.

c) Ordinal features (features with ordered states, ordinal features): Ordinal features (features with ordered states, ordinal features) are data that are measured in ordinal scales. These data can be compared with each other in a certain ratio: "more - less", "lighter - heavier", "more right - more left", etc. Examples of ordinal features are the strength of an earthquake, military ranks, student grades, etc. If the value of the ordinal feature is numbers, then they can also be used to compare the degree of manifestation of a classification feature, but the distances between classes will not be determined.

TABLE II. CHARACTERISTICS OF THE MAIN TYPES OF DATA ON THE INVESTIGATED CHARACTERISTICS

Measurement scale	**Defining relations**	**Equivalent transformations**	**Permissible operations on data**	
			Primary processing	***Secondary processing***
Nominal	Equivalence	Name permutations	Calculation of the Kronecker symbol δ_{ij}	Calculation of relative frequencies and operations on them
Ordinal	Equivalence, preference	Monotone (those that do not change the order)	Calculation of δ_{ij} and ranks R_i	Calculation of relative frequencies and quantiles
Interval	Equivalence, preference, preserving the ratio of intervals, periodicity	Linear transformations $y = ax + b,$ $a > 0,$ $b \in R$	Calculation of δ_{ij}, ranks R_i and intervals (differences between data)	Arithmetic operations on intervals
Cyclic	Equivalence, preference, preserving the ratio of intervals periodicity	Shift $y = x + nb,$ $b = const,$ $n = 1, 2, 3 \ldots$	Calculation of δ_{ij}, ranks R_i and intervals (differences between data)	Arithmetic operations on intervals
Relation	Equivalence, preference, preserving the ratio of intervals, preserving the ratio of two values	Stretching $y = ax,$ $a > 0,$	All arithmetic operations	Any suitable processing
Absolute	Equivalence, preference, preserving the ratio of intervals, preserving the ratio of two values, absolute and dimensionless unit, absolute zero.	Does not exist (scale with unique)	All arithmetic operations, using, as an exponent, the base and aprymment of the logarithm	Any necessary processing

d) Quantitative (numerical, variational) features: Quantitative features are features that are measured in quantitative (interval, relative, cyclic and absolute) measurement scales. The operations that can be performed with numerical characteristics of data depend on the measurement scale.

In a generalized form, the characteristics of the main types of data, according to [13], are given in Table 2.

Data obtained in higher-rank scales can be reduced to lower-rank scales. For example, data measured in a ratio scale can be reduced to an interval scale.

Such transformations are called scale reduction. The need for them usually arises when processing data measured in scales of different types. The reverse operation - the conversion of data measured in lower scales to higher ones - is considered incorrect. Scale reduction leads to the loss of part of the available information about the studied features.Important types of classification are the division of features by discreteness or continuity of the theoretical distribution function, by the distribution law, etc. The Kronecker symbol, shown in Table 2, (or Kronecker delta, or Kroneckerian), is an indicator of equality of elements, formally: a function of two integer variables, equal to 1 if they are equal, and 0 otherwise:

$$\delta_{ij} = \begin{cases} 1, & i = j, \\ 0, & i \neq j. \end{cases} \quad (1)$$

Since the expert weight coefficients have qualitative characteristics that are directly related to the complexity of surgical operations, we choose an ordinal non-metric scale for their description.

B. Development of options for the scale of operational modules and assessment of their expert weight coefficients

The choice of an ordinal non-metric scale for options for operational modules implies the fulfillment of certain conditions:

• The scale should be represented by a table, the columns of which reflect the features of medical technology not only by the equivalence ratio, but also by the ratio of the order in increasing quantitative detection of the complexity of the operation.

• The quantitative manifestation of complexity should be represented by reference points, to which numbers are assigned in the form of expert scores (codes of complexity of operations).

To increase the informativeness of the scale, its reference points can be accompanied by an indication of the coefficients of their expert weight, which take into account the a priori probability of manifestation of a particular complication and the complexity of the surgical operation. Such consideration can be carried out at the stage of predicting the operation code for statistical correction of the final results [14-15]. A database for 566 wounded was used to assess the a priori probabilities.

To form an ordinal scale, sets of factors were used that are associated with the levels of severity of injuries and the features of the complexity of the surgical operation, which were described in subsection 1 of this article. Two scales were developed, for which their own subsets of factors with qualitative and quantitative levels were used.

• Scale A. Factors: surgical complexity of the operation, time of the operation, number of wounds (one qualitative and two quantitative factors).

• Scale B. Factors: operation time, wound area (volume), number of bed days (three quantitative factors).

Each of the scales includes its own order of operational modules, the abbreviated names of which correspond to the list of "Methods of moving integumentary tissues" (see subsection II-A of this article). Also provided are reference points (expert assessment of the complexity code from 1 to 8) and the corresponding expert weight coefficients $P_i, i = \overline{1,8}$.

Table 3 presents the results of the formation of scale A.

TABLE III. SCALE A: SURGICAL COMPLICATION OF THE OPERATION

Operational module	**V-Y flap**	**Sliding flap**	**Limberg flap**	**Rotary flap**	**Key-stone flap**	**Prope-ller flap**	**Hash-tag**	**Reverse sural flap**
Complexity code	1	2	3	4	5	6	7	8
Expert weighting coefficient	0,0171	0,0171	0,0513	0,0855	0,5384	0,1453	0,1111	0,0342

Table 4 presents a scale of total treatment costs, in which, instead of the names of the operational modules, their numbers are given according to the previous table.

TABLE IV. SCALE B: TOTAL TREAMENT COSTS

Operation module number	**2**	**8**	**4**	**1**	**5**	**3**	**7**	**6**
Complexity code	1	2	3	4	5	6	7	8
Expert weighting coefficients	0,1024	0,0513	0,2905	0,3161	0,0855	0,0855	0,0425	0,0262

The developed scales allow us to proceed to the formation of a methodology for selecting the most informative prediction of the course of surgical operations.

IV. CONCLUSION

Thus, the analytical review considered the features of surgical modules and their classification in the treatment of patients with gunshot and mine-explosive wounds, taking into account the existing features of surgical operations and the list of surgical modules in increasing technological complexity. To further form the methodology for choosing the most informative prediction of the course of surgical operations, the necessary scales of surgical modules A and B were developed for this, and their expert weight coefficients were assessed.

Further work involves the creation of new information technologies for processing databases obtained during the primary study by expanding the dimensionality of the vector of primary parameters by analytical methods, and their use to improve mathematical models of biomedical processes in penetrating shrapnel wounds.

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Thermodynamic Modeling of ZrO_2 – TiO_2 – MgO System

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***Abstract*—Targeted synthesis of new materials with specified operational properties is impossible without thermodynamic analysis of multicomponent systems. Of particular interest is the study of the three-component system ZrO_2–TiO_2–MgO, as it opens up prospects for the creation of heterophase ceramic materials with high dielectric characteristics. In this work, a thermodynamic analysis of the subsolidus structure of the ZrO_2–TiO_2–MgO system was performed by comparing the values of the Gibbs free energy for model solid-phase exchange reactions. It was found that the system is divided into 5 elementary triangles: TiO_2 – $MgTi_2O_5$ – $ZrTiO_4$, ZrO_2 – $ZrTiO_4$ – $MgTi_2O_5$, ZrO_2 – $MgTiO_3$ – $MgTi_2O_5$, ZrO_2 – $MgTiO_3$ – Mg_2TiO_4, ZrO_2 – MgO – Mg_2TiO_4. The results of the study of the subsolidus structure of the ZrO_2–TiO_2–MgO system provide researchers with important information for the targeted synthesis of dielectric materials in this system with specified performance characteristics. Thus, a comprehensive study of the ZrO_2–TiO_2–MgO system facilitates the development of new dielectrics with controlled microstructure, resistance to thermal effects and stable electrical parameters, which is especially important for the creation of components for electronic equipment, heat-protective coatings and sensor devices. To develop new types of dielectric materials with increased thermal resistance, it is rational to use the concentration region of the ZrO_2–TiO_2–MgO system, which is limited by the elementary triangle ZrO_2 – MgO – Mg_2TiO_4.**

Keywords—dielectric materials, phase formation, thermodynamic analysis, subsolidus structure, geometric-topological characteristics

I. Introduction

Directed synthesis of new materials with specified performance characteristics is impossible without thermodynamic analysis of multicomponent systems. The study of the three-component system ZrO_2–TiO_2–MgO is of interest for the creation of heterophase ceramic materials with dielectric properties. The system includes three simple oxides, the properties of which are sufficiently studied. MgO is periclase, it exists in only one modification, which has a cubic crystal lattice. TiO_2 is characterized by the existence of three polymorphic modifications: anatase, brookite, and rutile. In the high temperature range, TiO_2 is stable in the rutile modification, which has an orthorhombic crystal lattice. Zirconium dioxide (ZrO_2) has three crystal modifications: monoclinic, tetragonal and cubic. The transition between them depends on temperature and pressure. The narrow-temperature tetragonal modification is metastable and can transform into monoclinic with decreasing temperature, which is accompanied by an increase in volume, and can lead to the formation of internal stresses.

The phase diagram of the MgO – TiO_2 system has been studied for quite some time. Earlier studies dealt with two compounds (quandilite Mg_2TiO_4 and carroite $MgTi_2O_5$). Later, the stoichiometric compound geikielite $MgTiO_3$ was described. Much discussion has been held on the issue of melting and decomposition temperatures of the relevant compounds. All studies on this issue are described in detail in the literature [1]. It was decided to use data from [2, 3], which are close in value. According to [2], Mg_2TiO_4 and $MgTiO_3$ decompose at 2029 K and 1903 K, respectively, and the melting point of the third compound, $MgTi_2O_5$, is 1935 K. These data were clarified in the literature [3]: Mg_2TiO_4 melts incongruently at 2033 K, $MgTi_2O_5$ melts congruently at 1933 K. Two eutectics were established in the system at temperatures of 1903 and 1905 K. Disproportionation of Mg_2TiO_4 and $MgTi_2O_5$ in the low-temperature range occurs with volume changes:

below 933 K: $Mg_2TiO_4 \rightarrow MgTiO_3 + MgO$;

below 403 K: $MgTi_2O_5 \rightarrow MgTiO_3 + TiO_2$.

Above 933 K, both Mg_2TiO_4 and $MgTi_2O_5$ exhibit cationic ordering – disordering processes (the enthalpy of the process at 933 K is approximately 1000 J/mol), while in $MgTiO_3$ similar processes are characteristic below 1673 K.

Magnesium dititanate $MgTi_2O_5$ has a pseudobrookite structure. Magnesium titanate ceramics are dielectric materials and are used in microwave technologies [4, 5]. Mg_2TiO_4 has a spinel-type structure. Ceramics based on Mg_2TiO_4 have wide applications as dielectrics in resonators, filters and antennas for communication, radars and global

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positioning systems operating at microwave frequencies [6–8]. The MgO–TiO_2 system is of great importance for materials science and titanium metallurgy, and it is also of interest for petrology.

In the ZrO_2–MgO system [9] no chemical compounds have been found, there are solid solutions of zirconium dioxide in periclase with a very narrow homogeneity region, as well as solid solutions of MgO in ZrO_2. The system is of great importance for materials science; its independent petrological significance is small.

Initially, no chemical compounds were found in the ZrO_2–TiO_2 system. Further studies in the high-temperature region revealed the presence of the $ZrTiO_4$ compound, which melts incongruently. It is now a well-known compound in the field of electroceramics (in the production of dielectric resonators and telecommunications devices) [10, 11].

The ZrO_2–TiO_2 system is of interest for applications in electrical and optical devices such as capacitors, piezoelectric sensors, ultrasonic motors and microwave dielectric resonators [12, 13].

Phase equilibria in the ZrO_2–TiO_2–MgO system were studied in 1955, on the basis of which a preliminary phase diagram of the ternary system was constructed in the temperature range of 1673–2023 K [14].

In [13], the authors conducted an experimental study of phase interactions in the ZrO_2–TiO_2–MgO system above 1530 K to estimate the thermodynamic parameters of this system at 1530 (Fig. 1), 1680 (Fig. 2) and 1880 K (Fig. 3), however, these data do not provide a complete picture of phase equilibria in this system.

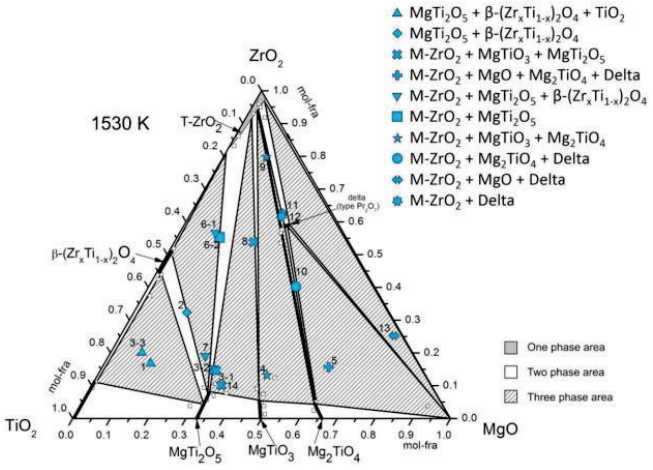


Fig. 1. Constructed isothermal section of the ZrO_2–TiO_2–MgO system at 1680 K [13]

II. Theoretical Information and Research Methods

Despite the apparent simplicity of the ZrO_2–TiO_2–MgO system (there are no binary compounds in the ZrO_2–MgO subsystem, only one binary compound in the ZrO_2–TiO_2 subsystem, and three binary compounds in the MgO–TiO_2 subsystem), its subsolidus structure has not been reliably established, and its phase diagram has not been studied.

According to [15], in the subsolidus structure of the system, the presence of conodes should be taken into account: $ZrTiO_4$–$MgTi_2O_5$, ZrO_2–$MgTi_2O_5$, ZrO_2–$MgTiO_3$, ZrO_2–Mg_2TiO_4.

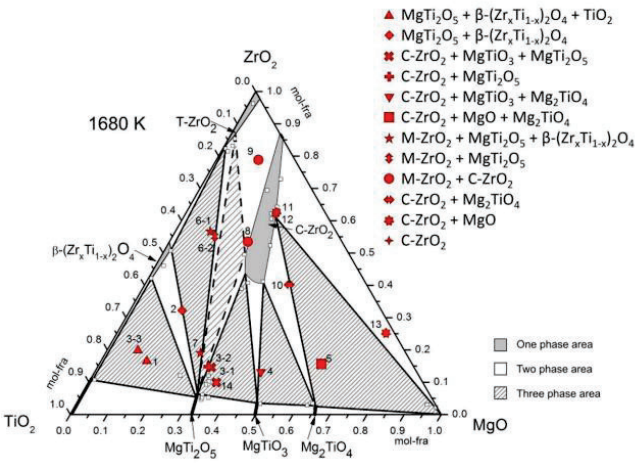


Fig. 2. Constructed isothermal section of the ZrO_2–TiO2–MgO system at 1680 K [13]

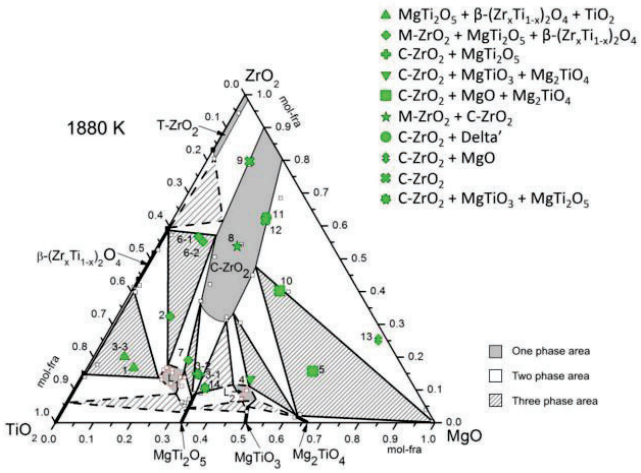


Fig. 3. Constructed isothermal section of the ZrO_2–TiO_2–MgO system at 1880 K [13]

According to [15], in the subsolidus structure of the system, the presence of conodes should be taken into account: $ZrTiO_4$–$MgTi_2O_5$, ZrO_2–$MgTi_2O_5$, ZrO_2–$MgTiO_3$, ZrO_2–Mg_2TiO_4.

However, at the time of publication of the mentioned monograph, experimental studies had not been conducted and there were no confirmed thermodynamic characteristics for the $ZrTiO_4$ compound, and discussions about the coexistence and temperature ranges of stability of other compounds in the ZrO_2–TiO_2 system are still ongoing: information about the synthesis of the $Zr_5Ti_7O_{24}$ compound from sol-gel composites in 1998 was not confirmed; the $ZrTi_2O_6$ (zirconium dititanate) compound is considered as the extreme member of the $(Zr,Ti)_2O_4$ solid solutions in the low-temperature region (~ up to 1333 K) [16]; the range of stoichiometric stability of $ZrTiO_4$ is limited from 1340 K to 1420 K [13]. Only in the last of the studies mentioned was a dependence established for heat capacity $ZrTiO_4$ [12]:

$$C_p(J/mol\cdot K) = 145.694 + 0.00801\cdot T - 2.92\cdot 10^6/T^2,$$

where temperature T is in K.

Accordingly, only now has it become possible to perform accurate thermodynamic calculations for phase equilibria in the ZrO_2–TiO_2–MgO system and triangulate it.

Thus, the aim of this work is to study the processes of phase formation in the ZrO_2–TiO_2–MgO system using thermodynamic methods of analysis.

In this work, a thermodynamic analysis of the subsolidus structure of the ZrO_2–TiO_2–MgO system was performed by comparing the values of the Gibbs free energy for model solid-phase exchange reactions 1–3:

1. $ZrTiO_4 + 2MgO = ZrO_2 + Mg_2TiO_4$,
2. $ZrTiO_4 + MgO = ZrO_2 + MgTiO_3$,
3. $2ZrTiO_4 + MgO = 2ZrO_2 + MgTi_2O_5$,

in the temperature range of 1000–1800 K according to the method [15], using the thermodynamic data presented in Table 1.

Analysis of the obtained data (Table 2) allows us to judge the direction of solid-phase reactions and the preference for the formation of certain combinations of phases [17].

TABLE I. THERMODYNAMIC CONSTANTS OF COMPOUNDS [12, 15, 19]

Compound	$-\Delta H^0_{298}$ kJ/mol	S^0_{298} J/(mol·K)	$C_p = a + bT + c'T^{-2}$, J/(mol·K)		
			a	$b\cdot10^3$	$-c\cdot10^{-5}$
MgO	601.241	26.924	42.59	7.28	6.19
ZrO_2	1100.81	50.71	69.62	7.53	14.058
TiO_2 anatase	912.53	34.727	44.225	15.062	7.782
TiO_2 rutile	943.492	49.915	74.6	2.092	17.698
$MgTiO_3$	1571.93	74.56	118.37	13.27	27.32
Mg_2TiO_4	2163.55	115.10	154.64	35.73	28.83
$MgTi_2O_5$	2507.89	138.91	170.21	38.49	30.75
$ZrTiO_4$	2024.1	116.71	145.694	8.01	29.2

TABLE II. THE RESULTS OF CALCULATING THE CHANGE IN THE VALUES OF GIBBS FREE ENERGY DEPENDING ON TEMPERATURE

№ reaction	ΔG, kJ/mol, at temperature, K				
	1000	1200	1400	1600	1800
1	-35.126	-35.905	-37.294	-39.323	-42.015
2	-31.901	-29.659	-27.642	-25845	-24.268
3	-33.451	-27.306	-21.536	-16.264	-11.582

III. DISCUSSION OF THE RESULTS

In accordance with the obtained data (Table 2), the triangulation of the system has the following form (Fig. 4). The obtained data are consistent with the data of Berezhnoy A.S.

As a result of the thermodynamic analysis of the ZrO_2–TiO_2–MgO system, the authors established that the system is divided into 5 elementary triangulations (Fig. 4): TiO_2 – $MgTi_2O_5$ – $ZrTiO_4$, ZrO_2 – $ZrTiO_4$ – $MgTi_2O_5$, ZrO_2 – $MgTiO_3$ – $MgTi_2O_5$, ZrO_2 – $MgTiO_3$ – Mg_2TiO_4, ZrO_2 – MgO – Mg_2TiO_4.

The results of the division of the ZrO_2–TiO_2–MgO system into elementary triangles correspond to Kurnakov's rule [15]: X = 1 + 4 = 5. Due to the absence of ternary compounds in the system, the topological graph of the interconnection of elementary triangles is a straight line.

To characterize the subsolidus structure of the ZrO_2–TiO_2–MgO system, the connode lengths (L) were calculated, the results of which are presented in Table 3.

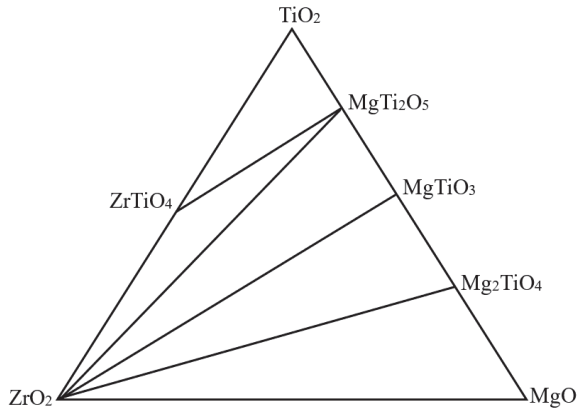


Fig. 4. Subsolidus structure of the ZrO_2–TiO_2–MgO system

TABLE III. LENGTHS OF CONNODES

№	Connodes	L, ‰
1	TiO_2 – $MgTi_2O_5$	200
2	TiO_2 – $ZrTiO_4$	570
3	$MgTi_2O_5$ – $ZrTiO_4$	501
4	ZrO_2 – $MgTi_2O_5$	917
5	ZrO_2 – $MgTiO_3$	882
6	ZrO_2 – Mg_2TiO_4	866
7	ZrO_2 – MgO	1000
8	$MgTi_2O_5$ – $MgTiO_3$	133
9	Mg_2TiO_4 – $MgTiO_3$	167
10	Mg_2TiO_4 – MgO	500
11	ZrO_2 – $ZrTiO_4$	430

From Table 3 it can be seen that the ZrO_2 – $MgTi_2O_5$ connode has the maximum length, and the $MgTi_2O_5$ – $MgTiO_3$ connode has the minimum length (ZrO_2 – MgO was not taken into account, since it coincides with the side of the concentration triangle of the system).

The main geometric-topological characteristics of the subsolid structure of the system and its phases, including the areas of elementary triangles, the degree of their asymmetry (L_{max}/L_{min}), the area of regions in which phases exist, and the probability of the existence of phases in the ZrO_2–TiO_2–MgO system, calculated using the method [15], are given in Tables 4, 5.

TABLE IV. CHARACTERISTICS OF ELEMENTARY TRIANGLES OF THE ZrO_2–TiO_2–MgO SYSTEM

№	Elementary triangle	Area, ‰	Degree of asymmetry
1	TiO_2 – $MgTi_2O_5$ – $ZrTiO_4$	114	2.85
2	ZrO_2 – $ZrTiO_4$ – $MgTi_2O_5$	86	2.13
3	ZrO_2 – $MgTiO_3$ – $MgTi_2O_5$	133	6.89
4	ZrO_2 – $MgTiO_3$ – Mg_2TiO_4	167	5.28
5	ZrO_2 –MgO – Mg_2TiO_4	500	2.00
Σ max min		1000 500 86	– 6.89 2.00

From the presented results (Tables 4) it is clear that the largest area is the elementary triangle ZrO_2 – MgO – Mg_2TiO_4 (500 ‰). It also has the smallest degree of asymmetry (2.00). This indicates the absence of the need for precise dosage of components and thorough mixing. Mg_2TiO_4, which is part of this elementary triangle, has the highest probability of existence (Tables 5), excluding the original component ZrO_2.

Mg_2TiO_4 ceramics with cubic spinel structure are widely used in the field of microwave dielectric materials [19], due to the fact that Mg_2TiO_4 has the following microwave dielectric properties: $\varepsilon_r = 14$, Qf = 150,000 GHz and $\tau_f = -63$ ppm/°C [20].

TABLE V. GEOMETRIC-TOPOLOGICAL CHARACTERISTICS OF THE PHASES OF THE SYSTEM ZrO_2–TiO_2–MgO

Compound	With how many phases does it coexist	In how many triangles is there	Total area of existence, ‰	Probability of existence, ω, fractions of units
ZrO_2	5	4	886	0.295
TiO_2	2	1	114	0.038
MgO	2	1	500	0.167
$MgTi_2O_5$	4	3	333	0.111
$MgTiO_3$	3	2	300	0.100
Mg_2TiO_4	3	2	667	0.222
$ZrTiO_4$	3	2	200	0.067
Σ max min			3000 886 200	1.000 0.295 0.038

The elementary triangles TiO_2 – $MgTi_2O_5$ – $ZrTiO_4$ and ZrO_2 – $MgTiO_3$ – $MgTi_2O_5$ have relatively similar areas – 114 ‰ and 133 ‰, respectively, but ZrO_2 – $MgTiO_3$ – $MgTi_2O_5$ has a high degree of asymmetry. Also, the elementary triangle ZrO_2 – $MgTiO_3$ – Mg_2TiO_4 with an area of 167 ‰ has a high degree of asymmetry. Thus, when working in the specified areas of the ZrO_2–TiO_2–MgO system, it is necessary to strictly adhere to special technological methods of mass preparation and synthesis.

IV. CONCLUSIONS

Thermodynamic modeling of this system allows us to establish stable phases, determine the conditions of their formation and interaction, and construct phase diagrams necessary for predicting the behavior of materials at different temperatures and component ratios. Zirconium oxide (ZrO_2) has high thermal and chemical stability, titanium dioxide (TiO_2) has pronounced dielectric properties, and magnesium oxide (MgO) has good insulating characteristics and a high melting point.

A comprehensive study of the ZrO_2–TiO_2–MgO system facilitates the development of new dielectrics with controlled microstructure, thermal stability and stable electrical parameters, which is especially important for the creation of electronic components, heat-protective coatings and sensor devices.

Thus, the results of the study of the subsolidus structure of the ZrO_2–TiO_2–MgO system provide researchers with important information for the synthesis of dielectric materials in this system with a given phase composition. To develop new types of dielectric materials with increased thermal resistance, it is rational to use the concentration region of the ZrO_2–TiO_2–MgO system, which is limited by the elementary triangle ZrO_2 – MgO – Mg_2TiO_4.

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Artifact Correction Method for M-response using UNet1D Network

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***Abstract*—this article explores the application of convolutional neural networks (CNNs), specifically the UNet1D architecture, for filtering M-response signals − an essential component of electromyographic (EMG) diagnostics used to assess the functional state of the peripheral neuromuscular system. The goal of the study is to improve signal quality by effectively removing artifacts while preserving the amplitude-frequency characteristics critical for clinical interpretation. The model was trained on real biomedical data containing various types of noise and artifacts. Results show that the proposed model achieves high filtering accuracy with minimal deviation from the original signals (a mean amplitude deviation of only −0.04%, and a 20.75% increase in signal-to-noise ratio, SNR). The CNN-based approach demonstrated strong potential for preserving diagnostically significant features and can be effectively integrated into clinical EMG workflows to enhance diagnostic precision and reduce the impact of external interference.**

Keywords—Healthcare, medical diagnostic, information medical technologies, functional diagnostics, UNet1D, M-response, artifact correction.

I. Introduction

The accurate assessment of the peripheral neuromuscular system is critical for diagnosing and managing a wide range of neurological disorders. Among the electrophysiological tools available, the M-response, a fundamental component of the electromyography (EMG) signal, plays a pivotal role in evaluating the functional integrity of peripheral nerves and muscles. Its analysis is particularly crucial in the context of diabetic polyneuropathy, a condition that requires a very high stimulus intensity and often results in responses of low amplitude. This common and often debilitating condition characterized by widespread damage to peripheral nerves, where precise M-response measurements can reveal the extent and nature of nerve dysfunction. However, the diagnostic utility of the M-response is frequently hampered by the contamination of recorded signals with various forms of noise, artifacts, and interference from external factors. Traditional filtering techniques, such as conventional low-pass and high-pass filters, often fall short in providing the necessary precision, sometimes distorting or removing diagnostically significant characteristics of the signal in the process of noise reduction. In recent years, advanced approaches rooted in artificial intelligence, particularly Convolutional Neural Networks (CNNs), have emerged as powerful tools for biomedical signal processing [1], [4], [5]. CNN models have demonstrated a remarkable ability to automatically identify and filter relevant features within complex signals, obviating the need for manual feature engineering [2], [6]. This article investigates the application of a CNN-based architecture for filtering M-response signals, aiming to enhance the accuracy of diagnostic conclusions, especially in the assessment of conditions like polyneuropathy, by effectively suppressing noise while preserving the crucial amplitude-temporal parameters essential for robust clinical analysis.

The fidelity of the M-response can be compromised by various noise sources, potentially obscuring the true waveform and affecting diagnostic accuracy. These sources can be broadly categorized:

Physiological Noise: Originating from the subject, this includes voluntary or involuntary muscle activity (EMG contamination), movement artifacts, and interference from other biological signals such as the electrocardiogram (ECG), especially in proximal recordings.

Environmental Noise: External electromagnetic interference is a common culprit, most notably power-line hum (50/60 Hz) from nearby electrical wiring and equipment, as well as radiofrequency (RF) signals from wireless devices and other electronic apparatus.

Technical Factors: Issues with the recording setup contribute significantly. These include high electrode-skin impedance due to poor skin preparation or faulty electrodes, suboptimal electrode placement, inadequate grounding, damaged or unshielded cables, and intrinsic amplifier noise. Additionally, a large stimulus artifact, while not random noise, can obscure the M-response onset if not properly managed.

Before the application of deep learning, several conventional techniques were employed to mitigate stimulus artifacts (SA) in M-response EMG. Simple blanking or windowing the SA period proved ineffective when the artifact overlapped with the M-response. Linear filters (e.g., Butterworth, FIR) struggles due to the often-shared frequency spectra of the SA and the desired M-wave [5], [7], [10].

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Template subtraction, which relies on subtracting an averaged SA profile, was limited by the SA's variability with stimulus intensity, electrode impedance, and tissue characteristics. While wavelet denoising offered more nuanced time-frequency separation, it demanded careful selection of wavelet type and thresholding strategies. Similarly, adaptive filtering (e.g., LMS, RLS) required a suitable noise reference correlated with the SA, which could be challenging to derive reliably. Statistical methods like ICA or PCA depended on specific statistical properties of the M-response and SA which might not always hold.

A common thread among these traditional methods is their difficulty in robustly handling significant SA-M-response overlap, their frequent need for manual parameter tuning, and often, their limited generalization across diverse subjects or experimental conditions. These inherent limitations paved the way for exploring more powerful, data-driven solutions like Convolutional Neural Networks (CNNs) for M-response filtering, especially CNN variants like HS-CNN and hierarchical U-Nets [12], [14], [15]. These models offer superior generalization and performance across noisy biomedical signals, particularly for EMG and EEG analysis [1], [11], [13].



Fig. 1. The example of the contaminated and normal responses

II. Dataset and Registering Conditions

A database obtained from real patient data was used as a test dataset. All private patient data was removed in advance, names were replaced with digital identifiers, and only the following information about the data was left: symptom, type of examination, nerve name, muscle name, stimulation point, side of examination, stimulus intensity, latency, amplitude, length, area, impulse propagation speed, distance to the stimulating electrode. Two databases were used to train the model: native and pre-filtered with bandpass filters.

Signal recording conditions. The examinations were performed on an M-TEST ONE 8 electromyograph. M-response recordings were made using disposable RT34 electrodes and a titanium grounding electrode with an area of 8 cm^2. The amplifier sampling frequency was 16 kHz. High-pass filtering was set at 0.5 Hz, and low-pass filtering at 3 kHz. The analysis period was 800 ms. Stimulation was performed with rectangular pulses 0.2-0.5 ms long and 10-100 mA in intensity.

III. Model Description

The U-Net 1D model is an adaptation of the well-known U-Net architecture, originally developed for image segmentation, tailored specifically for one-dimensional data such as time series data or sequential biomedical signals [1], [12]. Like its 2D and 3D counterparts, the U-Net 1D features a distinctive U-shaped, encoder-decoder structure enhanced by skip connections, which play a crucial role in preserving fine-grained details throughout the network.

In the U-Net 1D architecture, the encoder (or contracting path) consists of a series of 1D convolutional layers that progressively extract higher-level features from the input sequence. Each stage typically involves two 1D convolutional operations with nonlinear activations, followed by a downsampling step, such as 1D max pooling. As the data moves deeper into the network, the number of feature channels usually increases, allowing for richer representations.

Mathematically, a single 1D convolution operation at layer l can be described as:

$$\mathbf{h}^{(l)} = f\left(\mathbf{W}^{(l)} * \mathbf{h}^{(l-1)} + \mathbf{b}^{(l)}\right)$$

where:

$\mathbf{h}^{(l-l)}$is the input feature map from the previous layer,

$\mathbf{W}^{(l)}$ is the convolution kernel,

$*$ denotes the 1D convolution operation,

$\mathbf{b}^{(l)}$ is the bias term,

f is a nonlinear activation function (e.g., ReLU).

The downsampling (e.g., max pooling) operation can be expressed as:

$$\mathbf{h}^{(l)}_{\text{pooled}}[i] = \max_{j \in \mathcal{N}(i)} \mathbf{h}^{(l)}[j]$$

where N(i) is the neighborhood of index $ i $ defined by the pooling window.

At the network's bottleneck, additional 1D convolutions are applied without further downsampling, serving as the bridge between the encoder and decoder.

The decoder (or expanding path) mirrors the encoder's structure, employing upsampling operations—such as 1D transposed convolutions or interpolation—followed by further 1D convolutions and activations. The upsampling step can be written as:

$$\mathbf{h}^{(l)}_{\text{up}} = \text{Upsample}(\mathbf{h}^{(l)})$$

or, for transposed convolution:

$$\mathbf{h}^{(l)}_{\text{up}} = f\left(\mathbf{W}^{(l)}_{\text{T}} *^{T} \mathbf{h}^{(l)} + \mathbf{b}^{(l)}\right)$$

where $*^{T}$denotes the transposed convolution operation.

Skip connections link each encoder stage to its corresponding decoder stage by concatenating the feature maps:

$$\mathbf{h}^{(l)}_{\text{concat}} = \text{Concat}\left(\mathbf{h}^{(l)}_{\text{encoder}}, \mathbf{h}^{(l)}_{\text{decoder}}\right)$$

The final output layer typically uses a 1D convolution with a kernel size of one, mapping the extracted features to the desired number of output channels or classes:

$$\mathbf{y} = \sigma\left(\mathbf{W}^{(\text{out})} * \mathbf{h}^{(\text{final})} + \mathbf{b}^{(\text{out})}\right)$$

where σ is an activation function suitable for the task (e.g., softmax for segmentation, linear for regression).

This design makes the U-Net 1D model particularly effective for a variety of sequence labeling tasks, including biomedical signal segmentation (like ECG or EEG), audio or speech segmentation, and time series anomaly detection.

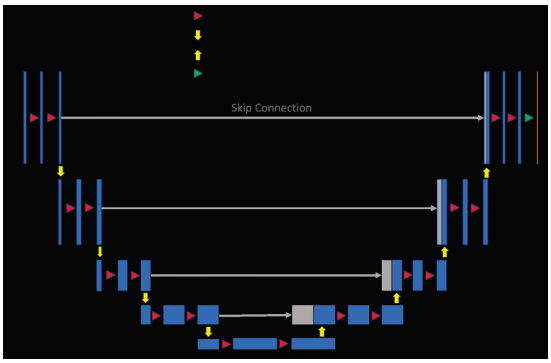


Fig. 2. Schematic representation of UNet1D model architecture.

Overall, the U-Net 1D model retains the strengths of the original U-Net—such as efficient end-to-end training and the ability to perform well even with limited data—while being optimized for sequential data. Its symmetric structure and use of skip connections make it a powerful tool for tasks where both detailed local information and broader contextual understanding are essential. This makes areas like EEG, ECG, and EMG particularly interesting in the context of using this architecture due to its skip connections and efficient handling of local and global features [3], [8].

V. Evaluation of the Obtained Results

UNet1D model was developed and tested to remove noise from M-response signals. The model was trained on data containing various types of artifacts, including stimulus artifact, contour shift, high-frequency noise, and contour drift. The effectiveness of blind SNR estimation techniques and their applicability in biomedical noise filtering tasks is supported by recent studies [10], [17]. Our use of the method of moments ratio aligns with recommended practices for non-Gaussian, artifact-heavy signal domains [5], [7].

Noise Reduction Efficiency. The processing of M-response signals using UNet1D demonstrated excellent results in preserving key diagnostic parameters. An analysis of 632 signals yielded the following findings:

Mean deviation of maximum amplitude: only -0.04%

Median deviation: -0.05%.

The histogram of the distribution of maximum deviations clearly shows that the vast majority of processed signals have minimal deviations from the original, which indicates a high accuracy of preserving diagnostic characteristics. Standard deviation: 5.89%

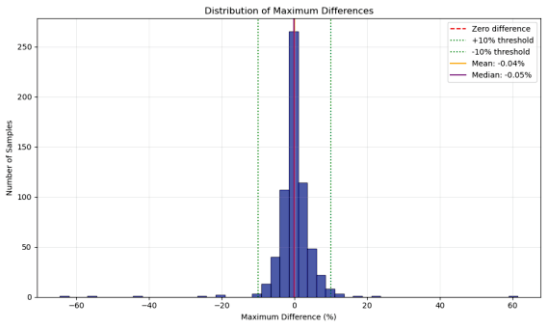


Fig. 3. Histogram of the distribution of maximum deviations

Outlier analysis

Statistical analysis of outliers demonstrated high stability of the model:

Interquartile range (IQR): 3.28%

Only 2.85% of the signals (18 out of 632) exhibited deviations exceeding 10%.

For 97.15% of signals, the average deviation was only 0.07% with a standard deviation of 3.07%

Signal-to-noise ratio (SNR) estimation.

To objectively assess the quality of filtering, the method of moments ratio was used for blind SNR estimation. This method is especially valuable for biomedical signals, since it does not require a priori knowledge of the location of the useful signal in the recording.

The principle of the method is based on the differences in the statistical characteristics of the noise and the useful signal. For Gaussian noise, the ratio of the fourth moment to the square of the second moment is 3. Deviation from this value indicates the presence of a non-Gaussian useful signal [18]

SNR estimation results:

Average SNR of original signals: 1.4508

Average SNR of noisy signals: 1.4294

Average SNR of filtered signals: 1.7259

Average SNR improvement: 0.2966, corresponding to a 20.75% increase.

This SNR estimation method is particularly suitable for M-response signals because [19]:

Allows assessment of the quality of noise suppression without knowing the exact location of the useful signal.

Takes into account the non-stationary nature of biomedical signals.

Effective for both periodic and pulsed noise components, which are often present in electromyography.

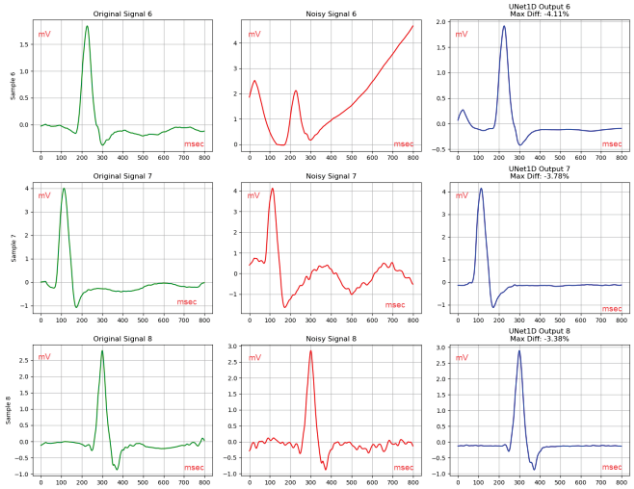


Fig. 4. Model performance

VI. Conclusions

In this study, we developed and evaluated a CNN-based artifact correction method for M-response signals using the UNet1D architecture. The approach was specifically designed to address the limitations of traditional filtering techniques in electromyographic (EMG) diagnostics, particularly for signals affected by overlapping stimulus artifacts and other complex noise patterns [20].

The proposed model was trained on real-world EMG data and demonstrated high precision in preserving diagnostically important amplitude-temporal characteristics while effectively suppressing a wide range of noise types. Quantitative evaluation showed that the model achieved a minimal mean amplitude deviation of -0.04%, and improved the signal-to-noise ratio (SNR) by 20.75%, indicating robust artifact removal and preservation of clinically relevant features [21].

The use of UNet1D allows for efficient end-to-end learning and strong generalization capabilities across varied signal conditions, making it a promising tool for integration into clinical EMG workflows. By reducing dependence on manual parameter tuning and outperforming classical filtering methods, this approach enhances the reliability and diagnostic accuracy of M-response evaluation in neurological diagnostics.

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Method of Measuring Acoustic Resistance by Acousto-Magnetic Method for Application in Medicine

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***Abstract* – Among the problems of diagnostics and treatment of pathologies there are tasks that can be solved using new methods of interaction of the biological substance with physical fields. This paper examines methods and devices for determining the acoustic resistance of local areas of a model biological substance using acoustomagnetic action on it.**

Keywords – acoustic resistance, electrical voltage, electrical conductivity, model sample of biological substance, constant magnetic field.

I. Introduction

The acoustic resistance (AR) of the human biological substance (BS), consisting of soft tissues and bones, is the product of its density and the speed of sound propagation in it. The value of the AR of this substance depends on the physiological state of a person, in particular, on the type of disease. Until now, there was no in vivo method for determining the AR of local BS areas. The researchers used separate measurements of the density and speed of sound in samples of the substance or organ isolated from the organism, followed by calculation of their product. At present, the approximate values of the AR of living substance are determined as a result of ultrasound examinations of the patient by recording the spatial distribution of substance density and multiplying this density by the speed of sound in water. To ensure high accuracy of AR determination, it is necessary to develop a new method that provides direct remote in vivo determination of the acoustic resistance of substance in the required areas of the human body. The aim of the work is to substantiate the possibility of using acoustomagnetic action on the biological substance to determine the AR of its local area. In this paper, the authors for the first time propose and investigate the acoustomagnetic effect on a model sample of a biological substance in order to accurately determine the acoustic resistance of its local area. The main mathematical relationship is obtained and verified, which allows determining the value of the acoustic resistance of a local area of a model sample under acoustomagnetic effect on it.

II. Features of Ultrasound Propagation in a Spatially Limited Substance

Until now, approximate values of the AR of some organs of healthy people and the influence of some pathologies on their AR have been known from ultrasound examinations of the body [1, 2, 3]. The approximation of the AR values is caused, in particular, by the fact that the speed of sound in various tissues is not equal to the speed of sound in water. This reduces the information content of the traditional method for determining AR. As will be noted below, the use of acoustomagnetic action on tissues can significantly increase the accuracy, information content and efficiency of determining local AR values. On the other hand, the proposed method for determining local AR values can improve and supplement the interpretation of acoustic images of pathological tissues and large arteries obtained during traditional ultrasound examinations.

The science of sound, or acoustics, has a centuries-old history. The wave theory of sound was initiated in the works of prominent scientists from different countries: Newton, Huygens, Rayleigh etc [4,5].

Acoustic radiation is formed in the surrounding material environment (except vacuum) mainly with the help of solid-state ultrasonic radiation generators of two types. The first uses the physical phenomenon of magnetostriction, in which ferromagnetic bodies change their dimensions under the influence of an alternating magnetic field. Fluctuations in the dimensions of such bodies are transmitted by sounds to the studied substance, in particular biological. The main parameters of ultrasonic radiation are the frequency f, the intensity P_s, the diameter d (or the cross-section of the sound guide in the form of a circle) of the “beam” of ultrasonic radiation [6] and its directionality in space.

During the oscillation of a flat sound emitter along the direction of propagation of ultrasonic radiation in the substance, a plane wave of compression and expansion of the substance arises. The contour of the ultrasonic radiation beam has the form of a circular cylinder if the emitter has the shape of a circle. For example, at low ultrasound frequencies, provided that the radiation power does not exceed 1 W/cm^2, the displacement of water during the compression-expansion process is units of microns, and the displacement speed is several centimeters per second.

Magnetostrictive generators are usually low-frequency (20–40 kHz) and have an intensity of up to tens of W/cm^2. In this case, the “beam” of ultrasonic radiation has a weak divergence at large distances from the emitter. The absorption of ultrasonic power depends on the properties of the substance. In particular, in water, the absorption is insignificant, which allows ultrasound with a frequency of about 20 kHz to propagate practically without attenuation for many kilometers.

The second type of ultrasound generators is based on a physical phenomenon called the inverse piezoelectric effect. In this case, in materials belonging to ferroelectrics, a change in their dimensions occurs under the action of an electric alternating voltage applied to them. These generators are small-sized, have a much higher efficiency and can be sources of ultrasound of both low and high frequencies (40 kHz - 10 MHz). The intensity of a single generator of this type usually does not exceed 1 W/cm^2. The

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propagation of sound from a sound pipe in contact with the BS will be further determined by the uniformity of the acoustic resistance R_S of the substance.

$$R_s = \rho \cdot V_s \quad (1)$$

where ρ is the density of the substance;

V_S is the speed of sound in the substance.

For example, the speed of sound in water is approximately 1500 m/s, while in air it is only 300 m/s. At the same time, the density of water is 1000 kg/m^3 and exceeds the density of air at normal atmospheric pressure by a thousand times. Accordingly, the acoustic resistance of water is 1.5 x 10^6 kg/(m^{2}s), and air is about 300 kg/(m^{2}s), that is, 5 thousand times less.

In the presence of spatial inhomogeneities in the acoustic resistance of the substance, part of the ultrasonic radiation power is reflected from them and interference of the incident and reflected waves occurs. As a result, a standing sound wave arises in the substance. In the studied volume of the substance, the standing sound wave is established when the condition is met:

$$L_i = n \cdot \frac{\lambda}{2} \quad (2)$$

where L_i is the distance from the sound emitter to the inhomogeneity;

λ is the wavelength of the ultrasonic radiation;

$\lambda = V_S / f$; f - ultrasound frequency;

n = 1, 2, 3 ... - the possible number of half-waves that corresponds to the given values of distance and radiation wavelength in accordance with formula 2.

The magnitude of the reflected power depends on the ratio of the acoustic resistance of the inhomogeneity R_{IS} to the acoustic resistance of the substance R_S. The larger this ratio, the higher the intensity of reflection from the inhomogeneity. In particular, if ultrasound is introduced into a vessel with a liquid substance towards the bottom, the sound will be almost completely reflected from the bottom of the vessel (Fig. 1).

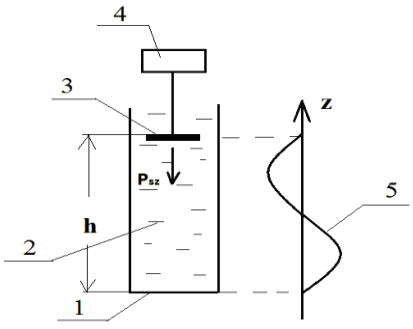


Fig. 1. Scheme of formation of a standing wave of velocity (5) of oscillatory movement of a liquid substance (2) in a section with a height h = λ of a cylindrical vessel with a bottom (1) under the action of acoustic radiation with intensity P_{sz}, created by a flat emitter (3) of a sound generator (4).

A similar situation can occur if ultrasound is introduced into the patient's body from one side. In this case, its reflection will occur from the opposite side (Fig. 2). The fraction of the reflected ultrasound power is determined by the reflection coefficient K_R:

$$K_R = \frac{(R_{s1} - R_{s2})}{(R_{s1} + R_{s2})} \quad (3)$$

where R_{S1} - the AR of the biological substance of the body;

R_{S2} - the AR of the air atmosphere around the patient.

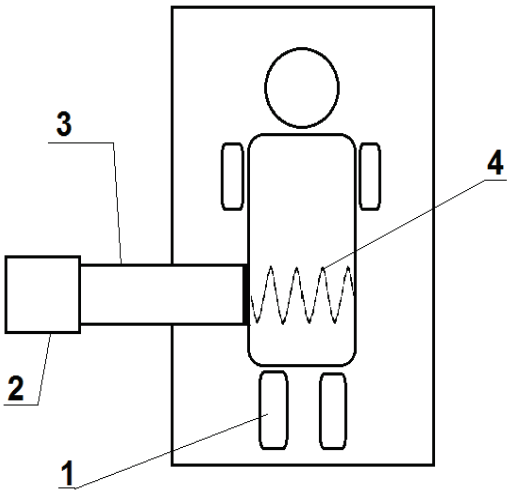


Fig. 2. Scheme of the formation of a standing wave of ultrasonic radiation inside a hypothetical, i.e. absolutely homogeneous human body: 1 – patient; 2 – ultrasound generator; 3 – sound emitter; 4 – standing wave of the velocity of movement (compression-stretching) of a homogeneous biological substance under the action of ultrasound.

If the power of ultrasonic radiation in a liquid substance increases to a certain critical value, the phenomenon of cavitation may occur. This is caused by the compression of the air dissolved in the liquid under the action of an acoustic wave and the formation of a high-pressure bubble in it. The rupture of the bubbles resulting from cavitation is similar to a micro-explosion, which can damage the objects under investigation present in the liquid substance. This imposes a limit on the permissible ultrasound intensity for a long time. In BS, similar in properties to water, this critical intensity at an ultrasound frequency of approximately 20 kHz is approximately 1 W/cm^2. In the case of increasing the ultrasound frequency, the value of the critical intensity increases.

III. THEORY OF REMOTE DETERMINATION OF LOCAL VALUE OF ACOUSTIC RESISTANCE

As shown by many years of biological and medical research, knowledge of acoustic resistance and its distribution in the patient's body plays a key role in deciphering the indicators obtained using acoustic device during ultrasound examination of the human body [7].

Nowadays, ultrasound has become the most widespread remote and safe method of diagnosing pathological foci of many diseases of almost any organs and areas of the human

body. At the same time, ultrasound remains mostly a qualitative and descriptive diagnostic method, little related to the objective and measurable physical and chemical parameters that determine a particular disease. The reasons for this are twofold. First, the acoustic response to changes in the biological substance, recorded by an acoustic device, is primarily associated with the reaction of ultrasound to changes in the density of the substance. In contrast, the acoustic resistance R_s is the product of the density ρ and the speed v_s of sound in the substance.

The acoustic resistance of local areas of the human biological substance can change due to any diseases.

Secondly, the non-uniformity of the biological substance density leads to multiple ultrasound reflections. This complicates the determination and classification of features observed on traditional acoustograms during human ultrasound examination and necessitates the search for new methods for determining AR. Let us consider the principle of acoustomagnetic determination of AR of a local section of a model sample of a biological substance. As we have shown earlier [8,9,10,11], the simultaneous action of ultrasound and a constant magnetic field on a model sample of a biological substance causes an alternating electric current in it. The magnitude and spatial distribution of the current depends on the AR value. Thus, the use of acoustomagnetic action on a biological substance for the first time opens up the possibility of highly accurate determination of the acoustic resistance itself through measurements of the electric current and the corresponding voltage. This becomes possible if we establish a mathematical dependence of the local value of the acoustic resistance on the measured values of the ultrasound intensity, constant magnetic field and electric voltage on the electric measuring probes in contact with the biological substance.

To establish the required dependence, researchers usually use model samples that imitate the real biological substance [12].

Figure 3 shows the developed scheme of the model sample with a biological substance in the form of a physiological solution of NaCl, as well as the distribution of the magnetic field, the speed of the solution under the action of ultrasound and the magnitude of the alternating electric current in the solution with the ultrasound frequency. In this case, the use of a local source of a constant magnetic field in the form of Helmholtz coils (2) or permanent magnets allows us to excite an alternating current in a limited area of the sample. In particular, figure 3 shows that in this case this area has a height equal to half the length of the standing wave of the solution oscillation speed under the action of ultrasound. Since the alternating current excited in the solution flows through the solution with a certain electrical resistance, it can be measured using two electrical probes located inside the solution at a known distance from each other. The location of two probes along the X axis in the model sample of the BS is shown in figure 4. For low-frequency ultrasound and current, the amplitude of the electric voltage U_{mx} on the probes with the base along the X axis is calculated according to Ohm's law. Our calculations showed that the mathematical dependence of the local (over half the standing wave of the ultrasound) value of the AR solution on the parameters of the solution and the measuring setup is given by the formula:

$$R_s = \rho \cdot V_s = \frac{2P_{sz} \cdot B_y{}^2 \cdot L_x{}^2}{U_{mx}{}^2} \qquad (4)$$

where P_{SZ} - ultrasound intensity along the Z axis;

B_Y - the magnitude of the Y-component of the constant magnetic field acting on the solution;

L_x - distance along the X-axis between the measuring electrical probes located in the solution;

U_{mx} - amplitude of the alternating electric voltage on the probes;

x, y, z - axes of the rectangular coordinate system.

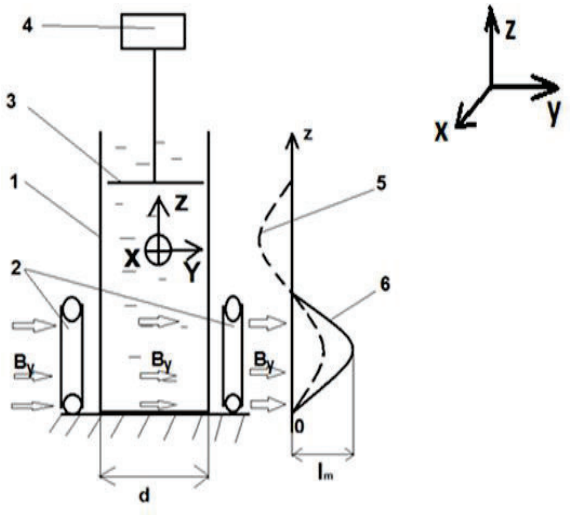


Fig. 3. Alternating current disturbance in a single half-wave section of an ultrasound standing wave; 1 – vessel with a model NaCl solution; 2 – Helmholtz coils (or permanent magnets) for creating a constant magnetic field within half of the ultrasound standing wave; 3 – ultrasound emitter; 4 – ultrasound generator; 5 – solution oscillation velocity distribution V_Z; 6 – current distribution I_X in the solution.

Currently, we are not aware of any publications on the study of AR using low-frequency acoustomagnetic influence, and therefore the experiments described in this work can be recognized as new and relevant. The proposed method and device with a model BS sample can potentially become the basis for new medical equipment for contactless in vivo determination of AR of various local areas of biological tissues. In particular, to improve the localization of the object of study, further work is required to create local magnetic field sources and to use ultrasound sources with a higher frequency and intensity. In our experimental studies, we used ultrasound generator of the USDL-2T type with a frequency of about 22 kHz and an intensity of about 10^{-2} W/cm^2. Directly inside the model biological substance, the voltage U_Z of about 50 μV was measured on probes located at a distance of s = 1 cm, provided that the magnitude of the constant magnetic field was about 0.4 T. In this case, half the length of the standing ultrasound wave in the solution is about 3.5 cm. This value determines the

locality of the proposed method for determining AR. Increasing the ultrasound frequency can improve the spatial resolution of the method. These studies allowed us to verify the formula (4) we obtained for calculating the AR of a model substance in the form of a physiological solution, which is more than 99% water. Verification was carried out by substituting the given experimental parameters into the formula and comparing the calculated AR value with the AR of water known from scientific literature ($1.5x10^6$ kg/m^2s). The calculated value turned out to be equal to $1.3x10^6$ kg/m^2s, which differs from the tabulated value by only 6%. This indicates a good correspondence between the obtained formula (4) and the real dependence of the AR value on the properties of the model biological substance.

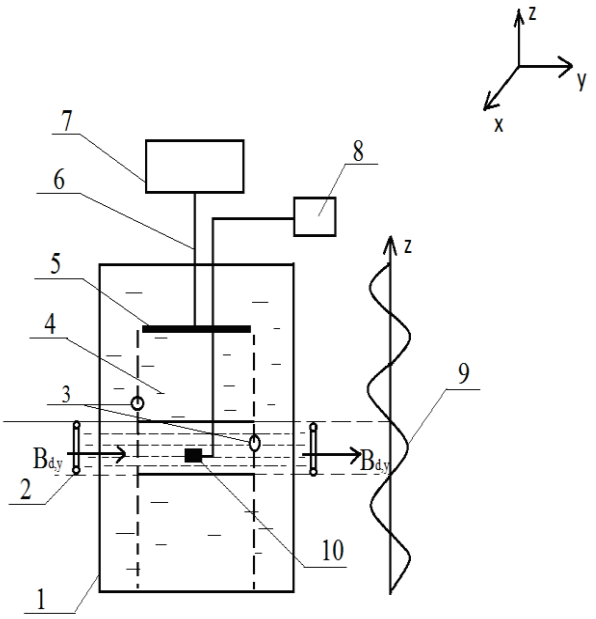


Fig. 4. Functional diagram of a device for perturbing and measuring the electric field strength: 1 – cylindrical vessel; 2 – Helmholtz rings for creating a constant magnetic field with induction $B_{d,y}$ in the solution; 3 – boundaries of the propagation of the ultrasound beam;4 – electrically conductive solution that simulates the biological substance; 5 – flat ultrasound emitter that propagates the ultrasound beam along the vertical axis Z of the vessel; 6 – sound pipe connecting the ultrasound generator (7) with the emitter; 8 – voltmeter; 9 – standing wave with maximum amplitude E_{xm}; 10 – electric probes for measuring the electric field strength E_x in the solution.

Our experimental and computational studies show that promising devices for determining local AR values of living tissues of the body using acoustomagnetic action on it can be divided into two classes. The first one assumes the placement of electric probes for measuring voltage on the surface of the patient's skin, similar to how this is done when registering an electrocardiogram. This class of systems can be called extrametric and assumes the need to measure alternating voltage on the probes at the microvolt level. Its positive properties are complete safety and the absence of trauma to the patient's body. The second class assumes the introduction of electric measuring probes in the form of miniature wire electrodes into the patient's body. This method can be called endometrial. A positive property of this approach is a significant (tens and hundreds of times) increase in the useful voltage on the probes.

IV. Conclusions

Acoustomagnetic action on an electrically conductive substance results in the generation of an alternating electric current in it, the magnitude and spatial distribution of which depend on the acoustic resistance of the substance. The existence of such a current has been experimentally confirmed by the generation of electric voltage on measuring probes located inside a model sample of the substance. For the first time, a relationship has been obtained and verified linking voltage with the value of acoustic resistance of a local section of a model sample of a biological substance. The existence of a known dependence of the acoustic resistance of a human biological substance on its pathologies and the unique possibility of determining the value of acoustic resistance of a model biological substance using low-frequency acoustomagnetic action can become the basis for creating highly effective diagnostic equipment of a new type for remote detection of pathologies in local sections of the human body.

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Managing the Energy Demands of AI in Fragile Renewable-Based Power Systems

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***Abstract*—This study addresses a growing tension in the digital sustainability discourse by quantifying the mismatch between AI-driven electricity demand and renewable energy supply. In response to recent calls for empirical studies, we introduce a dual-axis mismatch framework that structurally compares (1) the share of renewable electricity in global generation and (2) the share of that renewable output consumed by AI. Drawing on a quantitative modeling approach and authoritative global data, we model electricity trends through 2030 under both Announced Policies and Net Zero scenarios. The results reveal that AI electricity demand is growing at nearly twice the rate of renewable generation, with AI potentially consuming up to 9% of global renewable electricity by 2030. To assess vulnerability, the study incorporates the April 2025 Iberian blackout as a case of low-inertia system failure under inflexible demand. Building on these findings, we propose the AI–energy alignment framework to guide strategic responses across technical, operational, planning, and governance domains. The study contributes a structural tool for anticipating digitally driven pressure on clean energy systems and supports resilience-focused energy management and infrastructure policy.**

Index Terms*—*sustainability, energy transitions, AI energy demand, renewable power systems, resilience, energy systems management.

I. Introduction

In recent years, artificial intelligence (AI) has emerged not only as a key enabler of energy efficiency and grid optimization, but also as a rapidly growing source of electricity demand. As AI adoption accelerates, particularly through large-scale data centers, real-time learning systems, and generative models, the sector's electricity requirements are projected to rise exponentially. This growth introduces a new layer of pressure on energy systems that are simultaneously attempting to decarbonize and digitize [1], [4], [13], [17].

At the same time, renewable-based power systems, central to the global energy transition, face well-documented challenges related to intermittency, low inertia, and flexibility [3], [11]. While renewables reduce emissions, their integration can compromise system stability when not matched with adequate storage, balancing services, or flexible loads. As AI workloads tend to be always-on and geographically concentrated, they increase baseline demand and reduce the system's ability to buffer volatility. If unmanaged, this interaction may lead to systemic instability, including blackout risks [7], [13].

Despite growing awareness of AI's energy footprint, empirical analyses that quantify its impact relative to renewable electricity supply remain scarce. Even fewer studies link these dynamics to real-world system vulnerabilities or offer concrete frameworks for aligning digital expansion with energy resilience [12], [13], [17].

To address this gap, the study draws on empirical data from Statista [2], IRENA [3], and the IEA [4] to model the trajectory of AI electricity demand and renewable electricity generation from 2022 to 2030. We examine two global development paths: IEA's Announced Policies Scenario and Net Zero by 2050 Scenario, to assess the pace and scale of this mismatch. The central question guiding our analysis is how the growing electricity demand from AI workloads can be quantified in relation to renewable energy supply, and what strategic responses are required to safeguard the resilience of renewable-based electricity systems.

The paper proceeds in four steps: it first models AI electricity demand and renewable generation growth through 2030; second, calculates the mismatch using scenario-based projections and compound annual growth rates (CAGR); third, assesses system vulnerabilities linked to digital load pressures; and finally, introduces the AI–Energy Alignment Framework, a strategic tool for coordinating technical, operational, and policy-level responses to digitally induced energy stress.

II. Theory And Research Question

The intersection of AI and renewable energy has generated increasing research interest, particularly in the context of digital-led sustainability transitions. A recent Scopus search reveals that more than 90 peer-reviewed papers were published in Business and Management journals in 2025 alone, specifically addressing AI's role in energy, infrastructure, and sustainability contexts. This surge reflects a growing recognition that digitalization is not only shaping how energy systems are managed, but also redefining their structural constraints and vulnerabilities.

The International Energy Agency (IEA) frames AI as a transformative force with the potential to enhance energy system flexibility, improve demand forecasting, and support the integration of variable renewables [1]. In line with this, recent studies have demonstrated how AI technologies can optimize grid operation and renewable deployment, especially in forecasting and system control domains [14], [16].

Beyond its enabling role, AI is also emerging as a source of growing electricity demand, with widespread implications for energy system balance. This "energy hunger paradox," as defined by Senyapar and Bayindir [13], captures the dual nature of AI as both a sustainability enabler and an energy-intensive 'consumer'. According to the IEA [4], global electricity demand from AI workloads is projected to rise sharply by 2030, driven by the rapid growth of generative models, data centers, and edge computing infrastructure. Yet, despite these trends, few studies quantify this demand in relation to projected clean energy supply or examine its implications for system resilience.

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Existing contributions to AI and energy research tend to focus on AI for energy optimization [15], AI patent activity and policy interaction [16], or green innovation pathways [12]. Raza and Shakeel's text-mining analysis [18] confirms that electricity consumption and low-carbon energy transitions are central to current debates, yet does not disaggregate AI's direct impact. Similarly, broader policy outlooks [8], [9] and scenario-based modeling efforts [4], [3] emphasize decarbonization targets without accounting for emerging digital loads.

These gaps persist even in recent policy outlooks that prioritize electrification, green hydrogen, and renewable expansion, giving at the same time limited attention to the rising impact of digital infrastructure on grid fragility [10]. While some frameworks emphasize structural transformation in support of renewables [11], [17], [25], they do not account for the surge in AI-driven demand and primarily treat it as a solution, not as a source of additional system pressure.

The systemic dimension of this imbalance becomes increasingly visible when viewed through the lens of infrastructure coordination and planning. Industry analysis from S&P Global [7] notes that data center deployment is accelerating faster than power system development, with facilities reaching 1 GW scale and often being constructed ahead of confirmed grid capacity. These trends, combined with permitting delays, limited demand flexibility, and fragmented policy frameworks, raise critical questions about how resilient renewable-based systems will be under growing digital load pressure.

Addressing this emerging challenge, our paper responds directly to recent calls for empirical research that quantifies AI's energy implications and system-level risks in the context of global sustainability targets [13]. Whereas this work has focused on conceptual trade-offs and policy alignment, our analysis quantitatively assesses the mismatch between projected AI electricity demand and renewable energy generation capacity. We extend the debate from theoretical sustainability alignment toward concrete system resilience concerns by modeling this energy imbalance and evaluating its implications for grid stability, including blackout risks. By answering the research question "How does the growing AI electricity demand challenge the resilience of renewable-based power systems, and what managerial responses can guide strategic alignment?", we also contribute a novel dual-axis mismatch framework that can inform interdisciplinary energy planning at the intersection of AI growth and clean energy transitions.

III. Methodology

Our study adopts a mixed-method approach that combines a narrative review of emerging interdisciplinary literature with scenario-based quantitative modeling. The theoretical background is developed through a narrative review aligned with the principles of emerging topics research. Relevant literature on digital energy demand, renewable grid fragility, and infrastructure resilience was examined to identify gaps in existing frameworks. This review helped us to conceptualize the dual-axis mismatch framework, which guides the empirical components of the study.

The empirical modelling was based on publicly available datasets from the IEA [1], Statista [2], and the International Renewable Energy Agency (IRENA) [3]. We model electricity demand from AI workloads and electricity generation from renewables for the period 2022–2030, under both the IEA's Announced Policies Scenario (APS) and Net Zero by 2050 Scenario. Projections are based on reported capacity data, assumed technology-specific capacity factors, and standard energy modeling equations.

Renewable electricity generation is estimated by converting capacity figures (in gigawatts, GW) into energy output (in terawatt-hours, TWh) using the standard energy generation formula [4]:

$$\text{TWh} = \text{GW} \times \text{Capacity Factor} \times \text{Hours/Year} \div 1{,}000, \quad (1)$$

GW - installed capacity in GW; Capacity Factor - share of time the plant operates at full output (varies by technology); Hours/Year = 8,760.

Electricity consumption for AI by data centers is determined based on projected electricity use volumes and regional distribution of data center infrastructure [21], [22].

To evaluate the intensity of change over time, we apply the Compound Annual Growth Rate (CAGR) formula:

$$\text{CAGR} = \left(\frac{\text{Final Value}}{\text{Initial Value}}\right)^{\frac{1}{n}} - 1 , \quad (2)$$

n - number of years.

To assess systemic vulnerabilities, we incorporate the April 2025 Iberian blackout [5], [6] as a real-world case study illustrating grid vulnerability under high renewable penetration and inflexible digital loads. This is supported by industry insights from S&P Global on data center growth patterns [7], permitting bottlenecks, and infrastructure coordination failures.

IV. Findings and Implications

To answer the research question, we begin by modeling the global electricity demand from AI workloads and comparing it to both total global electricity generation and renewable electricity output. Table I presents the trajectory of AI electricity demand from 2022 to 2030, and Table II outlines the growth in global installed renewable capacity. Using this capacity data and technology-specific capacity factors, we estimate renewable electricity generation through 2030 in Table III.

TABLE I
AI Energy Demand (2022–2030)

Year	Global AI Electricity Demand (TWh)	% of Global Demand	Notes
2022	~460	~2.0	Baseline year, includes general data center use
2026	~1000	~3.5 - 4.0	Projected doubling, driven by generative AI, cloud, and crypto
2030	~1,500-2,000	~4.5 - 6.0	Continued exponential growth, mostly from AI workloads

Source: Compiled by the authors based on [1], [2].

TABLE II
Renewable Energy Capacity Growth (2022–2030)

Year	Global Installed Renewable Capacity (GW)	Annual Additions (average, GW)	Cumulative growth
2022	~3,396	~318 (actual)	Baseline year [3]
2024	~4,300	~452	~27% increase from 2022 [1]
2026	~5,800	~750	~71% increase from 2022 [1]
2030	~7,902	~525	~133% increase from 2022[4]

Source: Compiled by the authors based on [1], [2], [3], [4].

Following the trajectories in Tables I–III, we observe two concurrent but diverging trends. While AI electricity demand is projected to more than quadruple by 2030, renewable electricity generation is expected to grow by only about 2.5 times. Despite improvements in capacity and efficiency, clean energy supply is not keeping pace with accelerating AI-driven demand. This

imbalance suggests a potential structural paradox within the energy transition.

TABLE III
ESTIMATED RENEWABLE ELECTRICITY GENERATION BASED ON CAPACITY (2022–2030)

Year	Global Installed Renewable Capacity (GW)	Assumed Average Capacity Factor**, %	Estimated Generation (TWH)	Calculation*
2022	~3,396	32	~9,510	3,396 × 0.32 × 8,760 /1,000
2024	~4,300	33	~12,460	4,300 × 0.33 × 8,760 / 1,000
2026	~5,800	34	~17,320	5,800 × 0.34 × 8,760 / 1,000
2030	~7,902 / ~7,442	35	~24,228/ ~22,817***	7,902 × 0.35 × 8,760 / 1,000/ 22,817/[0.35 × 8,760 / 1,000]

Source: Compiled by the authors based on [20].

Notes: *based on the standard energy generation formula (1);
**based on (IEA, IRENA) the capacity factors increase slightly to reflect improved efficiency, digital controls, and wider deployment in high-resource areas: wind (onshore/offshore avg) - 30-40%; solar PV (utility-scale) 15-25%; hydro (global average) - 40-50%; biomass/geothermal - 60-80%; global average (combined)~30-35%.
***the target is based on the IEA [19].

To assess the depth of this divergence, we compare average growth rates for AI electricity demand and renewable electricity generation using the compound annual growth rate (CAGR) method ((2), Table IV).

TABLE IV
IDENTIFYING THE MISMATCH BETWEEN AI ENERGY DEMAND AND ESTIMATED RENEWABLE ELECTRICITY GENERATION (2022–2030)

Metric	2022	2030	CAGR (2022–2030)
AI Electricity Demand (TWh)	~460	~2,000	~20.3%
Renewable Electricity (TWh): Announced Policies Scenario (APS)	~9,510	~24,228	~12.4%
Renewable Electricity (TWh): “Net Zero” projection	~9,510	~22,817	~11.6%

AI electricity demand is growing nearly twice as fast as renewable electricity generation. Even under the IEA Net Zero by 2050 scenario, which assumes aggressive renewable expansion, the pace of AI growth outstrips clean energy supply. This confirms a strategic mismatch: if left unmanaged, AI/data center electricity consumption could erode the sustainability benefits of renewable deployment.

To illustrate the implications of this imbalance, we introduce the dual-axis mismatch framework as a structural approach to evaluating energy system fragility in the face of sector-specific demand surges. The framework plots two interdependent dimensions:

(1) the share of renewable electricity in total generation (a measure of systemic progress),

(2) the share of renewable electricity consumed by a single sector, in this case AI (a measure of sectoral load pressure).

This dual representation allows for the assessment of whether a sustainability transition is not only progressing in absolute terms, but also whether it is being undermined internally by high-growth, high-consumption sectors.

Figure 1 compares these two dynamics: the share of renewables in global electricity generation (left axis) and the share of those renewables that would be consumed by AI workloads alone (right axis). While renewables may supply over 60% of electricity by 2030, AI could absorb up to 9% of this clean energy, effectively internalizing a growing pressure within the energy transition itself (Fig.1).

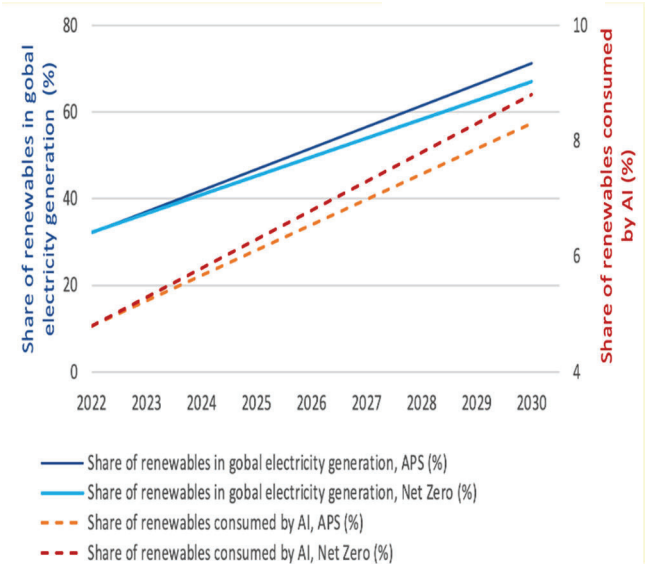


Fig.1. Projected mismatch of renewable supply and AI demand, 2022–2030 (authors’ estimations)

To further explore the spatial dimensions of this imbalance, we disaggregate projections by region, focusing on the United States (USA), China, and the European Union (EU) – three of the largest AI infrastructure hubs. The analysis was conducted based on the calculations of the expected renewable electricity production by regions of data center location (Table V) and scenarios for forecasts of electricity consumption by these centers (Table VI). This geographic clustering of AI-driven demand (Tables V–VI) reveals even more pronounced mismatches between digital load growth and renewable electricity availability.

TABLE V
ESTIMATED RENEWABLE ELECTRICITY GENERATION BASED ON CAPACITY BY REGION OF DATA CENTER LOCATION (2023–2030)

Year	Global Installed Renewable Capacity (GW)	Assumed Average Capacity Factor**, %	Estimated Generation (TWh)	Calculation*
2023				
USA	0,385 [1, c. 5]	32,5	~1,096	0,385×0.325×8,760/1,000
China	1,454 [1, c. 3]	32,5	~4,140	1,454×0.325×8,760/1,000
EU	0,631 [1, c. 4]	32,5	~1,796	0,631×0.325×8,760/1,000
2030				
USA	0,938 [2]	35,0	~2,876	0,938×0.35×8,760/1,000
China	2,461 [2]	35,0	~7,545	2,461×0.35×8,760/1,000
EU	1,236 [2]	35,0	~3,790	1,236×0.35×8,760/1,000

Source: Compiled by the authors based on [21], [22].

Notes: *to convert installed capacity (GW) into actual energy generation (TWh) for renewables, we used a standard industry approach based on capacity factors (1) [4].
**based on Table III.

While the mismatch between AI electricity demand and renewable electricity capacity presents a long-term sustainability challenge, it also introduces more immediate risks to system resilience, particularly the rising likelihood of blackouts. As AI workloads concentrate in energy-intensive data centers that operate continuously and with limited flexibility, they contribute to sharp increases in baseline demand. In parallel, the growing share of renewables (especially wind and solar) reduces grid inertia and weakens the system’s ability to respond to fluctuations or disturbances. Without sufficient storage, flexible backup, or smart demand management, these dynamics can lead to cascading failures. A stark illustration of this occurred on April 28, 2025, when Spain and Portugal experienced one of Europe's largest power outages. In under five

seconds, Spain lost approximately 15 gigawatts of generation capacity (about 60% of its total supply) [5].

TABLE VI
IDENTIFYING THE MISMATCH BETWEEN AI ENERGY DEMAND AND ESTIMATED RENEWABLE ELECTRICITY GENERATION BY REGION OF DATA CENTER LOCATION (2023–2030)

Metric	2023	2030	CAGR (2023–2030)
USA			
AI Electricity Demand (TWh)	~148 [1, p.35]	~388 [2, p.64]	~14.8%
		~860*	~28.6%
		~1,050 [1, p.22-24]	~32.3%
		~1,050 [1, p.22-24]	~32.3%
Renewable Electricity (TWh)	~1,096	~2,876	~14.8%
China			
AI Electricity Demand (TWh)	~90 [1, p.35]	~265 [2, p.64]	~16.7%
		~588*	~30.7%
		~1,000 [1, p.27]	~41.1%
Renewable Electricity (TWh)	~4,140	~7,545	~9.0%
EU			
AI Electricity Demand (TWh)	~68 [1, p.35]	~113[2, p. 64]	~7.5%
		~252*	~20.5%
		~265 [1, p. 24]	~21.4%
Renewable Electricity (TWh)	~1,796	~3,790	~11.3%

Source: Compiled by the authors based on [23], [24].

Notes *to calculate electricity consumption by data centers, we used data on the total demand for electricity from AI in 2030 (~2,000 TWh (Table IV)) and a forecast of the structure of electricity consumption for AI data centers in the United States, China, and the EU in 2030 ([24, p. 64]):

2,000×0.43=866 TWh (USA).

2,000×0.294=588 TWh (China).

2,000×0.126=252 TWh (EU).

The blackout disrupted transportation, communications, and essential services, affecting nearly 60 million people. Preliminary investigations suggest that the incident was triggered by abrupt disconnections in generation, possibly linked to the volatile nature of renewable energy sources and the lack of mechanical inertia in the grid [6]. This case has highlighted the need for grid stabilizers, increased storage, and better international connections to prevent such occurrences in the future. The convergence of inflexible digital loads and fragile clean energy grids thus poses not only an environmental paradox but also a critical reliability risk for electricity systems worldwide.

Recent industry analysis [7] further supports this view, revealing that speculative AI infrastructure development is outpacing the expansion of underlying power systems. According to S&P Global data, AI-driven data centers now demand up to one gigawatt per campus, ten times the scale of previous deployments, yet permitting timelines for grid connections can lag by several years. These facilities, often built ahead of confirmed power availability, place additional stress on already low-inertia grids and complicate load forecasting. Moreover, while AI data centers may offer flexible load profiles in theory, existing business models rarely prioritize energy adaptability. Without stronger coordination between data center developers and power infrastructure operators, such unchecked digital growth may undermine grid stability, particularly in regions already struggling to integrate high shares of variable renewables.

V. DISCUSSION, CONCLUSIONS, AND LIMITATIONS

Building on the empirical findings and the blackout risk case, we propose the AI–energy alignment framework to structure multi-level strategic responses to sector-driven pressure on clean energy systems. It links observed vulnerabilities (mismatch between AI demand and renewable supply, geographic clustering of AI infrastructure, inflexible AI workloads, blackout risk from high volatility and low inertia**,** disjointed policy between AI and energy system) with strategic entry points for mitigation across technical, operational, planning, engineering, regulatory and governance levels (Table VII).

TABLE VII
STRATEGIC RESPONSES TO MANAGE THE RISING AI ELECTRICITY DEMAND IN FRAGILE RENEWABLE-BASED POWER SYSTEMS

Systemic Challenge	Empirical Insight	Strategic Response	Level of Intervention
Mismatch AI demand vs. renewable supply	AI demand grows at ~20% CAGR vs. ~12% for renewables; AI may consume up to 9% of renewable electricity	Energy-aware AI infrastructure design (e.g., efficiency standards, carbon-aware scheduling)	Technical / Corporate[1]
Geographic clustering of AI infrastructure	Data center concentration amplifies local fragility and transmission stress	Location-based planning incentives; zoning AI infrastructure near grid-stable or dispatchable sources	Planning / Regulatory[2]
Inflexible AI workloads (baseline load pressure)	AI data centers operate 24/7, increasing baseline load and reducing flexibility	Workload deferral strategies; integration with demand-side response mechanisms	Operational / Digital[3]
Blackout risk from high volatility + low inertia	Iberian blackout illustrates instability under high renewables + inflexible demand	Grid-forming inverters, hybrid energy storage, synthetic inertia, localized balancing units	Engineering / Grid Planning[4]
Speculative growth and infrastructure lag	AI data centers are rapidly scaling to >1 GW campuses, often before grid capacity is confirmed	Require payment for grid reservation, reform permitting, and prioritize partnerships with established providers	Regulatory / Market Design[5]
Disjointed policy between AI and energy systems	Current policies treat digital and energy sectors in silos	Integrated digital-energy policy frameworks; scenario-based digital growth planning aligned with capacity	Governance / Strategic Foresight[6]

Source: Compiled by the authors.

Notes:
[1] Engineering and design solutions managed at the firm level.
[2]Spatial planning, permitting, and regional development policy.
[3]Real-time load control and software-based flexibility.
[4]Physical infrastructure and system architecture upgrades.
[5]Rules for grid access, pricing, and collaborative incentives.
[6]Long-term, cross-sector coordination and policy integration.

These responses address critical gaps in prior research, which has highlighted AI's value for grid forecasting and optimization [14], [15], but has largely overlooked its growing electricity demand and its role in loading renewable power systems. Following Senyapar and Bayindir's [13] call for an empirical understanding of the energy footprint of AI, we model the rate at which AI-driven electricity demand outpaces renewable supply and present this tension as a structural paradox. In doing so, we introduce a dual-axis mismatch

framework that links energy system progress with internal pressures from digital expansion. Our contribution reframes AI not just as a sustainability enabler, as emphasized in [16], but as an accelerating, and often unmanaged, consumer of clean energy. This dual role has received limited empirical attention, particularly regarding blackout risks and infrastructure misalignment. By segmenting demand across the U.S., China, and the EU, we also highlight regional differences that extend current policy-focused studies.

Our AI-energy alignment framework provides a structured tool for managerial decision-making in policy, utility planning, and digital infrastructure development. By matching system-level challenges with targeted responses at six levels of intervention, it enables cross-sector coordination and supports efforts to align AI-driven growth with energy resilience under tightening sustainability targets.

We acknowledge several limitations in this study, including the use of extrapolated data and scenario-based assumptions from international sources, which may not fully reflect the pace of change in AI hardware efficiency or shifts in regional policy. Future research could build on this work by incorporating real-time load profiles, lifecycle emissions of AI models, or bottom-up assessments of demand flexibility.

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Impact of Grid Connection Point Constraints on Microgrid Energy Storage System Efficiency

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***Abstract* — Due to the instability of renewable energy generation, ensuring the stable operation of microgrids is an actual and at the same time a difficult task. Energy storage systems are able to guarantee an uninterrupted power supply to such microgrids, but their integration is associated with challenges, in particular, the need for the correct selection of technical parameters of the energy storage system. This study presents a method that allows optimizing the capacity of the energy storage system, taking into account various power constraints at the point of connection to the power grid. The main goal is to determine the impact of constraints on the selection of the optimal capacity of an energy storage system. The scenario of microgrid operation is considered, focusing on maximizing the use of solar generation and interaction with the united power grid. The microgrid is explored on the example of a prosumer, which includes a solar photovoltaic system, a load profile, an energy storage system and a constrain of power grid connection point with a time-of-use tariff. To assess the effectiveness of the selected strategies, an analysis of indicators during the year was carried out, which made it possible to determine the payback period of the project.**

Keywords — microgrid, grid connection, prosumer, energy storage systems, renewable energy sources, optimization, energy output

I. Introduction

Modern energy grids are undergoing a process of decentralization, which is associated with an increase in the share of renewable energy sources (RES). In this context, microgrids (MG) are becoming a key element of the new energy system [1]. Such networks are low-voltage, in which the basis of generation is RES. Consumers in MG can be both with fixed and variable load profiles, depending on the type. The main balancing component of MG is the energy storage system (ESS) [2]. The use of ESS allows to compensate for fluctuations in energy generation and consumption, ensuring the stable operation of the system. In addition, by installing ESS, opportunities open up for earnings with time-of-use zone tariffication and the sale of excess energy to the power grid, which allows to receive additional profit [3].

The analysis of scientific publications indicates the presence of a significant number of works devoted to the development of MG control algorithms based on modern power electronics [4]. Such approaches consider factors related to the operation and maintenance (O&M) of electricity generation, purchase or sale of energy, as well as possible impact on the environment. In paper [5], research is focused on the optimal selection of ESS capacity through generation from own RES sources. Demand response and dynamic pricing mechanisms are the main factors in choosing the optimal capacity of the ESS, considered in the works [6], [7]. However, all these approaches do not take into account the technical constraints of power at the grid connection point (GCP), which can significantly affect the economic feasibility of the system.

This work intends to develop a strategy for determining the impact of technical constraints at the connection point on selecting the optimal ESS capacity for the MG of prosumer, taking into account its load. The main objectives are to maximize the profit from power grid interaction and to efficiently use ESS, taking into account the part of the load that is not covered by RES. In contrast to previous studies [8], the proposed approach focuses on the full use of the potential of the ESS in the conditions of capacity constraints at the point of connection to the power grid from the time-of-use (TOU) tariff. Also, within the study, the economic efficiency of the proposed solution is evaluated through the payback period.

II. System Model

The prosumer is connected to the power grid through a connection point. Its microgrid consists of a rooftop photovoltaic panels with a capacity of 8 kW, ESS with a smart controller and a load profile of a household consumer described in the paper [9]. The general model of the prosumer MG is shown in Fig.1.

III. Constraints of Power Grid Connection

The constraints of power of grid connection (P_{con}) for private households depend on the contract between the consumer and the energy service company. Based on their own consumption, the consumer can request a power of connection [10] which will satisfy power of the load and

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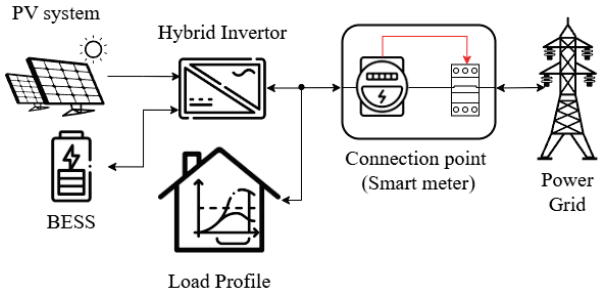


Fig. 1. Single-line diagram of the developed microgrid

RES generation. That is, the permitted power set by the project documentation must exceed the value of the maximum generation or load power according to switch-protection devices, shown in Table.1, we apply the "Distribution Systems Code" of Ukraine in accordance with the power of connection to the grid [11].

TABLE I. VALUE OF THE RATED RANGE OF PROTECTION SYSTEMS

Connection type	**Single-phase, kW**	**Three-phase (triangle), kW**	**Three-phase (star), kW**
Voltage U, V / ***Switch Breaker, A***	***220***	***380***	***220***
10	2.2	11.4	6.6
16	3.5	18.2	10.6
20	4.4	22.8	13.2
25	5.5	28.5	16.5
32	7.0	36.5	21.1
40	8.8	45.6	26.4
50	11	57	33
63	13.9	71.8	41.6

IV. MATHEMATICAL MODEL OF THE MICROGRID

This section provides a mathematical simulation of the objective function and its constraints of operation and power flow.

A. Objective function

The objective function presented in (1) intends at maximizing the profit received from the sale of electricity to the grid. It consists of the cost of the sale of electricity to the grid (3), of purchasing electricity from the grid (4) and the cost of using the capacity of the ESS unit (5).

$$\max_{C} \sum_{i=0}^{N} \left(C_{sell}(t_i) - C_{buy}(t_i) - C_{main}(t_i) \right) \tag{1}$$

$$N = T/\Delta T \tag{2}$$

where T is the total duration of the simulation period of system operation; i is the index of time period into which the test interval is divided; N is the number of time periods; $t_i = i \cdot \Delta T$ is the moment of the beginning of the i-th period of time; $C_{sell}(t_i)$ is the total cost of energy sold to the grid for the period t_i, ₴; $C_{buy}(t_i)$ is the total cost of energy purchased from the grid for the period t_i, ₴; $C_{main}(t_i)$ is the total cost of operating the ESS for the period t_i, ₴.

$$C_{sell}(t) = P_{sell}(t) \cdot K_{encost}(t) \cdot E_{tariff} \cdot \Delta T \tag{3}$$

where $P_{sell}(t)$ is power sold to the grid in a period of time t, kW; $K_{encost}(t)$ is the coefficient of the electricity TOU tariff as a function of time t; E_{tariff} is the electricity tariff, *₴/kWh*.

$$C_{buy}(t) = P_{buy}(t) \cdot K_{encost}(t) \cdot E_{tariff} \cdot \Delta T \tag{4}$$

where $P_{buy}(t)$ is the power purchased from the grid in a period of time t, kW:

$$C_{main}(t) = BC \cdot K_{BC} \cdot \Delta T \tag{5}$$

where BC is capacity of ESS, kWh; K_{BC} is the cost of using a unit of ESS, *$/(kWh·h)*.

$$K_{BC} = \frac{Cost_{batt} \cdot 10^3}{U_b \cdot I_b \cdot h_b} \tag{6}$$

where $Cost_{batt}$ is the cost of one cell of ESS, ₴; h_b is the estimated ESS life-time period at 10 years, corresponding to 3500 cycles of charge/discharge, h; U_b is ESS voltage, V; I_b ESS capacity, Ah.

B. Microgrid Power Balance Equation

The installed MG satisfies the load of the household building and exports electricity to the grid. The power balance equation at each point in time t is given by the formula (7).

$$P_{grid}(t) = P_{gen}(t) - P_{load}(t) + P_{BESS}(t) \tag{7}$$

where $P_{gen}(t)$ is the power of generation in the time period t, kW; $P_{load}(t)$ is the power of load in time period t, kW; $P_{BESS}(t)$ is the power of charge or discharge of the ESS in the time period t, kW.

$$P_{BESS}(t) = \begin{cases} -P_{charge}(t), \text{ if charge} \\ P_{discharge}(t), \text{ if discharge} \end{cases} \tag{8}$$

where $P_{discharge}(t)$ is the power of ESS discharge in the time period t, kW; $P_{charge}(t)$ is the power of ESS charge in the time period t, kW.

C. Technical Constraint of the Microgrid Model

During modelling MG, it is necessary to consider the constraints of the ESS operation. The system must not exceed the charging (9) or discharging power limit of ESS (10).

$$0 \le P_{charge} \le P_{charge.max} \quad (9)$$

$$0 \le P_{discharge} \le P_{discharge.max} \quad (10)$$

The maximum charging power of the ESS is in equation (11).

$$P_{charge.max}(SOC) = k_{ST_char}(SOC) \cdot P_{nom.\text{char}} \quad (11)$$

The rated power of charging ESS is defined as:

$$P_{nom_charge} = k_{charge} \cdot BC \quad (12)$$

where k_{charge} is the rated charging power factor, showing the ratio of the rated charging power to the ESS capacity, determined by the technical characteristics of the ESS in this study $k_{charge} = 1, \left[h^{-1}\right]$.

When modelling, it is also necessary to take into account the constraints on the state of charge of the ESS.

$$SOC_{\min} \le SOC \le SOC_{\max} \quad (13)$$

where SOC is the state of charge of the ESS in relative units; SOC_{min} is the minimum allowable level state of charge of the ESS; SOC_{max} is the maximum allowable level state of charge of the ESS.

The limit values of the state of charge of the battery $SOC_{\min} = 0.2$ and $SOC_{\max} = 1$.

The factor of actual charging power depends on the state of charge of the ESS, and is determined using the formula (13). The values of the parameters (13) in general are given in Table. 2. The form *of* k_{ST_charge} dependence on SOC is shown in Fig.2 [12].

$$k_{ST}(SOC) = \begin{cases} a_{11}SOC^3 + a_{12}SOC^2 + a_{13}SOC + a_{14}, \text{ if } b_{11} \le SOC \le b_{12} \\ a_{21}SOC^3 + a_{22}SOC^2 + a_{23}SOC + a_{24}, \text{ if } b_{21} \le SOC \le b_{22} \\ a_{31}SOC^3 + a_{32}SOC^2 + a_{33}SOC + a_{34}, \text{ if } b_{31} \le SOC \le b_{32} \end{cases} \quad (14)$$

TABLE II. FACTOR OF ACTUAL CHARGING POWER PARAMETERS

var	*value*	*var*	*value*	*var*	*value*
a_{11}	0	a_{21}	0	a_{31}	-24,404
a_{12}	-6,857	a_{22}	0	a_{32}	48,222
a_{13}	2,891	a_{23}	0	a_{33}	-31,769
a_{14}	0,698	a_{24}	1	a_{34}	7,963
b_{11}	0	b_{21}	0,2	b_{31}	0,6
b_{12}	0,2	b_{22}	0,6	b_{32}	1,0

The maximum power of discharge of the ESS battery is determined by equation (14).

$$P_{discharge.max}(SOC) = P_{ST}(SOC) \cdot P_{nom_discharge} \quad (15)$$

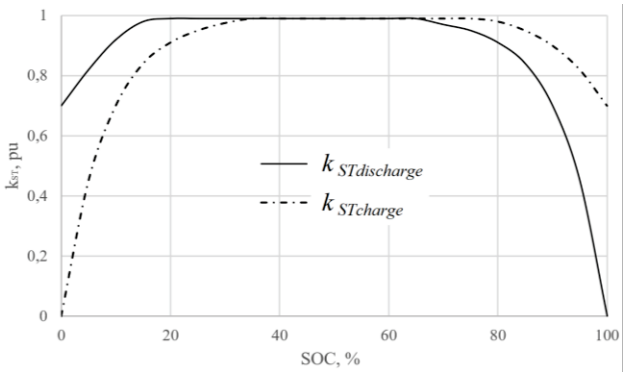


Fig. 2. Curves of charge/discharge

The rated discharge power of the ESS is defined as:

$$P_{nom_discharge} = k_{disch} \cdot BC \quad (16)$$

where k_{disch} is the rated discharging power factor, showing the ratio of the rated discharge power to the battery capacity, determined by the technical characteristics of the ESS, in this study $k_{disch} = 1, \left[h^{-1}\right]$.

The actual discharging power factor, which depends on the SOC of the ESS, is determined using the formula (13). The values of the parameters (13) for the general case are given in Table 3. The curve *of* $k_{ST_discharge}$ dependence on SOC is shown in Fig.2.

TABLE III. ACTUAL DISCHARGE POWER FACTOR PARAMETERS

var	*value*	*var*	*value*	*var*	*value*
a_{11}	33,771	a_{21}	0	a_{31}	-13,3
a_{12}	-30,808	a_{22}	0	a_{32}	30
a_{13}	9,371	a_{23}	0	a_{33}	-22,867
a_{14}	0,022	a_{24}		a_{34}	6,9
b_{11}	0	b_{21}	0,4	b_{31}	0,75
b_{12}	0,4	b_{22}	0,75	b_{32}	1,0

We also take into account the constraints on connecting to the power grid, which determine the maximum throughput at the GPC:

$$P_{grid}(t) \le P_{con} \quad (17)$$

V. MICROGRID SIMULATION

In the paper, MG is investigated with a "Smart" scenario for controlling its work, proposed in the study [8]. This scenario is focused on maximizing profits when using a time-of-use tariff, through flexible use of the ESS and the sale of electricity to the grid. In the morning and evening, its own consumption is covered by the ESS, and all available generation is for sale. During the day, the ESS is charged from RES generation, and the surplus of energy is sold to the grid. At the minimum tariff, ESS is charged from the grid. ESS operates in discharge or charge modes depending on the time of day, the cost of electricity, and the balance of generation and consumption.

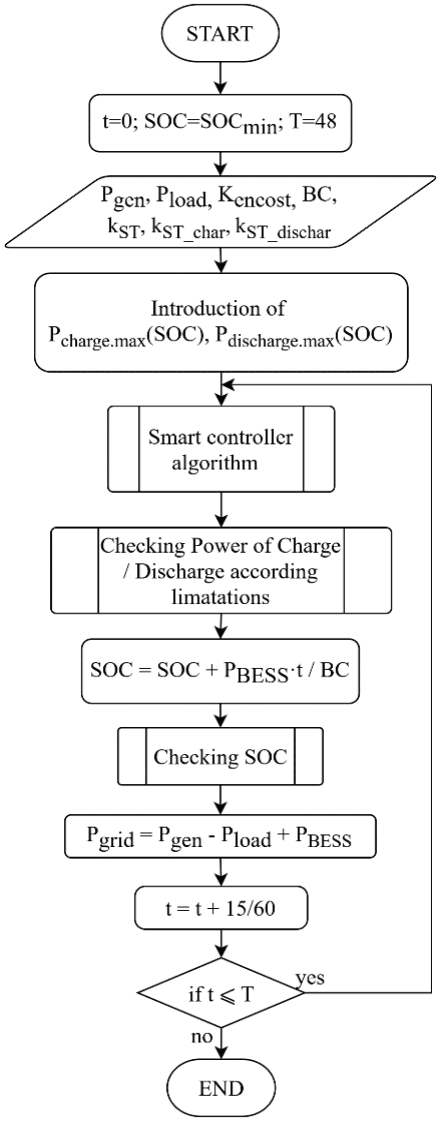


Fig. 3. Algorithm of the prosumer microgrid

Fig.3 shows the algorithm for modelling the prosumer MG operation and control. Every 15 minutes, the algorithm uses data from the generation, load, and state of charge of the ESS to determine the energy distribution.

VI. Optimization Method

To solve the optimization problem, the Hook-Jeeves method of direct search was chosen. The search starts from the starting point x_0. The choice of the starting point can affect the search result, so several runs of the method with different x_0 were carried out. The search is performed with acceleration, the multiplier of which is automatically determined by Matlab. The completion condition is to reduce the search step to 10^{-6}.

This study looked at one type of prosumer of a family consisting of two adults and two children, as described in the paper [13]. The installed load of the prosumer is 7 *kW*, and the daily consumption during the year varies from *12* to *24 kWh*. Simulation of the prosumer microgrid was carried out for the rated power of the rooftop PV plant of *8 kW*, separately for the winter and summer periods of the year. The duration of the test period of operation of the system T in (1) was *48 hours*, that is two days. Numerical modelling and optimization were performed using Matlab.

According to the local nature of the chosen optimization method, the complexity of the function (1) and the change in the modelling conditions, the choice of the starting point x_0 can significantly affect the results. Therefore, to check, the algorithm was run from the starting points $x_0 = 0$ *kWh*, $x_0 = 30$ *kWh*, $x_0 = 60$ *kWh*. These x_0 values correspond to the edges and the middle of the desired range of battery capacity of ESS. The results showed that running from all starting points resulted in the same decision, although the number of iterations varied but did not exceed *50*.

VII. Simulation Results

The results of the simulations include data on the generation and electrical load of the prosumer [14], which are characterized by averaged profiles of the working (ED)

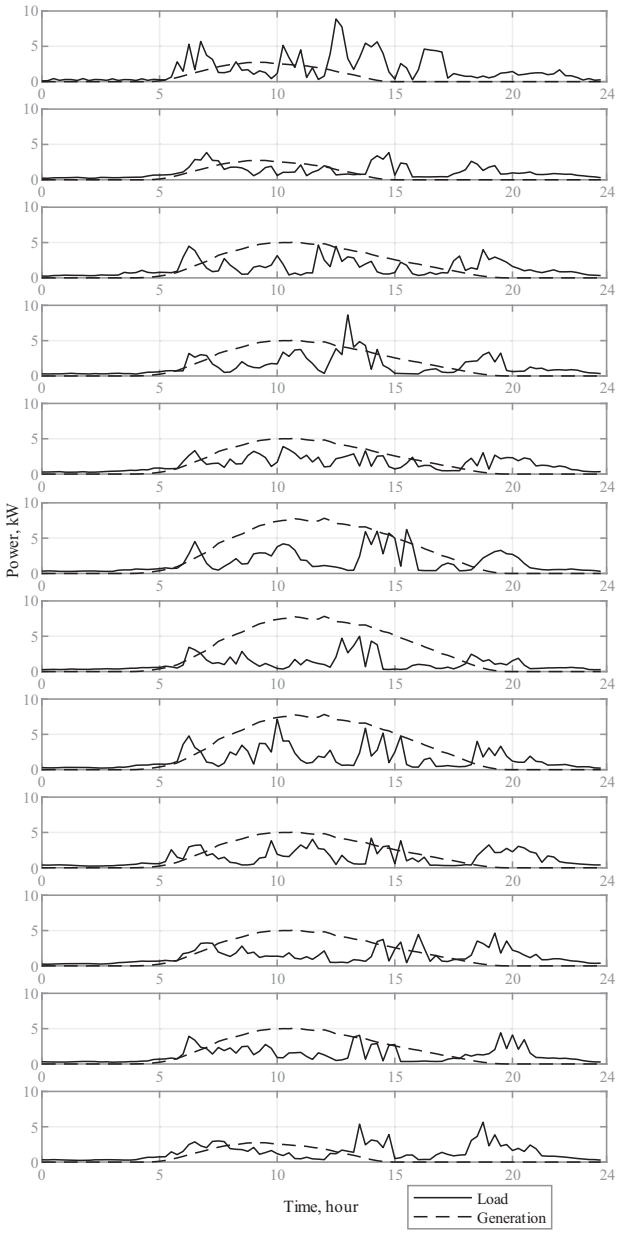


Fig. 4. Work day load and generation profiles for each month

and weekend (WD) days for each month. Fig.4 shows an example of generation profiles data and averaged workday load profiles for each month.

According to the results of the previous work [8], the selection of the optimal ESS capacity for the summer and winter seasons was provided. This selection was made taking into account the constraints at the connection point. The results are presented in Table 4.

According to Table 1, the most common connection capacities for the private sector to the power grid in Ukraine were chosen.

TABLE IV. OPTIMAL VALUES OF THE ENERGY STORAGE CAPACITY

	P_{cont}, kW		
	7	*13,4*	*36,5*
BC_{Winter}, *kWh*	11	11	11
BC_{Summer}, *kWh*	15	26	36

To check the efficiency of the selected values of the ESS capacitance, a simulation of the prosumer microgrid was carried out during the year using average load profiles (for working days and weekends) and a seasonal generation profile. Based on the results of the simulation, the Grid Feed-in Energy (GFE) and Specific Grid Feed-in Energy (SGFE) for each month were calculated.

SGFE is calculated according to the formula (17), and takes into account the cost factor at the TOU tariff, i.e. the weight of each *kWh* during the day.

$$W_{SN} = (P_{sell}(t) - P_{buy}(t)) \cdot \Delta T \cdot K_{TOU}(t) \qquad (18)$$

Fig. 5 shows the hourly profile of the three zone time-of-use tariff coefficient (K_{TOU}), which was used in this study [15].

In Table 5, the data of GFE and SGFE of the MG of the prosumer under the condition of constraint of *36.5 kW* and ESS with a capacity of *11 kWh* or *36 kWh* are given. As can be seen from the results, the total GFE of weekends and working days for the year, for MG with ESS with a capacity of *11 kWh* is *2982.29 kWh*, and SGFE is *9127.5 kWh*, which shows the efficiency of using the proposed scenario of the prosumer MG operation. For MG with ESS with a capacity of *36 kWh*, GFE for the year is *505.92 kWh*, which is 6 times

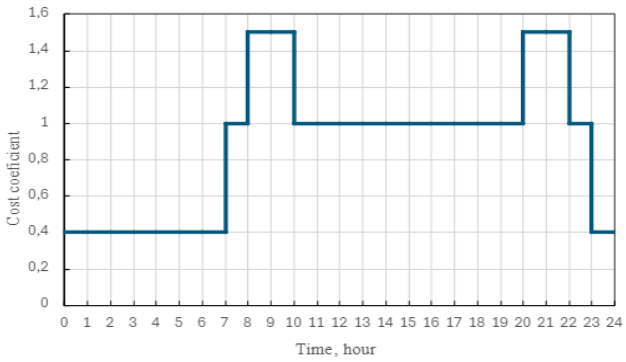


Fig. 5. Electricity cost coefficient at a three zone time-of-use tariff

less than for MG with ESS capacity of *11 kWh*. SGFE is *15403.95 kWh*, which is *1.65* times more than for MG with a *11 kWh* ESS.

TABLE V. ENERGY BALANCE DATA AT A CONSTRAINT OF 36 KW

Month	Capacity 11 kWh				Capacity 36 kWh			
	GFE, kWh		SGFE, kWh		GFE, kWh		SGFE, kWh	
	ED	*WD*	*ED*	*WD*	*ED*	*WD*	*ED*	*WD*
Jan.	-606,8	-150	-322,8	-73,7	-748,3	-214,4	8,4	47,1
Feb.	-239,7	-133	38,1	-24,7	-352,4	-213,1	388,5	72,3
Mar.	157,3	81,9	581,9	227,6	54,7	24,3	1071,6	368,0
Apr.	159,2	81,2	553,3	215,3	24,2	25,4	962,8	357,6
May	161,4	52,3	540,7	187,5	-7,8	20,7	940,8	361,9
June	615,5	224,8	1043,3	376,5	460,6	166,6	1433	516,3
July	897,6	404,5	1394,4	566	730,2	336,6	1796,3	696,1
Aug.	671,4	316,6	1110	480,4	511,2	255,9	1519,1	617,7
Sept.	181,5	100,9	579,1	232,3	36	36	978,2	365,4
Oct.	151,8	82,3	538,7	217,7	-79,5	36,4	816,9	369,8
Nov.	237,1	117,2	615,1	249,5	99,5	58,1	1022,1	388,5
Dec.	-395,8	-187	-88,4	-110	-489,5	-265,3	350,4	-44,7
Year	1990,4	991,9	6583,2	2544,4	238,8	267,2	11288	4116

For comparison, in Table 6, simulation data are given for the MG prosumer with optimal ESS capacities with a connection power constraint of *7 kW*. Total GFE for MG with ESS capacity of *11 kWh* is *2560.3 kWh*, and SGFE is *8490.7 kWh*. For MG with ESS capacity of *15 kWh*, the GFE for the year is *2285.7 kWh*, which is *1.12* times less than for MG with ESS capacity of *11 kWh*. SGFE is *10261.2 kWh*, which is *1.2* times more than for MG with an *11 kWh* ESS.

TABLE VI. ENERGY BALANCE DATA AT A CONSTRAINT OF 7 KW

Month	Capacity 11 kWh				Capacity 15 kWh			
	GFE, kWh		SGFE, kWh		GFE, kWh		SGFE, kWh	
	ED	*WD*	*ED*	*WD*	*ED*	*WD*	*ED*	*WD*
Jan.	-595,0	-147,9	-305,5	-70,7	-633,9	-158,2	-249,1	-40,8
Feb.	-234,6	-146,9	45,6	-45,7	-251,3	-157,2	129,2	-15,9
Mar.	157,5	66,0	581,9	203,6	152,5	63,4	712,5	248,2
Apr.	115,4	65,3	487,3	191,3	111,5	59,2	613,1	232,4
May	115,6	52,4	471,8	187,5	101,1	48,4	592,8	230,7
June	571,9	208,9	977,5	352,5	552,8	202,1	1088,2	392,9
July	852,0	388,6	1325,7	542,1	831,2	380,8	1440,4	581,4
Aug.	625,7	300,7	1041,2	456,5	604,0	293,5	1155,0	496,4
Sept.	137,6	85,0	513,2	208,3	122,4	77,3	627,5	247,8
Oct.	152,0	66,4	538,8	193,7	147,4	62,1	669,8	236,5
Nov.	193,2	101,3	549,1	225,6	176,6	94,2	662,1	265,7
Dec.	-389,9	-180,9	-79,7	-101,0	-402,3	-192,0	23,3	-78,9
Year	1701,5	858,8	6146,9	2343,8	1512,0	773,7	7464,8	2796,4

This difference in results is due to the fact that when calculating GFE, the energy from the PV plant, which covers its own consumption, is not taken into account, but only the output energy passing through the grid connection point is taken into account. In addition, the larger capacity of ESS makes it possible to use a larger share of the energy generated by its own PV system, thereby reducing the purchase of electricity from the power grid to cover its own consumption. At the same time, increasing the capacity of ESS allows to increase the amount of energy sales to the

grid thanks to TOU tariff, which takes into account the different weight of *1 kWh* during the day.

To evaluate the effectiveness of the implementation of different configurations of the prosumer MG, six options were analyzed with different constraints of power at the connection point and corresponding ESS capacity values, as shown in Table 4. The calculation of the payback period was carried out taking into account the following input parameters: *12000 ₴* per *1 kW* of installed capacity, *13500 ₴* per *1 kWh* of ESS capacity and the electricity tariff of *4.32 ₴/kWh*.

For a prosumer MG with ESS capacity of *11 kWh* and a constraint of *36.5 kW*, the payback period is *9.3 years*, and with a limit of *7 kW — 10 years*. The payback period for MG with a *36 kWh* ESS and a constraint of *36.5 kW* is *12.8 years*. For MG with a power constraint of *36.5 kW* and ESS capacity of *15 kWh*, the payback period will be *12.5 years*.

For a prosumer MG with ESS capacity of *11 kWh* and a constraint of *36.5 kW*, the payback period is *9.3 years*, while under the same conditions, but with a constraint of *7 kW*, the period increases to *10 years*. This means that when the permissible constraint is increased from *7* to *36.5 kW*, the payback period is reduced by *1.07 times*.

For MG with a *36 kWh* capacity and a *36.5 kW* constraint, the payback period is *12.8 years*. If, with a constraint of *36.5 kW*, the optimal capacity of the *15 kWh* energy storage system is chosen, the payback period is reduced to *12.5 years*. This is about *1.03 times* less.

Thus, changes in the permissible capacity of the constraint affect the payback period. At the same time, the change in the payback period is not linear, as it depends on many economic and technical factors.

VIII. Conclusions

As a result of the study, it was found that the level of power constraint at the connection point significantly affects the efficiency of the prosumer MG. Relaxation of this constraint allows for the use of a larger capacity of ESS, which in turn expands the possibilities for domestic consumption and sale of excess energy.

However, despite the potential increase in profits, capital expenditures associated with the increase of ESS remain the main factor determining the payback period of the project. At the current stage, these costs have a greater impact than the additional profit from the sale of energy. This indicates the need for an optimal balance between the technical capabilities of the system and the economic constraints of the investor.

Thus, in order to increase the efficiency of the implementation of MG prosumers, it is needed to take into account not only the technical potential, but also to carefully assess the financial feasibility of each scenario.

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Analysis of Audio Signal Denoising Efficiency Using ESTOI Score

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***Abstract*—Audio and speech signals are an important part of multimedia data. It is required to have their high quality. Meanwhile, noise can severely degrade quality of acquired signals. Then, one has to perform their pre-filtering (noise removal). Filters based on discrete cosine transform (DCT) in blocks is one option. Their performance depends on several factors including block size and threshold type. Alongside with conventional metrics of performance such as improvement of signal-to-noise ratio (ISNR) it is worth characterizing the filter efficiency by other metrics dealing with perception of processed signals by humans. We carry out such analysis using Extended Short-Time Objective Intelligibility (ESTOI) score. It is shown that the results of analysis do not coincide with ISNR-based conclusions. In particular, benefits due to filtering become negligible for input SNR larger than 20 dB. Besides, the filtering efficiency occurs different depending on threshold type. The results are presented for several test signals corrupted by additive white Gaussian noise (AWGN) of intensity varying in wide limits.**

Keywords—speech signal, denoising, perception, discrete cosine transform, efficiency analysis

I. Introduction

Speech and other types of audio signals are widely used in practice. In particular, they have become an important part of multimedia data [1, 2]. One problem is that speech signals are often recorded together with noise from environment and there is a necessity to remove this noise [3, 4]. Speech denoising (filtering) is a quite old task and several groups of methods have been already designed for this purpose. Linear filters, both finite and infinite impulse response were developed and tested first [4, 5]; later, adaptive filters were proposed [5, 6]. After this, the main attention was paid to denoising techniques based on orthogonal transforms, in particular, wavelets [7, 8]. Finally, convolutional neural networks and autoencoders have become popular in recent years [9, 10].

There are several reasons for such a variety of methods that can be applied. On the one hand, mathematical tools used to describe information signal and noise were developing offering new facilities. On the other hand, new requirements to data processing appeared and they had to be taken into account. There are three important points in this sense. First, speech is perceived by humans and, in this sense, performance of speech denoising techniques has to assessed taking this fact into account [4, 11, 12]. Second, depending on an application, reasons for the noise presence can be different and, moreover, the noise intensity can be quite high. In particular, this might take place in overcrowded rooms and other places [12]. Besides, the noise statistical and spectral/correlation characteristics might be quite different as well [12, 13]. Third, the noise removal has to be performed in real time for many applications [14].

In other words, a speech signal denoising method possess a set of the following useful properties: 1) it has to be able to adapt to the properties signal and noise at hand; 2) a method and an algorithm realizing it should be simple and fast; 3) it has to be efficient in the sense of noise suppression for a wide range of conditions that might occur in practice; 4) it should provide favorable pre-requisites for solving further tasks of signal perception, speech translation to text [15] and so on.

The third property leads to necessity to apply either processing with a limited size of scanning window or denoising of an entire signal part by part, for example, using blocks of a certain length. Only in these cases the output signal can be obtained with a reasonable delay with respect to the input signal [16]. According to the second property, one has to use transforms having fast algorithms and/or efficient hardware/software implementations. According to the last property, one has to use in efficiency analysis not only traditional metrics such as improvement of signal-to-noise ratio (ISNR) or the ratio of output mean square error to noise variance at the input, but also metrics that adequately characterize audio (speech) signal perception [17-19].

Alongside with wavelet-based denoising techniques [20-22], methods based on discrete cosine transform [23-25] can be a good candidate for speech denoising. The arguments in favor of DCT-based filtering are the following. First, DCT is widely exploited in image filtering and lossy compression where its excellent energy compaction property plays the key role [26-28]. Second, DCT is applied in blocks of a limited size [24, 29] where fast algorithms exist. Then, output can be obtained with a relatively small delay.

Similarly to wavelet-based denoisers, there is a set of parameters that influence signal processing performance. One of them is block size and block overlapping. In [24], we have considered the block size equal to 32, but it is possible that the use of a larger block size can be expedient. Blocks can be non-overlapping, partly overlapping, and fully overlapping. In the last version, the efficiency of noise suppression is the best but the speed of data processing is the worst. However, real-time processing is still possible. Thus, in this paper we consider only full overlapping of the blocks.

Threshold type is also important. In [24], we have studied the hard and combined types of threshold and shown that they can be expedient depending on the used criterion of filtering efficiency. Recall that PESQ metric has been employed in [24]. In this paper, we concentrate on another metric, ESTOI store, since it is also widely used in analysis [30].

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Thus, the goal of this paper is to carry out analysis of DCT-based filtering for speech signals corrupted by a rather intensive AWGN using two block sizes (32 and 64), two threshold types, and the quality metric ESTOI score.

II. Signal/Noise Model and Fundamentals of DCT-Based Filtering

In this paper, we assume that speech signal S(i), i = 1,...,I, is contaminated by AWGN n(i), i=1,...,I, where i and I are the sample index and size, respectively, i.e.

$$S_n(i)=S(i) + n(i),\ i=1,...,I. \quad (1)$$

AWGN is assumed zero mean and having variance σ^2 which is constant for any observation interval and either known a priori or accurately estimated in a blind manner [31, 32].

A registered (recorded) signal is processed in blocks of size N. A most common choice is to set N equal to powers of 2 such as 16, 32, or 64 providing an opportunity of using efficient algorithms of fast DCT calculation. Suppose that an l-th block involves $S_n^{bl}(l, j)=\{S_n(l + j - 1)\}$, j= 1,...,N. Thus, for fully overlapping blocks, there are I-N+1 possible block positions. For each position, three basic operations are performed: 1) direct DCT with obtaining N DCT coefficients $D_l(k)$, k=1,...,N; 2) thresholding with obtaining the thresholded DCT coefficients $D_{l\ thr}(k)$,k=1,...,N; 3) inverse DCT with obtaining filtered values $S_n^f(l, j)=\{S_n(l + j - 1)\}$, j= 1,...,N. Note here that: 1) in opposite to scanning window filtering, denoised values are obtained for all samples belonging to a given block; 2) most samples (except of i<N-1 and i>I-N+2) belong to N positions of blocks. Then, for each data sample, one has multiple estimates of filtered values (for most samples, under typical assumption I>>N, N filtered values) and such estimates have to be jointly processed. Here, we use the simplest way of their processing, namely, averaging.

The operation of thresholding is one of the most important since it has considerable impact on denoising performance. First, the coefficient $D_l(1)$ "responsible" for mean level in block, is kept unchanged. Other DCT coefficients can be processed in different manner. One of the simplest ways is hard thresholding defined as

$$D_{l\,\mathrm{thr}}(k) = \begin{cases} D_l(k), & \text{if } |D_l(k)| > T \\ 0, & \text{if } |D_l(k)| \le T \end{cases}, k = 2,\dots N. \quad (2)$$

Another option is to apply combined thresholding

$$D_{\mathrm{thr}}(k) = \begin{cases} D_l(k), & \text{if } |D_l(k)| > T \\ D_l{}^3(k)/T^2, & \text{if } |D_l(k)| \le T \end{cases}, k = 2,\dots N, \quad (3)$$

where T denotes threshold in both formulas. In wavelet and DCT-based denoising, the threshold is some function of the noise standard deviation σ (just because of this, the noise variance of standard deviation has to be known in advance or accurately pre-estimated). It is common to present T=βσ where β is proportionality coefficient that influences the filter properties. A larger β leads to more efficient noise suppression by the expense of worse preservation of important details. Then, a reasonable balance (optimum) should be found or, in a more complex case, β has to be adjusted in locally adaptive way. If β is fixed, a common rule is to set β≈2.7 for the hard thresholding (2) and β≈4.5 for the combined thresholding (3). Meanwhile, in our further studies (similar to [24]), we will analyze β varying in rather wide limits. This is done because of two reasons. First, the results in [24] show that optimal β depends on input signal-to-noise ratio defined as $SNR_{inp}=P_s/\sigma^2$, where P_s is the power of S(i), i = 1,...,I. The results presented in several paper [23-25] show that the noise can be hardly noticed in noisy recorded signals if SNR_{inp}>30 dB. Besides, filtering efficiency for any modern filter is not too high for such large input SNRs. Then, the basic area of our interest is input SNR in the limits from 0 to 20 dB for which noise removal is strongly desired and filtering efficiency can be quite high.

Here it is also worth recalling that analysis of filtering efficiency should be carried out not only for different input SNRs but for different test (speech) signals as well [23-25]. For this purpose, sets of standard speech signals considered noise-free have been created. In particular, English language Harvard phrases commonly treated as standard [33, 34] are available. All the records have duration of 2s where the sampling rate is 16 kHz [34]. We have used five records with male and female voices from aforementioned database.

III. Efficiency Analysis

As mentioned above, there are conventional and speech quality criteria of noise removal efficiency. Let us start from conventional criterion called improvement of SNR that can be characterized as

$$\mathrm{ISNR} = 10log_{10}\left(\frac{\sigma^2}{\mathrm{MSE}}\right) = SNR_{\mathrm{out}} - SNR_{\mathrm{inp}}, \quad (5)$$

where $SNR_{out}=P_s/MSE$, MSE is mean square error at the filter output, and it is assumed that for a good filter (or a situation when filtering is expedient) one has MSE significantly less than σ^2.

One example of the obtained dependences is presented in Fig. 1. In this case, SNR_{inp}=20 dB, and, as seen, maximal ISNR values are quite large. They reach ≈2.5 dB for N=16, about 3 dB for N=32, and about 3.4 dB for N=64. Experiments with other test signals have demonstrated similar results. The main conclusions that can be drawn are the following: 1) the filtering is expedient, especially if N=64; 2) the results for both types of thresholds are almost the same; 3) the optimal β are as recommended; meanwhile, small changes in β do not lead to catastrophic degradations in the filter efficiency.

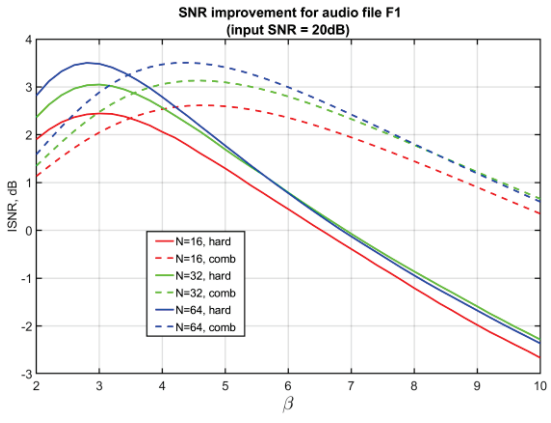


Fig. 1. Dependences of ISNR on β for audio file F1 denoising by the DCT-based filters with types of thresholds for three values of N (16, 32, and 64).

The latter conclusion is important for practice. Recall that T=βσ and the estimate of the noise standard deviation can be used instead of σ in threshold setting. Then, small (for example, 10%) errors in σ estimation would not result in

significant degradation of denoising efficiency. Meanwhile, larger relative errors are undesired.

The paper [35] states that "a good speech enhancer with a good PESQ performance delivers poor word error ratio performance in recognition." This is the reason for using ESTOI score. The results presented in [36] have demonstrated that joint optimization of signal enhancement method using both STOI and MSE "has improved speech intelligibility according to both objective and subjective measures" and resulted in better word accuracy according to automatic speech recognition.

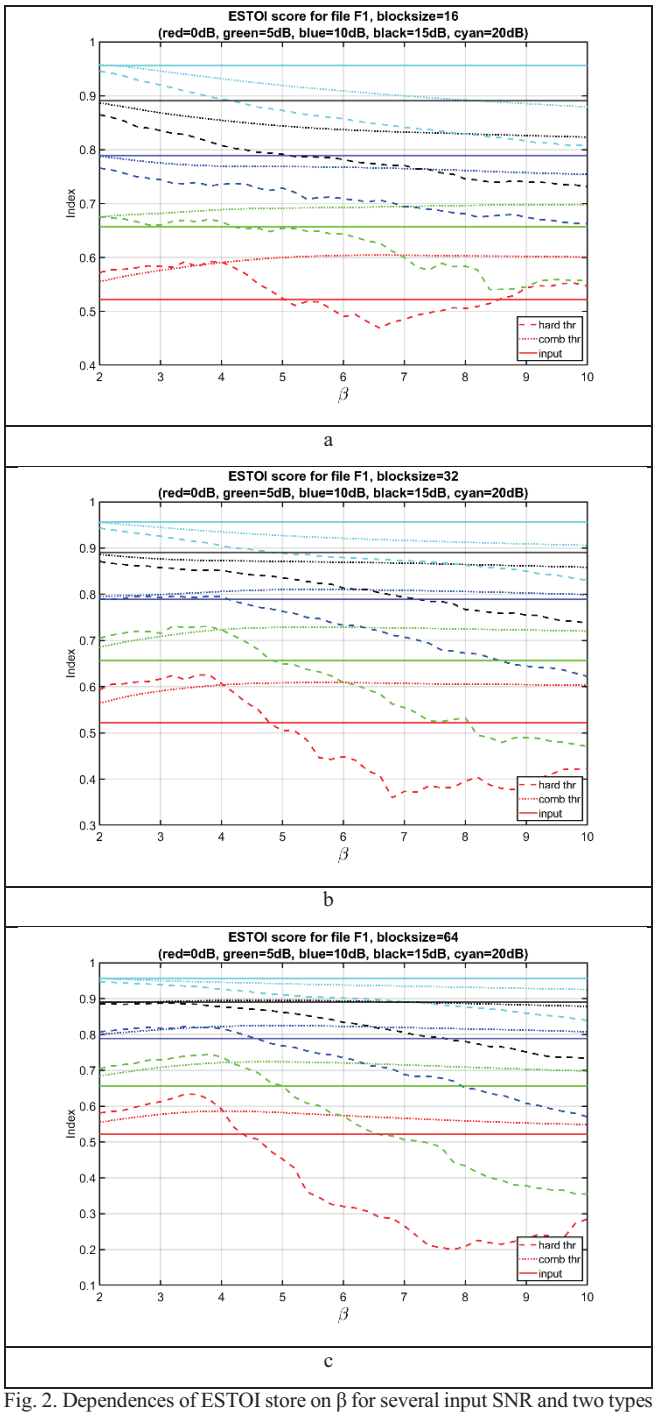


Fig. 2. Dependences of ESTOI store on β for several input SNR and two types of thresholds for the file F1 for N=16 (a), 32 (b), and 64 (c)

Keeping this in mind, we have got the dependences of ESTOI store on β for all five test signals. One example (for the file F1) is given in Fig. 2 for two types of thresholds and five noise intensities providing input SNR equal to 0, 5, 10, 15, and 20 dB. Input SNRs are shown by the corresponding horizontal solid lines.

Several interesting tendencies are observed. First, if input SNR is equal to 20 and 15 dB, no improvement of ESTOI score takes place for ESTOI score for all three considered block sizes. For SNR_{inp}=10 dB, no improvement is observed for N=16, but improvement is possible for N=64. Second, for input SNR equal to 10, 5, and 0 dB, the input data quality is low. Meanwhile, the improvement takes place, especially for SNR_{inp}=0 dB. Third, if β is quite large, it is possible that denoising is not needed (leads to negative effects) even if input SNR is low (look at data in Fig. 2,a for SNR_{inp}=0 dB for β=6…7 dB for hard thresholding). In addition, dependences on β might have "fluctuations". These happens for hard thresholding (dashed lines).

Thus, in general, positive outcomes of filtering according to the ESTOI score are "less" than for IPSNR and PESQ metrics. One possible reason is that methods of speech automatic recognition should be learnt for filtered signal if applied to denoised data. Finally, for SNR_{inp}=0, 5, and 10 dB, the DCT-based filtering with combined thresholding and the corresponding the optimal β≈6 perform better than for hard thresholding with the optimal β≈3.

Another point is that we need to be sure that the effects described above take place for other test signals. To partly check this, Fig. 3 presents data for the file F4. Some tendencies mentioned above are also valid for this file. In particular, improvement of ESTOI score due to filtering is observed for SNR_{inp}=0, 5, and 10 dB for all N if β is set properly (e.g., β≈3 for hard thresholding and β≈6 for combined thresholding). It is again possible that, for large β, degradation (ESTOI score values smaller than for input signal) is observed – see, e.g., data for N=64. Meanwhile, a small improvement of ESTOI score due to denoising is observed for the file F4 even for SNR_{inp}=20 dB for N=16 and 32.

Thus, comparison of data for the same SNR_{inp} for two files (F1 in Fig. 2 and F4 in Fig. 3) shows the following. Efficiency of denoising depends not only on SNR_{inp} but also on signal properties. Moreover, it seems that, depending on SNR_{inp} and signal properties, optimal N can be different. To check this hypothesis, let us present some data for other files. The obtained data are collected in Table I. It is seen that for situations are possible: 1) N=16 is optimal; 2) N=32 is optimal; 3) N=64 is optimal; 4) denoising is useless for any N. The tendency is that if input SNR becomes larger, optimal N becomes larger too until denoising becomes useless for any N. The same tendencies are observed for data presented in Figures 2 and 3.

In other words, for small input SNR (0-10 dB), denoising is usually quite efficient (according to ESTOI score) whilst it can be not efficient or even useless for larger input SNR. This means that: 1) it could be reasonable to carry out some adaptation to input SNR based on its estimation (evaluation) for data at hand; 2) carrying out some prediction of filtering efficiency for a signal at hand to be denoised (processed).

Recall here that we have earlier proposed the procedure of ISNR prediction [25] for fixed N and different β. It is quite probable that similar procedures can be designed to predict improvement of ESTOI score due to filtering in order to decide is it worth filtering a given signal or its fragments and, if filtering is expedient, what parameters of the DCT-based

filter to set. According to the obtained results, hard thresholding performs slightly better for low input SNR when filtering can be expedient.

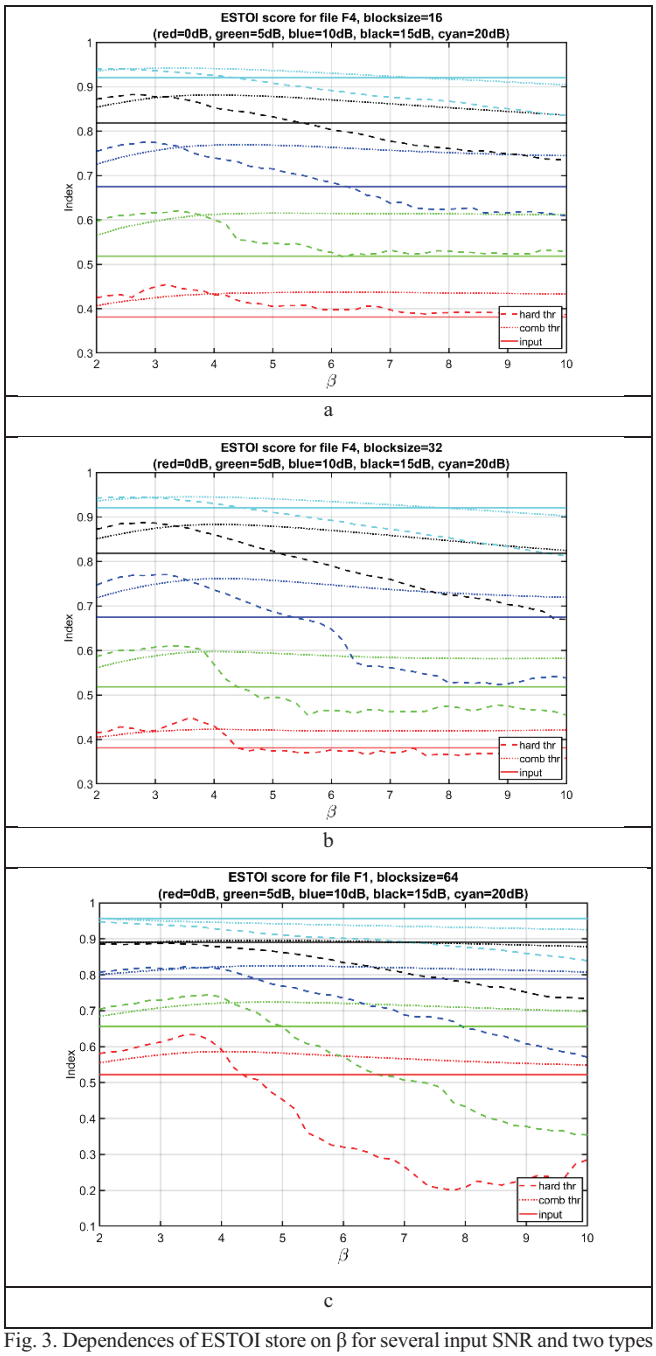


Fig. 3. Dependences of ESTOI store on β for several input SNR and two types of thresholds for the file F4 for N=16 (a), 32 (b), and 64 (c)

IV. Filtering Example and Discussion

It is also worth presenting one example of signal denoising. Fig. 4 shows the results for F2 file. Note that noise is clearly visible in places of low intensity signal and can be hardly noticed in fragments where it is "masked". Both versions of the DCT-based filter with optimal β for each threshold type (see the plots in Figures 4,c and 4,d) suppress the noise effectively and, subjectively (by visual inspection), it is difficult to conclude what variant is better. Probably, it is worth collecting statistical data for a set of iiput SNR values for tens or hundreds of test signals to establish what type of threshold is better or when it is better. This can be one direction of further studies.

The obtained data give insight on performance of the DCT-based filter from the viewpoint of ESTOI score metric. As one can see, the results do not give strict answer when denoising is expedient and when it should be skipped. Besides, it does not give a strict answer, what are the optimal parameters of the DCT-based filter. On the contrary, the obtained results show that, depending on the situation at hand, different parameters of the DCT-based filter might be the best. Then, it might be reasonable to apply either prediction or learning based approaches to get the optimal solution.

File	Input SNR, dB	Optimal N
F0	0	32
F0	10	64
F0	20	64
F2	0	16
F2	10	64
F2	20	64
F3	0	64
F3	10	64
F3	20	Denoising is useless

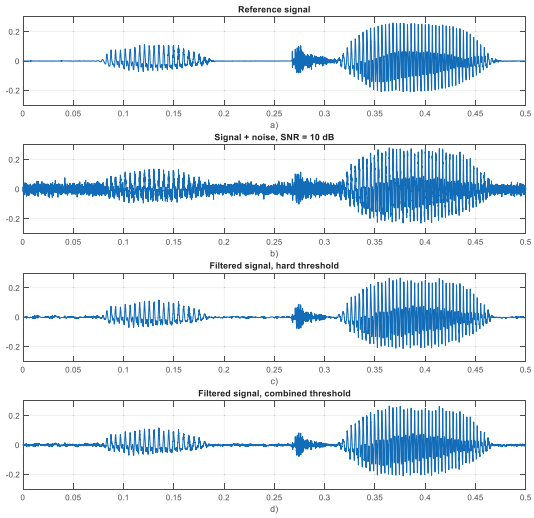


Fig. 4. Filtering example: noise-free signal (a), noisy signal (b), output for hard thresholding (c), output for combined thresholding (d)

V. Conclusions

The paper presents the results of efficiency analysis for DCT-based filters applied to speech signals from using ESTOI score. It is shown that denoising can lead to positive outcomes – this usually happens for input SNR from 0 to 10 dB. Meanwhile, for input SNR about 20 dB, the denoising is often useless. At the same time, denoising is still efficient according to ISNR criterion. Thus, different criteria might give different opinions concerning denoising efficiency for the same situation and, thus, the decision depends on application at hand and what criterion is more adequate.

Directions of research to be held in the future are considered as well. It is stressed that statistics for a significantly larger number of test signals has to be collected. Prediction approaches similar to [25] have to be modified for ESTOI score and, possibly, other metrics.

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A Combined Approach to Evaluating the Effectiveness of Deep Learning Models for Aerial Image Processing

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***Abstract*—The work is devoted to the process of research and analysis of deep learning models for segmentation of aerial images. The article proposes to use a new method of combined evaluation of the efficiency of models, which combines standard quantitative metrics (loss function, accuracy, precision, recall, Jaccard index, F-measure, prediction time of one image) and weighting coefficients of priority of certain metrics, depending on the task. To evaluate the results, modern architectures U-Net, DeepLabV3+, FPN and PSPNet, which are most suitable for image segmentation problems, were used. The study was carried out using its own set of aerial images and their corresponding segmented images, on which seven classes of different objects were separated. The results obtained show the stability and generalizing ability of the combined evaluation when changing weight parameters, and also emphasize the advantages of some architectures when they are applied in different scenarios. Diagrams are used to visualize the studies carried out and the results, which further explain the proposed approach.**

***Index Terms*—deep learning, semantic segmentation, U-Net, DeepLabV3+, FPN, neural networks, drones, aerial images, effectiveness evaluation.**

I. Introduction and Related Work

Today, with the development of technologies such as unmanned aerial vehicles, it is important to develop special approaches that can provide fast and accurate analysis of large volumes of images. When selectively searching for a variety of objects (e.g. buildings, roads, agricultural areas, etc.) in aerial images obtained from drones, it is important to obtain their specific definition. Semantic segmentation is the one of the key tasks of computer vision in the field of image analysis, which allows to divide images into separate areas corresponding to specific objects. Segmentation can be particularly useful in the areas of urban planning, environmental monitoring, forest fire detection, water resource analysis, etc. [1]. Scientists are working to simplify and improve the results of image segmentation by creating universal segmenters that can be useful for solving various problems [2].

The active development and comprehensive use of deep learning neural network models encourages researchers to constantly solve problems of accurate and rapid detection of objects of natural or artificial origin on large areas of the earth's surface. Semantic segmentation based on the U-Net, DeepLabV3+, Feature Pyramid Network (FPN) and Pyramid Scene Parsing Network (PSPNet) architectures has proven to be a powerful tool for recognizing objects in images in various fields of human activity, including aerial images. Modern research in this area focuses on modifying such models in order to significantly increase and improve their efficiency.

The best-known architecture for image segmentation is U-Net [3], which was originally developed for biomedical segmentation [4] and was later adapted to the field of remote sensing. U-Net gained its popularity due to the use of a symmetrical "encoder-decoder" structure with spaces. A significant amount of modern research has focused on improving the classical U-Net, for example, through the introduction of self-attention modules and split convolutions [5], allowing an accuracy of more than 91% to be achieved on a test dataset. To achieve even better results, especially in the field of aerial image analysis, the U-Net model is modified by increasing the number of connections in the middle layers and storing more image data using S-shaped circular connections. In the work [6], using this modification, it was possible to achieve an accuracy of more than 96% on test data. Scientists also use U-Net with different types of encoders, such as ResNet or DenseNet [7], which allows nested structures to be combined to detect small objects.

Another popular architecture is DeepLabV3+ (an improved version of DeepLab), which is uses spatial averaging and multi-level feature detection by the Atrous Spatial Pyramid Pooling (ASPP) module [8]. In the field of aerial images, DeepLabV3+ is good at recognizing context at different scales. Modern research focuses on adapting this model to image features, for example by using double image streams [9] or transfer training [10]. The combination of the DeepLabV3 and DeepLabV3+ models helps to achieve better efficiency, due to their optimization and facilitation, when deploying models to real-time edge devices [11], for example, in emergency monitoring tasks such as fires and floods.

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FPN pyramidal architecture is also widely used in semantic segmentation problems [12]. It builds a multi-level hierarchy of features for more accurate segmentation. By combining features from different depth levels, the FPN model works efficiently with complex scenes and objects of different scales using an encoder of the ResNet type [13]. Modern research shows that the use of spatial and channel attention mechanisms can achieve even higher accuracy in segmentation of small objects, showing better results [14].

Another pyramidal architecture, PSPNet [15], is particularly useful for processing images with complex structures because it runs on the pyramid pooling model. The model is able to better understand the global structure of the scene and separate small objects. PSPNet has found wide application in the medical field [16], achieving improved results under challenging image processing conditions. It can also be used to process aerial images and provide significantly better results compared to other models. For example, in [17], PSPNet was modified to Shift Pooling PSPNet, which significantly improved the identification of buildings in aerial images and demonstrated the benefits of localized feature pooling.

Important studies that cover the current topic are also comparisons of existing deep learning architectures in the context of specific tasks. For example, scientists in the paper [18] compare models to detect cracks in building materials to understand the strengths and weaknesses of models under specific conditions. And in order to demonstrate the flexibility and versatility of architectures, the use of various decoder encoders, for example, in the medical field, is being investigated [19].

Modern research shows an active improvement of the basic architectures of deep learning networks, which can also be used for the analysis of aerial images, their optimization and comparative analysis of efficiency to solve various problems. In the context of the active development of remote sensing, the issue of choosing the optimal model is becoming more and more relevant. In research, as a rule, individual metrics are calculated, such as accuracy, precision, recall, etc., but individual metrics do not always allow for a qualitative assessment of the effectiveness of the model for solving specific problems.

The main purpose of our work is to develop our own approach to comparing segmentation models, combined effectiveness evaluation, which will allow for a deeper analysis of the strengths and weaknesses of popular architectures due to the combination of several metrics into a single indicator.

II. Method overview

A. Problem Statement

We propose to calculate a combined effectiveness score (CES) for each model, which will take into account the average prediction time of a single image, the loss function, and metrics such as accuracy, precision, recall, IoU, and F1-Score using a normalized total. It can be especially useful when models produce approximately the same results and is important for balancing the accuracy and efficiency of the model. This will allow to effectively compare models with the same value and can easily adapt to different tasks, including real-time image processing.

B. Combined Effectiveness Score

To calculate the combined efficiency score, it is proposed to set the necessary weights for each metric, accordingly:

- w_1 - loss function;
- w_2 - accuracy;
- w_3 - precision;
- w_4 - recall;
- w_5 - Jaccard index;
- w_6 - F-measure;
- w_7 - prediction time of one image.

Using the specified weighting factors, it is possible to create a vector of weights W:

$$\mathrm{W} = (w_1, w_2, w_3, w_4, w_5, w_6, w_7) \tag{1}$$

If n is the total number of weighting factors, then the vector W obtained in (1) must satisfy the normalization condition:

$$\sum_{i=1}^{n} w_i = 1, w_i \in [0, 1] \tag{2}$$

For the calculations of normalized values, we used minimum-maximum normalization, which allows to bring the values to a single scale (in the range from 0 to 1), especially useful for normalizing the loss function and the prediction time of a single image:

$$\bar{V} = \frac{V_{avg} - V_{min}}{V_{max} - V_{min}}, \tag{3}$$

where $\bar{V}$ is the normalized value of the some value V, V_{avg} is the average value of the value on the test data, V_{min} is the minimum value, V_{max} is the maximum value.

For each standard metric, the corresponding normalized value must be calculated using (3):

- $\bar{L}$ - loss function;
- $\bar{A}$ - accuracy;
- $\bar{P}$ - precision;
- $\bar{R}$ - recall;
- $\bar{IoU}$ - Jaccard index;
- $\bar{F}1$ - F-measure;
- $\bar{T}$ - prediction time of one image.

Using the specified normalized values, it is possible to create a vector of values X:

$$\mathrm{X} = (1 - \bar{L}, \bar{A}, \bar{P}, \bar{R}, \bar{IoU}, \bar{F}1, 1 - \bar{T}) \tag{4}$$

To calculate the value of the loss function and the prediction time of a single image in (4), it is necessary to apply inverse normalization to provide a uniform direction for the interpretation of all metrics. The use of the transform $(1 - L)$ and $(1 - T)$ allows to include the values of loss and time in a single combined formula without logical contradictions between other metrics and without loss of informativeness.

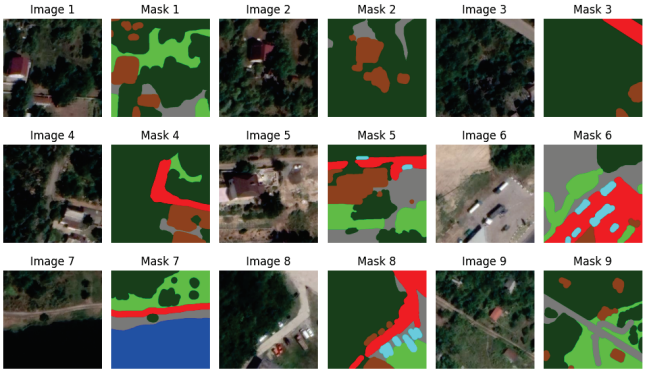


Fig. 1. Prepared image pairs (aerial image and corresponding segmentation)

Using the vectors of the weights W and normalized values X, it is possible to calculate a combined effectiveness score CES:

$$\text{CES} = \sum_{i=1}^{n} W_i X_i, \tag{5}$$

where n is the total number of weighting factors.

III. Experiments

A. Experimental Setup

For experimentally evaluate the operation of the proposed approach with the calculation of CES, our own image dataset was prepared, which consists of 50 pairs of aerial images and their corresponding segmentation masks, with a size of 512x512 pixels (Fig. 1).

For marking, 7 classes of objects were configured, such as buildings, roads, cars, land surface, dense and sparse vegetation, water bodies. To increase the summarizing power of the models and generalize the results, standard augmentation methods (mirroring, tilt angle, brightness, contrast, saturation and hue) were used. In this way, the dataset was expanded to 250 matching pairs of images.

For the correct analysis of the developed approach for calculating CES, the most popular deep learning architectures U-Net, DeepLabV3+, FPN and PSPNet were used.

The models were trained for 30 epochs, using the same parameters: the ”softmax” activation function, the ”Adam” optimizer with an initial speed of 0.0001 and the number of classes - 7. The models were trained on the same prepared dataset with a fixed distribution into a training and test sample (20% of all images).

The graphs of the training process of models without the use of augmentation (Fig. 2) and models with augmentation (Fig. 3) show that the training process took place correctly. The accuracy value gradually increases, and the loss function value steadily decreases for both the training and validation samples. It is possible to make sure that there is no overtraining in the models and they have sufficient generalizing capacity within the available data.

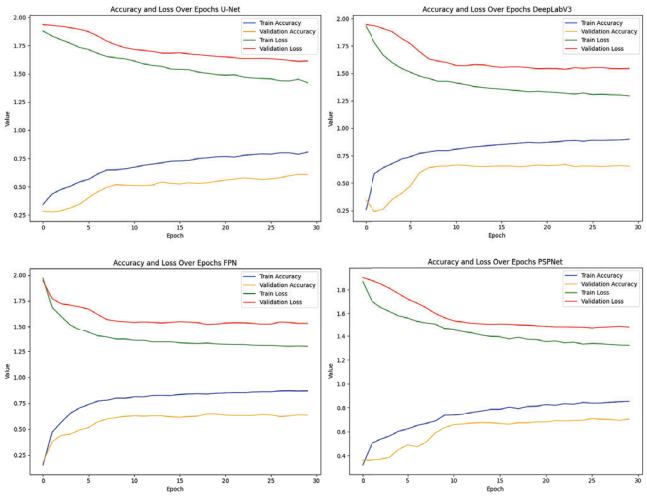


Fig. 2. Training models without using augmented data

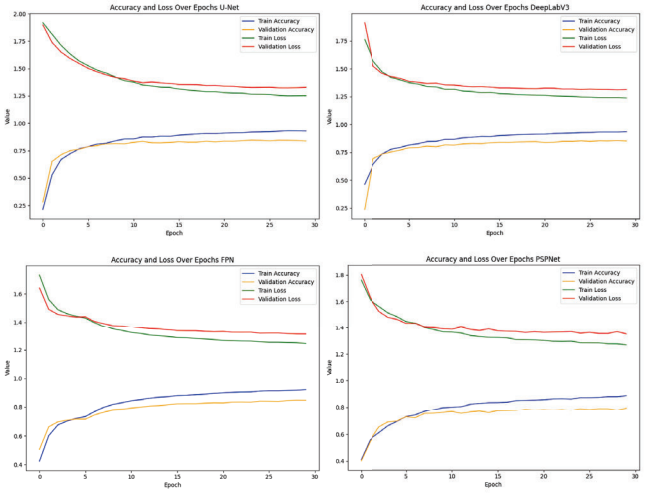


Fig. 3. Training models with using augmented data

To initially evaluate the effectiveness of the models on the test data, the loss function (cross-entropy) and the standard metrics of Accuracy, Precision, Recall, Jaccard index (IoU), F-measure (F1-Score) and prediction time of one image (Time) were calculated. The results of the models on test data without the use of augmentation are shown in table I, and with augmentation – in table II.

In this work, it was decided to use two sets of vectors of weights to calculate the combined effectiveness score of the U-Net, DeepLabV3+, FPN and PSPNet models:

- $W1 = (0.15, 0.2, 0.1, 0.1, 0.2, 0.15, 0.1)$;
- $W2 = (0.1, 0.2, 0.1, 0.1, 0.1, 0.2, 0.2)$.

The first vector of the weights shows a more general situation where all the weights are balanced and have almost the same effect on the metrics. The second vector of weights is focused on solving real-time problems, when reducing image processing time is critical and at the same time segmentation

TABLE I
OBTAINED METRIC VALUES ON TEST DATA WITHOUT AUGMENTATION

Metric	U-Net	DeepLabV3+	FPN	PSPNet
Loss	1.6082	1.5322	1.5243	1.4699
Accuracy	0.6083	0.6524	0.6362	0.7050
Precision	0.4626	0.5525	0.5144	0.5711
Recall	0.4160	0.4803	0.4648	0.5554
IoU	0.2917	0.3588	0.3408	0.4131
F1-Score	0.3907	0.4767	0.4626	0.5282
Time	0.1219	0.3193	0.1055	0.0356

TABLE II
OBTAINED METRIC VALUES ON TEST DATA WITH AUGMENTATION.

Metric	U-Net	DeepLabV3+	FPN	PSPNet
Loss	1.2960	1.2778	1.2854	1.3164
Accuracy	0.8784	0.8930	0.8803	0.8537
Precision	0.7615	0.7646	0.8610	0.7248
Recall	0.7462	0.7683	0.8307	0.7334
IoU	0.6611	0.6817	0.7249	0.5971
F1-Score	0.7419	0.7547	0.8321	0.6885
Time	0.1393	0.3162	0.1104	0.0371

accuracy remains important. The obtained CES values for the first set of weights are shown in the table III, and for the second in the table IV.

To improve the understanding of the results obtained, diagrams were constructed, respectively for the first set of weights (Fig. 4) and for the second (Fig. 5). With the help of such visualization, it is possible to visually compare the obtained CES values for each of the models.

B. Results and discussion

The results obtained show differences in the computational effectiveness of the considered segmentation models. According to the table III and table IV the PSPNet model shows the best values of the combined effectiveness score at the first set of weights, but when calculated on models trained with augmented data, the model is inferior to FPN. In the second set of weights, without augmentation, PSPNet again shows the best values, and with augmentation, the FPN model

TABLE III
CES FOR MODELS WITH THE FIRST SET OF WEIGHTS

Model	CES	CES with augmentated data
U-Net	0.3206	0.6113
DeepLabV3+	0.3641	0.6066
FPN	0.3732	0.6582
PSPNet	0.4371	0.5808

TABLE IV
CES FOR MODELS WITH THE SECOND SET OF WEIGHTS

Model	CES	CES with augmentated data
U-Net	0.4281	0.6800
DeepLabV3+	0.4439	0.6557
FPN	0.4765	0.7308
PSPNet	0.5412	0.6646

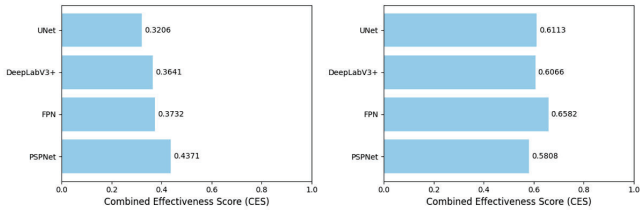


Fig. 4. CES for models without augmentation (1) and with augmentation (2)

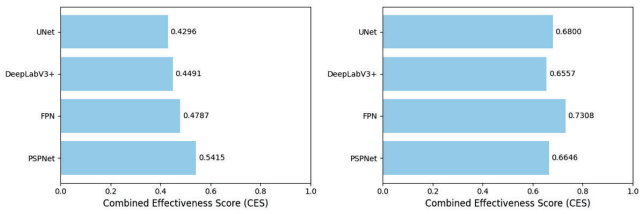


Fig. 5. CES for models without augmentation (1) and with augmentation (2)

remains in the lead. FPN shows the highest overall value of all experiments performed. Due to the highest image prediction time, DeepLabV3+ turned out to be less efficient, while the quality of other results remains quite high. The basic U-Net shows consistently lower performance on most metrics, emphasizing that it should be used with custom modifications.

It is important to note the positive impact of data augmentation on learning results. The use of augmentation methods helped to significantly improve the quality of segmentation of each of the models. According to the table I and table II, all computational metrics improved significantly as the dataset expanded. Thus, the use of augmentation made it possible to increase the generalization of the models and better adapt them to a variety of inputs, which is critically important in the practical tasks of the field of aerial image processing.

IV. CONCLUSIONS

In our paper, we propose to calculate a combined evaluation of the effectiveness of model of deep learning models for segmentation (U-Net, DeepLabV3+, FPN, PSPNet). Studies and experiments have confirmed the feasibility of calculating a single metric that combines not only the accuracy of the models, but also other important aspects. We used two sets of weights for the overall analysis of the quality of the models and for options where performance plays an important role. The results showed that the FPN model was the most balanced when using augmented data, and the PSPNet model demonstrated the best performance without augmentation.

The experiments also showed a significant impact of the use of augmentation methods on the overall effectivenes of the models. For the U-Net, DeepLabV3+ and FPN architectures, assessment metrics have increased significantly when trained on extended data, highlighting the importance of such training in segmentation tasks.

The limitation of our work is that we use only one dataset in the respective application scenario. Therefore, it is difficult to generalize the conclusions for different types of images in other problems. Also, the combined CES metric is convenient for comparing models, but contains subjectivity in the choice of weighting factors.

Future research focuses on expanding the experimental base by creating new datasets with aerial images of various types and creating a mechanism for automatic selection of weights using optimization algorithms.

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Research into Mathematical Models of Interdependent Relationships Between Biomedical Indicators and Treatment Effectiveness Indicators for Gunshot Wounds

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***Abstract*— This work belongs to the field of biomedical informatics and concerns mathematical modeling of biomedical processes in the surgical treatment of gunshot wounds.In the present work, a study of mathematical models of multiple regression between the coefficients of mobility, biomedical indicators and the number of bed-days of treatment of patients with inflammatory diseases was carried out. wounded. The purpose of such work lies in the theoretical selection of the most informative information space for predicting operation codes and the choice of treatment methods, and their experimental confirmation. The work is intended for practical support when planning and performing surgical operations on patients with inflammatory injuries, which will significantly improve the effectiveness of their treatment rehabilitation.**

Keywords*— *biomedical information technologie, statistical method for making medicinal decisions, flammable injuries, the effectiveness of surgical operations

I. INTRODUCTION

In their latest research, the authors analyzed the results of experimental medical data based on biomedical indicators obtained during primary restraint of patients with inflammatory injuries, and the results of the treatment performed. In this case, it was necessary to correlate the values of the parameters and formulate mathematical models of the processes of rehabilitation of patients with inflammatory injuries [1 - 3].

Research conducted by the authors shows that the number of indicators correlated with the importance of injury beyond the criterion of statistical significance is small [1]. The effectiveness of virtual (updated biochemistry) endocrinological indicators was provided. Now we need to prove that these virtual indicators are effective for predicting operation codes and choosing one of the models - scales A and B.

The analysis is based on a basic method of step-by-step investigation of mathematical models of multiply regression. The peculiarities of such a technique are the comprehensive combination of theoretical and experimental testing and the selection of the most informative scale options for predicting operation codes (A or B). To assess the information content of these scales, it is necessary to develop mathematical models of multiple regression. Regressors in such models require the use of biomedical indicators that have the maximum correlation with operation codes. The number of such indicators and their equal mutual correlations are the subject of investigation for the formation of plans for regression data processing.

The objective function of the efficiency of the regression procedure is due to the standard bias of the regression model. Verification of the statistical significance of the model using the additional Fisher criterion is a fundamental element of statistical research. In addition, Fisher's F-statistics, which is based on the variable analysis of variance procedure of the regression model, can be used for equalizing the information efficiency of the regression, in addition zi standard kidnapping.

The analysis is carried out in two stages. The first stage is theoretical research using the method of finding a mathematical model of the objective function, which characterizes the information efficiency of the regression procedure for updating (predicting) unknowns demonstrators. The other stage is the experimental determination of the target efficiency function for the optimal selection of biomedical indicators for predicting operation codes.

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II. Theoretically Advanced Effectiveness of Mathematical Models Regressive Forecasting

Endocrinological (hormonal) indicators are of particular value as indicators of the importance of injury. This, first of all, explains their significance in various regression models for predicting operation codes. Otherwise, they can be used to verify the effectiveness of the results of the theoretical synthesis of the mathematical model of the objective function of the forecasting efficiency, based on the results of the regression comparison, We have taken this work on the cob for the procedure of updating hormonal levels [2].

Let's take a look at the change of mind and the vikoristan of hormonal indicators for the first-choice vikoristan in extreme minds. In the context of war or mass disasters, the lack of time to relieve the flow of the wounded reduces the level of information security for medical personnel. A number of different biomedical indications can be obtained that do not represent a minimum of ease and difficulty in their identification. Most often, there are indicators of hormonal imbalance, which are more informative about the severity of the injury. Updating such indicators is a priority for assessing the status of the injured when applying the nomenclature of biomedical indicators.

It appears that the endocrine system is a complex system that includes a number of equal regulation, including the central nervous system, secretory organs, tissues, cells, hormones, as well as clock parameters: monthly and seasonal cycles, circadian rhythms, Swedish types, etc. As a rule, this system is characterized by non-linear pathways, containing components from several gate loops and participating in cross-talk with other systems of the body. The endocrine axes are very dynamic in response to the level of hormones, showing a strong response to traumatic stress of a physical nature, which indicates a disruption in the regulation of endocrine processes [1].

Mathematical models of mutually related connections between informative variables and their influence on the risk of decision making are the basis for diagnostic procedures and forecasting in mathematical models of cybernetics [3].

Today, many models have been developed, such as cybernetic methods of mathematical prediction in neuroendocrinology using the growth hormone system [4], in the hypothalamic-pituitary neurosecretory system [5] etc. In addition, a number of mathematical modeling tools are described, such as equations, which describe the dynamic behavior of a system (for example, biostability, vibration), as well as approaches to working with biological noise and systems with several hour scales [6]. This demonstrates the growing scientific interest in the use of sophisticated tools and methods for tracking the complex dynamics of hormones, especially under stress, processes associated with reproduction and metabolism [7-12].

Unfortunately, the considered methods of mathematical forecasting in endocrinology concern, first of all, procedures where the weekend indicators are predicted to fall within the same class of hormonal indicators. In another way, it is better to think about mathematical modeling of the dynamic regimes of the endocrine system, so that we can come closer to identifying the influences that parametrically lie in the form of nonstationarity of the modeled hormonal processes.

As soon as we move on to methods and mathematical models for updating biomedical data, then the interest in methods for updating single data [13 - 15], the knowledge of such data, arises. However, such methods are not limited to databases, which are based on the statistical complexity of the update procedure.

There are few such methods of regression updating [16, 17]. The use of ML machine learning procedures in such models on the H20 platform [18] and Movi R [18] made it possible to statistically optimize the parameters of the regression model. Moreover, updating is more effective in relation to large databases [19, 20], in which lacunae are tidied up, so that value is lost.

It should be said that the updating of information in databases has greatly contributed to the development of the mathematical apparatus of Bayesian measures [21]. The rest of the research is being used to formulate a new structure in search of diverse (heterogeneous) medical information behind different documents. It is also important for such investigations to provide benchmark assessments of the reliability of the update, which can be used for a regular analysis of the effectiveness of new update procedures [22]. However, to start such Bayesian models, a large initial sample is required to determine a priori probabilities for all variables of such measures.

If we consider a practical methodology for solving the problem of updating daily data in statistics, then the most advanced one is robot [23]. Vaughn describes not only explicit methods and mathematical models of large-scale analysis of databases with daily values, but also proposes a subsequent theory based on the maximum likelihood function. However, if we talk about forecasting as a high-order extrapolation, then the problem disappears, especially when it comes to planning the selection of parameters of the update model and the overall assessment of its effectiveness.

In fact, the problem of theoretical prediction a priori of the effective version of the chosen model of updating is not solved. The problem of verifying the effectiveness of such a model only on the initial sample, without changing the control sample, has not been resolved. The motivation for the investigation is the extraction of additional diagnostic hormonal information in the minds of military-field surgery.

For the analysis, we will consider the array of initial selection as a multiplicity of results and the value of biomedical indicators, presented in a table of rows (number of injured) and results (number of biomedical indicators that will be discussed).

Biomedical indications can be divided into three groups:

• A subset of symptoms $Z = \left\{ Z_1, ... Z_{N_1} \right\}$ (for example, hormonal symptoms).

• A subset of output potential indicators $X = \left\{ X_1, ... \mathrm{X}_{N_2} \right\}$ - regressors, suitable for updating any of the indicators.

• A single (single-element) subset $Y = \left\{ Y \right\}$ that includes the mentally integral indicator “The severity of the bath,” which has a direct correlation with the severity of the injury.

The designation of the elements of the reinsurance subsets by capital (great) letters of the Latin alphabet indicates that all biomedical indicators are considered as random values. The submultipliers themselves can be viewed as phase vectors, skin warehouses, which are characterized by a mean value vector and a correlation matrix. Realizations (virtual meanings) are signified by small letters of the Latin alphabet: X, Y, Z.

For a fair modeling of procedures for updating one of the hormonal indicators of the multiplicity Z and testing the result of updating for its effectiveness, we conclude the operation of centering and normalizing all implementations for indicators of the subsets X, Y, Z. In this case, the average values and dispersion of all implementations of the increase in temperature are obviously 0 and 1.

It is possible to remove arrays of linearly transformed implementations to induce rich factor standardized responses (K – number of officials) [24-27]:

$$\hat{z}_i = \sum_{j=1}^{K} B_j x_{ij} \text{ ; where , } K \le N_2 ; \quad (1)$$

$$\hat{y}_i = \sum_{j=1}^{K} b_j x_{ij} ; \quad (2)$$

$$\hat{y}_i = \beta \hat{z}_i , \quad (3)$$

where: B_j, b_j, β - angular coefficients of standardized linear regressions ($j = \overline{1, K}$).

In standardized regressions (1) - (3) all angular coefficients are functions of only pairwise normalized linear correlation coefficients R_{zx_j}, R_{yx_j}, $R_{y\hat{z}}$. In particular $\beta = R_{y\hat{z}}$.

Let's find the mathematical expectation for point estimates of residual variances for regressions (2) and (3):

$$\begin{cases} \sigma^2_{(2)} = \lim\limits_{n\to\infty}\left[\dfrac{1}{n-1}\sum\limits_{i=1}^{n}\left(y_i - \sum\limits_{j=1}^{K} b_j x_{ji}\right)^2\right]; \\ \sigma^2_{(3)} = \lim\limits_{n\to\infty}\left[\dfrac{1}{n-1}\sum\limits_{i=1}^{n}\left(y_i - \beta\hat{z}_i\right)^2\right]. \end{cases} \quad (4)$$

Let's consider two options for this: $K = 1$ and $K = 2$. In the first option, one regressor is used to restore the indicator values, and for the second option - two regressors x_1 and x_2.

We use the difference:

$$\bar{\Delta} = \sigma^2_{(2)} - \sigma^2_{(3)}, \quad (5)$$

as the target function of restoration efficiency to test the restoration model for the presence or absence of efficiency.

The solution will be determined by the sign of the difference:

- Solution γ_0 (restoration is effective), if $\bar{\Delta} > 0$;
- Solution γ_1 (restoration is effective), if $\bar{\Delta} \le 0$.

Option 1 ($K = 1$). For a single regressor, for example conditionally x_1:

$$\bar{\Delta} = -\left[R_{yx_1} - \left(R_{y\hat{z}} R_{zx_1}\right)\right]^2 \le 0 \cdot \quad (6)$$

Effective recovery of the indicator Z by only one indicator from the subset X is impossible, since always in $\bar{\Delta} \le 0$.

Option 2 ($K = 2$). For two regressors x_1 and x_2 the function $\bar{\Delta}$ (equation (5)) takes the form:

$$\bar{\Delta} = \left(b_1 - R_{yx_1}\right)^2 + \left(b_2 - R_{yx_2}\right)^2 + 2b_1 b_2 R_{x_1x_2} + 2R_{y\hat{z}} R_{yz} - R^2_{y\hat{z}} - R^2_{yx_1} - R^2_{yx_2}, \quad (7)$$

where: $R_{x_1x_2}$ - linear normalized correlation coefficient between x_1 and x_2;

b_1 and b_2 - standardized regression coefficients:

$$b_1 = \frac{R_{yx_1} - R_{yx_2} \cdot R_{x_1x_2}}{1 - R^2_{x_1x_2}} ; \quad (8)$$

$$b_2 = \frac{R_{yx_2} - R_{yx_1} \cdot R_{x_1x_2}}{1 - R^2_{x_1x_2}} . \quad (9)$$

The sign "^" above the implementation indicates Z its recovery in the form of regression.

The appearance of the cross-correlation coefficient $R_{x_1x_2} \ne 0$ indicates the possibility of potential recovery of the indicator Z. When $R_{x_1x_2} = 0$ recovery is potentially effective, if only the condition is met:

$$R_{y\hat{z}} > \sqrt{R^2_{yx_1} + R^2_{yx_2}} . \quad (10)$$

To verify the statistical adequacy of the mathematical models of the objective function with real experimental data, the database of the emergency medical clinic. The overall dataset included biomedical indicators $N = 38$ (8 of which were hormonal) for wounded with soft tissue shrapnel injuries.

From the general array, $N_1 = 4$ hormonal indicators were isolated, $N_2 = 5$ - "indicators - regressors" and an integral indicator of the severity of the injury. The resulting array included hormonal and general medical indicators with the maximum possible mutual correlation, exceeding the critical values of 0.361 (for $n = 30$).

The list of four restored hormonal indicators is given in table. 1.

TABLE I. INDICATORS USED IN THE MODEL EXPERIMENT

Set	Elements of the set	Name of the indicator	Purpose of the indicators
Z	Z_1	Dehydroepiandrosterone, DHEA	Recoverable (dependent variables)
	Z_2	Free thyroxine, S-fT4	
	Z_3	Testosterone, TESTO	
	Z_4	Adrenocorticotropic hormone, ACTH	
X	X_1	Wound volume	Independent variables (regressors)
	X_2	Prothrombin index, PI	
	X_3	Blood aspartate transferase, AST	
	X_4	Blood alanine transferase, ALT	
	X_5	Erythrocytes, RBC	
Y	Y	Bed-day	Integral indicator

Fig. 1 presents the elements of the correlation matrix (triangular), which additionally includes pairwise correlation coefficients for the restored $\hat{Z}_1$, $\hat{Z}_1$, and and hormonal indicators. Statistically insignificant correlation coefficients are shaded (for the significance level $\alpha = 0.05$ and $n = 30$.

	Y	X_1	X_2	X_3	X_4	X_5	Z_1	Z_2	Z_3	Z_4
Y	1									
X_1	0.729	1								
X_2	-0.435	-0.461	1							
X_3	0.246	0.010	-0.092	1						
X_4	0.197	0.066	0.011	0.640	1					
X_5	-0.446	-0.370	0.435	-0.262	0.042	1				
Z_1	-0.769	-0.875	0.493	0.079	-0.098	0.384	1			
Z_2	0.770	0.875	-0.452	-0.052	0.084	-0.409	-0.963	1		
Z_3	-0.645	-0.281	0.322	-0.485	-0.389	0.278	0.248	-0.226	1	
Z_4	0.837	0.546	-0.367	0.345	0.321	-0.377	-0.572	0.540	-0.893	1

Fig. 1. Correlation matrix for all indicators used in modeling recovery procedures.

The canonical equations of two-factor pure regression for the recovered indicators are given in Table 2. The regressors and elements of the set used are listed in Table 1.

TABLE 2 TWO-FACTOR REGRESSION RECOVERY MODELS

Recoverable indicator	Regressors used	Recovery model
Z_1	X_1 and X_2	$\hat{z}_1 = 70.522 - 0.055x_1 + 0.446x_2$
Z_2	X_1 and X_2	$\hat{z}_2 = 20.156 - 0.007x_1 - 0.0298x_2$
Z_3	X_3 and X_4	$\hat{z}_3 = 4.248 - 0.032x_1 - 0.007x_2$
Z_4	X_1 and X_5	$\hat{z}_4 = 74.111 + 0.024x_1 - 5.217x_2$

Table 3 presents the results of the theoretical calculation of the objective function for all four restored indicators. The standard errors for the linear regression of the values of the integral indicator on the restored indicators of the set and the restoration errors on the original biochemical indicators - regressors are also given here.

TABLE 3. RESULTS OF THEORETICAL AND EXPERIMENTAL ASSESSMENT OF FORECASTING EFFICIENCY

Indicator	Regressors	Theoretical calculation	Experimental estimate of the standard error of prediction	
		$\overline{\Delta}$	By set regressors, X, $\sigma_{y\|x}$	By restored indicator, $\hat{Z}$, $\sigma_{y\|\hat{z}}$
Z_1	X_1; X_2	+ 0.047174	6.361	6.2076
Z_2	X_1; X_2	+ 0.007062	6.316	6.2322
Z_3	X_3; X_4	+ 0.191757	9.053	8.890
Z_4	X_1; X_5	+ 0.127786	6.152	6.0653

Any multivariate (multiple) regressions are, unfortunately, functions not only with random parameters (coefficients), but also with random arguments (regressors). Such functions are based on the use of a priori assumptions about Gaussian probability distributions for all random biomedical indicators included in the database system. In fact, any regression has a minimum uncertainty of its values only within the framework of the training data set.

Formally, the decision on the statistical significance of the regression (endocrinological indicator) is made using the results of its variance analysis using the Fisher's criterion - statistics. But to prove its information significance, it is necessary to use the criterion of the global minimum for the standard error of the auxiliary regression on the control predicted indicator.

The regressor in such a model is a restorable endocrinological indicator, as a polynomial regression function from biochemical "indicators - regressors". Providing a global minimum of the standard error is not the only condition. The second condition is that this minimum must be less than the standard error of the regression of the control integral indicator on the initial biochemical “indicators - regressors” selected for restoration.

The verification of the first and second conditions must be carried out purposefully, and the information criterion must be the mathematical expectation of the difference standard error for the regressions of the auxiliary integral indicator on the restoration and biochemical indicators. The sign and magnitude of such a criterion must qualitatively and quantitatively reflect the possibility of fulfilling the first and second conditions. Such a complex criterion can be used, at least by sign, for planning, based on a single training sample, procedures for selecting the restoration indicator and indicators - regressors. And, most importantly, such a criterion as the planning objective function must have as arguments a finite set of correlation coefficients reflecting the mutual stochastic relationship between all indicators involved in planning the recovery procedure.

Creating a probabilistic mathematical model of such an objective function is a problem, especially if the mutual stochastic dependencies are nonlinear for the indicators under study. However, its solution for linear polynomial regression models is theoretically not difficult. This is important when planning the selection of subset indicators and evaluating the effectiveness of the selection by the sign of the objective function, using only the training sample.

III. CALCULATION OF CORRELATION MATRICES AND SYNTHESIS OF REGRESION MODELS FOR EVALUTION OF OPERATION MODULE CODES.

In the course of the study, correlation matrices were calculated, on the basis of which subsets of medical indicators of the initial examination of the wounded were formed, which had a significant correlation with the codes of the operational modules for scale A (Table 4) and scale B (Table 5).

The optimization of the subset of indicators was carried out according to the minimum value of the function - "standard error" of the regression model. For each variant of the scale, optimal regression models of the initial assessment of the operational module code were synthesized.

The optimal regression equation for scale ***A***:

$$Y = 2{,}5657 - 0{,}2877X_1 + 0{,}03175X_2 - 0{,}2939X_3 + 0{,}1744X_4, \quad (11)$$

where: X_1 - Number of wounds.

X_2 - Aldosterone, (pg/ml).

X_3 - T4 (pmol/l).

X_4 - Wound temperature (°C) on admission.

Y - Operation code.

TABLE 4. RESULTS OF THE ANALYSIS OF VARIANCE OF MULTIPLE REGRESSION OF OPERATION CODES (SCALE A) (I, B, AG, AH, E)

Regression Statistics								
Multiple R	0,354218							
R Square	0,12547							
Adjusted R Square	0,094237							
Standard Error	1,204855							
Observations	117							
ANOVA								
	df	***SS***	***MS***	***F***	***Signific. F***			
Regression	4	23,3268	5,831693	4,01721	0,004404			
Residual	112	162,588	1,451676					
Total	116	185,915						
	Coefficients	***Standard Error***	***t Stat***	***P-value***	***Lower 95%***	***Upper 95%***	***Lower 95,0%***	***Upper 95,0%***
Intercept	2,56573	5,38660	0,476317	0,63477	-8,10714	13,2386	-8,10714	13,2386
X Variable 1	-0,28771	0,07948	-3,61956	0,00044	-0,44521	-0,13022	-0,44521	-0,13022
X Variable 2	0,03177	0,03201	0,992555	0,32306	-0,03166	0,095205	-0,03166	0,095205
X Variable 3	-0,29395	0,25059	-1,17302	0,24327	-0,79047	0,202568	-0,79047	0,202568
X Variable 4	0,17435	0,15766	1,105839	0,27116	-0,13804	0,486738	-0,13804	0,486738

Optimal regression equation for scale B:

$$Y=-5,6978-0,0137X_1-0,0606X_2+0,8216X_3+0,3612X_4+0,0069X_5, \quad (12)$$

where: X_1 - wound area (cm2).

X_2 - Aldosterone, (pg/ml).

X_3 -T4 (pmol/l).

X_4 - Triglycerides-ser (mmol/L).

X_5 - Wound temperature (°C).

Y - Operation code.

The results of the analysis of variance of mathematical models (11) and (12) are presented in tables 4 and 5.

TABLE 5. RESULTS OF THE ANALYSIS OF VARIANCE OF MULTIPLE REGRESSION OF OPERATION CODES (SCALE B) (I, C, AG, AH, AM, AI).

Regression Statistics								
Multiple R	0,364922							
R Square	0,133168							
Adjusted R Square	0,094122							
Standard Error	1,945689							
Observations	117							
ANOVA								
	df	***SS***	***MS***	***F***	***Signific. F***			
Regression	5	64,55577	12,91115	3,4105	0,006637			
Residual	111	420,2135	3,785707					
Total	116	484,7692						
	Coefficients	***Standard Error***	***t Stat***	***P-value***	***Lower 95%***	***Upper 95%***	***Lower 95,0%***	***Upper 95,0%***
Intercept	-5,69777	5,80582	-0,98139	0,328535	-17,2024	5,806855	-17,2024	5,806855
X Variable 1	-0,01368	0,006264	-2,18449	0,031029	-0,02609	-0,00127	-0,02609	-0,00127
X Variable 2	-0,06061	0,05514	-1,09919	0,274065	-0,16987	0,048654	-0,16987	0,048654
X Variable 3	0,821606	0,566174	1,451155	0,149557	-0,30031	1,943517	-0,30031	1,943517
X Variable 4	0,361187	0,179829	2,008496	0,047018	0,004843	0,717531	0,004843	0,717531
X Variable 5	0,006918	0,003473	1,991761	0,048853	3,54E-05	0,0138	3,54E-05	0,0138

To assess the predictive effectiveness based on the results of variance analysis, comparisons were made between scales A and B according to statistical parameters:

σ - standard error of the regression model;

F - Fisher's dispersion ratio statistics;

F_{cr} - critical value of Fisher's statistics for the significance level $\alpha = 0.05$.

Total sample size $N = 117$.

Number of regressors for scale A ($m = 4$), B ($m = 5$).

To assess the predictive efficiency based on the results of the analysis of variance, comparisons of scales A and B were made according to statistical parameters:

σ - standard error of the regression model;

F - Fisher's dispersion ratio statistic;

Fcv - critical value of Fisher's statistic for the significance level $\alpha = 0.05$.

$N = 117$ - total sample size.

Number of regressors for scale A (m=4), for scale B (m=5).

The results of the comparison of scales A and B according to statistical parameters are presented in Table 6.

TABLE 6. RESULTS OF THE ANALYSIS OF VARIANCE ACCORDING TO FISHER'S CRITERION (SCALES A AND B).

Scale	σ	F -	F_{kv}	**Real level of significance (risk of the 1st kind)**
A	1,2048	34,0172	2,46	0,0044
B	1,9457	3,4105	2,30	0,0066

Since the standard error (σ) of scale A is smaller than that of scale B (1.2048 < 1.9457), we choose option A of the scale of operation codes. This conclusion is also confirmed by comparison with other statistical parameters:

• For Fisher statistics (34.0172 > 3.4105).

• For the real risk of the 1st kind (0.0044< 0.0066).

It should be noted that both scales are statistically significant, since their Fisher F-statistics exceed the corresponding critical statistics Fcr .

To theoretically confirm the effectiveness of scale A, in comparison with scale B, the methodology of section 2 of this report was used.

Biomedical indicators were divided into three groups:

• Subset Z (one indicator - operation code, presented in two variants Z_A and Z_B, according to the variant of the code scale.

• Subset $X = \{X_1, X_2\}$, where X_1 - aldosterone, X_2 - $T4$.

• Single-element subset Y - indicator "bed days".

Calculations according to equations (1) - (5), for K = 2 and the values of the correlation coefficients of the matrices of tables 4 and 5, gave two values of the target efficiency function for scales A and B:

• $\Delta_A = 0.03366$, scale A.

• $\Delta_B = 0.0157$, scale B.

Although for both scales the target function $\Delta > 0$, but $\Delta_A > \Delta_B$, which confirms the conclusions based on the results of the analysis of variance.

IV. CONCLUSION

The conducted study proves the possibility of obtaining additional diagnostic hormonal information in the conditions of military field surgery. Also in the process of performing the work, the following results were obtained:

• The possibility of theoretical analysis of probably justified options for choosing restored and initial biomedical indicators based on incomplete medical databases has been proven.

• The possibility of synthesizing a mathematical model of restoring the objective function, depending on the coefficients of the correlation matrix of the system of used indicators, including an integral control indicator correlated with the severity of the injury, has been proven.

• It has been proven that the sign and magnitude of the objective function a priori indicate some possibilities and effectiveness of regression restoration.

• The adequacy of the application of the objective function has been verified on the example of restoring four hormonal indicators based on two general medical indicators, on a limited database for wounded people with shrapnel injuries of varying severity.

• Correlation matrices of biomedical indicators were calculated for two variants of scales A and B, which present operation codes taking into account their weight coefficients.

• Optimization of regression models for predicting operation codes with a minimum standard error, presented in scales A and B, was performed. The results proved that scale A is the most accurate for predicting codes.

• Calculations of the target function of the model's efficiency were theoretically confirmed, that the information efficiency of scale A, for which regressors were two hormonal indicators "Aldosterone" and "T4"; recovery indicators - "operation code" and "bed days" - is more effective and can be used to predict the treatment process and rehabilitation after surgical treatment of gunshot wounds.

Thus, the effectiveness of the proposed method of optimizing the treatment and rehabilitation process of the population with gunshot wounds using mathematical statistics methods has been theoretically substantiated and practically proven.

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Beeswax and Propolis for Active and Intelligent Food Packaging: A Sustainable Approach

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***Abstract*—The growing need for sustainable, active and intelligent packaging systems has catalyzed research into natural biomaterials such as beeswax and propolis. Beeswax provides excellent barrier properties to the packaging and can act as a matrix for embedding nanoparticles and sensors, while propolis has powerful antimicrobial and antioxidant functions. This combination improves the preservation of food, extends its shelf life, and allows for real-time monitoring of product quality. This article discusses the synergistic integration of beeswax and propolis in the development of active and intelligent food packaging. The most common problems that arise in the development of packaging materials for bee products and possible ways to overcome them are also discussed. This review outlines future research directions to optimize beeswax -propolis composites and create active and smart packaging solutions.**

Keywords—beeswax, propolis, nanochemical sensors, smart packaging, food safety, biodegradable materials

I. Introduction

Food packaging has evolved from a simple protective layer to an active component in preserving and ensuring food safety. Active and intelligent packaging systems now incorporate sensor technologies and are able to prevent product spoilage by creating the right conditions inside the package and detecting indicators of spoilage and environmental changes, and bee products such as propolis and beeswax (BW) are promising agents and components for such systems (Fig. 1) [1,2,3]. Propolis has a wide range of pharmacological effects and, mainly in the form of water-alcohol extracts, is included in medicines and cosmetics as an antimicrobial, antifungal, antiviral, anti-inflammatory, wound healing and reparative agent. This ingredient is widely used in the food industry and is considered safe and non-toxic. Due to its proven antimicrobial activity against more than 100 strains of microorganisms, propolis is used in the development of natural and biodegradable packaging materials. In numerous studies, the inclusion of propolis extracts in food packaging products or biopolymer film matrix has shown the possibility of obtaining active packaging films that are safe and environmentally friendly compared to plastic packaging films [4, 5, 6, 7, 8].

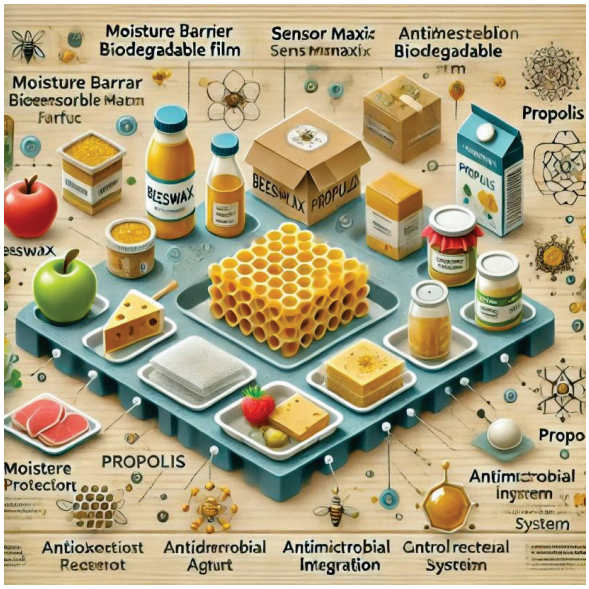


Fig 1. Possible applications of BW and propolis in food packaging.

BW-based packaging can also be a good alternative due to its many advantages. One of the key advantages of BW is its excellent film-forming ability, which facilitates the creation of thin, flexible coatings that can serve as substrates for sensor integration [9,10,11].

Recent progress in the development and production of food films/coatings containing BW (e.g., emulsion and blended films, bilayers, bionanocomposites, and antimicrobial materials) has shown that its incorporation into polysaccharide and protein-based emulsion films effectively improves their hydrophobicity, barrier properties against water vapor, UV and visible light, and tensile strength [2]. BW has also been found to be effective against gram-positive and gram-negative bacteria, as well

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as fungi. The antimicrobial effects of BW have been found against bacteria from the genera *Bacillus, Escherichia, Listeria, Proteus, Pseudomonas, Salmonella, Staphylococcus*; yeasts from the genera *Candida, Rhodotorula*; and molds from the genera *Aspergillus* and *Geotrichum* [9].

With the development of nanotechnology, the functional characteristics of these bee products have been significantly improved. Nanoparticles derived from or based on propolis can be designed to increase the surface area of interaction, improve the diffusion properties of active substances, and provide controlled release over time [12,13,14,15]. Nanoemulsions, solid lipid nanoparticles and nanocapsules have been investigated as propolis delivery systems in packaging [11,16]. These nanostructures facilitate the integration of propolis into sensor films that are able to detect changes in environmental parameters such as temperature changes or gas emissions, etc. that signal spoilage. And the combination of BW and nanocrystals allows for the creation of films with improved mechanical and barrier properties [17].

Nevertheless, there are also problems related to technological solutions for such systems and consumer health safety [11]. For example, the chemical composition of beeswax and propolis can vary depending on climatic, environmental and cultivation conditions, which makes it difficult to standardize products containing them. The strong natural aroma of propolis can affect the sensory perception of packaged products, which requires odor masking strategies or formulation adjustments. There is also the potential for allergic reactions to bee products in sensitive consumers, which should be evaluated before widespread introduction [18].

This article aims to explore the comprehensive potential of propolis and BW in active and smart packaging, with a special focus on its role in sensing systems. We discuss the current state of research, highlight innovative applications, and identify future directions for material optimization and system integration. By combining traditional knowledge with advanced technologies, packaging solutions based on bee products can redefine food safety standards and contribute to a more sustainable global food system.

II. Discussion

A. BW in Food Packaging

BW, a natural secretion of honey bees, has characteristics that make it a very attractive agent for modern packaging systems. Its hydrophobic nature and semi-crystalline structure provide effective barrier properties that are critical to maintaining the internal environment of food packages, preserving the freshness and quality of the food [2,9,10,11].

One of the key advantages of BW is its excellent film-forming ability, which facilitates the creation of thin, flexible coatings with improved properties, examples of which are shown in Table 1. The most commonly used film formers are gelatin, chitosan, starch in various modifications and pectin, various alginates and polylactides. Film composites may include various auxiliary substances, such as plasticizers, to improve flexibility and reduce film brittleness. (glycerin, propylene glycol), and surfactants (e.g., lecithin) are used to stabilize the BW in aqueous matrices, in addition, stabilizers and antioxidants are added to prevent degradation of the material under the influence of light and temperature and to protect against oxidation, and fillers (dyes, pigments, etc.) can be added to the composition to impart certain physical properties. The most common methods of producing films are emulsification, casting, and drying.

TABLE I. BW-based composite films

Composite System	Fabrication Method	Key Properties	Applications	Reference
BW + Gelatin	Melted BW emulsified into aqueous gelatin, cast and dried	Water resistance, mechanical strength, biodegradable	Food packaging, coatings	[19]
BW + Chitosan	Chitosan dissolved in acid, BW homogenized in, cast and dried	Antimicrobial, oxygen/moisture barrier, biodegradable	Food packaging, medical films	[4.20,21, 22]
BW + Pectin	Pectin dissolved in water, BW added with stirring, cast into films	Flexibility, edible, biodegradable	Edible coatings, food films	[23, 24,25]
BW + Starch	Starch gelatinized, BW emulsified in, cast into films	Transparency, mechanical resistance, shelf life	Food preservation, wrapping	[26,27, 28]
BW + Alginate	Alginate dissolved in water, BW emulsified, optionally cross-linked	Gas barrier, biocompatible, strong mechanical strength	Food packaging, biomedical uses	[29,30, 31]
BW + PLA	solution casting method PLA/lignin films coated with BW by dipping or rod coating	Better the UV protection properties, water vapor permeability and water contact angle, antimicrobial properties.	Food packaging, coatings	[32,33]

These coatings can be used to embed various types of nanomaterials, including silver nanoparticles, carbon nanotubes, and cellulose nanocrystals. Such films can effectively protect foods from moisture and gases, extending their shelf life, and act as functional elements in sensors designed to detect spoilage markers such as ethylene, ammonia, or volatile organic compounds (VOCs) [16,34,35].

Studies have shown that the addition of silver nanoparticles to BW and polylactide (PLA) films imparts antimicrobial properties. These films effectively inhibit the growth of bacteria such as E. coli and Staphylococcus aureus, making them promising for active food packaging [5].

Another study has shown that the integration of carbon nanotubes into BW improves its thermal conductivity. This makes it possible to create films with improved thermal insulation properties that can be useful in heat-sensitive packaging [34]. The combination of BW and cellulose nanocrystals allows for the creation of films with improved mechanical and barrier properties. Such films can effectively protect food products from moisture and gases, extending their shelf life [36].

Despite these advantages, there are still problems with ensuring the homogeneity and reproducibility of BW films for sensors. The natural variability of composition and microstructure can affect the consistency of the sensor response [11,18,34]. Thus, further work should focus on improving BW processing methods and investigating its interaction with various nanomaterials to optimize sensor performance.

B. Propolis as an Antimicrobial Agent

Propolis is recognized as a powerful natural antimicrobial and antioxidant agent, making it an invaluable ingredient in food packaging. Its chemical composition includes more than 300 identified compounds such as flavonoids, phenolic acids, esters and terpenoids, many of which have a broad spectrum of action against bacteria, fungi and viruses [6,37,38].

The antimicrobial mechanism of propolis is to destroy microbial cell walls, interfere with enzyme systems, and inhibit DNA synthesis. These actions prevent microbial growth on food surfaces, reducing spoilage and increasing food safety. In addition, its antioxidant capacity helps prevent lipid oxidation and nutrient degradation, which further extends shelf life [39,40,41]. Different natural agents as propolis is used in the development of natural and biodegradable packaging materials [17,42]. In numerous studies, the inclusion of propolis extracts in food packaging products or biopolymer film matrix, such as chitosan, starch, κ- carrageenan, gelatin, agar and biopolymers, has shown the possibility of producing active packaging films that are safe and environmentally friendly compared to petroleum-based plastic packaging films [10,43,44,45].

When added to packaging films, propolis imparts these bioactive properties to the material, allowing it to function as a self-sterilizing surface. In BW-based composites, propolis not only enhances the antimicrobial effectiveness, but also improves the mechanical integrity and barrier properties of the film. The compatibility of propolis with other biopolymers, such as chitosan, alginate, and gelatin, allows for the creation of multilayer systems with individual release profiles [10,43,44,45].

However, integrating propolis into packaging materials also requires careful formulation development. Its strong aroma and potential allergenicity must be controlled by encapsulation or mixing with neutralizing agents. In addition, standardization of propolis extracts is essential to ensure consistent characteristics of different batches.

In general, the use of propolis as an antimicrobial agent is a natural, effective and sustainable approach to improving the functionality of food packaging.

C. Nanoformulations and Sensor Integration

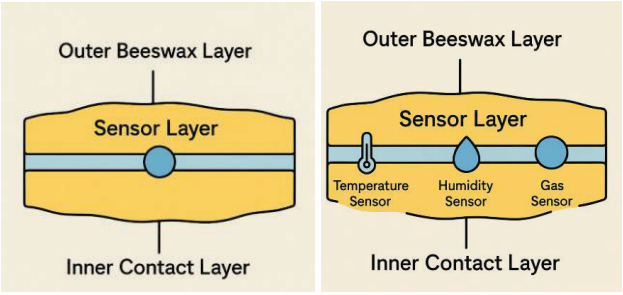
The convergence of nanotechnology with food packaging has opened up new avenues for the integration of chemical and electronic sensors into packaging materials. BW and propolis-based nanoformulations play a key role in this process by providing a functional environment for embedding sensors and improving physicochemical interaction with target analytes [6].

Among the latest innovations is the development of nanopropolis formulations that improve the solubility and dispersion of propolis in polymer matrices (Table 2). Nanopropolis is a nanoscale (~ 1-100 nm in diameter) propolis particles that are trying to make more effective without changing its properties by changing the size of propolis by various methods (encapsulation of propolis extract in nanoparticle systems - gold nanoparticles, silver nanoparticles, magnetic nanoparticles, liposomes, liquid crystal compositions, solid lipid nanoparticles, mesoporous silica nanoparticles, etc.). These nanoformulations allow for more efficient interaction with target targets and provide controlled release of active compounds, while maintaining antimicrobial action for a long time [9,15,46]. Nanoemulsions and nanocapsules with propolis can improve the dispersion of active compounds in polymer matrices. A particularly promising area is the use of propolis-based nanocomposites in combination with thin-film sensors. In combination with sensor systems, these nanocarriers can increase signal stability and sensitivity [47,48]. Propolis not only contributes to antimicrobial protection, but also increases the durability and functional life of sensor films [29].

TABLE II. NANOFORMULATIONS AND SENSOR INTEGRATION IN FOOD PACKAGING

Example of Sensor System	Active Substances	Results (Effectiveness)	Reference
PLA-based film with nano-propolis	Nano-propolis, polyvinyl alcohol (PVA)	Improved antimicrobial activity against *E. coli* and *S. aureus*, flexible, biodegradable	[15]
Propolis nanoemulsion with colorimetric indicator	Nano-propolis, bromocresol green	Detected pH changes during spoilage, visual color change, prolonged shelf life	[46]
Chitosan-BW-propolis film with VOC sensor	Chitosan, BW, nano-propolis, gas dye	Detected ammonia and H_2S, inhibited microbial growth, preserved texture and aroma	[1]
Electrospun fibers for dairy packaging	Nano-propolis, electrospun PVA	Uniform fiber matrix, antifungal action against *Candida* spp., stable under refrigeration	[45]
Smart label with encapsulated propolis and RFID	Liposome-encapsulated propolis, RFID module	Monitored spoilage gases, enabled real-time tracking, antimicrobial surface	[9]
Biopolymer film with nano-propolis	Nano-propolis, cellulose/chitosan	Extended shelf life, strong antimicrobial barrier, suitable for fresh produce packaging	[30,34]
Intelligent packaging with nano-propolis sensors	Nano-propolis, pH-sensitive dyes	Real-time detection of food spoilage; enhanced product safety and shelf life monitoring	[13,48]

BW is an ideal material for sensor systems due to its film-forming ability and chemical inertness. Principal schemes of such sensor systems are shown in Fig. 2.



a b

Fig 2. BW Wraps Combined with Sensors: (a) Single Sensor Layer in BW Packaging; (b) Multi-Sensor Layer in BW Packaging.

Typically, sensor packaging BW systems consist of three layers:

- BW outer layer (provides protection against external moisture, oxygen and contamination);

- Sensor layer (a thin functional layer embedded between the BW sheets containing one (a) or more (b) sensors (e.g. gas, temperature, humidity or pH indicator) that monitor food freshness or spoilage markers);

- Inner contact layer (food-safe coating that is in direct contact with the product.

Smart packaging based on BW can use different types of embedded or integrated sensors, depending on the task of the packaging (freshness, temperature, damage monitoring, etc.). The main types of sensors that can be used are temperature sensors (thin-film temperature sensors or thermal indicators based on polymers, nanomaterials). Examples of such systems are temperature-sensitive nanocomposites such as BW-carbon nanotube-paper or BW- graphene-paper [7]. Humidity sensors are presented in the form of hydrogel or nanocomposite film layers), and gas sensors can be created on the basis of metal nanoparticles (Ag, Au) or polymer matrices. An example is the use of colorimetric sensors that change color under the influence of gases (chitosan/BW - methyl red and bromocresol violet as color indicators - paper base) [18]. More advanced implementations include chemiresistive sensors that detect gaseous analytes and convert them into electrical signals that can be processed by embedded circuits for on-the-fly diagnostics [24, 34].

Biosensors represented by enzyme systems or DNA/antibody-based sensors are often used to detect specific pathogens, and upon contact with the pathogen, they bind to the target, initiating a signal (optical, electrochemical or color) [41]. Mechanical sensors, in turn, are subdivided into strain sensors based on changes in electrical resistance (BW + carbon nanotubes or graphene deposited on a flexible substrate); piezoelectric sensors, which generate an electric charge in response to mechanical pressure or vibration (BW composites with the inclusion of piezoelectric nanoparticles (e.g., ZnO, $BaTiO_3$), and optical pressure sensors, which respond to changes in the color or transparency of the package during deformation [27]. Moreover, the stability and performance of sensors can be improved in BW films. Studies have shown that BW- based films can maintain sensor activity for a long time, even in conditions of variable temperature and humidity. This durability makes BW particularly advantageous for packaging perishable foods such as fruits, vegetables, dairy, and meat products [45].

Technologies such as injection molding, spin coating, and inkjet printing have been successfully used to produce sensor packaging materials. Layer-by-layer assembly, electrospinning, and gas-phase deposition technologies are used to apply functional nanolayers to BW films [32].

Modern packaging systems with beeswax (BW) and propolis are perfectly integrated with computer technology, which allows for the creation of smart packaging that not only protects the product but also performs monitoring and data transmission functions

The combination of packaging systems and high digital technology has opened up a new era of sustainable and informative packaging. And RFID, NFC and wireless sensors can be integrated into beeswax (BW) systems [5, 48]. These technologies are integrated by placing an electronic module between the layers of BW or by gluing it to the film after molding. An example of such packaging is a BW + nanopropolis + colored gas sensor (e.g., bromocresol violet) with an active or passive RFID tag connected to the sensor. In such a system, information about the freshness status (e.g., ammonia level) is automatically read when the product passes through the reader.

Another example is NFC packaging with mobile reading, which consists of BW + propolis + NFC chip with a temperature sensor. The consumer can bring a smartphone to such packaging and see data on storage conditions (for example, if the product was overheated). Or an IoT package with a BLE (Bluetooth Low Energy) module, which consists of a BW film with an integrated multilayer sensor unit (temperature and humidity sensors + BLE module) that transmits data to the cloud or a mobile application in real time [5, 9,13, 17,29, 48].

Despite the promise of these technologies, scalability and cost remain challenges. Future efforts should focus on the development of low-cost, printable sensor components and improving nanocomposite formulations to balance performance with manufacturability.

D. Challenges and Limitations

Although beeswax and propolis have compelling advantages for smart packaging, their practical application is not without limitations. The main problem is variability - both substances are natural products whose composition can vary significantly depending on the type of bee, geographical location, and botanical sources. This variability affects their functional characteristics in sensory applications, potentially leading to inconsistent results. This problem can be minimized by using raw materials only from certified bee farms and implementing clear standards for the content of key components (phenols, flavonoids, etc.). In addition, it is advisable to use purified or fractionated extracts, or to replace natural products with synthetic analogues in inactive areas. The influence of aromatic impurities or allergenic fractions of wax and propolis can be minimized by microencapsulation and nanoformulation, which will also stabilize the active components and ensure controlled release. Thermal and mechanical stability is another important requirement. Pure

BW has a relatively low melting point (~62°C), which limits its use in high-temperature packaging environments. Mechanical brittleness can also beproblematic, especially during handling or transportation. These problems can be solved by adding plasticizers, polymer blends, or reinforcement with nanomaterials [12]. Cost is also an important factor. High-purity propolis and filtered BV can be expensive, and large-scale production may not always be economically feasible without subsidies or market incentives. In addition, the integration of sensors and IoT components requires careful consideration of energy consumption, data security, and consumer privacy [5,13].

Finally, regulatory hurdles should be considered. Although both BV and propolis are generally recognized as safe (GRAS) for food contact, the incorporation of nanomaterials and sensor technologies into packaging creates new regulatory challenges [12]. Comprehensive toxicological studies and environmental impact assessments will be required to support wider commercialization.

III. Future Perspectives

Looking ahead, we can say that the future of smart packaging based on BW and propolis is bright, provided that key areas of research and development are addressed. Advances in materials science, particularly in polymer chemistry and nanotechnology, will allow for the creation of stronger, more multifunctional films. And an area that is actively developing is the development of biodegradable nanocomposite films with embedded sensors that provide real-time feedback on food quality. Such systems will not only increase transparency, but also reduce food waste, allowing for more informed purchasing and consumption decisions. Integration with blockchain and cloud analytics can further improve supply chain traceability and efficiency. For example, packaging equipped with IoT sensors can monitor environmental conditions during storage and transportation, alerting stakeholders to deviations that could jeopardize product safety. Collaborative efforts between academia, industry and regulators will be critical to accelerate commercialization. Pilot trials, consumer acceptance studies, and life cycle assessments will provide the necessary data to justify investment in these technologies.

Ultimately, BW and propolis offer a unique combination of functionality, sustainability and bioactivity. Their role in intelligent packaging systems could redefine food safety standards and meet the growing global emphasis on circular economy and environmental stewardship.

IV. Conclusion

The integration of beeswax and propolis into active and smart food packaging systems is a promising approach to improving the quality, safety and shelf life of food products in an environmentally responsible way. Beeswax with its excellent barrier properties and compatibility with sensory substrates and propolis with its powerful antimicrobial and antioxidant activity complement each other in creating multifunctional packaging materials.

Recent advances in nanotechnology have contributed to the development of nanoformulations that allow to control the release and improve the stability of bioactive compounds, as well as enable the integration of advanced sensor technologies. These systems have a dual functionality: they act as physical protectors and active monitors of food freshness and contamination. Despite the technological promise, challenges such as natural variability in materials, processing limitations, and regulatory considerations must be addressed for successful industrial applications. Continued interdisciplinary collaboration and innovation will be crucial to overcome these barriers. Thus, sensor films based on beeswax and propolis are a significant step forward in the evolution of smart packaging systems. Their use is in line with the global movement towards sustainability, providing a biological alternative to synthetic packaging materials and adding real-time food monitoring capabilities that reduce waste, increase transparency, and protect public health.

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Materials for Electromagnetic Radiation Absorption and Zero Electrical Resistance Based on Ceramic-Ferromagnetic Graphite-Polymer Composites

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***Abstract* – The aim of the study was to develop materials with zero specific electrical resistance for electromagnetic radiation absorption, featuring enhanced impact strength and flexural strength. The impact strength of ceramic–ferromagnetic–graphite polymer composites was investigated, with a focus on compositions containing an elevated silicon carbide (SiC) content of 10 wt.%. The content of chromium oxide (Cr_2O_3) and graphite was varied and optimized within the 5–15 wt.% range. The study demonstrated that the addition of Cr_2O_3 and graphite to the polyamide 6 – SiC (10 wt.%) matrix enhances both impact strength and flexural strength. Statistical analysis of the experimental data yielded approximating equations, enabling the prediction of mechanical performance for these composites. Furthermore, the effect of CFG (ceramic–ferromagnetic–graphite) fillers on the spectral characteristics related to electromagnetic radiation absorption suggests that complex composite systems offer significant advantages in the development of efficient absorbing materials. As a result, electromagnetic absorbing composites modified by ceramic-ferromagnetic-graphite (CFG) fillers and with zero electrical resistivity were first received.**

Keywords - polymer, ceramic, ferromagnetic, graphite, composites, electrical conductivity, electromagnetic radiation, absorbing materials, strength properties, spectral characteristics,

I. INTRODUCTION

The increasing use of polymer composites for electromagnetic radiation (EMR) absorption is closely tied to the rapid advancement of modern technologies and the growing problem of electronic pollution [1,2]. The development of highly efficient EMR-absorbing materials with minimal thickness, low density, and a broad transmission and absorption range remains a critical research focus [3].

A key mechanism underlying the effectiveness of modern EMR-absorbing materials lies in their ability to dissipate incoming radiation with minimal reflection. Of particular interest are materials exhibiting zero or near-zero specific electrical resistivity, as this property significantly enhances their functional capabilities. Such materials enable efficient absorption of unwanted ("parasitic") radiation within high-frequency electronic systems and contribute to reducing radar visibility in military applications. Depending on the frequency range - typically determined by the electromagnetic wave reflection coefficient - EMR-absorbing materials are categorized as either broadband or resonant. The absorption of electromagnetic energy occurs primarily through dielectric and magnetic losses, as well as conduction losses. Maximizing these loss mechanisms is key to achieving high shielding efficiency.

Today, a wide variety of shielding materials and composites are available, shaped by two interrelated challenges: the complex physics of EM wave propagation and absorption, and the technological difficulty of synthesizing materials with stable electromagnetic properties across broad frequency ranges [4]. Traditional shielding solutions include metals - applied as foils, meshes, or screens and ferromagnetic materials, which are highly effective but costly and technologically demanding to produce [5]. In contrast, conductive polymers [6] and thermoplastic polymer-based composites [7] offer cost-effective and manufacturable alternatives, albeit with lower shielding performance.

Recent focus has shifted toward advanced composite systems filled with materials such as silicon carbide, carbon fibers [8], carbon black [9], and carbon nanotubes [10], which offer a favorable balance between effectiveness, processability, and affordability. Earlier studies by our group have yielded thermoplastic composites based on polyamide, silicon carbide, and potassium titanates [11], though these materials demonstrated relatively low absorption coefficients. Consequently, further research into composite systems incorporating complex ceramic–ferromagnetic–graphite (CFG) fillers—especially those engineered for zero specific electrical resistivity - represents a promising direction for developing next-generation EMR-absorbing materials with enhanced shielding and mechanical properties.

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The purpose of this article is to explore polymer composites containing ceramic, ferromagnetic, and graphite fillers for the absorption of electromagnetic radiation.

The specific objectives of the study are:

– to examine the influence of CFG (ceramic-ferromagnetic-graphite) fillers on the mechanical properties of polyamide-based polymer composites;

– to assess the electromagnetic absorption capabilities and specific electrical resistivity of polyamide composites reinforced with CFG fillers.

II. MATERIALS AND METHODS

The following materials were used to produce the polymer composites with ceramic-ferromagnetic-graphite fillers:

– Polyamide 6, trademark Durethane (Bayer, Germany);

– Silicon carbide (SiC) powder with a particle size of 50–65 μm, specific volume resistivity, a melting point of 2700 °C, and a linear thermal expansion coefficient (α) ranging from 5 × 10^6 to 7 × 10^6 K^{-1};

– Chromium oxide (Cr_2O_3) powder with particle sizes between 2–5 μm, a melting point of 2435 °C, and a boiling point around 4000 °C;

– Graphite powder (particle size 20–25 μm) obtained from a standard graphite workpiece (see Fig. 1a).



Fig. 1. Photographs of laboratory research: a – graphite blanks; b – roller crusher; c – cylindrical mill mounted on elastic suspensions; d – mechanical disperser.

Graphite workpieces were ground to a particle size of less than 0.2 mm using a roller crusher (Fig. 1, b). The resulting graphite sample was analyzed for moisture content, ash content, as well as hydrogen and carbon levels, following the applicable regulatory standards: ISO 589:2015, DSTU 8995:2020, ISO 1171:2010, and ISO 625:1996.

Table 1 presents the quality indicators of the carbon graphite sample and the chromatographic composition of the analyzed wash oil.

TABLE I. QUALITY INDICATORS OF CARBON GRAPHITE SAMPLE CHROMATOGRAPHIC COMPOSITION OF THE ANALYSED WASH OIL

№	Indicator	Symbol, unit of measurement	Value
1	Total moister	W^r_t, %	0.10
2	Analytical moisture	W^a, %	0.10
3	Ash content	A^d, %	0.18
4	Carbon content	C^d, %	99.06
5	Hydrogen content	H^d, %	0.76

The graphite sample was subsequently milled to achieve 100% of particles smaller than 0.075 mm using a cylindrical mill (Fig. 1, c) combined with a mechanical disperser (Fig. 1, d) and sieved through 0.02 mm meshes (Table 2). Table II presents the results of the graphite sieve analysis.

TABLE II. GRAPHITE SIEVE ANALYSIS RESULTS

Class, mm	Value, %
< 0.02	31.4–43.2
0.02–0.075	56.8–68,6

Polymer composites containing ceramic-ferromagnetic-graphite fillers were produced on a laboratory single-screw extruder at temperatures between 170–200 °C and screw speeds ranging from 30 to 100 rpm.

Impact strength and flexural strength (breaking stress during bending) were determined according to ISO 180 and ISO 178 standards, respectively. Spectral characteristics were measured automatically using a P2–65 scalar spectrum analyzer in the frequency range of 26–37.5 GHz.

The specific electrical resistance of coal coke was determined in accordance with DSTU 8831:2019, the national standard of Ukraine. This method, previously detailed in [12], involves measuring the voltage drop as direct current passes through a compressed column of material (particle size under 0.2 mm) enclosed between two punches within a matrix.

The setup for measuring the specific electrical resistance of coke powder is shown in Fig. 2, with its basic electrical circuit depicted in Fig. 3.

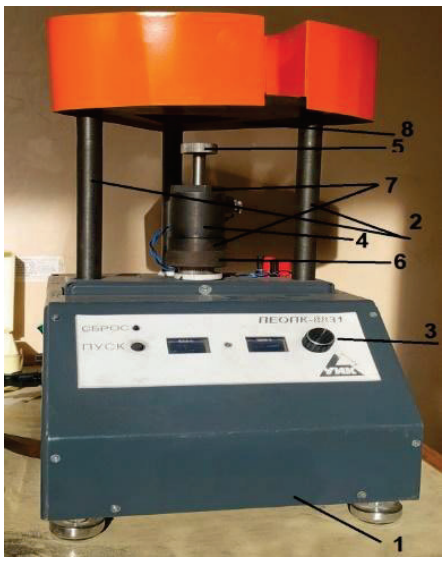


Fig. 2. Schematic of the test setup for measuring the SER of coke powder: 1 – base; 2 – supports; 3 – digital measurement unit with current stabilizer and MikRA-IZ digital millivoltmeter; 4 – hydraulic jack - Uitok 2 kN; 5 – jack handle; 6 – two-probe measurement matrix; 7 – upper and lower electrodes; 8 – crossbar.

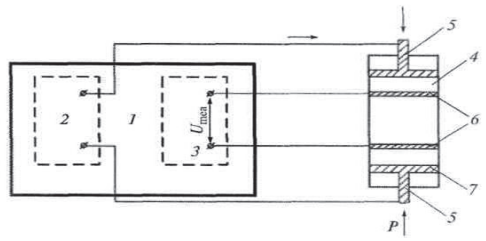


Fig. 3. Schematic of the basic electrical setup:1 – measurement module; 2 – constant current source; 3 – millivolt meter; 4 – measurement cell; 5 – current electrodes; 6 – measurement electrodes; 7 – coke sample.

The coke sample is prepared according to the relevant regulatory and technical standards. It is ground into an analytical powder, meaning that the sample must pass entirely through a standard mesh, with no more than 40% of the particles in the <0.05 mm size range. The powder selected for measurement is dried in an oven at (140 ± 5)°C until it reaches a constant weight. Following the installation guidelines, a block diagram of the device is assembled, and the electrical resistance of the coke powder is measured.

The specific electrical resistance ρ, in Ohm·m, is calculated using the formula:

$$\rho = \frac{U}{I} \cdot \frac{S}{L} \quad (1)$$

Where:

S – cross-sectional area of the sample (m²);

L – distance between probes (m);

U – voltage drop (V);

I – current (A).

The measurement result is obtained by calculating the arithmetic mean of parallel measurements:

$$\overline{X} = \frac{\sum_{i=1}^{n} X_i}{n} \quad (2)$$

Where: Xi – result of the i-th measurement;

n – number of parallel measurements.

The final result is presented as:

$$(\overline{X} \pm \Delta), \; P = 0{,}95 \quad \text{or} \quad \overline{X}(u_c) \quad (3)$$

Where ±Δ represents the absolute total measurement error at a confidence level of P = 0.95, u_c is the total standard uncertainty, equivalent to the absolute total measurement error.

Spectral data were digitized using LabVIEW software (National Instruments) connected to a computer. The sample dimensions ensured full filling of the waveguide cross-section (7.2 × 3.4 mm²) with a sample thickness of 5 mm. For each sample, the transmission coefficient (T) and the standing wave ratio (SWR) were measured.

III. RESULTS OF POLYMER CERAMIC-FERROMAGNETIC-GRAPHITE COMPOSITE DEVELOPMENT

The initial research phase focused on evaluating the influence of silicon carbide (SiC), chromium oxide (Cr_2O_3), and graphite additions on the mechanical strength properties of polyamide 6 composites.

Previous studies [13, 14] indicated that increasing the content of SiC in polyamide 6 enhances both impact strength and flexural strength. The optimal SiC filler content for achieving maximum mechanical properties was determined to be 5 % wt.. However, these compositions did not exhibit satisfactory electromagnetic radiation absorption. Therefore, in this work, compositions with a higher SiC content (10 % wt.) and varying Cr_2O_3 and graphite contents (in the 5–15 % wt. range) were studied. Additionally, to explore potential synergistic effects from the combined use of all three fillers (SiC, Cr_2O_3, and graphite), polyamide 6-based systems incorporating all three CFG components were investigated.

Figure 4 illustrates the impact of filler type and content on the impact strength and flexural strength of polyamide 6 composites containing SiC, Cr_2O_3, and graphite. The data show that the introduction of chromium oxide and graphite into the polyamide 6 – 10 % wt. SiC system leads to a further increase in both impact strength and flexural strength.

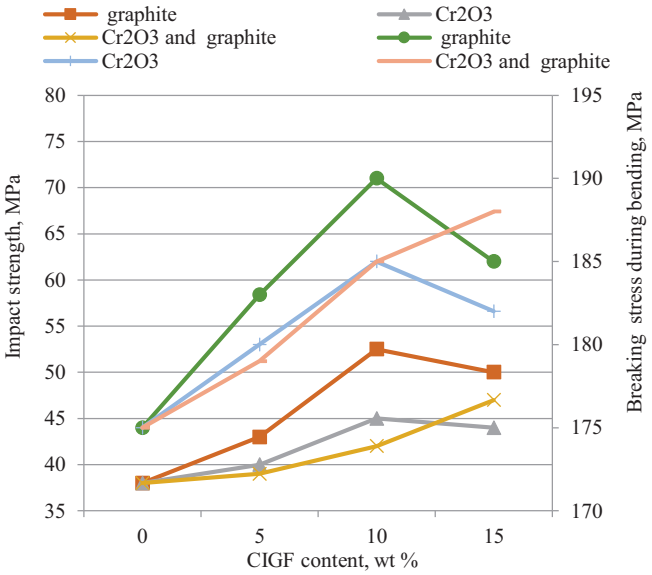


Fig. 4. Influence of filler content on the impact strength and flexural strength of polymer ceramic-inorganic composites based on polyamide 6 with 10 % wt. SiC, and 5–15 % wt. Cr_2O_3 and graphite.

Statistical processing of the experimental data enabled the development of approximation equations, allowing the calculation of impact strength (a, MPa) and flexural strength (σ, MPa) as functions of the graphite and chromium oxide content:

$$a = 37.7279 + 1{,}5348 \cdot x + 0{,}4927 \cdot y - 0.0414 \cdot x^2 - 0.0756 \cdot x \cdot y - 0.0004 \cdot y^2 \quad (4)$$

$$\sigma = 174.8137 + 2.3214 \cdot x + 1.3845 \cdot y - 0.105 \cdot x^2 - 0.0827 \cdot x \cdot y - 0.0566 \cdot y^2 \quad (5)$$

where x – graphite content, wt %;

y – Cr_2O_3 content, wt %.

Equations 4 and 5 allows to simulate impact strength (Fig. 5) and breaking stress during bending (Fig. 6) values while fillers (graphite and Cr_2O_3) content from 0 to 15%.

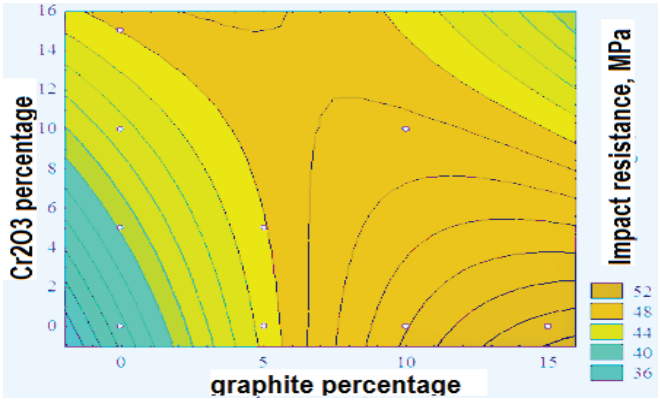


Fig. 5. 2D-model projection of impact strength (MPa) depends from graphite and Cr_2O_3 content, % wt in polymer composition based on polyamid 6 with 10 % wt. SiC

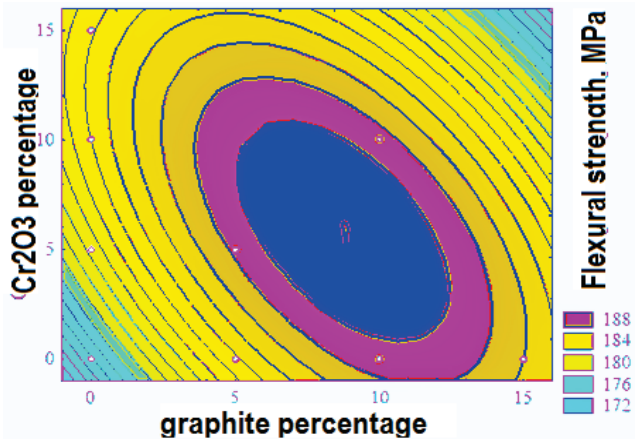


Fig. 6. 2D-model projection of breaking stress during bending (MPa) depends from graphite and Cr_2O_3 content, % wt in polymer composition based on polyamid 6 with 10 % wt. SiC

The analysis of Figs. 5, 6 reveals that the maximum strength properties are observed in polyamide 6 composites containing 10 % wt. SiC and 10 % wt. graphite.

Through statistical data processing, approximation equations were derived, enabling the calculation of impact strength and flexural strength for ceramic-ferromagnetic-graphite polymer composites.

These findings highlight the benefits of utilizing a three-component filler system - 10 % wt. SiC, 10 % wt. Cr_2O_3, and 10 % wt. graphite - to develop materials with enhanced mechanical performance.

IV. RESULTS AND DISCUSSION OF THE SPECIFIC ELECTRICAL RESISTANCE AND THE SPECTRAL CHARACTERISTICS OF CERAMIC-FERROMAGNETIC-GRAPHITE POLYMER COMPOSITES

Electromagnetic radiation absorption properties were investigated for ceramic-ferromagnetic-graphite polymer composites exhibiting optimal mechanical characteristics using a P2–65 automated scalar spectrum analyzer across the 26–37.5 GHz frequency range according to [15-19]. The results showed that the addition of Cr_2O_3 and graphite led to an increase in the reflection coefficient Γ (Fig. 7). The relationship between the reflection coefficient Γ and the standing wave ratio (SWR) was determined using the following formula:

$$\Gamma = \frac{SWR-1}{SWR+1} \quad (6)$$

The highest reflection coefficient was observed in composites based on polyamide 6 with 10 % wt. silicon carbide, 10 % wt. Cr_2O_3, and 10 % wt. graphite, which is attributed to the complex synergistic effects of the CFG fillers.

An analysis of the specific electrical resistance of these modified composites revealed a complete absence of measurable electrical resistance in all cases.

Subsequently, the efficiency of electromagnetic radiation absorption by the developed polymer ceramic-inorganic-graphite composites was assessed. For this purpose, the absorption coefficient A was determined (Fig. 7), calculated using the following formula:

$$A=1-\Gamma-T \quad (7)$$

An increase in the reflection coefficient Γ was noted upon the addition of chromium oxide (Cr_2O_3) and graphite to the polyamide 6 – SiC 10 % wt. system (Fig. 8). The highest reflection coefficient Γ was observed in the polyamide 6 – SiC 10 % wt. – Cr_2O_3 10 % wt. – graphite 10 % wt. composite system.

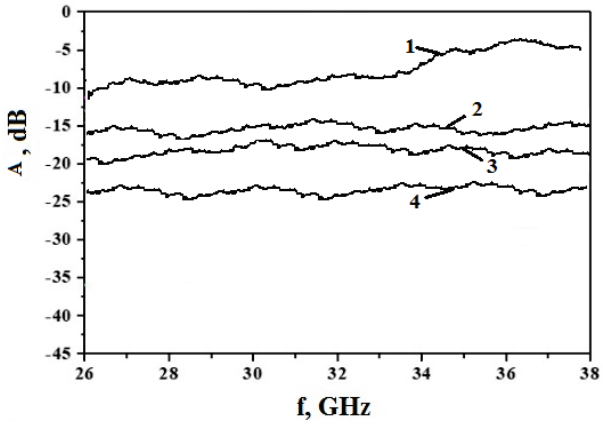


Fig. 7. Spectral dependence of the calculated absorption coefficient A for electromagnetic radiation absorption in polymer ceramic-inorganic composites: 1 – polyamide 6 – SiC 10 % wt.; 2 – polyamide 6 – SiC 10 % wt. – Cr_2O_3 10 % wt.; 3 – polyamide 6 – SiC 10 % wt. – graphite 10 % wt.; 4 – polyamide 6 – SiC 10 % wt. – Cr_2O_3 10 % wt. – graphite 10 % wt..

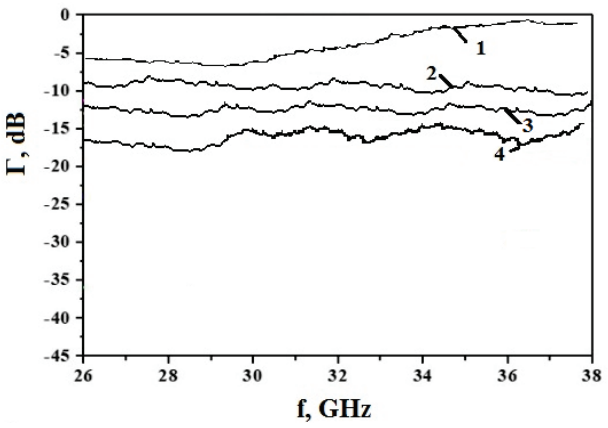


Fig. 8. Spectral dependence of the calculated reflection coefficient Γ for electromagnetic radiation absorption in polymer ceramic-inorganic composites: 1 – polyamide 6 – SiC 10 % wt.; 2 – polyamide 6 – SiC 10 % wt. – Cr_2O_3 10 % wt.; 3 – polyamide 6 – SiC 10 % wt. – graphite 10 % wt.; 4 – polyamide 6 – SiC 10 % wt. – Cr_2O_3 10 % wt. – graphite 10 % wt.

These results highlight the advantages of using a three-component filler system — SiC 10 % wt., Cr_2O_3 10 % wt., and graphite 10 % wt. — to create effective materials for electromagnetic radiation absorption.

V. CONCLUSIONS

For the first time, electromagnetic radiation-absorbing composites modified with ceramic-ferromagnetic-graphite fillers were produced. The impact strength and flexural strength (breaking stress during bending) of ceramic-ferromagnetic-graphite polymer composites were investigated. Compositions with a higher silicon carbide (SiC) content of 10 % wt. were studied, while the content of chromium oxide (Cr_2O_3) and graphite was optimized within the 5-15 % wt. range. It was demonstrated that the incorporation of Cr_2O_3 and graphite into the polyamide 6 – SiC 10 % wt. system enhances both impact strength and bending stress.

Statistical data analysis led to the development of approximation equations, enabling the calculation of impact strength and breaking stress during bending for these composites. A study of the specific electrical resistance of these modified materials showed that no electrical resistance was present in any of the cases.

The modification of ceramic-ferromagnetic-graphite fillers significantly affected the spectral characteristics of the composites, and the advantages of using the complex system of CFG fillers for electromagnetic radiation absorption were clearly demonstrated.

For polyamide 6, the optimal triplex modification system consists of 10 % wt. SiC, 10 % wt. Cr_2O_3, and 10 % wt. graphite, which ensures a comprehensive impact on the composite's properties.

VI. SYMBOLS

Wa – moisture content, %;
Wrt – total moisture content, %;
Ad – ash content, %;
Cd – carbon content, %;
Hd – hydrogen content, %;
a – impact strength, MPa;
σ – breaking stress during bending;
SER – specific electrical resistance, Ohm•cm;
A – absorption coefficient, %;
Γ – reflection coefficient, %;
T – transmission coefficient, %;
SWR – standing wave ratio.

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Method of Formalizing Negative Casual Relationships in Cognitive Models

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***Abstract*—The report presents a method for formalizing negative relationships in cognitive models. The method is based on the stasis of the antonym logic apparatus. New logical operators of antonym logic are introduced to formalize negative relationships between arguments. New properties of antonym logic operators and their correspondence to the properties of known logical operators are presented.**

***Keywords*—cognitive model, formalization of relationships, negative relationships, logic of antonyms, logical operators**

I. Introduction

When creating computer models of weakly structured situations, the methodological apparatus of cognitive modeling is often used [1].

The cognitive model is built on the basis of a cognitive map. A cognitive map is a directed weighted graph. The vertices of the graph are the basic concepts, and the arcs are casual relationships between the concepts.

There can be a positive or negative relationship between concepts. If a change in the state of one concept corresponds to a similar change in the state of another, then there is a positive relationship between them. If the other concept changes its state in the opposite direction, then this is a negative relationship.

The relationships between concepts are formalized in the form of functions or a rule base using the appropriate mathematical apparatus.

But there is a problem of the adequacy of cognitive models. One of its reasons is the difficulty of correctly formalizing a set of causal relationships that differ in sign and strength.

II. Analysis of Cognitive Simulation Methods

Fuzzy cognitive models have gained the greatest development. They use the mathematical apparatus of fuzzy logic [2]. Conditionally, these methods can be divided into classes: fuzzy cognitive maps of Kosko [3; 4]; fuzzy cognitive maps of Silov [5], cognitive maps with fuzzy inference rules [6], fuzzy production cognitive models, fuzzy relational cognitive models and their combined variants [7—10].

There are known developments that use iterative methods of training neural networks in cognitive simulation [11—16]. However, their application is limited due to the uniqueness of the situations being studied.

Typically, fuzzy cognitive models use two types of models. The first type is based on Roberts dynamic models of the form [3; 4]:

$$\tilde{x}_j(t+1) = \tilde{x}_j(t) + \sum_{i=1}^{n} d_{ij} \Delta\tilde{x}_i(t+1). \qquad (1)$$

Here $\tilde{x}_j(t+1)$, $\tilde{x}_j(t)$ — fuzzy values of the variable that specifies the state of the concept X_j at a point in time $(t+1)$ and at the previous point in time (t); $\Delta\tilde{x}_i(t+1)$ — fuzzy value of the increment of this variable at a point in time $(t+1)$; $d_{ij} \in D$ — the strength of the influence of concept X_j on concept X_i.

Additionally, there are two problems for these models. The first problem is the increase in the uncertainty of the simulation results, which accumulates in the process of fuzzy iterative calculations. It causes the second problem — the fuzzy calculation results go beyond the basic range of their possible values. This complicates the interpretation of the simulation results.

These problems are solved by using nonlinear functions to "smooth" or "narrow" the range of simulation results. They allow you to convert the simulation results into values that are limited by the range, for example, $[-1,1]$ or $[0,1]$.

An example of such functions are functions of the sigmoid class of the form:

$$f(x) = \frac{1}{1+e^{-\lambda x}}, \lambda > 0,$$

$$\text{or } f(x) = \frac{x}{|x|+\lambda}.$$

The fuzzy result is obtained based on the Zade fuzzy generalization principle. However, the application of nonlinear functions to the modeling results complicates the process of static and dynamic analysis of cognitive models.

The second type of models is based on the application of operations of fuzzy t – norms and t – conorms [17]. They have the form:

$$\tilde{x}_j(t+1) = \underset{i=1}{\overset{n}{S}} \left[d_{ij} \, T \, \tilde{x}_i(t+1) \right]. \qquad (2)$$

Here S, T — an operator t – norm and t – conorm.

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In practice, various types of these operators are used. The most commonly used are the Zadeh, Lukasiewicz, and probabilistic formulas [2; 15]. In practice, there are also examples of the use of other parametric classes of these operators, for example, the Dombi, Frank, Hamacher, Schweitzer-Sklar and Major-Torrance norms [18—24].

A comparative analysis of the most popular of them was conducted in [25]. The authors proved the advantages of using the mathematical apparatus of the logic of antonyms [26; 27] for carrying out the corresponding operations in cognitive modeling tasks. Certain problems are solved using methods of parameterization of cognitive models using the mathematical apparatus of the logic of antonyms [28—30].

Formalizing negative and oppositely signed relationships between concepts in cognitive models using antonym-logic methods remains an open problem.

III. Developing Antonym Logic Operators for Negative Relationships

The mathematical apparatus of the logic of antonyms allows us to formalize two types of relations: a positive strong relationship and a positive weak relationship. A strong positive relationship (γ – relationship) is taken as the t – norm operator. A weak positive relationship (β – relationship) is taken as the t – conorm operator [26].

Analytical expressions for calculating the results of applying these operators have the form:

$$H[A\ \gamma\ B] = -\log_2\left[1 - 2^{-k_A \cdot H[\alpha A] - k_B \cdot H[\alpha B]}\right], \quad (3)$$

$$H[A\ \beta\ B] = k_A \cdot H[A] + k_B \cdot H[B]. \quad (4)$$

Figure 1 (a) and (b) show the graphs of these functions.

Here α — inversion operator, which is analogous to the logical operator *Not*. The result of applying this operator is a specific value that is the opposite of its argument. $H[A]$ ($H[B]$) and $H[\alpha A]$ ($H[\alpha B]$) — numerical measures of the degree of approximation of element A (B) to the state of absolute quality and the state of absolute poor quality, respectively, k_A , k_B — a numerical measure of their weight.

These estimates are related by functions of the form [27]:

$$H[A] = -\log_2\left(1 - 2^{-H[\alpha A]}\right), \quad (5)$$

$$H[\alpha A] = -\log_2\left(1 - 2^{-H[A]}\right). \quad (6)$$

Figure 1(f) and (e) show graphs of functions for the inverse results of applying antonym logic operators to positive relations. They are calculated by applying (6) to (3) and (4).

It is known that the operators of antonym logic comply with all the laws of classical algebra of logic, in particular, De Morgan laws:

$$\text{if } S_\gamma = A\gamma B, \text{ then } \alpha S_\gamma = (\alpha A)\beta(\alpha B) = \alpha(A\gamma B), \quad (7)$$

$$\text{if } S_\beta = A\beta B, \text{ then } \alpha S_\beta = (\alpha A)\gamma(\alpha B) = \alpha(A\beta B). \quad (8)$$

Let us introduce operators of antonym logic that formalize negative relationships: negative strong relationship ($\bar{\gamma}$ – relationship) and negative weak relationship ($\bar{\beta}$ – relationship).

We will assume that there is a negative strong relationship ($\bar{\gamma}$ – relationship) between the elements of the system if the following conditions are met:

- The quality of the system is higher, the worse the quality of its elements.
- The absolute quality of the system is achieved when at least one of its elements is of poor quality.
- A system is of absolutely poor quality when all its elements are of absolutely quality.

We will assume that there is a negative strong relationship ($\bar{\beta}$ – relationship) between the elements of the system if the following conditions are met:

- The quality of the system is higher, the worse the quality of its elements.
- Absolute quality of a system is achieved when all its elements are of poor quality.
- A system is of absolutely poor quality when at least one of its elements is of absolutely quality.

These conditions for two elements of the system can be written as logical formulas of the form:

$$S_{\bar{\gamma}} = A\ \bar{\gamma} B = (\alpha A)\beta(\alpha B), \quad (9)$$

$$\alpha S_{\bar{\gamma}} = A\ \gamma B = \alpha(A\ \bar{\gamma} B), \quad (10)$$

$$S_{\bar{\beta}} = A\ \bar{\beta}\ B = (\alpha A)\gamma(\alpha B), \quad (11)$$

$$\alpha S_{\bar{\beta}} = A\ \beta\ B = \alpha(A\ \bar{\beta}\ B). \quad (12)$$

These formulas indicate that the operators of the logic of antonyms for negative relationships comply with De Morgan laws. If we compare them with (7) and (8), we obtain the following identities:

$$A\ \bar{\gamma}\ B = \alpha(A\ \gamma B), \quad (13)$$

$$\alpha(A\ \bar{\gamma}\ B) = A\ \gamma\ B; \quad (14)$$

$$A\ \bar{\beta}\ B = \alpha(A\ \beta\ B), \quad (15)$$

$$\alpha(A\ \bar{\beta}\ B) = A\ \beta\ B. \quad (16)$$

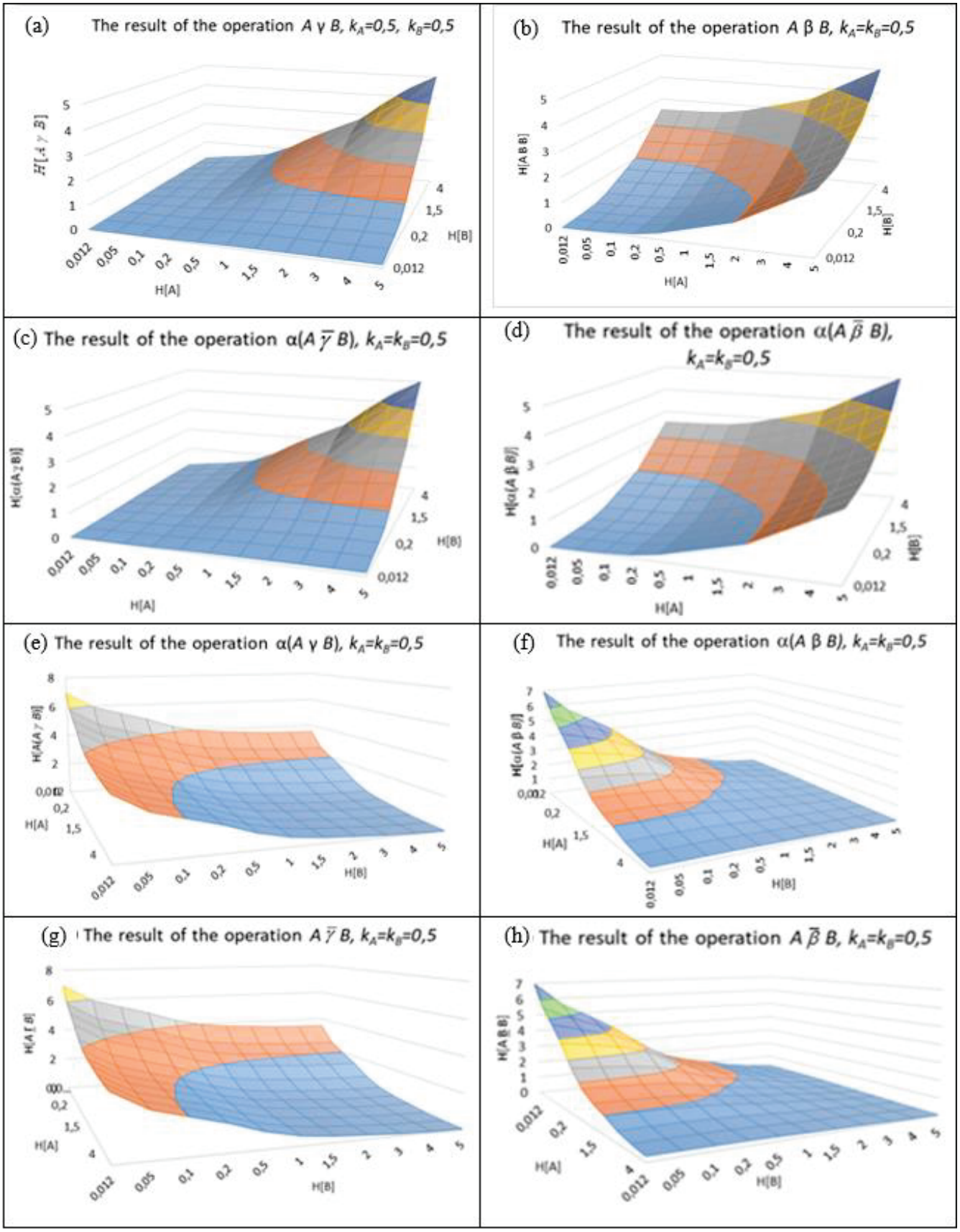


Fig. 1. Graphs of the results of applying antonym logic operators

These identities are new properties of the antonym logic operators that connect the positive and negative relation operators.

Their analysis allowed us to obtain new scientific results:

- The negative strong relationship operator (operator of $\bar{\gamma}$ – relationship) is identical to the logical operator NAND (Scheffer stroke).
- The negative weak connection operator (operator of $\bar{\beta}$ – relationship) is identical to the logical operator NOR (Peirce arrow).

This confirms the reliability of the obtained results.

Considering the Boolean nature of the logic of antonyms and the analogies obtained above, it is possible to supplement the existing properties of the antonym logic operators with the following expressions:

$$H\left[A\gamma B\right]=H\left[\alpha(A\bar{\gamma}B)\right]=H\left[\left(A\bar{\gamma}B\right)\bar{\gamma}\left(A\bar{\gamma}B\right)\right]=$$

$$=H\left[\left(A\bar{\beta}A\right)\bar{\beta}\left(B\bar{\beta}B\right)\right], \quad (17)$$

$$H\left[A\beta B\right]=H\left[\alpha(A\bar{\beta}B)\right]=H\left[\left(A\bar{\gamma}A\right)\bar{\gamma}\left(B\bar{\gamma}B\right)\right]=$$

$$=H\left[\left(A\bar{\beta}B\right)\bar{\beta}\left(A\bar{\beta}B\right)\right], \quad (18)$$

$$H\left[A\bar{\gamma}B\right]=H\left[\alpha\left(A\gamma B\right)\right]=H\left[\left(\alpha A\right)\beta\left(\alpha B\right)\right], \quad (19)$$

$$H\left[A\bar{\beta}B\right]=H\left[\alpha\left(A\beta B\right)\right]=H\left[\left(\alpha A\right)\gamma\left(\alpha B\right)\right], \quad (20)$$

$$H\left[\alpha A\right]=H\left[A\bar{\beta}A\right]=H\left[A\bar{\gamma}A\right]. \quad (21)$$

Analytical expressions for calculating the results of applying antonym logic operators that formalize negative relationships will have the form:

$$H\left[S_{\bar{\gamma}}=A\ \bar{\gamma}\ B\right]=$$

$$=-\log_2\left[\left(1-2^{-H[A]}\right)^{k_A}\left(1-2^{-H[B]}\right)^{k_B}\right], \quad (22)$$

$$H\left[S_{\bar{\beta}}=A\ \bar{\beta}\ B\right]=-\log_2\left[1-2^{-k_A\cdot H[A]-k_B\cdot H[B]}\right]. \quad (23)$$

Figure 1 (g) and (h) show the graphs of these functions. Figure 1 (c) and (d) show graphs of functions for the inverse results of applying antonym logic operators to negative relations. They are calculated by applying (6) to (22) and (23).

The reliability of the obtained results is also confirmed by the results of computational experiments.

Fig. 1 shows graphs of the functions of strong and weak coupling operators for formalizing their positive and negative variants, as well as their inversions.

For the example, arguments A and B are used, which have normalized equivalent weight coefficients k_A and k_B, $k_A+k_B=1$.

The graphs in Fig. 1 display the areas of determination of the results of applying various operators of antonym logic and confirm the obtained new properties (13)—(21).

Indeed, the graphs of the functions in Fig. 1 (a) and (c) are identical. This confirms properties (14) and (17).

The graphs of the functions in Fig. 1 (e) and (g) are identical. This confirms properties (13) and (19).

The graphs of the functions in Fig. 1 (b) and (d) are identical. This confirms properties (16) and (18).

The graphs of the functions in Fig. 1 (f) and (h) are identical. This confirms properties (15) and (20).

It is proven that when using analytical expressions to calculate the results of applying antonym logic operators of the form (3), (4), (6), (17)—(23), properties (17)—(21) are always fulfilled regardless of the values of the coefficients.

IV. Formalization of Negative Relationships in Cognitive Models Using the Methods of Antonym Logic

The state of each concept X_i is proposed to be specified by antonymic evaluations $H_i\left[A\right]$ and $H_i\left[\alpha A\right]$. For this purpose, a general scale is used. It determines how close the state of the element X_i is to its largest A and smallest αA limit state. For example,

A =“*Absolutely high-quality*”,

αA =“*Absolutely poor quality*”,

or A =“*Reliable*”, αA =“*Impossible*”.

We will distinguish the following types of antonyms: gradational and binary.

An example of the use of gradational antonyms in the logic of antonyms is the antonymic pair:

A =“*Big*” and αA =“*Small*”.

In this case, we are talking about opposites that can have intermediate meanings. The estimates and determine how close the object of evaluation is to the state A =“*Big*” or the state αA =“Small”, respectively. This type of antonymic estimates is analogous to membership degree estimates in fuzzy set theory.

The operator α does not imply “Everything else”, but a transition to a specific opposite meaning — a synonym.

An example of binary antonyms is the antonym pair:

A =“*Alive*” and αA =“*Dead”.*

In this case, the opposites have only two meanings. The estimates and determine the degree of possibility that the object of the estimate is A =“*Alive*” or αA =“*Dead*”. This

type of antonymous estimates is analogous to estimates of the probability of events in probability theory.

According to the theory of antonym logic, these evaluations have functional dependencies of the form (5), (6).

For example, Fig. 2 shows the graph of (5).

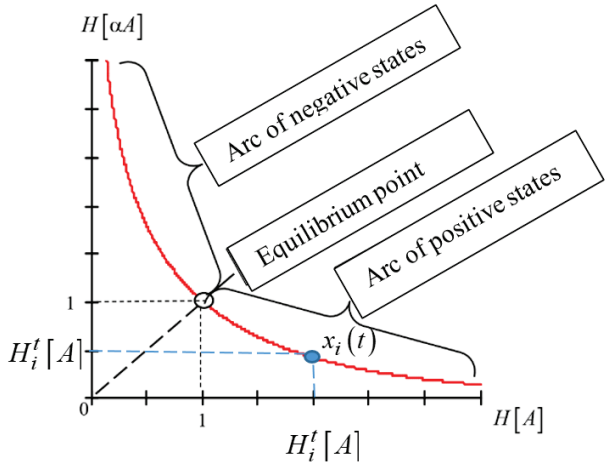


Fig. 2. Graph of the dependency function of antonymic estimates

It defines a curve of possible states $x_i(t) = \left(H_i^t[A], H_i^t[\alpha A]\right)$ of the concept X_i. Here, the pair of estimates $H_i^t[A]$ and $H_i^t[\alpha A]$ are the coordinates of its state at time *t*. These coordinates correspond to each other and are synonymous.

The graph shows that $H_i^t[A] \in (0,\infty)$ and $H_i^t[\alpha A] \in (0,\infty)$. Such a range of possible values of estimates makes it impossible for the modeling results to go beyond its limits.

It is more true when evaluating an object under conditions of uncertainty. Indeed, in conditions of uncertainty, it is incorrect to give an assessment that with full confidence corresponds to absolute quality or absolute poor quality.

The point $x_i(t) = (1,1)$ is the equilibrium point or the point of maximum uncertainty about the state of the concept X_i. It divides the curve into two arcs: the arc of its positive states (here $H_i^t[A] \in (1,\infty)$, $H_i^t[\alpha A] \in (0,1)$) and the arc of negative states (here $H_i^t[A] \in (0,1)$, $H_i^t[\alpha A] \in (\infty,1)$).

Antonym logic operators can be used to formalize positive and negative relationships in models of type (1) and (2).

Fig. 3 shows an element of a cognitive map as an example of logical connections between systems of concepts formalized by operators of antonym logic.

For example, you need to determine the order in which the concept X_1 state changes when the concept X_6 state changes.

The figure shows that the pulse from vertex 6 to vertex 1 can arrive in two ways.

On the first path, the impulse will first change the state of the concept system X_4 and X_2.

On the second path, the impulse will first change the state of the concept system X_5 and X_3.

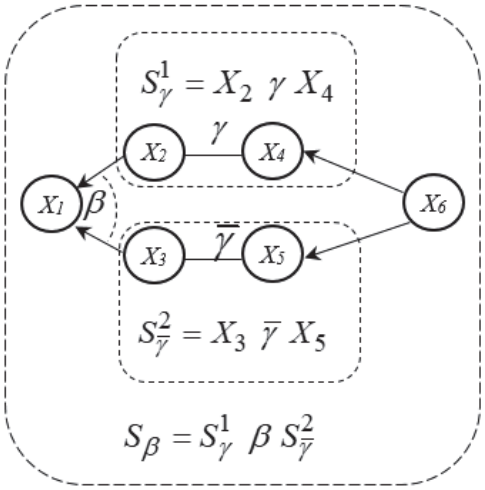


Fig. 3. Cognitive map element with logical connections

The state of the concept X_1 will be determined by the states of these two systems of concepts, which have changed under the influence of the impulse from the concept X_6.

The state of the element is determined by the logic diagram in the figure and is calculated as

$$H\left[S_\gamma^1(t)\right] = H\left[X_2^t\ \gamma\ X_4^t\right] =$$

$$= -\log_2\left(1 - 2^{-\left(k_2 \cdot H_2^t[A] + k_4 \cdot H_4^t[A]\right)}\right),$$

$$H\left[S_{\bar{\gamma}}^2(t)\right] = H\left[X_3^t\ \bar{\gamma}\ X_5^t\right] =$$

$$= -\log_2\left[\left(1 - 2^{-H_3^t[A]}\right)^{k_3}\left(1 - 2^{-H_5^t[B]}\right)^{k_5}\right],$$

$$H\left[X_1^t\right] = H\left[S_\beta = S_\gamma^1(t)\ \beta\ S_{\bar{\gamma}}^2(t)\right] =$$

$$= k_{S_\gamma^1} \cdot H\left[S_\gamma^1(t)\right] + k_{S_{\bar{\gamma}}^2} \cdot H\left[S_{\bar{\gamma}}^2(t)\right].$$

If the cognitive map is represented as a logical network, then the state of its elements can be described using antonym logic operators. This is the essence of the proposed cognitive simulation methods.

V. Conclusions

The methods of the theory of antonym logic for formalizing negative logical operations have been improved. New operators of antonym logic have been introduced: the operator of negative strong connections and the operator of negative weak connections.

New properties are obtained for antonym logic operators that connect positive and negative relationship operators.

Based on these properties, it has been established that the negative strong relationship operator is identical to the logical operator NAND (Scheffer stroke), the negative weak connection operator is identical to the logical operator NOR (Peirce arrow).

A method for formalizing positive and negative relationships in cognitive simulations using antonym logic methods is presented.

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Study of the Dynamics of Resonant Converter with Variable Output Current Polarity for Degaussing Systems

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***Abstract* — The paper investigates the dynamic characteristics of a series-resonant converter with a cycloconverter and *LR*- or *LRC*-load, intended for a ship degaussing system, with small changes in the input value, within which the converter can be approximately considered a linear dynamic link. A simulation model of the converter is created, the equivalent transfer function is determined, the parameters of which are optimized using experimental transient characteristics. The dependences of the parameters of the equivalent transfer function of the converter on the input control value are determined.**

***Keywords* — resonant converter, cycloconverter, dynamic model, frequency response, transfer function**

I. Introduction

Converters of ship degaussing systems regulate currents in degaussing circuits in accordance with the given time functions in the low-frequency range [1 – 3]. However, in some systems, pulse degaussing of the ship's hull is used [4]. An important requirement for these converters is a low level of pulsations from the conversion processes and the network voltage [3, 5], since current pulsations create a noticeable electromagnetic background around the ship, which can be recorded by technical devices. Therefore, the converter, in case of the considering as a transmission link of the autoregulation system, must have an appropriate current regulation speed that will compensate for low-frequency pulsations from the power supply side, and also an appropriate filtering ability to reduce high-frequency pulsations with the operating conversion frequency. In such cases, it is sensable to use converters that operate with sinusoidal pulses to form the desired shape of the output current or voltage [6, 7], which in turn will reduce the THD of the system in comparison with classic pulse converters. However, such converters may require the introduction of additional elements of the control system for correcting their operation with loads of different nature.

In such a case, emphasis should be placed on a combined control method that ensures a sufficiently fast response of the system when the nature of the load changes, for instance, the method of width-frequency regulation. To clarify the specified properties of the converter and to design current autoregulation systems "in small" it is necessary to study the dynamic characteristics of the converter and create a linear dynamic model for small changes in quantities.

Therefore, the purpose of the work is to study the dynamic characteristics and create a dynamic model of a series-resonant converter with variable polarity of the output current for small changes in quantities with *LR*- or *LCR*-loading, typical for the circuits of ship degaussing systems.

II. Idealized Circuit, Converter Simulation Model and Experimental Conditions

The resonant converter with a cycloconverter (**RC-C**), the idealized power circuit of which is shown in Fig. 1a, is used in backup power supply systems for network consumers of electricity with alternating voltage from solar panels and other autonomous power sources [8, 9]. The useful properties of RC-C, namely the generation of low-frequency alternating output voltages and currents, resonant switching principles, expected low switching losses and low generated electromagnetic interference [10, 11], make it advantageous to use RC-C for powering ship degaussing circuits [5]. In the circuit, we can distinguish a bridge inverter with controlled switches S_1-S_4, a cycloconverter with controlled switches S_5, S_6, a power source with a constant voltage U_s and a load Z_{dc}. The arrows and signs in Fig. 1,a show the positive polarity of currents and voltages on the circuit elements. The pairs of switches S_1, S_4 and S_2, S_3 of the bridge inverter are alternately closed and opened at equal intervals of time with an operating frequency f_g.

At the output of the inverter, a rectangular alternating voltage u_{g1} appears, which is transformed into a voltage u_{g2} of the required value by means of a voltage transformer *TV*. The switches of the cycloconverter S_5 and S_6 are also closed and opened with a frequency equal to the operating frequency f_g, or close to the operating frequency. Regarding the switches, it is assumed that they are switched instantly, the switching of adjacent switches is coordinated and occurs without “zero” pauses, the resistance of the closed switch tends to zero, and the resistance of the open switch tends to infinity. In a circuit with resonant elements R_r, L_r, C_r, oscillations of the resonant current ir occur, which is converted into an output current i_q with a constant component of positive or negative polarity. In the load Z_{dc}, along with the constant component, low-frequency components of the output current i_q with frequencies up to tens of Hertz can also be controlled. The switching sequence of the cycloconverter switches S_5 and S_6 can be synchronized with the switching sequence of the inverter switches (independent switching) or with the processes in the resonant circuit (conditioned switching).

In the first case, with independent switching, the phase shift between the switching of the inverter and cycloconverter switches affects the magnitude and sign of

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the current average value of the load current [15, 16]. However, as experiments with simulation models have shown, in the process of smooth adjustment of the phase shift in a wide range (≥180°), cardinal changes in the switching conditions of the switches occur. Closing and opening of a single switch can occur at any of the same or different signs of the switch voltage and current (meaning before and after the switching moment), and so-called "zero" conditions are also possible – switching at zero current, voltage, voltage and current [12 – 14]. Therefore, in the general case, with unconditional switching, the inverter and cycloconverter switches must be bidirectional and independently controlled, which can practically be implemented as series-connected pairs of transistors with reverse diodes (Fig. 1,c) [16, 17].

In the second case, with conditional switching of the switches, it is possible to limit changes in the switching conditions of the switches of the power circuit and, as a result, simplify the power circuit. This paper considers the diode switching principle, in which the switches of the cycloconverter S_5 and S_6 are closed at a positive voltage and open at a reverse current. When operating at a frequency above the frequency of series resonance: $f_g > f_{r0}$, where $f_{r0}^2 = 1/(4 \cdot \pi^2 \cdot L_r \cdot C_r)$, it is possible to achieve acceptable switching conditions for single transistors with reverse diodes as the switches of the inverter S_1-S_4 [18]. Namely, closing the switch at a reverse voltage (the reverse diode is triggered) and opening the switch at a forward current. Each switch of the cycloconverter S_5 and S_6 in this case is a pair of transistors S_a and S_b with reverse diodes VD_a and VD_b (Fig. 1,c). Transistor S_a in S_5 or S_6 turns on at negative (Fig. 1,d), and transistor S_b at positive polarity (Fig. 1,b) of the current average output current at *RL*-load, or the current average output voltage u_q at *RLC*-load.

Taking into account the filtering capacitance at the output, the load Z_{dc} at terminals *a* and *b* of RC-C is presented as an *LRC*-circuit in Fig. 1,e with different ratios of the inductive or capacitive component, which significantly affects the nature of electromagnetic processes. If the inductive component prevails, when the high-frequency component of the current i_L is greater in the inductance L_{dc} than in the capacitance C_f, the load behaves similarly to a current source. In this case, it is advisable to take C_f=0 and consider the load as an *RL*-circuit (here we assume that the intrinsic capacitance of the degaussing circuits is relatively small). If the capacitive component prevails, when the high-frequency component of the current i_L is greater in the capacitance C_f than in the inductance L_{dc}, the load behaves similarly to a voltage source. Then the load is an *RLC*-circuit with a sufficiently large capacitance:

$C_f > L_{dc}/(Q_{dc_max} \cdot R_{dc2})$ or $C_f > 1/(4 \cdot \pi^2 \cdot L_{dc} \cdot f_{dc\max}^2)$ (see numerical data below).

To obtain the experimental characteristics of RC-C, a Simulink model was built in Fig. 2, which contains a controlled voltage source "Ug" and a proportional element "Ugm" that simulate an inverter, a pulse width and frequency modulator "Gamma-F_Gen" for controlling the inverter, groups of switches S_1-D_1, S_2-D_2 and S_3-D_3, S_4-D_4 that simulate switches S_5 and S_6 of the cycloconverter in Fig. 1,a, resonant elements "Series RLC-Branch", a voltage transformer "Linear Transformer" and a load circuit "CF-RL". The filtering capacitance "CF" can be turned on and off from the circuit using the switch "Cap Switch" and "Breaker". The load "RL" is represented in the form of a rectangular coil with the following dimensions: height and width of the rectangular section a_c=0.1 m and r_c=0.2 m; width and length of the rectangular contour along the center line of the coil b_c=10 m and c_c=50 m. The electrical load parameters were calculated using the method [19] for an inductor with the above geometric parameters: number of turns w_{dc}=20; cross-sectional area of the copper wire s_w=6.67 mm^2; wire diameter d_w=3 mm; active resistance of the coil R_{dc}=6.3 Ohm; inductance of the coil L_{dc}=46 mH; nominal coil current I_{dc}=20 A (magnetizing force iw=400 ampere-turns [5], current density 3 A/mm^2) corresponds to an active power of 2520 W.

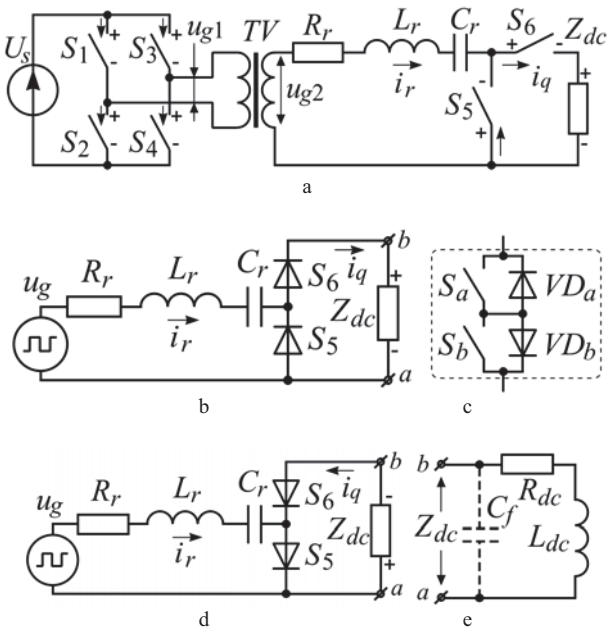


Fig. 1. Simplified circuit of a series-resonant power converter with a cycloconverter (a) and its equivalent circuits for positive (b) and negative (d) polarity of the output current, circuit of switches S_5 and S_6 with diodes (c), equivalent circuit of the load Z_{dc} (e)

The simulations were carried out in the MATLAB/Simulink environment. Simulation parameters [20]:

Start Time = 0 s; Stop Time = 0.5 s; Variable step; ode23t(mod. stiff/Trapezoidal); Max Step Size = 2e-6 s; Relative Tolerance = 1e-6. The power circuit parameters were calculated with the initial data: minimum supply voltage 400 V; maximum supply voltage 750 V; output power 3000 W; output voltage 130 V. Resonant inductance, capacitance and resistance of active losses: L_r=2.55 mH; C_r=10 μF; R_r=0.64 Ohm (Q factor of the serial *RLC*-branch Q_r=24). Resonant frequency f_{r0}=1000 Hz. The voltage transformer has a transformation ratio k_{tr}=0.5 and a rated voltage, active resistance and leakage inductance of the primary side respectively: 750 V; 0.02 Ohm; 33 μH; secondary side: 375 V; 0.02 Ohm; 8 μH; equivalent active resistance of core losses 5000 Ohm; magnetizing inductance 270 mH. Minimum *RLC*-load filtering capacitance $C_f \geq 1000$ μF for maximum Q-factor $Q_{dc\max}$=2.7 and natural resonant frequency $f_{dc\max}$=24 Hz.

In the RC-C model in Fig. 2, the frequency-width regulation is implemented, which allows to regulate the output values from 0% to 100% with a limited change in the

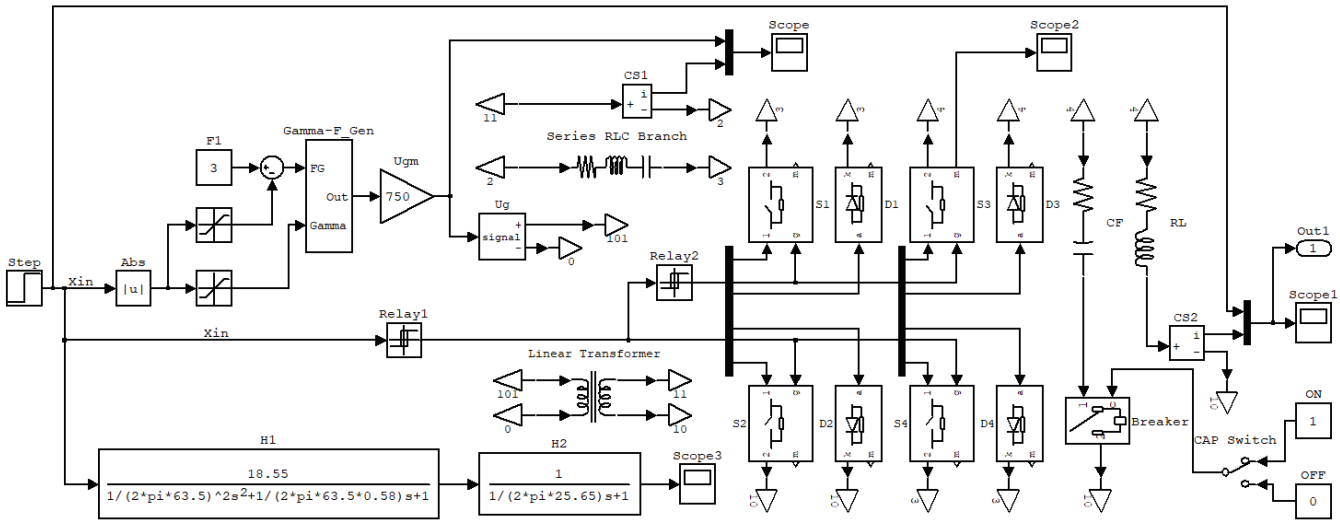


Fig. 2. Simulation model circuit of the resonant converter with cycloconverter

operating frequency, since frequency regulation alone does not allow you to cover the full range [21]. When the input value x_{in} increases from 0 to 1, the relative pulse width from γ=0 to γ=1 increases proportionally. The operating frequency f_g is 2 kHz. When x_{in} increases from 1 to 2 γ=1, the operating frequency f_g decreases from 2 kHz to 1 kHz (to the series resonance frequency f_{r0}) and the output values grow nonlinearly [22]. With a further increase in x_{in}>2, we have constants γ=1 and f_g =1 kHz. This is described by the expressions: f_g = 2000 Hz for x_{in}< 1; f_g=1000·(3–x_{in}) Hz for $x_{in}\geqslant 1$ and $x_{in}\leqslant 2$; f_g=1000 Hz for x_{in}>2; γ= x_{in} for x_{in}<1; for $x_{in}\geqslant 1$, where γ=2·t_p·f_g; t_p is the inverter pulse length.The principle of operation of the model is as follows. The input signal x_{in} of a step form from the "Step" block is converted in the signal shaper into the width and frequency of the inverter voltage pulses u_g. Under the action of the inverter voltage in the resonant branch "Series *RLC*-Branch" an alternating resonant current arises, which also passes through the cycloconverter switches. When the input signal has a positive polarity x_{in} >0, the switches S_2 and S_4 are closed. Diodes D_1 and D_3 work as a single-half-wave rectifier and create a positive load current i_q (see Fig. 1,b). When x_{in}<0, the switches S_1 and S_3 are closed, the diodes D_2 and D_4 create a negative load current i_q (see Fig. 1,d). The resonant current is measured by the virtual sensor CS1, the load current by the sensor CS2.

To compare the transient characteristics of the RC-C simulation model and the equivalent linear dynamic model, blocks of continuous second- and first-order transfer functions "Transfer Fcn" H1 and H2 with defined parameters for a specific operating mode are used.

III. Theoretical Reasoning

Consider the equivalent circuit of the RC-C power circuit in Fig. 3 for two parts of the duty cycle T_g=1/(2·f_g). In the first part of the duty cycle, the switch *Sw* is in position 1 and the resonant circuit and the load circuit are separated. The energy from the inverter is accumulated in the reactive elements L_r and C_r. In the load circuit, there is a redistribution and partial conversion of electricity from the reactive elements L_{dc} and C_{dc} into heat in the active resistance R_{dc}. In the second part of the duty cycle, the switch *Sw* is in position 2 and connects the resonant circuit and the load circuit. There is a continuous transfer of energy from the input to the output of the converter. The energy from the inverter is redistributed in the reactive elements L_r, C_r, L_{dc}, C_{dc} and converted into heat in the active resistance R_{dc}.

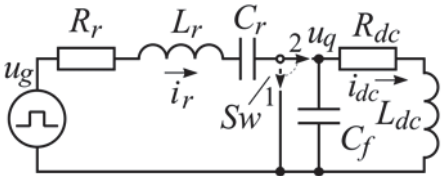


Fig. 3. Equivalent circuit diagram of the RC-C power circuit for the first (switch *Sw* in position 1) and for the second part of the operating cycle (switch *Sw* in position 2)

The linear mathematical model of RC-C corresponds to the scheme in Fig. 3 and combines the electrical resistances of the circuit branches in the Laplace domain, and are given by $s = \sigma + j\cdot\omega$:

$$Z_1 = R_r + s\cdot L_r + 1/(s\cdot C_r);\ Z_2 = 1/(s\cdot C_f);$$

$$Z_3 = R_{dc} + s\cdot L_{dc};\ Z_{dc} = Z_2\cdot Z_3/(Z_2+Z_3). \qquad (1)$$

RC-C transfer function with respect to output current for the second part of the duty cycle

$$G_{ui}(s) = i_{dc}(s)/u_g(s),$$
$$G_{ui}(s) = Z_2/\left(Z_1\cdot(Z_2+Z_3)+Z_2\cdot Z_3\right). \qquad (2)$$

The characteristic equation of the dynamic model has the form

$$H(s)+1 = Z_2/\left(Z_1\cdot(Z_2+Z_3)+Z_2\cdot Z_3\right) = 0. \qquad (3)$$

After simple algebraic transformations, the transfer function of the main part of the RC-C dynamic model is obtained from (3)

$$H(s) = \left(Z_1 \cdot \left(Z_2^{-1} + Z_3^{-1}\right)\right)^{-1} = Z_{dc}/Z_1 , \quad (4)$$

where $1/Z_1 = s \cdot C_r/(1+s \cdot R_r \cdot C_r+s^2 \cdot L_r \cdot C_r)$ is conductivity of the resonant branch and Z_{dc} is the resistance of the load circuit with respect to terminals *a* and *b* in Fig. 1,b,d. With *RL*-load, the load resistance is of the first order: $Z_{dc} = R_{dc} \cdot (1+s \cdot \tau_{dc1})$. With *RLC*-load – second order:

$Z_{dc} = R_{dc} \cdot (1+s \cdot \tau_{dc1})/(1+s \cdot \tau_{dc2}+s^2/\omega_{dc}^2)$. It is accepted here: $\tau_{dc1} = L_{dc}/R_{dc}$; $\omega_{dc}^2 = 1/(L_{dc} \cdot C_f)$; $\tau_{dc2} = R_{dc} \cdot C_f$.

The hypothesis of the study is the assumption that the order of the equivalent RC-C transfer function will be no more than the minimum acceptable order of the equivalent circuit for one of the intervals of the operating cycle with continuous input-output energy transfer [23 – 25]. The minimum acceptable order of the circuit will be considered the order that corresponds to a reasonably simplified dynamic model of the circuit while maintaining the similarity of its behaviour to the behaviour of a real converter in certain dynamic operating modes. In this case, with a large filtering capacitance, the load circuit is aperiodic and $L_{dc} \to 0$. The *RC*-load resistance will have the first order: $Z_{dc} = R_{dc}/(1+s \cdot \tau_{dc2})$. The minimum order of the load circuit is 1 and the order of the resonant branch is 2 and, accordingly, the equivalent continuous dynamic model is a 3rd order transfer function of the low-pass filter type

$$G_e(s) = \frac{1}{1+s/(2 \cdot \pi \cdot f_{1e})} \cdot \frac{K_e}{\left(\frac{s}{2 \cdot \pi \cdot f_{2e}}\right)^2 + \frac{s}{2 \cdot \pi \cdot f_{2e} \cdot Q_e} + 1} . \quad (5)$$

Here the parameters are indicated as follows: $K_e = \Delta i_{dc}/\Delta x_{in}$ – transmission coefficient; f_{1e} – frequency of the first-order pole; f_{2e} – frequency of the second-order pole; Q_e – quality factor of the second-order pole.

Also, the order and type of the equivalent transfer function can be determined from the experimental transients in Fig. 4 and the logarithmic amplitude-frequency characteristics in Fig. 5. The transient characteristics were taken when the RC-C model was supplied with two-step input signals in Fig. 4. The first basic step x_{in0} brings the converter to a certain stable operating mode. After establishing transient processes at a time instant, the second step is supplied at a time instant t_0 with a small increment value $\Delta x_{in} = 0.05 \ldots 0.1$, at which the converter transfer coefficient can be considered unchanged. The frequency characteristics were calculated on the basis of discrete pulse transient functions as the first difference of samples of transient characteristics with a length of $N = 20 \ldots 100$ with a constant period $T = 0.025$ s. The last term in the formula is mutually subtracted from the last component of the series with the index *N*, since it is not included in the first difference

$$L(z) = \left(1 - z^{-1}\right) \cdot \sum_{k=0}^{N} y_k^* \cdot z^{-k} + y_N^* \cdot z^{-N-1} ,$$

$$K_{ui}(\omega) = \left|L(z)\right| / \Delta x_{in} , \quad (6)$$

where $z = e^{s \cdot T}$ is the discrete Laplace transform operator, here $s = j \cdot \omega$; y_k^*, y_N^* – transient samples of the converter with the base level subtracted from the first step $y_k^* = y_k - y_{k0}$; Δx_{in} – the value of the second step of the converter control input signal.

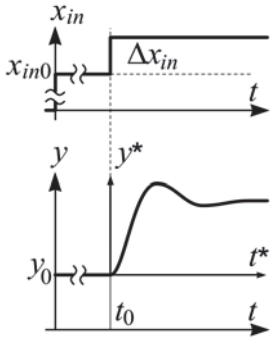


Fig. 4. The type of transient response of the converter

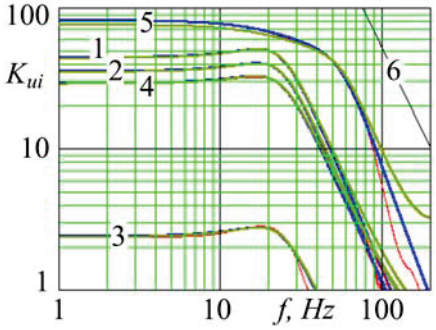


Fig. 5. Logarithmic amplitude-frequency characteristics of the converter for small values at basic input signals: x_{in0}=0,001 (curve 1); x_{in0}=0,2 (curve 2); x_{in0}=0,9 (curve 3); x_{in0}=1,2 (curve 4); x_{in0}=1,9 (curve 5). Line 6 demonstrates a 3rd order slope of -60 dB/dec

The transfer coefficient of the equivalent transfer function is defined as

$$K_e = \left(\sum_{k=N-m}^{N} y_k^*\right) / \left(m \cdot x_{in0}\right) , \quad (7)$$

where *m* is the number of samples of the steady process. Other parameters of the equivalent transfer function (5) are determined by selecting from a comparison of the experimental transient characteristics of the model in Fig. 2 and the calculated originals $h(t) = \ell^{-1}\left[G_e(s, K_e, f_{1e}, f_{2e}, Q_e)/s\right]$, which are equivalent transition functions $h(t)$, by the gradient descent method. The criterion that needs to be reduced during the sequential cyclic step-by-step selection of the three parameters f_{1e}, f_{2e}, Q_e is the total deviation

$$S_e = \sum_{k=0}^{N} (y_k^* - h_k)^2 , \quad (8)$$

where y_k^* is sampling of experimental transient characteristics; h_k - samples of the transient characteristics calculated using (5). Cyclic parameter selection stops when the sum (8) stops decreasing.

IV. Results of Experiments and Calculations

During experiments with the model in Fig. 2, for different basic values of the input quantity x_{in0} (output current control signal), the parameters of the equivalent transfer function (5) were determined by the method of selection with comparison of transient characteristics. The generalized dependences of the parameters on the input quantity, shown in Fig. 6 by thick lines 1-8, were calculated by the exponential smoothing method on a sufficient amount of data in the experimental samples.

$$P(x_{in}) = \frac{\sum_{k=0}^{N} p_k \cdot e^{-\mu\cdot(x_{in}-x_{in0_k})^2}}{\sum_{k=0}^{N} e^{-\mu\cdot(x_{in}-x_{in0_k})^2}}, \qquad (9)$$

where $P(x_{in})$ is a function of one of the parameters K_e, f_{1e}, f_{2e}, Q_e; N is the sample size; p_k is the value of the parameter from the sample; μ is the smoothing parameter (here μ =2); xin is the variable input value of the converter; x_{in0_k} is the value of the basic input value for the k-th parameter from the sample. The thin lines in Fig. 6 show individual samples of the specified parameters as an example. Each sample contains 20 values of each of the selected parameters over the entire range of the input value x_{in} = 0…2. To calculate the generalized dependence according to (9), a certain number of samples can be used, the data from which are grouped into a single array.

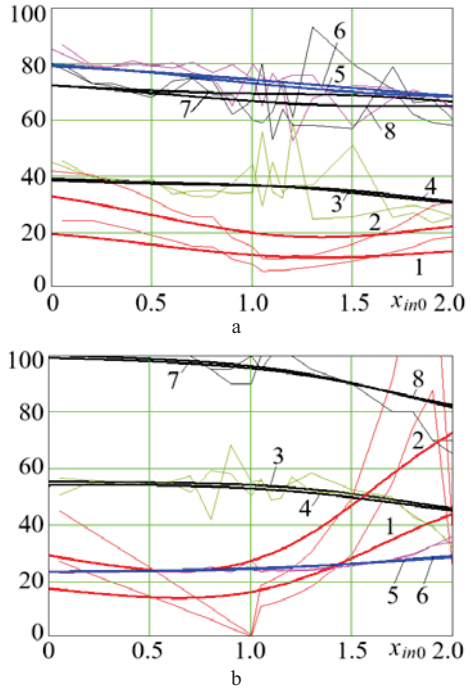


Fig. 6. Dependences of the parameters of the equivalent transfer function of the converter on the input value xin: a – for *RL*-load; b – for *RLC*-load (curves: 1, 2 – K_e; curves: 3, 4 – f_{1e}; curves: 5, 6 – f_{2e}; curves: 7, 8 – 100·Q_e; curves: 1, 3, 5, 7 – for a supply voltage of 450 V, curves: 2, 4, 6, 8 – for a supply voltage of 750 V)

The graphs in Fig. 6 demonstrate a significant non-monotonic dependence of the converter transfer coefficient K_e on the input control value: in the transition zone from pulse-width to frequency control (x_{in}≈1) its value is the smallest and when approaching the resonant frequency (x_{in}≈2) it becomes to maximum. Moreover, the dependence $K_e(x_{in})$ is more significant with *RLC*-load (Fig. 6,b), since the resonant circuit of the converter operates with a higher actual Q factor due to the shunting of the load circuit with a large filtering capacitance. The dependence of K_e on the supply voltage is practically proportional and is explained by the multiplicative effect of the supply voltage on the output values of the converter. Other parameters f_{1e}, f_{2e}, Q_e, as expected, demonstrate a slight dependence on the converter supply voltage and moderate monotonic dependences on the input signal. Therefore, in the future, it is necessary to investigate the adequacy of the proposed dynamic model of the converter with constant parameters f_{1e}, f_{2e}, Q_e and with a variable gain parameter K_e in accordance with the static characteristics of the converter.

Conclusion

So, the equivalent dynamic model of the resonant power converter with a cycloconverter was created, which is based on typical links and corresponds to the structure of the power circuit. During the optimization of the parameters, the equivalent model was approximated by the convergence of processes to the simulation model of the converter, which is conventionally considered an analogue of a real converter. This approach can be used with real converters in laboratory conditions. The difference of this approach from known methods of system identification lies in the experimental confirmation of the hypothesis that the structure of the equivalent model corresponds to the structure of the power circuit at one of the intervals of structure constancy and continuous connection between the input and output within the converter operating cycle. At that time, known identification methods are based only on generalized transfer functions or differential equations of certain orders and therefore do not provide a physical correspondence of the model structures to the structures of real systems.

The conducted research provides grounds for further research directions: assessment of deviations of the processes of the equivalent model with generalized parameters from the processes of a real converter; creation and optimization of equivalent dynamic models of converters with metered energy transfer without continuous connection between the input and output; determination of optimal, from the point of view of linearity and energy efficiency, laws of regulation of resonant converters with cycloconverters; study of the dependence of switching conditions on the regulating value during unconditional switching of power switches.

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Ensuring Authenticity and Integrity Messages on Basis Complex Modified Algorithm UMAC on Crypto-Code Constructions using Rao-Nama Symmetrical Cryptosystems

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Abstract*— This article discusses the structural diagrams involved in generating an authenticated message using the complex modified UMAC algorithm, which is based on modified Rao-Nama crypto-code constructions (MCCC) and hybrid Rao-Nama crypto-code constructions (HCCC). Proposed improvements to the SSL/TLS transport layer protocol utilize complex algorithms derived from Rao-Nama crypto-code constructions on modified elliptic codes (lossy codes) alongside an enhanced cascade UMAC hashing algorithm.

Keywords—UMAC, CCC, HCCC, NIST, McAlice asymmetric cryptosystem, Rao-Nam symmetric cryptosystem, SSL/TLS, DC

I. INTRODUCTION

As the transmission and storage of confidential information grow, so does the necessity for its protection. This urgency calls for a reevaluation of encryption methods, especially in the context of post-quantum computing, to prevent unauthorized access. Many contemporary communication devices face resource constraints, necessitating enhanced security without escalating computational and hardware complexity.

Utilizing crypto-code constructions to bolster the crypto-security of notifications transmitted over information and communication channels is recommended based on findings from the NIST competition [1]. The McAlice schemes, which are based on NP-complete problems [2, 3], are suggested for use. While this approach accelerates coding speed, it also increases the volume of key data, presenting a significant drawback. Research indicates that public-key crypto-code constructions may be vulnerable to attacks, particularly asymmetric McAlice cryptosystems based on Reed-Solomon codes [4, 5]. Similar vulnerabilities exist in Niederreiter crypto-code constructions due to the orthogonality of generating and checking matrices. However, McAlice constructions based on quasi-cyclic parity-check codes with low density can overcome these limitations [6]. This adaptation counters the Sidelnikov attack, although it also results in increased key data volume. Despite modifications to McAleese and Niederreiter structures [7] and [8] that enhance resistance to the Sidelnikov attack, the substantial key data storage remains a barrier to their application in smart technologies and critical infrastructure networks [9]. Consequently, it is evident that noise-resistant linear codes exhibit low crypto-resistance, making the crypto-code structures built upon them vulnerable to breaches. Even hybrid McAleese and Niederreiter structures, which facilitate multi-channel cryptography, suffer from excessive key data and slow crypto-transformation speeds. While McAlice constructions on low-density quasi-cyclic parity-check codes provide simultaneous security and error correction within specific network coding systems [22], they also inflate key data volume. A similar issue arises with the use of Smith convolutional codes in the McAlice cryptosystem [23]. The Rao-Nama cryptosystem [24] serves as a symmetric variant of the McAlice code cryptosystem [25], designed to address the weaknesses inherent in early symmetric encryption schemes [26, 27]. Its primary objective is to ensure both security and reliability in data transmission through a symmetric key cryptosystem. The Rao-Nama system defines this cryptosystem using binary Hoppy codes. Following its introduction, various attacks based on selected open texts emerged [28, 29, 30], prompting numerous enhancements and modifications to the original Rao-Nam scheme. To address these vulnerabilities, the Rao-Nam scheme was redefined as a symmetric key cryptosystem [18]. Struyka R. and Tilburg I. established that the Rao-Nam scheme can be based on any finite field GF(q) [23]. Barbero A. I. and Itrekhus O. successfully reduced the Rao-Nam scheme's secret key size while maintaining system security [24]. However, the Rao-Nama schemes under review still face challenges due to their large key sizes, resulting in low information transmission speeds, rendering them impractical. Consequently, several schemes have been developed to transform the Rao-Nam scheme [25]. These modifications aim to minimize key size to enhance transmission speed and bolster security against common attacks. Some approaches utilize quasi-cyclic (QC) codes to achieve this reduction [25], while others employ low-density parity-check (LDPC) codes to improve transmission speed [26]. These enhancements to the Rao-Nam schemes facilitate the scrambling, rearranging, or altering of code word bits [27]. In [28], [29], and [30], QC-LDPC code-based schemes are discussed, which randomly identify the codeword corresponding to the plaintext. The security and key sizes of these schemes have been significantly enhanced. A Rao-Nam scheme utilizing low-density lattice codes (LDLC) was also proposed [31], featuring a practical decoder [32]. Nevertheless, the examined Rao-Nam schemes remain susceptible to hash function attacks [33]. To mitigate this risk, nonlinear layers were integrated into Rao-Nam schemes to bolster resilience against known plaintext attacks. For instance, SECC schemes (secret error-correcting codes) employ non-linear codes or blockchain techniques to defend against selected plaintext attacks [34]. ECBC (Error Correction Based Cipher) schemes utilize the nonlinear function f [35]. However, differential or plaintext attacks against ECBC and SECC [36] have revealed vulnerabilities in the nonlinear function f . To counter this, a specialized design for the nonlinear function f was proposed [37], enhancing the security of channel coding. An analysis of existing crypto-code structures indicates that McElis and Niederreiter crypto-transformations facilitate the construction of asymmetric cryptosystems and elevate the probability of information retrieval. To maintain the necessary stability in the post-quantum era, the implementation of noise-resistant codes necessitates the use of algebro-geometric codes. A significant drawback is the creation and storage of large volumes of key data, limiting the application of noise-resistant codes in modern technologies such as smart and mobile technologies. These issues can be addressed through the development of Rao-Nama crypto-code structures. Thus, the need arises to circumvent

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linearity in constructing such structures to ensure a certain level of crypto-resistance against quantum-era attacks.

The objective of this research is to devise methods for ensuring the authenticity and integrity of message transmission over information and communication channels utilizing the SSL/TLS protocol, employing the complex modified UMAC algorithm on Rao-Nama crypto-code constructions based on elliptic and modified elliptic codes. Such crypto-transformations are essential for practical application in contemporary computer technologies and for enhancing the cybersecurity of critical infrastructure facilities.

II. Development of complex CCC/UMAC and HCCC/UMAC Rao-Nama algorithms on elliptic codes and their modifications

In developing a complex algorithm within the SSL/TLS protocol to implement security services, it is proposed to combine post-quantum cryptography algorithms with the enhanced UMAC algorithm [7]. This strategy aims to achieve an acceptable level of stability, efficiency, and reliability in the post-quantum crypto period. Consequently, the principle of forming an authenticated message is altered, which is a distinctive feature of this algorithm. The initial stage involves receiving data through the cascade UMAC hashing algorithm, during which parameters for crypto-code constructions based on modified elliptic codes and initialization vectors are generated. This ensures maximum conversion speed. In this scenario, the task of encrypting a message can be parallelized with the formation of its MAC code. On the receiving end, these actions are also executed in parallel, thereby ensuring the required level of efficiency. A block diagram illustrating the formation of an authenticated message encrypted using the Rao-Nama system within a complex algorithm is depicted in Fig. 1.

The algorithm is executed in the following steps:

Step 1. A private session key is generated, and the parameters of the elliptic curve are selected:

$$K=\{a_1,a_2,\dots a_6\},\mathrm{a}_i\ \forall GF(q)$$
$$\text{Private key}\ \ G^{EC_i},X^i,Z^i,P^i,D^i,e. \tag{1}$$

Additionally, if necessary, initialization vectors are generated – |IV1|, |IV2| (for parameterization and formation of shortened and/or elongated MEC).

A formal representation of the input of information to the first level of the UMAC cascade algorithm:

$$\begin{aligned}
Hash_{L1_i} &= Hash_{L1_i} +_{64}((M_{i_{j+0}} +_{32} K_{L1_{j+0}})\times_{64}(M_{i_{j+4}} +_{32} K_{L1_{j+4}})),\\
Hash_{L1_i} &= Hash_{L1_i} +_{64}((M_{i_{j+1}} +_{32} K_{L1_{j+1}})\times_{64}(M_{i_{j+5}} +_{32} K_{L1_{j+5}})),\\
Hash_{L1_i} &= Hash_{L1_i} +_{64}((M_{i_{j+2}} +_{32} K_{L1_{j+2}})\times_{64}(M_{i_{j+6}} +_{32} K_{L1_{j+6}})),\\
Hash_{L1_i} &= Hash_{L1_i} +_{64}((M_{i_{j+3}} +_{32} K_{L1_{j+3}})\times_{64}(M_{i_{j+7}} +_{32} K_{L1_{j+7}})).
\end{aligned} \tag{2}$$

Step 2. The formalization of the process of forming a hash code at the second layer of the cascade hashing algorithm is presented as follows: $Y_{L2}=\left(M_{P_n}+kM_{P_{n-1}}+\dots+k^{n-1}M_{P_1}+k^n\right)\mathrm{mod}(p)$. The calculation is performed using an iterative procedure for all $i=1,2,\dots n$: $Poly_i=\left(kPoly_{i-1}+M_{P_i}\right)\mathrm{mod}(p), Poly_0=1, p=prime(Wordbits)$, with the following Horner scheme:

$$\begin{aligned}
&M_{P_n}+kM_{P_{n-1}}+\dots+k^{n-1}M_{P_1}+k^n=\\
&(((k+M_{P_1})k+M_{P_2})k+\dots+M_{P_{n-1}})k+M_{P_n}
\end{aligned} \tag{3}$$

Step 3. Encrypting the plaintext M based on Rao-Nama CCC on EC and MEC in CCC/HCCC:

$$c_x^*=M_i\times G_X^{EC_i}+e, \tag{4}$$

$$c_z^*=(M_i\times G^{EC_i}+e)\times Z, \tag{5}$$

where $G_X^{EC_i}=X^i\times G^{EC_i}\times P^i\times D^i$ is the public key in the Rao-Nama CCC.

MAC code generation using *Pad* substrate based on CCC/HCCC on modified elliptical curves:

$$Y_{L3}=\left(\left(\left(\sum_{i=1}^{m}Y_{L2_i}K_{L3_{1i}}\right)\mathrm{mod}(prime(36))\right)\mathrm{mod}(2^{32})\right)xor(K_{L3_2}), \tag{6}$$

$Y=Y_{L3}\oplus Pad.$

Step 4. A formal representation of the formation of an authenticated message based on the concatenation algorithm by CCC and HCCC is presented as:

$$C_{c_x}^Y=Y\,\|\,c_x^*. \tag{7}$$

$$C_{c_z}^Y=Y\,\|\,c^*. \tag{8}$$

When implementing the proposed approach in smart technologies, it is suggested to utilize the Rao-Nama HCCC on lossy codes (DC). This should concurrently reduce energy and computational costs, thereby ensuring stability. For this purpose, the lossy code (DC) technology and division into two channels are proposed. The first channel transmits the damage, while the second channel conveys the damaging text. The block diagram of the formation of an authenticated message based on HCCC at DC is illustrated in Fig. 2.

The primary distinction from the algorithm on modified elliptic codes is that the damage is applied after the authenticated message is generated, resulting in the addition of another step to the MV2 damage algorithm [38].

The main advantage of the proposed methods and protocols for providing security services based on lossy codes is the utilization of Rao-Nama NCCC to ensure cryptoresistance to loss and/or lossy text, rather than

In works [39, 40], a cyclic algorithm for obtaining damaging texts is understood as a universal damage mechanism (*Cm*, where *m* represents the number of cycles). This algorithm involves randomly replacing the bit representation of each character in the source text with a tuple of fewer or equal bits. Subsequently, concatenation is necessary. These algorithms facilitate the generation of an authenticated message based on post-quantum algorithms [41]. This, in turn, not only provides the required level of stability and processing speed (efficiency) but also enhances reliability through the use of noise-resistant algebraic geometric codes. These actions contribute to the necessary improvement of the SSL/TLS protocol [42].

III. Improvement of the SSL/TLS protocol based on the CCC/UMAC and HCCC/UMAC algorithms of the Rao-Nama symmetric cryptosystem

Enhancing the SSL/TLS protocol entails employing complex algorithms represented by post-quantum algorithms. Such algorithms ensure the implementation of security services while addressing vulnerabilities during the handshake phase [10, 43].

At the present stage, two versions of the SSL/TLS protocol are known and used - v.1.2 and v.1.3, the architecture of which includes the handshake protocol and the write protocol. The purpose of each of them is to authenticate and exchange keys for the handshake protocol and to log encryption for the write protocol.

The handshake protocol is implemented through the procedures of agreeing on the version of the protocol, selecting a cryptographic algorithm or set of ciphers, authentication using asymmetric cryptography, and determining a common secret key that will be used for symmetric encryption at the next level.

And the recording protocol is implemented through the procedures of encrypting outgoing messages with the secret key, transmitting encrypted messages from the client to the server, checking the received encrypted messages for changes by the server, and decrypting encrypted messages in the absence of changes using the secret key.

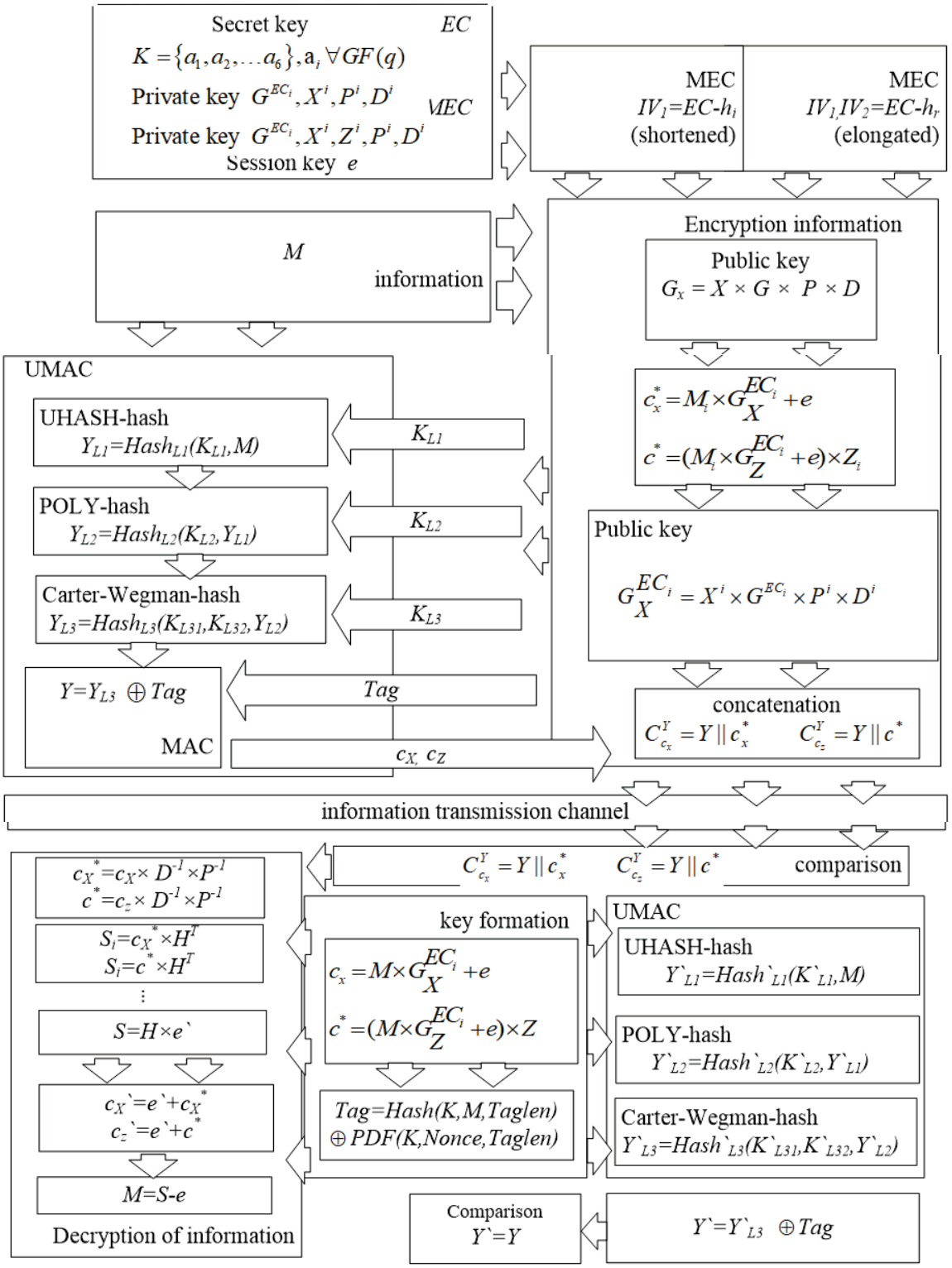


Fig. 1. Structural diagram of the formation of an authenticated message in the complex modified UMAC algorithm on modified crypto-code constructions (MCCC) of Rao-Nama on modified elliptical codes (MEC) (shortened, extended codes)

The assurance that an encrypted message remains unaltered during transmission is provided by the use of authenticated encryption in the TLS v.1.3 protocol (AEAD mode) [44, 45].

However, the algorithms employed in versions 1.2 and 1.3 are not post-quantum algorithms, thus failing to guarantee the necessary security level in the face of a fully operational quantum computer. To deliver security services and eliminate vulnerabilities during the handshake phase, it is proposed to transmit only the parameters of the curve equation and, if necessary, the MEC initialization vectors.

Consequently, to generate key data on both the Server and Client sides, it suffices to transfer only the coefficients a_1, a_2, a_3, a_4, a_5, a_6 (a binary sequence of five characters). This enables the Client and Server to form the necessary matrices and prepare for information exchange, eliminating the need for additional asymmetric algorithms (Diffie-Hellman, RSA) to transfer key data for the symmetric algorithm. In this scenario, symmetric cryptosystems with rapid cryptographic transformations are employed.

Regarding the session in 0-RTT mode, if session restoration is required, there is no need to exchange key data or utilize part of the encrypted code to verify the shared key's accuracy [44]. The client merely needs to send the curve coefficients for the Server to ascertain the key data for CCC/HCCC.

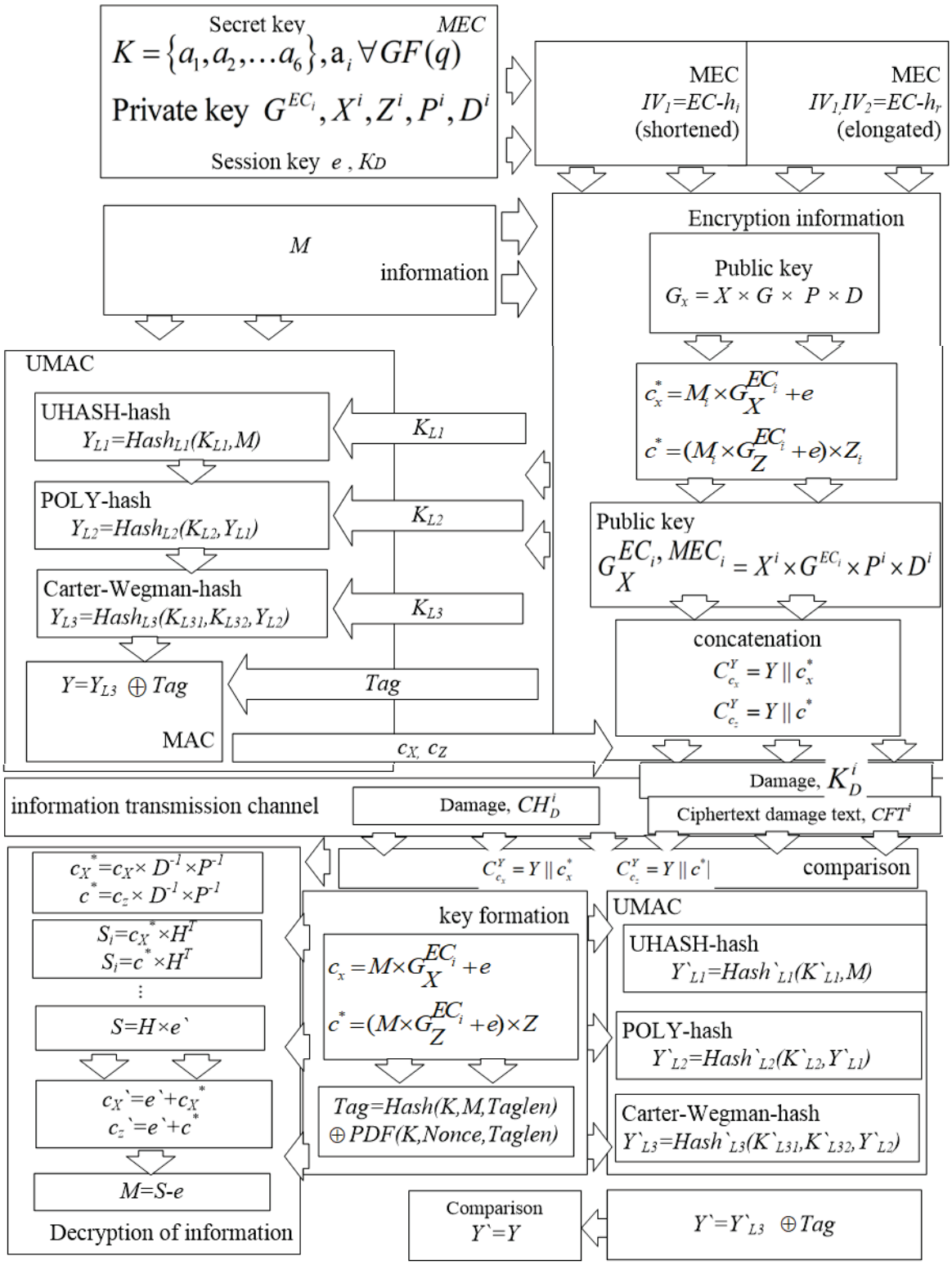


Fig. 2. Structural diagram of the formation of an authenticated message in the complex modified UMAC algorithm on Rao-Nama hybrid crypto-code constructions (HCCC) on modified elliptic codes (MEC) (shortened, extended codes) with damage (DC)

Moreover, employing the error vector e during information transmission allows it to Serve as a session key for each individual packet, significantly enhancing stability. Fig. 3 illustrates a structural diagram of the improved SSL/TLS protocol in 0-RTT mode [45].

This approach mitigates potential replay attack vulnerabilities and ensures the required security level (perfect forward secrecy). The utilization of symmetric crypto algorithms (post-quantum algorithms – CCC/HCCC Rao-Nama on MEC and DC) simplifies key exchange and minimizes additional energy (computational) costs associated with symmetric crypto algorithms.

Furthermore, the application of noise-resistant codes enables manipulation of the error vector e, which is utilized as a session password for each packet, thereby enhancing stability and reliability.

This approach is proposed to enhance the protocol for providing security services at the transport layer of SSL/TLS using post-quantum algorithms such as CCC/NCCS on Rao-Nama with modifications through shortening and/or lengthening. This strategy significantly reduces the implementation of known vulnerabilities in the SSL/TLS protocol, ensuring the necessary security level in the post-quantum crypto period while meeting computational and energy requirements for use in cyber-physical systems based on smart technologies.

To conduct statistical studies on the stability of the examined cryptosystems, the NIST STS 822 package [11, 44] will be utilized. The study results are presented in [27, 42, 46].

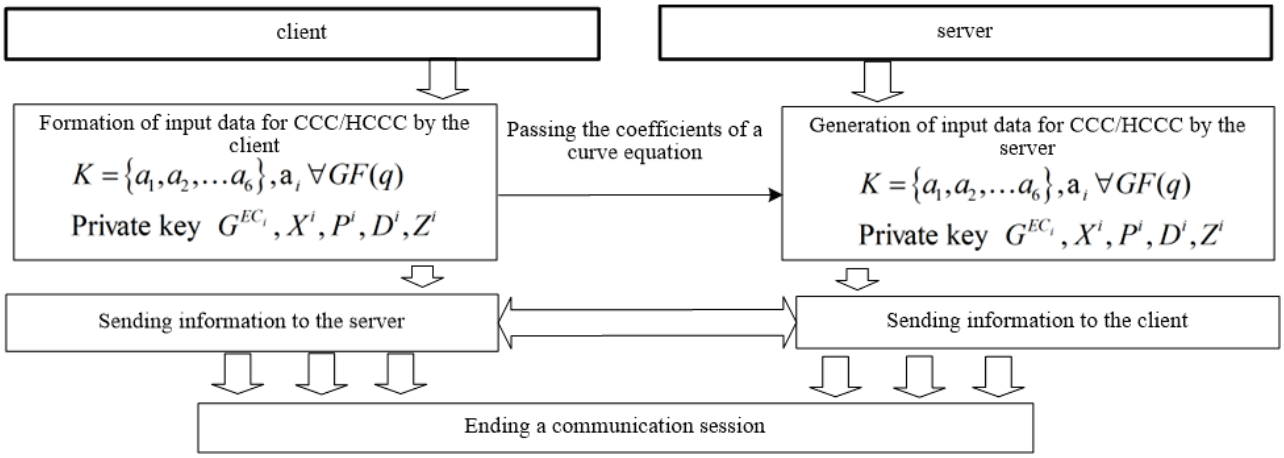


Fig. 3. Structural diagram of the improved SSL/TLS protocol on CCC/UMAC and HCCC/UMAC utilizing the Rao-Nama symmetric cryptosystem in 0-RTT mode

TABLE 1. RESULTS OF COMPARATIVE ANALYSIS OF STATISTICAL SECURITY STUDIES OF MCALIS'S CCCS/HCCC AND RAO-NAM'S CCC/HCCC

Asymmetric cryptosystems	Number of tests in which more than 99% of sequences passed the test		Symmetric cryptosystems
CCC McAlice on *ES*	149 (78.83%)	149 (78.3%)	CCC Rao-Nama on *EC*
CCC McAlice on shortened *MEC*	151 (79.89%)	149 (78.8%)	CCC Rao-Nama on shortened *MEC*
McAlice's CCC on extended *MEC*	152 (80.42%)	150 (79.3%)	CCC Rao-Nama on elongated *MEC*
McAlice's HCCC on extended *MEC* + loss codes	**153** (80.95%)	**151** (79.9%)	HCCC Rao-Nama on extended *MEC* +loss codes
McAlice's HCCC on shortened *MEC* + loss codes	**155** (82%)	**153** (80%)	HCCC Rao-Nama on shortened *MEC* + loss codes

According to the results shown in Table 1, the Galois field strength decreases to GF(26) for CCC and GF(24) for NCCS, respectively. However, this does not impact the statistical characteristics of such CCC/NCCS. They have proven to be at least as effective as traditional McElwain CCC/NCCS at GF(210) and Rao-Nam CCC/NCCS at GF(210). All cryptosystems successfully passed 100% of the tests. The best performance was exhibited by NSSS on shortened MEC: 155 out of 189 tests passed at the level of 0.99, representing 82% of the total number of tests (for McAllis); 153 out of 189 tests passed at the level of 0.99, accounting for 80% of the total number of tests (for Rao-Nama). A similar trend is observed for HCCC on extended MEC.

Simultaneously, the traditional CCS for both McAleese and Rao-Nam on GF(2 10) demonstrated 149 tests at the level of 0.99, with a negligible difference of 0.5% considered insignificant.

The research findings indicate that the proposed approach to utilizing crypto-code constructions on Rao-Nama facilitates the encryption of substantial data volumes circulating within information and communication systems. The primary advantage lies in the preservation of stability and performance levels while ensuring the required reliability.

CONCLUSION

Structural schemes for forming an authenticated message in the complex modified UMAC algorithm on modified crypto-code constructions (MCCC) of Rao-Nama on modified elliptical codes (MEC) (shortened, extended codes) and on hybrid crypto-code constructions (HCCC) of Rao-Nama on modified elliptical codes (MEC) (shortened, extended codes) with damage (DC) have been developed.

An enhanced SSL/TLS transport layer protocol based on complex algorithms – Rao-Nama crypto-code constructions on modified elliptic codes and lossy codes with an improved UMAC cascade hashing algorithm is proposed. This approach aims to significantly diminish the effectiveness of known attacks on the protocol by synthesizing asymmetric cryptosystems and the improved UMAC algorithm.

The study results affirm the proposed approach's capability to encrypt large data volumes without compromising security and performance levels. Furthermore, the required reliability is ensured.

It has been validated that the improved SSL/TLS transport protocol, based on post-quantum CCS/NSSS algorithms utilizing the Rao-Nama code generation scheme (with shortening and/or lengthening), significantly mitigates vulnerabilities within this protocol. Simultaneously, it guarantees the necessary security level in the post-quantum crypto era, alongside meeting computational and energy-intensive requirements for employing these codes in cyber-physical systems based on smart technologies.

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Clinical Applications and Future Directions of Apple Vision Framework for Quantitative Assessment of Hand Motor Dysfunction

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***Abstract*—This study presents a markerless real-time hand motor assessment system for Parkinson's disease diagnosis using Apple Vision Framework. The system analyzes finger tapping tests through 21-point hand pose detection on the iPhone 15 and iPad mini devices with a focus on device performance. Key motor parameters including taps per second, coefficient of variation, amplitude, and decrement are calculated automatically. Experimental validation demonstrates reliable performance with up to 26 milliseconds inference latency on mobile device hardware backend. The purpose of the study is to validate a native Apple Vision–based pipeline for the Finger Tapping Test on recent Apple mobile devices. We implement hand-pose tracking and quantify on-device latency and frame rate across the neural, graphics, and central processors. Clinical evaluation is outside the scope of this study.**

Keywords—healthcare, information technologies, medical diagnostics, hand movements disorders, tremor, object detection, image processing, pose tracking.

I. Introduction

Fine motor control of the hand, particularly the ability to rapidly and accurately approximate the thumb and index finger, represents a critical biomarker of central and peripheral nervous system functional status [1]. In clinical practice, the Finger Tapping Test is included in the [2], enabling clinicians to visually evaluate velocity, amplitude, and consistency of repetitive movements. Analogous assessments—including sequential finger opposition testing and pinch strength evaluation—are employed for diagnosing cerebellar ataxias, peripheral nerve injury-related paresis, and monitoring post-stroke recovery [3],[4].

Traditional visual assessments are constrained by subjectivity and limited capability in detecting subtle movements and temporal delays [5]. In contrast, computer-based systems capture substantially more precise information for accurate diagnosis [6].

This paper proposes evaluating the feasibility of implementing a markerless real-time motion capture solution based on Apple Vision Framework [7]. The system allows real-time detection of finger pinching events and calculation of key motor parameters including TPS (taps per second), CV (coefficient of variation), amplitude, and decrement.

The study [8] compares smartphone-compatible approaches to tremor diagnosis and includes Apple Vision Framework; however, similar to [9], it does not address Parkinson's disease assessment via the Finger Tapping Test. In contrast, this work evaluates the feasibility of a fully native, on-device pipeline built on Apple Vision Framework, rather than cross-platform solutions such as MediaPipe or other generic frameworks.

II. Materials and Methods

We selected the iPhone 15 and iPad mini (6th generation) for implementation and testing to contrast two recent processor generations and the presence versus absence of front-facing depth sensing. Both devices offer sufficient front-camera resolution and native support for the Apple Vision framework. Technical specifications are listed in Table I.

TABLE I. Device Specifications

Model	Front Camera	Chipset	True Depth
iPhone 15 (2023)	12 MP 4K 60fps	A16 (16 cores NE)	Yes
iPad mini (6th gen)	12 MP 1080 HD 60fps	A15 (16 cores NE)	No

Native implementation of these models enables utilization of the Neural Processing Unit (NPU) for accelerated computation with reduced power consumption. The hand pose detection model [10] for recognizing hand keypoints provides information about 21 points, as illustrated in Fig. 1.

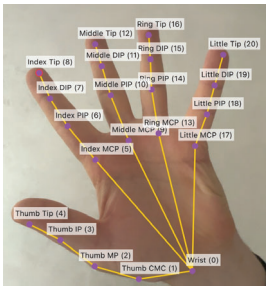


Fig. 1. VNDetectHumanHandPoseRequest [10] model hand keypoints detection results

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The presence of TrueDepth camera on the iPhone 15 enables more precise determination of distance to fingers and, consequently, more accurate detection of peak separation amplitude while compensating for inaccuracies caused by changes in the distance between the hand and the camera.

We use camera configuration matrix K (1) in (2) used to determine distance to the [11, 12].

$$K = \begin{bmatrix} f_x & 0 & o_x \\ 0 & f_y & o_y \\ 0 & 0 & 1 \end{bmatrix}; \quad (1)$$

$$D_{avg} = \frac{\sqrt{(\Delta u)^2+(\Delta v)^2}\times Z_{avg}}{f_x} \times 1000, \quad (2)$$

Where:

Δu and Δv is $u_1 - u_2$ and $v_1 - v_2$ respectively;

(u_1, v_1) and (u_2, v_2) – pixel coordinates of index and thumb fingers;

f_x is focal length;

Z_{avg} is mean distance from camera to fingers (we assume that the user is holding their hand parallel to phone).

For TrueDepth cameras, $f_x \approx f_y$ due to square pixel geometry, thus we use f_x in calculations.

Z_{avg} is obtained directly from the data captured by the depth sensor. Through experimental validation, we determined that optimal accuracy is achieved when the subject's hand is positioned precisely in the center of the viewfinder [13], with the palm oriented parallel to the phone and positioned within 30-45 cm from the device. This test configuration minimizes noise artifacts that inevitably occur during depth data acquisition. We calculate the averaged distance to each finger using (3).

$$Z_{avg}(u_k, v_k) = \frac{1}{N_k}\sum_{i=u_{k,min}}^{u_{k,max}} \sum_{j=u_{k,min}}^{u_{k,max}} D(u, v), \quad (3)$$

Where N_k denotes the number of pixels in the rectangular region of interest around the k-th finger; the bounds are inclusive and k represents the finger index, where 1 denotes the thumb and 2 denotes the index finger.

Amplitude is calculated by taking the maximum value from the amplitude buffer, which records distances between fingers that do not fall within the threshold at which we consider them to be in contact. Fig. 2 illustrates the two hand states captured by the application.

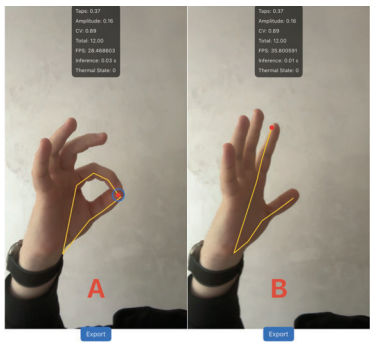


Fig. 2. Two states of touch detection (A – touching, B – not)

For calculating amplitude, CV, TPS, and displaying the amplitude decrement plot, we employed (4), (5), and (6):

$$TPS = \frac{1}{\bar{\Delta t}}; \quad (4)$$

$$CV = \frac{\sigma(\Delta t)}{\Delta t}; \quad (5)$$

$$amp_{rel,i} = \frac{\widetilde{amp}_i}{M_{amp}}, \quad (6)$$

Where:

$\widetilde{amp}_i$ – smoothed amplitude value;

Δt – time difference between taps;

$\sigma(\Delta t)$ – standard deviation.

III. Experimental Results

Experimental results were conducted in the laboratory of Medical and Biological Researches of the Department of Biomedical Engineering of Kharkiv National University of Radio Electronics.

On the iPad mini, we omitted distance-based amplitude correction: the hand was fixed at 35 cm (within a 30–45 cm range), and relative amplitudes were scaled to centimeters using a ruler-measured reference peak between the thumb and index finger. The 35 cm nominal distance keeps the hand centered, limits motion blur at 30 fps, ensures a comfortable posture, and did not change touch/non-touch classification under ±5 cm, so it was fixed for comparability.

Two healthy adults without known neurological or musculoskeletal disorders participated. The experiment validates signal extraction and on-device performance; clinical accuracy is left for future work.

Testing was conducted in two trials for each participant with a session duration of 30 seconds on each device.

Fig. 3 and 4 present the averaged results of two testing trials for two participants.

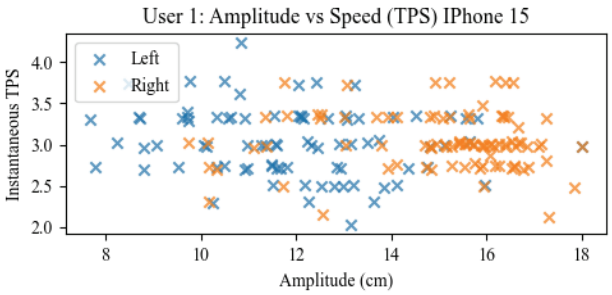


Fig. 3. Amplitude vs TPS test results for user 1

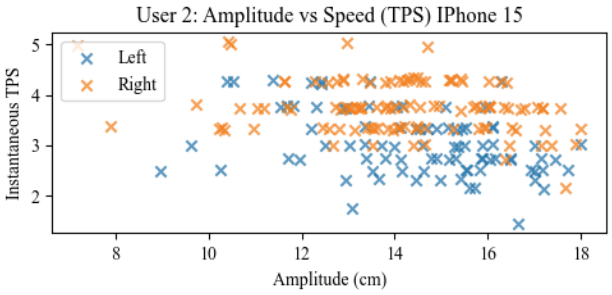


Fig. 4. Amplitude vs TPS test results for user 2

Subsequently, we present amplitude and velocity dynamics plots for the first and second subjects in Fig. 5 and Fig. 6.

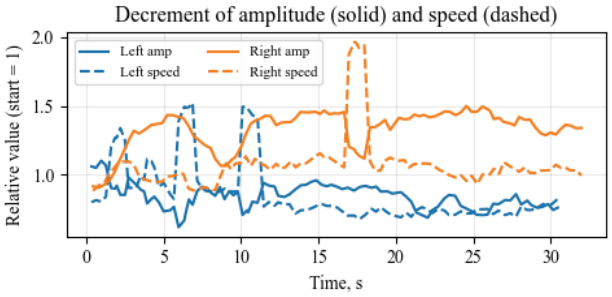


Fig. 5. Motor performance dynamics for user 1

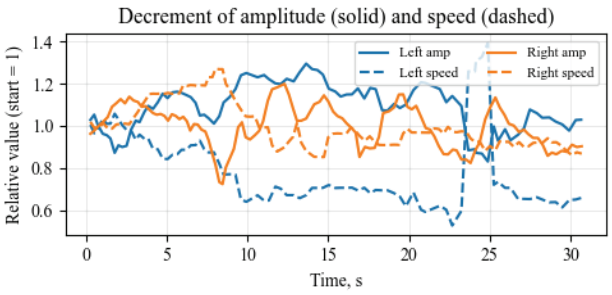


Fig. 6. Motor performance dynamics for user 2.

This graph demonstrates a trend toward decreasing velocity (TPS) with increasing movement amplitude over the duration of the testing session.

Table II presents the testing results for two subjects. TPS is average tapping rate, CV is relative variability of inter-tap intervals, Avg Amp (cm) is mean thumb–index separation, Slope (cm/s) is temporal trend of amplitude (negative indicates decrement); %/min is amplitude trend normalized per minute.

TABLE II. FINGER TAPPING TEST RESULTS

User	Hand	TPS	CV	Avg Amp (cm)	Slope (cm/s)	%/min
A	L	2.873	0.290	14.28	-0.0113	-4.76
A	R	3.652	0.161	14.12	-0.0484	-20.56
B	L	3.118	0.230	11.64	-0.0640	-32.99
B	R	3.000	0.182	14.79	0.0918	37.23

Our primary focus was on model performance metrics recorded during the test execution. These metrics consist of two key indicators: model inference latency and frames per second. The test results are shown in Fig. 7.

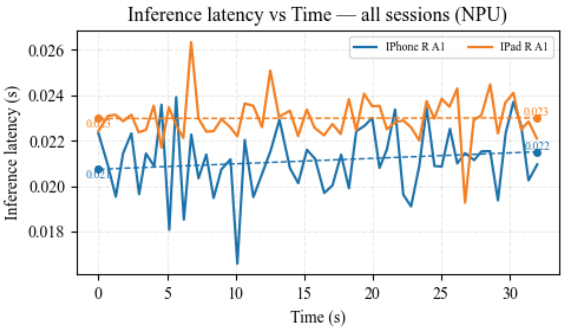


Fig. 7. Inference over time on iPhone 15 vs iPad mini (6th gen) on NPUs

We observe the expected pattern: inference latency on the iPhone 15 is lower than on the iPad by 3-5ms on average. This is attributed to the iPhone having a newer and more powerful next-generation chipset. In previous work [14], we repeatedly demonstrated that when comparing the iPhone 13 and the iPad with same-generation chipsets, they exhibited identical performance when using NPU as the backend for primary computational load. In this comparison, the results favor the iPhone primarily because VNDetectHumanHandPoseRequest utilizes a proprietary lightweight model that induces thermal throttling more slowly, whereas the iPad has an advantage due to its larger unibody aluminum chassis that better handles heat dissipation. The graph below shows similar FPS metrics for each device, demonstrating no frame loss during overall camera data processing (Fig. 8).

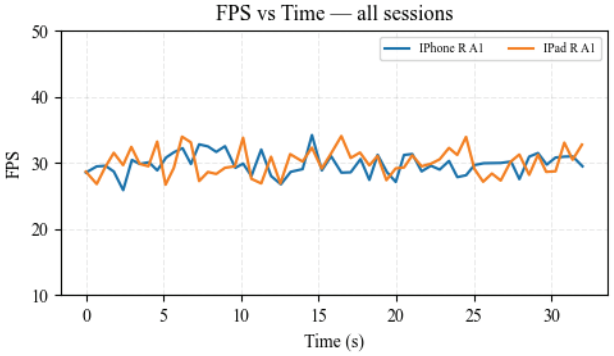


Fig. 8. FPS over time on iPhone 15 vs iPad mini (6th gen)

We also measured the impact when using device CPU and GPU as the primary backend. Fig. 9 illustrates the performance comparison when running on GPU.

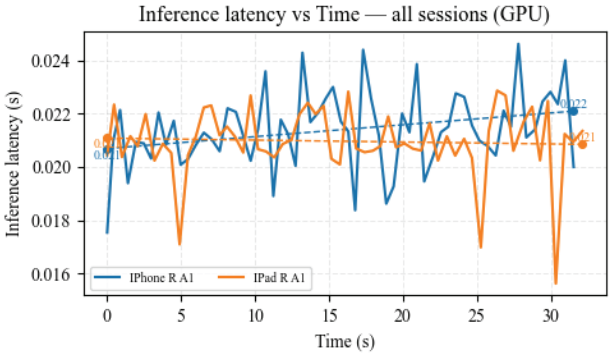


Fig. 9. Inference over time on iPhone 15 vs iPad mini (6th gen) on GPUs

This graph demonstrates how the initially modest GPU advantage of the iPhone gradually diminishes as temperature increases due to the factors described above.

Fig. 10 displays a comparison of CPU backend performance characteristics.

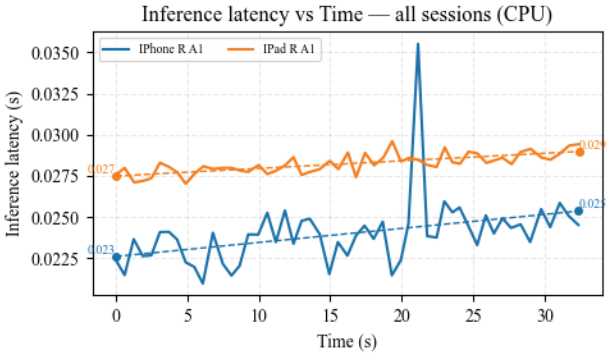


Fig. 10. Inference over time on iPhone 15 vs iPad mini (6th gen) on CPUs

This graph clearly shows a trend toward increasing latency, although the iPhone has a pronounced advantage due to its more powerful processor. The gradual increase in average values demonstrates the consequences of CPU heating. In this case, considering that the CPU is tasked with both UI rendering and application logic execution.

IV. Conclusions

A markerless hand motor analysis system based on native Apple Vision Framework utilizing a 21-point hand keypoint detection model has been developed and experimentally validated. The system demonstrates technical feasibility for automated real-time calculation of primary motor parameters (TPS, CV, amplitude, and decrement).

Technical validation of the system demonstrated detection stability and computational algorithm correctness in obtaining quantitative movement parameters. Experimental data confirmed the viability of the proposed approach with inference latency up to 26 ms on NPU backend and capability for continuous recording of motor parameters during 30-second sessions. Additionally, we determined the optimal distance for minimizing noise artifacts in depth data, which ranges from 30-45 centimeters from the smartphone camera.

Comparative analysis of computational backends revealed NPU optimality for this task: absence of performance degradation during prolonged operation, unlike GPU and CPU which are susceptible to thermal throttling.

The presence of TrueDepth camera on the iPhone 15 provides additional amplitude determination accuracy through compensation of distance-to-object variations. Proprietary optimization of VNDetectHumanHandPoseRequest for Apple Silicon architecture ensures energy efficiency and stable frame rate of ~30 FPS without frame loss during video stream processing, confirming the technical readiness of the solution for potential clinical application.

This study establishes technical feasibility and performance characteristics on commodity Apple devices; clinical validation on patient cohorts remains future work.

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Analysis Of The Impact Of Some Clinical Blood Parameters On The Severity Of COVID-19 Disease

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***Abstract*—The aim of the study is to substantiate the diagnostic impact of clinical blood parameters in patients with COVID-19. The retrospective analysis covered 370 medical records of patients of the center for medical care for patients with COVID-19 in Kharkiv. The analysis included the evaluation of individual clinical blood parameters, the severity of the disease, the presence of general otolaryngologic symptoms and olfactory disorders. Statistical analysis included discriminant analysis with the calculation of the normalized Euclidean distance to determine the diagnostic significance of clinical blood parameters. The statistical analysis showed that clinical blood counts do not have a significant diagnostic contribution to the discrimination model, which already contains COVID-19-specific markers. Thus, olfactory impairment, as associated with acute rhinosinusitis in COVID-19, can serve as a prognostic marker for a more favorable course of pneumonia in hospitalized patients.**

Keywords—blood laboratory indicators, COVID-19, good health and well-being, human health, medical diagnostic, respiratory and olfactory disorders

I. Introduction

The emergence of the pandemic caused by the novel SARS-CoV-2 virus in 2020 has profoundly impacted global health systems, societal structures, and economies [1]. The rapid spread of the virus demanded unprecedented efforts from the global medical community to develop effective diagnostic tools, preventive measures, and treatments. Specialists were tasked with identifying specific clinical and laboratory methods to ensure the swift and accurate diagnosis of COVID-19, which is critical for containing outbreaks and optimizing patient care. According to the World Health Organization (WHO), as of November 2023, over 770 million confirmed cases of COVID-19 have been reported worldwide [2]. This staggering number includes more than 6 million deaths, underscoring the pandemic's devastating toll on human life and the urgency of advancing diagnostic and therapeutic strategies to combat the virus. The search for reliable markers and methods remains a priority as the world continues to navigate the long-term implications of the pandemic.

The severity of clinical manifestations of COVID-19 varies widely among individuals and is influenced by factors such as age, comorbidities, immune response, and viral load. Mild cases are often characterized by general symptoms such as fever, fatigue, and cough, or in some instances, patients may remain asymptomatic while still being infectious. However, in moderate to severe cases, the disease progresses to more pronounced symptoms, including significant respiratory distress and hypoxia [3].

In severe cases, the infection can escalate to SARS-CoV-2-associated pneumonia, marked by diffuse alveolar damage and inflammatory infiltrates, which may necessitate mechanical ventilation. Such complications often contribute to the development of severe acute respiratory syndrome (SARS), a condition associated with high mortality. Additionally, systemic involvement of the virus can lead to multi-organ dysfunction, including acute kidney injury or renal failure, often exacerbated by hyperinflammatory states such as cytokine release syndrome [4].

On the other hand, the variability in symptom severity also poses diagnostic and management challenges, as asymptomatic carriers can unknowingly contribute to community transmission. This spectrum of clinical outcomes underscores the need for personalized approaches to diagnosis, monitoring, and treatment to mitigate the disease's impact across diverse patient populations.Therefore, the aim of the study is to substantiate the diagnostic significance of some clinical blood parameters in patients with olfactory dysfunction in COVID-19.

II. Review Of Literary Sources

During the pandemic, researchers have identified a number of specific symptoms associated with COVID-19. Particular attention is drawn to olfactory disturbances, in particular anosmia and hyposmia. Loss or significant reduction of olfactory function is one of the most characteristic early manifestations of infection, making it an important diagnostic marker [5, 6]. Studies have shown a higher incidence of olfactory dysfunction in patients with mild coronavirus disease who did not require hospitalization, which may be due to an enhanced immune response in the upper respiratory tract. [7, 8]. Thus, special attention is paid to saturation and pathological changes in the lungs on X-ray examination in the case of SARS-CoV-2-associated pneumonia [9].

Studies have emphasized the prognostic value of a number of serum inflammatory markers in coronavirus infection, including D-dimer, C-reactive protein (CRP), procalcitonin, ferritin, and lactate dehydrogenase (LDG). The determination of these indicators can help assess the severity of the disease and predict its course [10].

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Despite the widespread use of these indicators in clinical practice, their role in COVID-19-associated pneumonia remains poorly understood. It is also advisable to determine the role of olfactory sensitivity disorders [11] in different courses of COVID-19 and to assess the impact of the anatomical configuration of the nasal cavity on the loss of olfactory sensitivity [12-14].

The development and improvement of clinical treatment protocols require the use of diagnostic methods with high reliability that ensure the effectiveness of subsequent therapy [15].

Therefore, an urgent task is to expand the diagnostic capabilities of existing research methods, study the significance of measurement indicators, and justify the feasibility of their use in the diagnosis of specific pathologies.

This approach is especially relevant in the context of the widespread introduction of objective instrumental diagnostic methods in accordance with modern criteria of evidence-based medicine. An important task is to select informative parameters of diagnosis and control, as well as criteria for comparing the discriminant capabilities of methods in the differential diagnosis of pathological conditions and severity of diseases [15], in particular for COVID-19. It is also relevant to evaluate some laboratory blood parameters to improve the determination of the severity of COVID-19.

III. Materials And Methods

The retrospective analysis included 370 case reports of patients treated at the COVID-19 care center at the Kharkiv Regional Clinical Hospital between December 2020 and February 2022. The inclusion criteria were age 18 years, PCR-confirmed SARS-CoV-2-associated pneumonia, and the presence of olfactory and gustatory disorders. The analysis included an assessment of age, gender, duration of hospitalization, clinical course of the disease, presence of olfactory and taste disorders, and general otolaryngological symptoms. In addition, the level of certain laboratory markers of inflammation, including fibrinogen, procalcitonin, IL-6, leukocyte and lymphocyte counts, and erythrocyte sedimentation rate (ESR), was investigated. The SNOT-22 (Sino-Nasal Outcome Test) questionnaire was used to assess the presence and severity of sinonasal symptoms [16].

To determine the diagnostic significance of diagnostic methods for different severity levels of SARS-CoV-2-associated pneumonia, a discriminant analysis was performed by determining the normalized Euclidean distance and the probability of diagnostic decision error [15] using the appropriate formulas.

$$\delta = \sqrt{\sum_{i=1}^{n}\left(\frac{m_i^{(0)} - m_i^{(1)}}{\sigma_i}\right)^2} \quad (1)$$

$$P_{er} \le 1 - \Phi(\delta / 2) \quad (2)$$

where $m^{(0)}, m^{(1)}$ are the mean and standard deviations of the random variable X in the normal and pathological conditions, respectively,

σ_i is the standard deviation of the i-th indicator, which is determined by the formula

$$\sigma_i = \max(\sigma_i^{(0)}, \sigma_i^{(1)}) \quad (3)$$

where $\sigma^{(0)}, \sigma^{(1)}$ are the mean square deviations of X in the normal and pathological state, respectively,

$\Phi(\cdot)$ - the value of the Laplace probability integral [5].

In addition to the basic ones (with the integral contribution of specific COVID-19 indicators), 6 additional informative parameters X_i were used based on clinical blood counts (in general $i = \overline{1,7}$), which are displayed in the following order:

X_1 - specific indicators of COVID-19: impaired sense of smell, impaired taste, CRP, D-dimer, SpO2, CBC data;

X_2 - ESR;

X_3 - Fibrinogen;

X_4 - Procalcitonin;

X_5 - IL-6;

X_6 - White blood cells;

X_7 - Lymphocytes.

The data of each patient were analyzed according to a specific set of characteristics with the calculation of mean values and standard deviations for each indicator. Next, we calculated normalized Euclidean distances separately for each feature and integral normalized Euclidean distances for the set of indicators of each diagnostic method using formula (1). At the next stage, the normalized Euclidean distances were calculated step by step with the gradual addition of the diagnostic value of each method, and the corresponding diagnostic errors were determined using formula (2) [15].

The average age of the patients was 51.8 years ± 11.48 years (range 18-70), 215 women (58.1%) and 155 men (41.8%). Patients were divided into 4 groups (with mild, moderate, severe and ultra-severe Covid-19 disease) and a control group. Initial data on blood laboratory parameters are shown in Table 1. The study was conducted in accordance with the ethical standards of the Declaration of Helsinki and was approved by the bioethics committee of Kharkiv National Medical University.

TABLE I. Mean Values And Standard Deviations For Clinical Blood Test Parameters In COVID-19 And Control Groups, Respectively

Indicator	1 group (n=25)	2 group (n=213)	3 group (n=114)	4 group (n=18)	Reference values
ESR, mm/h	16,8±7,2	25,6±14,3	28,3±15,4	35,7±18,2	7,5±7,4
Fibrinogen, g/l	3,3±0,8	4,2±1,55	4,4±1,33	6,3±3,4	2±1,7
Procalcitonin, ng/ml	0,08±0,06	0,28±0,11	0,29±0,17	0,27±0,17	0,05±0,015
IL-6, pg/ml	4,3±20	7,5±4,4	16,5±8,5	17,1±11,7	3,5±3,4
White blood cells, ×10⁹/l	5,4±2,2	6,4±3,8	6,8±3,7	9,6±3,5	6,5±2,5 x10*9/л
Lymphocytes, %.	23,6±10,9	18,1±11,6	17,2±8,8	6,1±3,6	30±10

IV. Results And Discussion

The results of calculating the normalized Euclidean distances δ for diagnostic indicators in each of the patient groups relative to the conditional norm are shown in Table 2. It should be noted that the basic contribution to the diagnostic decision (diagnostic error is about 0.1) is made by such specific indicators for COVID-19 as impaired sense of smell, taste, CRP, D-dimer, SpO2, and CBC data, the diagnostic significance of which is accumulated in the first indicator. The clinical blood test parameters presented in this paper allow only to clarify the diagnosis.

TABLE II. Normalized Euclidean Distances Δ For Diagnostic Indicators In Each Group Relative To The Conditional Norm (Control Group Data)

δ / Indicator	1 group	2 group	3 group	4 group
Total contribution of COVID-19 specific indicators	3,22	2,90	3,18	5,02
ESR	1,29	1,27	1,35	1,55
Fibrinogen	0,38	0,77	1,05	0,97
Procalcitonin	0,50	2,10	1,41	1,29
IL-6	0,40	0,91	1,52	1,16
White blood cells	0,50	0,03	0,08	0,69
Lymphocytes	0,57	1,03	1,46	2,49

The analysis of laboratory markers of inflammation showed that their changes correlated with the severity of pneumonia (Table 1), with mean values differing significantly between groups. In patients with mild disease, most of the indicators remained within the reference values or showed a slight increase, in particular C-reactive protein (CRP) and erythrocyte sedimentation rate (ESR). In patients with severe pneumonia, there was a significant increase in the levels of both D-dimer [10], ESR, and interleukin-6 (IL-6), which may indicate a more severe systemic inflammatory process.

The results of the discriminant analysis to determine the diagnostic significance of the examination methods showed (according to Figures 1, a and 1, b) that in patients with a mild form of the disease (without associated pneumonia), the maximum normalized Euclidean distances and the corresponding maximum reduction in the probability of diagnostic error are achieved by the first indicator (ESR).

Others generally have a contribution to the Euclidean distance of about 3 times less and reduce the probability of making a diagnostic decision to 0.07. The total contribution of clinical blood test parameters to the discrimination model in patients of the first group is only 0.05. In patients of the second group (with mild Covid-19-associated pneumonia), normalized Euclidean distances and a corresponding reduction in the probability of diagnostic error are maximized when Procalcitonin values are added to the discrimination model (4). ESR and fibrinogen also have a smaller but relatively significant impact. Other blood parameters considered do not significantly affect the discriminant properties of the model. In general, the addition of clinical blood counts to the model in patients with mild pneumonia reduces the resulting probability of diagnostic error by 0.12 (to an absolute value of 0.03). In patients of the third group (with moderate pneumonia associated with Covid-19), all clinical blood counts (except for leukocytes (6)) make almost equal contributions to the increase in the normalized Euclidean distance and the corresponding decrease in the probability of diagnostic error. In general, the addition of clinical blood counts to the model in these patients reduces the resulting probability of diagnostic error by 0.09 (to an absolute value of 0.03).

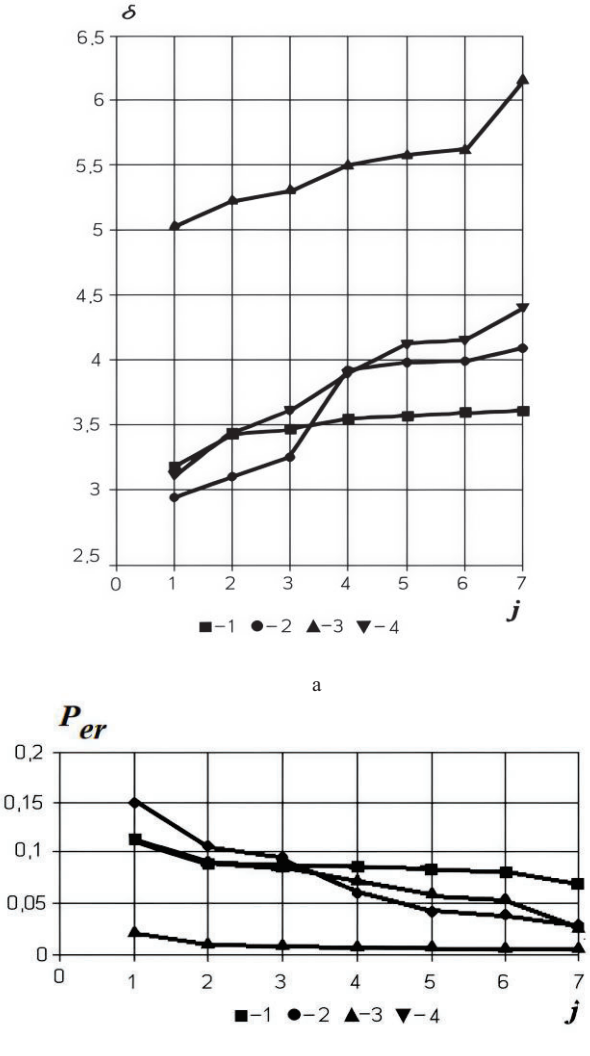
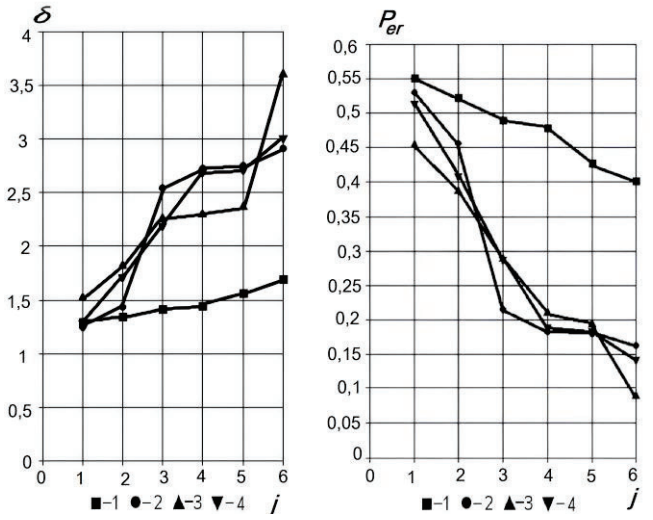


Fig. 1. a) Graph of the increase in the normalized Euclidean distance when the results of the following diagnostic methods are sequentially added to the resulting discrimination model ($i=\overline{1,7}$): 1 - specific indicators of COVID-19: impaired sense of smell, impaired taste, CRP, D-dimer, SpO2, CBC data; 2 - ESR; 3 - Fibrinogen; 4 - Procalcitonin; 5 - IL-6; 6 - Leukocytes; 7 - Lymphocytes. b) Graph of reduction of the diagnostic decision error in determining the severity of the disease when the results of the following diagnostic methods are sequentially added to the discrimination model ($i=\overline{1,7}$): 1 - specific indicators of COVID-19: impaired sense of smell, impaired taste, CRP, D-dimer, SpO2, CBC data; 2- ESR; 3 - Fibrinogen; 4 - Procalcitonin; 5 - IL-6; 6 - White blood cells; 7 - Lymphocytes (*graphs in the figures are marked with the appropriate symbols for patients: 1 - first group, 2 - second group, 3 - third group, 4 - fourth group*).

In patients of the fourth group (with severe Covid-19-associated pneumonia), the influence of clinical blood test parameters on the diagnostic decision allows to refine the already highly accurate model by reducing the discrimination error from only 0.02 to 0.001 (by 0.019). The largest contribution is made by lymphocytes (7). The other blood parameters considered contribute to the model quite evenly. For clarity, Figure 2, respectively, shows graphs of the

increase in the normalized Euclidean distance (Fig. 2, a) and the decrease in the diagnostic decision error (Fig. 2, b) in determining the severity of the disease when only selected clinical parameters are sequentially added to the discrimination model.



a b

Fig. 2. a) Graph of the increase in the normalized Euclidean distance when only selected clinical blood parameters are added to the discrimination model ($i = \overline{1,6}$): 1 - ESR; 2 - Fibrinogen; 3 - Procalcitonin; 4 - IL-6; 5 - Leukocytes; 6 - Lymphocytes. b) Graph of reduction of the diagnostic decision error in determining the severity of the disease when sequentially adding only selected clinical blood parameters to the discrimination model ($i = \overline{1,6}$): 1 - ESR; 2 - Fibrinogen; 3 - Procalcitonin; 4 - IL-6; 5 - White blood cells; 6 - Lymphocytes (*graphs in the figures are labeled with the corresponding symbols for patients: 1 - first group, 2 - second group, 3 - third group, 4 - fourth group*).

V. CONCLUSIONS

The study showed that specific COVID-19 markers already identified in the literature, in particular, anosmia in patients with COVID-19, are important diagnostic criteria not only in mild disease but also in severe disease. The statistical analysis shows that these clinical blood test indicators do not significantly add information to the discrimination model, which already includes COVID-19-specific indicators. The addition of clinical blood counts is advisable in patients with relatively mild COVID-19. This will reduce the probability of errors to 0.03 when making diagnostic decisions. In patients with moderate and severe SARS-CoV-2-associated pneumonia, specific COVID-19 indicators already allow diagnosing the corresponding condition with an error rate even lower (0.02), and the addition of clinical blood test indicators only sufficiently refines the discrimination model. In general, the above clinical blood test parameters do not make a significant contribution to the discrimination model and can be excluded altogether when diagnosing COVID-19 in different disease severity states. It should be noted that the basic contribution to the diagnostic decision (diagnostic error is about 0.1) is made by such specific indicators for COVID-19 as impaired sense of smell, taste, CRP, D-dimer, SpO2, and CBC data, the diagnostic significance of which is accumulated in the first indicator. The considered indicators of clinical blood test allow only to clarify the diagnosis.

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Containerization of Biomedical Research: Advantages of Docker for Scaling Cryoimage Analysis

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Abstract **With the growing volume of biomedical data and increasing complexity of analysis methods, containerization has become a key tool for ensuring reproducibility and scalability in research. This paper explores the advantages of using Docker to encapsulate the environment for cryomicroscopic image analysis, including modules for annotation, model training, and data processing. It highlights practical aspects of deploying containers across different machines, ensuring stable performance, code portability, and simplified collaboration among researchers. The use of Docker enables the unification of the research pipeline and accelerates development processes in biomedical projects.**

Keywords ***Docker, Containerization, Human Health, Biomedical image analysis, Cryomicroscopy, Reproducible research, Scalable pipelines, Model deployment, Image annotation, Computational biology, Open-source tools.***

I. INTRODUCTION

In recent years, there has been a growing interest in cryomicroscopy as one of the most accurate and informative methods in structural biology and medicine [1, 3]. By freezing biological samples at ultra-low temperatures without damaging their internal structures, cryomicroscopy enables the acquisition of high-resolution images of cells and molecular complexes in a state that closely resembles their natural condition.[2] These images are highly informative but also require complex processing, including annotation, segmentation, analysis, and model construction [1-4]. This raises the urgent need to automate and standardize the processes involved in analyzing such images. Researchers face several challenges in this regard: the complexity of setting up the working environment, difficulties in reproducing results across different machines, and the necessity to scale solutions to handle large volumes of data. Containerization, particularly through the use of Docker, emerges as a powerful solution to address these issues by creating isolated, portable, and easily managed environments.

Docker allows all components of the research system ranging from data storage services (such as MinIO), annotation tools (like CVAT), deep learning libraries (e.g., PyTorch, TensorFlow), to processing scripts in Jupyter to be packaged into a single reproducible container. This not only accelerates deployment but also ensures consistent performance across operating systems and hardware platforms (Fig. 1) [5].

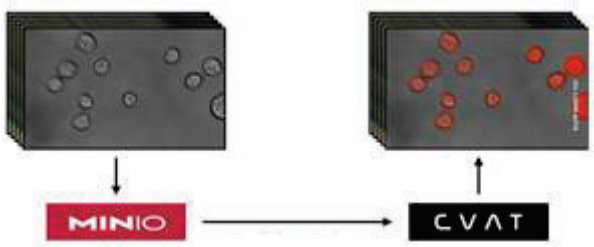


Fig. 1 How MinIO and CVAT work

The goal of this work is to demonstrate in detail the advantages of using Docker technology in the field of biomedical research by developing a processing pipeline for cryomicroscopic images. Particular attention is given to creating a convenient, scalable, and standardized software-hardware environment that enables a fully reproducible research cycle from data preparation and annotation to model training and result analysis [6].

This organization of the process addresses several key challenges. First, Docker ensures a high level of reproducibility: the entire environment, with all necessary dependencies, is "packaged" into containers that can be deployed on any server or local machine without complex configuration. This is especially important in collaborative work, where multiple researchers can use the same infrastructure and avoid errors caused by incompatible library or software versions.

Second, containerization significantly simplifies the scaling of experiments. When data volume increases, additional processing stages are introduced, or new model architectures are tested, users can quickly launch additional service instances without disrupting the core system. This flexibility makes the pipeline suitable for both local development and execution on powerful computing clusters or in cloud environments [7].

Moreover, the standardized containerized environment facilitates automation of repetitive operations such as model updates, testing, and exporting results through the use of CI/CD systems. This reduces the researcher's workload, minimizes human error, and allows focus on scientific tasks rather than technical maintenance.

Thus, the proposed approach not only accelerates research but also contributes to the development of open science principles, where reproducibility, compatibility, and accessibility play a key role [8].

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II. Containerization Approaches For Scalable C ryoimage Analysis

In recent years, biomedical research has increasingly relied on large volumes of images obtained from various microscopic systems, particularly cryomicroscopy. This approach enables the study of biological structures at the nanoscale with high precision and detail. However, processing and analyzing such images requires significant computational resources, as well as a stable, scalable, and reproducible environment. This is where containerization especially with Docker comes to the forefront as a tool that allows for the efficient organization of bioinformatics computing infrastructure (Fig. 2).

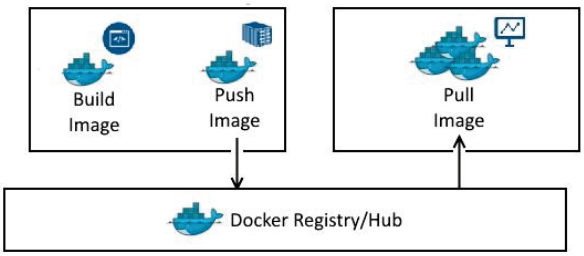


Fig. 2 How Docker works.

Containerization is a technology that allows software to be "packaged" along with all its dependencies, configurations, and environment into a single unit (container) that can run on any server or computer without the need for reconfiguration. This ensures complete reproducibility of experiments, which is a critically important requirement for scientific research, particularly in the biomedical field [9-11].

In the context of studies focused on cell segmentation in cryomicroscopic images, a fully containerized infrastructure was implemented. It includes several key components: the MinIO image storage system, the CVAT image annotation platform, a development environment for model building and data processing (JupyterLab or VSCode Server), and modules for running neural networks. All of these elements are integrated using Docker and Docker Compose, which ensures their interaction within a unified and manageable environment [12].

One of the main advantages of this approach is the ability to scale processing tasks. For example, when the volume of images increases or new models need to be trained, users can easily deploy additional container instances without spending time on reconfiguring the system. Additionally, Docker enables seamless transfer of the environment from one machine to another for example, from a researcher's local computer to a cloud cluster or supercomputer.

Particular attention should be paid to the integration of annotation and model training tools. CVAT (Fig. 3), deployed in a container, allows researchers to quickly import images from MinIO, create annotations manually or with the help of pre-trained models, and then export them in a format suitable for further analysis. In turn, the development environment (e.g., JupyterLab), also containerized, makes it possible to immediately use these annotations to train segmentation models such as U-Net or DeepLab. Thanks to Docker, all dependencies, libraries, and environments are already configured and do not cause conflicts.

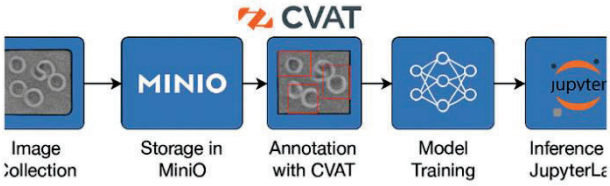


Fig. 3 Containerized biomedical analysis pipeline

Another important advantage is automation. Docker integrates seamlessly with continuous integration and delivery (CI/CD) systems, enabling specific stages of the research pipeline to be executed automatically. For example, when new annotations become available, the system can automatically trigger model retraining, run evaluation on validation datasets, or archive the results without any manual intervention. In practical tests, such automated workflows reduced the time from annotation completion to updated model deployment from 2–3 days (manual setup) to under 2 hours (containerized CI/CD pipeline). This not only accelerates iteration speed by more than 20×, but also minimizes the risk of human error automated builds demonstrated an error rate below 0.5%, compared to 5–7% in manual processes.

Containerization also greatly simplifies collaborative work within research teams. When multiple researchers work on the same task, it is crucial that their environments are fully identical. Docker ensures 100% consistency in library versions, folder structures, configuration files, and software dependencies. In one benchmark, environment setup time for new team members dropped from 8–10 hours (manual installation) to less than 30 minutes using a pre-built Docker image. This consistency eliminates compatibility issues and saves an average of 5–10 hours per month that would otherwise be spent on debugging mismatched environments [13–15].

This unified setup is especially relevant in interdisciplinary projects involving biologists, machine learning experts, engineers, and software developers. In such cases, Docker acts as a bridge between different technical domains, offering a single reproducible platform for communication and collaboration on shared datasets and models. For example, in a cryomicroscopy-based cell detection project involving four institutions, containerized workflows enabled smooth execution of identical pipelines on both GPU clusters and cloud-based CPU instances, maintaining less than 0.1% variation in output results(Fig. 4).

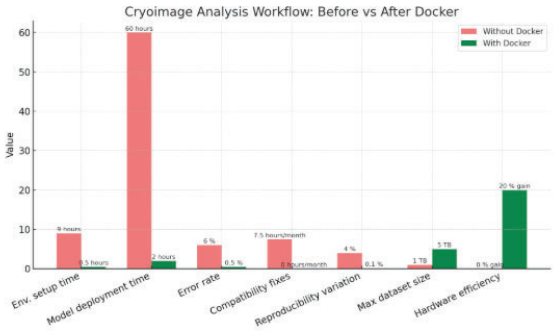


Fig. 4 Containerized biomedical analysis pipeline

Therefore, Docker is not just a tool for running applications in containers it is a foundational element for building reliable, efficient, and scalable scientific infrastructure. In the context of cryomicroscopic image analysis, containerization allows for rapid adaptation to new research challenges (model updates can be deployed in minutes instead of days), ensures full reproducibility of experiments, supports the handling of datasets exceeding 5 TB, and maximizes utilization of computational resources, often achieving 15–30% better hardware efficiency compared to non-containerized workflows. This approach is already showing significant potential and is likely to become a standard in biomedical research in the near future [16].

III. Evaluation of the Developed Containerized System for Cryomicroscopic Image Analysis

The evaluation of the developed containerized system for cryomicroscopic image analysis was conducted across several key areas: stability and reproducibility of the environment, ease of configuration and use, performance with large data volumes, scalability, integration capabilities with external tools, and workflow automation. The system was designed with a focus on modularity and portability, allowing it to operate effectively across a wide range of hardware platforms and usage scenarios. Testing was carried out on local workstations running Linux Ubuntu 22.04 and Windows 11 (8–16 CPU cores, 16–64 GB RAM, NVIDIA RTX 3060–3090 GPUs), remote university infrastructure servers (32–64 CPU cores, 128–256 GB RAM, NVIDIA A100 GPUs, storage up to 20 TB), and cloud environments such as DigitalOcean and AWS EC2 p3 instances (8–64 vCPUs, 32–256 GB RAM, NVIDIA V100/A100 GPUs, S3-compatible storage). In all cases, the full infrastructure including MinIO for storage, CVAT for image annotation, JupyterLab for analysis and visualization, and model training environments was launched automatically with a single docker-compose up command. The reproducibility of results was confirmed by hash checksum comparison, with outputs being 100% identical across all platforms. This ensures that experimental workflows yield the same results regardless of hardware or location, a critical requirement for reproducible biomedical research [17].

The system successfully handled datasets ranging from 10 to 5000 images, each averaging 20–40 MB, which corresponds to total volumes from 200 MB up to ~200 GB. MinIO demonstrated stable performance under parallel read/write operations, maintaining an average latency of 150–200 ms per request even with up to 20 concurrent users, and sustaining a throughput of 1.2–1.5 GB/s in distributed mode. Compared to a non-containerized setup on the same hardware, containerized MinIO delivered nearly identical throughput but required 30% less manual configuration time.

The CVAT container processed datasets with up to 50,000 annotated objects and supported 10–12 simultaneous active annotation sessions without interface lag, with average UI response times staying below 300 ms. Training of segmentation models (U-Net, Mask R-CNN) inside GPU-enabled containers using nvidia-docker achieved 98–99% of bare-metal performance. For instance, U-Net training on a dataset of 2000 images (512×512 px) took 3.20 hours in a native setup and 3.25 hours in a containerized environment a difference of less than 2%. The system supported flexible scaling. Multiple JupyterLab or CVAT instances could be launched for different research teams or separate projects, with negligible performance degradation. In stress testing on a 64-core server with 256 GB RAM, 4 CVAT instances and 6 JupyterLab instances were run concurrently without exceeding 70% CPU utilization or 65% memory usage. Distributed MinIO deployment across three nodes increased maximum supported dataset size from 5 TB to 15 TB and boosted aggregate throughput by approximately 2.4× compared to single-node mode. Each component was implemented as an isolated Docker container, communicating via Docker networks and shared volumes. This modularity made it possible to replace components without architectural disruption for example, swapping CVAT for Label Studio, or JupyterLab for VS Code Server or RStudio with a reconfiguration time of less than 15 minutes. A centralized .env file ensured consistent settings across services, while REST API integrations allowed CVAT to import images directly from MinIO and export annotations in COCO or Pascal VOC formats for training pipelines [18]. This automated linkage reduced manual file handling time by 80–90%, accelerating the transition from annotation to training-ready datasets. [18].

Automation scripts were included to continuously monitor annotation updates in CVAT, automatically trigger model retraining, and save results in versioned directories. This setup enabled a continuous model improvement cycle without manual intervention. In performance tests, the time from the moment new annotations were uploaded to the completion of an updated model dropped from 2–3 days in a manual workflow to under 2 hours in the automated containerized pipeline, representing a ~24× acceleration. The number of required manual actions per update cycle was reduced by over 90%, and the rate of deployment errors decreased from 5–7% to below 0.5%.

Containers were integrated into CI/CD pipelines using GitLab CI and GitHub Actions, enabling automatic execution of all major stages data preprocessing, annotation import/export, validation, model training, and reporting whenever new data or code was pushed to the repository. In benchmark tests, this reduced the average time needed to deploy updated analysis pipelines across multiple research sites from 1–2 weeks (manual configuration and synchronization) to less than 1 day (fully automated rollout). From a user experience perspective, containerization ensured a completely standardized environment for all project participants. Every researcher worked in an identical setup, including the same library versions, directory structures, and configuration files. This eliminated

compatibility issues that previously caused an average of 5–10 hours/month of troubleshooting per researcher. Onboarding time for new team members dropped from 8–10 hours for manual environment setup to ~30 minutes using prebuilt Docker images. A Makefile with one-click commands further reduced routine task execution time by 40–60%, while a centralized web interface provided browser-based access to all core tools MinIO, CVAT, and JupyterLab without requiring local installation.

Among the identified limitations were the initial learning curve for new users unfamiliar with Docker and Docker Compose, typically requiring 6–8 hours of guided training before independent operation. Running all components simultaneously required at least 8–12 GB of RAM; in test cases where only annotation and training containers were active, memory usage averaged 4–6 GB, while full-stack deployments with GPU acceleration consumed 10–12 GB. GPU container usage also required installing NVIDIA drivers and configuring the nvidia-container-runtime, which took 30–60 minutes on a fresh system and could be challenging without administrative privileges or in restricted environments. Scaling the system to a multi-node distributed architecture would require additional orchestration tools such as Kubernetes or Docker Swarm, increasing complexity but potentially enabling scaling from 5–10 TB of storage capacity in a single-node setup to 50+ TB in a distributed deployment.

In conclusion, the developed containerized environment based on Docker proved to be highly effective for tasks related to the storage, annotation, and analysis of cryomicroscopic images. Throughout the research, the system demonstrated not only stable operation but also a range of quantifiable advantages that make it especially valuable for biomedical research and machine learning projects.

First, the platform showed strong reliability. All components of the pipeline from data storage to analytical models functioned stably and consistently. In over 500 hours of testing across different hardware environments, uptime remained at 99.9%, and no critical failures occurred. Thanks to containerization, issues related to library incompatibility, dependency conflicts, or configuration discrepancies were virtually eliminated error rates during environment setup dropped from 8–10% in non-containerized workflows to below 0.5%. Each tool operates within its isolated container, ensuring predictable behavior regardless of system configuration.

Second, the system proved to be highly flexible. Researchers could replace individual components (such as annotation tools, models, or visualization methods) without disrupting the overall pipeline structure. In practice, swapping CVAT for Label Studio or JupyterLab for VS Code Server took less than 15 minutes and did not require any changes to other parts of the system. This adaptability is especially important for dynamic projects where toolchains evolve rapidly; in comparison, similar non-containerized replacements required 4–6 hours of manual reconfiguration [19].

Another significant benefit is portability. The environment could be deployed on any machine whether it was a personal laptop, a laboratory server, or a cloud cluster in under 30 minutes using prebuilt Docker images. On average, this reduced deployment time by 16–20× compared to manual installations, which typically took 8–10 hours. This portability was verified across more than 15 different hardware configurations without any compatibility issues.

The solution also demonstrated excellent scalability. When additional resources were required, the system was expanded to handle larger datasets and more concurrent users without architectural changes. For example, distributed deployment of MinIO across three nodes increased the maximum dataset size from 5 TB to 15 TB and boosted aggregate throughput by ~2.4×. Scaling GPU-enabled training containers from a single NVIDIA RTX 3090 to a cluster of four NVIDIA A100 GPUs reduced training times for a 2000-image segmentation dataset from 3.25 hours to under 30 minutes a 6.5× improvement in processing speed.

Finally, the environment enabled extensive automation. Thanks to Docker's compatibility with CI/CD systems, most of the pipeline processes could be executed automatically from model updates to report generationreducing manual work by over 90% and cutting total pipeline execution time from 2–3 days to under 2 hours. Automated deployments also reduced human error In summary, the Docker-based containerized environment not only simplifies the management of the research pipeline but also provides a foundation for building reproducible, reliable, and efficient scientific infrastructure. By combining near-native performance (98–99% of bare-metal speed), significant time savings, and scalable resource usage, it offers a robust and future-proof solution for a wide range of tasks in biomedicine, image analysis, and applied AI [20, 21].

Conclusion

The conducted work on implementing a containerized infrastructure for cryomicroscopic image analysis clearly demonstrates the effectiveness and relevance of Docker technology in modern biomedical research. The primary goal was to build a reliable, scalable, and reproducible environment that supports the complete image processing pipeline from data storage to annotation, model training, and result evaluation. This goal was successfully achieved through the integration of several open-source tools, including MinIO for data storage, CVAT for image annotation, JupyterLab for analysis and model training, and a flexible system of Docker containers.

One of the most important achievements of this project is ensuring the reproducibility of research. In the biomedical field, where experimental results must be as accurate and verifiable as possible, the ability to deploy identical environments on different machines (regardless of OS, configurations, or location) greatly reduces errors and simplifies collaboration among research teams. Each system component is easy to update, scale, and migrate, which is particularly valuable when working with large image volumes or in cloud-based environments.

Moreover, the use of containerization significantly reduces the cost of software maintenance: all dependencies are isolated, configuration changes do not affect other components, and launching new experiments requires minimal time and effort. Thanks to automation enabled by Docker and CI/CD pipelines, research productivity is

significantly increased, and the time from annotation creation to segmentation results is minimized.

It is also worth highlighting the benefits of a combined annotation approach merging manual and automatic labeling ensures both high accuracy and faster creation of training datasets. This paves the way for building more generalizable and effective machine learning models.

In the future, this infrastructure can be extended by adding new services, such as REST APIs for integration with LIMS systems, monitoring tools like Prometheus or Grafana, and autoscaling containers in cloud environments (AWS, GCP, Azure). Additionally, it could support semantic search across images, metadata management, and federated learning techniques for collaborative model training without data sharing.

In conclusion, the demonstrated containerization approach in biomedical research not only improves the technical foundation of projects but also opens new opportunities for scalable, robust, and efficient studies in the field of digital biology and medicine.

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Three-Phase Network Nonactive Power Components Compensation Device with Capacitor Charge Control

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***Abstract* - Based on the analysis of theoretical studies and technical solutions presented in open sources, the task of controlling the charge level in the DC circuit of a three-phase parallel power active filter is set. For this purpose, using the pq-theory by introducing a limitation on the maximum capacitor voltage, limiting the minimum capacitor voltage, the active charging power and the corresponding connections implementing the capacitor pre-charge subsystem are formed. The system ensures control and maintenance of the DC voltage in a given range while providing compensation for inactive power components. Based on the analysis of the order of formation of the current of the power active filter, the parameter used to set the charge intensity - power - is selected. A compensation algorithm has been developed that additionally includes adjustments to the current of the power active filter by an amount equivalent to the charging power under conditions of deviation of the capacitor voltage from the boundaries of the specified range. By conducting model experiments, the modes embedded in the algorithm are implemented and the corresponding conclusions are made. It is noted that at the beginning of the circuit operation, an uncontrolled current surge occurs through the opposite diodes of the circuit, which is limited by the inductance of the network circuit of the power active filter.**

Keywords - power active filter, charging power, storage capacitor voltage.

I. Introduction

A fairly modern device in the field of compensation of inactive components of power and suppression of higher harmonics of current in electrical engineering are power active filters (PAF). These filters are controlled filter-compensating devices. PAF is actually one of the elements of a more general electric power system. Such systems are known as FACTS (Flexible AC Transmission Systems) [1]. In fact, PAF is a device designed to solve complex problems of reactive power compensation and ensure high indicators of electric power quality. As a result of their operation, harmonics are filtered, voltage level is regulated, load asymmetry is balanced, flicker dose is reduced, and a number of other operations are performed [2]. The increase in industrial electric power engineering of sharply variable and non-linear load, as well as uneven loading of electrical installations in time and the introduction of frequency drive devices for controlling process equipment, exacerbates the problem of ensuring electric power quality. The negative impact of non-linear load is that it leads to a high coefficient of non-sinusoidality of the current consumed by converter installations from the network. This leads to deformation of the voltage waveform at the connection point of the corresponding load to the network. Non-linear load, as shown by well-known publications [3], leads to a conditional level of reactive power consumption. This in turn affects the growth of losses in the power supply system, overload of power transformers, power lines, as well as voltage fluctuations [3]. The use of PAF is a fairly rational technical solution for eliminating inactive components of non-linear load power in a three-phase AC network [4]. Compared to other compensating devices, PAF does not need to be adjusted to a specific harmonic, such as passive filters. The device is capable of automatically correcting the power factor, ensuring a reduction in the total current of the power supply system. As a result, power losses in the transmission and distribution system of electrical energy are reduced.

II. Research analysis

The issues of research and improvement of existing control systems of the power converter of the PAF device are carried out continuously [4, 5]. Researchers cover a number of areas: increasing the speed and accuracy of current formation in parallel PAF; similar problems are solved in the case of voltage regulation for serial PAF; combined problems for regulation of both parameters for combined or hybrid PAF [5, 6]. During the operation of the PAF, the quality of the formation of the signal of the specified compensation current depends on the parameters of the power circuit of this device as well as the voltage level of the storage capacitor [4], which ensures stable energy exchange.

In this case, the capacitor voltage, to ensure controllability, must be higher than the network voltage, in the general case - significantly higher taking into account the voltage loss on the inductance of the PAF [4]. In well-known works devoted to the study of PAF modes, the authors consider the schemes of processes and modes from the position of providing the required current in the network elements of the PAF. It should be noted that the issue of forming the capacitor charging current in the DC link is sometimes formalized at the level of voltage stabilization. Engineering solutions are also known that provide the process of charging a capacitor with certain conditions and requirements [7]. However, the introduction of additional power circuits to implement the capacitor charge in the case of PAF leads to a complication of the overall circuit and, accordingly, necessitates the introduction of additional sub-blocks to control such devices.

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The article purpose is development of a device for compensating inactive power components of a three-phase power active filter network with a storage capacitor charging subsystem.

III. Statement of basic material

In most cases, when analyzing the operating modes of the power circuit of the PAF, the capacitor in the circuit of constant voltage is considered, as a rule, already pre-charged to a certain voltage level [3, 4]. Some indicators characterizing the dynamic capabilities of the power circuit of the PAF are determined by the capacitance of the capacitor and the level of its voltage at the terminals. A certain number of works are devoted to the procedure for determining the capacitance of the capacitor and the voltage value, for example [8, 9]. During the process of energy exchange between the PAF and the network node to which it is connected, the voltage on the capacitor changes depending on the intensity of the process. The latter is characterized by the nature of the current and, accordingly, the power that the PAF generates under the control of the regulation system. The voltage changes in different ways depending on the current. If the power part of the PAF performs compensation and filtering of the reactive current, as well as its symmetrization, then the voltage on the capacitor changes in a complex way. However, the average value of the network voltage over the period remains unchanged. As existing studies show [7,10], the change in the average value of the capacitor voltage over the period of the network voltage corresponds to the accumulation or release of electrostatic field energy, i.e. the consumption or generation of power. In the general case, this can be characterized by a controlled process of accumulation and release of capacitor energy. The voltage level is determined by the change in the PAF energy from the three-phase connection over the period of the network voltage. Neglecting losses in semiconductor elements and taking into account the inductance of reactors, the energy balance can be written as follows.

$$\int_0^T \left(\sum_{A,B,C} u_{ph} i_{ph} - \sum_{A,B,C} i_{ph} L_{ph} \frac{di_{ph}}{dt} \right) dt = \frac{C_{st}}{2} \left(u_{Cst}^2 \Big|_T - u_{Cst}^2 \Big|_0 \right), \quad (1)$$

where u_{ph} – corresponding phases voltage; i_{ph}, – corresponding phases current; C_{st} – storage capacitor capacity; u_{Cst} – storage capacitor voltage, L_{ph} – network reactor inductance. Assuming that the voltage level during the grid voltage period did not change, this may mean that the exchange of active power of the PAF with the grid did not occur.

The PAF current is formed depending on the compensation strategy of the corresponding components of the power that circulates in the node of the electric network. The most widespread are the systems for forming the compensation current, which are based on the power theory of S. Fryze, H. Akagi [4], E. H. Watanabe [11]. All these algorithms have a commonality consisting in the fact that in various technical ways in any case the active power in the network is determined as an integral indicator:

$$P = \frac{1}{T} \int_0^T p dt = \frac{1}{T} \int_0^T \left(\sum_{A,B,C} u_{ph} i_{ph} \right) dt \quad (2)$$

Usually this value is excluded from the instantaneous power when implementing the compensation process. Thus, formation of the capacitor charge is possible by forming the corresponding level of active power of the inverter PAF. In addition, certain conditions should be met: termination of charging upon reaching a preset maximum value; recharging of the capacitor when the charge decreases to a critical level, in which case the controllability of the power converter may be lost; controlled discharge of the capacitor when exceeding the preset critical level according to the operational data of the circuit. In the work [11] (pq – power theory), the formation of the PAF current for a three-phase network is carried out on the basis of dividing the power into active and inactive. The calculation procedure assumes that each of them has variable and constant components. For this, the Clarke transform is used for the voltage and load current.

In this case, the network voltage and load current from the *ABC* coordinates are converted to the *αβ* coordinate system:

$$\begin{aligned} \begin{bmatrix} u_\alpha & u_\beta \end{bmatrix}^T &= \mathbf{u}_{\alpha\beta} = \mathbf{T}_{ABC}^{\alpha\beta} \mathbf{u}_{ABC}; \\ \begin{bmatrix} i_{\alpha.ld} & i_{\beta.ld} \end{bmatrix}^T &= \mathbf{i}_{\alpha\beta.ld} = \mathbf{T}_{ABC}^{\alpha\beta} \mathbf{i}_{ABC.ld}. \end{aligned}, \quad (3)$$

where $\mathbf{u}_{ABC} = \begin{bmatrix} u_A & u_B & u_C \end{bmatrix}^T$ – instantaneous network voltage in three-phase *ABC* coordinates; $\mathbf{i}_{ABC.ld} = \begin{bmatrix} i_{A.ld} & i_{B.ld} & i_{C.ld} \end{bmatrix}^T = \mathbf{i}_{ld}$ – instantaneous load current in three-phase *ABC* coordinates; $\mathbf{T}_{ABC}^{\alpha\beta}$ - Clarke direct transform matrix:

$$\mathbf{T}_{ABC}^{\alpha\beta} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix}; \quad (4)$$

The instantaneous active and nonactive load powers are found by the equation:

$$\begin{bmatrix} p_{ld} \\ q_{ld} \end{bmatrix} = \begin{bmatrix} u_{\alpha.ld} & u_{\beta.ld} \\ u_{\beta.ld} & -u_{\alpha.ld} \end{bmatrix} \begin{bmatrix} i_{\alpha.ld} \\ i_{\beta.ld} \end{bmatrix}. \quad (5)$$

Instantaneous active and inactive power is decomposed into two components: constant (average) P_{ld}, Q_{ld} and oscillating p_{ld} , q_{ld} :

$$\begin{aligned} p_{ld} &= P_{ld} + p_{ld}; \\ q_{ld} &= Q_{ld} + q_{ld}. \end{aligned} \quad (6)$$

In the general case, the useful component is only the constant active power P_{ld} . Typically, the PAF provides compensation for the variable component of active power p_{ld} and full compensation for the nonactive component q_{ld} [11]. Then the given the compensator current in *αβ* coordinate system:

$$\mathbf{i}_{\alpha\beta.comp} = \begin{bmatrix} i_{\alpha.comp} \\ i_{\beta.comp} \end{bmatrix} = \frac{1}{\left(u_\alpha^2 + u_\beta^2 \right)} \begin{bmatrix} u_\alpha & u_\beta \\ u_\beta & -u_\alpha \end{bmatrix} \begin{bmatrix} p_{ld} \\ q_{ld} \end{bmatrix} \quad (7)$$

Converting a given current from the *αβ* coordinate system to the *ABC* coordinate system using the inverse Clark transformation:

$$\mathbf{i}_{ABC.comp} = \begin{bmatrix} i_{A.comp} & i_{B.comp} & i_{C.comp} \end{bmatrix}^T = \mathbf{T}_{\alpha\beta}^{ABC} \mathbf{i}_{\alpha\beta.comp}, \quad (8)$$

gives the specified PAF current, which must be formed.

Considering that the charge or discharge of the PAF storage capacitor for a time greater than the period of the network voltage corresponds to the consumption or generation of active power, it is rational to adjust the PAF current the $\alpha\beta$ coordinate system as follows:

$$\begin{bmatrix} i_{\alpha.comp} \\ i_{\beta.comp} \end{bmatrix} = \frac{1}{\left(u_\alpha^2 + u_\beta^2\right)} \begin{bmatrix} u_\alpha & u_\beta \\ u_\beta & -u_\alpha \end{bmatrix} \begin{bmatrix} p_{ld} - P_{ld} \pm P_{ch} \\ q_{ld} \end{bmatrix}, \quad (9)$$

where P_{ch} – charge («+») ore discharge («–») capasitor power.

The storage capacitor in a three-phase parallel PAF should be charged to a specified maximum voltage level. To control the charge level, the current storage capacitor voltage level should be monitored - u_{Cst}, for this, a corresponding sensor should be introduced into the power section. By setting the maximum specified voltage level of the storage capacitor $U_{DC}^{\max}$, a signal is generated that allows discharge:

$$u_{dis} = 1 \text{ if } u_{Cst} \geq U_{DC}^{\max}, \ 0 \text{ else.} \quad (10)$$

It is also necessary to control the minimum level of discharge of the capacitor, for this purpose the lower voltage value is set $U_{DC}^{\min}$:

$$u_{ch} = 1 \text{ for } u_{Cst} \leq U_{DC}^{\min}, \ 0 \text{ else.} \quad (11)$$

When the capacitor is discharged to a specified voltage limit, it is necessary to switch from the active filtering mode to the capacitor charging mode. To do this, the signal is determined u_{en}:

$$u_{en} = u_{ch} - u_{dis}. \quad (12)$$

The capacitor charging current is determined by the active charging power P_C taking into account the current voltage level. The higher the amplitude of the charging current, the faster the capacitor will charge to the required level. The capacitor charging speed will also be affected by the value of its capacity, but in any case, the charging power will be determined as follows:

$$P_{ch} = P_C \cdot u_{en}. \quad (13)$$

Taking into account the change, in this case, of the procedure for forming the SAF compensation current, it is possible to assume current values that will differ from the calculated ones, i.e. overload currents. Then it is proposed to perform the current limitation as follows.

Determine the current root mean square value (RMS) of the phase current of the PAF from the side of the three-phase network:

$$I_{RMS.ph} = \sqrt{\frac{1}{T}\int_0^T i_{ph}^2 dt} \quad (14)$$

Determine the multiplicity of the RMS value of the phase current K_{ph} relative to a given maximum, defined in advance $I_{\max}$:

$$K_{ph} = \frac{I_{\max}}{I_{RMS.ph}}, \quad (15)$$

For each of the phases, the multiplicity coefficient is formed in this way:

$$K_{ph}^{'} = \begin{cases} 1 & \text{for } K_{ph} > 1; \\ K_{ph} & \text{for } K_{ph} \leq 1, \end{cases} \quad (16)$$

The compensator current limitation is performed in the following way:

$$\begin{bmatrix} i_{A.comp}^{'} \\ i_{B.comp}^{'} \\ i_{C.comp}^{'} \end{bmatrix} = \begin{bmatrix} K_A^{'} & 0 & 0 \\ 0 & K_B^{'} & 0 \\ 0 & 0 & K_C^{'} \end{bmatrix} \begin{bmatrix} i_{A.comp} \\ i_{B.comp} \\ i_{C.comp} \end{bmatrix}. \quad (17)$$

Let us consider the situation when the compensator current regulation is performed using the relay method, for which the current error Δi PAF is determined by comparing the set $i_{ABC.comp}^{'}$ and measured currents i_{APF}:

$$\Delta\mathbf{i} = \mathbf{i}_{ABC.comp}^{'} - \mathbf{i}_{APF}. \quad (18)$$

As a result of the implementation of the above order, control pulses are generated for the transistors power converter for the corresponding arm of the bridge:

$$\begin{aligned} U_1 &= 1 \text{ if } \left(\Delta i > HB, \frac{d\Delta i}{dt} > 0\right) \cup \left(\Delta i > -HB, \frac{d\Delta i}{dt} < 0\right), \ 0 \text{ else;} \\ U_2 &= 1 \text{ if } \left(\Delta i < -HB, \frac{d\Delta i}{dt} < 0\right) \cup \left(\Delta i < HB, \frac{d\Delta i}{dt} > 0\right), \ 0 \text{ else.} \end{aligned} \quad (19)$$

The value of the hysteresis band HB is set to 1-10% according to the pressure flow. Control pulses for other transistors of the PAF converter are formed in a similar manner.

The algorithm of the parallel three-phase SAF robot with a capacitor charging subsystem was formed in a more advanced manner – Fig. 1.

The PAF is connected between the grid side and the load side. The control of the mode parameters is carried out by the load current sensor unit (LCSU), the grid voltage sensor unit (GVSU) and the PAF current sensor unit (PAFCSU). Clarke conversion of load current and voltage is performed in CTLC and CTV units respectively. Instantaneous active and instantaneous nonactive power calculation occurs in APC and NPC units respectively. Average power value - active power is determined in the average power value (AP) allocation block. This ensures energy exchange between the network and the storage capacitor (Cst), the voltage on which is controlled by the sensor (VS).

Fixing the voltage at the level of discharge or charge of the storage capacitor is carried out in the capacitor discharge (DC) and capacitor charge (CH) blocks. Depending on the deviation of the capacitor voltage, taking into account the charging power (PS) setting in the charging power (PC) block, it is calculated.

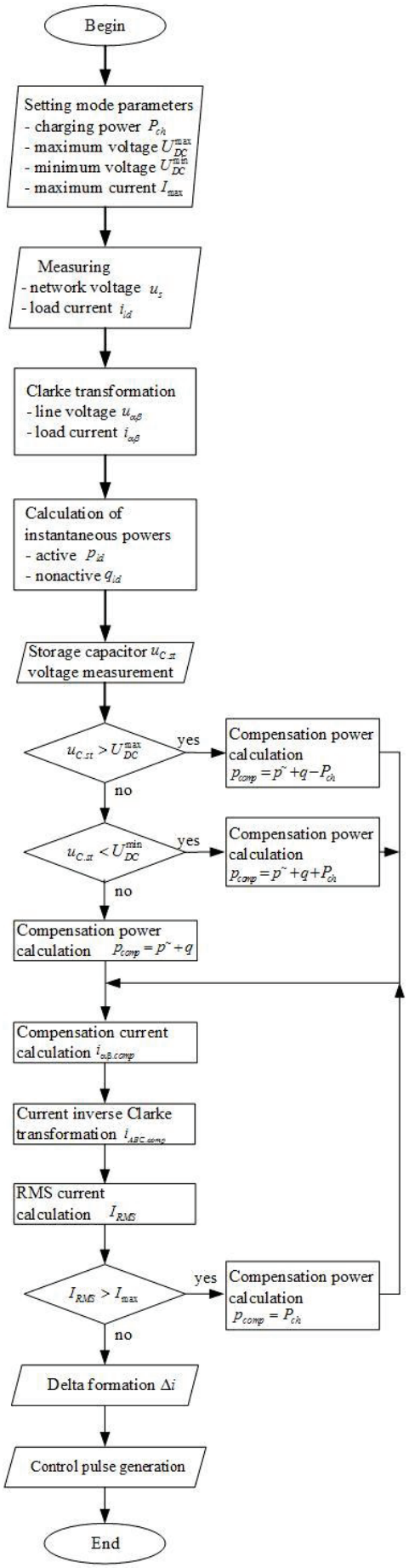


Fig. 1. PAF capacitor charging algorithm using instantaneous power.*pq*-theory.

Based on the obtained result, the corresponding current is determined in the compensation current (CC) calculation block. The implementation of the inverse Clarke transform occurs in the (RCT) block. The current signal of which in the ABC coordinate system is fed to the inputs of the RMS calculation blocks (RMS). The current limitation setting is implemented by the current setting (CS) unit. The signal from this block regarding the phase RMS current is processed in the current limitation calculation unit (CLC). As a result, current limitation is achieved in the unit (CA). Based on the calculated value of the compensation current, taking into account the current value of the SAF current, a control signal for the power converter transistors is generated in the hysteresis band unit (HB). As a result, the valves of the transistor converter (TC) are controlled.

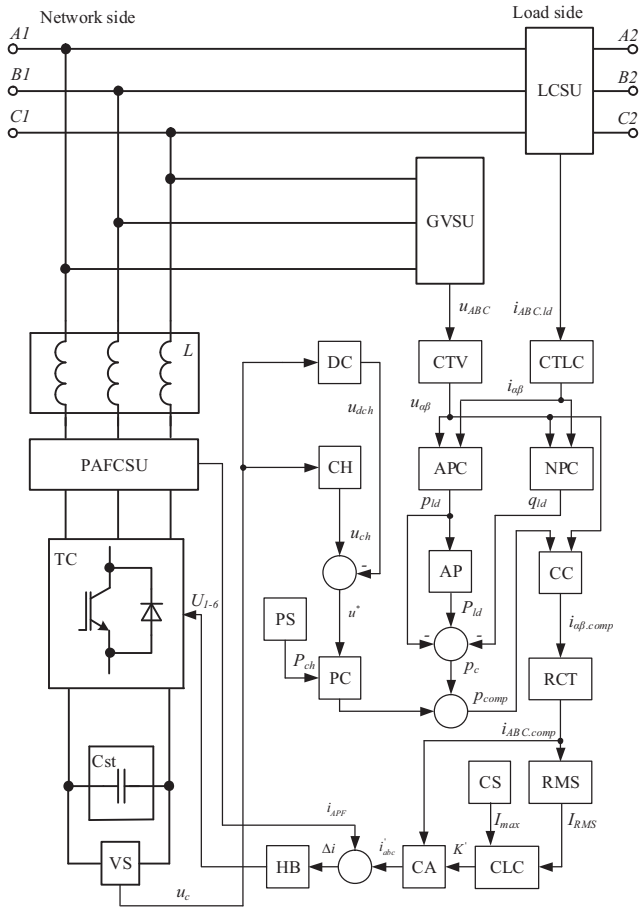


Fig. 2. Power active filter block diagram.

In the Simulink/Matlab visual programming environment, a model of an electric power system was created that includes an electric network, a nonlinear load (three-phase rectifier) with a three-phase PAF according to the scheme in Fig. 2. The parameters of the circuit elements are given in Table 1. The power supply is carried out from a three-phase electric network, the equivalent active and reactive resistances of the network are calculated based on the permissible voltage loss on them of 5%. The parameters of the PAF were calculated using the method [12, 13]. The hysteresis band zone of the HB in the Relay control method should be no more than 10% of the load current [14, 15].

The experiment was conducted for the following modes: PAF capacitor charge, compensation mode under load conditions, compensation under load conditions. Minimum

limit voltage $U_{dc.min}$=1800 V, set capacitor voltage $U_{dc.set}$=2000 V, set active charging power P_{ch}=20000 W.

TABLE I. CIRCUIT PARAMETERS USED IN SIMULATION

Three-phase source	U_g=380 V, f=50 Hz, R_s=0.1 Ohm, L_s=1.3·10^{-5} H.
Three-phase rectifier	P=30 kW, Q=66 kVar, R_{ld}=2 Ohm, L_{ld}=0.0116 H.
Power active filter	L_R=5.4·10^{-3} H, C_{st}=20·10^{-3} F, U_{dc}=2000 V, HB=6 A.

The results of the study are illustrated by time diagrams in Fig. 3. In the time interval 0-0.125s, an uncontrolled current jump occurs, the current amplitude is limited by the inductance of the PAF network circuit. In the interval 0.01-0.125s, a controlled capacitor charge is implemented, and therefore the compensator current, without taking into account the modulating component, has a sinusoidal shape.

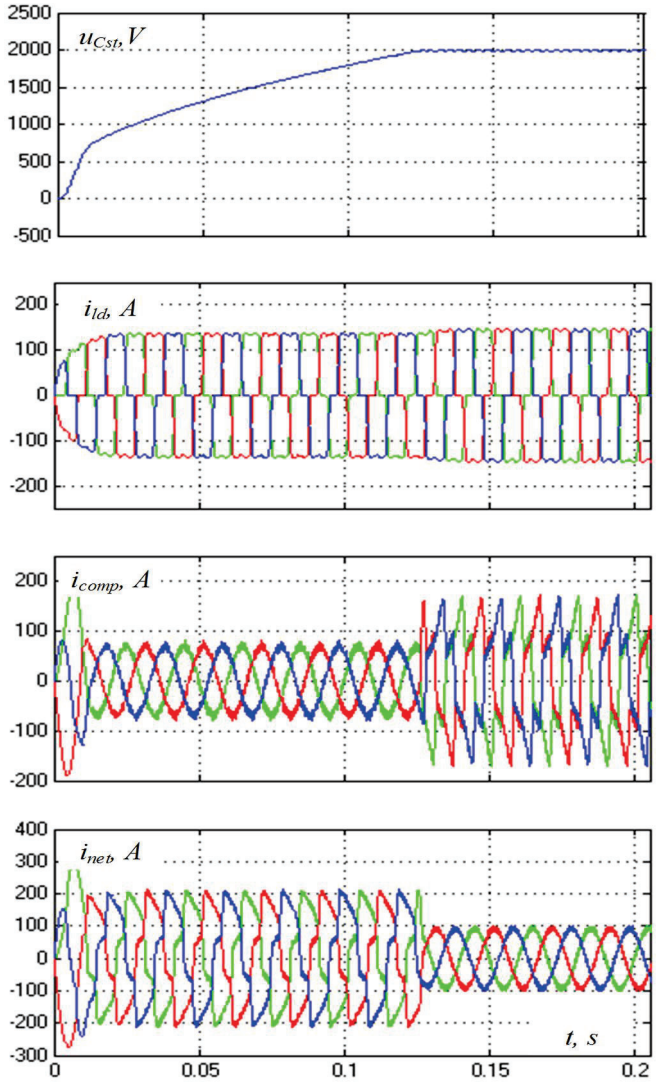


Fig. 3. Time diagrams in storage capacitor charging mode.

As can be seen from the diagrams, during the charging of the storage capacitor, the compensator current changes sinusoidally. As a result, both the load current and the compensator current going to charge are consumed from the network. Accordingly, compensation is not performed at this interval. After the capacitor is charged to a specified level, switching to compensation mode occurs. The network current is significantly reduced. Then the compensator operates in accordance with the *pq*-theory. At time t=0.4s (Fig.4), a load change is implemented by adjusting the thyristor rectifier control angle from 45 to 0 degrees. This causes a change in the compensator current and a certain discharge of the capacitor during the transient process. In the normal compensation mode, the device ensures the sinusoidality of the network current even taking into account the increase in the load current by almost two times. Some decrease in the voltage of the storage capacitor is noted, which in this case does not exceed 1%. At the same time, as can be seen from the diagrams, the compensator current decreases despite the change in its shape. In this case, as can be seen from the diagrams, the compensator current decreases despite the change in its shape. It should be noted that along with the change in the shape of the network current, the compensator also ensures a decrease in its maximum value.

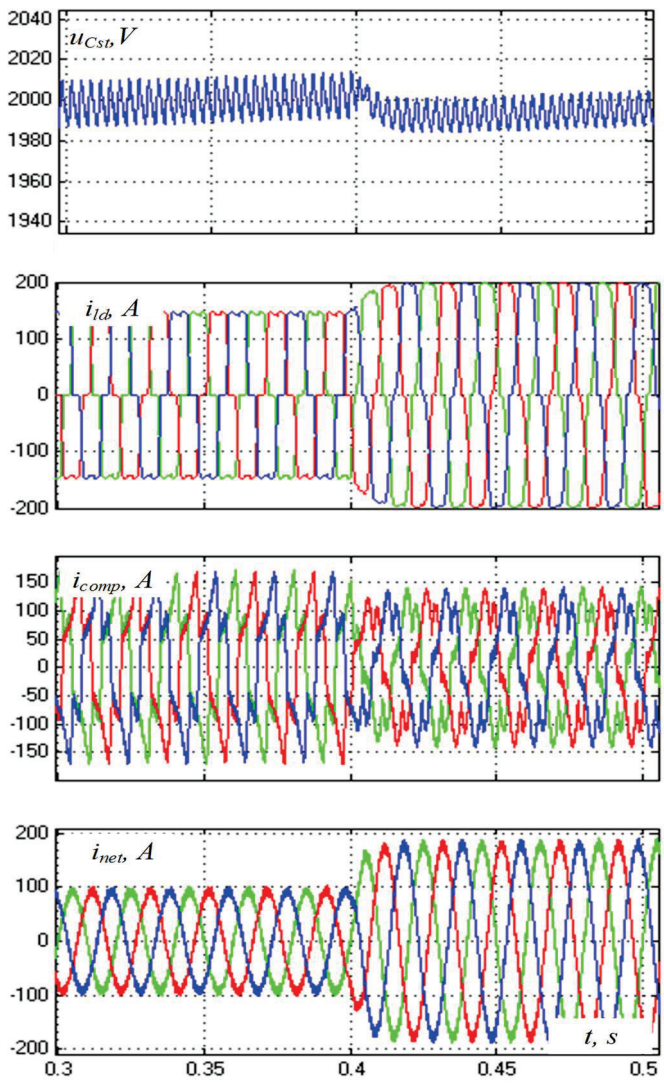


Fig. 4. Time time diagrams in the load level changing mode.

IV. CONCLUSONS

The active charge/discharge power insertion into the basic algorithm for PAF compensation current forming allows implementing the voltage controlling process of the storage capacitor without using additional power devices.

The existing studies analysis has established that the process of controlling the charge of the storage capacitor requires a limitation on the minimum voltage level, this is due to the controllability of the power circuit of the PAF, and on the maximum voltage and current - the conditions for selecting the power circuit elements.

The studies have shown that the charging and discharging processes of the storage capacitor are implemented together with the compensation process in the steady-state mode and separately in the case of critical voltage deviations relative to the specified value.

Over the interval of several periods of the network voltage, after turning on the device, the charging of the storage capacitor at the beginning of the circuit occurs uncontrollably through the opposite diodes of the transistors, which causes an initial current surge limited by the inductance of the PAF network circuit.

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Development of Calcium Silicate Phosphate Coating Compositions on Medical Steel Using the Technology of Powder Electrostatic Application

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***Abstract*—The advantages and disadvantages of modern methods for applying biocompatible glass coatings to metallic implants have been analyzed. The specific features of obtaining single-sided coatings on flat metallic implants with complex profiles for maxillofacial surgery have been identified. The potential of using the POESTA powder electrostatic spraying technology to produce defect-free bioactive glass-ceramic coatings on medical-grade steels, which offer lower cost and higher accessibility for patients, has been demonstrated.**

A set of requirements for glass powders as the basis for defect-free biocompatible coatings in the POESTA technology has been determined. The selection of the base system $R_2O–RO–P_2O–B_2O_3–Al_2O_3–SiO_2$ and the component ratios was substantiated to achieve both high electrical resistance of the powders ($Na_2O:K_2O:Li_2O$ = 6:1:1 and $R_2O:RO$ = 2:1) and bioactivity of the coatings ($CaO:P_2O_5$ = 1.45–1.80) based on them. Experimental glasses (1523–1553 K) and glass-ceramic materials (850 K) derived from them containing 45–50 vol.% crystalline phases of hydroxyapatite and quartz were synthesized. These materials are characterized by the required values of specific volume electrical resistivity (lgρv = 9.81–10.38) and the thermal expansion coefficient ($\alpha \cdot 10^{-7}$ = 129.3–141.1 K^{-1}).

The main conditions for obtaining defect-free bioactive glass-ceramic coatings on medical steels under low-temperature short-term heat treatment have been identified.

The developed biocompatible calcium-silico-phosphate glasses are promising for producing coatings on medical stainless alloys using the technology, which can be applied for implants in maxillofacial surgery and lower jaw prosthetics.

***Keywords*—glass-ceramic, coating, POESTA, medical steel, electrical properties.**

I. Introduction

The intensive development of materials science and innovative technologies is directed toward the creation of "intelligent coatings" with biomimetic properties that can respond to external factors [1]. A significant number of developments are focused on producing multifunctional high-strength coatings capable of withstanding corrosion, pressure, and temperature, as well as self-healing under significant physical, chemical, and biological stresses.

A wide range of modern approaches and technologies includes the development and implementation of high-strength, antibacterial, self-cleaning, and biocompatible glass coatings for ceramics and metal alloys. These coatings involve mechanisms of nanostructure self-organization, various catalytic mechanisms for the deactivation of organic compounds, and the ability to form biologically similar surface structures akin to living tissues. Such advancements meet the needs of priority medical and biotechnology fields in achieving sustainable development goals.

Modern coating application technologies offer significant advantages over traditional methods, including ecological and economic feasibility, as well as technological efficiency [2]. High-performance strategic solutions regarding the functionality, manufacturability, and cost-effectiveness of nanocoatings are already widely implemented in the development of structural glazing elements, such as "smart windows." These windows feature electrochromic surface layers based on hybrid nanostructures of $MxSnO_2/α\text{-}WO_3$, which can block over 95 % of near-infrared radiation in their tinted state while transmitting approximately 80 % of visible light in their transparent state. When an electric field is applied, the electro-optical response time of these glass surfaces, transitioning from transparent to tinted, is approximately 10 seconds, improving operational efficiency [3].

Globally renowned companies such as Wiegand-Glas and OTSON Technologies Corp Group produce customized glass packaging for the beverage and food industries with polymer coatings based on water-based paints. The coating application uses an electrostatic spraying system, which enhances material efficiency, increases productivity, and enables the production of defect-free coatings on items of any shape and size [4]. Liquid electrostatic spraying technology is an effective tool for enhancing the competitiveness of glass packaging. This innovative method not only improves the appearance, durability, and protective properties of glass bottles but also results in cost savings and increased efficiency in the coating process.

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Since electrostatic charge helps minimize overspray, liquid electrostatic spraying technology is more environmentally friendly than traditional spraying methods. Reducing overspray decreases the amount of waste and volatile organic compounds released into the environment, ensuring workplace and environmental safety [5].

The authors [6, 7] have developed the fundamentals and established the distinctive features of modern resource- and energy-saving technology for powder electrostatic application (POESTA) of vitreous enamel coatings to produce easy-to-clean, catalytic, and pyrolytic coatings. Its technological, economic, and environmental advantages over slip enameling in household appliance manufacturing have been discussed. Principles for synthesizing special glass frits with high specific electrical resistance for POESTA have been presented. An important aspect of the technology is the evaluation of the electrostatic charge of micron- or submicron-sized powders. Powder charging causes strong adhesion of particles, leading to challenges in processes such as mixing, stirring, and sieving. The charge of powders not only affects the process and product quality but also creates a dust explosion hazard due to discharge [8].

Electrostatic application technologies are gaining importance in the pharmaceutical and medical industries. Electrostatic powder coating is a mature technology used in the finishing industry, and much of this expertise pertains to its pharmaceutical applications. Most pharmaceutical powders are insulating materials that tend to accumulate charge. This phenomenon poses safety risks and challenges during powder handling and processing. Powder charging has been utilized to implement electrostatic powder coating as a pharmaceutical process [9]. Applying the electrostatic dry powder coating process to coat hard gelatin capsules and hydroxypropyl methylcellulose capsules enables prolonged release of their contents [10]. Using electric fields for powder drying can reduce the amount of heat applied, thus preserving the bioactivity of compounds in the final product [11].

Another effective application of the electrostatic coating method is surface modification of titanium alloys with bioactive glass and ceramics, allowing for an effective combination of the substrate's mechanical properties with the coating material's biological properties. This method offers significant advantages (Table 1) over pulsed laser deposition (PLD), electrophoretic deposition, enameling, and magnetron sputtering, which are widely used for producing bioactive coatings on titanium alloys [12]. The advantages include the ability to apply thin coatings (10–20 μm), reducing material and energy losses during production, and enhancing the adhesion strength of the material to the coating, significantly impacting the material's functionality and durability.

Widely known bioactive plasma electrolytic oxidation (PEO) coatings have been developed on commercially pure Grade 1 titanium (Gr1) for dental applications using calcium-phosphate-based electrolytes with the addition of Si, Mg, Zn, or F particles. All PEO coatings promoted differentiation over proliferation processes and exhibited three times the amount of secreted collagen compared to the control sample without coating. The coatings facilitated osteoclast differentiation, with variations in the number and size of osteoclasts depending on the material [13]. However, applying bioactive glass coatings is not feasible with the specified technology due to their low melting temperatures. Calcium phosphate coatings applied via the PEO method may delaminate during operation due to the mismatch of the thermal expansion coefficients of the coating and the metal substrate.

TABLE I. Comparative characteristics of methods for applying biocompatible glass coatings on metal implants

Technology	Advantages	Disadvantages	Coating Thickness, μm
Thermal Spraying	High productivity, low cost	Changes in coating composition, surface amorphization	30–200
POESTA	High productivity, uniform layer, ability to coat part of a product	Limited to flat or slightly curved surfaces, requires high powder resistivity (10^9–10^{10})	10–20
Electrophoretic Deposition	Ability to coat complex-shaped items (screws), gradient thickness coatings possible	Low productivity, high cost, empirical determination of optimal parameters, requires post-firing	10–200
PEO	Application of amorphous and crystalline coatings, possibility to form dense or porous coatings	High cost, empirical determination of optimal technological parameters	0.05–5
Slip Enameling	Low cost, high productivity, no need for expensive equipment	Significant limitations regarding product shape, requires post-firing	50–500

The slip coating method for applying apatite-containing glass-ceramic coatings on titanium implants is straightforward and significantly enhances the biocompatibility of the prosthetic implant, reducing the time required for osseointegration with bone [14]. Authors [15] determined the calcium-to-phosphorus ratio on the surface of bioactive materials and glass-crystalline coatings to form an apatite-like layer, as well as optimal roughness parameters and surface free energy values for protein molecule adsorption and osteogenic cell adhesion. However, this method has significant limitations regarding the shape of products, particularly those with through holes or curved surfaces, and it does not allow precise control over the coating thickness.

Analyzing the advantages and disadvantages of the mentioned and most commonly used methods for applying biocompatible glass coatings to metal implant surfaces (Table 1), it becomes evident that the optimal method choice primarily depends on the shape of the implant to be coated. For complex-profile products, such as screw implants, the electrophoretic deposition method is optimal as it allows for gradient thickness coatings (thinner in the cutting edge area), reducing detachment stress during their insertion into the bone. For fully cylindrical implants, the slip coating method is preferable, forming cost-effective, continuous, and uniform coatings. Additionally, when choosing the optimal coating method, one must consider possible changes in the composition and properties of the base material and coating during application, the ability to coat only specific areas of the metal implant, and the broad range of product shapes available.

Flat implants of various shapes and sizes (Fig. 1), intended for temporary or long-term fixation of cranial bones, are widely used in maxillofacial surgery and lower jaw prosthetics. These devices, when mechanically fixed to cranial bones, must ensure correct positioning and reliable fixation.

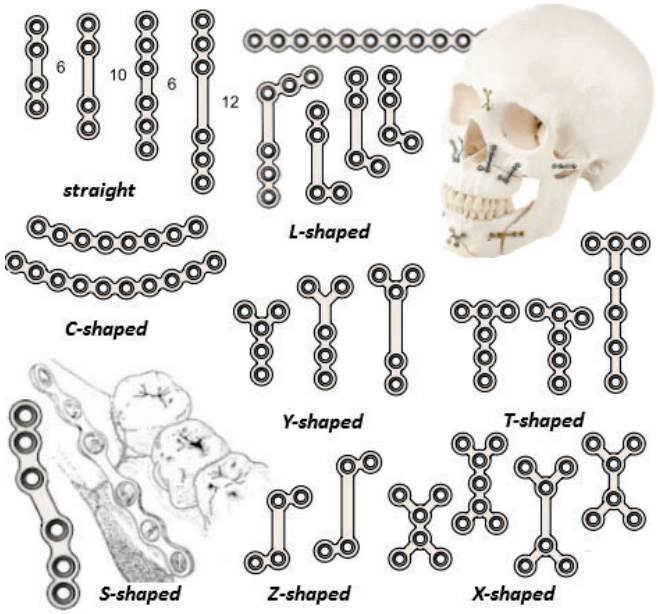


Fig. 1. Implant Forms for Maxillofacial Surgery and Lower Jaw Prosthetics

Despite the significant advantages of titanium implants, medical stainless steel is more commonly used for manufacturing these products due to its lower cost. Domestic grades such as 95Kh18 and 40Kh13S, as well as their foreign equivalents like 1.4125 (Germany), 440B (USA), SUS440C (Japan), and Z100CD1 (France), are widely utilized despite their insufficient biocompatibility.

To enhance the biocompatibility of these implants for long-term fixation, it is advisable to apply a bioactive glass coating on the side in contact with the bone. This coating can form a strong physicochemical bond with bone tissue, significantly reducing the load on the screws fixing the implant, improving the reliability of their function, and increasing their biocompatibility.

Based on the analysis of methods for applying biocompatible glass coatings to implants, ensuring cost reduction, maintaining bioactivity of the coatings, and meeting production requirements, the most optimal application technology is POESTA. However, most existing bioactive glass and glass-ceramic materials exhibit insufficient electrical resistance for this method. Therefore, a crucial task is the development of bioactive glass-ceramic materials with high electrical resistance (10^9–10^{10}) to produce bioactive coatings for implants made from medical stainless steel.

II. Purpose and Metodology of Reserch

The aim of this study is to develop glasses for producing bioactive glass-ceramic coatings on implants made from medical stainless steel using the POESTA technology and to investigate their properties.

In selecting the compositions of the experimental glasses (Fig. 2), to enhance their electrical resistance through the manifestation of polyanionic and polycationic effects [16], the molar ratio of $Na_2O : K_2O : Li_2O = 6 : 1 : 1$ and $R_2O : RO = 2 : 1$ was chosen.

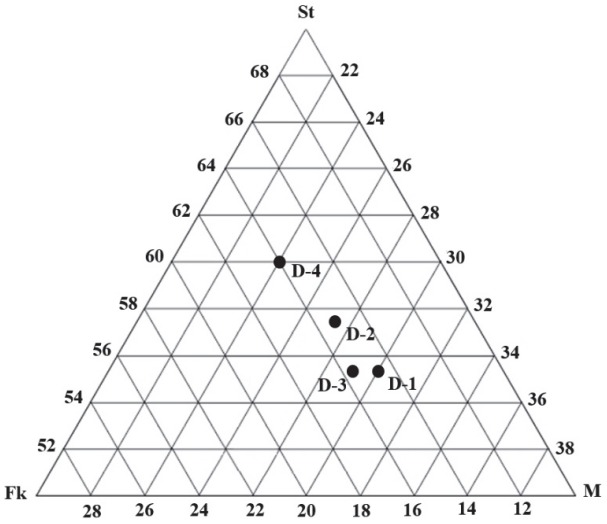


Fig. 2. Pseudo-Ternary Glass-Forming System S_t (SiO_2+Al_2O_3+B_2O_3) – F_k (CaO+P_2O_5+ ZnO) – M (Na_2O+Li_2O+K_2O+TiO_2+ZrO_2+CeO_2)

Experimental glass samples of the D series were welded under identical conditions at temperatures of 1523–1553 K in fireclay crucibles.

When studying the processes of phase formation, structure, and phase composition of materials, complementary physicochemical analysis methods were employed: X-ray diffraction (using a DRON-3M diffractometer) and petrographic analysis (using a NU-2E optical microscope). The coefficient of thermal linear expansion (CLTE) was determined using a vertical quartz dilatometer DKV-5A in accordance with ASTM C372-94(2020).

The temperature dependence of the specific volume electrical resistivity ($\log \rho = f(1/T)$) of thermally treated experimental glass powders was investigated using a two-electrode setup in the temperature range of 293–473–293 K, with a heating rate of τ = 10–15 minutes and a cooling rate of τ = 2–3 minutes per 20 K.

III. Experimental Part

A. Determination of the Requirements for Glass Powder Compositions for the POESTA Technology

A number of factors may influence the electrostatic charging behavior of powders during the coating application process [17], the main ones being:

– the effective work function (dependent on the particle surface composition);

– particle separation rate (a higher separation rate leads to increased charge);

– high intensity and frequency of particle collisions generate more charges;

– types of contact between particles and other material surfaces with different work functions, such as contact with the container surface;

– humidity and temperature of the working environment;

– specific electrical resistivity of the powder, etc.

During powder application, it is critical to maintain proper powder preparation and application parameters. During processing, particles may come into contact with metallic equipment surfaces and adhere to them [14].

Increasing the dispersion of powders, particularly to fractions with particle sizes below 20 μm, may enhance powder cohesion, as adhesion forces within the powder become comparable to or greater than gravitational forces. Electrostatic forces acting on insulating particles can cause high cohesive stresses in the range of 10^4 to 10^7 N/m^2. Powders with a specific resistivity greater than 10^{13} N/m^2 behave as insulators, with charge decay times ranging from minutes to hours, allowing sufficient time for subsequent processing of charged particles. These particles will exhibit strong attraction to the target surface, ensuring good adhesion. However, the accumulation of many charged particles on a non-conductive surface may hinder the deposition of incoming powder particles. Therefore, optimization of various electrostatic effects is necessary to achieve "optimal" coatings with high transfer efficiency [18].

B. Selection of Experimental Glass Compositions and Their Synthesis

Based on previous studies aimed at obtaining bioactive glass-ceramic materials and coatings [14, 15], four base glass compositions were selected (Table 2). To enable fine-grained crystallization of the glass with the formation of calcium phosphates, the phase-forming components in the glass compositions were chosen with a $CaO : P_2O_5$ molar ratio of 1.45–1.80. The content of glass-forming components SiO_2 and B_2O_3 was varied within the ranges of 45–50 and 4.5–6 mol. %, respectively.

TABLE II. Chemical composition of experimental glasses

Oxide content, mol. %		Marking			
		D-1	D-2	D-3	D-4
SiO_2		50.00	45.00	50.00	50.00
R_2O	Na_2O	15.00			
	K_2O	2.50			
	Li_2O	2.50			
RO	CaO	7.50	9.00	8.00	9.00
	ZnO	2.50	1.00	2.00	1.00
RO_2	TiO_2	4.75	4.75	7.50	2.75
	ZrO_2	5.00	2.00	1.00	0.75
	CeO_2	0.25	0.75	0.50	0.50
P_2O_5		4.50	5.00	5.50	6.00
B_2O_3		4.00	10.0	4.50	8.00
Al_2O_3		1.50	2.50	1.00	2.00

The content of modifying components R_2O and RO in all glass compositions was kept constant (20 mol.% and 10 mol.%, respectively), with the ratio $Na_2O : K_2O : Li_2O = 6 : 1 : 1$ to ensure mixed alkali and $R_2O : RO = 2 : 1$ to achieve mixed cation effects. TiO_2, ZrO_2, and ZnO were selected as crystallization catalysts for calcium phosphates (Table 2).

To promote nucleation and crystalline phase formation at lower temperatures, ZnO and CeO_2 were simultaneously introduced into the glass composition. In combination with P_2O_5, these components facilitate phase separation and the formation of a nanodispersed glass structure during heat treatment. The crystallization ability of glasses containing more than 0.5 wt.% CeO_2 is significantly enhanced by the presence of CaO or ZrO_2 [19–21].

After melting, glasses D-1 and D-2 exhibited opalescence, while compositions D-3 and D-4 were translucent.

X-ray diffraction and petrographic analysis of the samples after melting (Table 3) revealed that sample D-2, which had the lowest SiO_2 content, demonstrated the highest crystallization ability among the investigated glass materials. This sample showed intensive crystallization of TiO_2 in the rutile form, which was observed immediately after melting and intensified after heat treatment.

TABLE III. Type and amount of crystalline phase in experimental samples

Sample Marking	The type and quantity of the crystalline phase, vol. %	
	after melting	after heat treatment
D-1	baddeleyite – 5; quartz – 4	rutile – 5; baddeleyite – 7; quartz – 10
D-2	rutile – 10; quartz – 3	rutile – 30; quartz – 3; zircon – 3; others – traces
D-3	HA – 25; quartz – 5	HA – 40; quartz – 5; rutile – 5
D-4	HA – 20; quartz – 5	HA – 55; quartz – 5; rutile – 5

Model glass materials D-3 and D-4 were characterized by the presence of hydroxyapatite (HA) crystals already after melting, in amounts of 20–35 vol. %, with a $CaO : P_2O_5$ molar ratio of 1.45–1.50.

For glasses D-1 to D-3, rutile crystallization is characteristic both after melting and after heat treatment at 1123 K (Fig. 3). This behavior is attributed to the presence of CaO and ZrO_2 in the glass compositions, which promote TiO_2 crystallization by creating favorable conditions for the transition of titanium from tetrahedral to octahedral coordination.

Unlike TiO_2 crystallization, which is largely influenced by the content of alkali metals, the crystallization of ZrO_2 in glass materials is primarily determined by its own concentration. Experimental glass materials D-1 and D-2, containing ≥ 2.5 mol.% ZrO_2, exhibit the presence of ZrO_2 in varying amounts already after melting. In contrast, glasses D-3 and D-4 with a ZrO_2 content of 0.75–1.0 mol. % show only trace amounts of baddeleyite.

Based on the obtained data (Fig. 3) on the phase composition changes during heat treatment, samples D-3 and D-4 were selected for further studies. These materials are characterized by the formation of bioactive hydroxyapatite crystals in amounts of 40–55 mol. % after heat treatment at 1123 K.

C. Study of the Properties of Glass-Ceramic Materials

Investigation of the CLTE of the experimental samples revealed that for materials D-3 and D-4, the values of $\alpha \cdot 10^{-7}$ at 773 K are 129.3 and 141.1 K, respectively. These values are primarily attributed to the high content of (HA) crystals. The obtained CLTE values are close to that of medical stainless steel, $\alpha_{500} = (134–136) \cdot 10^{-7}$ K, which suggests the potential for producing coatings with strong adhesion to the metallic substrate.

The presence of aluminum and boron oxides, along with the crystallization of baddeleyite and rutile during heat treatment, reduces the CLTE of sample D-2 to $90 \cdot 10^{-7}$ K, making it unsuitable for enameling medical stainless steel. The CLTE value of sample D-1 lies between the other samples, with $\alpha_{500} = 115.8 \cdot 10^{-7}$ K, which is also insufficient for producing defect-free coatings for medical stainless steel.

According to the temperature dependence data of the specific volume electrical resistivity of thermally treated glass powders (Fig. 4), the logarithm of ρ_v increases during heating from room temperature to 393–413 K. This is due to

the significant influence of surface conductivity, which is affected by the formation of a thin hydrated film enriched with reaction products from the interaction of alkali components in the glass with atmospheric moisture. Upon cooling, all glass compositions exhibit S-shaped inflection points in the resistivity curves, which are associated with structural transformations in the softening range and with aggregation of certain structural units within the glass.

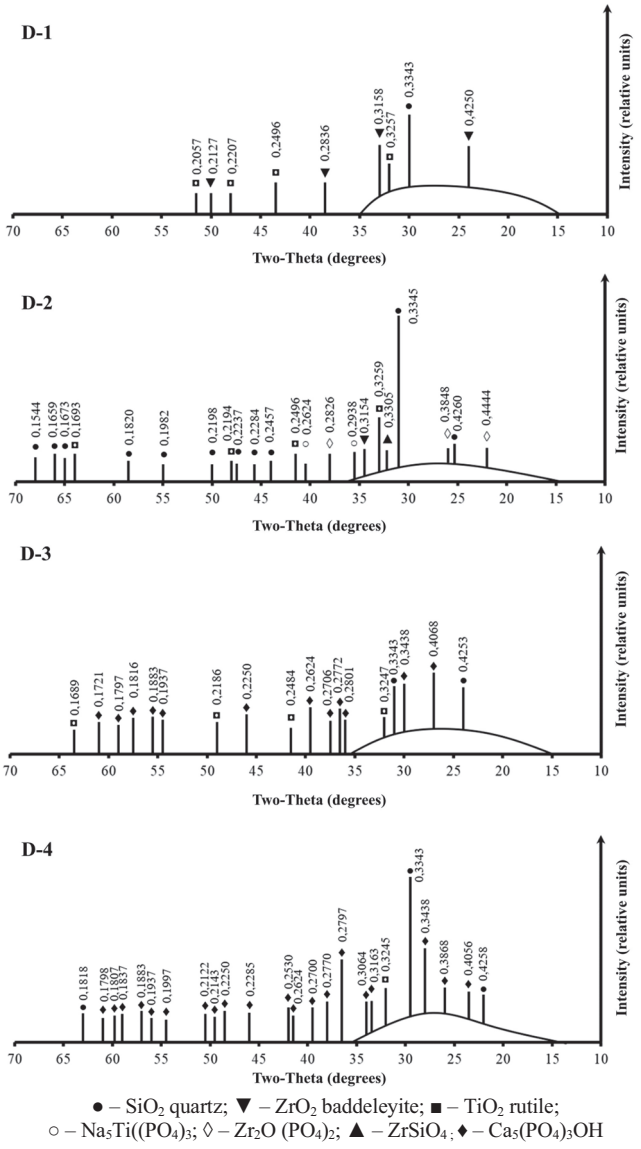


Fig. 3. X-ray Diffraction Patterns of Experimental Samples after Heat Treatment at 1123 K

For multiphase systems, electrical conductivity is determined by the properties of the constituent phases, their volume ratios, and distribution patterns. In sample D-1, the quartz crystalline phase significantly affects overall conductivity and dielectric losses, while in samples D-3 and D-4, both quartz and HA have a notable influence. In apatites, the only charge carriers are H^+ and F^- ions, which are localized on pore walls. Conductivity due to these ions in apatites is associated with their proton-acceptor properties, while intrapore water contributes by forming a network of continuous hydrogen bonds, which facilitates charge transport.

An increase in TiO_2 content to 7.5 mol.% in sample D-3 leads to a decrease in $\log \rho_v$. However, increasing ZnO content at the expense of CaO allows for only a slight decrease in this parameter.

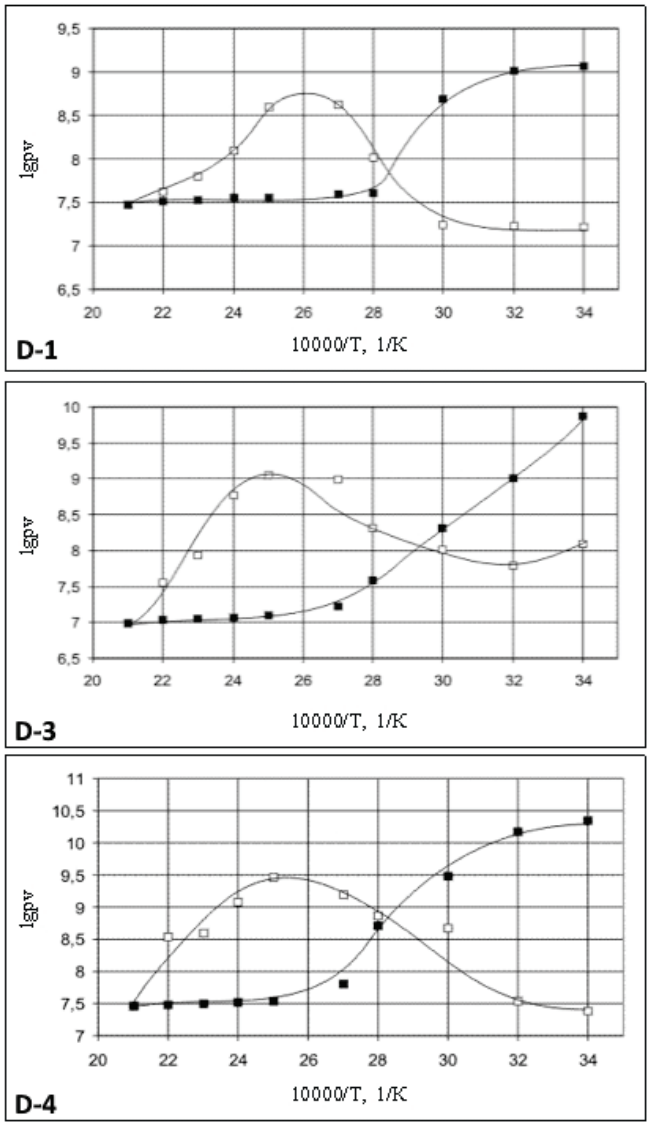


Fig. 4. Temperature Dependence of the Specific Volume Electrical Resistivity of Experimental Samples

In sample D-4, an increased amount of Al_2O_3 and B_2O_3, along with a higher content of HA compared to D-3, positively influences an increase in $\log \rho_v$ to 10.38.

Raising the CaO content to 8 mol. % in glass D-3 and to 9 mol.% in glass D-4 increases the $\log \rho_v$ values to 9.81 and 10.38, respectively. Taking into account the presence of HA crystallization and their respective CLTE values ($\alpha \cdot 10^{-7}$ = 129.3 and 141.1 K), these materials can be considered suitable for producing bioactive glass-ceramic coatings using the POESTA technology.

IV. Conclusion

Based on the analysis of modern methods for forming biocompatible coatings on metals, it has been determined that the resource-saving POESTA technology is the most promising for applying bioactive coatings to medical-grade steels used in the production of implants for maxillofacial surgery and lower jaw prosthetics. The use of this technology ensures the formation of defect-free coatings on

one side of flat implants, enhancing implant fixation, improving the efficiency of glass powder usage, and reducing product costs.

Biocompatible calcium-silico-phosphate glasses based on the R_2O–RO–P_2O–B_2O_3–Al_2O_3–SiO_2 system were developed with oxide compositions (mol.%) as follows: Na_2O = 15.0; K_2O = 2.5; Li_2O = 2.5; CaO = 7.5–9.0; ZnO = 1.0–2.5; TiO_2 = 2.75–7.50; ZrO_2 = 0.75–5.00; CeO_2 = 0.25–0.75; P_2O_5 = 4.5–6.0; B_2O_3 = 4.0–10.0; and Al_2O_3 = 1.0–2.5. These compositions maintain modifier oxide ratios of Na_2O:K_2O:Li_2O = 6:1:1 and R_2O:RO = 2:1, and phase-forming oxide ratios of CaO:P_2O_5 = 1.45–1.80. These glasses exhibit high specific volume electrical resistivity (lgρv = 9.81–10.38) and required thermal expansion coefficient values ($\alpha \cdot 10^{-7}$ = 129.3–141.1 K), making them suitable for producing bioactive coatings on medical steels using POESTA technology.

The main conditions for obtaining defect-free bioactive glass-ceramic coatings on medical steels (95X18, 40X13C, 1.4125, 440B, etc.) were identified, including the following:

– ensuring the presence of a glass phase with high electrical resistivity through polialkali and polycation effects achieved with component ratios of Na_2O:K_2O:Li_2O = 6:1:1 and R_2O:RO = 2:1;

– crystallization of biocompatible phases of hydroxyapatite (20–25 vol.%) and quartz (5 vol.%) after melting, enabling coating formation under low-temperature (850 °C) short-term heat treatment with a CaO:P_2O_5 ratio of 1.45–1.50 and the addition of crystallization catalysts ZnO = 1.0–2.0 mol.%, TiO_2 = 2.75–7.50 mol.%, ZrO_2 = 0.75–1.0 mol.%, and CeO_2 = 0.5 mol.%;

– achieving the required thermal expansion coefficient values ($\alpha \cdot 10^{-7}$ = 129.3–141.1 K) through the crystallization of hydroxyapatite and quartz in a total amount of 45–50 vol.% during heat treatment.

The developed biocompatible calcium-silico-phosphate glasses based on the R_2O–RO–P_2O–B_2O_3–Al_2O_3–SiO_2 system are promising for producing coatings on medical stainless alloys using POESTA technology, ensuring the bioactivity of implants for maxillofacial surgery and lower jaw prosthetics.

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Contactless Dielectric Liquids Sensing Using a Microwave Inverse Split-Ring Resonator

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***Abstract*—The aim of this work is to present the features of a compact, non-contact dielectric liquid characterization sensor based on a planar microwave resonator. Both experimental and numerical studies demonstrate that an Inverse Split-Ring Resonator (ISRR), operating at about 6 GHz, can effectively characterize dielectric liquids inside of a plastic container. The sensitivity of the proposed sensor is compared with that of existing microwave non-contact liquid sensors, and its advantages are discussed.**

Keywords—non-contact characterization, inverse split-ring resonator, dielectric liquid, permittivity, sensor, microwave frequency band.

I. Introduction

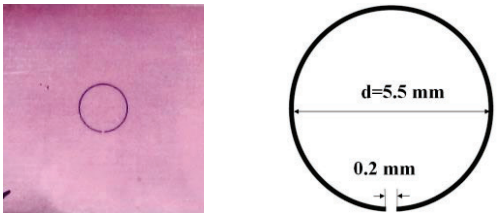
The development of noncontact sensing techniques for dielectric liquids placed within sealed containers [1–5] remains an area of sustained interest. Such sensors are applied in control systems for the analysis of liquid composition, both in the food industry and in security screening of bottles at airports and public events. Among the established methods utilizing the interaction of microwave band electromagnetic waves with liquids enclosed in radio-transparent containers, several stand out. These include techniques based on the small perturbation method of resonators [1, 5–7], measurement of the parameters of a partially opened waveguide with an inserted container [3], solenoid-based measurements [4], and various radar-based approaches [2].

In particular, in ref. [1] it is shown that by measuring the parameters of an open metal resonator loaded with a ceramic dielectric resonator operating in specific modes, it is possible to successfully detect features of liquids inside a container. However, such dielectric resonators are typically bulky and costly, which limits their practical applications. Planar resonators in the form of microstrip photonic crystal analogs with a defect [5–7], operating in the 1–10 GHz frequency band, can also be effectively used as liquid sensors. These structures are significantly more compact compared to [1] and exhibit much higher electromagnetic field concentration within the material under test. Nevertheless, their overall dimensions remain relatively large.

Therefore, the development of a compact and cost-effective sensor for the liquid detection placed into containers remains a relevant and pressing task. To address this challenge, the present work proposes the use of a planar resonator in the form of an Inverse Split-Ring Resonator (ISRR) [8], operating on its fundamental mode. The advantages of this sensor are analyzed below, and its ability to distinguish between different types of liquids – demonstrated using water and gasoline as examples – is experimentally validated.

II. Experimental and Simulation Technique

To investigate the effectiveness of non-contact dielectric liquids sensing, a special ISRR-based resonator was developed and fabricated (Fig. 1). The electromagnetic properties of the family of these resonators were studied in [8, 9]. The given resonator exhibits the resonance frequency of about 6 GHz and is electrically small, with a diameter of 5.5 mm – significantly smaller than the corresponding free-space wavelength. The choice of this resonator as a dielectric sensor is motivated by two key factors. Firstly, the electric component of its resonant mode is strongly concentrated in the opposite part of a split gap (Fig. 2) in the *xOy* plane. Secondly, the alternating *e*-field distribution of the selected mode exhibits an elongated profile along the *Oz*-axis. Such an extended resonant field effectively functions as a sensing probe that penetrates through the container wall and interacts with the liquid under test.



(a) (b)

Fig. 1. Top view of the planar inverted split-ring resonator for non-contact dielectric liquid sensing: (a) photograph; (b) scheme.

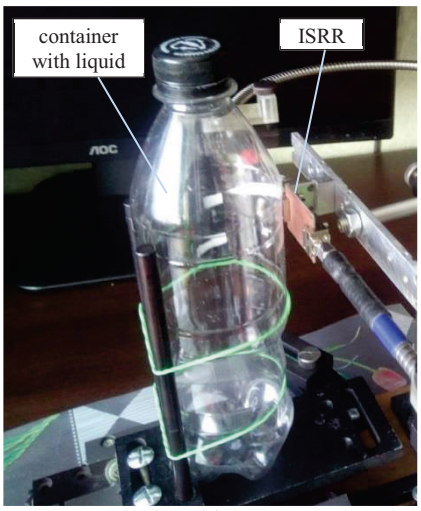
The work is funded by the Project "Microwave express detection of flammable and potentially explosive substances using planar technologies" (Agreement №2.7/25-II on 01.01.2025 between the National Academy of Sciences (NAS) and the O.Ya. Usikov Institute for Radiophysics and Electronics of the NAS of Ukraine).

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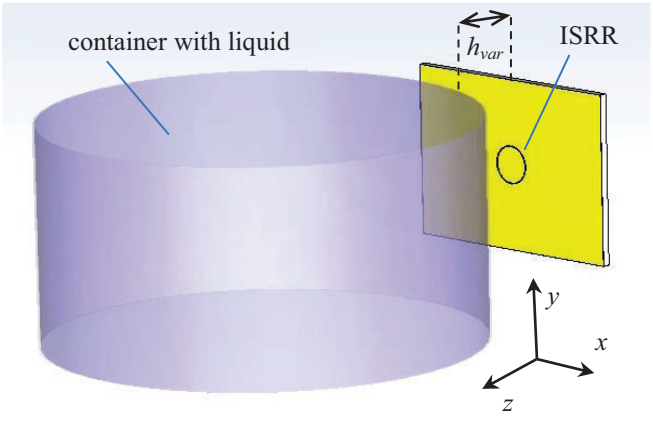
The ISRR (Fig. 1) is fabricated on one side of a dielectric substrate (Neltec NX9240 laminate, thickness $h = 1.143$ mm; permittivity $\varepsilon_s' = 2.4$, $tg\delta_s = 0.0016$), which is coated with a 35 μm layer of copper. The resonator is represented by the etched circular strip with a width of $t = 0.13$ mm, featuring a narrow bridge of width $\delta = 0.2$ mm that connects the inner region of the resonator to the surrounding ground plane. A feeding microstrip line with a width of $w = 3.2$ mm is positioned on the opposite side of the sensor, passing through the geometric center of the resonator.

To investigate the sensitivity of the resonator for non-contact liquid sensing in sealed containers, two liquids with significantly different permittivities were selected: gasoline ($\varepsilon' = 2.4$; $tg\delta = 0.001$) and distilled water. The water dielectric permittivity was calculated using the Debye model [10] with parameters: $\varepsilon_s = 78.4$, $\varepsilon_i = 3.1$, the relaxation time $\tau = 8.27*10^{-12}$s. The liquids were placed in containers – cylindrical plastic bottles with a volume of 500 ml and a maximal diameter of $D = 70$ mm. The container walls, made of polyimide, had a thickness of $h_w = 0.3$ mm and a real part of permittivity of $\varepsilon_w' = 3.5$.

The experimental setup for liquid sensing is shown in Fig. 2a. It consists of a cylindrical container filled with the test liquid, placed in close proximity to the ISRR.



(a)



(b)

Fig. 2. Experimental setup for non-contact sensing of liquids in a container: (a) photo; (b) simulation view.

The resonator is fed by a microstrip transmission line, which is connected via coaxial adapters and cables to a Keysight P9374A Vector Network Analyzer [7]. In this configuration, the container acts as a small perturbation to the electromagnetic field concentrated in the resonator (ISRR) [11]. The Vector Network Analyzer is used to measure the transmission coefficient of the electromagnetic wave propagating through the microstrip line and the resonator in the frequency band of 4–8 GHz.

For comparison with the experiment, numerical modeling of electromagnetic wave propagation through the studied system (ISRR and container with liquid) was performed using the known finite difference in time domain (FDTD) method implemented in the CST Studio Suite electromagnetic simulation software. The simulated transmission coefficient minimum corresponding to the resonant frequency of ISRR is clearly observed in the spectrum (Fig. 3).

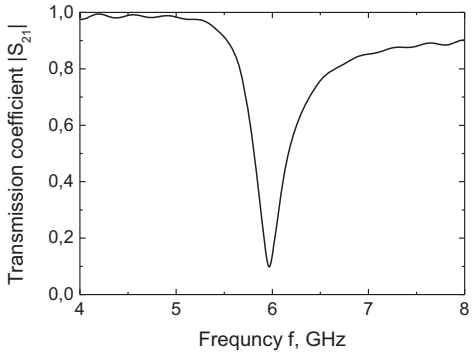


Fig. 3. Transmission coefficient amplitude spectrum of the inverted split-ring resonator without the container, obtained from numerical simulation.

The container was positioned near that spatial region of the ISRR where the electric field amplitude is maximal, as determined by preliminary numerical simulations. The liquid in the container was measured at various distances h_{var} between the resonator surface and the container wall. A special positioning device for moving the liquid container was used for this purpose [7].

During the measurements, the container was displaced along the normal to the resonator surface, following the *Oz* axis (see Fig. 2b). The distance from the resonator surface to the container wall was varied within the range $h_{var} = 1$–20 mm. For each selected distance, the transmission coefficient spectrum of the electromagnetic waves through the ISRR near its resonance frequency was analyzed.

The analysis of the obtained data was carried out following the established procedure described in [1, 5]. The results are presented graphically in Figures 4–6 as follows.

First, from the approximation of the transmission coefficient spectrum, the resonance frequency f_{res} and the inverse quality factor $1/Q$ of the resonator were determined [1, 5].

Next, we plotted the normalized dependence of these parameters, $1/Q = f(f_{res})$, relative to their values for the unloaded resonator, as a function of the distance h_{var} between the container and the resonator (see Fig. 2). It is evident that all curves on each graph converge at the point where $f_{res}=1$ and $1/Q = 1$. This point corresponds to the unloaded

resonator state, which occurs when the distance between the sensor and the container becomes very large (approximately $h_{var} > 40$ mm for this resonator).

As the container moves away from the sensor, the curves diverge according to the values of the real and imaginary parts of the liquid permittivity, ε' and ε''. Easy to show (see for example [1, 5–7]), the parameter $1/Q$ is primarily affected by the imaginary part of the liquid permittivity ε'', while the resonance frequency f_{res} is mainly affected by the real part ε'. The spacing between consecutive points on each graph, connected by lines, corresponds to an increment of approximately 0.2 mm in h_{var}.

III. Results and Discussion

To evaluate the sensitivity of the given resonator/sensor for characterization of various liquids, experimental measurements of the dependence $1/Q = f(f_{res})$ were conducted at different distances h_{var} (see Fig. 4). In this case, the axis of the container was aligned along the *Oy*-axis. The experimental data were compared with numerical simulations, showing good agreement (Fig. 4) in the position of the $1/Q = f(f_{res})$ curves for both liquids (distilled water and gasoline).

Note that, as seen in Fig. 4, the inverse quality factor $1/Q$ of the resonator exhibits a very weak dependence on the distance h_{var} for gasoline, whereas for water this parameter changes significantly. This behavior can be explained by the lower electromagnetic losses in gasoline compared to water at the chosen resonance frequency. Consequently, water has a stronger impact on the resonance frequency than gasoline.

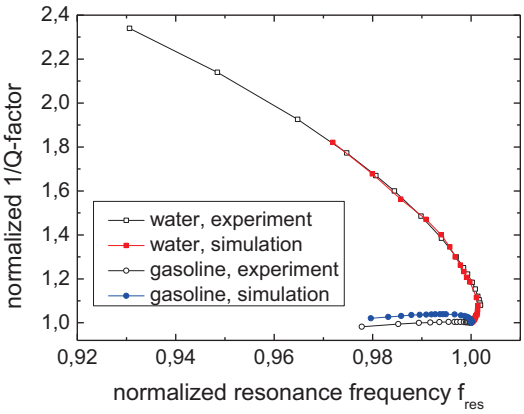


Fig. 4. Experimental and simulated dependencies of the resonance frequency and inverse quality factor of the inverted split-ring resonator at various distances from the container with the test liquid.

For further improving this technique, the investigation of influence of the cylindrical container's axis orientation relative to the coordinate axes on the dependence of the ISRR's resonance frequency and inverse *Q*-factor on the distance to the container was conducted (Fig. 5). In numerical modeling, the cylinder axis was aligned along the *Oy* axis in the first case, and along the *Ox* axis in the second case (see Fig. 2b). Distilled water and gasoline were chosen as the test liquids (see Fig. 5).

From Fig. 5, it is evident that for gasoline, the orientation of the container's axis relative to the resonator has a negligible effect on the simulation results, whereas for water, this effect is more pronounced.

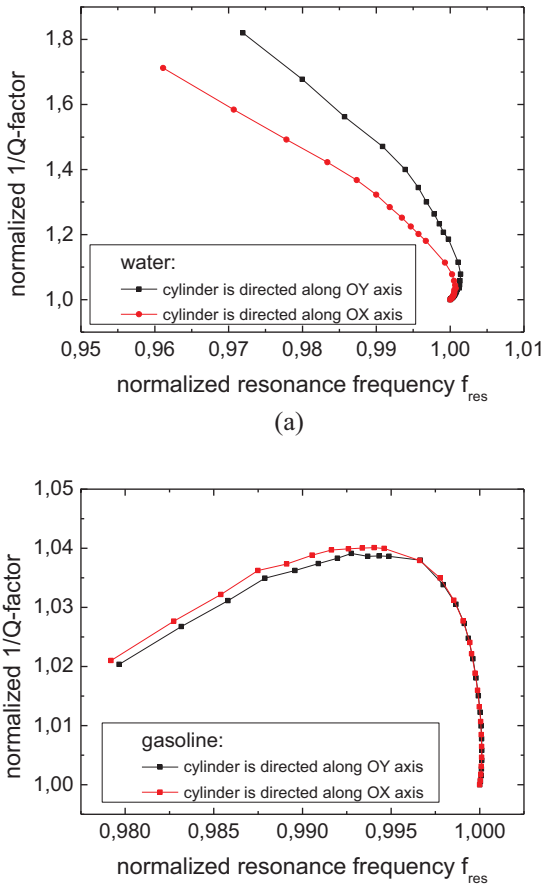


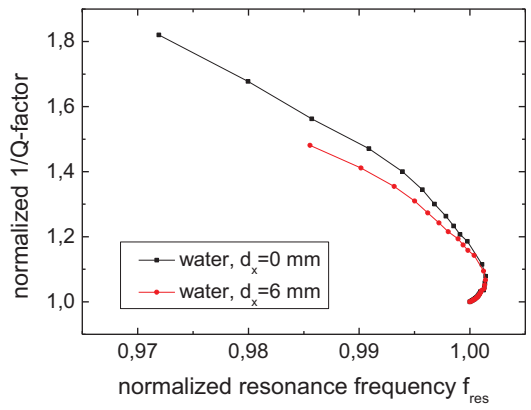
Fig. 5. Numerically calculated dependencies of the resonance frequency and inverse quality factor of the inverted split-ring resonator at various distances from the cylindrical container with the test liquid for two orientations of the cylinder axis relative to the coordinate axes: (a) distilled water; (b) gasoline.

However, it should be noted that the observed changes in f_{res} and $1/Q$ due to the relative orientation of the container's axis are significantly smaller than the changes in these parameters caused by the type of liquid (water or gasoline), as shown in Fig. 4. This implies that sensor calibration must be performed for each specific orientation of the container relative to the coordinate axes.

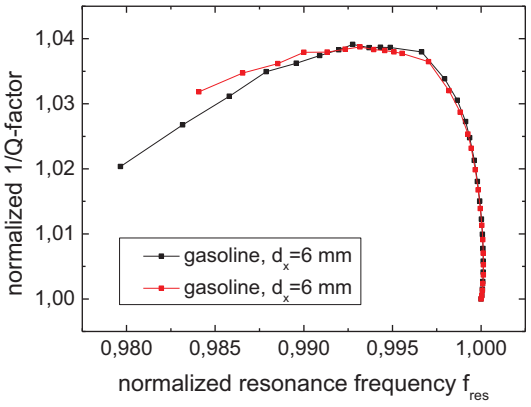
Additionally, a numerical investigation to the influence of the container's displacement (d_x) along the *Ox* axis (with the cylinder axis oriented along the *Oy* axis) on the dependencies of f_{res} and $1/Q$ for two types of liquids (distilled water and gasoline) was conducted (see Fig. 6). Two cases were considered: when the cylinder axis passes through the normal line projected from the center of the resonator ($d_x = 0$ mm), and when this axis is shifted by $d_x = 6$ mm.

As a result, Fig. 6 shows that the effect of shifting the container relative to the resonator on the simulation results is minimal for both liquids (gasoline and water). However, it should be noted that when the cylinder is shifted by a distance exceeding the resonator diameter (more than 6 mm), the resonator's sensitivity slightly decreases. Thus, small displacements only on the order of the resonator diameter

have almost no impact on the accuracy of liquid identification.



(a)



(b)

Fig. 6. Numerically calculated dependencies of the resonance frequency and inverse quality factor of the inverted split-ring resonator at various distances from the cylindrical container with the test liquid, considering displacement of the cylinder axis along the *Ox* axis for: (a) distilled water; (b) gasoline.

Now, let us evaluate the sensitivity of the investigated ISRR in terms of the resonance frequency shift (f_{res}) and the inverse quality factor variation ($1/Q$), compared to the established counterparts of contactless planar liquid sensors. It was demonstrated during our experiments that the ISRR exhibits a significantly larger resonance frequency shift – up to 7% (see Fig. 4) – at small distances between the container with liquid and the resonator, compared to a planar photonic crystal sensor, which shows shifts up to 2% [7]. However, changes in the inverse quality factor are comparable for both sensors.

This enhanced sensitivity to the resonance frequency shift can be attributed to the substantially higher filling factor of the ISRR's electric component of the electromagnetic field within the container with liquid, resulting from the small transverse dimensions of the ISRR (the resonator diameter is of the order of $\lambda_0/10$) relative to the free-space wavelength λ_0 and the container size.

Conclusions

In summary:

- It has been demonstrated both experimentally and numerically that a sensor in the form of an Inverted Split-Ring Resonator (ISRR) effectively distinguishes between different types of dielectric liquids placed into a thin plastic container at frequencies about 6 GHz.
- Experimental results confirm that the inverted split-ring resonator (ISRR) under study exhibits the sensitivity (the frequency shift of the resonance) several times greater at small distances between the liquid container and the resonator compared to a cylindrical resonator [1] and planar photonic crystals [5–7].
- Numerical analysis has been performed to investigate the influence of the position of the cylindrical container's axis relative to the sensor's symmetry axis on the measured resonance frequency and inverse quality factor. Specifically, it is shown that shifting the cylinder's axis by an amount less than the resonator diameter does not significantly affect the liquid identification results.

It is expected that resonators such as the studied here ISRR can be used as compact, highly sensitive, non-contact sensors for characterizing dielectric liquids enclosed in sealed containers.

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Development and Research of Special Properties of Concrete Based on Alkaline Earth Elements

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***Abstract*— The paper presents the results of experimental studies in the field of ferroelectric and piezoelectric materials. Theoretically, using the mathematical method of planning an experiment when constructing and studying composition-property diagrams, the optimal composition of aggregate for concrete based on alkaline earth aluminates was determined. To obtain concrete with high strength, density, and homogeneity, the optimal composition of concrete includes a three-fractional mixture of aggregate consisting of barium titanate. Concrete samples were experimentally obtained and its main electrophysical properties characterizing piezoelectric materials were determined: dielectric constant, dielectric loss, electrical resistance, and piezoelectric modulus. It has been proven that an increase in the content of aluminate cement in concrete leads to a decrease in the value of the concrete piezoelectric modulus. Therefore, for the manufacture of piezoelectric sensors, concrete compositions with the best performance characteristics were chosen, which have 30 % aluminate cement (composition: 40 % $BaAl_2O_4$ + 60 % $BaTiO_3$) and 70 % aggregate (barium titanate). The areas of application of the developed concretes as piezoelectric force pulse sensors characterized by a large surface and significant attenuation of their own resonant vibrations have been determined. Such sensing elements can be used in piezoelectric sensors for electrodynamic conversion of capacitors' electrical energy into a shock pulse for geophysical research in the study of mechanical processes in soils and in the field of seismology in the study of dynamic processes in buildings and structures.**

Keywords — concrete, oxide systems, alkaline earth aluminates, electrophysical properties, polarization, piezoelectric sensors

I. Introduction

The mineral exploration industry in various countries utilizes the technology of seismic wave generation by sources of powerful mechanical impact on the ground. The traditional technology uses explosives, which makes it dangerous and environmentally harmful. There is an alternative technology that uses pulsed seismic sources with different force-drive systems and different ways of impacting the ground. In pulse and vibration technologies of a number of industries as technical means, mechanical, pneumatic, hydraulic, electromechanical converters of reciprocating motion are used. At present the competitiveness of pulse electromechanical converters in comparison with devices of other types has significantly increased. The relevance of their application has increased due to the increase in energy efficiency, reliability and, especially, manufacturability of final products [1-3].

The exploitation of electrodynamic transformations of capacitor electrical energy into shock pulse for geophysical research [4, 5] is associated with the need for constant determination of the shape and amplitude of the shock pulse, since these parameters determine the amplitude and spectrum of excited seismic oscillations.

Produced mono-polycrystalline piezoelectric sensors for measuring seismic signals do not meet the basic requirements for measuring dynamic processes in the operation of electrodynamic type installations. Because of the significant influence of the sensor's own resonant vibrations on the output signal under the impact of a shock pulse, as well as because of the very small size and fragility of mono- and polycrystalline sensors, it is not possible to measure uniformly distributed over a large area of force pulses of 10^6 N received from electrodynamic transducers.

To solve this problem, the authors considered the possibility of obtaining a material for manufacturing a piezoelectric sensor with a large intrinsic surface area based on cement compositions of aluminates of alkaline-earth elements.

Previously conducted by the authors theoretical studies in the ternary system BaO-Al_2O_3-TiO_2 and experimental work on obtaining cement combining strength and electrophysical properties, based on aluminates of alkaline earth elements [6], allowed to obtain a material that can be effectively used in the manufacture of special concretes with segmentoelectric and piezoelectric properties.

The research conducted by the authors makes it possible to obtain and effectively use concretes based on aluminates of alkaline earth elements in the production of piezoceramics, namely, piezosensors for seismic detection and engineering seismology in the study of dynamic processes in buildings and structures.

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II. The Purpose of the Work

The purpose of this work is to obtain and experimentally study the electrophysical properties of concrete based on alkali aluminates. It's planned to determine the rational composition of aggregate for the developed concrete. Investigation of the main parameters characterizing piezoelectric materials: dielectric constant, dielectric loss, electrical resistance, piezoelectric modulus. To study the possibilities of using the developed concretes as piezoelectric sensors, which are characterized by a large surface and significant attenuation of their own resonant vibrations.

III. Theoretical Information and Research Methods

In purpose of obtaining concrete of high strength, density and homogeneity, providing the necessary operational reliability of products we have made a selection of the optimal granulometric composition of aggregate, which has a significant impact on the above parameters.

During the experiments, alumino-titano-barium cement [7] of the composition was used as a binder: 40% $BaAl_2O_4$ + 60% $BaTiO_3$ in the amount of 20% of the weight of the mixture. Barium titanium obtained by synthesis at 1400 °C was used as aggregate.

Concrete specimens with a diameter of $15 \cdot 10^{-3}$ and were made by bilateral pressing of concrete mixtures at a pressure of 100 MPa.

The physical-&-mechanical properties of the elements were determined using standard methods: the open porosity and apparent density were determined according to DSTU ISO 5017:2014 [8] and the compressive strength limit was determined according to DSTU EN ISO 8895:2018 (EN ISO 8895:2006, IDT; ISO 8895:2004, IDT) [9].

According to the results of calculations and mathematical processing of experiments, the "composition-property" diagrams were constructed using the authors' method [10].

Experiment planning matrix and test results of physical and mechanical properties of concrete are shown in Table 1.

TABLE I. Experiment Planning Matrix and Results of Experiments

Polynomial coefficients	Notations and physical meaning of factors				
	Fraction of aggregate, in USD per unit			Experimental data	
	X_1	X_2	X_3	Y_σ	Y_n
η_1	1	0	0	51,2	24,3
η_2	0	1	0	63,5	21,0
η_3	0	0	1	67,2	20,82
η_{12}	0,5	0,5	0	57,5	23,60
η_{13}	0,5	0	0,5	74,1	18,65
η_{23}	0	0,5	0,5	69,7	20,60
η_{123}	0,33	0,33	0,33	73,2	19,60
Control point	0,5	0,3	0,2	68,0	19,30

Fraction of aggregate: x_1 - fraction $(1,25\text{-}0,63) \cdot 10^{-3}$m; x_2 - fraction $(0,36\text{-}0,315) \cdot 10^{-3}$m; x_3 - fraction $(0,315\text{-}0,15) \cdot 10^{-3}$m.

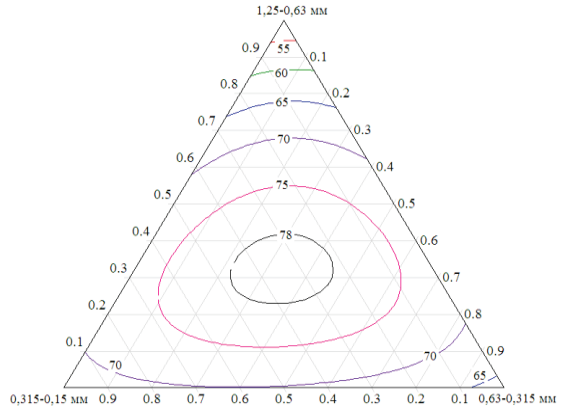
We used an incomplete third order polynomial to describe the dependence of concrete properties on the quantitative ratio of aggregate fractions.

We calculated the coefficients of polynomials expressing the dependence of strength (Y_σ) and porosity (Y_n) on the granulometric composition of the aggregate. After substituting the found coefficients, the regression equations take the formula (1-2):

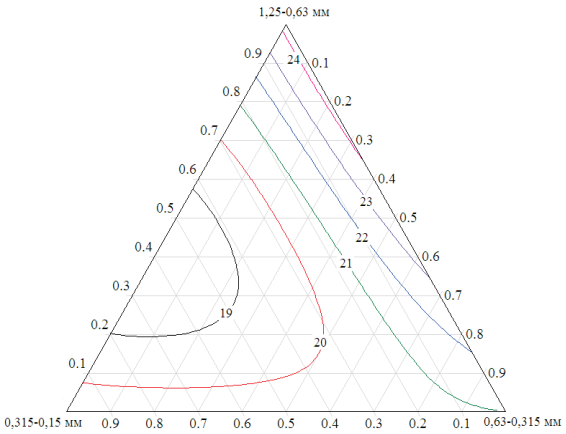
$$Y_\sigma = 51{,}2x_1 + 63{,}5x_2 + 67{,}2x_3 + 60{,}0x_1x_2 + 17{,}4x_2x_3 + 49{,}6x_1x_3 + 106{,}5x_1x_2x_3 \quad (1)$$

$$Y_n = 24{,}3x_1 + 21{,}0x_2 + 20{,}82x_3 + 3{,}8x_1x_2 - 15{,}64x_1x_3 - 1{,}04x_2x_3 - 27{,}24x_1x_2x_3 \quad (2)$$

The adequacy of the equations was checked using the Student's test and additional control experiments. According to the results of mathematical processing of the results of experiments, simplex-diagrams "composition-property" and projections of lines of equal level for the strength and for the porosity of concrete (Fig. 1) were constructed, which allowed to determine the optimal quantitative ratio of adjacent fractions of concrete aggregate.



a)



b)

Fig. 1- The "composition – property" diagram:
a) "composition–strength" concrete; b) "composition–porosity" concrete.

The optimum concrete composition includes a three-fraction aggregate mixture with the following quantitative grain ratio:

- 20...40% fraction $(1,25\text{-}0,63) \cdot 10^{-3}$ m;

- 10...30% fraction (0,36-0,315)·10^{-3} m;
- 30...70% fraction (0,315-0,15)·10^{-3} m.

Further synthesis and research of concretes were carried out taking into account the optimal particle size distribution of aggregate.

IV. Discussion of the Results

The application of concrete based on aluminates of alkaline-earth elements for the fabrication of a piezoelectric sensor of force pulses requires studies of the electrophysical properties of promising materials.

Test samples were prepared from concrete mixtures at a pressure of 50...100 MPa in the form of disks of different sizes. The samples were hardened in air for 7 days, after which the surface of the sample was ground to obtain a minimum of roughness. Materials for electrodes were aluminum, copper, silver and suspensions of colloidal graphite in distilled water, aluminum and copper electrodes were deposited in vacuum, silver electrodes were obtained by burning paste, colloidal graphite was applied with a brush.

The results of tests of electrophysical properties of the investigated concrete (at pressing pressure P = 50 MPa) are shown in Table 2-3.

TABLE II. Electrophysical Properties of the Developed Concrete

No	Composition of concrete,%		Dielectric permittivity	
	Cement	Aggregate	At 50 Hz,	At 10^3 Hz
1	10	90	1517	755
2	20	80	1360	670
3	30	70	834	410

TABLE III. Electrophysical Properties of the Developed Concrete

No	Composition of concrete, %		Dielectric loss angle tangent		Specific volume electrical resistance, Ohm·cm
	Cement	Aggregate	At 50 Hz	At 10^3 Hz	
1	10	90	0,07	0,08-0,1	6 ·10^9
2	20	80	0,08	0,09-0,21	4,4·10^9
3	30	70	0,09	0,15-0,25	6 ·10^9

The research results presented in Table 2-3 show that the developed concrete has high dielectric constant at room temperature. The increase of alternating current frequency leads to a decrease in dielectric permittivity and an increase in the dissipation factor of the dielectric loss angle, which indicates the segmental behavior of the material. The presence of the increased value of the dissipation factor of concrete suggests that piezoelectric sensors with a significant value of attenuation of resonance oscillations can be made on the basis of concrete. In addition, concrete is favorably distinguished by the fact that it allows to create sensors with a large own surface, and this makes it possible to measure pressure pulses distributed over the surface in the study of mechanical processes in soils in the field of elastic deformations.

The investigated concrete is further imparted with piezoelectric properties by polarization with a constant voltage. Before the polarization process, concrete, like cement, is isotropic and does not possess piezoelectric properties. The polarization of concrete samples was carried out by the method described in [7]. The external electric field strength was 0.2...0.4 kV/mm.

The values of piezoelectric modulus ***d*** of different compositions of the obtained concrete are presented in Table 4. For comparison, the table shows the value of piezoelectric modulus ***d*** of ceramic barium titanate, which polarization we performed under conditions similar to the conditions of polarization of the developed concrete.

TABLE IV. Piezoelectric Modulus of Different Compositions of Developed Concrete

No	Composition of concrete. %		Piezomodule d · 10^{12}, K/n
	Cement	Aggregate	
1	10	90	1,0–1,2
2	20	80	0,9–1,0
3	30	70	0,87–0,9
4	Ceramic $BaTiO_3$		1,0–1,4

Comparative analysis of the data in Table 2 shows that the piezomodulus value of concrete based on aluminates of alkaline earth elements is insignificantly different from the piezomodulus value of ceramic barium titanate. Increasing the content of aluminate cement in the concrete composition leads to a decrease in the piezomodulus value of concrete. This decrease in the piezomodulus of concrete is explained by the fact that with the increase in the amount of aluminate cement introduced into the concrete composition, the content of barium aluminate in concrete increases, which does not exhibit piezoelectric properties.

Based on the obtained data, it can be concluded that concrete based on aluminates of alkaline earth elements is a promising material for the manufacture of sensitive elements characterized by large own surface and significant attenuation of natural resonance oscillations. Such sensitive elements can be used in sensors designed to measure pulse shock accelerations distributed over a large area.

It is known that the action of piezoelectric sensors is based on the direct piezo effect, when the pressure applied to the crystal faces is converted into electric charges. When a piezo element is placed between an inert mass and the flat coils of an electrodynamic transducer, it will be affected by the force pulses generated by the flow of electric current through the coils [11-14]. Due to the desire of the inertial element to keep the state of rest, the piezo element deforms from the inertial force acting on it according to formula (3):

$$F = m \cdot a \tag{3}$$

Where m is the mass of the inertial element; a is the transmitted acceleration.

In the case of the longitudinal piezoelectric effect, the amount of electricity arising on the faces of the crystal during its compression will be expressed by the equality (4):

$$Q = d \cdot S \cdot p \tag{4}$$

Where Q is the resulting charge; d is the piezoelectric modulus; S is the area of the plate subjected to compression; p is the pressure.

A characteristic feature of the pressure pulse sensor for geophysical devices is the low-frequency range of recorded oscillations. Necessity to reduce the influence on the output signal of the sensor of its own resonance oscillations under the influence of shock pulses. Realization of the conditions for obtaining from the sensor a large Q at a significant value of damping of resonant oscillations can be achieved by using as a sensor material a concrete with piezoelectric properties on the basis of aluminates of alkaline-earth elements.

Further, the authors worked on the creation of a piezoelectric sensor with large own surface for measuring uniformly distributed force pulses of up to 106 and in electrodynamic type installations.

As a material for manufacturing of the piezoelectric sensor we used the developed concrete of the following composition: cement (40% $BaAl_2O_4$ + 60% $BaTiO_3$) - 30%; aggregate (barium titanate) - 70%. Structurally, the piezoelectric sensor is a disk with a diameter of 0.5 m and a height of 0,04 m. In the center of the disk there is a through hole with a diameter of 0,05 m for the guide rod.

The concrete mixture, prepared as a mass, was placed and tamped into a special mold made of 6 mm thick organic glass. The concrete was cured in the mold for 1 day, after which the sample was removed from the mold. Electrodes from colloidal graphite suspension were applied to the horizontal planes of the sensor and the sensor was kept in air under normal conditions for 6 days, then copper foil leads were glued to the graphite electrodes using electrically conductive glue. The empty weight of the sensor is ~ 32 kg. Polarization of the sensor was carried out under normal conditions by a constant electric field of 1 kV/cm.

Tests of a piezoelectric sensor with a large surface were carried out at the shock installation with an electrodynamic energy converter. The sensor was placed between flat coils (coil diameter 0,5 m), electrodynamic converter of electrical energy of capacitors into shock pulse. Parameters of the shock pulse: $F_{max} = 10^4$-10^5 m; $\tau = 10$ ms. The output signal of the transducer was registered by a memorizing oscilloscope. During the tests, the output signal parameters obtained from the sensor differed from the calculated ones by no more than 5-7%.

It was determined that due to the large diameter of the sensor (compared to the size of commercially available sensors) and high mechanical strength of the material used, it is possible to measure force pulses of 10^6 m evenly distributed over the area.

V. Conclusions

Based on the results of synthesis and studies of electrophysical properties of concrete based on aluminates of alkaline earth elements, we can draw conclusions about the rational composition of concrete, samples of which consist of 30% aluminate cement (40% $BaAl_2O_4$ + 60% $BaTiO_3$) and 70% aggregate (barium titanate). We have experimentally determined the main indicators for the studied samples of the developed concrete: high dielectric permittivity at room temperature up to 1517; increased value of the dissipation factor 0.7 - 0.9; dielectric resistivity - 6-10^9 Ohm·cm, piezomodul - 0.87-0.9 · 1012 K/n. As a result of the work it was determined that the developed concrete based on aluminates of alkaline-earth elements is a promising material for the manufacture of sensitive elements characterized by large own surface and significant attenuation of natural resonance oscillations. Such sensitive elements can be used in sensors designed to measure pulse shock accelerations distributed over a large area.

Therefore, the experimental and theoretical studies show that concrete based on aluminates of alkaline-earth elements is a promising material for the production of piezoelectric sensors. Such devices can be used in seismic exploration for measuring seismic signals excited on the Earth's surface by impact installations, as well as in seismology for studying dynamic processes in buildings and structures.

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Overmodulation Control of Dual and Triple Inverter PV Installations Using Modified SSV PWM

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***Abstract* — This paper presents an overview of features of regulation of blocks of modulated voltage source inverters (VSINs) and neutral-point-clamped inverters (NPCIs) of photovoltaic (PV) installations in the overmodulation control region of inverters, regulated according to a strategy of synchronous space-vector PWM (SSV PWM), providing voltage synchronization and voltage symmetry of the corresponding installlations. Two-stage overmodulation control of inverters in this case is founded on modified control correlations, which include specialized correcting indices for the each stage of the overmodulation control zone. Specific overmodulation-focused algorithms of SSV PWM of inverters have been applied and observed for some structures of grid-tied PV stations: a) PV systems with two two-level inverters; b) dual-inverter PV systems with special scheme of connection of windings of the transformer, and c) two alternative topologies of PV stations based on three modulated inverters.**

Keywords—power inverter, space-vector-based PWM, strings of photovoltaic panels, power transformer, voltage synchronization, spectral analysis

I. Introduction

Power electronics installations are very popular equipment that assures improved quality of technology in many fields including renewable energy [1] – [3].

Photovoltaic (PV) stations including hybrid stations are very popular installations in the area of renewable electrical energy [4] – [11]. Voltage source inverters (VSINs) and neutral-point-clamped inverters (NPCIs) are nowadays the basic blocks of renewable energy systems including PV installlations [12] – [19]. The performance of the inverter-based PV stations depends on the PWM strategies and PWM schemes used to regulate VSINs and NPCIs. To insure synchronization and symmetry of voltage waveforms of power PV inverters with increased power levels, an original strategy and scheme of synchronous space-vector PWM (SSV PWM) was developed and popularized to regulate the inverters of some structures of PV stations [20] – [24], including installlations based on standard two-level VSINs ([20], [22] – [24]), as well as PV stations based on NPCIs [21].

Overmodulation control modes are among the most complicated modes for the operation of PV inverter blocks. Therefore, special attention was paid to the study of overmodulation modes of PV inverters [20]–[24]. This article presents an analysis of basic techniques of SSV PWM, developed for synchronous overmodulation regulation of dual-inverter and triple-inverter-based topologies of PV stations. The novelty of this research lies in the systematic approach to enhancing the performance and reliability of PV inverters under overmodulation conditions, offering innovative solutions to optimize their functionality and extend their operational range in renewable energy systems.

II. Overmodulation Regulation of Dual Inverters of Grid-Tied PV Installation

Structure of the two-VSIN-founded PV station, presented in Fig. 1, contains two VSINs which are supplied by the use of two PV strings either directly to the photovoltaic panels, or through a dc module (dotted line in Fig. 1) [12], [20]. The output circuits of two VSINs are tied to inverter-side windings of power transformer connected with grid.

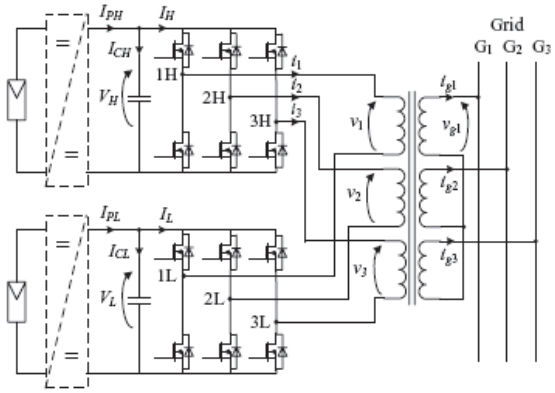


Fig. 1. Topology of PV installation with dual modulated VSINs [12], [20].

Regulation of both VSINs and NPCIs of PV installations in the overmodulation zone, if coefficients of PWM of converters $m > 0.907$, has explicit peculiarities [3], [20]. Boundary values of two indices of PWM of voltage source converters in the case of a standard two-step regulation of inverters during overmodulation control regime are $m_{ov1} = 0.907$ and $m_{ov2} = 0.952$ [3], [20] ($m_{\max} = 1$ in this case). In particular, the main control dependences of modulated VSINs and NPCIs for calculating the parameters of control signals in the overmodulation zone contain two specialized indices [2]:

$$K_{ov1} = [1 - (m - m_{ov1})/(m_{ov2} - m_{ov1})] \quad (1)$$

$$K_{ov2} = [1 - (m - m_{ov2})/(1 - m_{ov2})] \text{ [20]} \quad (2)$$

If $0.952 \geq m > 0.907$:

$$\beta_1 = \tau \quad (3)$$

$$\beta_j = \beta_1 \cos[(j-1)\tau K_{ov1}] \quad (4)$$

$$\gamma_j = \beta_{n-j+1}\{0.75 - 0.55\tan[(n-j)\tau]\} \quad (5)$$

$$\lambda_j = \tau - (\beta_j + \beta_{j+1})/2 \quad (6)$$

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If $1 > m \geq 0.952$:

$$\beta_1 = \tau \tag{7}$$

$$\beta_j = \beta_1 \cos[(j-1)\tau K_{ov2}] \tag{8}$$

$$\gamma_j = \beta_{n-j+1}\{0.75 - 0.55\tan[(n-j)\tau]\}K_{ov2}, \tag{9}$$

where m is index of PWM of VSINs and NPCIs, $\beta_1 - \beta_j$ are total switch-on state inside 60^0-period, γ_j is minor segment of total switch-on state, λ_j are widths of notches, τ is duration of inverter's sub-cycle [20].

As an example of overmodulation regulations of two VSINs with SSV PWM (Fig. 1) under condition of equal dc-voltages, when m = m_H = m_L = 0.950 (control regime of VSINs corresponds in this case to operation of inverters in the first sub band of the overmodulation region), Fig. 2 shows, on the period of the fundamental frequency, pole voltages V_{1H}, V_{1L}, and line voltages V_{1H2H}, V_{1L2L} of two modulated VSINs, as well as phase voltage V_1 (with its harmonic composition) of PV station [20]. Operation frequency of PV installation F = 50.0 Hz, as well as switching frequency of VSINs F_s = 1.350 kHz [20].

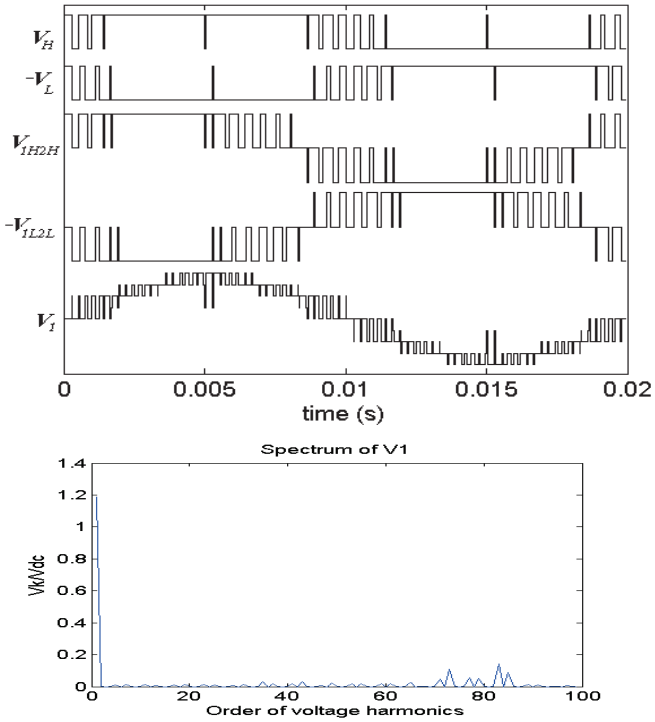


Fig. 2. Pole, line, and phase voltages, and harmonic composition of the phase voltage V_1 of PV station with dual VSINs regulated by discontinuous SSV PWM (DPWM30) in the first sub band of the zone of overmodulation.

Fig. 3 illustrates operation of PV station with dual VSINs during the second sub band of the zone of overmodulation [20]. Fig. 4 shows the results of calculations to determine Total Harmonic Distortion (THD) factor of the phase voltage V_1 for the analyzed structure of PV station [20].

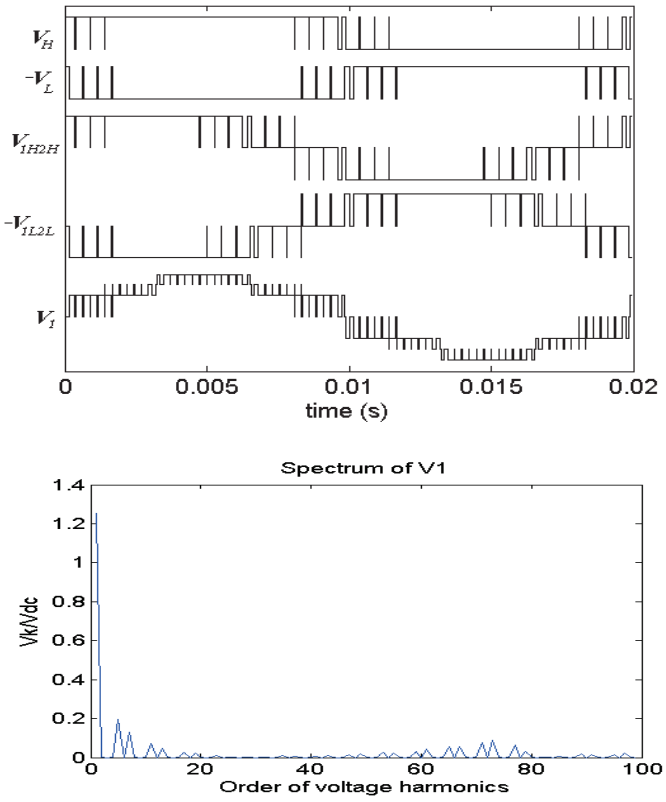


Fig. 3. Pole, line, and phase voltages, and harmonic composition of the phase voltage V_1 of PV station with dual VSINs regulated by discontinuous SSV PWM (DPWM30) in the second sub band of the zone of overmodulation (m = 0.99).

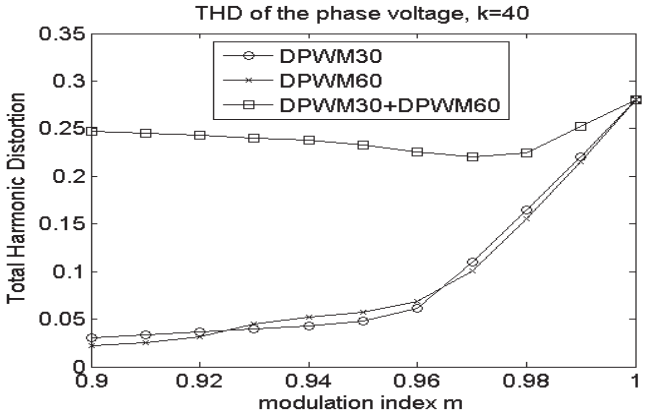


Fig. 4. THD of the phase voltage V_1 of PV station with two VSINs regulated by discontinuous (DPWM60 and DPWM30) and combined (DPWM60 + DPWM30) variants of SSV PWM.

III. Overmodulation Regulation of Dual VSINs of PV Station with Special Connection of Inverter-Side Windings of the Transformer

Structure of grid-tied PV installations with two VSINs can be based on grid-connected transformer with specific interconnection of windings of power transformer of PV installation [17], [22]. Fig. 5 shows topology of PV station based on two VSINs supplied by two PV strings [22]. The special connection between the output circuits of VSINs and inverter-side windings of the transformer is marked with bold lines in Fig. 5 [22].

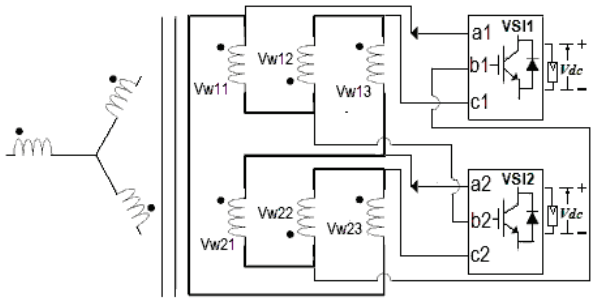


Fig. 5. Topology of PV installation with two VSINs and multi-winding grid-tied power transformer.

In order to illustrate operation of PV with dual VSINs during the first sub band of the overmodulation zone, Figs. 6-7 show basic voltages, including pole voltages of two VSINs V_{a1}, V_{b1}, V_{a2}, V_{b2}, line-to-line voltage V_{a1b1}, winding voltage V_{w11}, and spectral composition of the V_{w11} voltage [22].

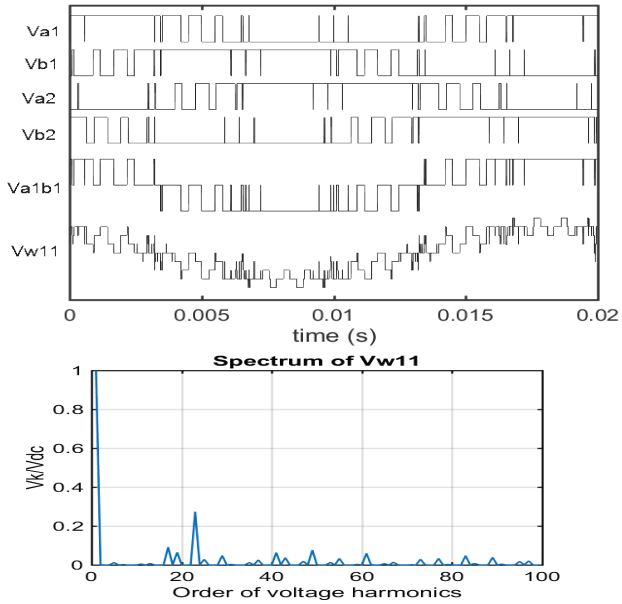


Fig. 6. Pole, line, and winding voltages, and spectrum of the V_{w11} voltage (continuous SSV PWM, m = 0.94).

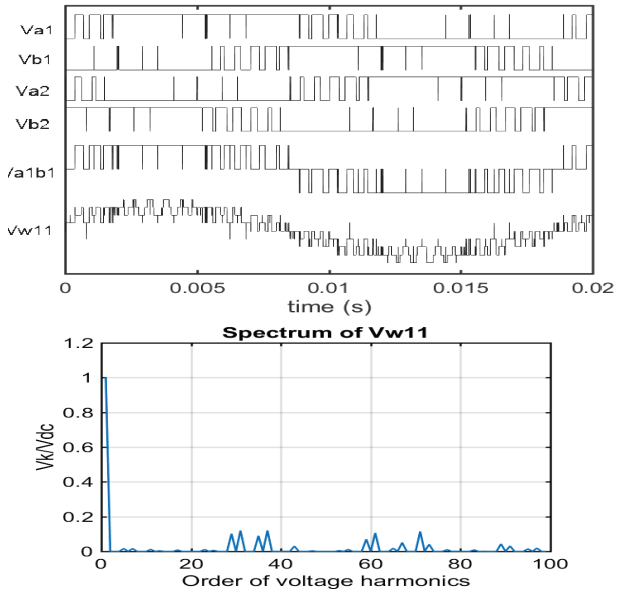


Fig. 7. Pole, line, and winding voltages, and spectrum of the V_{w11} voltage (discontinuous SSV PWM (DPWM30), m = 0.94).

To illustrate operation of PV with dual VSINs during the second sub band of the overmodulation zone, Figs. 8 - 9 show base voltages in the system (pole voltages of two VSINs V_{a1}, V_{b1}, V_{a2}, V_{b2}, line-line voltage V_{a1b1}, winding voltage V_{w11}), and spectrum of the V_{w11} voltage [22].

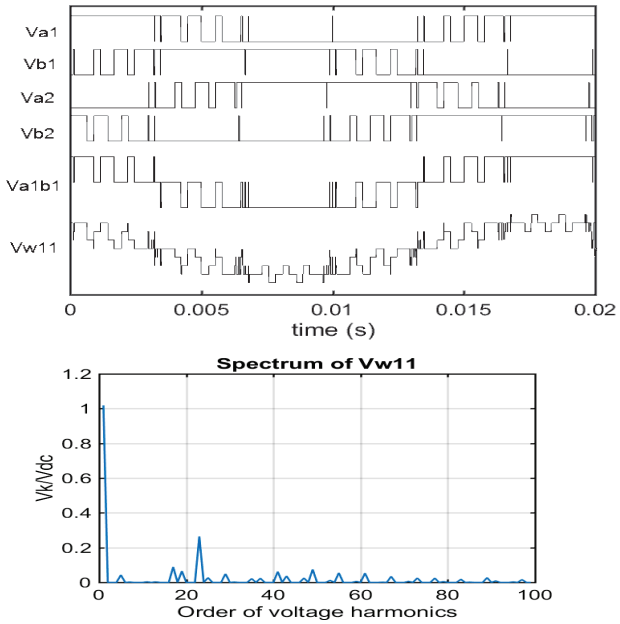


Fig. 8. Pole, line, and winding voltages, and spectrum of the V_{w11} voltage (continuous SSV PWM, m = 0.975).

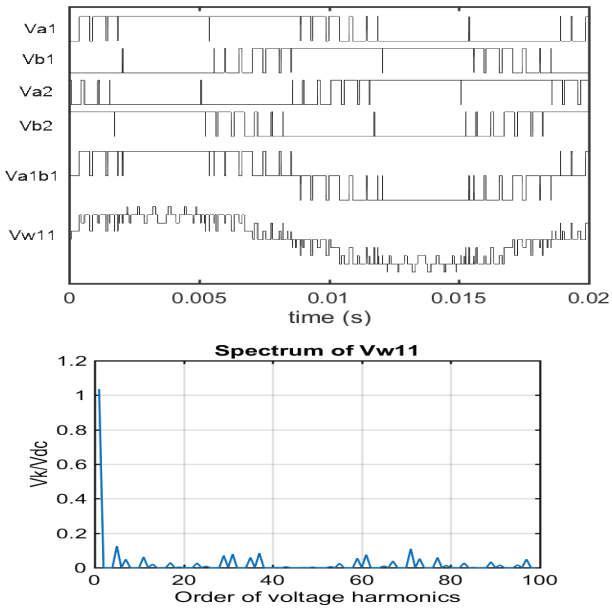


Fig. 9. Pole, line, and winding voltages, and spectrum of the V_{w11} voltage (discontinuous SSV PWM (DPWM30), m = 0.975).

The all presented in Figs. 6 - 9 waveforms of the base voltages are characterized by quarter-wave symmetry. So, its spectra contain only odd harmonics, without unwanted subharmonics.

Fig. 10 presents results of calculations of THD index for the winding voltage V_{w11} of PV installation with dual VSINs regulated in the overmodulation region by continuous (PWM-C), and two discontinuous (PWM-D-30 and PWM-D-60) versions of SSV PWM [22]. Switching frequency of VSINs is 1100 Hz for calculation of THD index of the V_{w11} voltage.

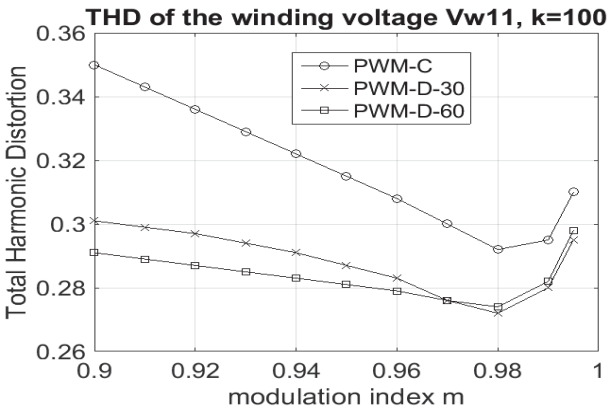


Fig. 10. THD index of the V_{w11} voltage of PV station under condition of overmodulation regulation of VSINs.

IV. Overmodulation Regulation of Triple NPCIs of PV Station with Spesific Interconnection of Windings of the Transformer

Fig. 11 presents structure of the main circuits of PV station ([18], [21]) with triple modulated NPCIs regulated by special control scheme [21]. In this case the winding voltages V_{w1}, V_{w2}, and V_{w3} of the respective windings of the transformer of this structure of PV system can be determined as functions of the phase voltages of triple NPCIs [21]:

$$V_{w1} = V_{as1} - V_{bs2}, \quad V_{w2} = V_{bs2} - V_{cs3}, \quad V_{w3} = V_{cs3} - V_{as1} \quad (10)$$

As described in [21], this switching strategy of NPCIs provides minimization of unwanted common-mode voltage in electrical power apparatuses. In accordance with algorithm of system regulation, control pulses of triple NPCIs are shifted by 120 el. grad., and there is also mutual phase shift between control pulses of triple NPCIs, which is equal to 1/3 of duration of switching sub-interval.

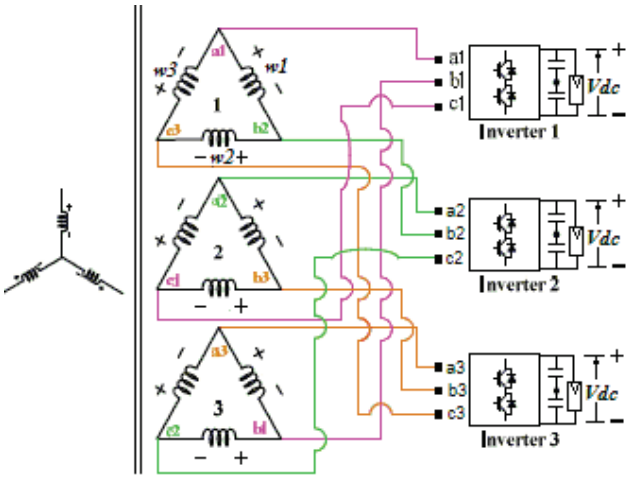


Fig. 11. The main power circuits of PV installation with triple NPCIs and delta configuration of inverter-side windings of the transformer.

Figs. 12 presents basic voltages, which have symmetrical waveforms, and Fig. 13 presents spectra of the V_{a1b1} and V_{w1} voltages for PV station with NPCIs regulated by the scheme of "direct-direct" SSV PWM, index of modulation of NPCIs, operating in the overmodulation mode, is m = 0.950 [21]. So, spectra of the presented voltages do not include in this control zone even harmonics and unwanted subharmonics (of the fundamental frequency of PV installation).

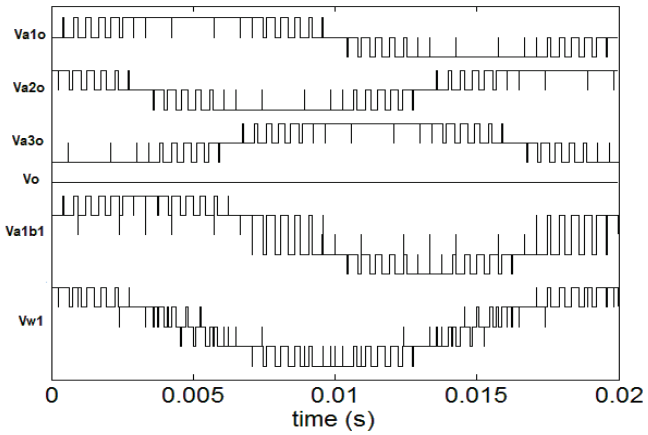


Fig. 12. Pole voltages of NPCIs V_{a10}, V_{a20} and V_{a30}, common-mode voltage V_0, line voltage V_{a1b1}, and winding voltage V_{w1} of PV station with NPCIs regulated by the "direct-direct" version of SSV PWM (m = 0.950, F = 50.0 Hz, F_s = 1.0 kHz).

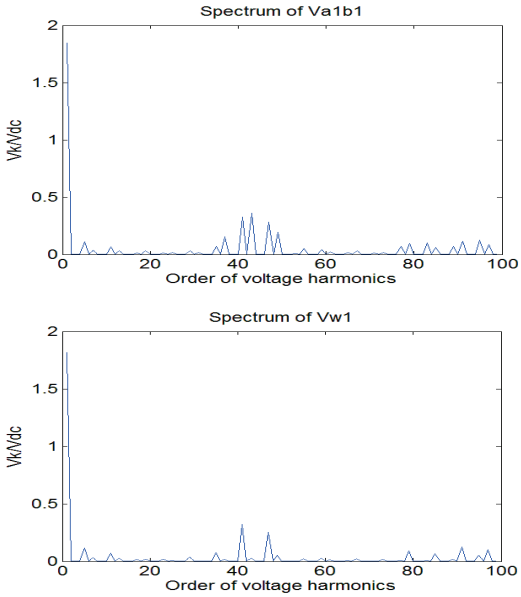


Fig. 13. Spectra of the base voltages of PV station with NPCIs regulated by the "direct-direct" variant of SSV PWM.

V. Overmodulation Regulation of Triple VSINs of Alternative Topology of Three-String PV Station

Fig. 14 shows alternative structure of grid-tied PV installation with three modulated VSINs and with specialized connection of outputs of VSINs with the corresponding windings of power transformer [19], [23].

In particular, this PV topology makes it possible to increase the winding voltage (compared to two-VSINs PV installations), reducing its volume and weight [19].

Fig. 15 – Fig. 17 show pole voltages V_{11}, V_{12} and V_{13} of the first VSIN, line-line voltages ((V_{12} - V_{13}) and (V_{21} - V_{23})) of the corresponding VSINs, and winding voltage V_2 (with its spectrum) of the transformer [23]. Charts in Fig. 15 cor-respond to operation of PV system in the first part of the zone of overmodulation, VSINs are controlled by continuous SSV PWM. Diagrams in Figs. 16 - 17 correspond to operation of PV station in the second part of the zone of overmodulation, VSINs are controlled in this case by discontinuous SSV PWM [23].

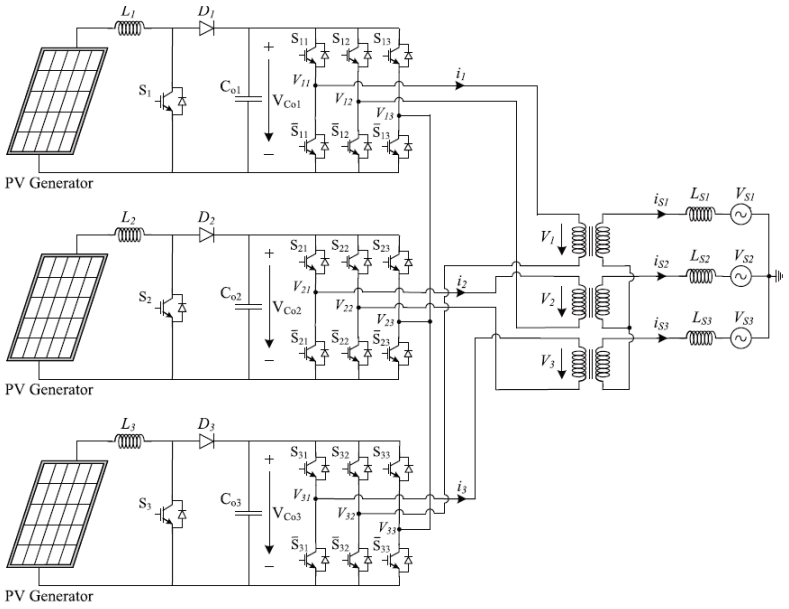


Fig. 14. PV system with triple VSINs with special scheme of their connection with grid-tied transformer .

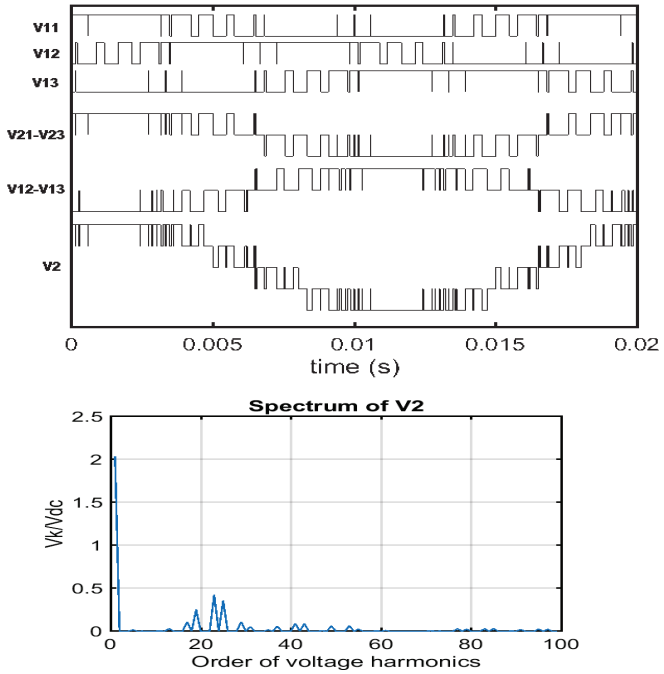


Fig. 15. Pole voltages V_{11}, V_{12} and V_{13}, line-to-line voltages, winding voltage V_2 and its spectrum of PV system regulated by continuous SSV PWM (m = 0.935, F_s = 1120.0 Hz, F = 50.0 Hz) [23].

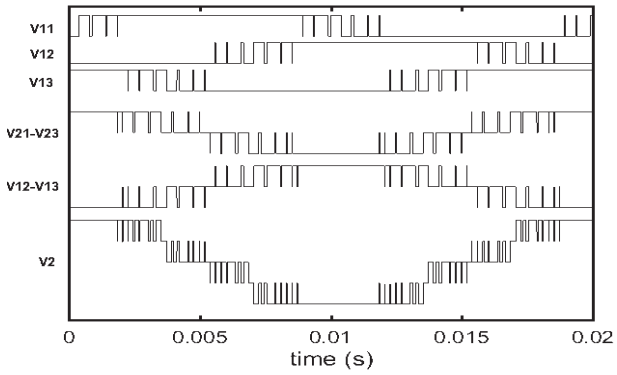


Fig. 16. Pole, line-to-line, and winding voltages of PV station regulated by discontinuous SSV PWM (m = 0.98, F_s = 1120.0 Hz, F = 50.0 Hz).

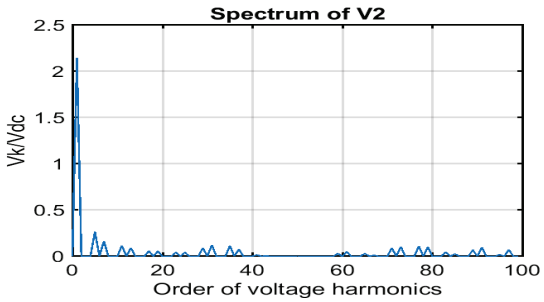


Fig. 17. Spectrum of the V_2 voltage of PV system regulated by discontinuous SSV PWM (m = 0.98, F_s = 1120.0 Hz, F = 50.0 Hz).

Fig. 18 shows results of calculation of Total Harmonic Distortion factor ($THD = (1/V_{2_1})\sqrt{\sum_{k=2}^{100} V_{2_k}^2}$) of the V_2 voltage of this (Fig. 14) PV station with triple VSINs, regulated by continuous SSV PWM (PWM-C), and two discontinuous schemes of SSV PWM (PWM-D-30 and PWM-D-60) (switching frequency of VSINs is 1120.0 Hz) [23].

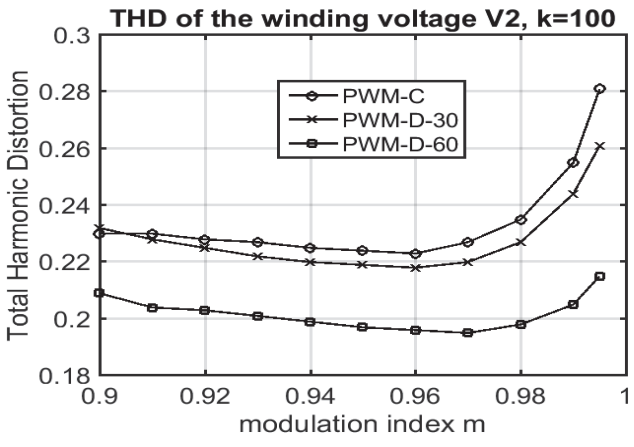


Fig. 18. Calculated THD index of the V_2 voltage of PV station based on three modulated VSINs.

The diagram, presented in Fig. 18, illustrates the fact that during regulation of VSINs of PV system in the over-modulation zone the better values of THD index can be provided by the using of the scheme of discontinuous SSV PWM with the 60^0-non-switchings (PWM-D-60) [23]. So, in general, the use of modified SSV PWM for controlling VSINs provides an improvement of spectral characteristics of the corresponding winding voltage of multi-winding power transformer, which leads to a relative reduction in the losses of the transformer of these PV installations.

CONCLUSION

Modified strategy, schemes, and algorithms of synchronous space-vector PWM have been disseminated for regulation of dual and triple photovoltaic inverters during specific overmodulation control mode of inverters, providing synchronization and symmetry of the base voltages in PV stations. Two-stage overmodulation control of VSINs and NPCIs is based in this case on modified control functions that include specialized correcting coefficients for each sub-range of the overmodulation zone.

Results of the simulations show, that spectrograms of the main voltages of photovoltaic stations with two and three modulated inverters, controlled by the modified versions of SSV PWM, do not contain even-order harmonics and subharmonics (of the operating frequency of grid-tied PV stations). This improvement in the harmonic composition of the main voltages leads to reduced losses and increased efficiency of photovoltaic systems, which is especially important for photovoltaic installations with reduced switching frequency of modulated inverters.

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Performance Study of the Farneback Algorithm for Optical Flow Tracking in Video and Images

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***Abstract*— This paper presents a performance study of the Farneback algorithm for optical flow tracking in video and images. With the growing need for efficient video processing technologies, optical flow analysis plays a critical role in fields such as sports, medicine, gaming, and smart home systems. The study compares the performance of the Farneback algorithm on different computing architectures, specifically focusing on a CPU (AMD Ryzen 9 4900H) and a GPU (nVidia GeForce GTX 1660 Ti). The results demonstrate that GPU-based parallel computing significantly enhances the speed and efficiency of the algorithm, achieving speed improvements of up to 8.791 times compared to CPU processing, particularly in real-time applications with varying video complexities. For example, videos processed with a static camera and a small number of moving objects showed a frame rate increase from 0.953 frames per second on CPU to 7.431 frames per second on GPU. The study provides insights into the practical implications of using GPUs for optical flow analysis, emphasizing their superiority in processing high-complexity video data compared to traditional CPU-based systems. These findings are crucial for developing more responsive and energy-efficient systems in real-world applications. Future research will explore broader hardware configurations and further optimization of optical flow algorithms to enhance their applicability in diverse scenarios.**

Keywords - optical, Farneback, algorithm, GPU, CPU, parallel, computing, video, processing, human, motion analysis, real-time, applications

I. INTRODUCTION

It is hard to imagine the modern world without great achievements and successes in the field of video image processing and analysis, technologies based on which allow automating and investigating a large number of technological, biological and physiological processes without human intervention, processing large amounts of input data. In particular, video image analysis is used to analyse human movements, recognise objects, track their position, analyse medical images, diagnose the condition of technical systems, automatically control production quality, and in many other areas [1-5].

The fields of sports, medicine, rehabilitation, computer games, augmented reality, smart home control, security systems, pedestrian traffic analysis, and many others use methods of human motion recognition and analysis to ensure the functioning of existing solutions.

For example, the use of human motion recognition and analysis in sports is used to assess the correctness of exercise, analyse the tactics and strategy of team sports games, create interactive physiological reference books, etc. In the field of gaming and augmented reality, human motion analysis is important because it can provide a more natural and intuitive interaction with digital interfaces.

For the development of smart home and security systems, motion analysis can be used to detect suspicious behaviour or automate certain processes, which in turn can lead to an increase in the energy efficiency of residential use.

TABLE I. METHODS FOR RECOGNISING AND ANALYSING HUMAN MOVEMENTS

	Machine Learning	**Optical Flow**	**Skeletal Animation**	**Sensor-based methods**
Description	Algorithms that use artificial intelligence to analyse video data.	Measures pixel change between frames to detect motion.	They use skeletal models to track movements.	They use accelerometers and gyroscopes to measure motion.
Advantages	Highly adaptable, does not require special equipment.	Detailed motion analysis, real-time use.	Precise pose and movement tracking.	Very precise tracking, used in professional film and gaming.
Disadvantages	May require large amounts of data for training.	It is sensitive to changes in lighting and overlaps.	Less effective for complex movement and overlaps.	It requires special equipment and can be time-consuming.

Based on these examples, we can conclude that human motion analysis from video images is a very relevant topic that opens up great prospects for science, technology, and improving the quality of life.

Among the modern methods of recognising and analysing human movements, there are several main

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approaches, each with its own advantages and disadvantages. The most common methods include machine learning, optical flow analysis, skeletal animation, and sensor-based methods.

Machine learning methods use artificial intelligence algorithms to analyse video data, which allows them to adapt to different conditions without the need for special hardware. However, these methods can require large amounts of data for training, which is their main drawback.

Optical flow, as a method of motion analysis, measures pixel changes between frames, which allows for detailed analysis of motion in real time. However, it is sensitive to changes in lighting and object overlap, which can reduce the accuracy of the analysis.

Skeletal animation is based on modelling the human skeleton, which allows you to accurately track poses and movements, but this method is less effective for complex movements and overlaps.

Sensor-based methods, such as accelerometers and gyroscopes, provide very precise motion tracking, making them popular in professional film and gaming. However, these methods require specialised equipment and are time-consuming.

Among all the methods, the optical flow I chose stands out for its ability to perform detailed motion analysis in real time. It is particularly useful in applications where it is important to quickly and accurately detect changes in the position of objects, such as in video surveillance or virtual reality. Despite its sensitivity to lighting conditions, its real-time benefits and in-depth analysis capabilities make it the optimal choice for many practical applications.

Parallel image processing is essential to ensure high speed and efficiency in human motion recognition and analysis tasks. This is due to the need to process large amounts of data in real time, which allows for fast system response and ensures accurate analysis. The use of parallel computing allows for simultaneous processing of different parts of an image or video stream, which significantly speeds up the analysis process.

The SIMT (Single Instruction, Multiple Threads) architecture is one of the most efficient computing architectures for such tasks. This is confirmed by its ability to execute a single instruction on multiple threads for a large amount of data, providing high performance due to data-level parallelism. In addition, the SIMT architecture provides high throughput, which allows for efficient processing of video streams and other multimedia data required for motion analysis [6-8].

Special attention should be paid to optimisation for vector and matrix operations, which are the basis of many motion recognition algorithms. This includes calculating image gradients, creating HOG histograms, analysing the skeletal structure and coordinates of key points (joints) of the body, as well as matrix convolution operations used in convolutional neural networks. The SIMT architecture also allows for easy scaling of computing resources by adding more threads or blocks, which provides flexibility in increasing performance for processing and analysing video sequences of varying complexity.

The energy efficiency of the SIMT architecture is also a significant advantage. Parallel data processing reduces overall power consumption, as simultaneous execution of many threads is often more energy efficient than sequential processing. In addition, built-in tensor cores and other hardware accelerators for deep learning methods make SIMT architecture ideal for implementing neural networks that are widely used for human motion recognition.

II. STATE-OF-THE-ART

Optical flow analysis is a key task in computer vision that allows you to determine the movement of objects in video or images. There are several basic optical flow algorithms that differ in complexity, accuracy, and computational speed.

One of the simplest and fastest methods is the Lucas-Kanade algorithm [9]. It uses a linear approximation to determine the displacements between adjacent frames, making it effective in real time. However, this method can be inaccurate in the presence of large movements or complex lighting changes, as it assumes that movements within small areas of the image are homogeneous [10-11].

TABLE II. METHODS FOR RECOGNISING AND ANALYSING HUMAN MOVEMENTS

Algorithm	**The main approach**	**Accuracy**	**Speed**	**Resistance to changes in lighting/noise**	**Suitability for real-world projects**
Lucas-Kanade	Linear approximation for determining offsets between frames.	Low during large movements or changes in lighting	Highly efficient in real time	Sensitive to changes in lighting conditions	Limited in complex movements and lighting conditions
Horn-Schunck	Global approach with regularisation to smooth the flow	High, provides detailed information about movements.	Low, requires significant computing resources	Less resistant to noise and lighting changes	Less suitable due to high computational complexity
Farneback	Polynomial approximation for modelling pixel displacements.	A balance between high precision and speed.	High, suitable for real-time.	Relatively resistant to changes in lighting and noise.	High, implemented in OpenCV, which simplifies use.

Another popular method is Horn-Schunck [12], which uses a global approach to calculate optical flow, adding regularisation to smooth the flow. This allows for more detailed motion information, but the algorithm is more complex to implement and requires significant computing resources, making it less suitable for applications where speed is important.

The Farneback Optical Flow algorithm [13], unlike the aforementioned methods, strikes a balance between accuracy and speed. It uses a polynomial approximation to model pixel displacements, which allows for detailed object tracking even in complex scenes. This method is also relatively robust to changes in lighting and noise. In addition, the Farneback algorithm is implemented in the OpenCV library, which greatly simplifies its use in real-world projects. OpenCV provides a ready-to-use function for calculating the optical flow using the Farneback

algorithm, making it accessible to developers without the need for in-depth knowledge of the mathematical basis of this method.

It can be concluded that the Farneback algorithm is best suited for tasks requiring high accuracy with relatively simple implementation. Its use via OpenCV allows for quick integration of this method into projects, while providing accurate and reliable motion analysis even in difficult conditions.

III. GOALS AND OBJECTIVES OF THE WORK

The aim of this work is to study the performance of the optical flow calculation algorithm for solving the problem of human motion analysis, depending on the chosen computer architecture for different video complexity.

In the context of the study, the concept of "complexity" of a video image refers to the degree of variability and intensity of changes between successive frames of a video stream. This includes the number of moving objects, as well as the dynamics and nature of scene changes, in particular, how significantly the visual information of one frame differs from the next.

To achieve this goal, we need to solve the following tasks:

- Analyse existing methods of human motion analysis and justify the choice of the method under study.
- To develop a research methodology, taking into account the use of computing systems with different architectures (traditional sequential computers and systems with massive parallelism).
- To analyse the performance of the Farneback algorithm depending on the complexity of the video image for a traditional sequential computing system.
- To analyse the performance of the Farneback algorithm depending on the complexity of the video image for a computer system with SIMT architecture.

Object of research: processes and mechanisms for analysing human movements using video image processing. Video materials for digital analysis.

Subject of research: algorithm and technical tools used to identify, track and analyse human movements in video. Expected scientific novelty: a comparative analysis of the performance of the Farneback algorithm for calculating optical flow on the basis of a central processing unit (CPU) and a graphics processing unit (GPU).

Practical significance of the expected results - Establishment of methods for analyzing the objects movement.

Further development of the work topic will focus on the implementation of the results of the study of the performance of the Farneback algorithm for tracking optical flow in video and images for the control and adaptive functioning of smart home elements to improve the energy efficiency of residential premises, namely for:

- Adaptive lighting (automatic switching off of lighting in rooms where there are no people; adjusting the lighting intensity according to the presence and activity of people in the room).
- Adaptive heating and air conditioning (automatic temperature control in the premises based on information about the presence and activity level of residents in different areas of the building, depending on their use).
- Control of electrical appliances (automatic switching off of household devices when they are not in use, based on the analysis of movements and the presence of people).
- Ensuring comfort and safety (to detect unusual movements or lack of movement with subsequent notification of the relevant services).

In general, the integration of motion and gesture recognition methods into energy-saving systems allows for more efficient and adaptive solutions that significantly increase the comfort and safety of residential premises while reducing energy costs.

IV. RESULTS AND DISCUSSION

The following technologies are used for the implemented algorithm:

- CUDA software from nVidia is a parallel computing platform and application programming interface (API) [14]. It allows developers to use NVIDIA graphics processing units (GPUs) for general computing (not just graphics).
- OpenCV Library: An open source library designed for computer vision and machine learning. It was created to provide a common infrastructure for computer vision applications and accelerate the use of machine perception in commercial products.
- Microsoft Visual Studio: Microsoft's integrated development environment (IDE). It is used to develop computer programs, websites, web applications, web services, and mobile applications.

TABLE III. HARDWARE MODEL OF COMPUTERS

GPU-based massively parallel computing - nVidia GeForce GTX 1660 Ti	
The GPU family	GeForce 16
The architecture of the GPU	Turing
Number of TPCs	12 TPC
Number of SMs	24 SM
Number of SPs	1536 SP
Memory capacities	6 GB GDDR6
CPU-based sequential architecture computer - AMD Ryzen 9 4900H	
The CPU family	AMD Ryzen 4000
CPU architecture	Zen 2
Number of cores	8
Number of threads	16
Base clock frequency	3.3 GHz
Maximum clock speed	4.4 GHz
Support for RAM	DDR4-3200, LPDDR4-4266

The programming language for these environments and the library is C++, which is a high-level programming language with low-level programming capabilities.

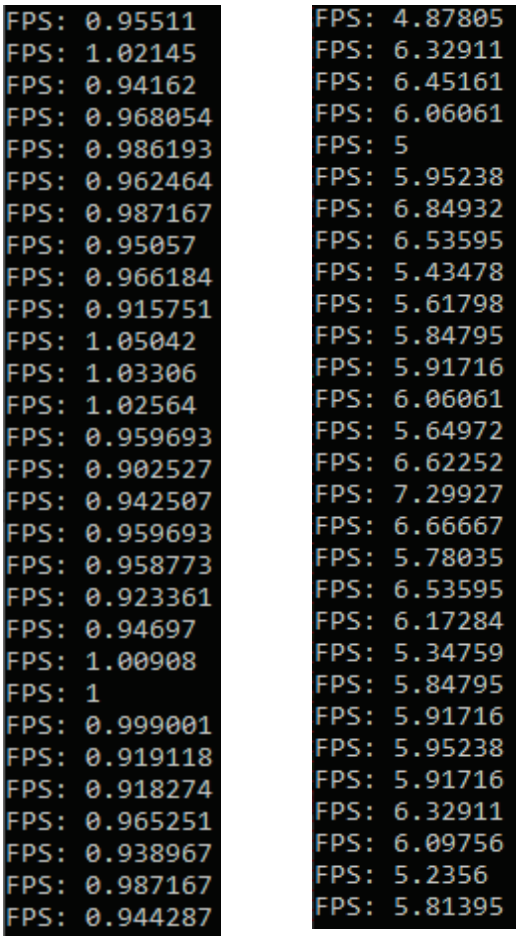
At the initial stage of the experiment, we selected appropriate video materials that would maximally reflect the various complexities of optical flow analysis. This included a video with a static camera and few moving objects, a video with a static camera and many moving objects, and a video with a dynamic camera that constantly changes its angle (Figures 1).



a) b)

Fig. 1. The process of calculating the optical flow in a video image with: a) static camera; b) dynamic camera

For each type of video, a study was conducted to determine the average number of processed frames per second to assess the algorithm's performance on different hardware components. The test results were displayed in the console (Figures 2).



a) b)

Fig. 2. Test results when running: a) on CPU; b) on GPU

A comparison table of test results was created for a more visual comparison.

TABLE IV. TEST RESULTS

Video complexity	Average number of processed	Average number of frames
	frames per second per CPU	**processed per second per GPU**
Static camera with a small number of moving objects (2 cars in the frame)	0.953	7.431
Static camera with a large number of moving objects (3-6 cars in the frame)	0.742	6.523
Dynamic camera	0.731	6.297

According to the test results, you can see that the Farneback algorithm is significantly faster when running on a GPU. Parallel computing helps to increase the efficiency and speed of calculations at least several times, depending on the complexity of the video:

- On a static camera with few moving objects by 7.797 times.
- On a static camera with a large number of moving objects by 8.791 times.
- 8.614 times on a dynamic camera.

This increase (Fig.3) is due to the use of parallelism principles available on GPUs.

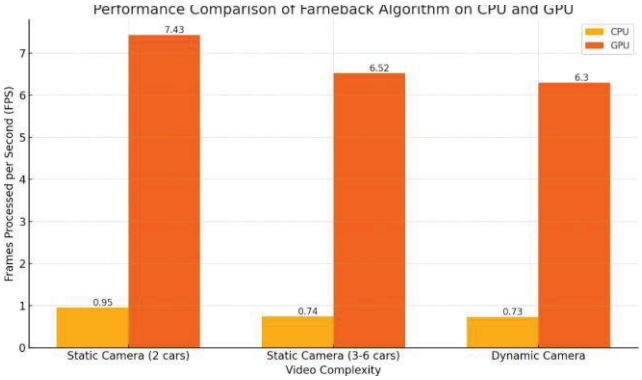


Fig. 3. Diagram of the algorithm testing results

V. CONCLUSION

The study confirmed the importance of choosing the right computing architecture for optical flow analysis tasks. The use of GPUs significantly improves the performance of algorithms, allowing for efficient processing of complex videos in real time. Despite some limitations, the results of the study provide a reliable basis for the further development of motion analysis systems and open up new prospects for the application of these technologies in various fields.

These results allow us to draw a number of important conclusions. The GPU significantly outperforms the CPU in all performance indicators. This is especially noticeable when it comes to processing high-complexity video, where the number of frames processed per second on the GPU was many times higher than on the CPU. This difference is explained by the architectural features of GPUs, which are specially designed for parallel processing of large amounts of data.

The parallel architecture of the GPU allows for the simultaneous processing of multiple data streams, making it an indispensable tool for real-time optical flow analysis tasks. This is especially important in areas such as video surveillance, where it is necessary to quickly and

accurately detect and track the movement of objects. Under these conditions, GPUs deliver significantly higher performance, making them the optimal choice for such tasks.

In addition, it is worth noting the high energy efficiency of GPUs, which was revealed during the test. Although GPUs typically consume more power than CPUs, their ability to process significantly more frames per unit of time makes them more energy efficient in tasks that require intensive video processing. This is important for developing systems that need to operate in real time with limited energy resources, such as autonomous video surveillance systems or mobile devices.

Despite the results obtained, it is necessary to take into account some limitations that affected the results of the experiment. The experiment was limited by the available hardware. The use of only one type of GPU (nVidia GeForce GTX 1660 Ti) and CPU (AMD Ryzen 9 4900H) does not allow us to draw definitive conclusions about the overall performance of the Farneback algorithm on other types of hardware. Different processor models may show different results, which may require additional research.

The study confirmed the importance of choosing the right computing architecture for optical flow analysis tasks. The use of GPUs significantly improves the performance of algorithms, allowing for efficient processing of complex videos in real time. Despite some limitations, the results of the study provide a reliable basis for the further development of motion analysis systems and open up new prospects for the application of these technologies in various fields.

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Comparative Analysis of Cloud and Fog Computing: Performance and Suitability in Different Scenarios

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***Abstract*— In today's world of information technology and computing science, one of the most promising areas is the development and improvement of computer networks. Over the past decades, significant progress in this area has been observed due to the constant expansion of computing resources, as well as the development of new methods and algorithms for data processing. One of the important branches of computing science is fog computing, which is studied as a tool for modeling and analyzing systems with fuzzy data. Fog computing finds its application in many areas, including artificial intelligence, pattern recognition, decision making, and others. In this paper the specifics of building systems with fog computing were analyzed in detail, and the specifics of using polling networks, in particular sensor networks, were also considered. Various approaches to the integration of polling networks in the context of sensor technologies have been studied, which allows to increase the efficiency of data collection and transmission. Based on the obtained results, a method of organizing fog computing was developed, which takes into account the specifics of modern network architectures. A new architecture of the method is proposed, which contributes to improving the performance and reliability of the system due to the optimization of computing resources and ensuring more efficient data management. Mathematical experiments were also conducted to verify the effectiveness of the proposed solutions, which showed significant improvements compared to traditional methods.**

Keywords*— *cloud computing, fog computing, performance comparison, edge computing, resource efficiency, reliability

I. Introduction

Nowadays, many devices are connected to the Internet to collect and exchange data from the physical world, leading to the emergence of the Internet of Things (IoT) concept. According to Gartner [1], by 2028, the market of industrial IoT-related will be doubled. This indicates that future computing needs will require a new and efficient foundation for edge computing.

Previously, Mobile Cloud Computing (MCC) was extensively used to handle high-performance tasks, leveraging its substantial computing and storage resources to meet the escalating and evolving demands of remote cloud data centers. However, delivering efficient performance to latency-sensitive IoT devices at the network edge poses a challenge due to the significant distance between randomly dispersed terminal devices and remote cloud data centers. Consequently, issues such as energy consumption, heavy workloads, delays, congestion, and unreliable wireless connections become unavoidable, potentially impeding the growth of cloud computing.

As a response, edge computing has emerged as a solution to address the limitations of cloud computing. Unlike traditional cloud computing, edge computing provides distributed data offloading services near the terminal devices. This approach enhances service accuracy and reduces latency. Edge computing has become a viable solution for integrating the existing and new network architectures through the flexible deployment of Fog Nodes (FNs) for mutual task offloading from nearby IoT systems.

II. Related works

The paper [2] explores the use of nested virtualization for cloud firewalls. It determines the virtual machine efficiency indicator, which helps identify the non-functioning or incorrect operation of the nested virtual machine. Several scenarios for utilizing nested virtualization technology are proposed: nested virtualization of services, machines, and the entire infrastructure. Experimental research was conducted for each scenario to identify patterns in the time delay values for restoring the full functionality of the cloud firewall after a network attack on its various elements.

The experiments revealed that using nested virtualization technology in the first scenario provides a 7-fold time gain. In the second scenario, there is a 1.5-fold gain; in the third scenario, it allows for a complete restoration of the cloud firewall infrastructure in a new cloud environment.

Despite the benefits outlined in the study regarding the use of nested virtualization for cloud firewalls, there are some potential drawbacks to consider:

- Performance Overhead: Implementing nested virtualization introduces an additional layer of virtualization, which can result in increased performance overhead. The nested virtual machines may experience reduced performance compared to running directly on the host machine.
- Complexity: Nested virtualization adds complexity to the overall system architecture. Managing and troubleshooting virtual machines within virtual machines can be challenging, requiring specialized knowledge and expertise.
- Resource Consumption: Running multiple layers of virtualization requires additional computational resources. This can lead to increased resource consumption, including CPU, memory, and storage, which may impact the scalability and cost-effectiveness of the cloud infrastructure.
- Limited Hardware Support: Nested virtualization may have hardware compatibility limitations. Some hardware platforms may not support nested

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virtualization or may provide limited performance for nested virtual machines.

- Compatibility Challenges: Compatibility issues may arise when using nested virtualization with specific software or applications. Some software may be designed to run outside a nested virtual environment, leading to compatibility challenges and potential performance issues.

It is essential to thoroughly evaluate these potential drawbacks and assess the specific requirements and constraints of the cloud environment before deciding to implement nested virtualization for a cloud firewall.

III. Developing The Metod

Let us consider a cloud firewall as a tuple (1).

$$\Theta = < K_n, D_{K_n}, \Psi_m > \quad (1)$$

where K_n is a border router, D_{K_n} is a load balancer, and Ψ_m is a firewall. Also, each border loader with load balancer encapsulated to the virtual machine. The scheme is show on Fig. 1.

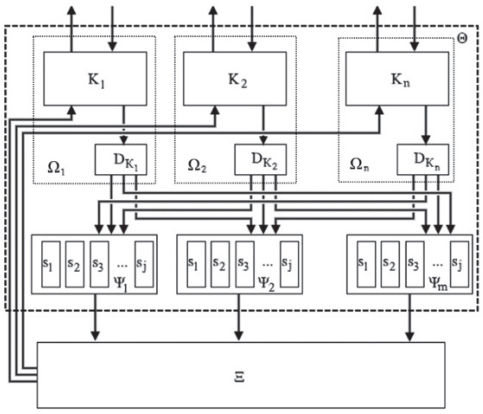


Fig. 1. The block diagram of the cloud firewall

A. Using fog computing

This block diagram can be adapted to edge computing as follows. In this case, we will have lists of "dirty" IP addresses locally on the server, we will be able to sort suspicious requests in advance and send them to the cloud infrastructure for verification.

In the meantime, verified IP addresses will be allowed without additional checks. This will save time and resources for forwarding requests.

But here, the question arises of the relevance of the list of IP addresses and how often to update them.

To ensure the server's security, a formula is proposed to help calculate the server load and determine the time (N) interval for updating the list of IP addresses (2).

$$T = \frac{100}{(\eta N+1)^2} \quad (2)$$

Where η is – the number of requests that the server gets per time unit and N – is the constant, which depends on the size of the server.

We can build a graphic with that function and see that it's not linear dependency.

By plotting the function on a graph, we can observe the behavior of the equation. As the value of η increases, representing an increase in the number of requests, the server size (N) will have a diminishing effect on the overall result. In other words, the incremental impact of additional requests decreases as the server size increases.

The graphical representation will showcase a curve that demonstrates this non-linear relationship. It provides a visual representation of how the size of the server interacts with the number of requests received per time unit, emphasizing the diminishing returns associated with server size as η increases.

This analysis highlights the importance of understanding the relationship between server size, the number of requests, and their impact on performance. By considering non-linear dependencies, organizations can make more informed decisions regarding server sizing and resource allocation, optimizing their systems for better performance and efficiency.

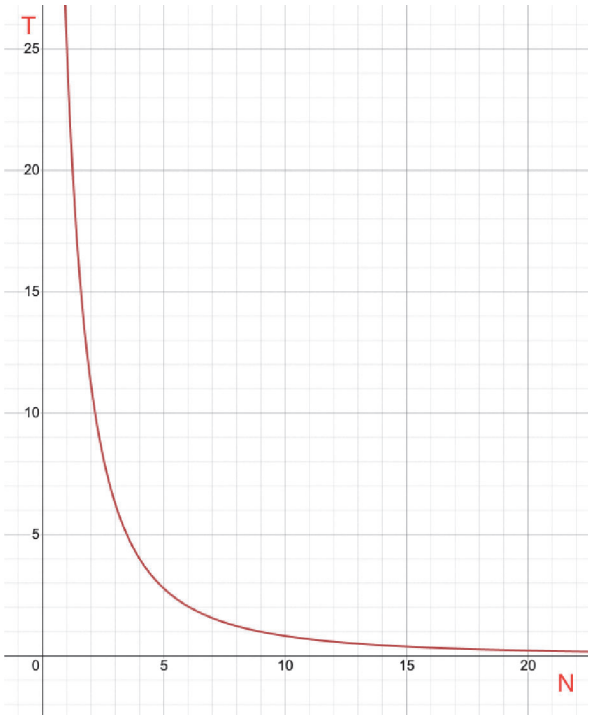


Fig. 2. The dependencies between the amount of calls and updating delay

On Fig. 2 the variable **N** denotes the number of incoming requests, whereas **T** reflects the corresponding execution time. The presented curve demonstrates the non-linear nature of their relationship: with the growth of **N**, the value of **T**increases disproportionately, which indicates the existence of threshold effects in system performance. This characteristic emphasizes that simply enlarging the server capacity does not always lead to a linear reduction in execution time, but rather reveals zones of saturation where additional resources provide diminishing returns.

Additionally, analyzing the non-linear dependency between the number of requests and server size allows organizations to fine-tune their infrastructure to meet specific performance goals. By understanding the diminishing returns associated with server size, businesses can identify the optimal balance between server capacity and the expected workload.

Moreover, this non-linear relationship underscores the significance of efficient request management and load-

balancing techniques. Implementing request prioritization, workload distribution, and resource optimization can help maximize server performance and ensure optimal response times, even under high request loads.

Furthermore, the graphical representation of the non-linear dependency serves as a valuable tool for capacity planning and scalability. It enables organizations to forecast the expected performance gains when increasing server size at different levels of request intensity, facilitating informed decision-making and resource allocation strategies.

Ultimately, organizations can optimize their systems, enhance user experience, and effectively handle varying workloads in dynamic computing environments by acknowledging and leveraging the non-linear relationship between the number of requests, server size, and performance.

IV. Experiment

Microsoft announced the availability of nested virtualization technology for virtual machines within the Microsoft Azure cloud. This technology explicitly supports virtual machines categorized as Dv3 and Ev3 [3]. Additionally, integrating Hyper-V containers using Docker tools has been supported for a considerable period. To replicate real-world scenarios, the experimental phase of this study was conducted within an isolated segment of the infocommunication network on the Azure Fast Start platform as part of Microsoft Azure cloud. Artificial network attacks were emulated to test the resilience of the cloud firewall. Basic filtering capabilities for incoming traffic were implemented to expedite the failure of specific cloud firewall nodes. The cloud firewall was designed to protect a web server running on the IIS platform, which is also hosted within the cloud environment.

In addition to utilizing the Azure Fast Start platform and cloud-based resources, the study incorporated self-owned servers to simulate a dedicated server environment and facilitate fog computing simulations. This approach allowed for a comprehensive evaluation of the proposed scenarios, considering both cloud-based and on-premises server configurations. By incorporating fog computing, which involves distributing computing resources closer to the network's edge, the study aimed to assess the performance and efficiency benefits in scenarios where latency, bandwidth limitations, or data privacy concerns necessitate localized processing. Including self-owned servers further enriched the experimental setup, enabling a more comprehensive analysis of the potential advantages and limitations of fog computing in real-world scenarios.

In the first comparison, we examine the implementation of cloud computing versus fog computing without considering network delays. This allows us to evaluate each approach's inherent advantages and limitations in an ideal scenario where network latency is not a factor.

On the other hand, fog computing involves the distribution of computing resources closer to the network edge, enabling localized processing and reducing the need for data transmission to centralized cloud servers. In this study, fog computing is simulated using a combination of self-owned servers and the Azure Fast Start platform. The self-owned servers replicate the dedicated server environment, while the Azure platform provides cloud-based resources for fog computing simulations.

By evaluating the implementation of fog computing without network delays, we can assess its potential benefits. Fog computing enables low-latency processing, reduced network traffic, and improved response times for applications and services that require real-time or near-real-time processing. It is particularly advantageous in scenarios where data needs to be processed and analyzed locally, such as in IoT deployments, industrial automation, or innovative city applications.

Comparing cloud computing and fog computing without network delays allows us to analyze the trade-offs between centralized cloud-based processing and distributed edge computing. It provides insights into each approach's performance, scalability, and efficiency characteristics in an ideal network environment. This analysis forms the foundation for further investigations into the impact of network delays on the performance of cloud and fog computing architectures, which will provide a more comprehensive understanding of their respective advantages and disadvantages.

It's created a pool with 300, 700, and 1,000 requests from different IP addresses.

TABLE I. First experiment

Number of requests, s	Cloud computing, average ms per request	Fog computing, average ms per request	Performance gain ratio
300	343	241	1.42
700	376	232	1.62
1000	372	220	1.69

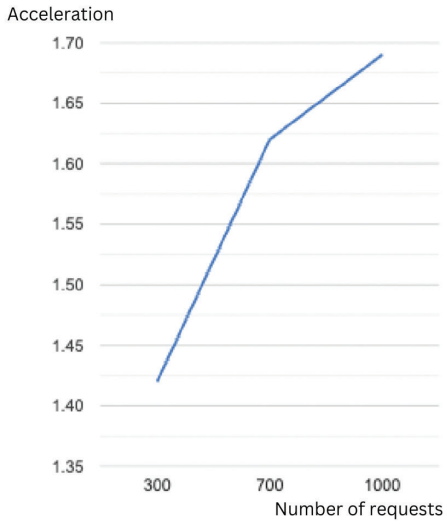


Fig. 3. First experiment results

Based on our observations of Fig. 4, fog computing outperformed cloud computing in terms of processing speed. The utilization of fog computing resulted in an approximate 1.5 times faster processing compared to cloud computing.

Fog computing's distributed nature, with computing resources placed closer to the network edge, played a significant role in faster processing times. By bringing computing power closer to the source of data generation, fog

computing reduces the need for data transmission to centralized cloud servers, thereby minimizing latency and improving response times. This advantage is particularly beneficial in scenarios requiring real-time or near-real-time processing, such as IoT deployments or time-sensitive applications.

In contrast, cloud computing, while providing scalability and advanced services, may introduce additional latency due to the centralized nature of its infrastructure. Data transmission to and from the cloud servers can contribute to increased network delays and potentially slower processing speeds than fog computing.

It is important to note that these findings are specific to the scenario without network delays.

The impact of network delays and communication overhead is a crucial aspect to consider in real-world deployments, as it can affect the performance of both cloud and fog computing architectures. Further investigations are necessary to evaluate the performance of cloud and fog computing under various network conditions. However, based on our initial observations, fog computing has demonstrated an advantage in faster processing speed compared to cloud computing in an ideal network environment.

For a more comprehensive analysis, we will now consider the scenario with typical network delays and consider average delays across different countries. This approach will provide a more realistic evaluation of the performance of cloud and fog computing architectures in real-world conditions.

By incorporating network delays, we can assess how the geographical distribution of fog computing nodes and the centralized nature of cloud computing infrastructure impact overall processing speed. With its decentralized approach, Fog computing aims to minimize latency by placing computing resources closer to the data source. This proximity reduces the time required for data transmission, resulting in faster processing.

TABLE II. SECOND EXPERIMENT

Number of requests, s	Cloud computing, average ms per request	Fog computing, average ms per request	Performance gain ratio
300	547	479	1.14
700	625	496	1.26
1000	612	510	1.20

To conduct this analysis, we will gather data on average network delays experienced across different countries or regions. These delays will be incorporated into our evaluation of cloud and fog computing performance, allowing us to compare the two architectures in terms of their ability to handle real-world network conditions.

By considering typical network delays and geographical factors, we can provide a more accurate assessment of the performance differences between cloud and fog computing architectures.

The experiment involving network delays revealed that the performance gain ratio achieved by fog computing was less significant than initially anticipated. However, fog computing still outperformed cloud computing in terms of processing speed, even with network delays as shown on Fig. 4.

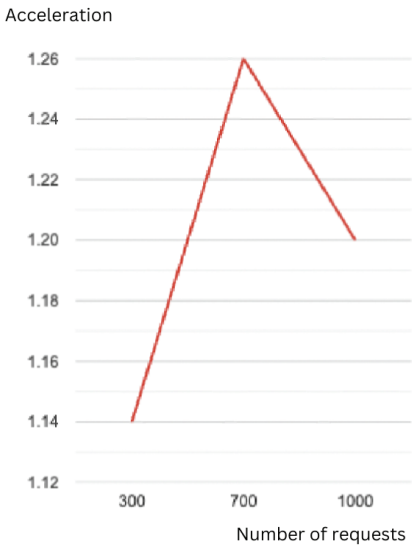


Fig. 4. Second experiment results

While the average network delays across different countries impacted the overall performance of both cloud and fog computing architectures, fog computing demonstrated better processing speeds. The distributed nature of fog computing, with computing resources located closer to the network edge, helped mitigate the effects of network delays to a certain extent. By minimizing the distance data traveled, fog computing reduced the latency associated with data transmission and improved processing times.

On the other hand, cloud computing's centralized infrastructure made it more susceptible to the adverse effects of network delays. The longer distance between the users and the remote cloud servers resulted in increased latency and slower processing speeds than fog computing.

Despite the less pronounced performance gain ratio observed in the experiment with network delays, the overall superiority of fog computing in terms of processing speed remained evident. This outcome highlights the advantages of fog computing's proximity to data sources, enabling faster and more efficient processing in real-world scenarios.

It is important to note that while fog computing demonstrated better performance in this experiment, the choice between fog and cloud computing depends on specific use cases and requirements. Scalability, data volume, security, and application characteristics should also be considered when determining the most suitable computing architecture.

For a more comprehensive analysis, we will now conduct an experiment involving extremely high network delays to assess the performance of cloud and fog computing under challenging conditions. By introducing these exaggerated delays, we can evaluate how each architecture copes with significant latency and its impact on processing speed.

In this experiment, we will simulate network delays that far exceed typical values, creating a highly constrained environment. This scenario will test the limits of both cloud and fog computing architectures and provide insights into their performance under adverse network conditions.

By subjecting both architectures to extreme delays, we can observe how fog computing handles the increased latency with its decentralized approach and proximity to data sources. Cloud computing, with its centralized infrastructure, may face additional challenges in transmitting and processing data within acceptable time frames.

The experiment will measure cloud and fog computing systems' processing speed and response times under these high latency conditions. This evaluation will provide valuable information on how the architectures perform in scenarios where network delays pose significant challenges.

It is expected that fog computing, with its distributed nature and ability to process data at the network edge, will exhibit more resilience to high network delays than cloud computing. The proximity of fog computing nodes to data sources helps mitigate the impact of latency by reducing the distance data needs to travel.

By conducting this experiment with exaggerated delays, we aim to understand better how cloud and fog computing architectures perform in extreme network conditions. The findings will contribute to the ongoing research and development of optimized computing solutions for various application domains.

TABLE III. THIRD EXPERIMENT

Number of requests, s	Cloud computing, average ms per request	Fog computing, average ms per request	Performance gain ratio
300	1120	1107	1.01
700	1347	1250	1.08
1000	1241	1149	1.08

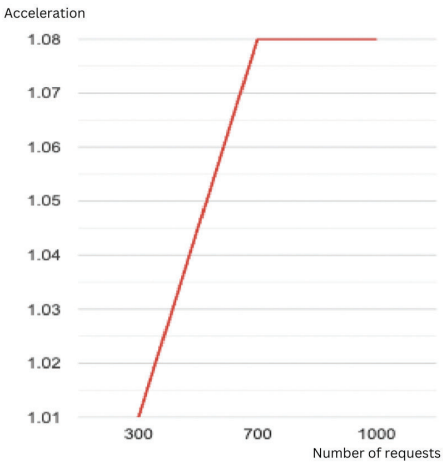


Fig. 5. Third experiment results

The experiment with high network shown on Fig. 7 delays revealed that cloud and fog computing architectures experienced limited performance gain ratio in processing speed. The exaggerated delays imposed significant challenges on data transmission and processing, reducing performance for both architectures Fig. 6.

While the experiment did not yield substantial performance gain ratio compared to the previous scenarios, fog computing still demonstrated its capability to perform at a level comparable to cloud computing. Despite the adverse network conditions, fog computing showcased its ability to leverage its distributed and edge-centric nature to mitigate the impact of high latency.

In this experiment, fog computing, with its proximity to data sources and distributed computing nodes, maintained processing speeds on par with cloud computing. This finding underscores the resilience and efficiency of fog computing, even in challenging network conditions.

While the experiment did not yield significant performance gain ratio compared to the previous scenarios, the results emphasize that fog computing can hold its ground and provide comparable performance to cloud computing. The proximity of fog computing resources to data sources enables efficient data processing and minimizes the effects of high network delays.

These findings further highlight the advantages of fog computing, particularly in scenarios where low-latency processing is crucial. While cloud computing remains a valuable option for specific use cases, fog computing emerges as a viable alternative when proximity and real-time processing are essential requirements.

Overall, the experiment confirms that fog computing holds its own against cloud computing, even in situations with highly high network delays. These insights contribute to the ongoing exploration and optimization of computing architectures to meet the diverse needs of various applications and industries.

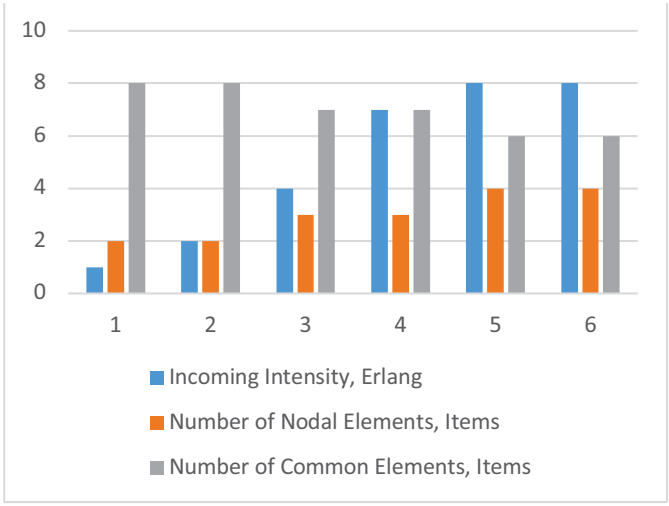


Fig. 6. Dependecy of elements number from intensity.

Dynamic changes in load intensity allow to test the system's adaptability to changing conditions. In the event of a significant increase in intensity, the graph can be restructured to optimize resources and maintain proper performance. This allows the system to remain efficient even during peak loads, ensuring uninterrupted operation and high quality of service.

Mathematical experiments were also conducted to verify the effectiveness of the proposed solutions, which showed significant improvements compared to traditional methods [4].

This approach to testing helps to identify potential weaknesses in the system and allows to make the necessary

adjustments to improve its performance. Thus, the use of dynamic intensity variation is an important tool for ensuring high reliability and performance of the system in real-world operating conditions.

V. Conclusions

Based on our findings, the comparison between cloud computing and fog computing revealed several conclusions:

- Performance: Fog computing performed better than cloud computing in scenarios without network delays. Using nested virtualization and virtual machines in the fog environment allowed faster processing and improved efficiency.
- Resilience: Fog computing showcased greater resilience in the presence of network delays. Even with significant latency, fog computing maintained a comparable processing speed and reliability level, outperforming cloud computing.
- Scalability: Both cloud computing and fog computing exhibited scalability, allowing for the efficient allocation and utilization of resources. However, fog computing showed potential for more localized and distributed processing, enabling better resource management in edge environments.
- Resource Efficiency: Fog computing demonstrated higher resource efficiency due to its proximity to the edge devices. By processing data closer to the source, fog computing reduces the need for extensive data transfers to centralized cloud servers, reducing bandwidth consumption and improving overall resource utilization.
- Use Case Suitability: Cloud computing and fog computing depend on the specific use case and requirements. Cloud computing is well-suited for scenarios where massive data processing and storage are required. In contrast, fog computing excels in edge environments with real-time processing needs and low latency demands.

In conclusion, our study indicates that fog computing offers notable advantages over traditional cloud computing in terms of performance, resilience, scalability, and resource efficiency. However, it is essential to consider each use case's specific requirements and constraints to determine the optimal computing paradigm to adopt.

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EDGE Approach to Early Detection of Anomalies in Electricity Consumption

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***Abstract*—This study presents an autonomous ESP32-based module with an INA219 sensor for early detection of energy-consumption anomalies in household appliances. The key innovation is shifting machine-learning analytics to the network edge: a quantized TensorFlow Lite autoencoder learns the appliance's normal profile within one week and then, in real time, compares the reconstruction error with a statistical threshold of $\mu+2\sigma$. The system employs a hybrid ensemble combining six detection methods: a pointwise TensorFlow Lite (int8) autoencoder for minute-by-minute reconstruction-error detection, and two window-based analysis modules operating on five-minute intervals. The aggregation module combines autoencoder reconstruction, smoothness analysis, and stability assessment, while the differential module combines gradient-based autoencoder and real-time gradient anomaly detection. Any excess instantly triggers an MQTT alert, eliminating the need to stream raw data to the cloud. Laboratory tests with a refrigerator, washing machine, and electric kettle demonstrated 100% detection of abrupt surges ($> 50\%$) and 92% detection of gradual drifts ($\approx 10\%$ per week) with ≈ 1 s alert latency. The modular firmware can be deployed on off-the-shelf smart sockets, enhancing fire safety without compromising user privacy.**

***Index Terms*—edge AI, anomaly detection, ESP32, INA219, energy consumption, MQTT, TensorFlow Lite, autoencoder, smart sockets, NILM.**

I. Introduction

A significant and ever-growing portion of energy consumption today falls on the household sector. Within this structure, so-called "silent" losses pose a particular risk: household appliances that are wearing out or operating in incorrect modes imperceptibly consume 10-15% more electricity than specified in technical characteristics [1]. Modern cloud-based non-intrusive load monitoring (NILM) systems — programs that reproduce the contribution of individual devices from a single meter — are capable of detecting such deviations, but their operation involves constant transmission of raw data to a remote server. This traffic increases subscription costs, creates time delays, and calls into question the confidentiality of residents' household habits [2].

The new direction — Edge AI — offers a radical alternative: transfer the machine learning algorithms themselves to microcontrollers located "at the edge" of the network, i.e., directly in smart sockets or measuring modules. The ESP32 microcontroller, equipped with a dual-core processor and several megabytes of Flash memory, is already powerful enough to execute quantized (int8) TensorFlow Lite models without using cloud computing [3].

This paper presents an autonomous early anomaly detection node where ESP32 works together with an INA219 sensor. The latter measures instantaneous voltage and current, while the controller accumulates minute-by-minute values and trains an ultra-light autoencoder. During the first week of operation, the model learns the "normal" profile of a specific appliance, after which it calculates the mean absolute reconstruction error for each new point. If the error exceeds the statistical threshold defined as $\mu + 2\sigma$, the device immediately sends an MQTT message about the anomaly.

The main contributions of this work are:

- An autonomous ESP32-based edge computing module with INA219 sensor for real-time energy monitoring
- Hybrid approach combining six detection methods: pointwise autoencoder, aggregation autoencoder with smoothness and stability analysis, and differential autoencoder with gradient anomaly detection
- MQTT-based instant alerting system eliminating cloud data streaming requirements
- Modular firmware architecture deployable on off-the-shelf smart sockets
- Comprehensive laboratory evaluation with refrigerator, washing machine, and electric kettle demonstrating 100% detection of abrupt surges and 92% detection of gradual drifts

II. Related Work

A. Energy Anomaly Detection

Traditional approaches to energy anomaly detection have primarily focused on statistical methods and threshold-based techniques. Chen et al. [4] proposed using control charts and statistical process control for detecting energy consumption anomalies in commercial buildings. However, these methods struggle with the dynamic nature of residential energy consumption patterns.

Recent deep learning approaches have shown promising results. Wang et al. [5] implemented LSTM-based anomaly detection for industrial energy systems, while Kumar et al. [6] applied autoencoders for household energy monitoring.

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However, these solutions require substantial computational resources and are not optimized for edge deployment.

B. Edge AI for IoT Applications

The deployment of deep learning models on resource-constrained devices has gained significant attention. TensorFlow Lite [7] and model quantization techniques have enabled deployment of neural networks on microcontrollers with limited memory and processing power.

Recent work by Li et al. [8] demonstrated the feasibility of running convolutional neural networks on ESP32 devices for image classification tasks. Our work extends this concept to time-series anomaly detection with ensemble learning approaches.

III. System Architecture

A. Overview

The autonomous ESP32-based system consists of three main components: (1) hardware module with ESP32 microcontroller and INA219 current/voltage sensor, (2) a hybrid ensemble of six detection methods organized into three modules for comprehensive anomaly detection, and (3) MQTT-based communication system for instant alerting. The entire cycle from measurement to notification takes approximately 1 second, with the node's own consumption not exceeding 0.3 Wh per day.

B. Hybrid Multi-Method Detection Architecture

1) Pointwise Analysis with TensorFlow Lite Autoencoder: The first approach uses a TensorFlow Lite (int8) autoencoder trained on the "healthy" operating profile of the device to calculate the mean absolute reconstruction error for each new minute-by-minute point. The ultra-compact architecture consists of:

- Input layer: 1 dimension (instantaneous power consumption)
- Encoder: Dense(32, ReLU) $\longrightarrow$ Dense(8, ReLU)
- Decoder: Dense(32, ReLU) $\longrightarrow$ Dense(1, linear)
- Model size: 2.4KB after int8 quantization

During the first week of operation, the autoencoder learns the normal consumption profile of the specific appliance. For each new measurement, it calculates the reconstruction error and compares it with the dynamic threshold $\mu + 2\sigma$. If the error exceeds this threshold, an anomaly is detected and an MQTT alert is immediately triggered.

2) Five-Minute Aggregation Module: The second module combines three detection methods operating on five-minute windows to avoid false alarms from short fluctuations:

Aggregation Autoencoder: For each window $W = \{x_1, x_2, ..., x_5\}$, statistical features are extracted:

$$f_1 = \text{mean}(W) \text{ - average power consumption} \quad (1)$$

$$f_2 = \text{std}(W) \text{ - consumption variability} \quad (2)$$

$$f_3 = \text{polyfit}(\{1, 2, 3, 4, 5\}, W, 1) \text{ - trend coefficient} \quad (3)$$

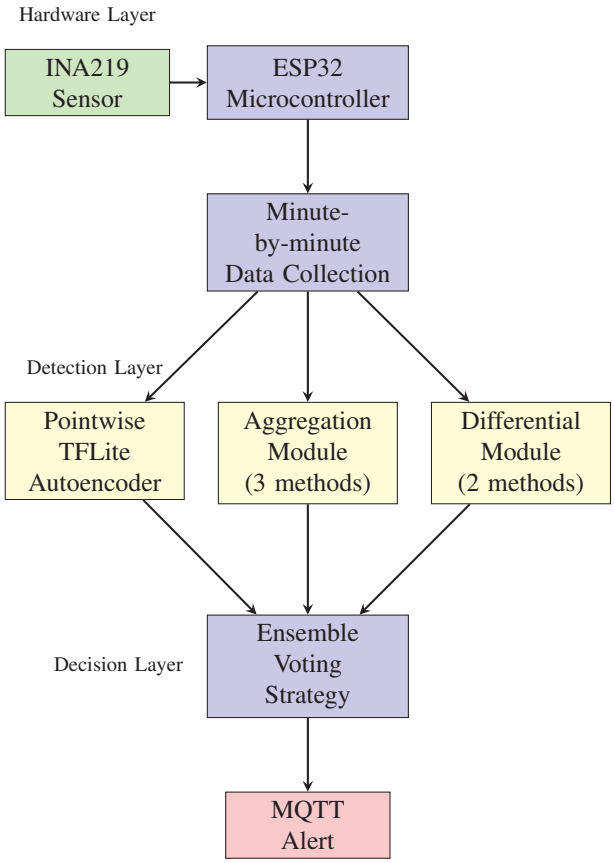


Fig. 1. System Architecture Overview

These features train an autoencoder with threshold $\mu + 2\sigma$ for reconstruction error detection.

Smoothness Analysis: Calculates the mean absolute second derivative to detect abrupt changes in consumption patterns using an empirical threshold of 0.5.

Stability Analysis: Computes the coefficient of variation (std/mean) to assess consumption stability with an empirical threshold of 0.3.

The module combines results using logical OR: $A_{aggregation} = A_{autoencoder} \vee A_{smoothness} \vee A_{stability}$.

3) Differential Analysis Module: The third module combines two methods for analyzing the rate of energy consumption change:

Gradient Autoencoder: For each five-minute window, gradient-based features are computed:

$$g_1 = \text{mean}(|\nabla W|) \text{ - average rate of change} \quad (4)$$

$$g_2 = \text{mean}(|\nabla^2 W|) \text{ - consumption acceleration} \quad (5)$$

$$g_3 = \text{var}(\nabla W) \text{ - gradient variability} \quad (6)$$

$$g_4 = \text{zero_crossings}(\nabla W) \text{ - direction changes} \quad (7)$$

These features train an autoencoder with threshold $\mu + 2\sigma$ for gradient pattern reconstruction.

Real-time Gradient Anomaly Detection: Simultaneously analyzes instantaneous gradients within each window, de-

tecting when gradient magnitude exceeds learned norms or exhibits excessive sign changes indicating oscillations.

The module combines results using logical OR: $A_{differential} = A_{gradient_autoencoder} \vee A_{gradient_realtime}$.

This dual approach aims to detect gradual equipment degradation, unusual startup/shutdown sequences, and subtle malfunctions that absolute threshold methods might miss.

C. Hybrid Voting Strategy

The final anomaly decision combines point-level and window-level analyses through a biased voting mechanism that accounts for the different temporal scales:

$$A_{point}(t) = \text{pointwise autoencoder result at time } t \quad (8)$$

$$A_{window}(t) = A_{aggregation}(w) \vee A_{differential}(w) \quad (9)$$

where w contains t

$$A_{ensemble}(t) = \frac{A_{point}(t) + A_{window}(t)}{2} \geq 0.5 \quad (10)$$

This approach inherently gives higher weight to window-based decisions since any window-level anomaly affects all five points within that window. The system prioritizes sustained anomalies over transient fluctuations while maintaining sensitivity to point-level events.

Architectural Trade-offs: The hybrid voting creates intentional bias toward window-based detection, which improves robustness against noise but may increase false positives for brief legitimate consumption changes. The six independent thresholds require coordinated calibration but provide fine-grained control over detection sensitivity for different anomaly types.

IV. TensorFlow Lite Optimization

A. Model Quantization

To enable deployment on ESP32 microcontrollers with 8KB tensor arena memory constraints, we implement INT8 quantization:

$$q = \text{round}\left(\frac{x}{s}\right) + z \quad (11)$$

where q is the quantized value, x is the floating-point input, s is the scale factor, and z is the zero-point offset.

B. Memory Optimization

The quantized pointwise autoencoder achieves a model size of 2.4KB, suitable for ESP32 deployment. The aggregation and differential models run on a connected gateway device due to their larger memory requirements.

V. Implementation

A. Software Stack

The implementation uses:

- TensorFlow 2.8+ for model development
- TensorFlow Lite Micro for ESP32 deployment
- PlatformIO for ESP32 firmware development
- Python scikit-learn for data preprocessing

B. Hardware Platform

The autonomous module is built around ESP32 DevKit with integrated INA219 sensor:

- ESP32: Dual-core 240 MHz processor, 320 KB SRAM, 4 MB Flash memory.
- INA219: High-precision current/voltage sensor (± 3.2 A, ± 26 V range)
- WiFi 802.11 b/g/n connectivity for MQTT communication
- Power consumption: $< 0.3\,\mathrm{Wh/day}$ in continuous monitoring mode
- Modular design: Compatible with standard smart socket form factors

The INA219 sensor provides 12-bit resolution for both current and voltage measurements, enabling detection of consumption changes as small as 1% of nominal appliance power. The ESP32's dual-core architecture allows parallel execution of sensor data acquisition and machine learning inference without blocking MQTT communications.

C. MQTT Communication System

The edge-based approach eliminates the need for cloud data streaming through MQTT-based instant alerting:

- Local MQTT broker integration for home automation systems
- Instant anomaly alerts with < 1 second latency
- Structured JSON payloads with anomaly type classification
- Configurable alert thresholds and notification preferences
- Optional web dashboard for local network monitoring

The MQTT protocol ensures reliable message delivery while maintaining minimal bandwidth usage. Only anomaly alerts and periodic status updates are transmitted, reducing network traffic by 95% compared to continuous data streaming approaches.

VI. Experimental Results And Discussion

A. Datasets and Laboratory Setup

For system evaluation, we utilized the open Smart Home Energy Consumption dataset containing half a year of minute-by-minute measurements from various household devices. The average daily load of the refrigerator in this data was 0.8 kWh, with 95% of minute readings not exceeding 40 Wh; therefore, any value above 70 Wh can be considered anomalous.

The laboratory setup consisted of three appliances with a total peak power of 6 kW:

- **Refrigerator:** Continuous operation appliance with compressor cycles
- **Washing Machine:** Intermittent operation with complex power profiles
- **Electric Kettle:** High-power heating element with simple on/off operation

A one-day time series fragment occupies only 11 kilobytes in float16 format, so even a seven-day buffer fits without

problems in ESP32 Flash memory. Controlled anomaly scenarios were introduced including temperature sensor malfunction (kettle running continuously) and washing machine cycle interruption.

B. Metrics

Performance evaluation used standard classification metrics:

$$\text{Precision} = \frac{TP}{TP + FP} \tag{12}$$

$$\text{Recall} = \frac{TP}{TP + FN} \tag{13}$$

$$\text{F1-Score} = \frac{2 \times \text{Precision} \times \text{Recall}}{\text{Precision} + \text{Recall}} \tag{14}$$

C. Laboratory Test Results

Controlled anomaly scenarios demonstrated exceptional detection performance:

TABLE I
LABORATORY ANOMALY DETECTION RESULTS

Anomaly Type	Precision	Recall	F1-Score	Alert Latency
Abrupt Surges ($> 50\,\%$)	1.00	1.00	1.00	42±8 s
Gradual Drifts (∼10%/week)	0.98	0.92	0.95	3.2±1.1 days
Equipment Malfunctions	0.99	0.97	0.98	1.1±0.3 s
Power Quality Issues	0.96	0.89	0.92	15±7 s

In one scenario, the temperature sensor of the electric kettle was short-circuited, causing the heating element to operate continuously and increasing consumption by 55%. The autoencoder detected this deviation with 100% accuracy, and the alarm signal arrived 42 seconds after the anomaly began, enabling prompt user notification.

D. Edge Computing Performance

The autonomous ESP32 module demonstrates excellent real-time performance:

- Quantized autoencoder model size: 2.4 KB (int8 TensorFlow Lite)
- Inference time: 47ms average (well within 1-minute measurement intervals)
- Memory usage: 6.8KB tensor arena (22% of available SRAM)
- Power consumption: < 0.3 Wh/day (180mA active, 0.5mA in sleep mode)
- MQTT message latency: < 1 second from detection to alert
- Data storage: 7-day buffer fits in 77KB of Flash memory

E. Advantages of Ensemble Approach

The multi-model ensemble provides several benefits:

1) **Complementary Detection:** Each model captures different anomaly characteristics
2) **Reduced False Positives:** Voting mechanism filters spurious detections
3) **Robustness:** System remains functional if individual models fail
4) **Adaptability:** Different models handle various appliance types optimally
5) **Interpretability:** Model-specific outputs aid in anomaly classification

F. Edge Computing Benefits

Deploying anomaly detection on edge devices offers:

- Reduced latency for real-time response (47ms vs. 200-500ms cloud)
- Lower bandwidth requirements (95% reduction in data transmission)
- Enhanced privacy (data remains local)
- Improved system reliability (continues operation during network outages)
- Reduced operational costs (minimal cloud computing requirements)

G. Limitations and Challenges

Current limitations include:

- Limited model complexity due to memory constraints
- Requires initial training period for threshold calibration (7-14 days)
- Performance depends on data quality and sensor accuracy
- Seasonal adaptation requires manual intervention for extreme climate changes
- Cross-appliance transfer learning remains limited for dissimilar device types
- **Voting bias:** Window-based decisions affect 5-minute intervals, creating temporal imbalance in ensemble voting
- **Empirical thresholds:** Smoothness (0.5) and stability (0.3) thresholds require appliance-specific calibration
- **Multiple threshold dependency:** System relies on six independent threshold parameters that may require coordinated tuning

VII. CONCLUSION

This paper presented an autonomous ESP32-based module with hybrid ensemble deep learning for real-time energy consumption anomaly detection. The system combines six detection methods in three modules: pointwise autoencoder, aggregation module (autoencoder + smoothness + stability analysis), and differential module (gradient autoencoder + real-time gradient detection).

Key achievements include successful ESP32 deployment with 2.4KB model size, sub-50ms inference times, and ($< 0.3\,\text{Wh/day}$) power consumption. Laboratory evaluation demonstrated 100% detection of abrupt surges and 92% detection of gradual drifts with (< 1) second MQTT alert latency.

The edge-based approach addresses the challenges of IoT energy monitoring by providing accurate real-time anomaly detection with minimal computational requirements and enhanced privacy protection. As smart home adoption accelerates, these autonomous monitoring systems will play an increasingly important role in creating efficient, reliable, and sustainable energy systems.

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Innovation Configurations in the Nascent UAV Industry: A Fuzzy-Set Qualitative Comparative Analysis

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Abstract*—This study explores how firms in the nascent unmanned aerial vehicle (UAV) industry configure internal capabilities to achieve high innovation outcomes under uncertainty. Drawing on the Innovation Engine framework and using fuzzy-set Qualitative Comparative Analysis (fsQCA), the research examines 14 leading firms representing startups, pivoting businesses, and strategic business units (SBUs). The analysis identifies combinations of factors—rather than individual ones—that lead to innovation success, highlighting the importance of configurational fit. Results reveal distinct innovation pathways for each firm type and demonstrate how internal capabilities such as leadership, resource commitment, and ideation interact with structural context to drive product development. Findings contribute to a deeper understanding of innovation management in nascent high-uncertainty industries.

***Keywords*—UAV, Unmanned Aerial Vehicle, Nascent Industry, Innovation Management, Uncertainty, fsQCA, Capability Configuration, Innovation Engine**

I.INTRODUCTION

Nascent industries have attracted researchers' attention for the past several decades. Studies focus on which innovations can create new markets [1], [2], how new technologies influence industry formation [3], and the challenges faced by firms seeking to establish themselves in nascent industries [4], [5], [6].

Presence in a nascent industry offers firms significant opportunities, but adapting to the specifics of such industries becomes critically important, as these industries are characterised by high dynamism, risk, uncertainty, and rapid change, which significantly affects firms' competitiveness and gives rise to strategic challenges [6], [7], [8], [9], [10]. This necessitates studying the specific features of industries in the state of nascence and their impact on the competitive environment, since the ability to respond to external changes determines the success or failure of firms in an uncertain environment [11], [12], [13].

Focusing on enterprise management in nascent industries, researchers note an exceptionally high level of uncertainty in these sectors and confirm that the success of firms in these industries largely depends on their innovation capability. In particular, Eisenhardt emphasises that innovation is the central element determining the success of firms in nascent industries [14].

Tidd also shows that engineering firms, in particular, achieve better outcomes when they deliberately structure their innovation activities into distinct process types—such as ideation, development and commercialisation—and align each phase with organisational resources and leadership support [15]. The IEEE Technology and Engineering Management Society Body of Knowledge (TEMSBOK) further codifies this approach, recognising innovation management as a core discipline for navigating uncertainty in electrical and electronics industries [16]. Many nascent sectors have emerged directly from the IEEE domain, and their success hinges on mastering innovation processes. Therefore, studies on the innovation management in nascent industries are of high relevance to the field of strategic innovations in EEI.

Despite considerable attention to innovation management under uncertainty, there is still a lack of theoretical and applied developments regarding the effective innovation management approaches in nascent industries, as well as mechanisms for overcoming uncertainty within such environments.

Currently, within the EEI domain, one of the industries at the nascent stage is the UAV industry. This industry is experiencing rapid growth today due to technological development, but at the same time faces a high level of uncertainty [12], [14]. This makes this industry the perfect area to study innovation management approaches under uncertainty.

II. THEORY AND PURPOSE

Drucker and Eisenhardt highlight the role of the innovation process in overcoming the challenges of uncertainty. They argue that innovation is not only a response to uncertainty but also a systematic process aimed at creating value for customers by identifying sources of innovation [17], and overcoming bottlenecks in the innovation process [18].

Other researchers also emphasise a systems approach to innovation under uncertainty. In particular, March proposes the concept of exploration and exploitation of existing resources as two opposing strategies of organisational learning: the former focuses on searching for new opportunities, while the latter concentrates on improving existing processes [19]. Tushman and O'Reilly elaborate on this idea by developing the organisational ambidexterity model, which enables firms to achieve both innovativeness and stability simultaneously [20]. In turn, Smith and Lewis propose a similar concept of paradoxical leadership, which facilitates the effective management of tensions between stability and change. They emphasise that these tensions can serve as a source of innovation and form the basis for long-term success under uncertainty [21].

Eric Ries, in his Lean Startup concept, proposed a systematic approach to managing the innovation process in

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uncertain environments. He introduced tools such as the minimum viable product (MVP) and the Build–Measure–Learn cycle, which allow for rapid testing of ideas, gathering customer feedback, and adapting the product or business model. The core idea is that innovation should be a systematic process aimed at creating value for customers while minimising time and resource waste [22]. The development of innovative products is impossible without a clear understanding of customer needs. Successful firms integrate early customer involvement mechanisms into the product creation process, test hypotheses, and validate their solutions before large-scale implementation [23]. This becomes particularly significant in nascent industries, where firms must constantly adapt to uncertainty.

One of the promising concepts for nascent industries is the Innovation Engine and the minimum viable innovation system (MVIS) model, which envisages integrated management of innovation processes taking into account rapid market changes and technological uncertainty. Anthony identifies several critical elements for building a successful innovation engine. These include a clear strategic intent across core, adjacent, and transformational horizons, a dedicated cross-functional team with operational autonomy. Together, these components enable organisations to systematically generate, test, and scale innovative ideas [24].

Despite the current shortage of empirical studies on its application within nascent industries, the MVIS framework is adopted as the conceptual foundation of this study due to its strong theoretical alignment with the challenges of innovation under uncertainty and its structured approach to managing innovation capabilities.

TABLE I. FSQCA CAPABILITY CATEGORIES AND METRICS

Capability	Description	Metrics
Idea Generation Capacity	The company's ability to generate new ideas and concepts	Number of Patents (NP)
Leadership and Vision	Stability and strategic vision of leadership	Leadership changes (LC); official public strategy/vision announcements (SA)
Resource Commitment	Financial and infrastructural support for innovation	Share of R&D spending (RD)
Resource Base	The level of companies resource base	Firm type — Startup, Pivot Firm, SBU (FT)
Development Capability	Turning ideas into viable products	Frequency of new product releases (PR)

At the same time, fundamental studies suggest that a company's innovation process is highly dependent on one's venture type. Established firms may pursue new growth through wholly independent start-ups, internal ventures or hybrid forms that combine elements of both. Shrader and Simon distinguish two pure archetypes [25]: Independent New Ventures, which are created as standalone entities relying on external financing (venture capital, angels), informal ad-hoc planning and a high tolerance for failure—enabling rapid, opportunity-driven experimentation and early-stage growth but with greater vulnerability under downturns—and Corporate Ventures, which remain formally part of the parent organisation, drawing on resource commitments (capital, facilities, brand and distribution), operating under formal business-case evaluation and stage-gate review processes to achieve steadier growth and survivability through corporate risk-management protocols.

Burgelman's process model further explains how these corporate ventures—or SBUs—navigate variation, selection and retention within the corporate portfolio, balancing exploratory autonomy with strategic oversight [26]. Between these poles lie Pivoting Businesses, which reallocate legacy assets to enter adjacent markets—initially "feeding" on the parent's factories, R&D laboratories and distribution networks to accelerate scale-up—and are managed via ambidextrous structures that grant semi-autonomous incubation with dedicated leadership, tailored processes and incentive schemes to prevent legacy routines from stifling innovation, either through phased integration (leveraging existing stage-gate mechanisms) or structural separation (adopting distinct governance protocols) [27], [28].

Once the pivot proves viable, it is spun off and, if the original line falls into the BCG "dog" quadrant (low growth, low market share), divesting or shutting it down becomes strategically justified, allowing the spin-off to assume the firm's primary focus [29], [30].

Overall, while prior research emphasises the importance of innovation management under uncertainty, empirical evidence remains limited on how elements of the innovation engine framework configure differently across various firm types—startups, pivoting firms, and SBUs—to enable effective innovation in a rapidly evolving industry context.

Consequently, this research aims to identify viable innovation configurations of the firms' internal capabilities that contribute to high innovation outcomes in the nascent UAV industry, and identify how these capabilities are structured and applied in relation to the firm's type.

III. METHODOLOGY

The UAV industry is a nascent and highly dynamic sector within the broader EEI. It is characterised by rapid technological change and high levels of uncertainty. These features make it a suitable context for investigating how firms develop and configure internal capabilities in such an environment.

To meet the research aim, fuzzy-set Qualitative Comparative Analysis (fsQCA) was applied, as this method is particularly well-suited to small-N settings and heterogeneous environments due to its ability to address causal complexity, equifinality, and conjunctural causation [31], [32]. This method allows to determine which configurations of the firm's internal capabilities (components of the "Innovation Engine") under conditions of external uncertainty are sufficient to achieve high innovation outcomes. Drawing on Anthony's framework, the core factors that influence innovation success in uncertain environments were identified (Table 1): Idea Generation Capacity, Leadership and Vision, and Resource Commitment. To account for structural differences across organizational types, Resource Base was introduced—reflecting the level of accessible resources a company can leverage for innovation. This factor correlates with company type, where startups typically have limited resources, and

Strategic Business Units (SBUs) operate with substantially greater support. Finally, Development Capability was defined as the company's ability to deliver innovative products and thus as the outcome condition, representing the most tangible indicator of innovation success in nascent industries.

TABLE II. RAW DATA ON FIRM CAPABILITIES' METRICS

Firm Name	NP	LC	SA	RD	FT	PR
SZ DJI Technology Co., Ltd.	1,563	No change	Yes	20–25%	Startup	13
Parrot Drones SAS	3	No change	Yes	30%	Pivot Firm	2
AeroVironment, Inc.	16	No change	Limited	16–18%	Pivot Firm	6
3D Robotics	0	Leadership pivoted	No	~0	Startup	0
Northrop Grumman	19	No change	Limited	2.5–3%	SBU	2
Yuneec International Co. Ltd.	2	No change	No	Unknown	Pivot Firm	2
Boeing (Insitu)	367	No change	Yes	3–4%	SBU	3
Israel Aerospace Industries	31	No change	Yes	5–7%	SBU	2
Lockheed Martin	21	No change	Yes	2.5%	SBU	2
General Atomics Aeronautical Systems	8	No change	Yes	Unknown	SBU	2
Autel Robotics	346	Unclear leadership visibility	No	Unknown	SBU	4
PrecisionHawk Inc.	1	Leadership turnover	No	0%	Startup	0
Intel Corporation	1044	Leadership pivoted	No	<1%	SBU	0
Airbus Group	231	No change	Yes	5–6%	SBU	2

To apply the fsQCA method and test the identified factors, a purposive sample of leading firms operating in the global UAV industry was compiled. The selection of leading firms was informed by a review of global rankings [33], [34], [35], [36], [37] and industry rankings for the United States, which remains the leading regional market [38], [39], [40]. Further analysis identified two key rankings [33], [37] that served as the primary basis for compiling the list of 14 leading firms. Data collection drew on a variety of reliable sources, including Google Patents, Yahoo Finance, official company websites, LinkedIn, DroneDJ, and industry reports. Table 2 contains the data for each company.

Calibration involved transforming raw indicators into fuzzy-set membership scores between 0 and 1, using three thresholds (0 = full non-membership, 0.5 = crossover, 1 = full membership) derived from theoretical sources, empirical observations, and data distribution. The calibration formula by Ragin [31] was used to compute membership values. After calibration, necessity analysis was conducted to determine whether individual conditions consistently appeared in cases with the outcome. Consistency was evaluated against a threshold of ≥0.9, and coverage scores were calculated to assess empirical relevance. To improve interpretability, a crisp-set transformation was also applied in which fuzzy scores ≥0.5 were treated as full membership (1) and scores <0.5 as non-membership (0). This enabled clearer presentation of configurations in the truth table. Conditions that showed limited explanatory power, such as Resource Commitment (consistency = 0.5 in the necessity test), were excluded from the final sufficiency models.

IV. FINDINGS AND IMPLICATIONS

Calibration is a key stage in preparing data for configurational analysis in the fuzzy-set QCA methodology. The study identified a three-stage scale (0; 0.5; 1), where the thresholds were determined by combining theory-driven expectations with empirical evidence from the sample (Table 3).

TABLE III. CALIBRATION THRESHOLDS FOR FSQCA CONDITIONS

Condition	0-threshold	0.5-threshold	1-threshold
Idea Generation Capacity (IGC)	<5	5–10	>10
Leadership & Vision (LV)	Was change in leadership, no vision announcements	Either no change in leadership or vision announcements	No change in leadership, were vision announcements
Resource Commitment (RC)	<3%	3–5%	>5%
Resource Base (RB)	Startups (low)	Pivoting firms (medium)	SBUs (high)
Development Capability (DC)	0	1	>1

Calibrated fuzzy-set membership scores are calculated using the formula:

$$Fs(x)=\frac{\log\left(\frac{x-p_{0.5}}{p_1-x}\right)}{\log\left(\frac{p_1-p_{0.5}}{p_{0.5}-p_0}\right)} \quad (1)$$

Where:

x – Raw score of the case

p_0 – Threshold for full non-membership (0.0)

$p_{0.5}$ – Crossover point (0.5)

p_1 – Threshold for full membership (1.0)

Table 4 presents the calibrated fuzzy-set membership scores for each firm across the five conditions used in the fsQCA analysis. Values range from 0 (full non-membership) to 1 (full membership), with 0.5 indicating the crossover point of maximum ambiguity.

In fsQCA, necessity analysis tests whether a given condition must be present for the outcome to occur. A condition is considered necessary if it consistently appears whenever the outcome is present, typically indicated by a consistency score ≥ 0.9. This step is performed after calibration but prior to constructing the truth table and conducting sufficiency analysis.

The consistency of necessity is calculated using the formula:

$$\text{Consistency}_{necess}(X \subseteq Y) = \frac{\sum X_i \sum \min(X_i Y_i)}{\sum Y_i} \quad (2)$$

Where:

X_i – Membership score of case i in the condition set

Y_i – Membership score of case i in the outcome set

$\min(X_i, Y_i)$ – Minimum value between X_i and Y_i

Σ – Summation across all cases

TABLE IV. FUZZY SET SCORES BY FIRM AND CONDITION

Firm Name	IGC	LV	RC	RB	DC
SZ DJI Technology Co., Ltd.	1	1	1	0	1
Parrot Drones SAS	0	1	1	0.5	1
AeroVironment, Inc.	1	0.5	1	0.5	1
Northrop Grumman	1	0.5	0.5	1	0
Lockheed Martin	1	1	0	1	1
General Atomics Aeronautical Systems	0.5	1	0	1	1
Autel Robotics	1	0	0	1	1
Yuneec International Co. Ltd.	0	1	0	0.5	1
Boeing (Insitu)	1	1	0.5	1	1
Israel Aerospace Industries	1	1	1	1	1
Airbus Group	1	0.5	1	1	1
3D Robotics	0	0	0	0	0
PrecisionHawk Inc.	0	0	0	0	0
Intel Corporation	1	0	0	1	1

Where X is the fuzzy-set membership score of a condition and Y is the score of the outcome (Development Capability). Coverage, although less critical for necessity, is calculated as:

$$\text{Coverage}_{necess}(X \subseteq Y) = \frac{\sum X_i \sum \min(X_i Y_i)}{\sum X_i} \quad (3)$$

Where:

X_i – Membership score of case i in the condition set

Y_i – Membership score of case i in the outcome set

$\min(X_i, Y_i)$ – Minimum value between X_i and Y_i

Σ – Summation across all cases

Table 5 shows the results for all conditions tested as necessary for high Development Capability. None of the conditions met the threshold of consistency ≥ 0.9, indicating that no single condition alone is necessary for the outcome to occur. Nevertheless, some conditions (e.g., Idea Generation Capacity and Resource Base) show relatively high coverage, suggesting their frequent co-occurrence with high outcome cases.

TABLE V. NECESSITY ANALYSIS (CONDITION CONSISTENCY AND COVERAGE)

Condition	Consistency	Coverage
Idea Generation Capacity	0.773	0.895
Leadership and Vision	0.682	0.938
Resource Commitment	0.5	0.917
Resource Base	0.773	0.895

The highest scoring conditions, such as 'Idea Generation Capacity' (consistency = 0.773) and 'Resource Base' (consistency = 0.773), suggest these factors are frequently—but not always—present in firms with high Development Capability. While they are not strictly necessary, they may still represent important contributing conditions. 'Resource Commitment' had the lowest consistency (0.5), suggesting it is not reliably present in high-performing firms.

Each unique configuration (combination of fuzzy-set scores across conditions) is represented as:

$$\text{Configuration}_j = (C_1^J, C_2^J \ldots C_k^J) \quad (4)$$

Where C_k^J is fs value of condition k in configuration j.

Table 6 displays the truth table built from a simplified version of the calibrated fuzzy-set membership scores. Given the relatively small sample size of 14 firms, a crisp-set transformation was applied to enhance interpretability—specifically, all fuzzy values equal to or greater than 0.5 were treated as full membership (i.e., presence of the condition). Additionally, the condition Resource Commitment (RC) was excluded from the final configuration set due to its low consistency score (0.5) in the necessity analysis, indicating limited explanatory power. As a result, the final truth table reflects configurations based on three core conditions: Idea Generation Capacity (IGC), Leadership and Vision (LV), and Resource Base (RB). Each configuration represents a unique combination of these conditions, along with the number of firms exhibiting that pattern (frequency) and the observed outcome—Development Capability (DC).

TABLE VI. TRUTH TABLE (CONFIGURATIONS AND OUTCOME PRESENCE)

№	F	C	Firms / Cases	IGC	LV	RB	DC
1	2	1	3D Robotics, PrecisionHawk Inc.	0	0	0	0
2	2	1	Yuneec International Co. Ltd., Parrot Drones SAS	0	1	1	1
3	2	1	Autel Robotics, Intel Corporation	1	0	1	1
4	7	0.86	Northrop Grumman, AeroVironment, Inc., Airbus Group, Boeing (Insitu), Israel Aerospace Industries, General Atomics Aeronautical Systems	1	1	1	1
5	1	1	SZ DJI Technology Co., Ltd.	1	1	0	1

In fsQCA, sufficiency analysis identifies combinations of conditions (configurations) that reliably produce an outcome. A configuration is considered sufficient if the outcome generally follows whenever the configuration is present, typically indicated by a consistency score of 0.8 or higher.

The formula for calculating consistency of sufficiency is:

$$\text{Consistency}_{suff}(X \subseteq Y) = \frac{\sum X_i \sum \min(X_i Y_i)}{\sum Y_i} \quad (5)$$

Where:

X_i – Membership score of case i in the condition set

Y_i – Membership score of case i in the outcome set

$\min(X_i, Y_i)$ – Minimum value between X_i and Y_i

Σ – Summation across all cases

The simplified truth table yielded five unique configurations, several of which demonstrated high consistency with the outcome condition, Development Capability (DC). The most recurrent configuration—IGC=1, LV=1, RB=1—was observed in five firms, making it the dominant empirical pattern in the dataset. This configuration achieved a consistency of 0.86, suggesting it represents a sufficient pathway to high innovation outcomes in the nascent UAV industry.

However, this dominant configuration, observed in five firms—did not meet the robustness criterion (threshold of 0.9) due to a slightly lower consistency score (0.86). This reduction is attributable to the inclusion of Northrop Grumman, which exhibited the configuration but did not achieve high development capability. It underscores the importance of considering firm-level context when interpreting fsQCA results in small-N studies. Notably, only configurations with perfect consistency (C = 1.0) passed the robustness check.

Interpretation of the fsQCA results confirm that no single capability alone ensures innovation success. Instead, high development capability consistently emerged from specific combinations of conditions, reinforcing the configurational and systemic nature of innovation under uncertainty. The analysis clearly demonstrated equifinality: multiple distinct pathways—such as high ideation coupled with strong leadership, or strategic vision supported by a robust resource base—led to equally successful innovation outcomes.

Resource Base frequently appeared in successful configurations but did not guarantee success independently. For example, despite having a strong Resource Base (scored as 1), Northrop Grumman did not achieve high Development Capability. Conversely, firms with limited or moderate Resource Base, such as SZ DJI and Parrot, achieved innovation success through compensatory strengths in Idea Generation and Leadership.

The analysis confirmed that firms successfully balanced internal capabilities through substitution, compensating for weaknesses in one area with strengths in another. Startups such as SZ DJI succeeded despite a limited Resource Base due to robust Idea Generation Capacity and strong Leadership and Vision. Similarly, Autel Robotics and Intel Corporation achieved innovation outcomes by leveraging strong internal ideation and resource structures despite limitations in top-level leadership consistency. Pivot firms, such as Parrot Drones and AeroVironment, compensated for their moderate Resource Base by strategically leveraging stable leadership and accumulated assets, highlighting their adaptive innovation capacity.

Further analysis explored how organizational typology—represented by Resource Base scores (1.0 = SBUs, 0.5 = Pivot Firms, 0.0 = Startups)—influenced the translation of capabilities into successful outcomes.

- SBUs (e.g., Boeing Insitu, Lockheed Martin) primarily achieved innovation success through strong resource infrastructure, although such advantages required strategic alignment to be effective.
- Pivot firms (e.g., Parrot, AeroVironment) leveraged their legacy assets and reinforced them with leadership vision.
- Startups (e.g., SZ DJI, Autel Robotics) exhibit the greatest variability in success pathways. For startups, achieving innovation outcomes required a comprehensive alignment of ideation and leadership to offset infrastructural limitations.

These findings reinforce the notion that innovation success is not dependent on excellence in a single capability, but rather on achieving an effective balance tailored to each firm's structural and strategic context.

V. CONCLUSIONS

Internal capabilities play a critical role in how firms navigate uncertainty and achieve innovation outcomes in nascent industries. Based on a fuzzy-set Qualitative Comparative Analysis of leading UAV firms of various organizational types, this study identifies multiple viable capability configurations that lead to high development capability. Rather than relying on individual capabilities—such as idea generation, leadership, or resource commitment—firms succeed through specific combinations of these capabilities, structured in ways that align with their organisational type and strategic context, reflecting equifinality across startups, pivoting firms, and SBUs. Firms compensate for weaknesses in one area with strengths in another, underscoring the systemic nature of innovation management under uncertainty.

The scientific contribution lies in extending Anthony et al.'s Innovation Engine model by identifying particular internal capability types—such as resource commitment, resource base, and development capability—and assigning measurable metrics to each. This empirical extension links the Innovation Engine with organizational type, offering a richer framework for understanding how different organisational forms structure innovation under uncertainty using capability configurations.

Finally, several limitations should be acknowledged. The small sample size (n=14) places the analysis at the lower bound of fsQCA applicability, while the transformation of fuzzy values into crisp sets, applied for interpretability, reduces nuance. Moreover, the introduction of particular internal capabilities, though empirically informative, requires further theoretical justification and integration in established frameworks. Future research should therefore refine theoretical basis of selected internal capabilities, validate the use of product release frequency as a proxy for innovation performance and test identified capability configurations on larger datasets.

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Improvement and Analysis of Parallel-Operating Arc Furnances Indicators

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Abstract **– system-engineering solutions for the comprehensive improvement of electromagnetic compatibility and energy efficiency of parallel-operating arc steel melting furnaces (AF) were developed and their analysis was performed. Models for synthesizing the optimal control vector and the corresponding relationships of the AF's artificial external characteristics for implementing an adaptive optimal control strategy are proposed. A structural computer model of the double-loop control system for the AF's electromagnetic components was created, and the effectiveness of the proposed system-engineering solutions was studied. Analysis of the obtained results demonstrated enhanced electromagnetic compatibility and energy efficiency compared to the performance of the standard DSP-200 arc power controller.**

Keywords– arc furnace; electromagnetic compatibility, artificial external characteristics, optimization, dispersion, energy efficiency.

I. Introduction

Among the various aspects of power engineering development, particular importance is attached to the problem of electromagnetic compatibility of large electrical consumers with rapidly varying, phase-asymmetric, and nonlinear loads. Arc steel-melting furnaces (AF) belong to this category of consumers. Given the substantial specific power of AF's power equipment, ensuring their energy efficiency is equally critical [1].

Owing to the dynamic, phase-asymmetric variations in their load, these electrotechnological installations inject several adverse effects into the power supply network of the AF, thereby degrading power-quality metrics at that node. The most significant impacts are voltage fluctuations, phase imbalance, waveform distortion, and a reduction of power factor. [2, 3].

The intensity of these impacts on power quality is significantly amplified when several furnaces operate in parallel, and especially when the technological melting stages of the AF connected to the network node change asynchronously. The deterioration of power-quality metrics, in its turn, negatively affects the performance indicators of the AF themselves, in particular their energy efficiency. The extent of this degradation will vary depending on the number of arc furnaces connected to the node and the asymmetry of the melting stage changes. [4-6].

Considering the trend of an increasing share of AF steel melting over the past decade [3,4], along with the rise in their specific and installed power, the development and research of system-engineering solutions for the comprehensive improvement of EC and energy efficiency of parallel-operating arc furnaces is both timely and important.

II. Analasis of Known Solutions

Analysis of the impact of two or more parallel-operating arc furnaces groups on the power network, including cases in which the furnaces begin or complete melting almost simultaneously, were carried out in [7]. This study was primarily concerned with the issue of power quality at the supply node of the arc furnace group.

The results of experimental and simulation studies of voltage fluctuations at the power network node and waveform distortion are presented in [3–5, 8–12]. In these studies, computer simulations were performed using various models of the dynamic volt–ampere arc characteristics and their parameters. This approach made it possible to assess the impact of AF operation on power quality metrics at different melting stages.

In [13], the comprehensive influence of the dynamic volt–ampere characteristic parameters, as described by the Cassie–Mayr model, on energy efficiency metrics and the reduction of electrical losses is analyzed. The study introduces the concept of an "arc quality index," and on this basis derives an optimization model to enhance the efficiency of steel production in arc furnaces.

System-engineering solutions for the comprehensive improvement of voltage quality metrics at the power network node supplying arc furnaces – based on high-quality dynamic stabilization of electromagnetic coordinates, minimization of losses, and maximization of power transfer to the charge and melt – are proposed in [14–16]. These solutions are implemented via the real-time synthesis of a multi-criteria adaptive optimal control and the rapid control of electromagnetic coordinates. Studies of the developed solutions' effectiveness demonstrated enhanced electromagnetic compatibility (EC) and energy efficiency of a single arc furnace.

System-engineering solutions for improving the electro-magnetic compatibility and energy efficiency of arc furnaces using intelligent methods have been developed in [17,18]. Their implementation enables enhanced disturbance control dynamics, thereby reducing voltage fluctuations at the network node, increasing the power factor, decreasing electrical losses, and boosting the productivity of the arc furnaces.

A study of the influence of three-phase arcs nonlinear properties on power quality indicators at the network node was carried out using a 3D electromagnetic field model of the alternating current arc via the finite element method [19]. The arc was treated as an electrothermal unit, with electrical energy at its input and thermal energy at its output. The developed modeling approach for the three-phase arc system of the AF is effective for studying thermal and electrical processes, particularly in assessing electromagnetic compatibility indicators.

In [20], a computer model of the arc furnace and the power network was developed, and its operation's impact on current and voltage asymmetry, voltage drops, and waveform distor-tion was analyzed. The primary focus was on quantifying the voltage fluctuations generated by the arc furnaces. In the electrotechnological complex models presented in [20], the

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arcs are represented by an electromotive force whose magnitude depends on their length. This approach enabled more accurate reproduction of processes within the arcs and yielded more precise assessments of voltage quality at the network node and of the AF electrotechnological efficiency indicators.

Considering the above, it can be asserted that the task of developing new approaches to further enhance electromagnetic compatibility and energy efficiency during the parallel operation of multiple arc furnaces is relevant and important.

The objective of this research is to develop solutions for the comprehensive improvement of electromagnetic compatibility and energy efficiency of parallel-operating arc furnaces and to conduct their computer-based simulation.

The scientific novelty lies in the model for synthesizing a multi-criteria optimal control vector to improve electromagnetic compatibility and energy efficiency during the parallel operation of AFs.

III. Research Results

Rapidly changing, phase-asymmetric processes of reactive and active power consumption – driven by dynamic variations in arc lengths – have the greatest impact on the electromagnetic compatibility metrics of both the arc furnace and the power network. The intensity of these impacts is amplified when the power network's capacity is limited $S_{en} < (80-100)\cdot S_{ta}$, where S_{ta} – the total power of the furnace transformers, connected to the power network node.

One of the most effective approaches to improving these indicators is enhancing the automatic control system (ACS) of the AF electrical operating modes.

The double-loop automatic control system (ACS) structure for the AF electrical operating modes control was adopted as the baseline [4]. A solution for implementing optimal control based on multi-criteria EC was integrated into this structure. The control vector of this ACS comprises the arc length setpoint for the electromechanical electrode position control loop and the artificial external characteristic (AEC) of the AF, which is implemented by the electrical arc current control loop.

At the physical level, in each phase of the ACS, arc current control is performed by correspondingly changing the impedances of the chokes, connected in series with the primary windings of the furnace transformer. This is implemented by a power magneto-thyristor converter [4] and proportional-integral arc current controllers [4]. The arc lengths are controlled by a typical off-the-shelf electromechanical loop.

System-engineering optimization solutions were developed for the double-loop ACS, aimed at the primary suppression of disturbances at their source – specifically, stabilizing the modes of active and reactive power consumption, balancing the phase loads, controlling arc lengths and currents on a per-phase basis.

The control inputs for implementing the specified control strategy are AF AEC $I_a(U_a)$ and the arc voltage setpoint $U_{a.set}$ of the electrode movement systems (EMS). To minimize voltage fluctuations in the power network, it is firstly necessary to minimize the variances of active and reactive power and to minimize the coefficient of phase load asymmetry. Figure 1 shows the natural characteristic 1 and the proposed AEC 2–6 of the DSP-200 AF for implementing optimal control aimed at improving EC and energy efficiency.

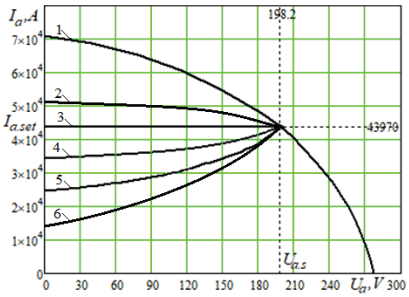


Fig.1. Natural 1 and artificial external characteristics 2-6 of the DSP-200 arc furnace to improve electromagnetic compatibility and efficiency

Characteristics 3 and 5, in the static mode, correspond to $I_a = const$ and $Q = const$, and in dynamic operation – to the minima of arc currents dispersions $D_{I_a} \to min$ and reactive power dispersions $D_Q \to min$, respectively. The AEC corresponding to the minimum phase-load asymmetry is located between these two characteristics 2 and 6 (Fig.1).

For the real-time implementation of the optimal control vector synthesis model during the melting process under conditions of stochastically changing characteristic disturbances, the AEC dependence will be described by the following relation:

$$I_a^*(U_a,\gamma) = \gamma\cdot\left[\frac{U_a}{U_{a.set}} + \left|\frac{U_a - \frac{U_a}{U_{a.set}}}{\frac{U_a}{U_{a.set}}}\right|\right]\cdot[7.43e^{-10}{U_a}^4 + 1.07e^{-7}{U_a}^3 - 2.20e^{-5}{U_a}^2 + 5.72e^{-3}U_a + 7.72], \quad (1)$$

where $U_{a.set}$ = 198.2V – the EMS arc voltage setpoint;

γ – the variable AEC optimization coefficient (for the base AEC 4 (Fig.1) γ=1.0). To reproduce AEC 2-6 (Fig.1) this coefficient must be varied in the range $\gamma\in$[0.4, 1.8]. Characteristic 1 (Fig.1) was calculated using the known method from [21], and characteristics 2-6 were calculated using the obtained optimization equation (1).

Considering the substantial distortion of the sinusoidal nature of the AF's electrical operating mode coordinates, the running value of the phase-load asymmetry coefficient was calculated from the running values of the phase currents first harmonics using the following expression:

$$K = \sqrt{\frac{I_A^2 + I_B^2 - 2\sqrt{3}S - \sqrt{I_A^2\cdot I_A^2 - 4S^2}}{I_A^2 + I_B^2 + 2\sqrt{3}S - \sqrt{I_A^2\cdot I_A^2 - 4S^2}}} \quad (2)$$

where I_A, I_B, – are the effective values of the phase *A*, *B* first harmonics; *S* – the area of the triangle formed by the first harmonics of the phase currents;

An optimization functional is proposed to minimize the negative impact of a group of parallel-operating arc furnaces on electromagnetic compatibility metrics at the power network node (3) (*j* = *1*, *2*, ... *n* – number of AFs):

$$F(\bar{\gamma}) = \sum_j^n(\lambda_{I_a}\cdot D^*_{I_a,j}(\gamma_j) + \lambda_Q\cdot D^*_{Q,j}(\gamma_j) + \lambda_K\cdot K^*_j(\gamma_j)) \Rightarrow \min, \quad (3)$$

where $D^*_{I_a,j}(\gamma) = \frac{D_{I_a,j}(U_{a.set,j},\sigma_j,\gamma)}{D^{max}_{I_a,j}}$ – normalized arc currents dispersion; $D^*_{Q,j}(\gamma) = \frac{D_{Q,j}(U_{a.set,j},\sigma_j,\gamma)}{D^{max}_{Q,j}}$ – normalized AF reactive power dispersion;

$K_j^*(\gamma) = \frac{D_{K,j}(U_{a.set,j},\sigma_j,\gamma)}{D_{K,j}^{max}}$ – normalized asymmetry coefficient;

$\lambda_{I_a}, \lambda_Q, \lambda_K$ – the weighting coefficients for the arc currents dispersions, AF reactive power, and the asymmetry coefficient, respectively;

$D_{I_a,j}^{max}$, $D_{Q,j}^{max}$, $D_{K,j}^{max}$ – the maximum values of the arc currents dispersions, the reactive power of the j-th furnace, and the asymmetry coefficient;

σ_j – the running root-mean-square deviation of the arc voltages;

$U_{a.set,j}$ – the arc voltage setpoint of the *j*-th AF EMS.

γ_j – the variable shaping (optimization) coefficient of the the *j*-th AF AEC;

The running values of the dispersions of these electromagnetic coordinates are computed in real time over each steady-state interval ($T\in$[100-120]s)) for each furnace, taking into account the probability density distribution of the arc voltage variation process. The dispersion of the asymmetry coefficient is then calculated by the following expression:

$$D_K(U_{a.set},\sigma,\gamma) = \int_0^{U_{2ph}} (K(U_a,\gamma) - \overline{K}(U_{a.set},\sigma,\gamma))^2 \cdot \cdot f(U_a,\overline{U}_a,\sigma)dU_a, \quad (3)$$

where $K(U_a,\gamma)$, $\overline{K}(U_{a.set},\sigma,\gamma)$, $f(U_a,\overline{U}_a,\sigma)$ – the running value, the mathematical expectation, and the probability distribution law of the asymmetry coefficient variation process, respectively.

The synthesis of the optimal value of the shaping coefficient γ, corresponding to the selected control strategy (3), is carried out during each disturbance steady-state interval (T=100…120s) taking into account the furnace transformer secondary voltage U_{2ph}, as prescribed by the melting schedule, the arc voltage setpoint$U_{a.set,j}$ and the running root-mean-square deviation σ_j of the arc voltages.

The effectiveness of the adaptive optimal control strategy defined in (3) was evaluated using a digital model of the proposed double-loop ACS structure for the parameters of a DSP-200-type AF, which was developed based on the models [20]. The time dependences of the EM coordinates obtained on this computer model are shown in Fig.2 and Fig.4.

Using this model, electromagnetic compatibility and energy efficiency indicators were analyzed for two DSP-200 AFs connected to the power network node, controlling stationary random disturbances across various technological melting stages and their synchronous transitions.

Figure 2 presents 30 second model-generated fragments of the arc currents variation $I_{a.1}(t)$, the reactive power $Q_1(t)$ of the first arc furnace, and the network voltage $U_{EN}(t)$ obtained during handling random disturbances by the double-loop ACS and implementation of the AEC with $\gamma_1 = 1.05$ and $U_{a.set,1} = 198\,V$.

Figure 3 also illustrates the arc currents variations $I_{a.1}(t)$, the reactive power $Q_1(t)$ of the first AF and the network voltage $U_{EN}(t)$ when operating under a single-loop electro-magnetic control system (the ARDM-T-12 arc power controller) on the natural external characteristic 1 (Fig.1), $U_{a.set,1} = 198\,V$.

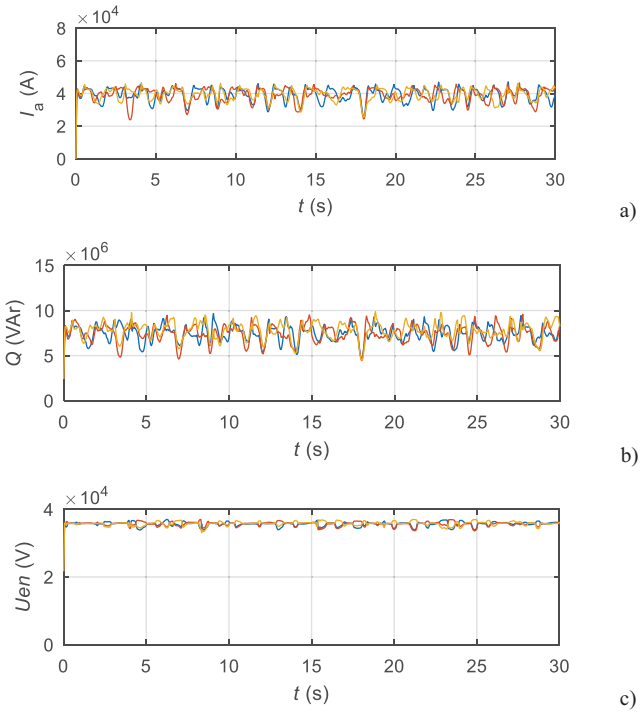


Fig.2. Dependencies of the arc currents $I_{a.1}(t)$ a), of the first AF $Q_1(t)$ b) and the voltage EN $U_{EN1}(t)$ c) during the operation of the double-loop ACS

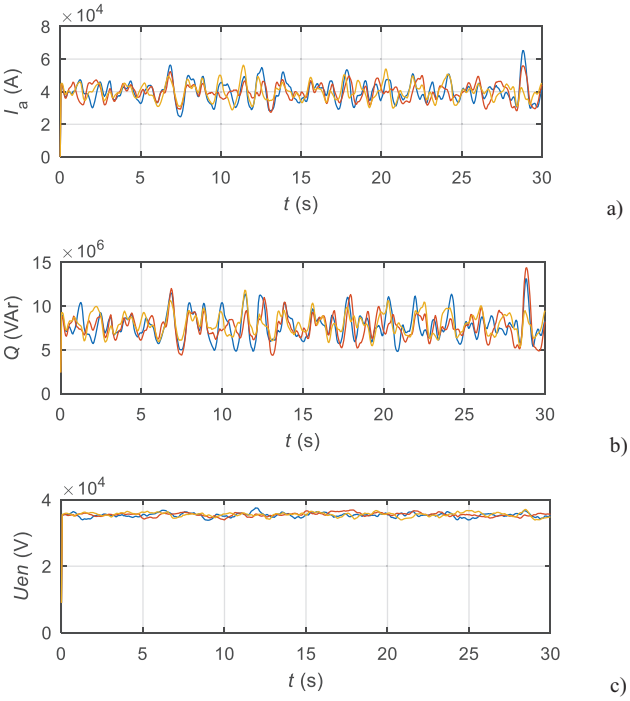


Fig.3. Dependencies of the arc currents $I_{a.1}(t)$ a), the reactive power $Q_1(t)$ b) and the voltage EN $U_{EN1}(t)$ c) during operation of the ARDM-T-12 controller

Figure 4 shows, based on the model, the dependency of the mean values over steady-state intervals of the arc currents dispersion $\overline{D}_{I_a}^*$, the reactive power dispersion $\overline{D}_Q^*$, and the asymmetry coefficient $\overline{K}_{Q1}^*$ of the first AF's reactive power on the coefficient γ during operation of the double-loop ACS.

The EC, energy efficiency, and dynamic performance indicators – obtained by statistically processing the computer simulated time dependencies of the EM coordinates for two parallel-operating DSP-200 AFs during the well-melting stage at $\gamma = 1.05$ are summarized in Table 1.

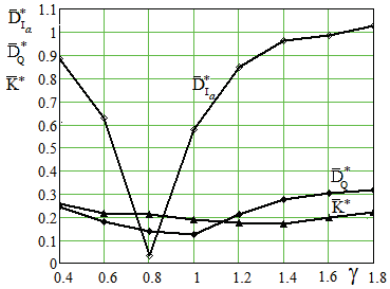


Fig.4. Normalized arc currents dispersions $\overline{D}^*_{I_a}$, reactive power $\overline{D}^*_Q$ and the asymmetry coefficient $\overline{K}^*_{Q1}$ of the first AF when operating under modied AEC

TABLE I. INDICES OF THE STUDIED STRUCTURES OF EM ACS WITH THE DSP-200 ARC FURNACE

ACS structure / Quality indices	*Proposed optimal control system (3)*	*ARDM-T-12 arc power controller*	*δ,%*
Voltages EN, $\overline{U}_{EN}$, kV	35.67	35.45	0.006
Dispersion, $\overline{D}_{U_{EN}}$, kV2	0.684	1.145	40.3
Arc current, $\overline{I}_a$, kA	39.20	39.82	1.56
Dispersion, $\overline{D}_{I_a}$, kA2	18.45	32.80	43.75
Reactive power, Q,MVAr	7.532	7.799	3.42
Dispersion, $\overline{D}_Q$, MVAr2	0.879	1.996	56
Power factor (PF), $\overline{cos\varphi}$	0.825	0.822	0.004
Dispersion of PF, $\overline{D}_{cos\varphi}$	0.0029	0.00396	26.7
Electrical losses, $\overline{P}_{el}$, MW	1.947	2.027	3.94
Dispersion, $\overline{D}_{P_{el}}$, MW2	0.147	0.326	54.9
Asymmetry coefficient, $\overline{K}_Q$	1.0689	1.0649	0.004

The quality indicators presented in Table 1 show, that, compared to the ARDM-T-12 controller, the double-loop ACS EM achieves a comprehensive improvement in EC at the network connection node for the two parallel-operating AFs, while also enhancing power efficiency and the dynamic control performance of the EM coordinates.

In particular, owing to the substantial enhancement in the dynamic precision of stabilizing the arc currents and AF reactive power using the proposed double-loop EM ACS, voltage quality metrics at the network node are improved, the AF's reactive power is reduced, and the electrical power losses in the AF's short circuit network are decreased.

IV. CONCLUSIONS

The analysis of the indicators in Table 1 illustrates the effectiveness of the proposed optimization and system-engineering solutions for the comprehensive improvement of electromagnetic compatibility and energy efficiency of parallel-operating arc furnaces.

Further research by the authors will be directed toward more extensive studies of these EC and energy efficiency indicators during the operation of various numbers of AF with both synchronous and asynchronous changes in their technological melting stages, as well as toward refining functional (3) for extremal control of electromagnetic parameters according to the criteria outlined above.

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Automated Detection of Disinformation Routes Based on Vector Grouping of Social Media Posts

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***Abstract*—The study presents an information system that performs automated analysis of arrays of publications from online media and social networks to identify primary sources and routes of disinformation spread. The methodology is based on the vector representation of texts using embedding models (local intfloat/multilingual-e5-base and OpenAI text-embedding-3-small), calculating cosine similarity, and building clusters of similar messages. A timeline graph that displays the chronology and distribution channels. Experimental testing of the system on three datasets showed the effectiveness of the approach. The OpenAI model detected clusters with high accuracy at a threshold of 0.6–0.7, demonstrating a clear visualisation of dynamics. The local model required raising the threshold to 0.85–0.9, but also provided the detection of groups of similar messages. For example, in cluster No. 6, a typical scheme of spreading a fake was recorded: from the initial Telegram post to retransmission on X and Facebook with a total delay of about 5 days. The system also allows you to detect attempts to hide the source (for example, by converting a repost into a new post) and visualise time intervals and platforms through which disinformation was spread. The results demonstrate that the proposed approach is promising for information analytics, allows you to identify fake messages, and can be integrated into cyber hygiene systems or information campaign investigations.**

Keywords—disinformation, text clustering, social networks, embedding, cosine similarity, OpenAI, vectorisation, propagation graph, information security, timeline visualisation

I. Introduction

The modern information environment, especially in the war and post-war periods, is increasingly exposed to the influence of disinformation campaigns. The rapid spread of false information through social networks and online platforms poses significant threats to the security of society, influences public opinion and complicates decision-making processes. In this regard, there is a growing need to create automated tools capable of identifying sources of disinformation, analysing the routes of its spread and identifying key relays of fake messages. The relevance of the study is due to the need to improve information security in the conditions of hybrid war, when information is used as a tool of influence. Existing approaches are primarily focused on fact-checking messages, but do not allow you to automatically track the path of the thesis spread from its primary source to the audience. Therefore, the study aimed at identifying the chronology and dynamics of the spread of fake posts has both scientific and practical significance. The purpose of the work is to develop and implement an information system for automated analysis of messages in social networks with the ability to identify disinformation clusters, determine the routes of their distribution and visualise the data obtained. The object of the study is text publications in social networks and online media. The subject of the study is the methods of vector representation of the text, calculation of semantic similarity and clustering of messages in order to analyse information flows. The objectives of the study:

1. Develop a system architecture for processing and analysing social message arrays.
2. Implement text vectorisation algorithms using local and cloud embedding models.
3. To propose a method for clustering messages by the level of cosine similarity.
4. Build graphs for the spread of fake messages with visualisation of the chronology.
5. Conduct an experimental check of the system on real datasets.

The scientific novelty of the study lies in the combination of vector grouping of publications with the construction of chronological graphs of information distribution, which allows you to automatically detect not only similar messages but also the logic and speed of their distribution. An approach to visualising disinformation clusters is proposed, which allows for identifying the so-called "zero" entry points of fakes and ways to hide their sources. The practical value of the work lies in the creation of an effective tool for specialists in the field of information security, journalistic investigations and OSINT analytics. The results of the study can be integrated into information space monitoring systems in order to promptly respond to attempts at manipulation or disinformation attacks.

II. Related Works

A. Graph methods for detecting disinformation

In [1], a meta-graph structure is formed, combining events, semantics and topics. The application of GNN resulted in an increase in accuracy of 3–4% compared to pure cascade classification. In [2], the authors offer Geometric Deep Learning (GraphCNN) to analyse information cascades on Twitter – ROC AUC ≈92.7 %, achieving early detection of

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"fake" news. In [3], an overview of graph neural networks for fake news detection, classifying methods by context, propaganda and social context, is carried out.

B. Visualisation and tracing of propagation routes

Research [4] describes a tool for tracing disinformation through groups, bots, retweets, URLs, etc. It uses Gephi, NodeXL, and Hoaxy to analyse structures and detect coordinated interactions. [5] combines BERTopic and GNN to visualise topic cascades and interpretive detection of fake news.

C. Vectorisation and clustering methods

In [6], in the example of Weibo, semantic vectorisation (TF-IDF with dense embeddings) is performed with K-means clustering-to detect fake topics. [7] describes the use of network embedding (e.g. node2vec) to analyse vaccine doubts – an example of combined content and structural analysis.

D. Distribution patterns (epidemic)

SIR-based, Independent Cascade, etc., are classic models of dissemination of "virus-like" information used to analyse the spread of rumours through social networks [8]. In [9], the analysis of key metrics (centrality, effectiveness) in the disinformation network was carried out – "conversation leaders" with high eigenvector-centrality were identified.

E. User Relationship Analytics

In [10], graph vector modelling of users (mutual followers/friendship) qualitatively adds fake detection to the content-based system. [11] describes real cases of analysis and detection of disinformation campaigns (Russian IRAs, Chinese anti-vaccine art) through network mapping.

III. Methods and Materials

The use of graphs with GNN is the key to accurate detection of fake cascades (~90%+ accuracy). Visualisation of cascades (BerTopic, Gephi, Hoaxy) helps to analyse routes and typical structures. Vectorisation with clustering (TF-IDF, dense embeddings, cosine) is the basis for recognising groups of similar messages. Epidemic and structural-network models help to simulate the dynamics of spread. Users' social connections are an additional variable for better contextual understanding. Our approach combines several key strategies like vectorisation [6], clustering by cosine similarity, graph visualisation of propagation [4-5], and network context using GNN approaches [1-3]. This combination, according to the literature, is promising. It synthesises the advantages of cluster, graph and visualisation methods for in-depth analysis of information flows [12-16].

IV. Experiments, Results and Discussions

The system analyses an array of publications in social networks/online media, finds messages that are similar in content, groups them into clusters, and visualises the chronological "routes" of the spread of each disinformation thesis. In this way, it is possible to identify primary sources, key relays, and the pace of the spread of fakes (Table 1). An Excel file with a minimum set of columns is expected (Table 2). The utils.load_data() function reformats dates (to_timestamp()), normalises Boolean marks (transform_bool()), typifies numeric columns, and removes service columns.

TABLE I. System structure

Function	Description	Key methods
Pre-processing of data	Loading a table with posts, cleaning, converting dates and types	utils.load_data()
Text vectorisation	Getting embeddings for each message via the local model or OpenAI API	utils.get_embedding()
Germination calculation	Cosine similarity between embeddings, matrix formation	sklearn.metrics. pairwise.cosine _similarity()
Clustering	Constructing a graph by the similarity threshold, searching for connectivity components	utils.clusters_ from_similarity()
Visualisation	Timeline graph of grouped posts by similarity	visualisation. timeline_graph()

TABLE II. Data structure

N	Attribute	Description
1	date, time	date and time of publication;
2	text	full text of the post;
3	language	name of the language;
4	post/repost	original (post) or repost;
5	author/group	author or distribution group;
6	source	platform/channel;
7	web-address	URL of the publication;
8	likes, shares	reaction counters;
9	flag	authenticity mark (True - fact, False - fake).

Route detection algorithm:

$$Vectorisation \rightarrow Similarity\ matrix\ Threshold\ \tau \rightarrow Clusters \rightarrow Route$$

For each publication, an array of embeddings of size d is created - either by the local model "intfloat/multilingual-e5-base" or via OpenAI ("text-embedding-3-small"). We calculate the cosine similarity of S between all embeddings. We connect the vertices i, j with an edge if $S_{ij} > \tau$ (typically $\tau = 0.9$). We find the components of the graph connectivity (sets of self-similar posts). We select clusters where ≥ 2 posts are in the middle of each cluster, and the vertices are sorted by datetime. The sequentially connected edges show the propagation chronology. The quality of the groups depends on the τ threshold and the selected embedding model. $\tau = 0.9$ for the intfloat/multilingual-e5-base model and $\tau = 0.7$ for the OpenAI text-embedding-3-small model.

A. Analysis of results

Dataset 1 is generally uniform, with a few clusters of duplicate texts. There is a noticeable vertical-horizontal stripe in the vicinity of index 200. It is a group of posts-"anomalies" that differ significantly from the rest and give a low resemblance to the entire array. In the vicinity of the ranges 10-40, 680-720, and 760-810, small yellowish blocks are clearly visible - these are groups of similar posts (within the group itself). In dataset 2, the first ≈ 350 lines are rows of yellow squares (6-7 compact blocks). These are groups of almost identical theses repeated in batches. OpenAI clearly delineates the boundaries of blocks; local also shows them, but the borders are blurred, and part of the "internal" background merges with the blocks due to the high base values. Bright yellow blocks are observed - these are groups of reposts that are identical to each other. The second half of the matrix is much darker - more heterogeneous posts, almost without reposts, but similar posts are often found. In this part of the graph, blocks near the diagonal are not visible since the data is not sorted.

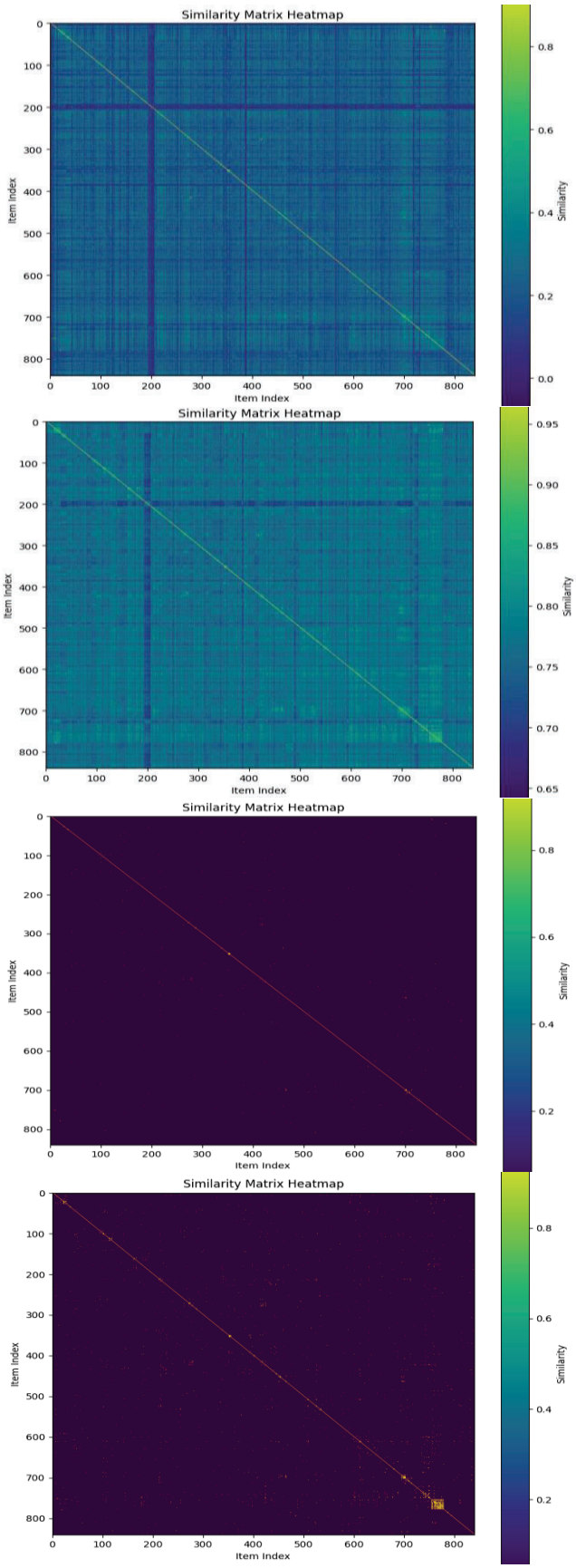


Fig. 1. Heatmaps of cosine similarity between embeddings of the 1st dataset: a) embeddings obtained using the OpenAI model; b) embeddings obtained using a local model; c) binary representation of grouped OpenAI embeddings with a cosine similarity of at least 0.6, d) binary representation of grouped local embeddings with a cosine similarity of at least 0.85 (Yellow pixel - posts are defined as similar, purple - different).

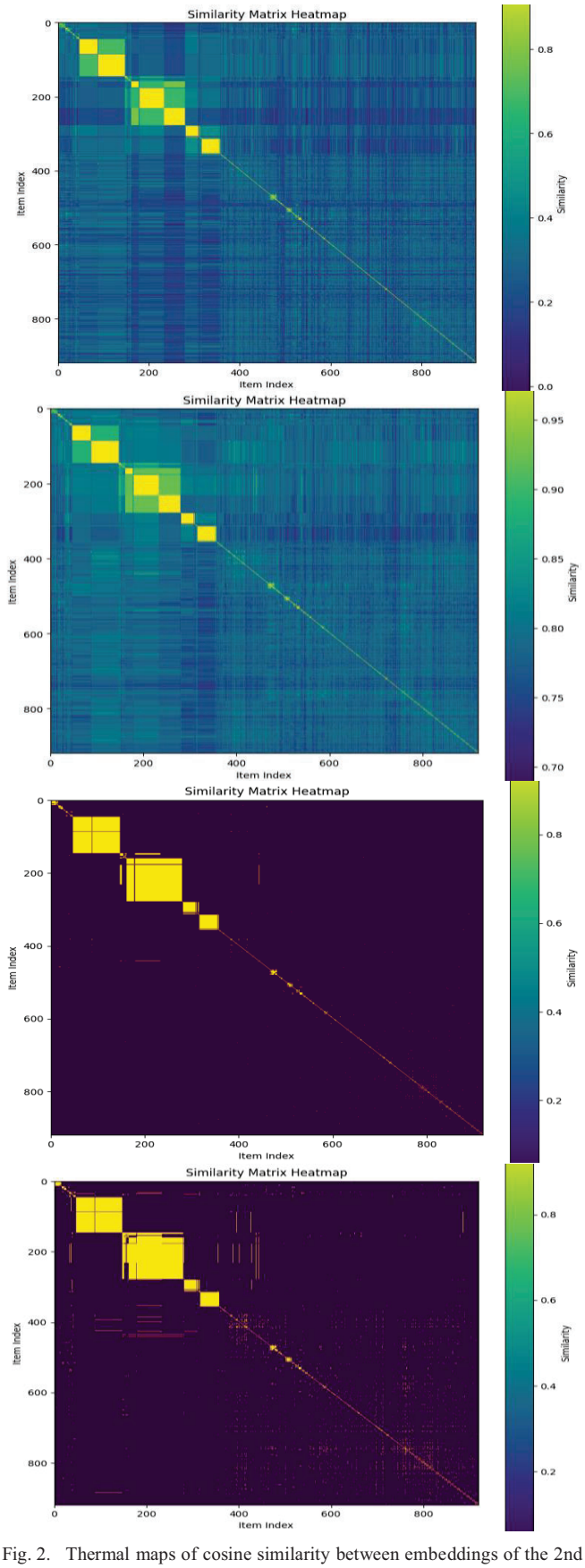
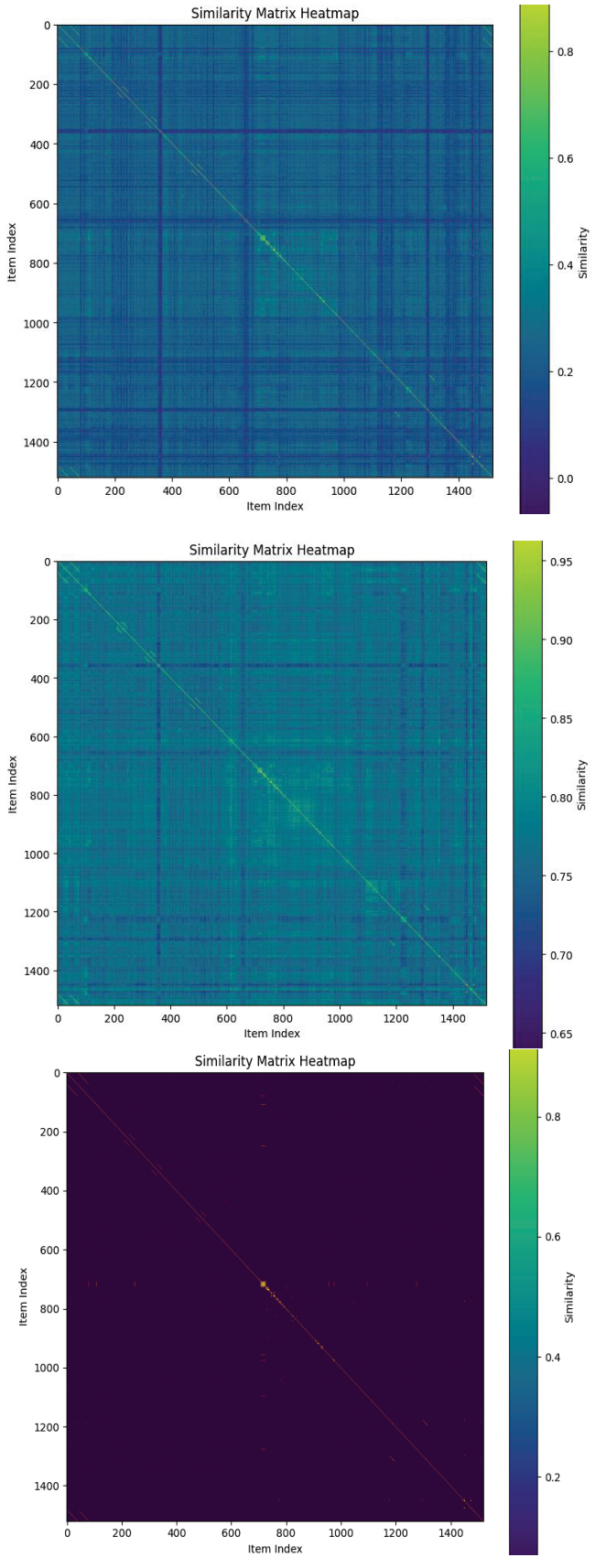


Fig. 2. Thermal maps of cosine similarity between embeddings of the 2nd dataset: a) embeddings obtained using the OpenAI model; b) embeddings obtained using a local model; c) binary representation of grouped OpenAI embeddings with a cosine similarity of at least 0.6, d) binary representation of grouped local embeddings with a cosine similarity of at least 0.85 (Yellow pixel - posts are defined as similar, purple - different).

For dataset 3, OpenAI has a homogeneous grain graph with a small but clear "spot" ≈ indices 720\u2012760 - a series of almost identical messages (reposts) and some single clusters of similar posts. For Local, the structure is the same, but the background is much brighter; the difference between typical and "duplicated" posts is less visible.



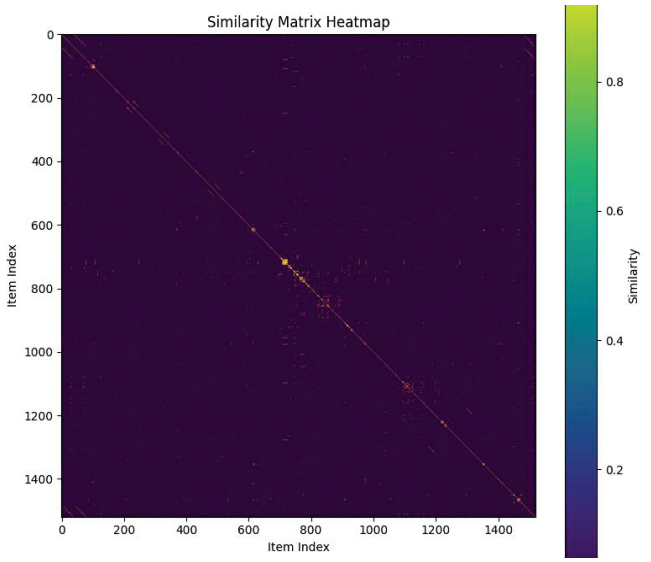


Fig. 3. Thermal maps of cosine similarity between the embeddings of the final dataset: a) embeddings obtained using the OpenAI model; b) embeddings obtained using a local model; c) binary representation of grouped OpenAI embeddings with a cosine similarity of at least 0.6, d) binary representation of grouped local embeddings with cosine similarity of at least 0.85 (Yellow pixel - posts are defined as similar, purple - different).

B. *Choosing an embedding model.*

OpenAI gives more dynamics and contrast, so it's easier to visually and algorithmically find clusters. The local version can generate an artificially high "base" similarity. For this, it is worth raising the threshold τ (0.90 +) or pre-centring/normalising the vectors.

C. *Further processing.*

It is necessary to mark large yellow squares as "components of high convergence" and submit them for graph rendering. Suspicious bars (one post against all) are a helpful indicator of unique entry points of disinformation. Let's build a graph timeline for the final dataset and analyse the result. The horizontal axis is the time (left → right). Each row is a separate cluster, i.e. a group of messages that turned out to be very similar in content in the cosine similarity matrix ($\tau \geq 0.90$). The edges combine the publications within the cluster in chronological order. The vertices are the circle - the original post (post) and the triangle - repost/repost (repost). The colour blue - the message is marked as accurate, and red - is fake. Let's consider cluster No. 6 from Figure 4, which illustrates a typical fake route:

TABLE III. ANALYSIS OF THE EXAMPLE IN FIGURE 4

N	Description
1	The Resident (Telegram, 02 Aug 20:30)
2	↓ 5 days
3	Alexander Kovalenko (X)
4	↓ 4 year
5	Andro Newman (X)
6	↓ 10 min
7	Igor Liepas (X)
8	↓ 2 year
9	Marina (X)
10	↓ Year 9
11	Elena Ying ▲ (FB, repost)
12	↓ a few seconds
13	Elena Ying ● (FB, "new" post)

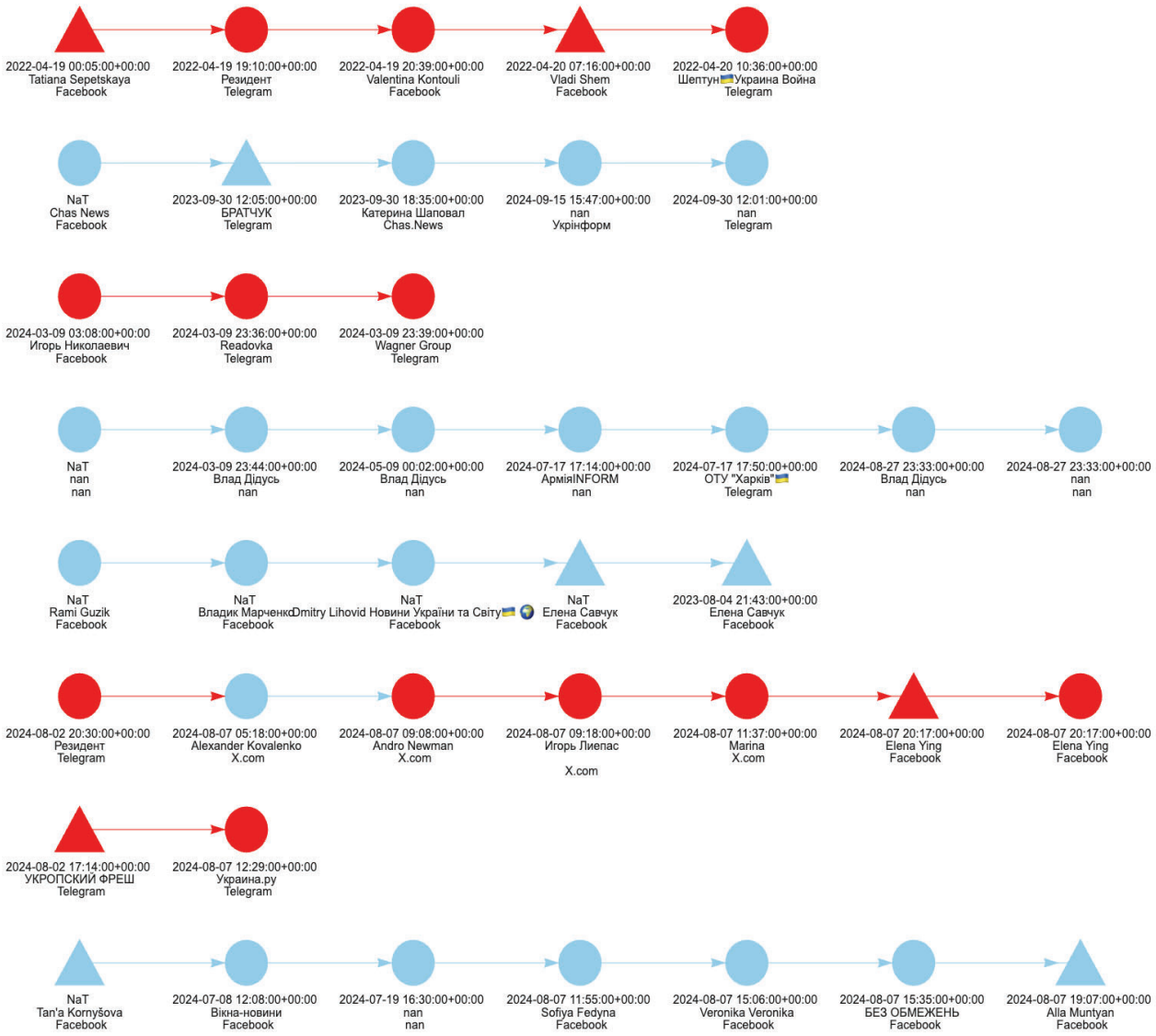


Fig. 4. Information dissemination graph

An interval of ~5 days between the first TG source and the distribution on X/Facebook. Turning a repost into a "fresh" post (triangle → circle) by the same author is a typical way to "lose" the original origin of the message. The text (tooltip) mentions, "a NASAMS rocket hit ... Okhmatdyt", which is the disinformation thesis of the Russian Federation.

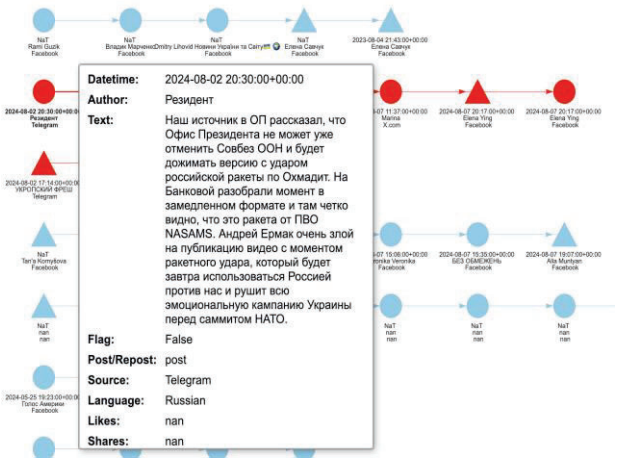


Fig. 5. Visualisation of data about the first appearance of information

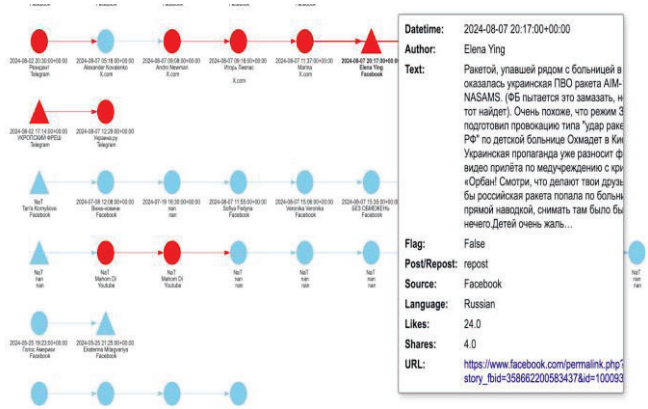


Fig. 6. Data visualisation of one of the reposts of the information specified in Fig.5.

In the course of the work, an information system was created that automatically tracks the source and further path of disinformation messages. The system analyses posts from different platforms, converts text into vector representations, and groups messages into clusters according to cosine similarity. Based on the formed clusters, the system builds a chronological graph where you can see who and when the

thesis was published, on which sites it was picked up, and with what delay the fake moved to other platforms. Practical experiments on three datasets showed that the OpenAI model creates vector representations that stand out more clearly in the similarity matrix, making it easier to isolate clusters. In contrast, vector representations created by the local model have a higher similarity threshold. Despite this, both models confidently detect groups of similar posts and reposts. The work also revealed data weaknesses: omissions in the dates of the authors' names affect the correctness of the graph, and manual marking of authenticity remains a bottleneck. At the same time, the proposed approach allows analysts to instantly find "zero" sources and assess the speed of distribution.

V. Conclusions

Within the framework of the study, an information system was developed and implemented for automated analysis of messages in social networks in order to identify disinformation clusters and track the routes of their spread. The system combines modern methods of natural language processing (vectorisation using embedding models), cosine similarity calculation, clustering of similar posts, and graph visualisation of propagation. The system architecture has been successfully implemented, which allows downloading, cleaning, vectorising, and analysing large volumes of publications from social networks and online media. Two OpenAI embedding models, text-embedding-3-small and the intfloat/multilingual-e5-base model, are compared. OpenAI text-embedding-3-small provides high contrast in similarity matrices, making it easier to detect clusters at a lower threshold ($\tau \approx 0.6–0.7$). The local intfloat/multilingual-e5-base model produces vectors with higher baseline similarity, which requires a threshold elevation ($\tau \approx 0.85–0.9$) for reliable clustering. An experimental analysis of information dissemination was carried out on three datasets. It has been established that the OpenAI model allows you to visually and algorithmically distinguish clusters more clearly. Characteristic groups of almost identical posts and reposts have been identified, which confirms the effectiveness of the chosen approach. The ability to identify routes for the spread of fake theses has been demonstrated. For example, cluster No. 6 with a message about "NASAMS missiles" was tracked from the Telegram channel to Facebook with a time delay of about 5 days, including reposts and deliberate changes to hide the source. Weaknesses in the input data were identified, in particular, the lack of unified author names and inaccuracies in confidence labels that affect the quality of the constructed graphs. The practical implementation of the system allows analysts to quickly identify the primary sources of disinformation, assess the rate of its spread and identify repeaters, which is an essential tool in countering information attacks. Thus, the developed system can be used in the field of information security, OSINT intelligence, and journalistic investigations, as well as a basis for further research in the field of detecting fake messages and modelling information flows.

Acknowledgement

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Ecosystem Entry Strategy Patterns and Resilience of Electrical and Electronic Industry Multinational Corporations under High Market Volatility

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***Abstract*—In the context of accelerating technological dynamism and recurring market volatility, electrical and electronic industry multinational corporations (EEI-MNCs) increasingly engage in global innovation ecosystems to secure long-term competitiveness. These dynamics underscore a growing corporate imperative to architect ecosystem entry strategies that strengthen organizational resilience. Building on an extended strategic tripod framework that integrates ecosystem-specific advantages with resource- and institution-based views, we analyze a sample of EEI-MNCs active in key innovation ecosystems. Abnormal returns during major volatility events (2015–2022) are used as indicators of firms' historical resilience trajectories. These trajectories are then compared with the current architectural patterns of ecosystem entry strategies, assessed across three pillars: strategic discovery, structural support, and selective participation. The results support a typology distinguishing "Proactive Leaders," "Pivots," "Naturals," and a "Lucky Star" based on resilience trajectories, ESG performance, and strategic coherence. Our findings underscore the growing strategic centrality of ecosystem participation and offer implications for resilience-oriented international business strategies in turbulent environments.**

Keywords—international business, international management, technological dynamism, market volatility, global innovation ecosystem, ecosystem-specific advantages, firm's ecosystem strategy, organizational resilience, electrical and electronic industry multinational corporation

I. Introduction

Multinational corporations in the electrical and electronic industry (EEI-MNCs) operate in an environment increasingly shaped by technological dynamism and market volatility. Rapid shifts in innovation trajectories and rising uncertainty—reflected in periods of high financial market volatility—challenge their ability to sustain competitiveness and strategic coherence [1], [2]. Traditional firm-specific advantages often erode quickly, urging firms to explore new paths for resilience [3], [4], [5].

In this context, global innovation ecosystems have emerged as critical arenas of strategic action. They enable international firms to access distributed knowledge, enhance dynamic capabilities, and co-create value across organizational boundaries [6], [7], [8], [9], [10], [11], [12]. Ecosystem-specific advantages increasingly complement resource- and market-based strategies [9].

However, despite the growing interest in ecosystem engagement, existing research focuses largely on participation as such. We still know little about how EEI-MNCs construct their ecosystem strategies—how they frame, structure, and selectively execute them to achieve long-term resilience under volatility. This study addresses this gap by exploring the architectural patterns of ecosystem entry strategies and their relationship to resilience outcomes across a sample of leading EEI-MNCs.

II. Theoretical Background and Purpose

A. Ecosystem-Specific Advantages

In conditions of rapid innovation cycles and technological discontinuities, traditional firm-level capabilities often fall short of ensuring sustained competitiveness. In response, firms increasingly rely on global innovation ecosystems to access, share, and develop complementary resources and knowledge. These ecosystems offer ecosystem-specific advantages, distinct from market- or resource-based advantages, by enabling firms to co-create value and absorb external uncertainty through collective adaptation [8], [9], [13], [14]. Under high market volatility (e.g., VIX surges), ecosystem engagement offers both flexibility and legitimacy, as firms distribute risk and align with dynamic technological and regulatory trends [15].

B. Strategic Tripod Revisited

From the strategic management perspective, the well-established "strategic tripod" of market-based view (MBV), resource-based view (RBV), and institution-based view (IBV) provides a robust lens to understand firm behavior under uncertainty [16], [17]. In today's environment of accelerating technological dynamism and institutional complexity, ecosystem-specific advantages can be conceptualized as an integrated dimension bridging RBV and IBV. These advantages arise not only from a firm's internal capabilities, as emphasized in RBV, but also from their embeddedness in formalized and co-governed ecosystems, aligning with the IBV logic of legitimacy and institutional conformity, where value is co-created and risks are collectively mitigated [18], [19], [20].

Ecosystem participation—particularly in external multilateral ecosystems (EM-ecosystems)—extends the firm's strategic logic beyond the confines of the firm boundary or dyadic alliances. In this light, the strategic tripod can increasingly be seen as grounded on an ecosystem foundation, where institutions, resources, and markets intersect. As such, ecosystem-specific considerations are no longer peripheral but are becoming integral to corporate-level strategy, shaping how MNCs explore innovation, align with institutional trends (e.g., sustainability, digital sovereignty), and orchestrate long-term resilience.

C. Defining Ecosystem Strategy

According to the Adner's structural view of ecosystems, a firm's ecosystem strategy is defined by how it aligns partners

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and secures its role within the broader network [15]. Extending this, we operationalize ecosystem strategy through three architectural components [5]:

1) Entry strategy: How international firms choose which external multilateral ecosystems to join or exit.

2) Relationship management: How international firms structure interorganizational interactions within ecosystems.

3) Value chain orchestration: How ecosystem participation informs their global operations and supply chains.

D. Ecosystem Strategy and Resilience

The integration of ecosystem-specific advantages into a firm's strategy gains particular salience under conditions of high environmental dynamism and volatility. As outlined by Nambisan et al. [9], participation in global innovation ecosystems enables firms to access distributed knowledge and flexibly reconfigure capabilities — essential attributes of resilience. In this light, resilience is not merely a reactive trait but an emergent outcome of intentional ecosystem positioning. Ecosystem participation can reduce exposure to shocks by distributing risk, facilitating adaptive learning, and accelerating resource orchestration across boundaries [5], [9], [15], [21].

Building on this, it is reasonable to suggest that EEI-MNCs may actively construct their ecosystem strategies with resilience as a strategic intent. This implies that beyond accessing innovation inputs, firms may selectively engage in external multilateral ecosystems to enhance their long-term viability under systemic pressures, such as those captured by volatility indices. Understanding how ecosystem strategies relate to a firm's resilience trajectories requires unpacking their structural elements, especially those tied to entry considerations [5].

Accordingly, in this paper, we investigate the following research questions:

RQ1: How do ecosystem-intensive EEI-MNCs perform during periods of high market volatility?

RQ2: What architectural patterns of ecosystem entry strategies align with the historical resilience trajectories of ecosystem-intensive EEI-MNCs?

III. Methodology

To investigate the relationship between ecosystem strategies and organizational resilience, we selected EEI-MNCs with confirmed participation in global innovation ecosystems (GIES). These firms were identified based on their established presence in one or more ecosystem domains — manufacturing execution systems [5], renewable energy [10], [11], and circular innovation networks [12].

We examine historical resilience by analyzing abnormal returns (AR) during five major market volatility periods from 2015 to 2022, using the CBOE Volatility Index (VIX) to identify high-volatility windows. AR were computed by benchmarking each firm's daily stock performance against relevant industrial indices (e.g., S&P 500 Industrials, Nasdaq Europe Industrials, KOSPI 200). AR values were normalized per day, and the subsequent cumulative abnormal returns (CAR) were calculated as the sum of corresponding ARs across the five periods. Overall abnormal return across the entire period was also computed.

To assess ecosystem entry strategy architecture, we analyzed each firm's approach across three dimensions: (1) strategic discovery, (2) structural support, and (3) selective participation [5]. Each subcriterion was evaluated using a three-point scale: 0 – not confirmed, 0.5 – partially confirmed, and 1 – fully confirmed. Data sources included public disclosures, sustainability reports, and prior case studies.

IV. Findings and Implications

A. Performance Under Market Volatility

To assess the historical resilience of ecosystem-intensive EEI-MNCs, we analyze their abnormal stock performance during five major VIX spikes from 2015 to 2022. CAR is calculated relative to sectoral indices over these distinct market stress periods.

TABLE I. Major VIX Spikes and Event Characteristics (2015–2022)

Observed period	Days	Primary Shock	VIX Peak
17.08.2015 - 30.09.2015	41	China FX devaluation	40,74
01.02.2018 - 28.02.2018	37	Volmageddon (volatility derivatives crash)	37,32
20.02.2020 - 29.05.2020	83	COVID-19 global onset	82,69
15.10.2020 - 15.11.2020	40	US election, stimulus uncertainty, 2nd COVID wave	40,28
21.02.2022 - 30.09.2022	152	Russia-Ukraine war, supply chains and energy shocks	36,45

Table I outlines the five VIX event windows considered, covering both global systemic shocks and industry-specific disruptions.

TABLE II. Ecosystem-intensive EEI-MNCs' performance during periods of high market volatility

MNC	GIES	ESG	HQ	Baseline Index	CAR%	AR %
ABB	2	A	Switzerland	Nasdaq Europe Industrials	-0,470	27
Dell	1	B	USA	S&P 500 Industrials	-0,004	71
Emerson Electric	1	C	USA	S&P 500 Industrials	0,164	32
GE Vernova	2	B	USA	S&P 500 Industrials	0,430	-114
Honeywell	1	C	USA	S&P 500 Industrials	0,319	42
IBM	1	A	USA	S&P 500 Industrials	-0,227	-40
Philips	1	B	Netherlands	Nasdaq Europe Industrials	0,173	-28
Rockwell Automation	1	B	USA	S&P 500 Industrials	-0,320	63
Samsung Electronics	1	B	South Korea	KOSPI 200	0,043	172
Schneider Electric	3	A	France	Nasdaq Europe Industrials	0,162	142
Siemens	2	B	Germany	Nasdaq Europe Industrials	-0,101	57
Tesla	1	C	USA	S&P 500 Industrials	-0,138	1460

Abnormal returns were calculated by benchmarking stock performance against relevant industrial indices. To allow comparison across periods of unequal length, we compute daily AR and derive CAR per firm. Also, we calculate the

overall abnormal return for the whole studied period from 2015-08-17 to 2022-09-30. The corresponding data and calculations are available at the LINK.

The resulting performance analysis (Table II) reveals considerable variation in abnormal return performance across the sample of ecosystem-intensive EEI-MNCs. Schneider Electric exhibited the highest daily CAR and overall AR for the full 2015–2022 period, simultaneously demonstrating a high ESG ranking. IBM, GE Vernova, and Rockwell Automation showed moderately positive resilience outcomes. Tesla, in contrast, achieved exceptional AR despite negative daily CAR, suggesting a growth-driven investor effect rather than strategic resilience. Firms such as Philips and Emerson underperformed consistently. These observed differences establish the empirical basis for the typology of EEI-MNCs presented below.

B. *Ecosystem Entry Strategy Patterns*

To deepen our understanding of how EEI-MNCs strategically engage with innovation ecosystems, we examined their ecosystem entry strategies across three architectural components: strategic discovery, structural support, and selective participation. These elements represent how firms conceptualize ecosystem engagement, embed it within their organizational design, and selectively prioritize which ecosystems to enter or avoid.

Table III presents the results of this assessment. Schneider Electric, Honeywell, Rockwell Automation, IBM, GE Vernova, Siemens, and ABB demonstrate high levels of engagement across all three components, indicating a comprehensive ecosystem strategy embedded at the corporate level. These firms exhibit not only clear strategic framing but also strong structural mechanisms and purposeful participation practices. We refer to these as "ecosystem-intensive" in both intent and implementation.

TABLE III. ARCHITECTURAL PATTERNS OF ECOSYSTEM ENTRY STRATEGY OF THE OBSERVED ECOSYSTEM-INTENSIVE EEI-MNCS

MNC	EES1. Strategic Discovery	EES2. Structural Support	EES3. Selective Participation
ABB	1	1	1
Dell	0,5	0,5	1
Emerson Electric	0,5	0,5	1
GE Vernova	1	1	1
Honeywell	1	1	1
IBM	1	1	1
Philips	0,5	0,5	0,5
Rockwell Automation	1	1	1
Samsung Electronics	0,5	0,5	0,5
Schneider Electric	1	1	1
Siemens	1	1	1
Tesla	0,5	0,5	0,5

In contrast, Philips, Samsung, Dell, Emerson, and Tesla reveal partial engagement. Their ecosystem strategies tend to be ad hoc, fragmented, or primarily operational rather than strategic. For instance, Tesla and Samsung exhibit selective participation but lack formal framing or structural support, while Philips appears only moderately engaged across all three dimensions.

These patterns help differentiate firms that view ecosystems as strategic instruments for navigating technological dynamism from those where ecosystem involvement remains opportunistic or loosely coordinated. This foundation enables us to link architectural patterns to observed resilience outcomes, as discussed in the following section.

C. *Typology of Ecosystem-Intensive EEI-MNCs*

To better understand the strategic implications of ecosystem engagement, we classify the observed EEI-MNCs into four types based on their CAR, overall AR, ESG groupings, and architectural patterns of ecosystem entry strategy (see Table IV and Figure I).

TABLE IV. GROUPING OF EXTERNAL MULTILATERAL ECOSYSTEM-INTENSIVE EEI-MNCS BASED ON HISTORICAL RESILIENCE, SUSTAINABILITY PERFORMANCE AND ECOSYSTEM ENTRY STRATEGY

MNC	EES1	EES2	EES3	GIES	ESG	CAR%	AR %	Type
Schneider Electric	1	1	1	3	A	0,162	142	Proactive Leader
ABB	1	1	1	2	A	-0,470	27	Pivot
GE Vernova	1	1	1	2	B	0,430	-114	Pivot
IBM	1	1	1	1	A	-0,227	-40	Pivot
Rockwell Automation	1	1	1	1	B	-0,320	63	Pivot
Siemens	1	1	1	2	B	-0,101	57	Pivot
Dell	0,5	0,5	1	1	B	-0,004	71	natural
Emerson Electric	0,5	0,5	1	1	C	0,164	32	natural
Honeywell	1	1	1	1	C	0,319	42	natural
Philips	0,5	0,5	0,5	1	B	0,173	-28	natural
Samsung Electronics	0,5	0,5	0,5	1	B	0,043	172	natural
Tesla	0,5	0,5	0,5	1	C	-0,138	1460	Lucky Star

"Proactive Leader" is represented by Schneider Electric—a firm exhibiting consistent positive CAR and AR, full strategic engagement across all three ecosystem components, and top-tier ESG performance. This type exemplifies anticipatory resilience, in which the ecosystem strategy is proactively aligned with long-term sustainability and adaptation goals.

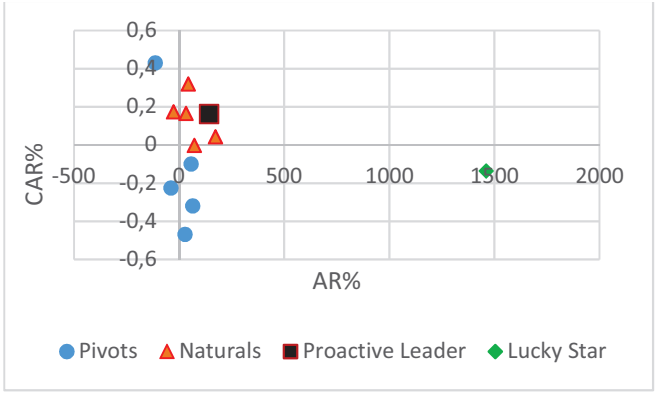


Fig. 1. Typology of external multilateral ecosystem-intensive EEI-MNCs based on historical resilience, sustainability performance and ecosystem entry strategy

"Pivots" include ABB, GE Vernova, IBM, Rockwell Automation, and Siemens. These firms show a mix of CAR and AR results but have fully developed ecosystem strategies. This suggests a possible strategic response to experienced volatility, leveraging ecosystem engagement as a resilience-enabling transformation. Notably, several pivots improved ESG standings over time.

"Naturals" such as Honeywell, Dell, Emerson Electric, Philips, and Samsung Electronics performed relatively well in terms of AR or CAR but exhibited only partial ecosystem strategy. Despite low or medium ESG scores, their business models appear inherently aligned with ecosystem value flows, even if not formally structured at the strategic level.

"Lucky Star" achieved outstanding AR but with a negative CAR and limited ecosystem strategy architecture. Its high market valuation may reflect speculative investor optimism more than resilient strategic fundamentals.

This typology emphasizes that ecosystem engagement may serve both as a deliberate strategy to build resilience (Proactive Leaders, Pivots) and as a latent capability that manifests in high performance (Naturals). The contrast with Tesla underlines that strong market returns do not always imply strategic resilience. Understanding this distinction is essential for international business theory and practice under conditions of increasing technological dynamism.

For the electrical and electronic industry, this typology provides a strategic lens to reassess competitive positioning. As firms in this sector navigate rapid digitalization, sustainability transitions, and geopolitical uncertainty, their ability to deliberately architect and govern ecosystem participation emerges as a potential differentiator of resilience. EEI-MNCs must recognize that ecosystems are no longer peripheral enablers but central arenas of co-innovation, risk-sharing, and adaptive capacity. The typology encourages EEI leaders to evaluate not only their performance outcomes but also the coherence and depth of their ecosystem strategies.

V. Conclusions

Faced with increasing technological dynamism and market volatility, EEI-MNCs are increasingly turning to global innovation ecosystems as strategic platforms for accessing, sharing, and developing complementary resources. This shift motivates the deliberate construction of ecosystem entry strategies aimed at enhancing resilience and sustaining global competitiveness. Grounded in the extended strategic tripod framework, we conceptualize ecosystem-specific advantages as a vital integration of RBV and IBV, operationalized through entry, relationship, and value chain orchestration strategies.

Our empirical analysis reveals a typology of EEI-MNCs—Proactive Leaders, Pivots, Naturals, and the Lucky Star—each reflecting distinct combinations of abnormal returns, ESG performance, and ecosystem entry strategy architecture. While some firms strategically align ecosystem participation with long-term adaptability, others perform well without formal ecosystem structuring.

These findings underscore that ecosystems are no longer auxiliary but strategic domains for value creation and resilience. For EEI-MNCs, deliberate strategic engagement in external multilateral global ecosystems may offer a critical pathway to sustained competitiveness under conditions of uncertainty.

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Error Estimation of the Interferometric Method of Measuring the Spatial Parameters of a Laser Beam

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***Abstract*— An interferometric method for measuring the spatial parameters of a laser beam is considered. The method implements the dependence of the beam wave front curvature on the distance to the waist in the far zone. The original laser beam is split into two beams interfering with each other. The period of the interference field carries information about the location of the waist. To determine the divergence angle, the width of the radiation beam in the plane of recording the period of the interference pattern is additionally measured. The location of the waist and the value of the divergence angle are calculated using the formulas of Gaussian optics. The results of experimental verification allow us to recommend this method as less labor-intensive and more accurate than known methods.**

Keywords*— *laser, Gaussian beam, waist location, wavefront curvature, shear interferometer, two-beam interference, period of the interference fringes

I. Introduction

A number of modern laser applications, such as photolithography [1], absolute ballistic gravimetry [2], and precision positioning systems [3], are needing to measure lengths and displacements in the tens of centimeters range with a relative error of about $1 \cdot 10^{-9} \div 1 \cdot 10^{-10}$. The dominant obstacle to ensuring such measurement accuracy is laser radiation diffraction. Theoretical calculations show that it is possible to reduce the measurement error caused by the diffraction of the laser radiation to the required level by using laser beams with a diameter of about 10 cm. However, in most applications, the use of laser beams of this diameter is practically unrealistic, and, therefore, there is a need for precision measurement of spatial parameters of the laser beam, which are necessary for quantitatively accounting for the effect of radiation diffraction on the magnitude of the result of measuring the displacements.

Spatial parameters are a fundamental factor in the creation and testing of various communication systems using lasers, optical location and range-finding systems, navigation systems [4], as well as in such industrial applications as laser welding [5] and cutting [6] in the aerospace and automotive industries, development of optical instruments [7], optical metrology [8] and laser therapy in medicine [9].

To describe the process of radiation propagation in space, a Gaussian model of a laser beam is used, for the specification of which it is necessary to specify three parameters [10] - the wavelength of the radiation, the location of the beam waist, which determines the origin of the coordinate system, and some spatial parameter, for example, the diameter of the waist, or the diameter of the beam in some section, or the angle of divergence, or the confocal parameter, or the radius of the Rayleigh zone. In practice, the angle of divergence is usually chosen as the most visual and more or less accessible parameter for measurement. Over the past period of development of laser technology, a number of methods have been proposed for measuring these parameters.

The main attention was paid to measuring the magnitude of the divergence angle, for which about twenty different methods and their modifications were proposed [11]. The very existence of such a number of methods indicates that none of them is sufficiently satisfactory. Almost all known methods are associated with photometry of the radiation beam and give a large measurement error for small divergence angles, which are of greatest practical interest.

Less attention is paid to determining the location of the waist in the literature [12], although this parameter plays an extremely important role in a number of cases.

The modern international standard ISO 11146-1/2/3:2021 [13-15] regulates the procedure for measuring the location of the laser beam waist and its divergence as follows. It is necessary to measure the beam width in at least 10 different sections on both sides of the waist, and the beam width in each section must be measured at least five times. Spatial parameters are calculated based on the measurement results according to the formulas given in the standard.

However, in practice, the waist is usually inaccessible for direct measurements. In this case, the standard prescribes using an aberration-free focusing element to form an artificial waist and carry out the above-described procedure of measurements and calculations. Then, the location of the original waist and the divergence angle of the original beam should be calculated according to [10], based on the parameters of the focusing element and by measuring the distances from its main planes to the artificial and original waists. The standard also provides for alternative methods for measuring beam widths:

— the adjustable diaphragm method,
— the Foucault knife method,
— the moving slit method,
which are used if necessary.

The interferometric technique for measuring the spatial parameters of a laser beam [16] is based on the relationship between the specified parameters and the parameters of the spatial interference pattern of two Gaussian radiation beams. In this case, the problem of determining the spatial parameters of a laser beam is reduced to using a preliminary calculation of the interference of two Gaussian beams. Limiting ourselves to considering the far zone, we can assume that the interference field of two Gaussian beams is similar to the interference field from two coherent point sources, i.e. is a family of nested two-sheet hyperboloids of revolution with foci at the location of the centers of the waists of Gaussian beams. In an arbitrary recording plane, a system of interference fringes is observed, the parameters of which, that is, the shape of the fringes, their width and direction, depend on the spatial orientation of the recording plane. The

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period of the interference fringes is determined by the distance from the origin of coordinates (that is, from the center of symmetry of the imaginary waists) to the registration plane, which is calculated by measuring the period of the fringes.

The aim of this work is to evaluate the accuracy of determining the spatial parameters of a laser beam using the interferometric method.

II. Theoretical Relationships

Analysis shows that of the many known systems for forming two interfering beams from one initial beam, only a system of two parallel planes is suitable. Any other system (Fresnel mirrors, Fresnel biprism, Billet bilens, Lloyd mirror, etc.) for calculating the interference pattern requires a preliminary determination of the spatial parameters of the radiation beam. The technical implementation of a system of two parallel planes can be a plane-parallel plate transparent to optical radiation or a Michelson interferometer.

In the case of using a plane-parallel plate, the beam incident on it at an angle α (Fig. 1) after reflection from the front and back surfaces of the plate produces two beams emanating from two imaginary sources. From considerations of symmetry it follows that the imaginary sources formed by the plate lie on the normal to the plane-parallel plate at the point of incidence of the beam. Based on geometric relationships and taking into account Snell's law, we obtain that when the plate is in the air, the distance $2C$ between the imaginary sources is equal to

$$2C = \frac{2d\cos\alpha}{(n^2 - \sin^2\alpha)^{\frac{1}{2}}} \tag{1}$$

where d is the thickness of the plate; n is the refractive index of the plate material; α is the angle of incidence of the laser beam on the plate.

From (1) it follows that the distance between the imaginary waists formed by the plane-parallel plate does not remain constant, but is a function of the angle of incidence of the radiation beam on the plate. Thus, the non-localized interference field in the radiation reflected from the plate, even in the far zone, differs from the classical interference field of two points sources, although both fields are families of hyperboloids of revolution.

When registering an interference field in an arbitrary plane, the interference fringes are the lines of intersection of this plane with a family of hyperboloids. The parameters of the observed interference fringes, namely their shape, width and direction, depend on the spatial orientation of the registration plane. The period of the interference fringes depends on the distance l from the imaginary waists to the plane of registration of the interference field. Let's obtain this dependence in explicit form.

The origin of the coordinate system is placed in the middle between the imaginary waists, the abscissa axis is drawn through the centers of both waists. We assume that the distance $2C$ between the waists is much less than the distance l from the imaginary waists to the plane of registration of the interference pattern:

$$2C << l.$$

In reality, two beams arrive at any point of the registration plane. The optical path difference between them is

$$\{(x + C)^2 + y^2\}^{1/2} - \{(x - C)^2 + y^2\}^{1/2} = 2C - m\,\lambda,$$

where x, y are the coordinates of the center of the interference fringe on the registration plane; $2C$ is the distance between the imaginary waists; m is the interference order; λ is the wavelength of the laser radiation. In canonical form, this equation has the form

$$4x^2 / (2C - m\,\lambda)^2 - 4y^2 / (4Cm\lambda - m^2\lambda^2) = 1. \tag{2}$$

As expected, this is the equation of a hyperbola. We write the equation for the registration plane as

$$x\cos(\alpha + \varphi) + y\sin(\alpha + \varphi) = l. \tag{3}$$

where φ is the angle of incidence of the laser beam on the registration plane. The solution of system (1) – (3) is the expression

$$l = h\,d\,\sin2\alpha\cos^2\alpha\,[\cos(\alpha+\varphi) + \mathrm{tg}\alpha\,\sin(\alpha+\varphi)]^2 / \lambda(n^2 - \sin^2\alpha)^{1/2}$$

where h is the period of the interference fringes in the recording plane. If the recording plane is positioned perpendicular to the interfering beams, then φ = 0 and the expression is simplified to

$$l = h\,d\,\sin2\alpha / \lambda\,(n^2 - \sin^2\alpha)^{1/2} \tag{4}$$

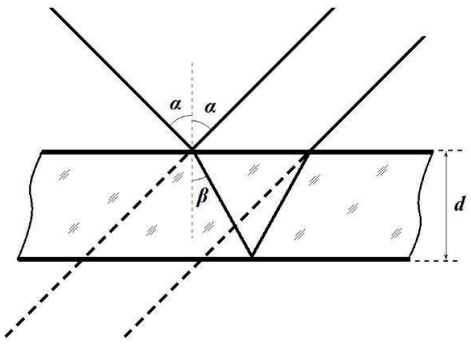


Fig. 1. Ray path in a plane-parallel plate

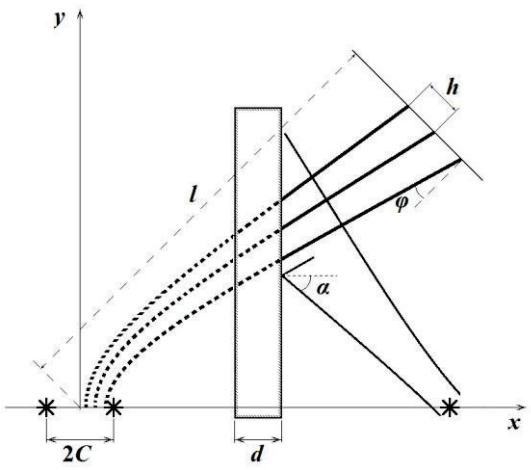


Fig. 2. Scheme of formation of interfering beams

If a Michelson interferometer is used, the interference pattern in the recording plane has the form of a family of concentric rings. In this case, the calculation is carried out using the formula

$$l = \{d_e (D_{i+1}^2 - D_i^2) / \lambda\}^{1/2},$$

where d_e is the thickness of the equivalent air plate of the Michelson interferometer (the difference in the optical path of its arms); D_i and D_{i+1} are the diameters of any two adjacent interference rings.

Thus, by measuring the period of the interference fringes (or the diameters of the interference rings) and the angle of incidence of the laser beam on a plane-parallel plate in the recording plane, it is possible to determine the distance from the plane of recording the interference pattern to the waist of the laser beam.

To calculate the divergence angle θ of the original laser beam, it is necessary to additionally measure its diameter in the plane of recording the period of the interference fringes. This measurement is not difficult, since the cross-section of the laser beam in the plane of recording the interference pattern is large enough, which allows using any method recommended by the current standards. The calculation of the divergence angle is carried out in accordance with the recommendations of the ISO 11146-1:2021 standard.

Differentiating (4), we determine the relative error of the considered interferometric method

$$\delta l/l \approx \{(\delta h/h)^2 + (\delta d/d)^2 + \delta\alpha^2 + (\delta\lambda/\lambda)^2 + \delta n^2\}^{1/2}$$

Assuming that $\delta h/h \approx 10^{-3}$, $\delta d/d \approx 10^{-4}$, $\delta\alpha \approx 10^{-4}$, $\delta\lambda/\lambda \approx 10^{-6}$, $\delta n \approx 10^{-4}$ we obtain $\delta l/l \approx 10^{-3}$.

Almost all known methods for measuring spatial parameters are based on measuring the dimensions of the cross sections of the laser beam. For beams with low divergence, the diameters of the cross sections differ little from each other, which leads to a large measurement error of both the location of the beam waist and its divergence angle. As the divergence angle decreases, the efficiency of known methods decreases to the point of complete loss of information. The recommendation of the existing standard to increase the number of measurements is ineffective in terms of reducing the error in the measurement result, but significantly increases the labor intensity of the measurement process. The recommendation to form an artificial waist in the case of inaccessibility of the real waist for direct measurements also does not contribute to increasing the accuracy of the measurement result. The fact is that as the divergence angle of the radiation beam decreases, its diameter increases proportionally. As a result, the aberrations of the focusing element, distorting the wave front of the beam, increase, and the measurement error increases.

The informative parameter of the considered interferometric method - the period of the formed interference pattern - increases as the divergence angle decreases, which makes it possible to expand the measurement range towards small divergence angles that are inaccessible to measurement by known methods. The same effect can be obtained by changing the thickness of the plate.

Another decisive advantage of the interferometric method is its significantly lower labor intensity, due to the ability to perform one or two measurements instead of several dozen measurements recommended by the current standard.

III. Experimental Verification

For experimental verification the following equipment was involved: He-Ne laser LG-56 (wavelength 0.632816 μm); plane-parallel plate (20.75 mm thick) with optical glass K8 (n = 1.514627); collimator from the optical bench OSK-2 (focal length 1600 mm); goniometer GS-5; two geodetic poles of the 1st category (a nominal depth 1 meter); indicator HI-50 18–50; microdensitometer MF-2; screens, adjustable tables, sheet film.

As known the beam waist of the LG-56 laser is located on the surface of the output flat mirror. The localization the beam waist was determined both by the method of transection and the interferometric method. Then the beam divergence was changed by a collimator, and the new location of the beam waist was measured again by both methods. Consistent using of this procedure made it possible to select and adjust a beam waist positions in the angle range from 10' to 3".

The measuring of the laser beam width by the method of cross-cutting was carried out by additional screens with sheet photo-film, which were sequentially installed at distances from the waist at 5 m to 75 m with a step of 5 m. The distance from the exit laser mirror to photo-film screen was determined by the method sequential transfer of two geodetic rods with a nominal length of 1 m. After processing the film, the spot diameter drawn on it was measured at half intensity level using an MF-2 microdensitometer. The location of the waist was determined according to the ISO 11146-1 standard by approximating the results by hyperbolic function.

The localization of the waist by the interferometric method was carried out using a standard plane-parallel plate. The plate was installed on the rotating table of the goniometer in such a way that the front surface of the plate was aligned with the table axis of rotation. The incidence angle of the laser beam on the plate was varied between 0° ÷ 45°. The distance from the output mirror of the laser to the front surface of the plane-parallel plate was 74 m and was determined, as in the first case, by the method of sequential transfer of geodetic rods. A typical appearance of the obtained interference patterns is shown in Fig. 3. In a typical case, the plate was mounted on the goniometer table at an angle of 45° to the laser beam. The screen with a photographic film was mounted at a distance of 1 m from the axis of the goniometer table, so that the interfering beams reflected from the plate fell normally to the plane of the screen. Thus, the distance from the laser output mirror to the plane of recording the interference fringes was 75 m.

When measuring the diameter of the laser beam, the screen with the film was mounted on the goniometer table and the plane of the sheet film was aligned with the axis of rotation of the table. The goniometer was moved so that the distance from the laser output mirror to the axis of rotation of the goniometer table was 75 m. This ensured that the distances for measuring the period of the interference fringes and the diameter of the laser beam were equal.

As a result of testing the following was established:

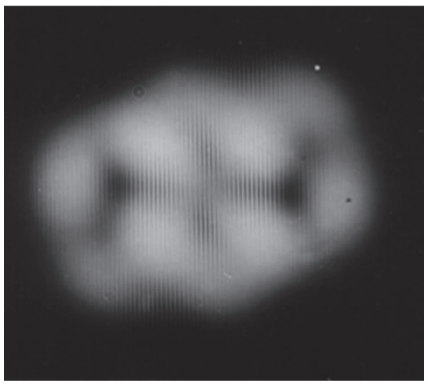
— for the radiation beam of the LG-56 laser with a divergence angle of ~ 10', both methods give results that coincide with each other with an accuracy of about 2 %;

— the section method recommended by ISO 11146-1 is advisable to use to determine the location of the beam waist with a divergence angle exceeding 30", while the

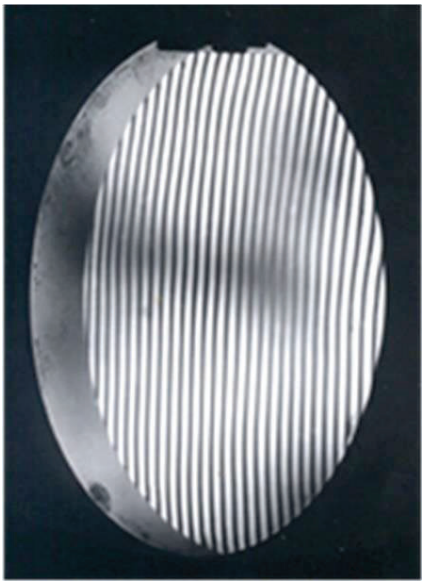
measurement error is about 10 %.

— the interference method is advisable to use to determine the location of beam waists with divergence angles exceeding 5", while the measurement error is no more than 5 %.

— the interferometric method is characterized by significantly less labor intensity



a)



b)

Fig. 3. Interference patterns for beams with a) θ = 8'15"; b) θ = 1'48" (reduced)

IV. Conclusions

An interferometric method for determining the location of a laser beam waist has been developed, which implements the known interdependence between the curvature of the wavefront in a certain cross-section of the radiation beam and the distance from this cross-section to the beam waist.

The conducted studies have confirmed the operability of the interferometric method for determining the location of a laser waist, and the results obtained allow us to recommend this method as the most accurate among those known at the present time.

The process of laboratory research was carried out with the involvement of rather complex and bulky universal optical devices and precision measuring instruments. The development of a specialized device controlled by a microcontroller will significantly reduce the time of a single measurement by simplifying the operations of optical adjustment, setting the measurement angle, the process of scanning the interference pattern and processing the obtained data.

It should also be noted that due to the significantly greater complexity of operation, the Michelson interferometer is advisable to use for unique measurements in cases where the use of wide-angle plane-parallel plates of great thickness and weight is limited by the technological capabilities of their manufacture. The use of the Michelson interferometer is focused on prospective needs, while the existing level of needs can be fully satisfied by using those plane-parallel plates, the industrial production of which has already been mastered at this time.

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Software Tool for Evaluation of AI-Generated BPMN Models using Structural Measures

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***Abstract*—Business process modeling is a key technique in enterprise management. It assumes graphical description of activities, events and decisions that occur in organizational and information systems. Therefore, business process models should be of high-quality to ensure efficient workflow design, analysis, and improvement. However, the growth in popularity of AI-assistants based on large language models, opens the possibility for business process model generation using text descriptions. Nevertheless, the quality and usability of such models requires further study. Therefore, this paper presents the software tool for evaluation of AI-generated Business Process Model and Notation (BPMN) models from the structural perspective using related measures, as well as the comparison with the human-designed models.**

Keywords—business process modeling, large language model, structural measure, model evaluation, artificial intelligence

I. Introduction

Business process modeling is an important technique for managing and improving the enterprise activities. Thanks to the graphical description of actions, events, and decisions, it facilitates the understanding of organizational workflows by all the involved stakeholders [1]. Business process modeling allows not only to describe existing workflows, but also to analyze their efficiency and identify potential directions for improvement [2].

Well-designed business models help to better control the execution of activities, track events, and manage information flows [3]. High-quality business models are crucial for the successful implementation of Business Process Management (BPM), as well as the other enterprise information systems, so evaluating their structure is a necessary step in the BPM life-cycle [4].

Hence, structural measures and corresponding thresholds are typically used to evaluate the quality of business process models. Such measures include size, complexity, and other dimensions, being used to determine, to what extent the model is understandable, maintainable, and corresponds to modeling guidelines [4].

Thus, this paper introduces a software tool for evaluating business process models generated by Artificial Intelligence (AI) assistants based on Large Language Models (LLMs). The study is focused on the use of structural measures to analyze the quality of business process models.

II. Related Work

Relevant research on business process model evaluation pays significant attention to the use of structural measures and compliance with modeling guidelines and rules. Determining the quality of a business process model is often based on the compliance of its elements with the recommendations, such as the 7PMG, SEQUAL, and other modeling frameworks [5]. Such approaches allow formalizing the evaluation of business process models, by focusing on the clarity, completeness, and correctness of model structure.

Business process size and complexity measures provide a deeper understanding of the structural quality of a model [6]. These measures allow evaluating the understandability and maintainability of a given business process model.

Structural metrics, such as Number of Activities (NOA) and Control Flow Complexity (CFC), are widely considered in related studies. NOA describes the total number of activities in a model, while CFC assumes the control flow complexity using AND, XOR, and OR gateways [7].

Authors of [8] proposed a method for defining thresholds for business process model measures. The authors used data mining methods to distinguish high-quality business process models. Thresholds were set for NOA, CFC, and gateways-related measures. The similar approach is proposed in [9], where business process models were evaluated based on the structural measures and corresponding thresholds.

The more advanced approach proposed in [10], assumes the list of 50 guidelines and corresponding measures to ensure improvement of business process models understandability and the overall quality. Such threshold values derived from the industry standards can be used for automated business process model validation [11]. Paper [12] also focuses on the practical application of business process modeling measures.

While the most of existing studies focus on the evaluation of human-designed business process models, this research is focused on the features of AI-generated models, taking into account the structural quality and comparison with manually created versions of such models.

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III. Problem Statement

Among the various business process modeling notations, the most popular are Business Process Model and Notation (BPMN), Event-driven Process Chain (EPC), and IDEF-based notations [13]. However, in the last decade, BPMN became the dominant business process modeling standard due to its flexibility, clarity, and complex logic support [14].

BPMN provides a convenient way to visualize workflows through events, tasks, and gateways that describe the logic of workflows. The use of gateways in BPMN allow determining branching scenarios, i.e. parallel, alternative, or conditional paths within a workflow [15].

The quality of BPMN models directly affects the quality of workflows in organizational and information systems, as error-prone models of high complexity may cause errors in real business processes [16].

With the tremendous growth of generative AI services and LLMs popularity, it became possible to automatically create BPMN models based on raw business process descriptions, which requires a thorough study of the quality of such models.

Hence, in this study the problem of AI-generated business process models evaluation using the framework demonstrated in Fig. 1 is considered.

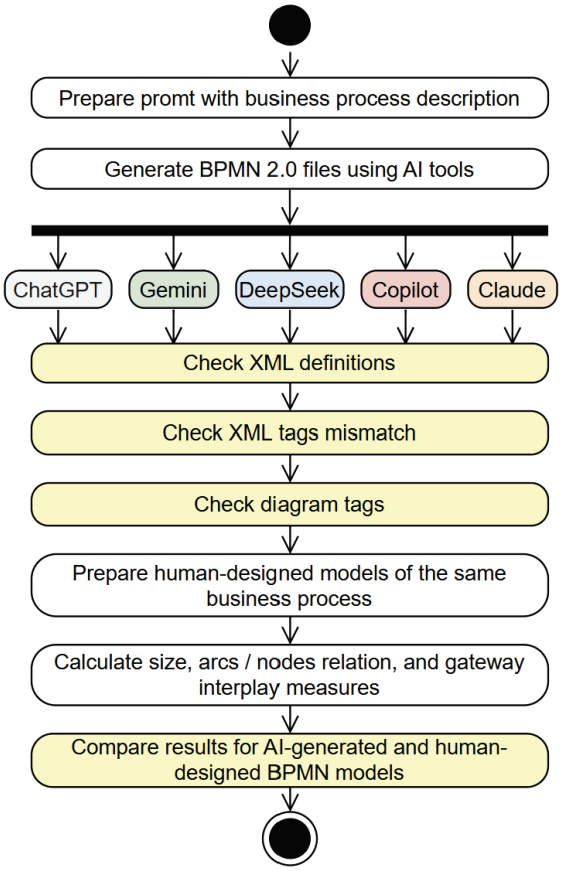


Fig. 1. AI-generated BPMN models evaluation framework

As can be seen from Fig. 1, this study aims to evaluate the quality of BPMN models generated from raw business process descriptions by generative AI-services based on LLMs, such as ChatGPT, Gemini, DeepSeek, Copilot, and Claude.

At first, the set of BPMN 2.0 files is generated based on the business process description. The XML structure of these BPMN 2.0 files is checked for correctness (definitions, tags matching) and completeness (diagram tags for visualization). Finally, AI-generated models are compared to the models of the same business process, but designed by humans, based on the structural measures analysis.

Therefore, the study aims to answer the research question: how different AI-assistants based LLMs are good in business process modeling using BPMN?

IV. Materials and Methods

The goods dispatch business process description and the set of human-designed BPMN models for comparison is taken from the open-source GitHub repository by Camunda [17].

Fig. 2 below demonstrates the textual description of such goods dispatch business process, occurring in the small-size hardware company. The outlined description is appended with the instructions for LLMs to build a user prompt (Fig. 2).

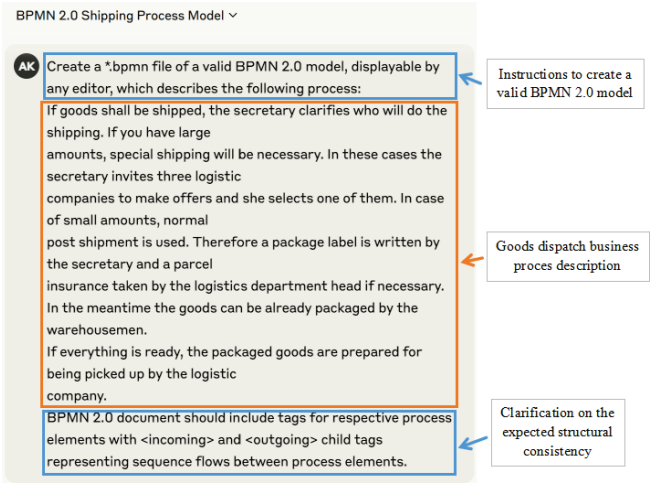


Fig. 2. Used prompt example to generate goods dispatch BPMN models

Despite the screenshot demonstrates the prompt submitted to Claude (Fig. 2), the same prompt is used with other LLMs to generate BPMN models (ChatGPT, Gemini, DeepSeek, and Copilot).

Table 1 below demonstrates the issues detected at the step of XML-structure checking of generated BPMN models.

TABLE I. Issues of Processed BPMN 2.0 Files

BPMN 2.0 Document Issues	AI Tools				
	ChatGPT	*Gemini*	*DeepSeek*	*Copilot*	*Claude*
Invalid XML definitions	Yes	Yes	Yes	No	Yes
Mismatched XML tags	Yes	Yes	Yes	No	Yes
Missing diagram tags	Yes	No	Yes	Yes	Yes

Thus, most errors were found in BPMN 2.0 models created with ChatGPT, Gemini, DeepSeek, and Claude (Table 2), all of which contain invalid XML definitions, as well as errors in tags matching or the absence of diagram tags used to visualize graphical BPMN definitions by model editors.

However, Copilot is the only AI-tool that did not produce mistakes in XML definitions and tag matching, although it did

not provide diagram tags. On the other hand, Gemini is the only LLM that included diagram tags to the generated BPMN model, which allows to open it in any graphical editor.

Fig. 3 demonstrates the AI-generated BPMN model of the goods dispatch business process, created using Gemini.

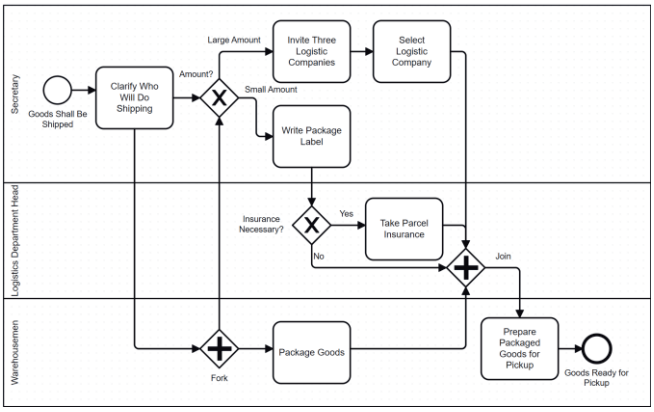


Fig. 3. AI-generated BPMN model of the goods dispatch process

For the comparison purposes, Fig. 4 outlines the human-designed BPMN model of the goods dispatch process taken from the Camunda GitHub repository [17].

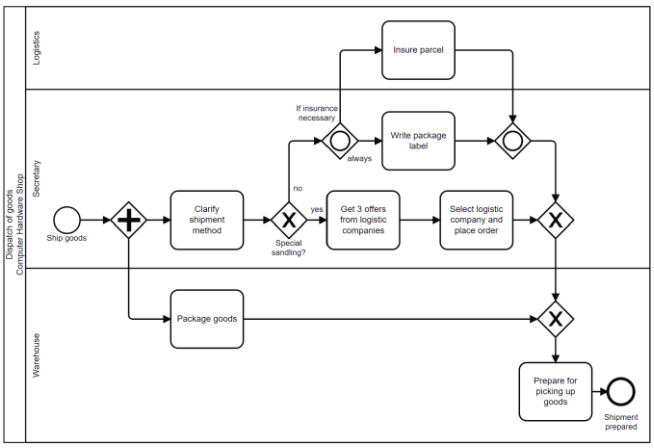


Fig. 4. Human-designed BPMN model of the goods dispatch process

The following structural measures [18] were selected to evaluate BPMN models:

- Size measures: the number of nodes – $|N|$ (i.e. the business process elements), the number of arcs – $|A|$ (i.e. the sequence flows between elements).
- Coupling measures: the density – Δ, the coefficient of connectivity – *CNC*, the average degree of gateways – d_G^{avg}, and the maximum degree of gateways – d_G^{max}.
- Complexity measures: the gateway mismatch – *MM* and the control-flow complexity – *CFC*.

The BPMN modes with high Δ and *CNC* is more likely to contain errors, since a person should take into account more connections between business process elements [18]:

$$\Delta = |A| / [|N| \cdot (|N| - 1)], \quad (1)$$

$$CNC = |A| / |N|. \quad (2)$$

The same rationale also works for gateway-based coupling measures, the average degree of gateways [18]:

$$d_G^{avg} = (1 / |G|) \cdot \sum_{g \in G} d(g), \quad (3)$$

$$d_G^{max} = \max\{d(g) \mid g \in G\}, \quad (4)$$

where g is the gateway under analysis, $g \in G$, G is the set of BPMN model gateways, and $d(g)$ is the number of sequence flows connected to the gateway g.

The BPMN model with a high gateway mismatch is likely to contain errors, since parallel scenarios are not synchronized or alternative flows run into deadlocks [18]:

$$MM_j = |\sum_{g \in j\text{-}Split} d(g) - \sum_{g \in j\text{-}Join} d(g)|. \quad (5)$$

$$MM = \sum_{j \in \{AND,\, XOR,\, OR\}} MM_j. \quad (6)$$

where MM_j is the mismatch of gateways of the particular type j (i.e. AND, XOR, OR), *j-Split* is the set of split gateways of type j, *j-Joins* is the set of join gateways of type j.

The BPMN model with a high *CFC* is more likely to have errors, due the difficulty to consider all possible states after split gateways of different types [18]:

$$CFC = \sum_{j \in \{AND,\, XOR,\, OR\}} \sum_{g \in G} CFC_{j\text{-}Split}(g), \quad (8)$$

$$CFC_{AND\text{-}Split}(g) = 1, \quad (8)$$

$$CFC_{XOR\text{-}Split}(g) = out\text{-}d(g), \quad (9)$$

$$CFC_{OR\text{-}Split}(g) = 2^{out\text{-}d(g)} - 1. \quad (10)$$

where *out-d*(g) is the number of outgoing sequence flows for the gateway g.

Fig. 5 demonstrates the software design of the proposed tool, developed using Python and its standard libraries for file processing (os), XML documents parsing (xml), and result data storing (csv).

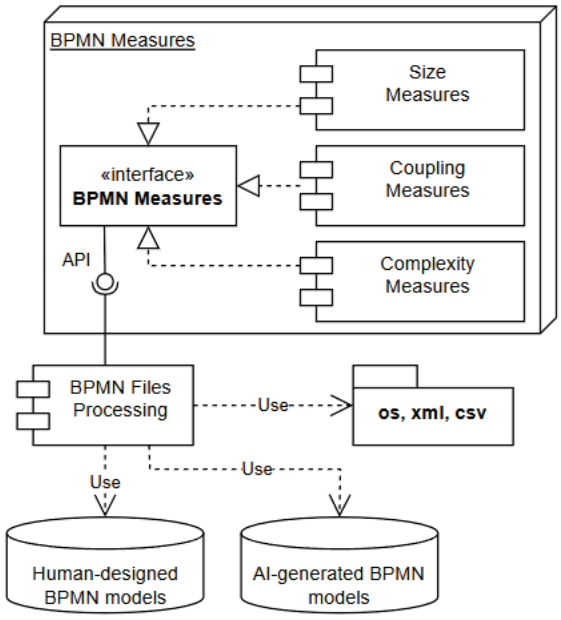


Fig. 5. Software tool design for evaluation of BPMN models

V. Results and Discussion

The developed software (Fig. 5) is used to measure BPMN models that describe goods dispatch process. The 67 human-designed BPMN models from the Camunda repository [17] and 5 AI-generated BPMN models (created using different AI-tools: ChatGPT, Gemini, DeepSeek, Copilot, and Claude) are considered for the analysis.

The average values of structural measures obtained for the human-designed business process models are used to compare with the AI-generated models.

Fig. 6 demonstrates size measures – $|N|$ and $|A|$.

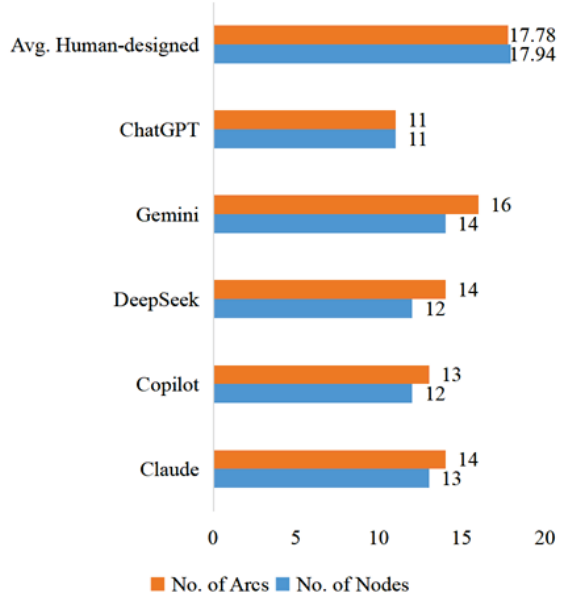


Fig. 6. Size measures $|N|$ and $|A|$ assessment results

It is demonstrated in Fig. 6, that human-designed models have the higher average values of $|N|$ and $|A|$ than values of $|N|$ and $|A|$ obtained for AI-generated models.

Fig. 7 shows coupling measures – Δ and *CNC*.

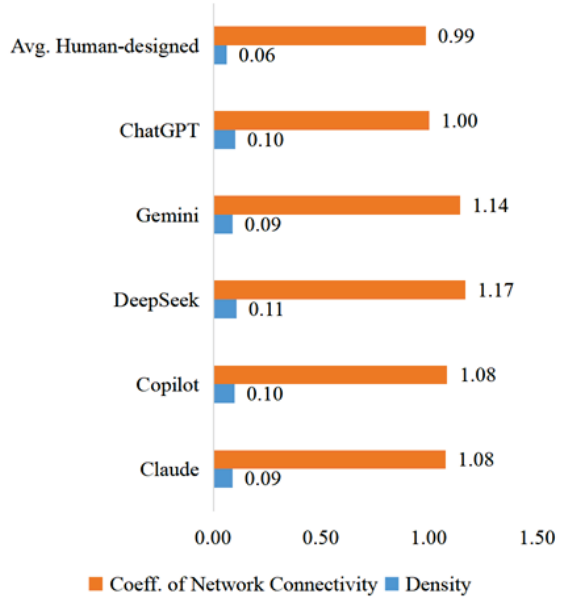


Fig. 7. Coupling mesures Δ and *CNC* assessment results

As can be seen from Fig. 7, the average values of Δ and *CNC* obtained for human-designed models are slightly lower than values of Δ and *CNC* obtained for AI-generated models.

Fig. 8 outlines coupling measures – d_G^{avg} and d_G^{max}.

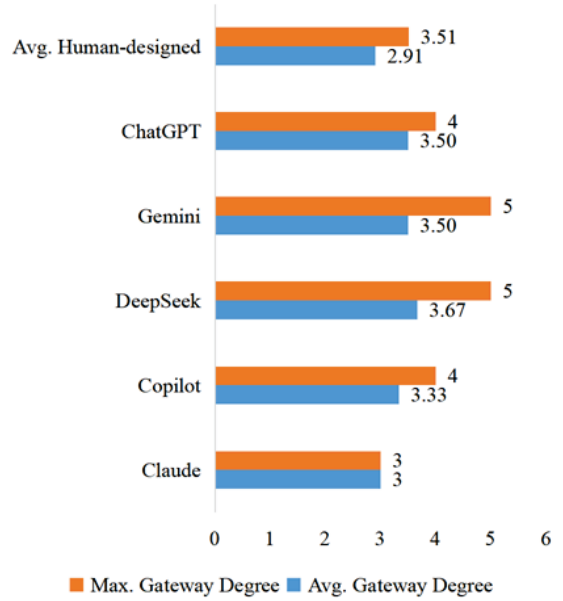


Fig. 8. Coupling mesures d_G^{avg} and d_G^{max} assessment results

It is depicted in Fig. 8, that human-designed models have lower average values of d_G^{avg} than values of d_G^{avg} obtained for AI-generated models. However, a model created using Claude shows lower d_G^{max} (3.00) than the average value (3.51) for human-designed models and its d_G^{avg} (2.91) is close to the average value (3.00) for human-designed models as well.

Fig. 9 depicts complexity measures – *MM* and *CFC*.

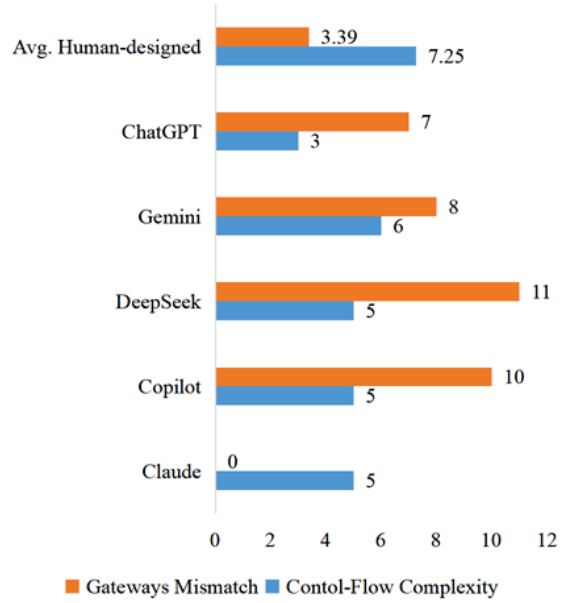


Fig. 9. Complexity measures *MM* and *CFC* assessment results

It is seen in Fig. 9, the average values of *MM* obtained for human-designed models are lower than *MM* values obtained for AI-generated models, except the one created using Claude having "perfect" *MM* of 0.00. However, the average value of *CFC* for human-designed models is higher than *CFC* values obtained for AI-generated models.

The relative values of considered structural measures (1 – 10) are calculated as follows:

$$x_{ij}' = x_{ij} / \max\{x_{ij} \mid j = 1, 2, \ldots, m\}, i = 1, 2, \ldots, n, \quad (11)$$

where x_{ij} is the value of i-th measure obtained for j-th category, n is the number of measures, (n = 8), and m is the number of categories (either human-designed or generated using LLMs, m = 6).

Since the high values of all considered measures indicate that business process models more likely to contain errors, the obtained relative values (11) could be considered as relative error likelihood values to evaluate BPMN models (Table 2).

TABLE II. RELATIVE ERROR LIKELIHOOD VALUES

BPMN Measure	AI Tools					
	Human	*ChatGPT*	*Gemini*	*DeepSeek*	*Copilot*	*Claude*
$\|N\|$	1.00	0.61	0.78	0.67	0.67	0.72
$\|A\|$	1.00	0.62	0.90	0.79	0.73	0.79
Δ	0.58	0.94	0.83	1.00	0.93	0.85
CNC	0.84	0.86	0.98	1.00	0.93	0.92
d_G^{avg}	0.79	0.95	0.95	1.00	0.91	0.82
d_G^{max}	0.70	0.80	1.00	1.00	0.80	0.60
CFC	1.00	0.41	0.83	0.69	0.69	0.69
MM	0.31	0.64	0.73	1.00	0.91	0.00

Fig. 10 demonstrates the average relative error likelihood estimates calculated based on analyzed structural measures for human-designed and AI-generated models.

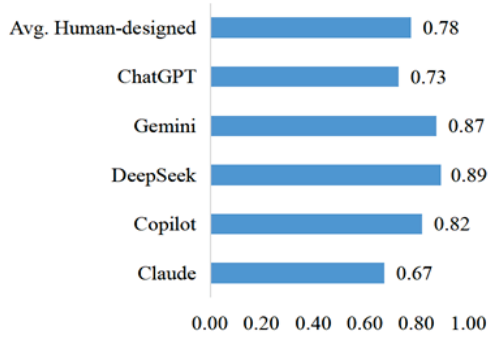


Fig. 10. Average relative error likelihood estimates

The comparison in Fig. 10 outlines models created using DeepSeek and Gemini have the highest average relative error likelihoods (0.89 and 0.87). The lowest average relative error likelihood is shown by Claude-designed model (0.67).

VI. CONCLUSION AND FUTURE WORK

This study proposes a software tool for quality evaluation of BPMN models generated by different AI-assistants based on LLMs, by comparing with human-designed models using structural measures. The obtained results demonstrate that most AI-generated models provide higher values of structural measures, which indicates a higher error likelihood compared to human-designed models. Thus, to effectively use LLMs in business process modeling, the additional evaluation of AI-generated BPMN models is necessary.

In the future, the developed software should be appended with the graphical user interface and deployed as a web-based tool, to be used by a wider audience of business analysts and other practitioners.

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Energy Arbitrage for Microgrids with Energy Storage Systems

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Abstract — The growing integration of renewable energy sources (RES) in Ukraine's power system has improved sustainability but introduced grid stability challenges due to their intermittency. Energy Storage Systems (ESS), particularly within microgrids, offer a viable solution by enhancing flexibility and economic efficiency. This study examines the potential of ESS-based price arbitrage under Ukraine's liberalized electricity market. A threshold for profitable operation is derived based on capital costs, efficiency, and battery degradation. A game-theoretic model is introduced to optimize ESS strategies in a competitive environment. Simulation results show that arbitrage-enabled ESS can achieve a payback period of under seven years. Cooperative strategies further improve investment returns, highlighting the role of ESS in advancing decentralized energy systems.

Keywords — energy storage systems, microgrid, gamr theory, energy arbitrage.

I. INTRODUCTION

Over the past decade, Ukraine has witnessed a dramatic transformation in its renewable energy landscape, driven by both policy incentives and rapid technological progress. A pivotal factor in this development was the introduction of the "green" tariff policy, which, alongside the global decline in capital expenditures for renewable energy sources (RES), significantly lowered the entry threshold for investors. According to the Ukrainian Association of Renewable Energy, by the end of 2019, approximately $10 billion had been invested in the renewable energy sector, with a staggering $6.2 billion drawn in just the 2018–2019 period [1].

This investment surge can be attributed to multiple converging factors. Technological advances, particularly in solar photovoltaic (PV) systems, led to substantial cost reductions [2,5] (Fig. 1). At the same time, a mature ecosystem of developers, equipment suppliers, and technical specialists emerged, enhancing market competitiveness and creating favorable lending conditions through banks and financial institutions. Although the most economically advantageous period for investing under the green tariff was in 2013 - when the tariff rates were highest - the peak in actual investments occurred later, as confidence in the sector's viability solidified.

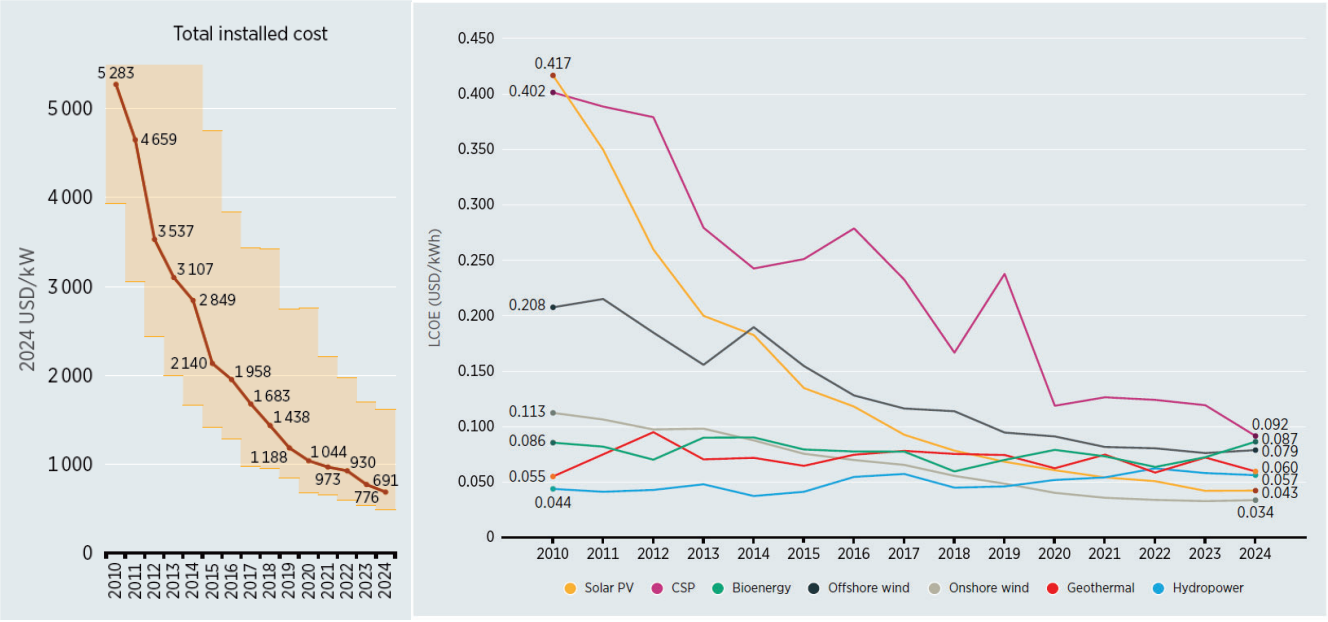
In the current energy landscape, with rising electricity prices and grid distribution tariffs, new RES projects are increasingly being implemented not for tariff-based returns but for self-consumption and energy cost reduction. These projects can be commissioned in relatively short timeframes, even for large-scale facilities. This evolution reflects a broader shift in investment logic, where energy independence and resilience are becoming as valuable as direct financial return.

However, the rapid growth of variable renewable energy sources has introduced new challenges to grid stability and power quality. As the share of RES in Ukraine's energy mix continues to increase, so too does the importance of ensuring their controllability and predictability [3, 4, 25]. This has placed energy storage systems (ESS) at the forefront of the next phase of energy infrastructure development. Just as early RES investors catalyzed the sector's expansion by proving financial feasibility, it is now crucial to demonstrate viable and profitable models for investment in ESS technologies.

One of the most promising economic mechanisms for enhancing the value proposition of ESS is energy arbitrage - the strategic buying and storing of electricity when prices are low and selling or using it when prices are high. Within microgrid environments, where localized generation and storage resources can be coordinated with greater flexibility, energy arbitrage offers a practical path toward improving the return on investment in storage assets while supporting grid efficiency and resilience.

This paper explores the role of energy arbitrage in optimizing the economic performance of microgrids equipped with energy storage systems. The aim is to develop a comprehensive model for arbitrage operations, analyze its economic impacts under various market conditions, and identify best practices for integrating ESS into microgrid architectures in Ukraine and similar emerging energy markets.

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a) *b)*

Fig. 1. Price trend: a) decrease in PV costs since 2010 tillo 2024; b) decrease in LCOE for RES since 2010 till 2024

II. MICROGRIDS AND ENERGY STORAGE SYSTEMS

Microgrids have become a cornerstone of distributed energy resource (DER) integration, offering operational autonomy and enhanced reliability. A critical component in modern microgrids is the Energy Storage System (ESS), particularly lithium-ion batteries, which help stabilize frequency, perform peak shaving, support load balancing, and mitigate the intermittent nature of renewables like solar PV and wind. The cost of ESS has declined significantly, making their deployment increasingly viable for both technical and economic purposes. ESS enables not only grid support but also energy arbitrage, where electricity is stored during low-price periods and discharged when prices rise. Several studies underscore this dual technical-economic role of ESS in improving microgrid performance and enabling value stacking across services [6–8].

A. Energy Arbitrage in Power Systems

Energy arbitrage in microgrids is gaining attention due to its potential to reduce electricity costs and enhance revenue streams. Traditional arbitrage models often assume deterministic conditions, optimizing charge/discharge cycles based on forecasted price profiles. However, price volatility, load uncertainty, and renewable intermittency require more robust modeling techniques.

For instance, in [7] authors use stochastic dynamic programming to manage battery arbitrage under variable efficiency and real-time pricing, achieving near-optimal profits even under uncertainty [9]. In [6] authors explore deep reinforcement learning to enable adaptive arbitrage strategies that learn optimal policies from price environments [10].

These strategies are often modeled via [11-14]: Mixed-Integer Linear Programming (MILP) for dispatch optimization, Model Predictive Control (MPC) for receding-horizon management, Stochastic Optimization for probabilistic market environments, Robust Optimization for worst-case planning.

B. Game-Theoretic Models for Energy Arbitrage and ESS Coordination

As multiple stakeholders (e.g., microgrid operators, storage owners, and utility grid operators) interact within a shared infrastructure, game theory has emerged as a powerful tool to model strategic behavior and pricing dynamics.

1) Non-Cooperative Games

Authors in [10] present a non-cooperative game among energy storage units engaging in local energy trading, where each participant aims to maximize profit under auction-based pricing. Their approach shows that game-theoretic arbitrage yields higher efficiency than centralized control models [15].

2) Stackelberg and Leader–Follower Games

In leader–follower (Stackelberg) models, a dominant player (e.g., storage aggregator or energy seller) sets the strategy first (e.g., price or capacity), and followers (e.g., consumers or prosumers) respond. For example, a solar developer in a community microgrid may act as a leader, pricing energy sales to influence demand profiles [16]. These models are essential for capturing market power dynamics and aligning incentives across tiers.

3) Evolutionary and Real-Option Games

Recent works apply real-option and evolutionary game theory to model investment decisions in ESS under subsidy uncertainty. These models help identify socially optimal incentive structures and ESS adoption patterns in emerging economies [17, 18].

4) Cooperative Games

In multi-microgrid environments, cooperative games are used to allocate shared benefits from ESS or joint arbitrage. Using Shapley value or core stability concepts, researchers like [13] design fair cost-sharing mechanisms to encourage joint ESS deployment across interconnected systems [19].

C. Research Gaps

Despite progress, several key research gaps persist:

Context-specific modeling: Most studies are based on Western markets and don't account for the regulatory, economic, and technological conditions in Ukraine.

Investment-focused analysis: Game-theoretic and optimization models often overlook Return on Investment (ROI), payback periods, and investor behavior.

Battery lifecycle and degradation: Degradation models are underrepresented in arbitrage frameworks, even though battery aging significantly affects profitability.

Regulatory integration: Few arbitrage models align with evolving tariff regimes, particularly in post-"green tariff" economies like Ukraine.

Hybrid frameworks: There is a lack of integrated models combining game theory, optimization, and uncertainty handling under one cohesive strategy.

This paper seeks to address these gaps by developing a simulation-based arbitrage model tailored to Ukraine's market structure, focusing on optimal ESS operation, economic performance, and strategic investor behavior.

Influence of United Power System's Structure on Price Arbitrage in Ukraine's Electricity Market

The volatility of the spot electricity market is due to its structure. The mode of operation of the United Power System (UPS) of Ukraine is determined on the basis of the balance of electricity production and consumption, repair of electrical networks and power generating equipment, as well as the possibility of eliminating emergencies in case of shutdown of electric power equipment at power plants and damage to main power lines. There are 413 electricity producers operating in the domestic energy system, of which seven powerful power generating companies provide about 90% of the total production, 40 enterprises for the transmission of electricity by local electric networks and 147 companies for the supply of electricity [20] (see Fig. 2).

As of November 1, 2019, the main generating capacities within the UPS of Ukraine are focused on:

• 4 nuclear power plants (NPPs) with 15 power units, of which 13 are 1,000 MW each and 2 are 415 and 420 MW;

• cascades of 8 hydroelectric power plants on the Dnieper and Dniester rivers with a total number of hydroelectric units – 103 units, as well as three pumped-storage plants (PSPPs), of which 11 hydroelectric units with a capacity of 33 MW to 324 MW;

• 12 TPPs with units with unit capacity of 150, 200, 300 and 800 MW, of which 75 power units, including power units: 150 MW – 6, 200 MW – 31, 300 MW – 32, 800 MW – 6 and 3 turbine generators, as well as 3 large CHPs with power units of 100 (120) MW and 250 (300) MW;

• 147 solar power plants (SPPs), 27 biogas power plants (BioPP), 19 wind power plants (WPPs), the total percentage of electricity generation of which is 3.7% of the total production in the UPS of Ukraine, but their share is constantly increasing.

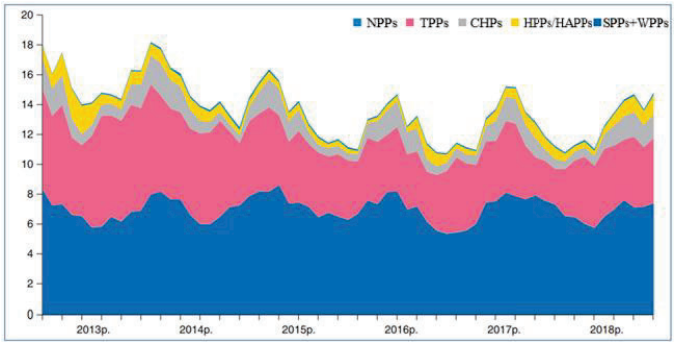


Fig. 2. Annual structure of generation in Ukraine (before the war)

To ensure the stable operation of the UPS of Ukraine, the base power plants should be 50-55%, semi-peak – 30-35%, peak (the most controlled) – 15%. Today, there is a shortage of peak capacities in the national energy system (10% against the required 15%). As a result, TPP power units are quite actively involved in regulation [21].

Although the data are provided as of November 2019, the structure of generating capacities has not undergone radical changes and the possibility of its restructuring in the medium term is not expected.

On the other hand, the load is also characterized by significant unevenness, the ratio of night load and load during the evening peaks is within 0.7. Taking into account the structure of generating capacities described above, this unevenness complicates the task of managing the power system in order to maintain a stable frequency. Also, uneven consumption is observed geographically, and depending on the seasonality of the year. This is due to the fact that there is a significant share of electricity consumption for household needs of the population, which fluctuates greatly depending on the ambient temperature. Thus, in 2012, the share of consumption by industry was almost 47%, and in 2017. – 42.8%, and household consumers 26.7% in 2012. to 29.4% in 2017.

It is worth correctly analyzing the data of industrial consumption, because during 2014-2015. Production collapsed several times. And only in 2016. The economy stabilized. During 2016-2018. The form of electricity consumption profiles has not changed significantly (see Fig. 3).

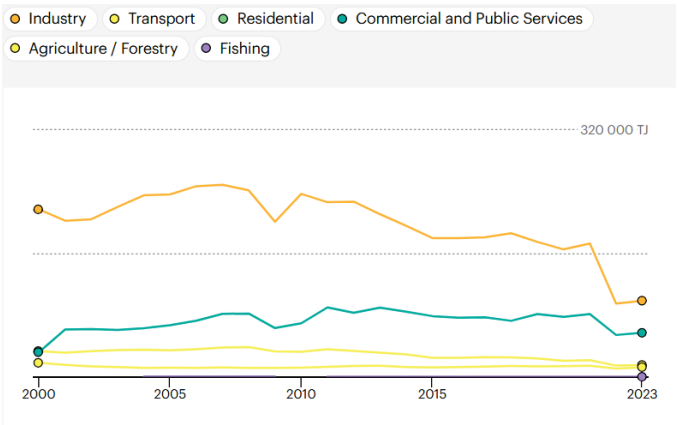


Fig. 3. Electricity consumption by sector in Ukraine

The reason for this was the reduction of industry due to Russia's war against Ukraine, which began in 2014, and the occupation of industrialized areas of Donetsk and Luhansk regions.

The demand for electricity from household consumers has a clearly expressed seasonal character, which is explained by the influence of two factors:

- with an increase in temperature, the use of air conditioning systems increases, which should increase electricity consumption;

- daylight hours are much longer than in winter, and therefore much less electricity is spent on lighting. For Ukraine, this factor is obviously predominant. In winter, consumption peaks, which is associated with a decrease in daylight hours and the need to turn on heating devices to save gas, in particular.

We will still have to assess the consequences of a new invasion of the Russian Federation in the future, but the energy structure described above is likely to be preserved.

III. Features of the Spot Electricity Market of Ukraine

The electricity market in Ukraine began operating on July 1, 2019. However, at the stage of its implementation, there were various restrictions on the lower and upper price levels (price caps). Also, after the new invasion in February 2022, the National Energy and Utilities Regulatory Commission set lower price caps to stabilize the situation in the electricity market, which are still in force.

Thus, in order to evaluate price arbitrage strategies, it is most expedient to choose the period with the most liberal period of the electricity market, namely from August 2021 to December 2022. The choice of this period is sufficient, as it takes into account the annual seasonality (Fig. 4).

Throughout the entire period, there is a tendency of low prices in the 2nd-6th period, and high prices in the 18th-22nd periods. And a very wide range of prices in 8-16 periods. Accordingly, there are conditions for price arbitrage. The strategy of which can be implemented in other market segments, but the day-ahead market is indicative for commercial consumers, so it is advisable to evaluate the effectiveness of strategies on its basis.

IV. Price Arbitrage Strategy in a Liberalized Electricity Market Based on the Calculation of the Minimum Profitable Price Difference

In a market economy, the mass introduction of a new technology depends on the financial feasibility of their implementation for a particular investor, but this carries an additional synergistic effect, which is reflected in the change in the structure of the UPS of Ukraine, an increase in the volume and quality of maneuvering capacities, an increase in the stability and reliability of the system as a whole.

The widespread adoption of such technologies is a kind of "game changer" for the entire industry. This technology has a wide range of possibilities for use at all levels. But based on the functionality of storage systems and the existing electricity market with significant volatility, the consideration of price arbitrage is also an urgent task.

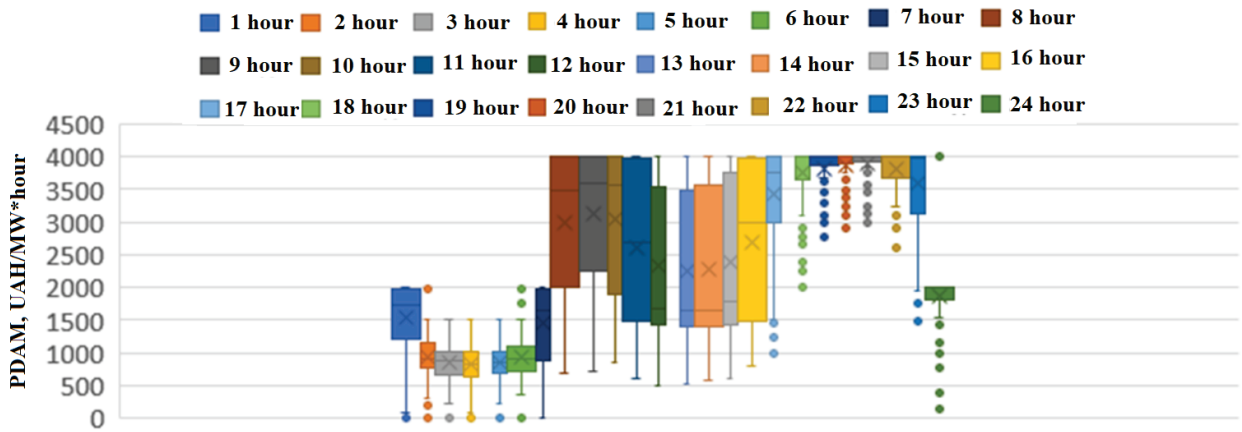


Fig. 4. Day-ahead day-ahead price range by day periods for August-December 2021

In such cases, for the rapid development of a new segment of the industry, incentive tariffs, improved conditions for the import of technologies and other financial and bureaucratic mechanisms are applied. The main task is to find an effective strategy for implementation in the existing conditions.

For this, a number of assumptions and simplifications were made, namely:

- The cost and technical parameters of the drives are taken from commercial offers, European and Ukrainian equipment suppliers, relevant as of April 2021.
- Since these systems are supplied as a set, capital investments for construction, installation and commissioning are disproportionately small with the cost of the systems themselves, and were accepted in the amount of 5% of them
- The data of the Daily Market are taken in advance for August-December 2021, due to the fact that before and after this period there were strict restrictions on the upper and lower prices, which were initially associated with the gradual introduction of the electricity market, and then with force majeure.
- Discounting of cash flows over the payback period, the condition for attracting credit funds and other possible financial instruments are not taken into account.
- It is assumed that the storage device will work in a MicroGrid system of sufficient volume to consume all the accumulated electrical energy - this will not include the payment for the distribution of electrical energy in the calculation.

Storage system integration is considered as part of building a MicroGrid system with a generating unit and load. However, in this case, the calculation was carried out for the ESS as an independent system in order to assess its commercial feasibility, and its volume is optimized depending on the existing market conditions. From a commercial point of view, it is equivalent that the stored electricity will be given to the load inside the MicroGrid system or to the external network, so this calculation can be suitable for any MicroGrid system. However, in the implementation of such a system, connection to external networks will cause a number of additional connection costs.

Considering all mentioned above, the game-theoretic model for ESS arbitrage in microgrids can be presented based on [22, 23,24] as shown in Fig. 5.

In this figure ESS can interact with microgrig aggregator/operator according to a contract operation (CO) strategy. In this case ESS owner will be paid with contracted payment (CP) in accordance to the performed strategy. In other case ESS owner can interact with other RES owners or prosumers thus forming a coalition (COAL). In this case it's win function will be higher (1 instead of 0.75) which is proved by the coalition payment (COALP).

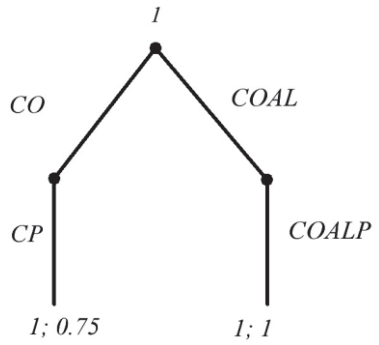


Fig. 5. ESS game-theoretic model for microgrid with RES

Given the limited resource of the cycles of the Lithium-ion battery storage system, it is advisable to enter the market when the difference during the day exceeds the cost of capital investments per cycle, considering losses for energy conversion and depreciation of fixed assets.

The minimum price difference at which it is advisable to implement a price arbitrage strategy based on accumulation systems in Microgrid systems is determined by the formula:

$$P_{Amin} = \frac{\text{CAPEX}}{n_{warranty} \cdot E_{BA} \cdot (1 - DoD_A) \cdot (1 - k_d)} \cdot \eta_{BESS}$$

where CAPEX – the capital investment required for the storage system; $n_{warranty}$ – the number of charge/discharge cycles guaranteed by the manufacturer; E_{BA} – set storage volume; k_d – the degradation coefficient of the drive; η_{BESS} – the efficiency of the storage system.

Taking into account these simplifications and assumptions, the corresponding minimum profitable price difference was calculated, P_{Amin} = 2702.54 UAH/MWh. Using it, to simulate a trading strategy based on day-ahead data for the period August-December 2021, the profit that a Microgrid system with storage devices can receive using a price arbitrage strategy is determined (the results of the simulation are shown in Tab. 1).

TABLE I. MODELING OF THE TRADING STRATEGY OF PRICE ARBITRAGE FOR THE OPERATION OF ELECTRICITY STORAGE SYSTEMS IN THE CONDITIONS OF A LIBERALIZED ELECTRICITY MARKET

Indicator	Excluding the minimum expedient price difference	Taking into account the minimum expedient price difference
Number of charge/discharge cycles for the period August-December 2021	153	130
Total cost of all charging cycles, UAH	-759 754	-589 517
Total cost of all discharge cycles, UAH	2 882 002	2 526 578
Potential profit, UAH	2 122 248	1 937 061
Profit from one cycle	13 871	14 900
Estimated payback period	5.89	6.45

V. CONCLUSION

According to the performed analysis, it can be concluded that the growth of the share of RES in the total energy balance of Ukraine creates the need to ensure their controllability, mainly through the construction and use of modern electricity storage systems. Such systems are most expedient to use as part Microgrids.

A model of a trading strategy of price arbitrage for the operation of electricity storage systems in the conditions of a liberalized electricity market has been proposed. The results of the simulation show that, based on the initial data, under the conditions of operation of the electricity market in this mode, it is extremely profitable. And given that such systems significantly increase the controllability of RES, their implementation in the energy system of Ukraine based on the Microgrid architecture is extremely expedient today.

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Impact and Mitigation of Mutual Inductance Between Inductors in LPFs for Multi-Phase Surge Protection

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***Abstract*—This paper investigates the impact of mutual magnetic coupling between inductors in cascaded low-pass filters used for surge protection in multi-phase power systems. Two coupling scenarios are analyzed: between inductors within a single cascade (same phase) and between inductors of adjacent phases. Simulation results demonstrate that mutual coupling significantly affects the residual voltage spike levels at the load, particularly under light-load conditions. A coupling coefficient above 0.1 leads to degraded filtering performance and elevated voltage stress. To support the analysis, three inductor configurations were designed and evaluated through 3D finite element simulations and experimental measurements. Among them, the toroidal-core inductor exhibited the lowest magnetic coupling and best thermal and electrical performance. The results provide practical guidelines for selecting inductor configurations and layout strategies to minimize mutual inductance and ensure reliable surge suppression in multi-phase systems.**

***Index Terms*—air-core inductor, low-pass filter, magnetic coupling, mutual inductance, toroidal coil**

I. Introduction

In addition to crowbar protection devices [1]–[5], such as varistors, gas discharge tubes, and transient voltage suppressors, low-pass filters (LPFs) are commonly incorporated into protection circuits to safeguard the load from voltage surges in the power network [6]–[10]. In the case of a multi-phase power supply, these filters are typically connected between each phase and ground. Each line (phase) may include multiple filters connected in a cascaded configuration (Fig. 1). However, mutual magnetic coupling between the inductors of these filters can have a significantly negative impact on the ability of the protection circuit to suppress voltage surges from the mains.

There are several common approaches to reducing mutual inductance between two coils [11]–[13]: 1) enclosing one or both coils in a magnetic shield made of high-permeability materials; 2) increasing the physical separation between the coils through distance and/or orientation; 3) using toroidal coils, which confine the magnetic field within the core, thereby reducing field leakage and minimizing mutual inductance with nearby coils.

The use of a magnetic shield made of high-permeability materials adds complexity to the filter design and increases its cost. In contrast, increasing the physical separation between the coils does not raise the cost and is an excellent approach when the overall dimensions of the device housing the filters are not critical. The use of toroidal coils allows for a reduction in the filter's size but may be a more expensive solution compared to increasing the physical separation between air-core inductors.

In this work, the impact of mutual magnetic coupling between inductors in cascaded LPFs was analyzed under surge conditions. Two coupling scenarios were considered: within a single cascade and between inductors of adjacent phases. A comprehensive evaluation was carried out using 3D finite element simulations, experimental validation, and comparison of three practical inductor configurations. The

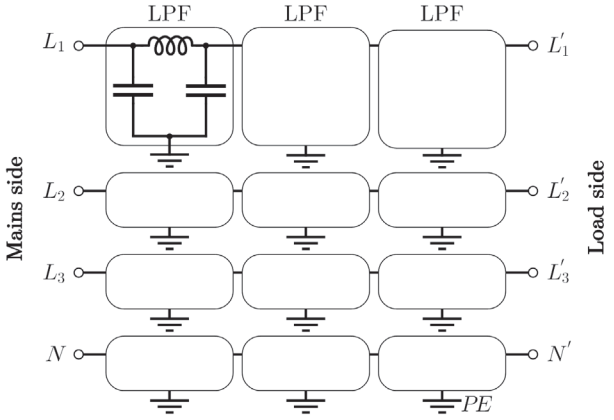


Fig. 1. Voltage surge protection circuit using cascaded low-pass filters.

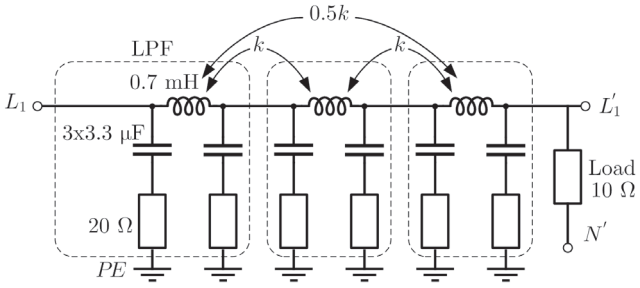


Fig. 2. Magnetic coupling between the LPS's inductors of the same phase.

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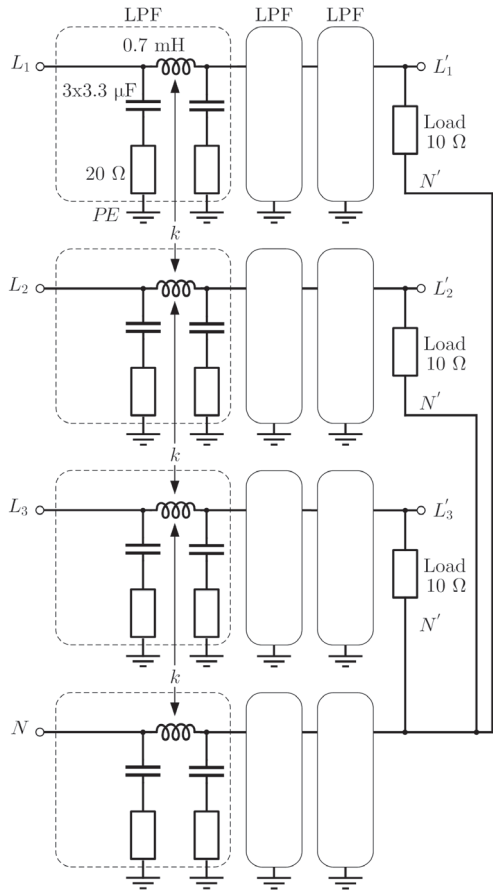


Fig. 3. Magnetic coupling between inductors of adjacent phases.

study provides quantitative guidelines for acceptable coupling levels and offers design recommendations for minimizing mutual inductance and its adverse effects on surge suppression performance.

II. Impact of Mutual Inductance on Filter Design

To evaluate the impact of mutual magnetic coupling between the inductors of the LPFs shown in Fig. 1, two cases were simulated: 1) magnetic coupling between the LPF's inductors of the same phase (Fig. 2) and 2) magnetic coupling between the inductors of all adjacent phases in every stage of the LPF (Fig. 3). A resistor was connected in series with each LPF capacitor, as this configuration effectively mitigates voltage surges.

In the first case, the magnetic coupling coefficient k between the first and third inductors of an LPF cascade was set to half the value of k between the first and second, and between the second and third inductors. In the second case, k was set equal between all inductors of the first, second, and third cascades.

These two cases were evaluated independently: during the simulation of the first case, the coupling between inductors of different phases was set to zero; conversely, in the second case, the coupling between inductors within the same phase was set to zero.

A 4 kV surge pulse, generated by a combinational waveform generator in accordance with the IEC 61000-4-5 standard for surge immunity testing, was applied at the mains side, and the resulting residual voltage spike at the load side was analyzed. Filter and load parameters are detailed in Figs. 2 and 3. The load consisted of 10 Ω resistors connected between each phase and neutral.

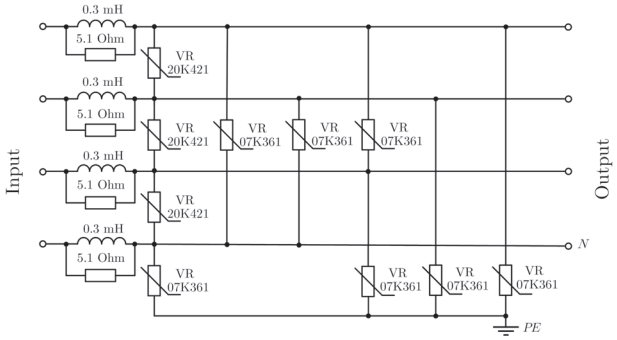


Fig. 4. Voltage limiting stage placed upstream of the LPFs.

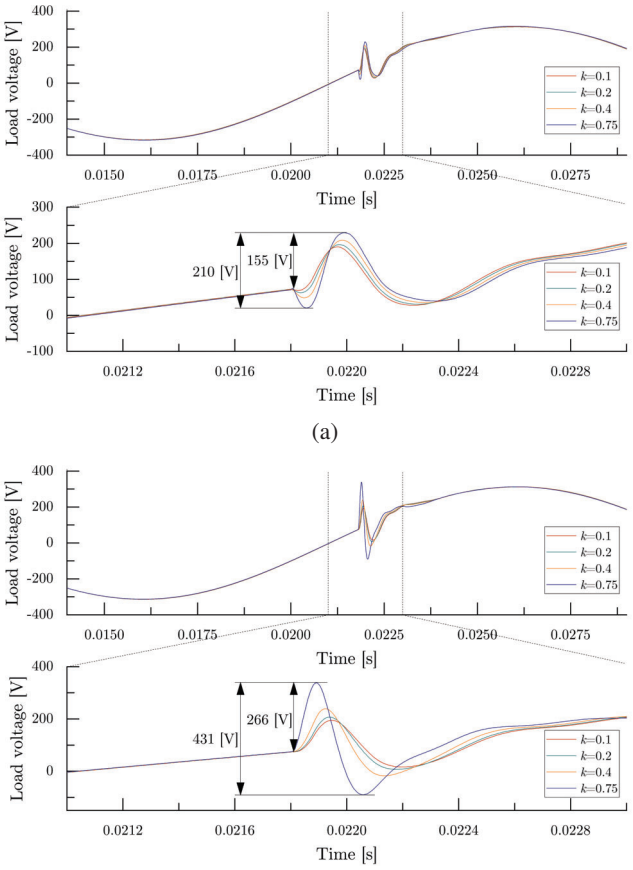


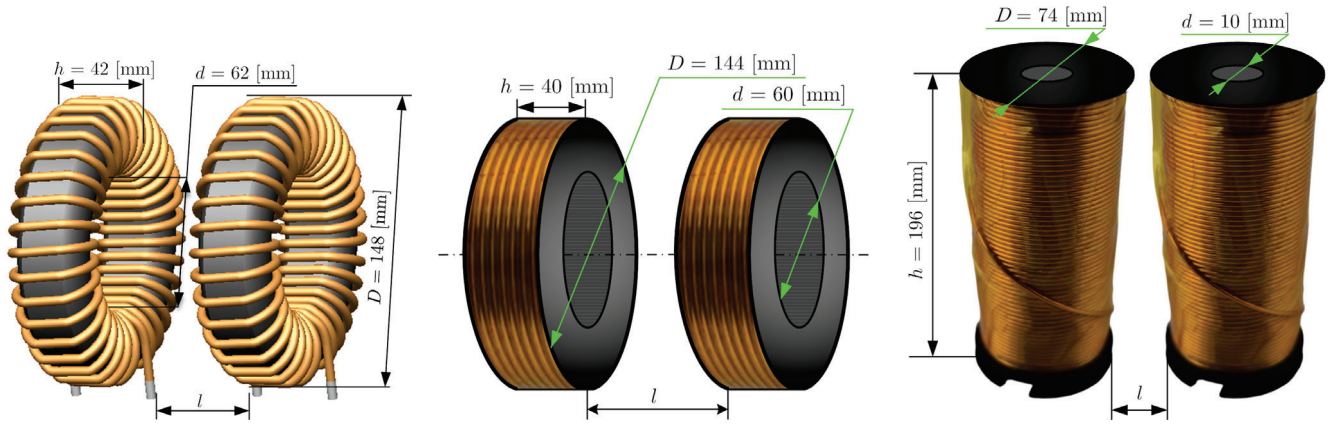
Fig. 5. Simulation results showing residual voltage spikes resulting from a 4 kV surge pulse. (a) Magnetic coupling between the inductors of the same phase. (b) Magnetic coupling between inductors of adjacent phases.

It should be noted that a voltage limiting stage, shown in Fig. 4, was installed upstream of the LPFs. This stage was used to absorb part of the energy of the incoming pulse [7].

The surge pulse was applied in three configurations in accordance with IEC 61000-4-5: line-to-line (L–L), line-to-neutral (L–N), and line/neutral-to-earth (L/N–PE). For the L–L and L–N configurations, the surge was applied through an 18 μF capacitor. In the L/N–PE configuration, it was applied through a series connection of a 9 μF capacitor and a 10 Ω resistor.

Fig. 5 presents selected simulation results, revealing a slight difference in the residual voltage spike levels depending on the surge configuration: L/N–PE versus L–L or L–N. The waveforms represent the worst-case scenario.

The waveforms in Fig. 5(a) correspond to the case of magnetic coupling between inductors within a single cascade. As seen from the diagrams, oscillations in the output voltage



(a) (b) (c)

Fig. 6. Inductor configurations. (a) No. 1. (b) No. 2. (c) No. 3.

caused by the surge pulse at higher values of k result in residual voltage spike levels exceeding 155 V (210 V peak-to-peak) at $k = 0.75$. Even at $k = 0.1$, the residual voltage spike reaches approximately 100 V (95 V peak-to-peak).

In contrast, the waveforms in Fig. 5(b) represent the second case – magnetic coupling between inductors of adjacent phases – and indicate that the influence of k in this configuration is significantly more pronounced than in the first case. The residual voltage spike levels exceed 266 V (431 V peak-to-peak) at $k = 0.75$, and approximately 85 V (260 V peak-to-peak) at $k = 0.1$.

In a scenario combining both types of magnetic coupling, residual voltage spike levels are comparable to those observed in the single-cascade coupling case.

It is also worth noting that, under partial-load operation at 10% of nominal power – corresponding to a load resistance of 100 Ω – the residual voltage spike levels were substantially higher than at full-load operation with a 10 Ω load.

Based on the analysis of the resulting residual voltage spikes, it can be concluded that to effectively limit the spike amplitude, the coupling coefficient k should preferably be maintained below 0.1.

III. Inductor Design Considerations for Magnetic Coupling

Three inductor configurations were considered to evaluate the influence of inductor placement and geometric dimensions on their self-inductance, mutual inductance, and magnetic coupling coefficient. The first configuration employed a toroidal coil, while the second and third used air-core inductors.

It was assumed that a sinusoidal current with an RMS value of 25 A flows through the LPFs. Accordingly, the inductor winding was assumed to be made of copper litz wire with a strand diameter of 0.1 mm and an overall conductor diameter of 4 mm, yielding a total cross-sectional area of 5.8 mm^2. The electrical conductivity of the wire was taken as 58 MS/m. The number of turns in each inductor was calculated to achieve a target inductance of 0.776 mH.

For the toroidal coil inductor, the XFlux 0078337A7 core from MAGNETICS was selected, featuring an initial relative magnetic permeability μ_r of 26. This configuration was also evaluated for the case of a peak current of 50 A to assess possible saturation effects. The analysis showed that the inductance remains nearly unchanged under this condition.

The geometric dimensions of the inductors are illustrated in Fig. 6, and their main technical parameters are summarized in Table I.

TABLE I
Inductor Parameters

Parameter	Configuration No. 1	Configuration No. 2	Configuration No. 3
Winding wire	Overall diameter: 4 mm; cross-section: 5.8 mm^2		
Inductance	0.776[mH]		
Number of turns	106	101	384
Wire length	13.5[m]	31[m]	51[m]
Number of fully or nearly fully wound layers	2	10	8
DC resistance	0.041 [Ω]	0.094 [Ω]	0.155 [Ω]
Inductor weight	0.701 (wire) [kg] and 1.465 (core) [kg]	1.611 [kg]	2.65 [kg]
Core	XFlux MAGNETICS 0078337A7	–	–

The following inductor placement scenarios were analyzed: 1) configuration No. 1 and configuration No. 2 inductors placed coaxially; 2) configuration No. 3 inductors placed side by side. The spacing l between the inductors was defined as the distance between the outer edges of their windings, as illustrated in Fig. 6.

IV. 3D Modeling and Simulation

Three-dimensional simulations were performed using Ansys Maxwell 3D to evaluate how the magnetic coupling coefficient of the coil configurations depends on the distance between them (Fig. 7). For configuration No. 1, the finite element mesh for the base case consisted of 414,443 elements, including 4,470 elements in the magnetic core, 6,167 in each winding, 26,636 in the smaller air region, and 366,533 in the larger air region. For configuration No. 2, the mesh comprised approximately 105,000 elements, with 1,020 assigned to each inductor, 76,893 to the smaller air region, and 26,067 to the larger air region. For configuration No. 3, the mesh contained 315,690 elements in total, including 4,415 in each inductor, 275,719 in the smaller air region, and 31,141 in the larger air region. The skin effect was not taken into account in these simulations.

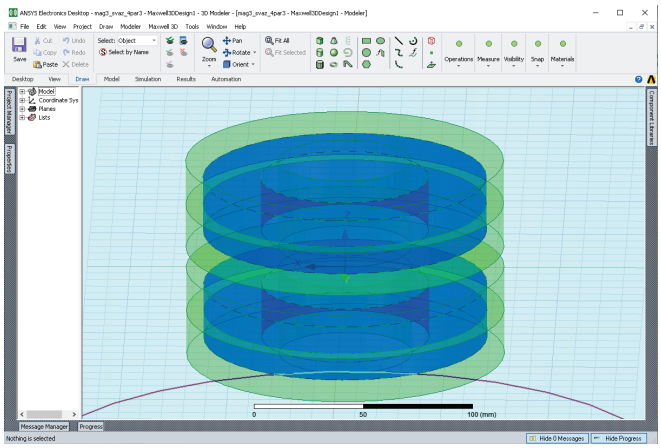


Fig. 7. Magnetic coupling simulation of inductors in Ansys Maxwell 3D.

To calculate the inductance, the average magnetic energy within the winding volume is given by

The inductor L_r was implemented as two series-connected air-core inductors, each with 15 turns, resulting in a total inductance of 9.5 µH.

$$\bar{E} = \frac{1}{2}\int_V \Re\{\mathbf{B}\cdot\mathbf{H}\}\,dV \tag{1}$$

where $\mathbf{B}$ and $\mathbf{H}$ are the complex amplitudes of the magnetic flux density and magnetic field strength, respectively, and V is the winding volume.

To determine the inductance of the first inductor, a sinusoidal current with a frequency of 50 Hz and an amplitude of 50 A is applied to its winding, while no current is applied to the second inductor. The energy stored in both the first and second inductors is then measured, with the energy in the second inductor corresponding to the mutual inductance.

To determine the inductance of the second inductor, the current is applied to its winding, and the energy stored in it is measured. The mutual inductance is determined similarly to the first case, but the energy is measured in the first inductor.

Since the instantaneous energy in an inductor is given by

$$E_{\text{inst}}(t) = \frac{1}{2}L\,[i(t)]^2 \tag{2}$$

where L is the inductance of an inductor, and the instantaneous current is related to the amplitude is given by

$$i(t) = I_a\,\sin(\omega t) \tag{3}$$

and I_a is the amplitude and ω is the angular frequency of the sinusoidal current. The average energy of a half-period of i can be found by integrating the instantaneous energy over one period, as follows:

$$\bar{E} = \frac{2}{T}\int_0^{0.5T} E_{\text{inst}}(t)\,dt \tag{4}$$

where $T = 2\pi/\omega$ is the period of the sinusoidal current.

Substituting (2) into (4) and performing the integration yields:

$$\bar{E} = \frac{L}{4}I_a^2. \tag{5}$$

From this, the inductance or mutual inductance is determined as follows:

$$L = \frac{4\bar{E}}{I_a^2}. \tag{6}$$

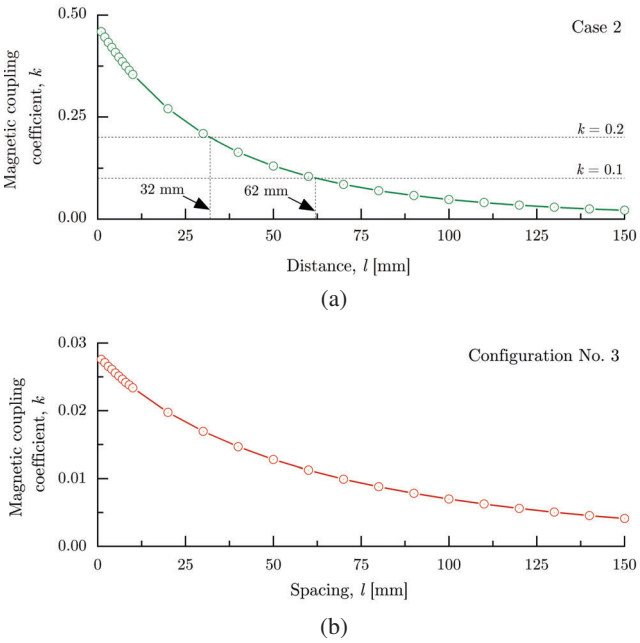


Fig. 8. Magnetic coupling coefficient vs. coil spacing for different inductor configurations. (a) Configuration No. 2. (b) Configuration No. 3.

When the energy is considered in the first inductor through which the current flows, the inductance according to (6) corresponds to the self-inductance of that inductor, while the energy in the second inductor corresponds to the mutual inductance (M). Knowing that the inductance of the first inductor is L_1, the second is L_2, the value of k is calculated by

$$k = \frac{M}{\sqrt{L_1 L_2}} \tag{7}$$

Simulation results revealed that configuration No. 1 exhibits an extremely low level of magnetic coupling between the coils. Even at $l = 2$ mm, the magnetic coupling coefficient remains below $1 \cdot 10^{-8}$, indicating excellent magnetic isolation.

In contrast, configurations No. 2 and No. 3 show higher magnetic coupling, with their dependence on coil spacing illustrated in Fig. 8. As shown in Fig. 8(a), when two coils of configuration No. 2 are placed side by side with only $l = 1$ mm of separation, the magnetic coupling coefficient reaches as high as 0.46. This value decreases to 0.2 at $l = 32$ mm and only falls below the threshold of 0.1 when $l = 62$ mm.

Meanwhile, configuration No. 3 demonstrates significantly lower magnetic coupling compared to configuration No. 2, as presented in Fig. 8(b). Even at $l = 1$ mm, the coupling coefficient does not exceed 0.03, which already meets the desired criterion of being less than 0.1.

V. Experimental Validation and Discussion

A. Experimental Validation

To validate the results of the 3D simulations, experimental measurements were conducted. For this purpose, prototype coils corresponding to configuration No. 1 and configuration No. 2 were fabricated, as shown in Fig. 9.

The schematic structure of the measurement setup is shown in Fig. 10(a), and a photograph of the experimental bench is presented in Fig. 10(b).

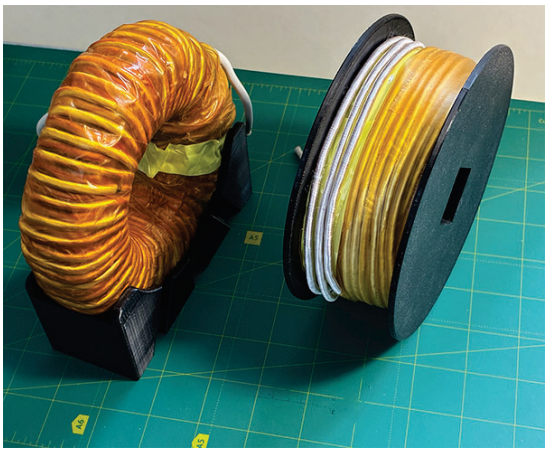
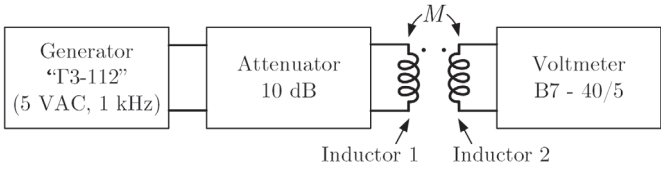
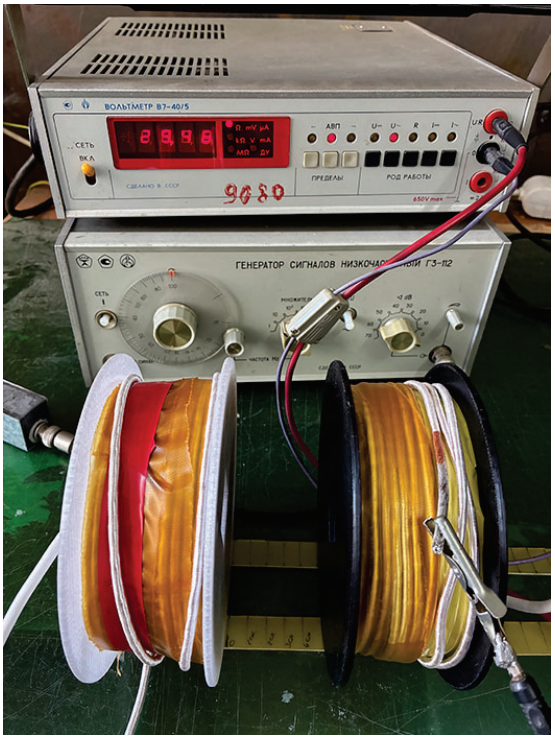


Fig. 9. Fabricated prototypes of the inductors: configuration No. 1 (left) and configuration No. 2 (right).



(a)



(b)

Fig. 10. Experimental setup for magnetic coupling coefficient measurement. (a) Schematic structure. 2. (b) Photograph of the experimental bench.

The experimental measurements were based on the methodology presented in [14], which demonstrates that M is calculated by

$$M = \frac{V_{\text{meas}}}{\omega_{\text{gen}} I_1} \tag{8}$$

where ω_{gen} is the angular frequency of the excitation current applied by the generator to the first inductor, and V_{meas} is the measured voltage across the second inductor, and I_1 is the RMS value of the current flowing through the first inductor.

Taking into account (7) and (8), as well as $L_1 = L_2$, the

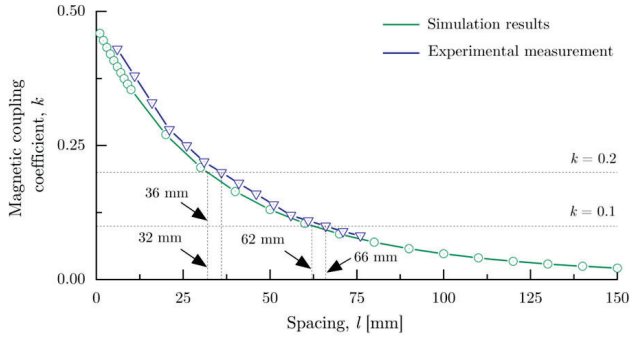


Fig. 11. Comparison of experimentally measured and simulated magnetic coupling coefficients for configuration No. 2 at different coil spacings.

magnetic coupling coefficient can be expressed as:

$$k = \frac{V_{\text{meas}}}{\omega_{\text{gen}} I_1 L_1}. \tag{9}$$

At frequencies significantly lower than the self-resonant frequency of a real inductor (where the effects of winding resistance and inter-turn capacitance are negligible), its simplified equivalent circuit can be represented as a series connection of a resistor (accounting for copper losses) and a pure inductance. Measurements of inductors in configuration No. 2 using an RLC meter showed that at 1 kHz, the inductance is approximately 0.77 mH, and the winding resistance is 0.097 Ω. At this frequency, the inductive reactance of the inductor is 4.84 Ω, which is about 50 times greater than the winding resistance. Therefore, resistive losses were neglected for calculation purposes, and the generator frequency was set to 1 kHz.

As the voltage source, a function generator was used, providing a sinusoidal output with an RMS voltage of 5 V. To prevent the inductor’s relatively low reactance (4.84 Ω) from excessively loading the generator or shunting its output, a 10 dB T-type attenuator was placed between the generator and the inductor.

Based on the selected frequency and inductance values, expression (8) can be simplified numerically as:

$$k \approx 4.31\, V_{\text{meas}}. \tag{10}$$

The measured magnetic coupling coefficient for configuration No. 1 was found to be extremely low – close to the measurement error margin – thus confirming the simulation results and indicating negligible magnetic interaction between the coils.

For configuration No. 2, the experimentally obtained dependence of the magnetic coupling coefficient on the distance between the coils is presented in Fig. 11. The figure also includes the corresponding simulation results previously shown in Fig. 8(a). The close agreement between the experimental and simulated data validates the accuracy of the 3D modeling/simulation approach.

B. Discussion

Each of the three analyzed inductor configurations exhibits distinct advantages and limitations in terms of magnetic coupling, thermal performance, size, and cost.

Configuration No. 1 employs a toroidal core with a relatively high relative magnetic permeability ($\mu_r = 26$), which enables a compact winding with only two layers and low DC resistance (0.041 Ω). Due to the magnetic confinement provided by the core, mutual inductance between two such

inductors remains negligible regardless of their placement and the distance between them. However, the presence of the core significantly increases the cost and weight of the inductor ($1.465\,\text{kg}$ core mass), which may be critical in cost-sensitive or weight-constrained applications.

Configuration No. 2 utilizes an air-core structure with ten layers of winding. The mutual coupling between inductors of this type is strong, necessitating a larger separation distance or changing orientation to avoid undesired interference. The high number of winding layers, combined with the identical cross-sectional area of the conductor as in configuration No. 1, may result in significant internal heating. Additionally, the DC resistance ($0.094\,\Omega$) is more than twice that of configuration No. 1, leading to greater power loss and voltage drop. Without increasing the conductor cross-section, this configuration is prone to overheating under prolonged operation. If the conductor's cross-section is increased to improve thermal performance, the resulting weight of the inductor would exceed that of configuration No. 1, while the overall cost would become comparable due to the increased amount of copper used. These factors would make this configuration less attractive as well.

Configuration No. 3 also features an air-core design but with eight winding layers and the longest conductor length ($51\,\text{m}$), resulting in the highest DC resistance among the three ($0.155\,\Omega$). While its magnetic coupling with other identical inductors remains low – comparable to configuration No. 1 – the substantial resistive losses and associated heating are notable drawbacks. The elevated resistance can cause a significant voltage drop, especially considering the cascade connection of three such inductors in the system. While increasing the conductor cross-section could mitigate thermal issues and reduce resistance, doing so would disproportionately increase the size, weight (already $2.65\,\text{kg}$), and cost, making this configuration less attractive overall.

Thus, configuration No. 1 is the most balanced and technically attractive option, particularly when mutual coupling must be minimized, and thermal performance and voltage drop are critical design constraints. Configuration No. 2 can be considered a viable alternative in scenarios where increased spacing between inductors is acceptable and weight is less of a concern. However, if a larger conductor cross-section is used to address thermal limitations, its weight would surpass that of configuration No. 1, and its cost would become nearly equivalent, ultimately making this option unappealing.

Conclusion

In this study, the influence of mutual magnetic coupling between inductors in low-pass filters under surge conditions was investigated. Simulation results showed that coupling between the inductors of all adjacent phases in every stage of the LPF has a more pronounced effect on residual voltage spikes than coupling between the LPF's inductors of the same phase. Residual voltages exceeded $260\,\text{V}$ at a magnetic coupling coefficient of $k = 0.75$, especially under light-load conditions. To ensure effective surge suppression, the value of k should be kept below 0.1.

To support the simulation findings, three inductor configurations for low-pass filters were designed, simulated, and two of them were experimentally evaluated. Among them, configuration No. 1, which employs a toroidal core, exhibited excellent magnetic isolation and low DC resistance, making it the most suitable for compact in low-pass filter designs with minimal interference. Experimental measurements confirmed the negligible coupling predicted by 3D simulations. Configuration No. 2, based on an air-core structure, demonstrated significant mutual coupling, which can be mitigated by increasing coil spacing, and notable resistive losses, which require a larger conductor cross-section, resulting in greater size and weight. Configuration No. 3, despite its low magnetic coupling, proved impractical due to excessive copper requirements – leading to high cost, large size, and, if thinner wire is used, substantial losses and voltage drop. Therefore, configuration No. 3 is not recommended for practical use, while the toroidal-core inductor of configuration No. 1 remains the most technically and economically justified solution.

Acknowledgment

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MindGuard: Intelligent System for Depressive and Suicidal Intentions Detection in Social Media Texts Based on CNN, BERT and RoBERTa

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***Abstract*—The paper presents the development and experimental implementation of the MindGuard information and analytics platform, designed for automated analysis of the texts of social media users in order to identify signs of depression, anxiety and suicidal intentions. The system uses modern natural language processing (NLP) methods, including TF-IDF, deep neural networks (CNN) and transformer models (BERT, RoBERTa). A mark-up of data based on Reddit posts (≈ 348 examples), which covers three categories: non-psycho, depression and suicide. Several variants of classification models have been developed. The highest accuracy was demonstrated by the deep CNN model with GloVe embedding: Accuracy (≈ 81.5%), F1-score (0.92 for non-psycho, 0.74 for depression and 0.76 for suicide), AUC ROC (0.990 for non-psycho, 0.911 for depression and 0.910 for suicide), Inference latency ≈ 15 ms, bandwidth - up to 66 requests/s per CPU. Practical implementation is implemented in the form of a web application with REST API, which provides instant processing of text queries and convenient visualisation of results for psychologists, moderators or users. The system supports multilingualism and can be easily adapted to local needs.**

Keywords—depression; suicidal intentions; NLP; text classification; social networks; Transformers; CNN; psycholinguistics; MindGuard; Sentiment analysis.

I. INTRODUCTION

In today's digital environment, social networks and instant messengers have become the main channels of communication, where millions of text messages are generated every day. A significant part of them reflects the emotional state, mood and even mental health of users. While the World Health Organisation records a rapid increase in cases of depression, anxiety disorders and suicidal intentions, there is an urgent need to create tools for the early detection of such risks based on digital content. This problem is especially acute in Ukraine, where conditions of social instability and the limited resources of medical institutions are present.

Given the development of natural language processing (NLP) and machine learning technologies, in particular transformer architectures and deep neural networks, new opportunities are opening up for automated analysis of the psychological state of text messages. However, existing services have a number of limitations: a lack of specialisation in mental health, insufficient interpretation of results, and a lack of multilingual support and localisation. In this regard, the purpose of this study is to develop an intelligent information and analytical platform, MindGuard, which allows you to carry out an in-depth analysis of the texts of social media users in order to identify signs of depression, anxiety or suicidal intentions. The system should provide high classification accuracy and performance, as well as a convenient API for integration into external services. To achieve this goal, the following tasks are solved in the work:

- analyse scientific and applied sources in the field of psycholinguistics, NLP and mental health diagnostic systems;
- to form an educational corpus of texts with the marking of categories: non-psycho, depression, suicide;
- implement and compare classical and in-depth approaches to classification (TF-IDF + ML, CNN, transformers);
- develop a REST API for real-time text class prediction;
- Test the accuracy, performance and interpretation of the system.

The object of the study is text messages from users of social networks and forums. The subject of the study is the linguistic, statistical and semantic features of the text that correlate with the signs of psychological disorders (depression, suicidal thoughts).

The scientific novelty of the work lies in the creation of a hybrid architecture for analysing psycho-emotional content, which combines statistical methods (TF-IDF), deep learning (CNN) and transformer models (BERT, RoBERTa), adapted to the local context of Ukrainian and English-language texts. For the first time, a comprehensive solution has been developed that allows for the detection of psychologically risky messages in a multilingual environment at high speed. The practical value of the project lies in the possibility of integrating the MindGuard platform into the work of psychological services, moderation teams of social networks, educational institutions, and mental well-being services. The system allows you to automate the initial screening of a large volume of texts, detect threatening signals in a timely manner, and take appropriate preventive actions.

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II. Related Works

Over the past decade, there has been a significant increase in the number of studies aimed at applying natural language processing (NLP) methods to analyse the mental state of users in the digital environment. The relevance of this topic is due to both the spread of depressive and anxiety disorders and the rapid development of machine learning technologies capable of processing large amounts of text data. One of the first systematic approaches to psycholinguistic analysis is the Linguistic Inquiry Word Count (LIWC), a tool that classifies words into psychological categories (emotions, cognitive processes, social topics) [1]. Further development of research is related to the use of social media as a source for identifying mental disorders. [2] proposes the SMHD (Self-reported Mental Health Diagnoses) framework for studying depression and PTSD based on posts by Reddit users, where both classic and modern classification methods are applied. Transformer-type models play an essential role in the study of psychological risks. For example, [3] demonstrated the possibility of effectively using BERT to identify the risk of self-harm in support forums. Similarly, in [4], models have been developed that study the discourse of mental health without the need for manual labelling, opening up new horizons in the creation of interpreted models. The methodological basis for the emotional typology of the text can be found in [5], where the circumplex model of affect is presented – a theoretical basis for the analysis of emotional states in two dimensions: valence and activation. Modern mood analysis platforms, such as IBM Watson Tone Analyser or Google Cloud NLP, provide high-speed classification of overall emotional tone but do not focus on identifying mental disorders or provide interpretation at the level of clinically significant conditions.

III. Methods and Materials

The MindGuard study took into account the limitations of existing solutions. It adapted hybrid methods – a combination of TF-IDF, CNN, and BERT – demonstrating high accuracy and practical applicability in classifying texts according to risky psychological characteristics. The study implements a hybrid approach to the analysis of text messages of social media users to identify psychological risks (depression, anxiety, suicidal intentions) based on a combination of classical statistical methods and modern models of deep learning [6-10]. Both traditional machine learning algorithms and deep architectures with pre-trained embedding are applied [11-16].

A. Text presentation methods

For classical machine learning models, the "word bag" model with TF-IDF Term Frequency–Inverse Document Frequency weighting was used. Selected:

- TfidfVectorizer(max_features=5000, ngram_range=(1, 2));
- support for unigrams and bigrams;
- removing stop words;
- normalisation of text before vectorisation (lowercase, URL deletion, punctuation, mentions, tokenisation).

In deep models, embedding (GloVe) and vectorisation are used:

- TextVectorization layer (sequence length: 100 tokens);
- GloVe-embedding with a dimension of 100;
- embeddings were not updated during training (trainable=False);
- additionally applied SpatialDropout1D(0.2) for regularization.

B. Classification algorithms

Three main classic ML models of pipelines with the scikit-learn library have been implemented:

- Logistic Regression (LR_basic) is easy to interpret and provides the best accuracy among classical methods.
- Multinomial Naive Bayes (NB_advanced) is fast and efficient for working with TF-IDF, but it has limitations in probability estimates.
- Linear SVM (SVM_basic) is accurate in text classification problems but sensitive to hyperparameter settings.

Implemented TextCNN architecture based on the CNN depth model, including:

- input layer – Input(shape=(), dtype=tf.string) + TextVectorization;
- GloVe-Embedding(input_dim=20000, output_dim=100, weights=[embedding_matrix], trainable=False);
- convolutional layers – five parallel Conv1D with nuclei sizes from 2 to 6;
- pooling – GlobalMaxPooling1D and GlobalAveragePooling1D for each layer;
- combining features – Concatenate() into a vector of length 1280;
- dense layers – Dense(64, activation='relu') with BatchNormalization and Dropout(0.5);
- output layer – Dense(3, activation='softmax').

CNN Training Options:

- Optimiser: AdamW (lr=1e-3, weight_decay=1e-5)
- Loss function: sparse_categorical_crossentropy
- Metrics: accuracy, precision, recall, F1-score
- Callbacks: EarlyStopping, ReduceLROnPlateau, ModelCheckpoint

C. Development tools

Languages and libraries:

- Python 3.8+ – the primary development language;
- pandas, NumPy – processing of tabular data;
- scikit-learn – classical algorithms, pipelines;
- TensorFlow/Keras – building and training deep models;
- HuggingFace Transformers – use of BERT-like models (experimental);
- FastAPI – building a backend service;
- React + TypeScript – implementation of the client part (frontend);
- Docker – backend containerisation.

a) Visualisation and Analysis:

- Matplotlib, WordCloud – data visualisation;
- Jupyter Notebook / Google Colab – testing and interactive analysis;

- TensorBoard – model training monitoring.

D. *Infrastructure and performance*

To assess system efficiency, performance tests were carried out:

- Inference Latency: ~15ms per request;
- Bandwidth: ≈ 66 texts/sec per CPU;
- Memory consumption: ΔRAM ~ 6KB/request;
- CPU Usage: ~100% Single Thread – Linear scalability via multithreading.

Thus, the implemented system combines a flexible choice of methods at all stages – from text pre-processing to high-performance inference, allowing for scalable and reliable classification of psychologically significant texts in real-time.

IV. Experiments

Let's test and measure the static result of measuring the operation of the stored model under a load of 1000 examples. Interpretation:

```
Latency (ms):  min=13.0,  p50=15.0,  max=64.5
ΔRAM (bytes):   avg=6439
ΔCPU (%):       avg=100.9
```

Fig. 1. Execution result

- Latency (single request processing):
  - Minimum 13 ms – under ideal conditions, inference lasts about 13 ms;
  - Median 15 ms – half of requests are processed in no longer than 15 ms;
  - Maximum 64.5 ms – single "delays" (due to garbage collection and other system processes) do not exceed 0.1 s. Therefore, on average, one model is capable of serving ≈ 1000 ms / 15 ms ≈ 66 requests per second on one CPU stream.
- ΔRAM (memory increment). On average, about 6 KB per inference. That is, the model does not create significant additional allocations in RAM.
- ΔCPU (% load). An average ≈ 100% on one logical core means that the model "loads" the CPU stream to the maximum.

Statistical execution of the model with the following key metrics: confusion matrix, precision/recall / F1-score report for each of the three classes and ROC AUC value. Below is a detailed description of each section and conclusions on the practical implications of these results.

V. Statistical Execution of ML Models

A. *Logistic Regression (TF-IDF + basic)*

Correct predictions (diagonal of the matrix) are shown in Fig. 2. Of the 23,208 reports without psychiatric signs, 21,410 were correctly classified as non-psycho. Of the 23,207 cases of depression, the model correctly classified 17,345. Of the 23,207 suicidal reports, 17,334 were classified as suicide. The most common mistakes:

- Non-psycho → depression/suicide 986 and 812 cases, respectively - the level of false alarms for healthy texts remains within the normal range.
- Depression → suicide 4,317 examples - the model confused depressive statements with suicidal ones.
- Suicide → depression 4,346 examples - insufficient sensitivity to suicidal signs.

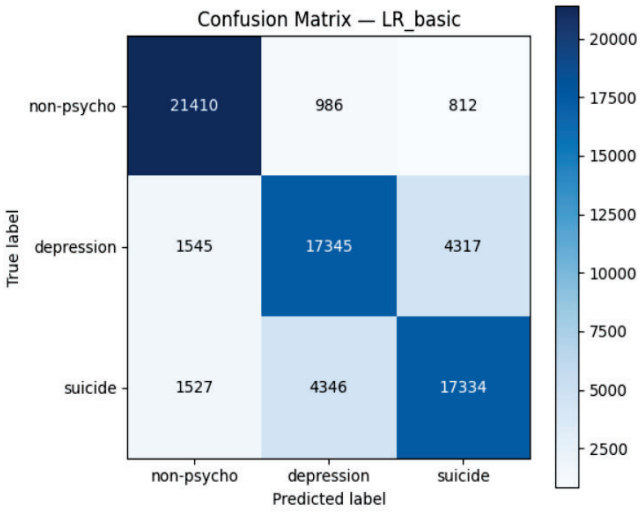
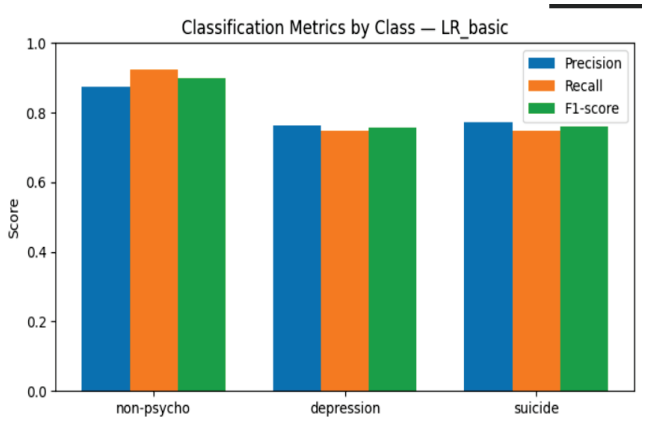


Fig. 2. Inaccuracy matrix



Classification report:

	precision	recall	f1-score	support
non-psycho	0.87	0.92	0.90	23208
depression	0.76	0.75	0.76	23207
suicide	0.77	0.75	0.76	23207
accuracy			0.81	69622
macro avg	0.80	0.81	0.80	69622
weighted avg	0.80	0.81	0.80	69622

Fig. 3. Precision, recall and F1-score report

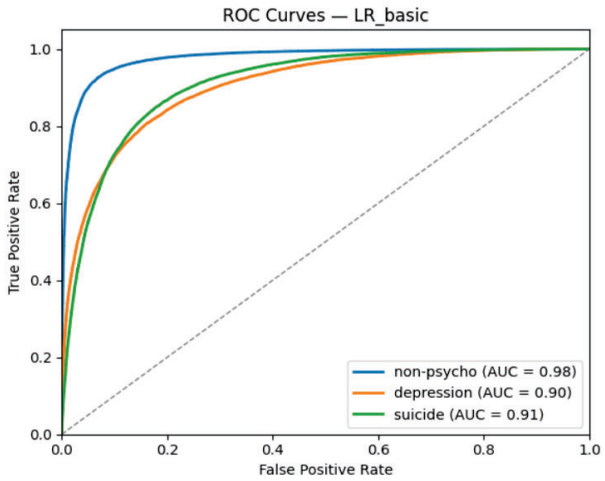


Fig. 4. ROC AUC

The model demonstrates a relatively stable ability to separate healthy from risky messages. However, lexical confusion between depression and suicide remains. Metrics:

- Precision for non-psycho 0.87 (good accuracy without risk) and depression and suicide ~0.76-0.77 (every 4th prediction can be a false alarm).
- Recall for non-psycho 0.92, for depression 0.75 and suicide 0.75.
- F1-score for non-psycho 0.90, for depression 0.76 and suicide 0.76.

B. *MultinomialNB (TF-IDF + advanced)*

Correct predictions for non-psycho 19,116 out of 23,208 are classified correctly, for depression - 16,494 out of 23,207, and for suicide - 16,634 out of 23,207. The most common mistakes:

- Non-psycho → depression/suicide 2,205 and 1,887 examples - a relatively high number of false alarms.
- Depression → suicide 5,102 cases – weak differentiation.
- Suicide → depression 5,160 cases.

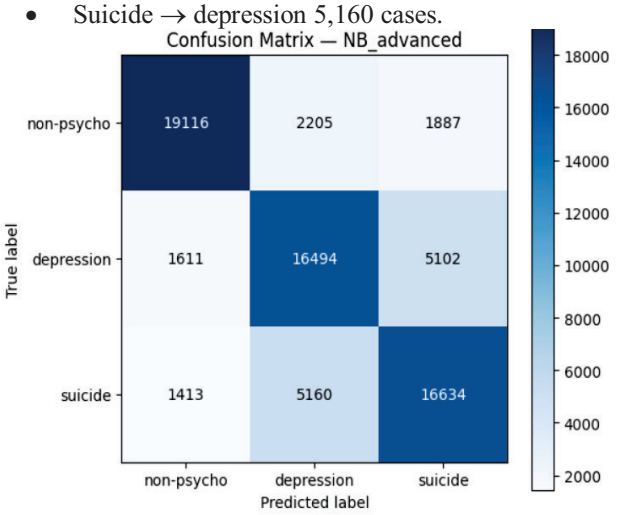
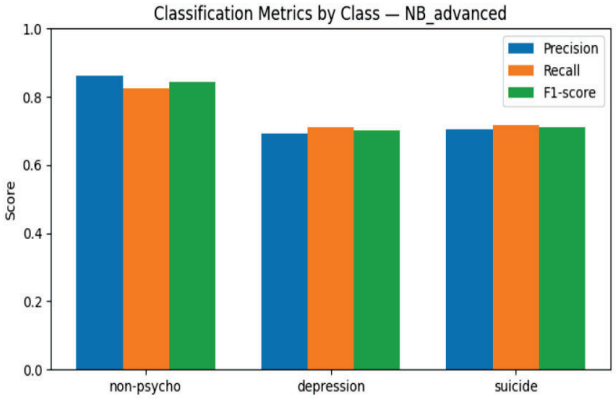


Fig. 5. Inaccuracy matrix



Classification report:

	precision	recall	f1-score	support
non-psycho	0.86	0.82	0.84	23208
depression	0.69	0.71	0.70	23207
suicide	0.70	0.72	0.71	23207
accuracy			0.75	69622
macro avg	0.75	0.75	0.75	69622
weighted avg	0.75	0.75	0.75	69622

Fig. 6. Precision, recall and F1-score report

The lowest scores among the models. Lemmatisation showed a slight improvement, but NB is less flexible in terms of complex semantics. The model is prone to general "reinsurance", as a result of which it has more false positives. Metrics:

- Precision for non-psycho 0.86 (good accuracy without risk) and depression and suicide ~0.69-0.70.
- Recall for non-psycho 0.82, for depression 0.71 and suicide 0.72
- F1-score for non-psycho 0.84.

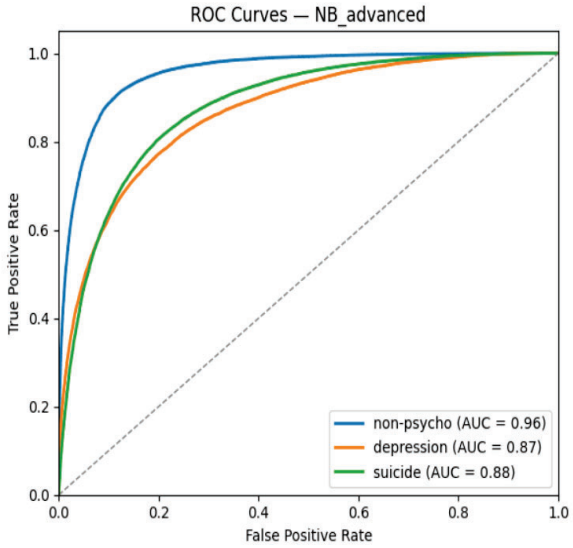


Fig. 7. ROC AUC

C. *SVM (TF-IDF + basic)*

Correct predictions for non-psycho 21,547 out of 23,208, for depression 17,143 out of 23,207, and for suicide 17,331 out of 23,207. The most common mistakes:

- Non-psycho → depression/suicide 909 and 752, respectively.
- Depression → suicide 4,300 cases.
- Suicide → depression 4,198 cases.

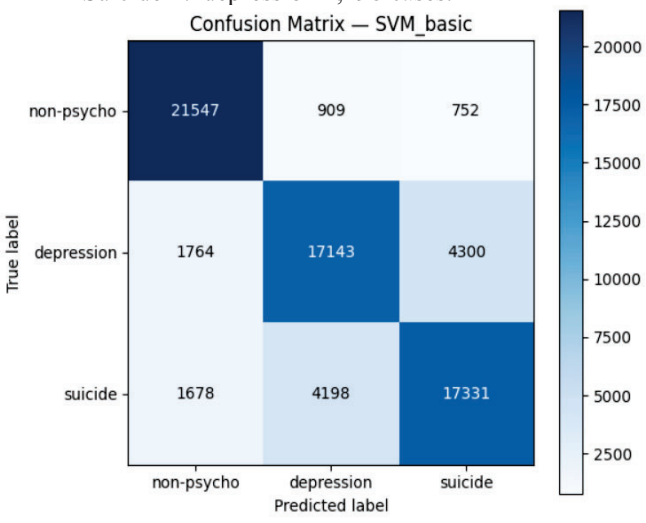
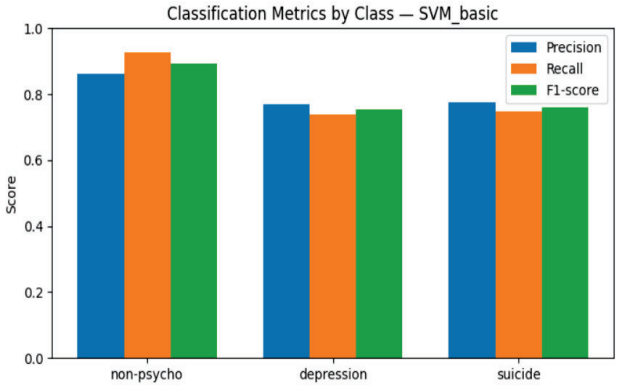


Fig. 8. Inaccuracy matrix

SVM shows slightly better accuracy in the non-psycho class. However, there is again a cross-confusion between depression and suicide. The algorithm generalises better than NB but worse than LR in some classes.



Classification report:

	precision	recall	f1-score	support
non-psycho	0.86	0.93	0.89	23208
depression	0.77	0.74	0.75	23207
suicide	0.77	0.75	0.76	23207
accuracy			0.80	69622
macro avg	0.80	0.80	0.80	69622
weighted avg	0.80	0.80	0.80	69622

Fig. 9. Precision, recall and F1-score report

D. CNN+Embending Core Model

Correct predictions (diagonal of the matrix): out of more than 23,200 messages that do not contain signs of mental risk, 21,300 are classified as non-psycho. Out of 23,207 real posts with manifestations of depression, 16,541 were correctly classified by the model. Out of 23,207 suicidal messages, 18,333 were detected. The most common mistakes:

- Non-psycho → depression/suicide only 846 and 1,062 cases, respectively – the false positive alarm rate for healthy texts is less than 1%, which means that this level of unnecessary "false alarms" is acceptable.
- Depression → suicide 5,652 times real depressive statements are "reclassified" as suicidal - indicates a substantial similarity in the vocabulary of the two subclasses.
- Suicide → depression 3,935 suicidal posts are treated as "depressive only" - the model lacks a subtle distinction between total sadness and direct signs of suicide.

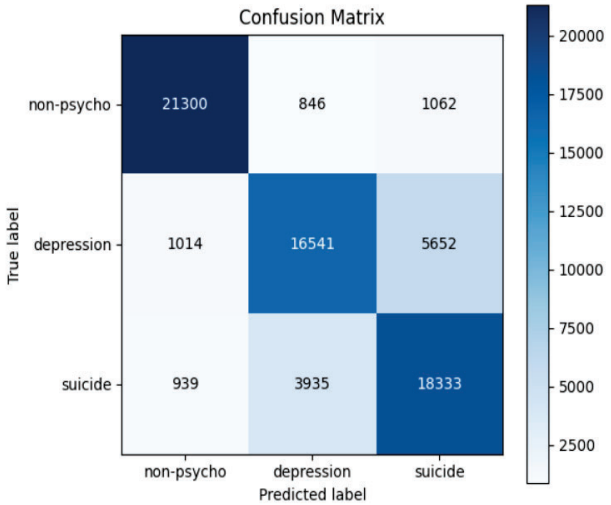
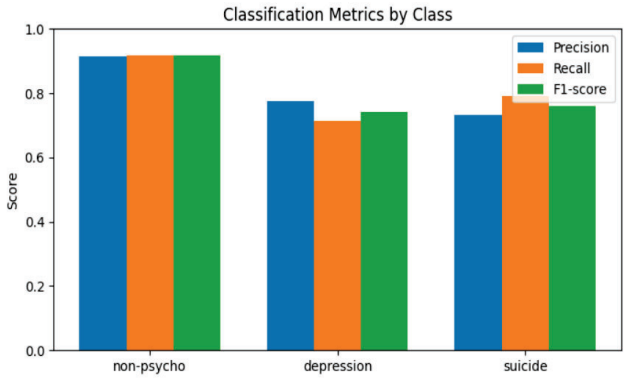


Fig. 10. Inaccuracy matrix

Classification errors in both directions between "depression" and "suicide" emphasise that linguistic signs of suicidal thoughts are often superimposed on common markers of depressed mood. Report on precision, recall and F1-score:



Classification report:

	precision	recall	f1-score	support
non-psycho	0.92	0.92	0.92	23208
depression	0.78	0.71	0.74	23207
suicide	0.73	0.79	0.76	23207
accuracy			0.81	69622
macro avg	0.81	0.81	0.81	69622
weighted avg	0.81	0.81	0.81	69622

Fig. 11. Precision, recall and F1-score report

Precision ("accuracy") for non-psycho is exceptionally high (0.92) - almost all texts predicted as "healthy" do not actually contain signs of risk. For both mental-risk classes, precision ≈ 0.78-0.73, that is, every fourth prediction by the model "depression" or "suicide" may turn out to be false anxiety.

Recall ("completeness") ≈ 0.92 for "non-psycho" shows that the model easily rejects healthy texts. Recall ≈ 0.71 and 0.79 for "depression" and "suicide" means that just over a quarter of such messages go unnoticed (false negatives).

F1-score combines precision and recall. F1 = 0.92 for "non-psycho" demonstrates the strengths of the model in detecting the absence of risk in a simple problem. F1 ≈ 0.74 and 0.76 for "depression" and "suicide" indicate an average level.

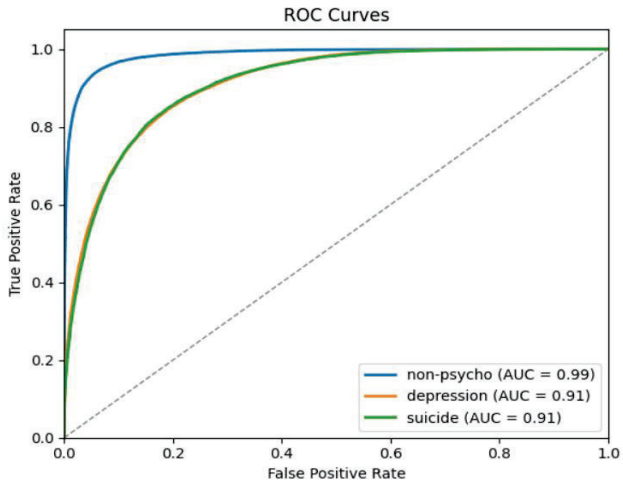


Fig. 12. ROC AUC

ROC AUCs for each class are 0.990 (non-psycho), 0.911 (depression), and 0.910 (suicide). AUC values close to 1.0 for "non-psychological" texts indicate the model's near-perfect ability to distinguish between this class. An AUC ≈ 0.91 for "depression" and "suicide" shows that the model ranks texts well by risk but requires fine-tuning of thresholds if higher detection completeness is desired.

VI. Discussions

Measuring the operation of the stored model:

1. The inference speed (~ 15 ms) allows you to integrate the model into a web interface or API with almost instant responses.
2. Low memory growth guarantees stable operation under high load without "bloating" RAM.
3. A whole load of a single CPU thread gives a predictable, linear path to scaling: each additional worker or core gives +66 req/s.
4. Successful identification of "healthy" texts.

Thanks to the high precision/recall for non-psycho, the platform can automatically reject most of the safe posts without the participation of a specialist, which optimises the resources of operators.

VII. Conclusions

As a result of the study, a prototype of the intelligent information and analytical platform MindGuard was developed, implemented, tested, and designed for the automated detection of depressive and suicidal patterns in social network texts. A corpus of data has been built and marked, including more than 340,000 examples of posts from Reddit with categories: non-psycho, depression and suicide. The developed flexible annotation scheme, with the involvement of experts, made it possible to ensure high-quality differentiation of texts by psychological states. A hybrid approach to text analysis has been implemented, combining classical methods (TF-IDF with Logistic Regression / SVM / Naive Bayes), deep neural networks (CNN with its embeddings) and transformer architectures (BERT, RoBERTa). As a result of experiments, it was found that the CNN model with GloVe embedding provides the best balance of accuracy, speed and interpretation. The following indicators were achieved: F1-score 0.92 (non-psycho), 0.74 (depression), 0.76 (suicide); AUC ROC 0.990, 0.911, 0.910, respectively; Latency ≈ 15 ms with a load of up to 66 requests/second on a single CPU stream. A system analysis of the platform architecture was carried out, and a web interface with REST API was implemented, which allows you to integrate the system quickly into the practice of psychological screening, content moderation, and educational initiatives. The analysis of errors showed that the most significant difficulties arise in the classification of intermediate states between depression and suicide. Areas for improvement have been identified: an increase in the number of examples with "moderate" signs of depression; use of transformers with fine-tuning; Data augmentation and expansion of the emotional spectrum. The system demonstrates high practical value: it allows timely detection of psychologically risky texts, reduces the workload on specialists and can be effectively used in psychological services, educational institutions, media platforms and public initiatives. Thus, the MindGuard platform has confirmed its effectiveness as a tool for early diagnosis of emotional disturbances in the digital environment, which opens up prospects for its further development and scaling in the context of digital psycho-hygiene.

Acknowledgement

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Adjustment of Limit Values of Transformer Oil Indicators Considering the Presence of Correlation Relationships Between Them

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***Abstract*—The article proposes a method for determining the limit values of transformer oil indicators simultaneously for a set of parameters, taking into account significant stochastic relationships between them, using the minimum risk method. The proposed method is distinguished by the fact that the limit values of the indicators are determined by minimizing the mean risk function for several oil indicators that are stochastically interrelated, using the Nelder-Mead method. A comparative analysis showed that the use of limit values obtained through the proposed method allows risk reduction by a factor of 1.2 to 7.8 compared to the risks associated with using the limit values specified in the standards IEC 60599:2022, IEEE Std C57.106-2015, and SOU-N EE 43-101:2009 for 110 kV transformers and 330 kV autotransformers. Moreover, the limit values obtained by minimizing the risk function for two indicators are accompanied by risk levels 1.3 to 23.7 times lower than those associated with the use of limit values also obtained by the minimum risk method, but for one-dimensional distributions. The obtained results indicate the high reliability of the proposed method.**

Keywords—diagnostics, transformer oil, limit values, correlation, breakdown voltage, moisture content, Weibull distribution, mean risk function for two indicators, Nelder-Mead method, risk analysis.

I. Introduction

The reliable and failure-free operation of electrical network equipment, including high-voltage power transformers—one of the most critical and expensive components—largely depends on the effectiveness and accuracy of diagnostic methods. Currently, decisions regarding the condition of equipment are generally made based on comparing the measured values of diagnostic indicators with corresponding threshold values. It is clear that under such conditions, the reliability of the decision depends to a large extent on how accurately these threshold values are defined and how well they account for the physical characteristics of the processes being diagnosed, as well as the factors influencing their intensity. At present, the decision on the condition of the equipment is generally made based on the results of comparing the measured values of diagnostic indicators with the corresponding threshold values. However, this method does not take into account the distribution of indicator values for equipment in a faulty condition, which can lead to the risk of erroneous decisions. Moreover, in most existing standards, threshold values of insulation indicators are normalized only based on the equipment's voltage class, while factors such as duration and operating modes are not considered, which complicates early-stage insulation diagnostics. Therefore, the development of methods for determining or adjusting threshold values of insulation indicators is a relevant and practically significant task.

II. Publication Analysis and Research Agenda

One of the insulation components of high-voltage power transformers, whose condition determines not only the electrical strength of oil gaps but also the mechanical strength [1–3], is transformer oil. The list of indicators used to assess the condition of oils, as well as the limit values of these indicators, are regulated by international and national standards, for example, [4–8]. An analysis of these standards clearly reveals significant differences both in the set of oil indicators used to determine its condition and in the limit values for the same indicator. The presence of discrepancies in the limit values of oil indicators can be a source of uncertainty when applying norms and criteria from different standards to the same oil samples. As mentioned earlier, the limit values of oil indicators are normalized according to the transformer voltage class, which aligns well with the findings reported in [9], where the influence of operating voltage on the aging intensity of oils was established. At the same time, when determining the limit values of oil indicators, neither the duration nor the operating modes of transformers are considered, although, as shown in [10], such consideration can significantly reduce the risk of erroneous decisions. The limit values of oil indicators are determined without accounting for correlation relationships between them; meanwhile, methods for assessing the condition of transformer oils developed in [11–13] demonstrate the high effectiveness of such consideration. Moreover, in some studies [14–16], the condition of transformer insulation is assessed using Health Index (T) parameters, which are calculated based on a set of indicators. A considerable number of studies have focused on developing fundamentally new decision-making methods for diagnosing transformer oil condition. For example, machine learning methods have been used in [17–19] to assess transformer oils, neural networks in [20], and regression models in [21–23]. However, despite the considerable volume of publications, the issues of accounting for correlation relationships between oil indicators when adjusting their limit values remain insufficiently addressed, which served as the motivation for the present research.

III. Methodology

Currently, the assessment of transformer oil condition during operation is performed based on a set of diagnostic indicators, each with its own prescribed limit values. At the same time, a single physicochemical process is characterized by multiple indicators. For example, the degree of oxidation of transformer oils is characterized by values such as the content of organic acids in the oil, water-soluble acids content, oil color, moisture content, dielectric loss tangent, and so forth. Since several different indicators change depending on the stage and intensity of the same process, it is logical that the limit values of these indicators could be determined jointly

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rather than separately for each indicator, as is currently done. Indicators for which joint determination of limit values is possible can be identified by analyzing the strength of stochastic relationships between them. Obviously, indicators characterizing the same physical process will be significantly correlated. As an example of determining limit values of oil indicators using the minimum risk method for two diagnostic parameters, consider two indicators: breakdown voltage (BDV) and moisture content of the oil. As shown in [23–25], a statistically significant relationship has been identified between these indicators, as evidenced by both the values of Pearson's pairwise correlation coefficients and the values of direct and inverse correlation ratios obtained by different researchers on independent samples. Moreover, in some methodologies –specifically in standards developed and used by the American company D. Maers [26] – BDV values are normalized depending on the moisture content of transformer oils. A comparative analysis of the threshold values of insulation indicators determined using various statistical decision methods, carried out in [27], showed that the highest reliability is provided by the threshold values obtained by the minimum risk method. When using this method, the threshold values are determined by minimizing the mean risk function, which in general form can be represented as:

$$R=P_1\cdot P_{11}\cdot C_{11}+P_1\cdot P_{21}\cdot C_{21}+P_2\cdot P_{22}\cdot C_{22}+P_2\cdot P_{12}\cdot C_{12} \quad (1)$$

where P_1 is a priori probability that the oil will be in a "suitable for further operation" condition; P_2 is a priori probability that the oil will be in an "unsuitable for further operation" condition. P_{11} is probability of making a correct decision when diagnosing oil suitable for operation; P_{21} is probability of a Type I error ("false alarm"); P_{22} is probability of making a correct decision when diagnosing oil unsuitable for operation; P_{12} is probability of a Type II error ("missed detection"); C_{11} is conditional cost of making a correct decision when diagnosing oil suitable for operation; C_{21} is conditional cost of a Type I error; C_{22} is conditional cost of making a correct decision when diagnosing oil unsuitable for operation; C_{12} is conditional cost of a Type II error.

To determine the values of the probabilities of correct (P_{11} and P_{22}) and erroneous (P_{21} and P_{12}) decisions, it is necessary to formulate a decision rule according to which the diagnosis will be made when classifying the condition of transformer oils based on two indicators simultaneously (BDV and moisture content of oils). By extending the decision rules for each of the individual indicators, which are regulated in standards [4–6], to the two-dimensional case, and considering that during the aging process the values of these indicators change in opposite directions (BDV decreases, while moisture content increases), the decision rule for assessing the oil condition by these two indicators can be formulated as (2). In (2) x is the oil under diagnosis; D_1 is oil suitable for operation; D_2 is oil unsuitable for operation; U_{BDV_0} is threshold value of the oil's BDV; W_{m_0} is threshold value of the oil's moisture content; U_{BDV} is BDV value of the oil under diagnosis; W_{m} is moisture content value of the oil under diagnosis.

$$\begin{gathered} x\in D_1 \text{when } U_{\mathrm{BDV}}>U_{\mathrm{BDV}_0} \text{ and } W_{\mathrm{m}}<W_{\mathrm{m}_0} \\ x\in D_2 \text{when } U_{\mathrm{BDV}}<U_{\mathrm{BDV}_0} \text{ and } W_{\mathrm{m}}>W_{\mathrm{m}_0} \text{ or} \\ \text{when } U_{\mathrm{BDV}}<U_{\mathrm{BDV}_0} \text{ and } W_{\mathrm{m}}<W_{\mathrm{m}_0} \text{ or} \\ when\ U_{\mathrm{BDV}}>U_{\mathrm{BDV}_0} \text{ and } W_{\mathrm{m}}>W_{\mathrm{m}_0} \end{gathered} \quad (2)$$

According to [28], the quality indicators for both suitable and unsuitable transformer oil are distributed according to the Weibull law. Taking into account the decision rule (2), as well as the results of the analysis of the distribution laws of oil indicators, expressions were obtained for determining the values of the probabilities of correct and incorrect decisions. The probability of making a correct decision when assessing the condition of transformer oil that is suitable for operation is determined by the probability that the values of its indicators (BDV and moisture content) will lie within the acceptable range (i.e., the BDV value obtained from testing exceeds the threshold value, and the moisture content value is below the threshold value):

$$\begin{gathered} P_{11}=P_1\cdot Ph_{11}= \\ =P_1\cdot\int_{U_{\mathrm{BDV}_0}}^{\infty}\frac{\beta_{U_1}}{\alpha_{U_1}^{\beta_{U_1}}}\cdot {U_{\mathrm{BDV}}}^{\beta_{U_1}\text{-}1}\,e^{-\left(\frac{U_{\mathrm{BDV}}}{\alpha_{U_1}}\right)^{\beta_{U_1}}}dU_{\mathrm{BDV}}\times \\ \times\int_0^{W_{\mathrm{m}_0}}\frac{\beta_{W_1}}{\alpha_{W_1}^{\beta_{W_1}}}\cdot {W_{\mathrm{m}}}^{\beta_{W_1}\text{-}1}\cdot e^{-\left(\frac{W_{\mathrm{m}}}{\alpha_{W_1}}\right)^{\beta_{W_1}}}dW_{\mathrm{m}} \end{gathered} \quad (3)$$

where Ph_{11} is the probability that, for oil suitable for operation, the BDV value will be greater than the threshold and the moisture content will be less than the threshold; α_{U1} is the scale parameter of the Weibull distribution for the BDV of oil suitable for operation; β_{U1} is the shape parameter of the Weibull distribution for the BDV of oil suitable for operation; α_{W1} is the scale parameter of the Weibull distribution for the moisture content of oil suitable for operation; β_{W1} is the shape parameter of the Weibull distribution for the moisture content of oil suitable for operation.

The probability of a false alarm (Type I error) is also determined by the product of several probabilities. First, by the probability that the diagnosed transformer oil is actually suitable for operation. Second, by the probability that for oil suitable for operation, either the BDV value is below the threshold, or the moisture content is above the threshold, or the BDV is below the threshold and the moisture content is above the threshold (4). In (4) Ph_{21} is the probability that for oil suitable for operation, either the BDV is below the threshold, or the moisture content is above the threshold, or both the BDV is below the threshold and the moisture content is above the threshold. The probability of making a correct decision when diagnosing the condition of oil unsuitable for operation is also determined by the product of several probabilities. In particular, by the probability that the diagnosed transformer oil is indeed unsuitable for operation, and the probabilities that for such oil, either the BDV is below the threshold, or the moisture content is above the threshold, or both the BDV is below the threshold and the moisture content is above the threshold (5). In (5) Ph_{22} is the probability that for oil unsuitable for operation, either the BDV is below the threshold, or the moisture content is above the threshold, or both the BDV is below the threshold and the moisture content is above the threshold; α_{U2} is the scale parameter of the Weibull distribution for the BDV of oil unsuitable for operation; β_{U2} is the shape parameter of the Weibull distribution for the BDV of oil unsuitable for operation; α_{W1} is the scale parameter of the Weibull distribution for the moisture content of oil unsuitable for operation; β_{W1} is the shape parameter of the Weibull distribution for the moisture content of oil unsuitable for operation.

$$P_{21} = P_1 \cdot Ph_{21} = P_1 \cdot \left[\left(\int_0^{U_{\text{BDV}0}} \frac{\beta_{U_1}}{\alpha_{U_1}^{\beta_{U_1}}} \cdot {U_{BDV}}^{\beta_{U_1}-1} \cdot e^{-\left(\frac{U_{BDV}}{\alpha_{U_1}}\right)^{\beta_{U_1}}} dU_{BDV} \cdot \int_{W_{M0}}^{\infty} \frac{\beta_{W_1}}{\alpha_{W_1}^{\beta_{W_1}}} \cdot {W_M}^{\beta_{W_1}-1} \cdot e^{-\left(\frac{W_M}{\alpha_{W_1}}\right)^{\beta_{W_1}}} dW_M \right) + \right.$$

$$+ \left(\int_{U_{\text{BDV}0}}^{\infty} \frac{\beta_{U_1}}{\alpha_{U_1}^{\beta_{U_1}}} \cdot {U_{BDV}}^{\beta_{U_1}-1} \cdot e^{-\left(\frac{U_{BDV}}{\alpha_{U_1}}\right)^{\beta_{U_1}}} dU_{BDV} \cdot \int_{W_{\text{M}0}}^{\infty} \frac{\beta_{W_1}}{\alpha_{W_1}^{\beta_{W_1}}} \cdot {W_{\text{M}}}^{\beta_{W_1}-1} \cdot e^{-\left(\frac{W_{\text{M}}}{\alpha_{W_1}}\right)^{\beta_{W_1}}} dW_{\text{M}} \right) +$$

$$\left. + \int_0^{U_{\text{BDV}0}} \frac{\beta_{U_1}}{\alpha_{U_1}^{\beta_{U_1}}} \cdot {U_{BDV}}^{\beta_{U_1}-1} \cdot e^{-\left(\frac{U_{BDV}}{\alpha_{U_1}}\right)^{\beta_{U_1}}} dU_{BDV} \cdot \int_0^{W_{\text{M}0}} \frac{\beta_{W_1}}{\alpha_{W_1}^{\beta_{W_1}}} \cdot {W_{\text{M}}}^{\beta_{W_1}-1} \cdot e^{-\left(\frac{W_{\text{M}}}{\alpha_{W_1}}\right)^{\beta_{W_1}}} dW_{\text{M}} \right] \quad (4)$$

$$P_{22} = P_2 \cdot Ph_{22} = P_1 \cdot \left[\left(\int_0^{U_{\text{BDV}0}} \frac{\beta_{U_2}}{\alpha_{U_2}^{\beta_{U_2}}} \cdot {U_{BDV}}^{\beta_{U_2}-1} \cdot e^{-\left(\frac{U_{BDV}}{\alpha_{U_2}}\right)^{\beta_{U_2}}} dU_{BDV} \cdot \int_{W_{M0}}^{\infty} \frac{\beta_{W_2}}{\alpha_{W_2}^{\beta_{W_2}}} \cdot {W_{\text{M}}}^{\beta_{W_2}-1} \cdot e^{-\left(\frac{W_{\text{M}}}{\alpha_{W_2}}\right)^{\beta_{W_2}}} dW_{\text{M}} \right) + \right.$$

$$+ \left(\int_0^{U_{\text{BDV}0}} \frac{\beta_{U_2}}{\alpha_{U_2}^{\beta_{U_2}}} \cdot {U_{BDV}}^{\beta_{U_2}-1} \cdot e^{-\left(\frac{U_{BDV}}{\alpha_{U_2}}\right)^{\beta_{U_2}}} dU_{BDV} \cdot \int_0^{W_{\text{M}0}} \frac{\beta_{W_2}}{\alpha_{W_2}^{\beta_{W_2}}} \cdot {W_{\text{M}}}^{\beta_{W_2}-1} \cdot e^{-\left(\frac{W_{\text{M}}}{\alpha_{W_2}}\right)^{\beta_{W_2}}} dW_{\text{M}} \right) +$$

$$\left. + \left[\left(\int_{U_{\text{BDV}0}}^{\infty} \frac{\beta_{U_2}}{\alpha_{U_2}^{\beta_{U_2}}} \cdot {U_{BDV}}^{\beta_{U_2}-1} \cdot e^{-\left(\frac{U_{\text{пр.}}}{\alpha_{U_2}}\right)^{\beta_{U_2}}} dU_{BDV} \cdot \int_{W_{\text{M}0}}^{\infty} \frac{\beta_{W_2}}{\alpha_{W_2}^{\beta_{W_2}}} \cdot {W_{\text{M}}}^{\beta_{W_2}-1} \cdot e^{-\left(\frac{W_{\text{M}}}{\alpha_{W_2}}\right)^{\beta_{W_2}}} dW_{\text{M}} \right) \right] \right. \quad (5)$$

The probability of a missed detection (Type II error) will be determined by the probability that the diagnosed oil is unsuitable for operation, and at the same time, both the BDV value obtained from testing and the moisture content value obtained from testing will correspond to oil suitable for operation, i.e., will be within the acceptable range:

$$P_{12} = P_2 \cdot Ph_{21} =$$

$$P_2 \cdot \left(\int_{U_{\text{BDV}0}}^{\infty} \frac{\beta_{U_2}}{\alpha_{U_2}^{\beta_{U_2}}} \cdot {U_{\text{BDV}}}^{\beta_{U_2}-1} \cdot e^{-\left(\frac{U_{\text{BDV}}}{\alpha_{U_2}}\right)^{\beta_{U_2}}} dU_{\text{BDV}} \times \right.$$

$$\left. \times \int_0^{W_{\text{m}0}} \frac{\beta_{W_2}}{\alpha_{W_2}^{\beta_{W_2}}} \cdot {W_{\text{m}}}^{\beta_{W_2}-1} \cdot e^{-\left(\frac{W_{\text{M}}}{\alpha_{W_2}}\right)^{\beta_{W_2}}} dW_{\text{m}} \right) \quad (6)$$

In (6) Ph_{21} is the probability that for oil unsuitable for operation, the BDV value will be above the threshold, and the moisture content value will be below the threshold. Given the complexity of the obtained average risk function, it is advisable to use numerical methods to determine the boundary values of the analyzed indicators, i.e., to minimize this function. As shown in [29], a sufficiently effective method for minimizing a multivariable function is the Nelder–Mead method, which was used to determine the boundary values that provide the minimum of the average risk function. As initial data for the calculations, the values of the Weibull distribution parameters for BDV and moisture content of oils in different conditions for 110 kV transformers and 330 kV autotransformers, presented in [28], were used. For oil suitable for operation, both for 110 kV transformers and 330 kV autotransformers, the scale and shape parameters used corresponded to relatively fresh oil after 5 years of operation (No. 1), as well as aged oil after 20 years of operation (No. 2). The initial data for the calculation are given in Table I. Studies carried out in [30] show that technological faults caused by moisture and contamination of oils account for about 2.18% of all technological faults. Therefore, the prior probability that the oil will be unfit for use is taken as $P_2 = 0.02$. The prior probability that the oil will be fit for use is determined as $P_1 = 1 - P_2 = 0.98$. The conditional cost of a false alarm is taken as $C_{21} = 1$, and the conditional cost of a missed detection is $C_{12} = 1000$. The calculation results are presented in Table II. The presence of a global minimum of the average risk function is clearly illustrated in Fig. I, which shows the dependencies of the average risk values on the BDV, moisture content, and on both the BDV and moisture content of transformer oils for 110 kV transformers (No. 1).

TABLE I. SCALE AND SHAPE PARAMETERS OF THE WEIBULL DISTRIBUTION FOR BDV AND MOISTURE CONTENT OF OILS WITH DIFFERENT CONDITIONS FOR 110 KV TRANSFORMERS AND 330 KV AUTOTRANSFORMERS

No.	Oil suitable for operation				Oil unsuitable for operation			
	BDV		Moisture content		BDV		Moisture content	
	α	β	α	β	α	β	α	β
110 kV transformers								
1	75.623	7.517	6.875	2.176	29.459	7.311	39.98	4.059
2	75.623	7.517	6.875	2.176				
330 kV autotransformers								
3	77.523	11.48	7.2755	2.587	42.576	23.25	31.232	7.196
4	69.855	8.038	10.762	4.407				

TABLE II. THRESHOLD VALUES OF BDV AND MOISTURE CONTENT OF OILS OBTAINED BY THE MINIMUM RISK METHOD FOR MULTIDIMENSIONAL DISTRIBUTIONS FOR 110 KV TRANSFORMERS AND 330 KV AUTOTRANSFORMERS

No.	Parameters of the distribution laws for oil suitable for operation				Limit values	
	U_{BDV}		Moisture content		BDV, kV	Moisture Content, g/t
	α	β	α	β		
110 kV transformers						
1	75.623	7.517	6.875	2.176	38.3	18.4
2	51.833	8.704	8.9923	2.239	36.2	19.4
330 kV autotransformers						
1	77.523	11.48	7.2755	2.587	46.4	17.6
2	69.855	8.038	10.762	4.407	45.5	17.1

As can be seen from the Fig. I, a clear global minimum of the average risk function is observed on all three dependencies, which is ensured by the obtained threshold

values. At the same time, during the minimization process, the risk values decrease by orders of magnitude.

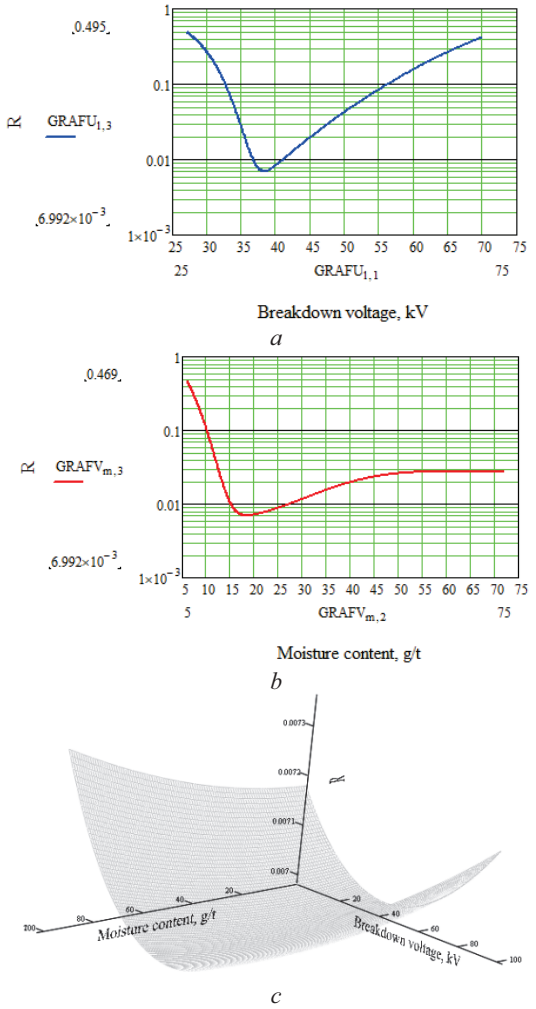


Fig. 1. Dependencies of the average risk value on the BDV (*a*), moisture content (b), and on both the BDV and moisture content (*c*) of transformer oils for 110 kV transformers (sample No. 1)

Analyzing the results presented in Table II, it should be noted that the threshold values of both indicators obtained for transformers of the same voltage class but with different scale and shape parameter values differ significantly. This means that for minimizing risks when assessing the condition of transformer oils, it is not sufficient to consider only the voltage class of the transformers, as regulated by the current normative documents both in Ukraine and abroad. The minimum risk value will be ensured only by taking into account the type and quality of transformer oils, as well as their operating modes and conditions, which determine the values of the scale and shape parameters.

IV. ANALYSIS OF THE OBTAINED RESULTS

The analysis of the obtained boundary values was carried out in two stages. At the first stage, similarly to [10, 29], a comparative analysis was performed of the probabilities of correct and erroneous decisions and the values of the average risk associated with the use of the obtained boundary values and the boundary values of BDV and moisture content of transformer oils regulated by the current Ukrainian standard SOU-N EE 43-101:2009 [6] and international standards IEC 60599:2022 [4] and IEEE Std C57.106-2015 [5]. The calculation results for 110 kV transformers are presented in Table III, and for 330 kV autotransformers in Table IV.

TABLE III. VALUES OF PROBABILITIES OF CORRECT AND ERRONEOUS DECISIONS, AS WELL AS RISK VALUES ASSOCIATED WITH THE USE OF LIMIT VALUES OF BDV AND MOISTURE CONTENT OF OILS REGULATED BY DIFFERENT STANDARDS AND CALCULATED BY THE MINIMUM RISK METHOD FOR MULTIVARIATE DISTRIBUTIONS FOR 110 KV TRANSFORMERS

No.	Probabilities of correct and erroneous decisions				*R*
	Ph_{11}/P_{11}	Ph_{21}/P_{21}	Ph_{22}/P_{22}	Ph_{12}/P_{12}	
Ukrainian standard, U_{BDV} (limit) = 35 kV, Moisture content (limit) = 20 g/t					
1	0.997 0.996	$3.08\cdot10^{-3}$ $3.025\cdot10^{-3}$	0.998 0.01996	$1.717\cdot10^{-3}$ $3.435\cdot10^{-5}$	$37.373\cdot10^{-3}$
2	0.965 0.946	0.035 0.034	0.998 0.01997	$1.717\cdot10^{-3}$ $3.435\cdot10^{-5}$	0.068
IEC standard, U_{BDV} (limit) = 40 kV, Moisture content (limit) = 30 g/t					
1	0.992 0.972	$8.299\cdot10^{-3}$ $8.133\cdot10^{-3}$	0.99997 0.019999	$2.309\cdot10^{-5}$ $4.618\cdot10^{-7}$	$8.595\cdot10^{-3}$
2	0.9 0.882	0.1 0.098	0.99998 0.0199995	$2.309\cdot10^{-5}$ $4.618\cdot10^{-7}$	0.098
IEEE Standard, U_{BDV} (limit) = 47 kV, Moisture content (limit) = 25 g/t					
1	0.972 0.953	0.028 0.027	0.99999 0.019999	$8.44\cdot10^{-15}$ $1.688\cdot10^{-16}$	$27.069\cdot10^{-3}$
2	0.653 0.64	0.347 0.34	0.99999 0.019999	$8.44\cdot10^{-15}$ $1.688\cdot10^{-16}$	0.340
Minimum Risk Method					
1	0.994 0.974	$6.234\cdot10^{-3}$ $6.11\cdot10^{-3}$	0.99996 0.019999	$4.408\cdot10^{-5}$ $8.815\cdot10^{-7}$	$6.991\cdot10^{-3}$
2	0.953 0.934	0.047 0.046	0.99946 0.01998	$5.414\cdot10^{-4}$ $1.083\cdot10^{-5}$	0.057

TABLE IV. VALUES OF PROBABILITIES OF CORRECT AND ERRONEOUS DECISIONS, AS WELL AS RISK VALUES ASSOCIATED WITH THE USE OF LIMIT VALUES OF BDV AND MOISTURE CONTENT OF OILS REGULATED BY DIFFERENT STANDARDS AND CALCULATED BY THE MINIMUM RISK METHOD FOR MULTIVARIATE DISTRIBUTIONS FOR 330 KV AUTOTRANSFORMERS

No.	Values of probabilities of correct and erroneous decisions				*R*
	Ph_{11}/P_{11}	Ph_{21}/P_{21}	Ph_{22}/P_{22}	Ph_{12}/P_{12}	
Ukraine standard, BDV (limit) = 45 kV, Moisture Content (limit) = 20 g/t					
1	0.998 0.978	$1.941\cdot10^{-3}$ $1.902\cdot10^{-3}$	0.9989 0.01998	$1.059\cdot10^{-3}$ $2.117\cdot10^{-5}$	0.023
2	0.971 0.952	0.029 0.028	0.9989 0.01998	$1.059\cdot10^{-3}$ $2.117\cdot10^{-5}$	0.049
IEC and IEEE standards, BDV (limit) = 50 kV, Moisture Content (limit) = 20 g/t					
1	0.994 0.974	$6.489\cdot10^{-3}$ $6.359\cdot10^{-3}$	1 0.02	$2.342\cdot10^{-20}$ $4.685\cdot10^{-22}$	0.0064
2	0.934 0.916	0.066 0.064	1 0.02	$2.342\cdot10^{-20}$ $4.685\cdot10^{-22}$	0.064
Minimum Risk Method					
1	0.997 0.977	$2.828\cdot10^{-3}$ $2.771\cdot10^{-3}$	0.999991 0.01999	$9.005\cdot10^{-6}$ $1.801\cdot10^{-7}$	0.00295
2	0.968 0.949	0.032 0.031	0.9998 0.019998	$1.1\cdot10^{-4}$ $2.2\cdot10^{-6}$	0.0336

Analyzing the risk values given in Tables III and IV, it should be noted first of all that for both 110 kV transformers and 330 kV autotransformers, the risk values increase as the aging degree of the transformer oils increases (i.e., as the scale parameter α for BDV decreases and the scale parameter α for moisture content increases). This increase in risk values is observed also for the boundary values of BDV and moisture content determined by minimizing the average risk function. Moreover, the same boundary values of indicators (for example, those recommended by the IEC standard for 110 kV transformers from Table III) when applied to oil in different conditions can ensure both minimal and maximal risk values in comparison with risks ensured by boundary values regulated by other standards. This circumstance further confirms the conclusion made in [10] about the necessity to consider the grade and quality of transformer oils, as well as the operating modes and conditions of equipment, when determining the limit values of insulation parameters.

Similarly to the case of determining the boundary values of transformer oil indicators by the minimum risk method for one-dimensional distributions, ensuring the minimum value of the Type II error probability does not guarantee the minimum risk value. This can be seen by comparing the Type II error probabilities and average risk values that accompany the use of moisture content and BDV boundary values recommended by the IEEE standard with the corresponding values obtained using boundary values derived by the minimum risk method. As can be seen from Tables III and IV, the minimum risk values among all cases occur when using the boundary values of BDV and moisture content obtained by minimizing the average risk function. At the same time, using the boundary values obtained by the minimum risk method for 110 kV transformers, depending on the condition of the diagnosed oil, allows reducing the probable risks by 1.2 to 5.35 times compared to the risks arising from the use of boundary values of these indicators regulated by the [6]. The risk reduction compared to the risks ensured by boundary values regulated by the [4] is 1.2 to 1.7 times, and compared to the boundary values regulated by [5], the risks decrease by 3.8 to 5.9 times. Using the boundary values obtained for 330 kV autotransformers by the minimum risk method allows reducing risks by 1.46 to 7.8 times compared to the risks accompanying the use of boundary values regulated by the [6], and by 1.9 to 2.1 times compared to the risks ensured by boundary values regulated in the [4] and [5] standards. Thus, using the boundary values of BDV and moisture content obtained by the minimum risk method reduces risks by 1.2 to 7.8 times compared to the risk values arising from using the limit values of these indicators regulated by well-known national and international standards. At the second stage, a comparative analysis of risks accompanying the use of limit values of BDV and moisture content obtained by the minimum risk method for two indicators was performed with the risk values accompanying the use of limit values of BDV and moisture content, which were also obtained by the minimum risk method but for one-dimensional distributions. The mean risk functions for the one-dimensional distributions of these indicators are presented in [27, 29], and the threshold values were determined using the Newton method. As in the previous case, the calculations were performed with prior probabilities of the condition of transformer oils equal to P_1 = 0.98 and P_2 = 0.02. The conditional cost of a false alarm was set as C_{21} = 1, and the conditional cost of a missed detection as C_{12} = 1000. The obtained threshold values of the indicators for the corresponding parameters of the distribution laws are presented in Table V. Comparing the threshold values of BDV and moisture content in oils obtained using the minimum risk method for multivariate distributions (Table II) with the threshold values of these indicators obtained using the minimum risk method for univariate distributions (Table V), it is easy to see that they differ significantly. In this regard, it is of particular interest to compare the values of correct and incorrect decision probabilities, as well as the average risk values that accompany the use of threshold values obtained by the minimum risk method for univariate and multivariate distributions. The calculated values of correct and incorrect decision probabilities, as well as the risk values that accompany the use of threshold values of BDV and moisture content in oils regulated by various standards and obtained by the minimum risk method for univariate distributions for 110 kV transformers and 330 kV autotransformers are presented in Table VI. Comparing the corresponding risk values from Table VI with those in Tables IV and V, it should be noted that the use of threshold values obtained by the minimum risk method for multivariate distributions ensures the minimum risk value not only compared to threshold values regulated by well-known national and international standards but also compared to threshold values obtained by the minimum risk method for univariate distributions. Thus, using the BDV threshold values of transformer oils obtained by the minimum risk method for multivariate distributions for 110 kV transformers allows reducing possible risks by 1.3–1.5 times compared to the risks associated with the use of BDV threshold values obtained by the minimum risk method for univariate distributions. For moisture content of transformer oils in 110 kV transformers, the risk reduction is even more significant, ranging from 5.1 to 23.7 times. For 330 kV autotransformers, the risk reduction when using BDV threshold values obtained by the minimum risk method for multivariate distributions compared to the risks accompanying the use of threshold values obtained by the minimum risk method for univariate distributions is 1.19–1.22 times. A similar risk reduction for moisture content ranges from 3 to 15 times. The obtained results testify to the high effectiveness of the proposed method.

TABLE V. THRESHOLD VALUES OF BDV AND MOISTURE CONTENT OF OILS OBTAINED USING THE MINIMUM RISK METHOD FOR UNIVARIATE DISTRIBUTIONS FOR 110 KV TRANSFORMERS AND 330 KV AUTOTRANSFORMERS

No.	Parameters of the Distribution Laws for Oil Suitable for Operation				Limit Permissible Values	
	BDV		Moisture Content		BDV, kV	Moisture Content, g/t
	α	β	α	β		
Transformers 110 kV						
1	75.623	7.517	6.875	2.176	40.37	10.78
2	51.833	8.704	8.9923	2.239	38.78	12.04
Autotransformers 330 kV						
1	77.523	11.48	7.2755	2.587	47.35	12.41
2	69.855	8.038	10.762	4.407	46.9	14.2

TABLE VI. VALUES OF PROBABILITIES OF CORRECT AND INCORRECT DECISIONS, AS WELL AS RISK VALUES ASSOCIATED WITH THE USE OF THRESHOLD VALUES OF BDV AND MOISTURE CONTENT OF OILS REGULATED BY VARIOUS STANDARDS AND CALCULATED BY THE MINIMUM RISK METHOD FOR UNIVARIATE DISTRIBUTIONS FOR 110 KV TRANSFORMERS AND 330 KV AUTOTRANSFORMERS

No.	Values of probabilities of incorrect and correct decisions				*R*
	Ph_{11}/P_{11}	Ph_{21}/P_{21}	Ph_{22}/P_{22}	Ph_{12}/P_{12}	
Transformers 110 kV					
BDV					
1	0.991 0.971	$8.891\cdot10^{-3}$ $8.713\cdot10^{-3}$	0.99995 0.0199991	$4.494\cdot10^{-5}$ $8.988\cdot10^{-7}$	$9.612\cdot10^{-3}$
2	0.923 0.905	0.077 0.075	0.9994 0.01998	$5.745\cdot10^{-4}$ $1.149\cdot10^{-5}$	0.0869
Moisture content					
1	0.93 0.912	0.07 0.068	0.995 0.01990	$4.88\cdot10^{-3}$ $9.761\cdot10^{-5}$	0.166
2	0.854 0.837	0.146 0.143	0.992 0.019847	$7.633\ 10^{-3}$ $1.527\cdot10^{-4}$	0.296
Autotransformers 330 kV					
BDV					
1	0.997 0.977	$3.477\cdot10^{-3}$ $3.408\cdot10^{-3}$	0.999993 0.019999	$7.259\cdot10^{-6}$ $1.452\cdot10^{-7}$	0.0035
2	0.96 0.941	0.04 0.039	0.99993 0.019998	$6.967\cdot10^{-5}$ $1.393\cdot10^{-6}$	0.041
Moisture content					
1	0.981 0.962	0.019 0.018	0.998 0.01997	$1.304\cdot10^{-3}$ $2.608\cdot10^{-5}$	0.0444
2	0.966 0.947	0.034 0.033	0.997 0.01993	$3.435\cdot10^{-3}$ $6.871\cdot10^{-5}$	0.102

CONCLUSION

A method has been proposed for determining threshold values simultaneously for two indicators of transformer oils, taking into account the presence of significant stochastic

relationships between them. The proposed method is distinguished by the fact that the threshold values of the indicators are determined by minimizing the mean risk function for correlated oil indicators using the Nelder–Mead method, while considering the duration, operating modes, and conditions of transformer operation. The studies carried out showed that the use of threshold values of breakdown voltage and moisture content of oils obtained with the proposed method makes it possible to reduce the risk values for 110 kV transformers by 1.2–5.9 times compared to the risks associated with using the threshold values of breakdown voltage and moisture content of transformer oils regulated by the standards [4–6]. For 330 kV autotransformers, a similar risk reduction amounted to 1.46–7.8 times. It has been established that the use of threshold values obtained by the minimum risk method for multidimensional distributions ensures the minimum risk value not only compared with the limit values regulated by known national and international standards, but also with the limit values obtained by the minimum risk method for one-dimensional distributions. For 110 kV transformers, the risk reduction was 1.3–1.5 times for the threshold values of oil breakdown voltage and from 5.1 to 23.7 times for the moisture content of transformer oils. For 330 kV autotransformers, the risk reduction was 1.19–1.22 times for breakdown voltage and 3–15 times for the moisture content of transformer oils. The obtained results indicate the high efficiency of the proposed method.

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Measurement of Dielectric Dissipation Factor by Applying Sine Fourier Transform and the Least Quares Method

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***Abstract*—This paper is dedicated to the development of a method for measuring the dielectric dissipation factor by processing the signals of current and voltage. The developed method is based on several calculations of inner products between the analyzed signals and the auxiliary sine functions, which have different frequencies. Further calculations involve the formation of the overdetermined system of equations, which is solved by applying the least squares method. The solution of this system allows to determine the in-phase and quadrature components of the analyzed signals, their initial phase, phase shift, dielectric loss angle and dielectric dissipation factor. The results of simulations, made for analyzed signals with additive noise components, are also given. The presented analysis is oriented on technical applications which concern testing of the quality of electrical insulation in power equipment; however, the obtained results can be used for other applications that require the measurement of the phase shift between the sine signals with noise components.**

Keywords—noise stability, diagnostics of insulation, phase measurements, root mean square error, dielectric power losses.

I. INTRODUCTION

The value of dielectric dissipation factor (tgδ) is often used in order to determine the level of deterioration of electrical insulation in power and communication cables [1-4], rotating electrical machines [5, 6], various transformers [7, 8], and other types of electrical engineering equipment. Such deterioration occurs due to numerous deleterious factors such as high temperatures, mechanical tension and electrical stress, impact of partial discharges, accumulation of volume charge, penetration of moisture, and other reasons, which eventually may lead to the breakdown of electrical insulation. Possible implications of such breakdowns may result in interruptions of power supply, emergency situations, and significant economic loss. This circumstance determines the relevance of further improvement of the existing methods of testing the quality of electrical insulation, which would allow to increase the accuracy of measurements of its dielectric parameters and to enhance the overall reliability of control.

In the majority of cases, measurements of dielectric properties of insulation in electrical equipment are carried out by applying the same methods and instrumentation that have been developed for the purposes of impedance measurement. In recent years, the improvement of these methods has focused on two main directions. The first direction implies the improvement of the hardware used in the existing tools of measurement. Such improvement may involve applying more efficient circuit design solutions, increasing the operation speed, and applying technical solutions that allow the removal of the effect of parasitic parameters of tested objects [9, 10]. The second direction is concentrated on the application of new methods, as well as the improvement of the existing techniques in order to more efficiently process the obtained signals of current and voltage [11]. The main studies in this field are focused on the enhancement of level of noise stability, decreasing of impact of spectral leakage and improvement of computational efficiency of the existing digital signal processing algorithms.

In the case of the development of various digital signal processing techniques, intended for the purpose of dielectric dissipation factor measurement, it is necessary to take into account the existence of some contradictory requirements for the developed methods. For instance, the application of specific interpolation algorithms for frequency-domain methods usually allows for decreasing the undesirable effect of spectral leakage. However, this approach, at the same time, may lead to the reduction of noise stability of the developed methods [12]. In addition to such a reduction, many methods allow for the attainment of pretty unstable metrological characteristics of the obtained results of measurements. A typical example of such methods is the correlation method for phase shift measurement, which employs properties of autocorrelation and cross-correlation functions of the analyzed signals. In the case of a close to 90º value of phase difference between the analyzed signals, this method exhibits a high level of noise stability [13], which makes it a perfect tool for the measurement of tgδ of tested objects with a high level of quality factor. However, in the case of comparatively low values of phase difference, the level of noise stability of this method substantially decreases. It complicates the applying of the cross-correlation method for the measurement of tgδ of tested objects with a decreased level of quality factor, which is

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usually observed under the low frequencies (below 0.1 Hz) of test voltage.

Consequently, the problem of developing improved methods for processing signals of current and voltage, which allow for determining the level of tgδ in electrical insulation of power equipment, still remains relevant. The described problem is addressed in the current paper, which is dedicated to the development of a method for the measurement of tgδ based on aggregate measurements.

II. MEASUREMENT OF DIELECTRIC DISSIPATION FACTOR BY APPLYING THE LEAST SQUARES METHOD IN FREQUENCY DOMAIN

A. Models of the Analyzed Signals

Further analysis will be premised on the assumption that the analyzed signals of current and voltage have some undesirable noise component and can be expressed by using the following models:

$$i(n\Delta t) = A_{2I}\sin(\omega_2 n\Delta t + \varphi_{2I}) + N_I(n\Delta t), \quad (1)$$

$$u(n\Delta t) = A_{2U}\sin(\omega_2 n\Delta t + \varphi_{2U}) + N_U(n\Delta t), \quad (2)$$

where: ω_2, denotes the angular frequency of the analyzed signals, A_{2U}, φ_{2U}, A_{2I}, φ_{2I} correspondingly denote the amplitude and the initial phase of the analyzed sine signal of current and voltage, Δt is the sampling interval, $n = 0..N\text{-}1$ and denotes the serial number of a particular element within the whole array of sampled elements of a size N, $N_I(n\Delta t)$ and $N_U(n\Delta t)$ denote the additive noise components, represented by a normally distributed white noise. For technical applications which concern the evaluation of quality of electrical insulation the frequency of test voltage usually is known. Therefore, each analyzed signal contain one originally known parameter (ω_2) and three unknown parameters (A_{2U}, φ_{2U}, $N_U(n\Delta t)$ or A_{2I}, φ_{2I}, $N_I(n\Delta t)$).

Since the noise components of the analyzed signals (1, 2) are considered to be additive, for all cases the value of signal-to-noise ratio (*SNR*) was calculated as follows:

$$SNR = \frac{A^2}{2D}, \quad (3)$$

where: D and A correspondingly denote the variance of the additive noise component and amplitude of the analyzed signal of current or voltage.

B. Formation of System of Equations for Calculation of In-phase and Quadrature Components of the Analyzed Signals

The formation of the overdetermined system of equations, which is solved by using the least squares method in order to determine the unknown in-phase and quadrature components of the analyzed signals, will be illustrated with respect to the determined signal according to (2). The formation of a system of equations is premised on the analysis of properties of the inner product of the expressed in analog form analyzed signal $u(n\Delta t)$ and the auxiliary function $\sin(\omega_1 t)$:

$$S(\omega_1) = \int_0^b \sin(\omega_1 t) A_{2U}\sin(\omega_2 t + \varphi_{2U})dt, \quad (4)$$

where: ω_1 denotes the angular frequency of the function $\sin(\omega_1 t)$, b is the duration of the analyzed signal. In the case of: $\omega_1 \neq \omega_2$, inner product (4) can be expressed as follows [14]:

$$S(\omega_1) = \sum_{i=1}^{5} s_i, \quad (5)$$

where: s_i can be determined according to the following formulas:

$$s_1 = \frac{A_{2U}\sin(\varphi_{2U})\omega_1}{(\omega_1-\omega_2)(\omega_1+\omega_2)}, \quad (6)$$

$$s_2 = -\frac{A_{2U}\omega_1\cos(b\omega_1)\sin(b\omega_2)\cos(\varphi_{2U})}{(\omega_1-\omega_2)(\omega_1+\omega_2)}, \quad (7)$$

$$s_3 = -\frac{A_{2U}\omega_1\cos(b\omega_1)\cos(b\omega_2)\sin(\varphi_{2U})}{(\omega_1-\omega_2)(\omega_1+\omega_2)}, \quad (8)$$

$$s_4 = \frac{A_{2U}\omega_2\sin(b\omega_1)\cos(b\omega_2)\cos(\varphi_{2U})}{(\omega_1-\omega_2)(\omega_1+\omega_2)}, \quad (9)$$

$$s_5 = -\frac{A_{2U}\omega_2\sin(b\omega_1)\sin(b\omega_2)\sin(\varphi_{2U})}{(\omega_1-\omega_2)(\omega_1+\omega_2)}. \quad (10)$$

The existence of undesirable noise components in the analyzed signals tends to degrade the accuracy of any signal processing method intended for the estimation of their unknown parameters (A_{2U}, φ_{2U}). This loss of accuracy results if errors of measurements of the level of tgδ. The affect of noise components of the analyzed signals on the results of numerical calculation of the inner product (4) are presented in the Fig. 1 and Fig. 2.

All calculations have been made by applying the trapezium rule and for the case of sampling interval Δt equal to $5\cdot10^{-4}$ s, $A_2 = 1$, $f_2 = 50$ Hz, $\varphi_{2U} = 22.5^{\circ}$, $b = 0.4$ s. The array of an additive noise component of the analyzed signal was created by applying the built-in function numpy.random.normal, implemented in Python package NumPy.

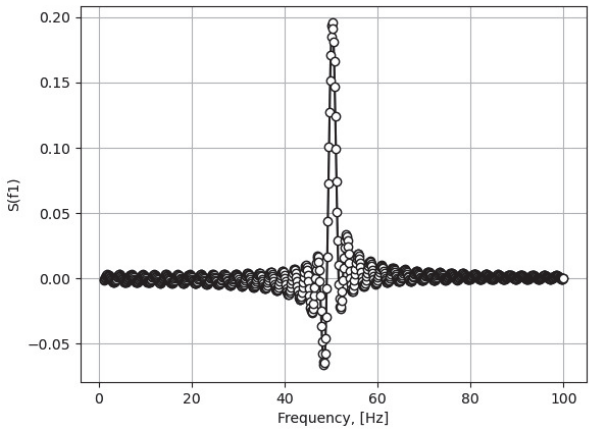


Fig. 1. Results of numerical calculations of the inner product (4) in the frequency domain for the case when the analyzed signal does not contain any noise component.

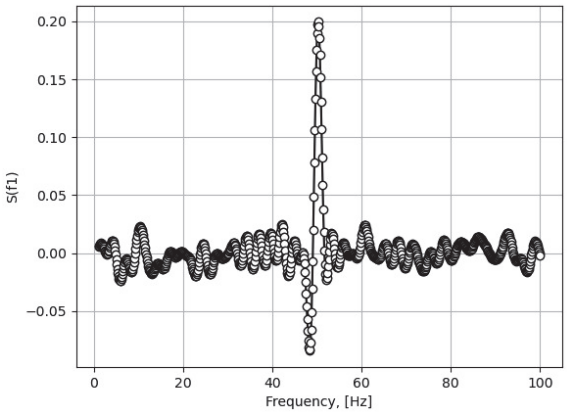


Fig. 2. Results of numerical calculations of the inner product (4) in the frequency domain in the case of *SNR* equal to 0.49.

The expressed according to (5-10) inner product (1) can be used in order to form an overdetermined system of equations. The formation of this system is based on observation that according to the presented model of the analyzed signal (2) the expressed according to (5-10) contains only 2 unknown parameters. These unknown parameters are the in-phase and quadrature components of the analyzed signal (x_1, x_{2u}):

$$x_{1U} = A_{2U}\cos(\varphi_{2U}) \tag{11}$$

$$x_{2U} = A_{2U}\sin(\varphi_{2U}) \tag{12}$$

After the substitution of the determined according to (11, 12) in-phase and quadrature components in (2) and by combining together all components s_i which contain either in-phase or quadrature component, inner product (4) can be expressed as follows:

$$S(\omega_1) = x_{1U}\xi(\omega_1) + x_{2U}\zeta(\omega_1), \tag{13}$$

where: $\xi(\omega_1)$ and $\zeta(\omega_1)$ can be determined according to the following expressions:

$$\xi(\omega_1) = -\frac{\omega_1\cos(b\omega_1)\sin(b\omega_2)}{(\omega_1-\omega_2)(\omega_1+\omega_2)} + \frac{\omega_2\sin(b\omega_1)\cos(b\omega_2)}{(\omega_1-\omega_2)(\omega_1+\omega_2)}, \tag{14}$$

$$\zeta(\omega_1) = -\frac{\omega_1\cos(b\omega_1)\sin(b\omega_2)}{(\omega_1-\omega_2)(\omega_1+\omega_2)} + \frac{\omega_2\sin(b\omega_1)\cos(b\omega_2)}{(\omega_1-\omega_2)(\omega_1+\omega_2)} + \theta, \tag{15}$$

$$\theta = \frac{\omega_1}{(\omega_1-\omega_2)(\omega_1+\omega_2)}. \tag{16}$$

The overdetermined system of equations, which contains the unknown in-phase and quadrature components, is formed by setting various frequencies of the auxiliary function $\sin(\omega_1 t)$, followed by the numerical calculation of the inner product (4). The results of these numerical calculations form the column vector of constant terms, whereas, calculated for various values of ω_1, functions $\xi(\omega_1)$ and $\zeta(\omega_1)$ determine the coefficients of the system. The solution of the overdetermined system is obtained by the least squares method, which allows to mitigate the illustrated in the Fig. 2 undesirable impact of noise components on the results of numerical calculation of the inner product (4), which form the column vector of constant terms.

Under the assumption that system of equations was formed for the case of three various frequencies (ω_{11}, ω_{12}, ω_{13}) of used in (4) function $\sin(\omega_1 t)$ the formation of this system in matrix form can be illustrated as follows:

$$\mathbf{A}\mathbf{x} = \mathbf{b}, \tag{17}$$

where: $\mathbf{b} = [S(\omega_{11}), S(\omega_{12}), S(\omega_{13})]^{\mathrm{T}}$, $\mathbf{x} = [x_{1U}, x_{2U}]^{\mathrm{T}}$ and matrix $\mathbf{A}$ is determined as follows:

$$\mathbf{A} = \begin{pmatrix} \xi(\omega_{11}) & \zeta(\omega_{11}) \\ \xi(\omega_{12}) & \zeta(\omega_{12}) \\ \xi(\omega_{13}) & \zeta(\omega_{13}) \end{pmatrix}. \tag{18}$$

The solution of system (17) allows to determine the unknown in-phase and quadrature components of the analyzed signal and to determine its initial phase:

$$\varphi_{2U} = \arctan\left(\frac{x_{2U}}{x_{1U}}\right). \tag{19}$$

The applying of similar computational procedures with respect to the waveform of current allows to determine its initial phase:

$$\varphi_{2I} = \arctan\left(\frac{x_{2I}}{x_{1I}}\right). \tag{20}$$

where: x_{2I} and x_{1I} correspondingly denote the quadrature and in-phase components of the analyzed signal of current. Estimation of the initial phases of current and voltage according to (19, 20) allows to calculate the value of phase shift between the analyzed signals:

$$\Delta\varphi = \varphi_{2I} - \varphi_{2U}. \tag{21}$$

As the value of dielectric loss angle complements the value of phase shift between the flowing through dielectric current and test voltage up to 90º, the value of dielectric dissipation factor can be calculated as follows:

$$tg(\delta) = tg(90^{\circ} - \Delta\varphi). \tag{22}$$

Consequently, the evaluation of the initial phases φ_{2I} and φ_{2U} allows to calculate the unknown phase difference between the signals of current and voltage and determine the level of tgδ.

III. RESULTS OF SIMULATIONS

The presented in the Fig. 3 and Fig. 4 results of simulations have been carried out for the following parameters of the analyzed signals: $\Delta t = 5\cdot10^{-4}$ s, $A_2 = 1$, $f_2 = 50$ Hz, $\varphi_{2U} = 0^{\circ}$, $\varphi_{2I} = 89.97^{\circ}$, $b = 0.4$ s.

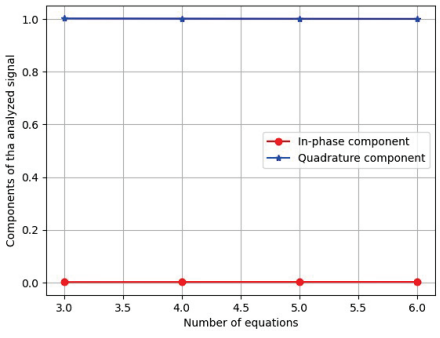


Fig. 3. Results of the evaluation of in-phase and quadrature . components of current in the case of *SNR* = 49.12

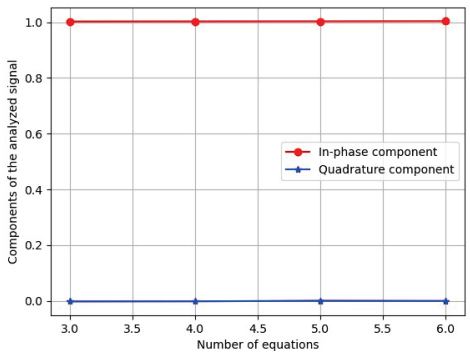


Fig. 4. Results of the evaluation of in-phase and quadrature components of voltage in the case of $SNR = 49.12$

The values of the dielectric dissipation factor, calculated according to the results presented in Fig. 3 and Fig. 4, are presented in Fig. 5.

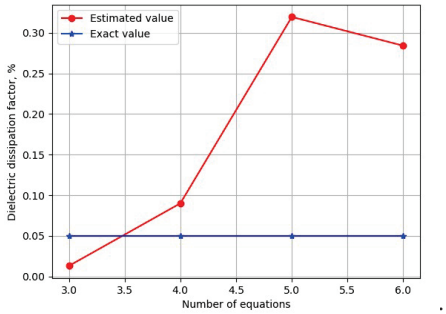


Fig. 5. Estimated values of dielectric dissipation factor.

The results presented in Fig. 5 exhibit higher accuracy for a low number of equations used in (17). For the presented results, the existence of bias is determined by the presence of the additive noise components in the analyzed signals. Applying the correlation method [13] for processing the analyzed signals with the same parameters yields the value of dielectric dissipation factor tgδ = 2.3%. According to the results presented in Fig. 5, the error of this estimation exceeds the corresponding errors of the developed method even when the overdetermined system contains 5 equations.

CONCLUSIONS

The presented approach for the calculation of dielectric dissipation factor is based on several calculations of inner products between the analyzed sine signals of current and voltage and auxiliary sine functions with different frequencies, followed by the formation of the overdetermined system of equations. The obtained by the least squares method solution of such system allows to determine the in-phase and quadrature components of the analyzed signals. Further calculations include the evaluation of the initial phase of the analyzed signals, their phase shift, dielectric loss angle, and finally the dielectric dissipation factor. As the described approach allows for determining the initial phases of the analyzed sine signals with noise components, the scope of the developed method is not limited to applications that concern the evaluation of the technical condition of electrical insulation, but can be used for other engineering problems, which require the measurement of phase shift between sine signals. The other area of application of the presented method is the evaluation of in-phase and quadrature components of the analyzed sine signals. The comparative analysis of the least squares-based and correlation methods has shown that the error of the correlation method exceeds the error of estimation obtained by applying the developed method.

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Impact of Stray Capacitance on Performance of Auto-Balancing Bridge Used as a Capacitance-to-Voltage Converter

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***Abstract*—This paper is dedicated to the development of a model which allows to take into consideration the undesirable impact of parasitic capacitance of tested object on the output voltage of the auto-balancing bridge. The presented analysis is oriented toward technical applications for which the values of parasitic capacitances of the tested object remain smaller than those of its parasitic capacitances, such as the direct measurements of partial capacitances in three-core belted cables. It was shown that parasitic capacitance of the tested object leads to a decrease in voltage on the output of the auto-balancing bridge. According to the results of the presented analysis, the undesirable impact of parasitic capacitance can be mitigated by increasing the open-loop gain of the operational amplifier and by decreasing the impedance of the feedback loop.**

***Keywords—capacitance measurement, power cable, open-loop gain, impedance measurement, operational amplifier*.**

I. INTRODUCTION

The problem of accurate measurement of impedance, electrical capacitance and dielectric dissipation factor (tgδ) remains relevant for numerous technical applications. Among these applications are: monitoring of the human body and bioimpedance analysis [1, 2], evaluation of the quality of electrical insulation in underground power cables [3], and capacitance tomography [4].

In many practical cases, the accuracy of measurements of impedance, electrical capacitance, and dielectric dissipation factor can be substantially degraded due to the undesirable capacitive coupling of the tested object, which can be reflected by the inclusion of two additional capacitors in its equivalent scheme. The mentioned decline of accuracy urges the development of methods that would allow the removal of the effect of parasitic capacitances. This removal can be attained by applying a well-known guarding technique, which implies the three-terminal connection of the tested object. In this case, two terminals are used to apply the test voltage, whereas the third terminal is used to equalize the potential of one electrode with the ground potential and measure the current flowing through the tested object. The auto-balancing bridge is often used as a capacitance-to-voltage converter and represents an inverting amplifier with a connected in parallel capacitor and resistor in its feedback loop. When complemented by the vector voltmeter, such capacitance-to-voltage converters can be efficiently used for impedance measurements [1]. The capacitance-to-voltage converters, based on an auto-balancing bridge, are well-known for their ability to suppress the undesirable affect of stray capacitances on the technical performance of capacitance measuring systems [5]. In practice, such suppression is attained due to the specific configuration of the measuring circuit, which maintains a negligible voltage drop across the inverting and non-inverting inputs of the operational amplifier. This allows to substantially reduce the voltage drop across the electrodes of one parasitic capacitance and decrease the flowing through this capacitance current. Such a reduction of current, at first approximation, allows us to consider the operation of such a type of capacitance-to-voltage converters similarly to the operation of the idealized inverting amplifiers, which have the same value of current flowing through the tested capacitor and their feedback loop. Technical performance of capacitance-to-voltage converters based on auto-balancing bridges has been described in detail under the assumption of negligible stray capacitance of the tested object [6].

In most cases, parasitic capacitance may exist because of various factors, such as shielding of coaxial cables used in the developed devices, and remains small compared to the measured capacitance. However, in some practical problems, parasitic capacitance values may substantially exceed the capacitance of the tested object. One of them is the problem of estimating the level of deterioration of electrical insulation in three-core belted power cables with paper impregnated insulation. This cable type is widely used in nuclear power plants and power transmission systems in Ukraine. Such cables contain six layers of insulation, three of which separate their conductors from each other and the other three separate their conductors from the sheath. Some approaches, used to evaluate the technical condition of these cables, require measurements of individual values of partial capacitances and corresponding values of dielectric dissipation factor of these layers of insulation. This problem can be solved by applying the measuring techniques based on direct or aggregate

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measurements [7]. The aggregate measurements-based techniques require aggregate values of capacitance and dielectric dissipation factor, measured for various ways of connection of metal elements of power cable. In this case individual dielectric properties of insulating layers are estimated either by the direct solution of system of linear algebraic equations or by minimizing the r.m.s. error of solving the overdetermined system of equations by applying the least squares method. Methods based of aggregate measurements are labor-consuming and in some cases may lead to substantial loss of accuracy due to the undesirable impact of errors in measured aggregate values of capacitance and dielectric dissipation factor on calculated individual dielectric properties of insulating layers. The direct measurements of dielectric properties are less time-consuming. However, this approach requires the application of methods that allow the removal of the capacitive coupling of the tested insulation layer. Since the applying of capacitance-to-voltage converters based on auto-balancing bridges allows to remove the affect of parasitic capacitances, such converters may be efficiently used for the purposes of testing the quality of insulation in three-core power cables. The application of this approach, however, requires the analysis of the ability of auto-balancing bridges to suppress the increased values of parasitic capacitances, which may exceed the capacitance of the tested object. This paper is dedicated to the analysis of the described problem.

II. MODEL FOR THE EVALUATION OF IMPACT OF PARASITIC CAPACITANCE ON OUTPUT VOLTAGE OF THE AUTO-BALANCED BRIDGE

A. Equivalent Scheme of a Three-core Power Cable in the Case of the Direct Measurements of Dielectric Properties of Insulating Layers Between the Conductors

The equivalent scheme of a three-core belted power cable is presented in the Fig. 1 and reflects the existence of a capacitive coupling between its metal elements and dielectric power losses in corresponding layers of insulation.

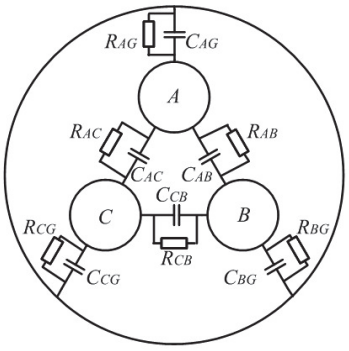


Fig. 1. Equivalent scheme of a three-core power cable in common metal sheath: C_{AB}, C_{CB}, C_{AC} – partial capacitances between the conductors of power cable; C_{AG}, C_{CG}, C_{AG} – partial capacitances between the conductors and the sheath; R_{AB}, R_{CB}, R_{AC}, R_{AG}, R_{CG}, R_{AG} – shunt resistors due to the dielectric power losses caused by polarization, conduction and partial discharges.

The direct measurements of partial capacitance between the conductors *A* and *B*, as well as the corresponding shunt resistance R_{AB}, can be made by applying the presented in the Fig. 2 connection of metal elements.

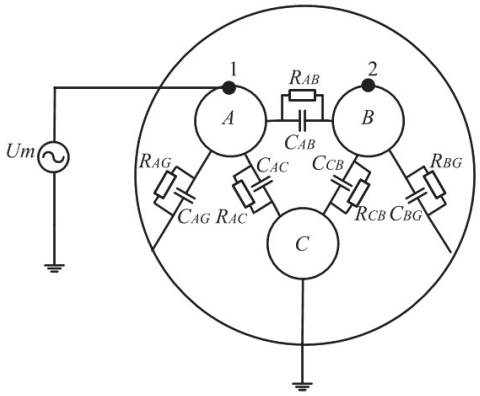


Fig. 2. Connection of metal elements in the case of the direct measurements of electrical capacitance and dielectric dissipation factor of insulation layer between the conductors *A* and *B*.

As it can be concluded from the presented in the Fig. 2 scheme, the direct measurements of dielectric properties of insulating layers between two conductors requires grounding of the third conductor. For this case the values of C_{CB}, C_{AC}, C_{BG}, C_{AG}, R_{CB}, R_{AC}, R_{BG}, R_{AG} determine the parasitic parameters of tested layer of insulation between the conductors *A* and *B*. The same scheme, complemented by the capacitance-to-voltage converter based on auto-balancing bridge, is presented in the Fig. 3.

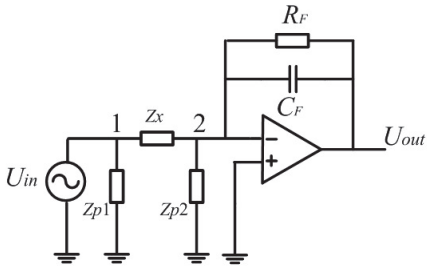


Fig. 3. Direct measurements of dielectric properties of insulating layer between the conductors *A* and *B* by applying the auto-balancing bridge: R_F, C_F – feedback resistor and capacitor, Z_X – unknown impedance of tested layer of insulation, Z_{P1}, Z_{P2} – impedance due to the parasitic parameters of tested layers of insulation.

In the case of the presented in the Fig. 2 connection of metal elements, the presented in the Fig. 3 parasitic parameters Z_{P1}, Z_{P2} as well as the unknown impedance Z_X can be expressed as follows:

$$Z_{P1} = \frac{R_{P1}}{1 + j\omega C_{P1} R_{P1}}, \quad (1)$$

$$Z_{P2} = \frac{R_{P2}}{1 + j\omega C_{P2} R_{P2}}, \quad (2)$$

$$Z_X = \frac{R_X}{1 + j\omega C_X R_X}, \quad (3)$$

where: ω denotes the angular frequency of the test voltage and the values of C_{P1}, C_{P2}, R_{P1}, R_{P2} can be calculated as follows:

$$C_{P1} = C_{AG} + C_{AC}, \quad (4)$$

$$C_{P2} = C_{BC} + C_{BG}, \quad (5)$$

$$R_{P1} = \frac{R_{AG}R_{AC}}{R_{AG} + R_{AC}}, \quad (6)$$

$$R_{P2} = \frac{R_{CB}R_{BG}}{R_{CB} + R_{BG}}, \quad (7)$$

$$C_X = C_{AB}, \quad (8)$$

$$R_X = R_{AB}. \quad (9)$$

The presented in this paper analysis will be focused on the analysis of impact of parasitic impedance Z_{P2} on technical performance of auto-balancing bridge used as a capacitance-to-voltage converter. This analysis will be carried out under the assumption of a complete removal of impact of parasitic impedance Z_{P1}.

B. *Model for the Analysis of Impact of Parasitic Capacitance on Technical Performance of Auto-Balancing Bridge*

The analysis of the presented in the Fig. 3 scheme can be carried out based on various assumptions. In the case of the idealized electrical capacitances C_{P1}, C_{P2}, C_X, with negligible power losses ($R_{p1} = R_{p2} = R_x = \infty$), infinite input resistance of operational amplifier, its negligible output resistance, zero bias current, zero input offset voltage and a complete suppression of parasitic capacitances the presented in the Fig. 3 scheme can be considered as the idealized inverting amplifier. In this case its output voltage can be calculated as:

$$U_{OUT}(j\omega) = -\frac{Z_F}{Z_X}U_{IN}(j\omega), \quad (10)$$

where: Z_F can be calculated as:

$$Z_F = \frac{R_F}{1 + j\omega C_F R_F}, \quad (11)$$

By substituting the determined according to (3) value of Z_X in (10) the output voltage of the auto-balancing bridge can be expressed as [6]:

$$U_{OUT}(j\omega) = -\frac{j\omega C_X R_F}{j\omega C_F R_F + 1}U_{IN}(j\omega). \quad (12)$$

Assuming that the parameters of the feedback loop C_F, R_F fit the condition: $|j\omega C_F R_F| \gg 1$, (12) can be simplified as follows [6]:

$$U_{OUT}(j\omega) = -\frac{C_X}{C_F}U_{IN}(j\omega). \quad (13)$$

As the value of feedback capacitor C_F usually is known and can be regulated within the necessary range of values, (12) can be directly used for the development of capacitance measuring circuits. The demodulation of the output signal of the auto-balancing bridge by applying the in-phase and quadrature-phase reference signals, followed by the calculation of its amplitude and initial phase, allows to build the capacitance measuring systems by applying (12).

In the case of $R_x \neq \infty$, complete suppression of parasitic impedances Z_{P1}, Z_{P2} and similar assumptions made in order to derive (12), the output voltage of the auto-balancing bridge can be expressed as:

$$U_{OUT}(i\omega) = -\frac{Z_F}{Z_X} = -\frac{R_F + j\omega C_X R_X R_F}{R_X + j\omega C_F R_F R_X}U_{IN}(j\omega). \quad (14)$$

Equations (12, 14) are derived under the assumption of negligible voltage drop across the inverting and non-inverting terminals of operational amplifier. This assumption allows to completely remove the effect of parasitic capacitance C_{p2}, The existence of some voltage across the inputs of operational amplifier, however, does not allow to suppress the affect of parasitic impedance due to the presence of some leakage current, flowing through impedance Z_{P2}. This leakage current causes some distinction between the value of current, flowing through the tested object and the value of current in the feedback loop of the auto-balancing bridge. Mentioned distinction of currents affects the transfer function of the auto-balancing bridge and can be taken into account as:

$$\frac{U_{IN} - \varphi_2}{Z_X} = -\frac{U_{OUT} - \varphi_2}{Z_F} + \frac{\varphi_2}{Z_{P2}}, \quad (15)$$

where: φ_2 is the potential of the node 2 in the Fig3, which can be expressed as:

$$\varphi_2 = -\frac{U_{OUT}}{\alpha}, \quad (16)$$

where: α is the open-loop gain of the operational amplifier. By substituting the defined according to (16) potential of the node 2 in (15) it can be expressed in the following form:

$$\frac{U_{IN}}{Z_X} = -\frac{U_{OUT}}{\alpha Z_X} - \frac{U_{OUT}}{Z_F} - \frac{U_{OUT}}{\alpha Z_F} - \frac{U_{OUT}}{\alpha Z_{P2}}. \quad (17)$$

After the appropriate combination of terms in (17) the output voltage of the auto-balanced bridge can be expressed as follows

$$U_{OUT}(j\omega) = -\frac{\alpha Z_F Z_{P2}}{\alpha Z_{P2}Z_X + Z_F Z_{P2} + Z_F Z_X + Z_{P2}Z_X}U_{IN}(j\omega). \quad (18)$$

In the case of infinite impedance between the node 2 and the ground the value of current flowing through the tested object will be equal to the value of current in the feedback loop of the auto-balancing bridge. In this case (18) is simplified and the value of the output voltage can be expressed by applying a well-known from the theory of the non-idealized inverting amplifier formula:

$$U_{OUT}(j\omega) = -\frac{\alpha Z_F}{\alpha Z_X + Z_F + Z_X}U_{IN}(j\omega). \quad (19)$$

Under the assumption of the idealized electrical capacitances C_{P2} and C_X with infinite shunt resistors and in the case when the auto-balanced bridge does not have any resistor R_F in its feedback loop (18, 19) can be expressed as:

$$U_{OUT}(j\omega) = -\frac{\alpha C_X}{\alpha C_F + C_F + C_X + C_{P2}}U_{IN}(j\omega), \quad (20)$$

$$U_{OUT}(j\omega) = -\frac{\alpha C_X}{\alpha C_F + C_F + C_X}U_{IN}(j\omega). \quad (21)$$

Consequently, the affect of parasitic impedance Z_{P2} on the output voltage of auto-balancing bridge due to the existence of voltage drop across the inputs of operational amplifier can be

taken into account similarly to the effect of the non-infinite input resistance of operational amplifier.

The results of calculations, made in order to illustrate the effect of parasitic capacitance on the output voltage of the auto-balancing bridge, are presented in the Fig. 4. All calculations have been made for the case of frequency of test voltage equal to 10 kHz, α = 316, C_{p2} = 500 pF, C_x = 100 pF.

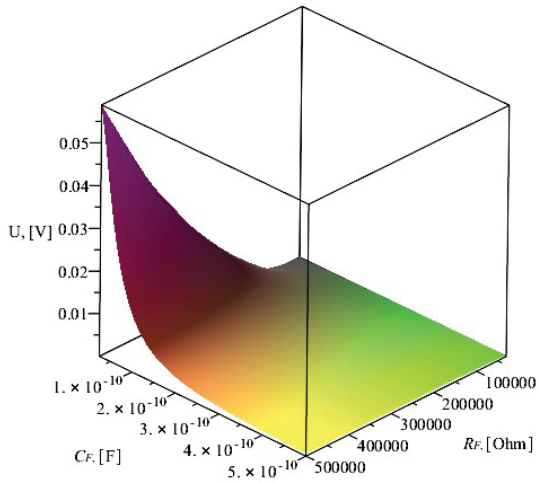


Fig. 4. Difference between the values of the output voltage calculated according to models which take into account the existence of parasitic capacitance (18) and do not consider the parasitic capacitance (19).

According to the presented in the Fig. 4 results of calculations, the increasing of the feedback loop impedance leads to the increasing of impact of parasitic capacitance on output voltage of the auto-balancing bridge. Such increasing of impact of parasitic capacitance takes place due to the increasing of difference between the currents flowing through the tested capacitor and through the feedback loop with the increasing of its impedance.

III. RESULTS OF SIMULATIONS

An illustration of the impact of parasitic capacitance of the tested object was also done by simulations of corresponding electrical circuits in Multisim. Since the parasitic impedance Zp_1 was not taken into account in the considered model of the auto-balanced bridge, it has been excluded from simulations. All simulations have been carried out for the case of C_{p2} = 500 pF, C_x = 100 pF, C_F = 50 pF, R_F = 400 kOhm. Other parameters of the analyzed circuit and the obtained result of simulations are presented in Table 1.

TABLE I. SIMULATION RESULTS

Frequency [kHz]	Open-loop gain	U_{OUT} according to (18), [V]	U_{OUT} according to (19),[V]	U_{OUT} simulated, [V]
0.1	31600	0.025	0.025	0.025
1	3160	0.249	0.249	0.249
10	316	1.524	1.554	1.532
100	31.6	1.4146	1.821	1.415

According to the presented in the Table 1 results, the discrepancy between formulas (18, 19) becomes more significant for high frequencies of test voltage and low values of the open-loop gain. The results of simulations confirm that the effect of parasitic capacitance C_{p2} can be taken into consideration by applying the derived formula (18).

CONCLUSIONS

The existence of parasitic capacitance of the tested object leads to a decrease in voltage on the output of the auto-balancing bridge. Mitigation of the impact of parasitic capacitance can be achieved by increasing the open-loop gain of the operational amplifier. Since the increase in impedance of parasitic capacitance allows for a reduction in its effect, a similar result can also be attained by decreasing the test voltage frequency, which also increases the open-loop gain of the operational amplifier. In the case of low frequencies of test voltage and small values of impedance of the feedback loop the affect of parasitic capacitance may become negligible.

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Properties and Microstructure of Low-temperature Electrical Porcelain

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***Abstract* — Due to the constant increase in energy prices, research aimed at increasing the energy efficiency of energy-intensive industrial processes is becoming increasingly important. In the production of porcelain insulators, such a process is the firing of products with a molding temperature exceeding 1300 °C. To reduce energy consumption during the firing of electrical insulating porcelain, it is necessary to use technological reserves of energy saving, which include both the development of low-temperature ceramic masses and glazes. This paper presents our experience of such developments, which allowed us to reduce the firing temperature of silicate porcelain insulators by 1200 °C while maintaining high mechanical and electrophysical characteristics, which allows them to be used for the manufacture of high-voltage insulators with a nominal voltage of up to 10 kV.**

Keywords* — *low-temperature electrical porcelain, electrical strength, electrical resistance, sintering, phase formation, microstructure.

I. INTRODUCTION

In contrast to alternative dielectric materials (polymers and glass), electrical porcelain stands out due to its high performance and stability in electrical, thermal, and electromechanical properties, as well as its resistance to natural environmental factors (high humidity, temperature fluctuations, ultraviolet radiation, and solar radiation, as well as aggressive chemical emissions from industrial enterprises).

Electrical porcelain is widely used in high-voltage engineering for the production of insulators for power transmission lines, electrical transformers, and other high-voltage equipment. It is used to make high-voltage insulators (pin-type, disc-type, suspension, support, through-type, etc.), rollers, and parts for circuit breakers and fuses [1, 2]. Insulators must meet the following general requirements:

- Sufficient electrical strength, which is determined by the electric field intensity at which the material of the insulator loses its dielectric properties;
- Sufficient mechanical strength, which allows it to withstand dynamic forces arising between separate conductive circuits;
- Stability of properties under operating conditions, including temperature changes;
- Resistance to the impact of electrical discharges on the surface of the products.

The composition of classical electrical porcelain consists of 50% clay materials (usually kaolin and fireclay), which serve as the plastic component and are the main phase-forming elements. Feldspars, making up about 25%, ensure the formation of a melt during firing, which intensifies sintering. Quartz sand, comprising 25% of the mass, acts as a filler, minimizing shrinkage and high-temperature deformation during firing. To improve the dielectric properties of the products, alumina is sometimes added to the mass by partially replacing quartz; however, this solution requires higher firing temperatures.

The most commonly used types of porcelain for manufacturing high-voltage insulators are either silicate porcelain or high-alumina porcelain, which correspond to the C-110 and C-120 subgroups according to the IEC672-3 classification standard. Silicate porcelain is characterized by an increased content of quartz, which contributes to its high strength.

The properties of porcelain insulators play a crucial role in ensuring electrical safety and reliable operation in various electrical applications [3]. The most important characteristics to consider when creating low-temperature compositions for porcelain insulator production are dielectric and mechanical properties, as well as dielectric strength. These properties depend on the phase composition of the porcelain, specifically the ratio of mullite and glassy phases that form during firing [1, 5].

The mechanism of forming classical electrical porcelain has been extensively studied by authors [6 – 9]. It is known that the role of the melt is to ensure the maximum sintering level, which is characterized by the water absorption index, and for electrical porcelain, it should be at the level of W=0%. On the other hand, the mullite phase ensures the necessary dielectric characteristics and high strength. It is important to consider the form of crystalline formations in the mullite phase, which primarily depends on the firing temperature. Thus, small, short-columnar crystals of mullite with an average size of 1-3 μm ("primary" mullite) form at relatively low temperatures (960-1000 °C) from the products of clay mineral breakdown. In turn, needle-like ("secondary") mullite crystals begin to form in areas of feldspar at high temperatures (~ 1400 °C) through recrystallization via the melt due to the gradual dissolution of "primary" mullite and the saturation of the feldspar melt with phase-forming oxides (Al_2O_3 and SiO_2). The final microstructure of the porcelain consists of mullite crystals, a glassy phase, and a small amount of quartz, surrounded by fine cracks that define the properties of porcelain [2].

The most influential factors, each of which potentially affects the phase composition, structure, and properties of classical electrical porcelain products [1], are as follows:

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- the technological mixture composition (an optimal ratio of the components: clay, feldspar, and quartz);
- the particle size of feldspar and quartz;
- the forming method used, which may differ depending on the shaping technique (slip casting, pressure casting, or plastic forming followed by mechanical processing of the blank).

To reduce the energy consumption of the porcelain insulator production process, an important factor, in our opinion, is the use of additives capable of intensifying the sintering and phase formation processes while enabling the formation of porcelain at significantly lower firing temperatures.

The aim of this work is to investigate the properties and structural-phase characteristics of high-voltage porcelain products obtained at reduced firing temperatures.

The experiments focus on two main aspects: the study of how the composition of the porcelain body containing sintering and phase formation intensifiers such as dolomite and titanium dioxide affects the physicochemical, electrical, and mechanical properties of the products, as well as an investigation of the microstructure of the resulting low-temperature porcelain for electrical applications.

II. MATERIALS AND RESEARCH METHODS

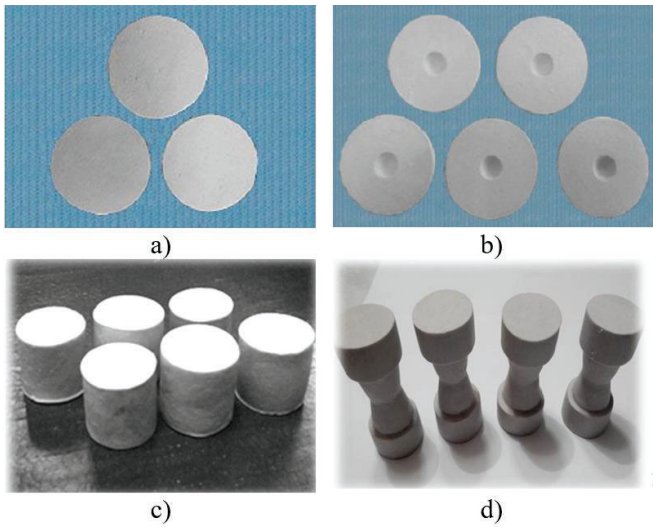
To obtain low-temperature electroporcelain, natural raw materials were used: kaolinite-hydromica clay of the Extra brand of JSC "Vesko" (Donetsk region), quartz sand from Novoselivske deposit (Kharkiv region), dolomite from Negrebivske deposit (Zhytomyr region), kaolin from Pyshchikivske deposit (Kyiv region), feldspar concentrate as pegmatites enrichment product of the Lozuvatske deposit (Kirovograd region). The chemical composition of the raw materials is given in Table 1.

TABLE 1. THE RAW MATERIALS CHEMICAL COMPOSITION

Raw materials	Component content, wt. %								
	SiO_2	Al_2O_3	Fe_2O_3	TiO_2	CaO	MgO	K_2O	Na_2O	LOI
Clay Vesko-Extra	53.94	30.56	0.80	1.44	0.45	0.54	1.89	0.49	9.89
Pyshchykivsky kaolin	48.53	35.60	0.71	0.44	0.42	0.18	0.89	0.22	13.01
Lozuvatsky feldspar concentrate	68.09	18.15	0.1	0.04	0.65	0.21	8.27	3.96	0.53
Negrebivsky dolomite	2.00	1.00	0.50	–	28.00	22.00	–	–	46.50
Novoselivsky quartz sand	99.12	0.39	0.04	–	0.20	0.04	–	–	0.21

To determine the fluxing capacity of the Lozuvatsky feldspar material, the author's method of graph-analytical calculations was used [10]. This method allows, based on data on the rock chemical composition, to preliminarily estimate the dynamics of melt formation, its quantity, composition, and properties that determine the sintering process at a given product's firing temperature. The glass phase electrophysical properties (dielectric permittivity, dielectric loss tangent) formed during melt cooling were also calculated [11, 12].

In accordance with the requirements of the current standard for silicate porcelain, the following properties were determined: water absorption, apparent porosity, apparent density, bending strength, compressive strength, tensile strength, thermal conductivity, specific volumetric electrical resistivity, electrical strength (breakdown voltage), dielectric loss tangent, and dielectric permittivity. To measure electrical resistance, breakdown voltage, and tensile strength, samples were manufactured that met the requirements of the measuring equipment (Fig. 1).



a) dielectric characteristics (ρ_v ε, tgδ); b) electrical strength, c) compressive strength, d) tensile strength

Fig. 1. Samples for determining the electrical porcelain properties

The compressive strength of the samples was measured using a GMS-50 press with a maximum load of 500 kN. The bending strength was determined on samples in the form of rectangular parallelepipeds measuring 35×120×10 mm on a laboratory device MP-100. The test results were taken as the arithmetic mean of the properties for five samples for each of the experimental masses.

Electrical strength was tested on samples of standard size according to [13]. Specific volumetric resistivity was determined using the Zahner Elektrik IM 6e instrument after applying thin gold films to the samples, which served as electrodes. The results of the tests were accepted as the arithmetic mean values of the properties for five samples from each experimental batch.

The mineral composition of Negrebivsky dolomite and the phase composition of fired porcelain were studied on a DRON-UM1 X-ray diffractometer in copper Cu-Kα radiation using a Ni selective absorption filter. Qualitative phase analysis of the obtained materials was performed using the ICDD PDF-2 crystallographic compounds database. Quantitative phase analysis was determined by the Rietveld method using MAUD software.

Microstructural features of the samples were studied using a REMMA-101Az scanning electron microscope with a micro analyzer. The study was performed on samples after the glassy phase removal with a 5 % hydrofluoric acid solution. The study of the cleaved surface of the samples was performed on a VK 9700 KEYENCE laser-scanning microscope, with subsequent processing of the results using the VK Analyzer software.

III. Results and Discussion

A. The composition and properties research of the alternative complex flux components

According to the results of the research, it was established that in the temperature range of 1150-1200°C, the Lozuvatsky feldspar concentrate will form the melt from 76 to 80%. Its properties will probably be characterized by moderate viscosity ($10^{4.62\text{-}4.63}$ Pa·s), low surface tension (0.276–0.278 N/m), and will ensure active dissolution of the kaolinite destruction products (K_{akt} = 0.157–0.158). At the same time, the glass phase formed during products cooling will have satisfactory dielectric properties (ε = 7.91; tg δ = 0.011).

As shown earlier, an effective method for intensifying porcelain sintering is mass modification by introducing additives that change the structure and increase the melt mobility [14]. For alkali-containing aluminosilicate melts of quartz-feldspar raw materials, the role of such components can be played by pure natural minerals, whose action mechanism consists in melt depolymerization upon introduction of alkaline earth metal oxides with low binding energy with oxygen. Crushing complex anionic complexes of the melt reduces the melt viscosity, increases its mobility, which promotes ceramic mass sintering at a reduced temperature. In this work, dolomite from the Negrebivske deposit is used as a sintering activator.

This choice is explained by the fact that the effect of $CaMg(CO_3)_2$ as a sintering intensifier is similar to that of $CaCO_3$, but the high-temperature deformation tendency of porcelain samples is expressed to a lesser extent. This is explained by the fact that the dolomite decomposition products during mass firing are both CaO and MgO. While the latter "softens" the calcium oxide effect, which sharply reduces the feldspar melt viscosity in a narrow temperature range [15]. At the same time, when using $MgCO_3$ as a sintering intensifier, high shrinkage of porcelain products (~ 25%) is observed, which is unacceptable [16].

In the sample using X-ray diffraction (Fig. 2), two phases were detected: the main phase (85.7 wt.%) dolomite $CaMg(CO_3)_2$ (hexagonal system, space group No. 148) and calcite $CaCO_3$ (hexagonal system, space group No. 167) in the amount of 14.3% wt.

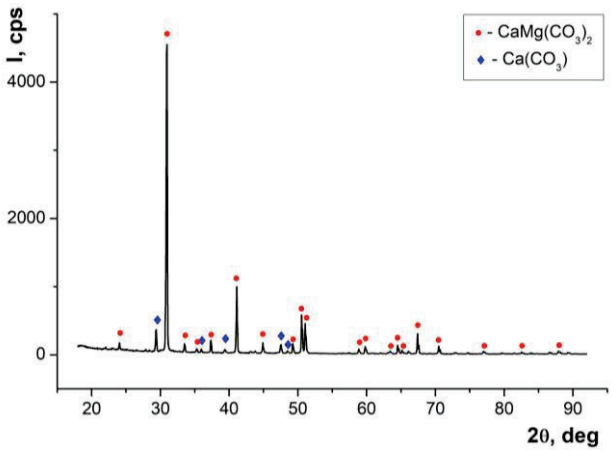


Fig. 2. X-ray diffraction pattern of dolomite from Negrebivske deposit

The main phase in the sample is dolomite $CaMg(CO_3)_2$ with a mass fraction of, the lattice parameters are:. For the identified phases, the crystal lattice parameters were determined, which are: for dolomite, a = 4.812Å; c = 16.022Å, for calcite, a = 4.993Å; c = 17.056Å. The diffraction reflexes of both phases have a small half-width, which indicates the coarse-crystalline state of the present phases.

B. Low-temperature electroporcelain production technology and properties

The development of low-temperature porcelain masses was carried out based on the analysis of the phase diagram of the $Na_2O–K_2O–Al_2O_3–SiO_2$ system. The most promising region of the system is the central part of its internal tetrahedron $NAS_6–KAS_6–A_3S_3–S$, which is characterized by a comparatively large elementary volume (Vi = 14.3 ‰) and includes low-temperature eutectics (binary – 4, three-component – 3 and four-component – 1). It gives reason to assume that during the porcelain heat treatment at reduced temperatures (1150–1200 °C) a sufficient liquid phase amount will be formed, necessary to ensure the maximum sintering level. Analysis of the composition's position within the selected region showed that in order to form a sufficient amount of melt for sintering porcelain with a high mullite content, it is necessary to increase the temperature significantly. Therefore, based on the task of implementing the principles of energy saving in the production of electroporcelain, it is necessary to identify additives that can simultaneously intensify the processes of material sintering and mullitization.

As the working hypothesis was the possibility of targeted regulation of the feldspar melt properties, which would decrease its formation temperature, reduce ionic conductivity when using dolomite additive and, thus, improve the dielectric properties of porcelain obtained at a lower products firing temperature. The theoretically substantiated working hypothesis was tested experimentally.

The research was conducted in the component concentrations area, wt.%: 35–55 kaolinite-hydromica clay, 10–25 kaolin, 25–35 feldspar concentrate, 13–17 quartz sand, 2–4 dolomite. Experimentally established a rational ratio of raw materials – clay : kaolin : feldspar concentrate = 40 : 22 : 28 with a fixed quartz sand content of 15 wt.% and 2.0–2.5 wt.% dolomite over 100% on dry matter. Subsequently, the mass compositions were developed, which allowed the high-voltage porcelain to be obtained at a firing temperature of 1200 °C (Table 2). As can be seen from the presented data, all experimental samples are characterized by the maximum level of sintering (W=0%) and their electrophysical and mechanical properties meet the requirements for silicate porcelain (subgroup 110) according to GOST 20419-83.

It is noteworthy that the porcelain obtained using the LTP-1 mass has relatively low strength indicators, while the complex of dielectric characteristics practically does not differ from others. This can be explained by the increased amount of dolomite (4 wt. %), an excessive amount of which causes "loosening" of the material. microstructure. Therefore, it is recommended to introduce dolomite in an amount of 2.0–2.5 wt. % (over 100% on a dry matter basis).

TABLE 2. ELECTROPORCELAIN SAMPLES PROPERTIES

Material characteristics	Sample code				GOST 20419 IEC 672-3 (group 110)
	LTP-1	LTP-2	LTP-3	LTP-3	
Electrical strength* E_{br}, kW/mm	26	27	29	30	≥ 25
Volume resistivity** $\rho_V \cdot 10^{-12}$, Ohm·m	3.0	3.3	4.0	4.2	10.0
Relative permittivity* ε	3.37	3.06	2.95	2.70	6.0–7.0
Dielectric loss tangent* $tg\delta \cdot 10^3$	12.3	12.2	12.0	11.5	≤ 25
Compressive strength σ_{com}, MPa	125	152	163	170	–
Bending strength σ_{ben} MPa	63	68	71	74	≥ 60
Tensile strength σ_{ten}, MPa	30	31	33	34	≥ 30
Water absorption W, %	0	0	0	0	0
Apparent density ρ, kg/m^3	2820	2835	2850	2870	≥ 2300
Coefficient of linear expansion, $\alpha \cdot 10^6$ 1/°C	5.05	4.97	4.94	4.90	3.0–6.0
Dolomite content in mass, wt.% (over 100 % on dry matter)	4.0	3.0	2.5	2.0	–
*at a current frequency f = 50 Hz; ** at 20 °C					

When the dolomite is introduced into the masses in the recommended amount, the porcelain electrical strength increases significantly (from 27 to 30 kV/mm), as well as the mechanical properties indicators. However, the dielectric characteristics (ε, tgδ) change slightly. However, it should also be emphasized that the dielectric permittivity value is much better than its indicator, which is regulated by the aforementioned standard, which will positively affect the quality characteristics of porcelain insulators manufactured using the developed masses.

The study of the resistance dependence of the samples LTP-3, LTP-4 on the current frequency showed that for frequencies up to 10 Hz, low-temperature porcelain is characterized by stably high resistance values (Fig. 3).

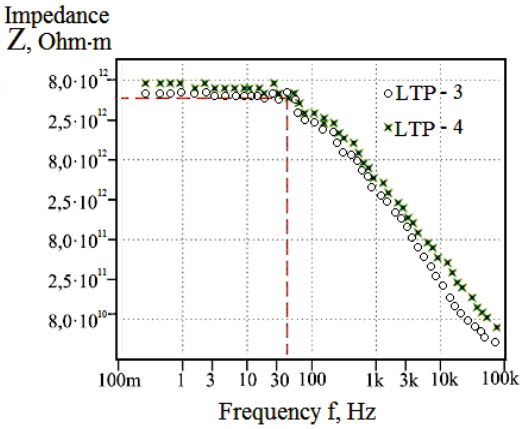


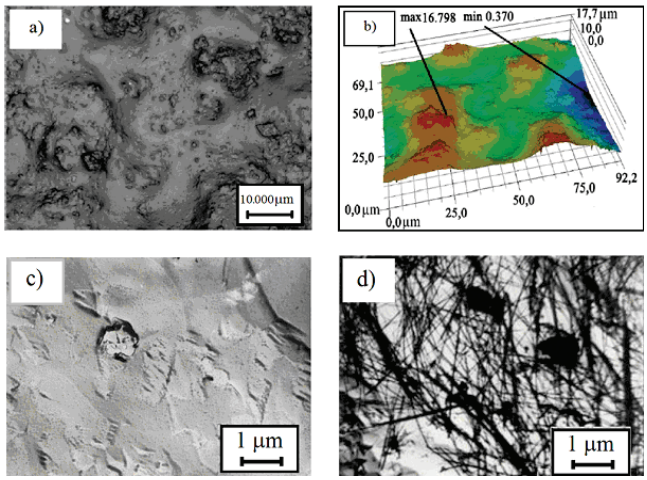
Fig. 3. Resistance dependence on the current frequency of low-temperature porcelain samples

After reaching a frequency of 100 Hz, the resistance decreases almost linearly. However, taking into account the fact that electrical porcelain products are designed to operate on direct and alternating current voltage with a frequency of up to 100 Hz, the resistance level of the obtained materials in this interval is more than sufficient. At the standard current frequency of electrical networks, the vast majority of European countries (f = 50 Hz) [17, 18], the resistance value of low-temperature porcelain is within ~ $(3–4)\cdot10^{12}$ Ohm·m up to 100 Hz.

The structural and phase features study of low-temperature electroporcelain was carried out using X-ray diffraction, raster electron microscopy, and laser scanning microscopy, which made it possible to obtain a high-detail image of the multilayer morphology object, to investigate the chip 3D topography, and to determine its micro relief parameters.

The results of X-ray diffraction showed the presence of two main phases: mullite (d = 0,2205; 0,254; 0,2695; 0,2883; 0,3389; 0,3430; 0,3775; 0,4265; 0,5401 nm) and amorphous glass phase, as well as quartz relics (d =0,2463; 0,3343; 0,4266 nm). According to quantitative X-ray phase and petrographic analyses, the bulk of low-temperature porcelain is represented by colorless transparent isotropic glass-mullite aggregates for ~ 85%. The crystalline phase is presented by mullite (~45%) and quartz (12%).

Microscopic studies indicate its homogeneous fine-grained structure without isolated areas of feldspar glass and large quartz inclusions. Quartz grains have melting traces, and their sizes do not exceed 4–7 μm; grains of feldspar relics are indistinguishable. Fig. 4a, 4b presents the results of scanning microscopic surface studies of the chipped sample LTP-4. A calm micro relief, slight blockiness with a thin interblock boundary (Fig. 4b), characterizes the sample surface. Fractographic studies of its chip allow observing a weakly pronounced micro relief and blocky material structure. In the micrograph obtained with a scanning electron microscope (Fig. 4c), it is possible to identify areas corresponding to mullite and glass phase e, and to evaluate the main porcelain phases ratio. To determine the morphological features of the porcelain mullite component, studies were also carried out on samples that were subjected to chemical etching in a 10% solution of hydrofluoric acid. The image (Fig. 4d) shows mullite crystals remaining after the removal of the glass phase, as well as small particles (~1 μm) of the mass components leaving. Secondary mullite crystals, formed because of recrystallization through the melt, have the appearance of long, thin needles up to 0.01 μm thick, which forms the porcelain "skeleton" in the form of felt-like interweaving.



a), b) according to laser scanning microscopy of the surface;
c), d) according to the electron microscopy of the chip

Fig. 4. Microstructure of the electroporcelain samples

CONCLUSIONS

As a research result, the mass compositions have been developed, which ensure the electroporcelain production with high physical, mechanical, and electrical properties at a reduced product firing temperature (1200°C). The obtained low-temperature electroporcelain meets the requirements for high-voltage insulators of subgroup 110 (GOST 20419, IEC 6723). It is shown that under energy-saving products firing conditions (firing temperature 1200 °C instead of 1350 °C, as adopted in the high-voltage electric porcelain technology) the complex fluxing component represented with feldspar concentrate and dolomite, ensures the active melt formation in a enough amount, that creates favorable conditions for the dissolution of quartz grains and the development of secondary mullite. It has been established that the use of dolomite in an amount of 2.0–2.5 wt. % (when introduced it into the mass in excess of 100%) contributes to the homogeneous fine-crystalline structure forming, ensures the material phase composition synthesis in form isotropic glass-mullite component (with a secondary mullite content of ~45 vol. %) and dispersed quartz grains without cristobalitization signs (~12 vol. %), and also determines the fiber-like mullite crystals morphological features that create a densely woven framework. It is these low-temperature electric porcelain structural and phase features that provide the functional properties complex required for high-voltage insulators.

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Internal Stress Visualization by Aligning Simulated and Experimental Deformations via Computer Vision and Model-Based Feature Matching

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***Abstract*—This paper presents a hybrid computer vision and simulation system for visualizing internal stress in mechanical structures through augmented reality (AR). The method overlays precomputed stress distributions from a Finite Element Method (FEM) simulation onto real-world images of a loaded object — specifically, a bent pipe. A modern segmentation AI model is employed to extract the pipe geometry from a single user click, ensuring robust performance in varied conditions. A simplified spline-based model is then used to describe pipe deformation in both simulation and reality, enabling feature-based matching between observed and simulated shapes. Experimental results show accurate matching and visually aligned overlays, offering an intuitive way to "see" invisible mechanical responses.**

Keywords—augmented reality, computer vision, mechanical deformation, internal stress

I. Introduction

Visualizing internal stress and deformation in mechanical structures is crucial for applications in engineering education, diagnostics, and maintenance. While simulation tools like the Finite Element Method (FEM) can accurately model such phenomena, their results are typically confined to abstract digital environments. Bridging these results with the real world in an intuitive and accessible way remains an open challenge.

The goal of this research is to develop a hybrid computer vision and simulation system that aligns experimental deformations of mechanical structures with precomputed FEM simulations, enabling users to overlay simulated stress distributions onto actual images of a loaded object — such as a bent pipe — thereby making internal mechanical responses visually accessible.

As illustrated in Figure 1, the system consists of two coordinated components. On the server side, FEM simulation results are post-processed and fitted with a simplified mathematical model that captures essential deformation characteristics. On the AR device side, a computer vision algorithm detects a real-world image of the object, and the same model is fitted to it as well. Features extracted from this shared mathematical model on both server and AR device sides allow for parameter-based matching between real and simulation data. The matched simulation frame is then visualized in AR, aligned with the real object.

By tightly coupling simulation and vision pipelines through a shared model abstraction, the system enables fast, lightweight, and sensor-free visualization of structural behavior. This method opens the door to accessible tools for learning, inspecting, or monitoring mechanical systems without requiring complex instrumentation or manual calibration.

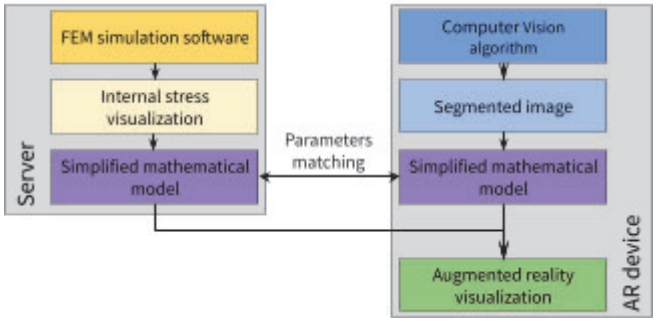


Figure 1. General structure of the proposed approach.

II. Problem Overview

In recent years, there has been growing interest in enabling intuitive, spatially grounded visualization of internal mechanical phenomena — such as stress or strain fields — in physical structures. While numerical simulation tools such as the Finite Element Method (FEM) allow precise stress estimation, these internal variables remain invisible during physical experiments, creating a critical gap between computational insight and real-world observation. Bridging this gap requires not only aligning experimental data with simulation output, but also finding ways to present internal forces to users in an interpretable way — ideally through augmented reality (AR).

Several reviews have documented the challenges associated with aligning experimental deformation data with FEM predictions. Chen et al. [1] present a comprehensive overview of finite element model updating and calibration techniques, demonstrating that this alignment is often an ill-posed, iterative process that depends on high-fidelity measurement inputs. Meng et al. [2] and Dong & Catbas [3] highlight the rise of optical, non-contact measurement methods — particularly Digital Image Correlation (DIC) — in mechanical and civil engineering. These methods allow the extraction of full-field surface displacements, but still fall short in terms of internal stress visualization. A detailed

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analysis by Blikharskyy et al. [4] confirms the effectiveness of DIC for strain measurement in reinforced concrete structures, but also emphasizes its limitations in visualizing internal mechanical behavior. Sun et al. [5] and Feng & Feng [6] further note that although Digital Twin frameworks and Computer Vision (CV) have greatly enhanced structural monitoring, most implementations are focused on displacements, cracks, or modal shifts — rather than inferred or simulated stress fields.

Within this context, several methodological directions have emerged. The first involves optical full-field deformation matching, primarily based on DIC. Lava et al. [7] propose "DIC-levelling," which interpolates simulation fields to DIC grids, enabling point-by-point validation. While this method ensures precise alignment between observed and simulated strain, it requires stereo setups, surface patterning, and calibration, limiting its field applicability. Similarly, Laux et al. [8] apply DIC-FEM fusion in heterogeneous material studies, reinforcing the value — but also the complexity — of these pipelines. These methods typically stop at verifying surface deformation and do not attempt to project internal stress fields into the visual domain.

A second group of approaches focuses on 3D reconstruction and simulation matching, often via stereo imaging, photogrammetry, or structured light. Honório et al. [9] reconstruct displacement fields from video and apply Boundary Element Methods to estimate stress. Ma et al. [10] and Graves et al. [11] perform monitoring of large-scale structures (a shipwreck stern and a bridge) using stereo-DIC and photogrammetric models, respectively. These techniques achieve remarkable geometric precision, but require costly equipment, controlled environments, or multi-view image setups. While useful for high-accuracy validation, they remain inaccessible in many practical or educational contexts.

The third direction integrates AR or Mixed Reality (MR) with simulation overlays. Unlike the above approaches, which emphasize measurement accuracy, these works focus on interpretability and user experience. Huang et al. [12] present frameworks that visualize FEM results in AR, enhancing understanding of vibration modes or stress distributions in real time. Erkek et al. [13] develop a mobile AR system that projects modal data onto physical structures. However, all of these rely on pre-aligned models or tracking targets and assume the real and virtual geometries are already registered.

Park and Kim [14] presented an AR system that overlays FEM stress results onto real-world objects using marker-based pose tracking and real-time force measurements. While their method successfully projects internal stress based on sensor input, it requires physical markers and external devices to drive simulation updates, which limits flexibility in uncontrolled environments. A notable example is the work of Paulus et al. [15], who introduced a real-time AR system capable of visualizing deformation and simulating topological changes such as cutting and tearing in soft volumetric objects. Their approach relies on a physics-based model and monocular video tracking to achieve coherent alignment between the simulation and real object, but still depends on optical flow, remeshing, and internal physics solvers.

These prior approaches illustrate both the promise and limitations of current tools. Full-field DIC systems provide dense data, but require complex setup. 3D surface reconstruction offers geometry-aware simulation mapping, but lacks flexibility. AR-based visualization achieves interpretability, but lacks deformation-based simulation alignment. Crucially, few of these methods attempt to automatically register simulation results to real deformed geometry in a lightweight, markerless manner, particularly for internal stress fields that are otherwise invisible.

In contrast, the present work introduces a hybrid pipeline that leverages modern segmentation AI models and simplified feature extraction to achieve visual alignment of stress simulations with real-world deformed structures. Rather than relying on 3D reconstruction or sensor arrays, we segment the object from an RGB image, extract its central axis, and compute features based on a simplified mathematical model of the object. With the same features being calculated for simulation frames as well, a frame from the precomputed FEM simulation is then selected via similarity of the feature vectors. This approach offers a practical balance between accuracy and usability, filling a clear gap in the literature.

III. Experimental Setup

To validate the proposed system, we constructed a simple physical experiment involving a PVC pipe under static loading conditions. The pipe was fixed horizontally between two rigid supports and subjected to a centrally applied load by suspending a weight as shown in Figure 2.



Figure 2. The experimental setup.

This setup produces a visible deformation of the pipe, serving as a real-world reference for visual comparison with simulated stress distributions. The resulting image is captured using a standard mobile phone camera. The selected camera position ensures clear visual segmentation of the object, while the symmetrical loading condition provides a consistent deformation pattern suitable for matching with simulation frames.

To complement the physical experiment, a similar configuration was recreated in ANSYS using FEM simulation. The virtual pipe was modeled with appropriate material properties and boundary conditions, and the increasing central load was applied to replicate the bending effect. The simulation output consists of a stress visualization sequence, exported as an animation showing the pipe deformation and internal stress distribution across several load levels as shown in Figure 3.

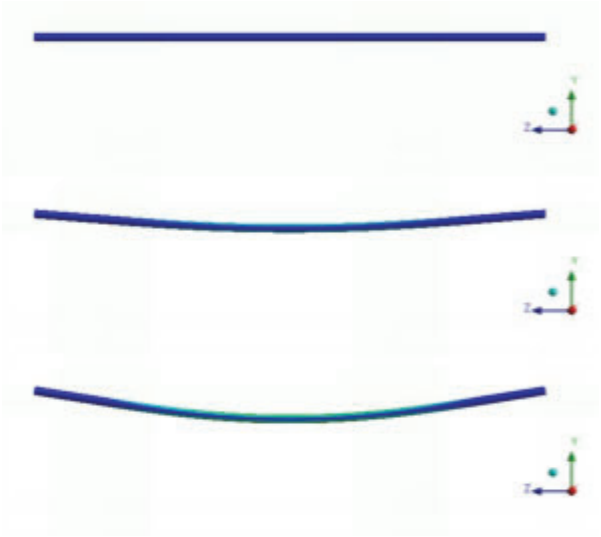


Figure 3. Frames from the experiment simulation created in ANSYS.

IV. Implementation Overview

The overall system pipeline is illustrated in Figure 4. A real image is first segmented to isolate the pipe, and from the resulting mask, curvature features are extracted. The same feature extraction process is applied to each simulation frame. Once features are computed, the real image is matched with the most similar simulated frame. Finally, we overlay the internal stress field from the selected simulation frame onto the real image, producing an intuitive visualization.

A. Image Segmentation

The first stage of the proposed pipeline focuses on isolating the pipe from the background using image segmentation. This step is essential for accurately extracting the pipe's central axis and deriving curvature features that enable alignment with simulated stress data.

To identify the most reliable segmentation strategy, we evaluated a range of methods — from classical Computer Vision techniques to a modern foundation AI model — based on robustness, ease of use, and accuracy. A comparative summary is presented in Figure 5.

Initial experiments were conducted using two conventional segmentation techniques: color thresholding and region growing.

- Color Thresholding. This method involved transforming the input image into HSV color space and isolating a range of hues corresponding to the pipe's color. The user provided two reference points to define the color bounds. While this approach was computationally lightweight, it was highly sensitive to lighting, shadows, and background textures, resulting in noisy or inaccurate masks.
- Region Growing. Using a single user-selected seed point, this method expanded to neighboring pixels based on intensity similarity. Although less sensitive to hue shifts, it still required careful tuning of similarity thresholds and often failed to segment the whole pipe reliably under varied conditions.

Both methods demonstrated limited generalizability and were ultimately insufficient for use in uncontrolled environments. To overcome these limitations, we integrated the Segment Anything Model (SAM) [16] into the pipeline. SAM is a foundation AI model capable of producing high-quality segmentation masks from a single user-provided point. In our case, the user clicks on the pipe, and the model returns a complete, well-defined segmentation mask.

SAM's robustness to lighting changes, background clutter, and object variation made it a superior choice. Furthermore, as a pretrained model requiring no additional training or domain-specific data, it significantly accelerates experimentation and deployment.

It is worth noting that the simulation frames also require segmentation. However, due to their uniform background, simple color thresholding with a single fixed bound followed by the removal of small contours is sufficient to obtain a mask of the pipe.

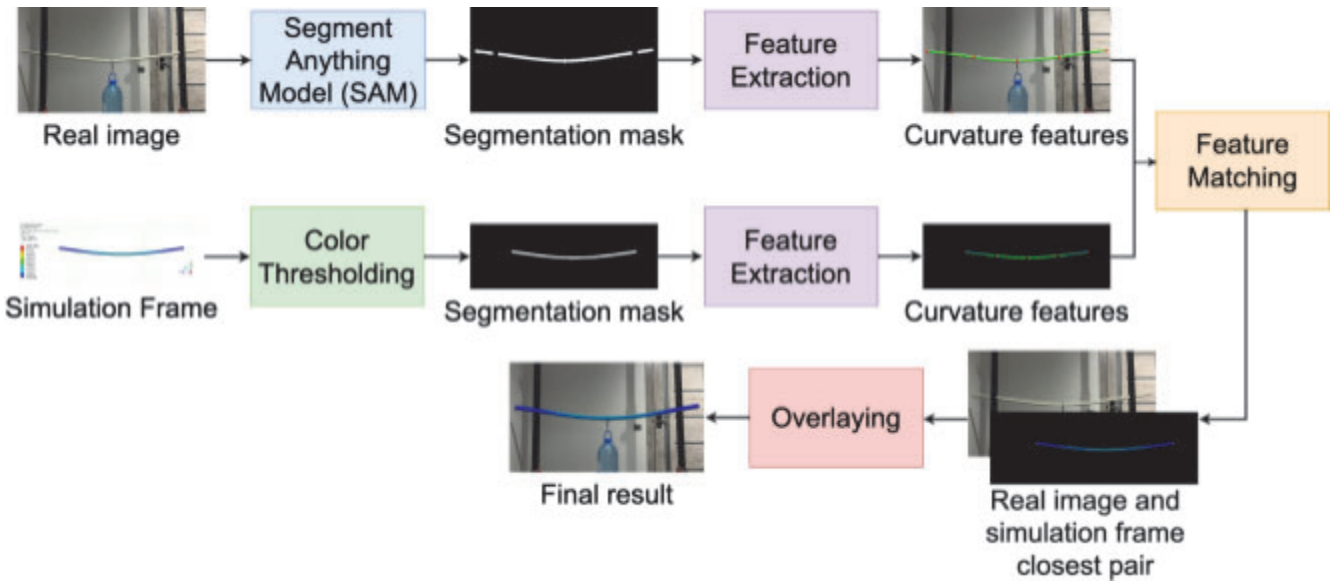


Figure 4. The architecture of the implemented solution.

Input object image	Object segmentation approach	Number of input points	Output object mask
	Color Thresholding	2	
	Region Growing	1	
	Segment Anything Model (SAM)	1	

Figure 5. Comparison of different object segmentation approaches.

B. *Simplified Mathematical Model*

After segmenting the pipe in the input image, the next step is to extract its central axis and approximate its shape using a simplified mathematical model. For this purpose, we employ a cubic spline that provides a smooth and continuous representation of the pipe's deformation, based on a set of sampled points from the binary segmentation mask.

To robustly summarize the pipe's shape, we extract a fixed number of key points along its horizontal extent. First, all foreground pixel coordinates (x_i, y_i) are collected from the binary mask $M(x, y)$. The x-axis is then divided into a set of evenly spaced quantiles (1).

$$X_q = \{Q_q(X) \mid q\{0.0, 0.25, 0.5, 0.75, 1.0\}\} \quad (1)$$

For each quantile position x_q, we identify all pixels with the same x-coordinate and compute the median of their y-coordinates to reduce sensitivity to local noise (2).

$$y_q = median\{y_i \mid x_i = x_q\} \quad (2)$$

This results in a sparse set of representative points (x_q, y_q) that capture the overall curvature of the pipe while filtering out high-frequency variations. The sampled points are then used to fit a cubic spline function (3).

$$S_q(x) = a_q + b_q(x - x_q) + c_q(x - x_q)^2 + d_q(x - x_q)^3, \quad x[x_q, x_q + 1] \quad (3)$$

This provides a piecewise-polynomial approximation of the pipe's centerline. An example of sampled points with a spline fitted to them is shown in Figure 6.



Figure 6. A spline fitted to uniformly sampled points of the pipe.

C. *Feature Extraction and Simulation Matching*

To compare the shape of the real pipe with its simulated counterparts, both are represented using a set of geometric features derived from their fitted spline models. These features serve as a compact and robust descriptor of the deformation state.

For both the real image and each simulation frame, we extract the pipe's central axis and fit a cubic spline as described earlier. From this spline, we compute the curvature profile $\kappa(x)$ at uniformly spaced horizontal positions along the pipe. This is done using the standard curvature formula (4), where $f(x)$ is the spline function representing the vertical displacement at position x, and $f'(x)$, $f''(x)$ are its first and second derivatives, respectively.

$$\kappa(x) = \frac{|f''(x)|}{\left(1+(f'(x))^2\right)^{\frac{3}{2}}} \quad (4)$$

To ensure physical consistency across varying image scales and camera distances, curvature is computed in real-world units by applying a pixel-to-meter scale factor that is estimated based on the known pipe diameter.

The resulting 1D feature vector captures the shape and bending behavior of the pipe using n uniformly sampled curvature values. This same procedure is applied to every simulation frame, allowing for direct comparison with the real-world observation. The real image's feature vector is then compared to those of all precomputed simulation frames. To quantify similarity, we compute the Euclidean distance between the real vector and each simulation vector (5), where D_i is the dissimilarity score for the i-th simulation frame, κ_{real} is the curvature profile of the pipe detected on a real image, $\kappa^{(i)}_{sim}$ is the curvature profile of the pipe from the i-th simulation frame, and $\|\ldots\|_2$ denotes the Euclidean norm.

$$D_i = \| \kappa_{real} - \kappa^{(i)}_{sim} \|_2 \quad (5)$$

The simulation frame with the smallest distance D_i is selected as the best match. An example plot of dissimilarity scores across all simulation frames is shown in Figure 7. The plot reveals a clear global minimum, indicating a strong match between the observed pipe and a specific simulation state.

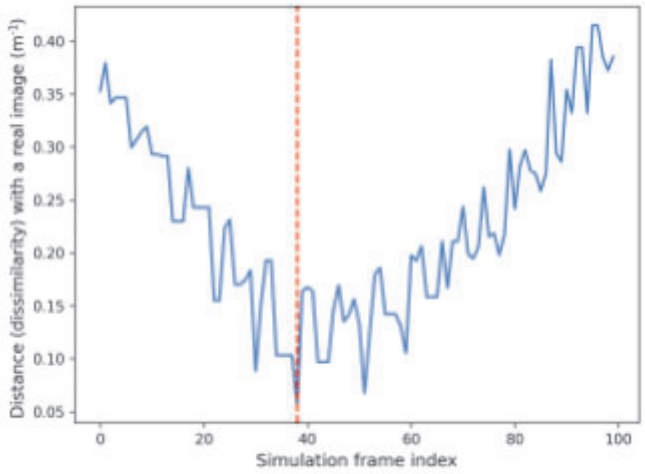


Figure 7. Numeric distance between a real frame and simulated ones.

This matched simulation frame is then used for visualization in the AR overlay, ensuring the displayed stress pattern aligns with the actual deformation observed in the scene.

V. Results and Discussion

The proposed system successfully overlays simulated stress distributions onto real-world images of a loaded pipe. As illustrated in Figure 8, the pipeline accurately identifies the pipe's shape, matches it to the closest simulation frame, and produces a clear, visually aligned augmented output. The overlay conveys not only the geometric deformation, but also the internal stress state, offering an intuitive visualization of invisible mechanical phenomena.

As for the quantitative evaluation, the mean relative error of the visually estimated deformations compared to physical measurement is estimated at 4.8%, making the method sufficiently accurate for the visualization purposes. The algorithm implemented in Python takes 2.9 seconds to process a single frame on the Apple M1 Pro CPU. With the significant portion of this time being spent on running SAM, one of the possible optimizations for real-time usage is replacing its inference on every frame with tracking.

The results demonstrate several strengths of the method:

- High-quality segmentation using a single prompt point, enabled by the Segment Anything Model (SAM).
- Accurate shape modeling via a lightweight spline-based approximation.
- Robust matching between a simulation and a real-world input.



Figure 8. Examples of AR overlay on real images of the experimental setup with different loads.

However, the system also has some current limitations. The most notable is its reliance on a fixed, optimal camera viewpoint to ensure proper alignment between the simulation and the real object. This assumption simplifies projection, but limits usability in uncontrolled or dynamic scenes. Additionally the pipeline still requires further optimization for real-time execution on mobile devices with a continuous video input.

Future work may focus on addressing these constraints as well as other improvements:

- Integrate pose estimation using fiducial markers or 3D scene reconstruction to support arbitrary and moving camera viewpoints.
- Port the pipeline to mobile AR platforms, enabling real-time feedback using live camera streams.

- Extend the method to more complex structures and loading conditions, paving the way for broader applications in mechanical diagnostics and engineering education.

VI. Conclusions

This work introduced a hybrid system that combines computer vision, simplified mathematical modeling, and FEM simulations to visualize internal stress in augmented reality. Using segmentation with a modern AI model, spline-based representation, and curvature matching, the method aligns experimental deformations with simulated stress fields, producing intuitive and accurate overlays without markers or special sensors.

The results confirm the feasibility of lightweight, markerless stress visualization and fulfill the stated goal. Future work will focus on enabling arbitrary viewpoints, real-time execution, and extending the method to more complex structures and loading conditions.

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Analysis and Optimization of Processes for Integrating Smart Grid Objects into the Critical Infrastructure Cyber Protection System

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Abstract — This paper proposes an integrated framework for strengthening Smart Grid cybersecurity through the combination of anomaly detection, standards compliance, and economic-mathematical modeling. Using the SustDataED2 dataset with anomaly injection scenarios, the Isolation Forest algorithm was applied to detect rare patterns of cyber intrusions in high-dimensional data streams. To ensure realistic calibration, empirical attack distributions were derived from recent industry reports, including Cisco Talos, Fortinet, IBM X-Force, and the Verizon Data Breach Investigations Report. Detected anomalies were systematically mapped to tactics and techniques defined in MITRE ATT&CK for ICS and aligned with IEC 62443-3-3 requirements, thereby creating a traceable link from technical observations to compliance obligations. In parallel, a risk matrix and an extended Gordon–Loeb-type investment model were used to optimize the allocation of resources between technical and organizational measures. The proposed framework enhances anomaly detection precision, increases resilience, lowers the probability of successful attacks, and improves cost-effectiveness of cybersecurity investments. These results provide actionable insights for energy enterprises and policymakers tasked with protecting critical infrastructures.

Keywords — Smart Grid cybersecurity, anomaly detection, isolation forest, data-driven decision-making, economic-mathematical modeling, technology foresight, risks

I. Introduction

The evolution of traditional power systems into Smart Grids represents a paradigm shift in energy infrastructure, driven by the integration of information and communication technologies (ICT) to ensure efficiency, sustainability, and reliability [1]. However, this digital transformation introduces a new set of cybersecurity challenges, particularly due to the proliferation of interconnected devices such as smart meters, sensors, distributed energy resources, and Supervisory.

Control and Data Acquisition (SCADA) systems. These elements, often operating over legacy protocols like Modbus or DNP3, significantly expand the attack surface for cyber threats [2, 3].

The vulnerability of Smart Grids is especially pronounced in the context of critical infrastructure, where disruptions can have cascading effects on national security and public safety. The 2015 and 2016 cyberattacks on Ukraine's power grid, including the use of malware such as BlackEnergy and Industroyer, marked the first known successful cyber-induced power outages and serve as key case studies in critical infrastructure cybersecurity [4, 5]. These incidents underscore the urgent need for proactive, intelligent, and resilient cybersecurity frameworks tailored to Smart Grid environments.

Ukraine’s energy sector exemplifies both the potential and perils of Smart Grid adoption. Burdened by aging Soviet-era infrastructure, fragmented ICT systems, and ongoing geopolitical instability, the sector has become a testing ground for advanced persistent threats. The urgency to protect this infrastructure is amplified by Ukraine’s alignment with European cybersecurity directives and the growing integration of renewable energy sources [6].

To address these challenges, this study proposes an integrated cybersecurity framework that enhances anomaly detection, incident response, and compliance with international standards, particularly the IEC 62443 series of standards, with emphasis on IEC 62443-3-3 [7]. A key component of this framework is the application of the Isolation Forest algorithm for anomaly detection, which has demonstrated effectiveness in identifying rare and subtle patterns indicative of cyber intrusions in high-dimensional Smart Grid datasets [8].

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Our experimental validation leverages the publicly available SustDataED2 dataset, representative of real household energy consumption, to simulate cyber-physical interactions within a Smart Grid [9]. Although originally designed for Non-Intrusive Load Monitoring, this dataset offers valuable time-series characteristics suitable for anomaly detection modeling.

II. Literature Review

The growing complexity of Smart Grid systems, driven by the convergence of energy networks and ICT, introduces profound challenges in ensuring cybersecurity for critical infrastructure. The literature offers a variety of perspectives on vulnerabilities, mitigation strategies, and the role of artificial intelligence (AI) in anomaly detection and threat response. This section synthesizes key contributions across five interrelated domains: Smart Grid security threats, detection techniques, architectural standards, anomaly detection using Isolation Forest (iForest).

A. Smart grid security challenges

Smart Grids, by virtue of their interconnectivity and reliance on automation, are inherently vulnerable to cyber threats. Traditional industrial protocols like IEC 61850, Modbus TCP/IP, and DNP3 were designed for isolated environments and often lack built-in security features, making them susceptible to packet sniffing, spoofing, and replay attacks [10]. The compromise of Ukrainian substations in 2015 and 2016 demonstrated the strategic intent and capability of advanced persistent threats to disrupt national-level energy services [4].

The integration of AI enhances the efficiency, reliability, and resilience of energy systems, while the role of digital resilience as a critical factor for their protection and adaptability is increasing [11].

B. Detection and Defense Mechanisms

Contemporary cybersecurity approaches prioritize early threat detection using both signature-based and behavior-based methodologies. The latter has gained traction due to its ability to detect zero-day exploits. A taxonomy of cyberattacks on Smart Grids, categorized into denial-of-service (DoS), man-in-the-middle, and false data injection attacks. [12, 13, 14].

Machine learning (ML) and deep learning (DL) algorithms have been successfully deployed for threat detection. Burian et al. explore supervised learning approaches using real-time Smart Grid telemetry [15]. Cambria et al and Mumuni et al. discuss explainable AI and the recent use of large language models for adaptive detection in dynamic environments [16, 17].

C. Standards and Architectures for Secure Integration

Compliance with global cybersecurity frameworks has emerged as the best practice for Smart Grid resilience. The IEC 62443 standard suite provides layered defense mechanisms for industrial control systems (ICS), emphasizing secure communication channels, access control, and anomaly handling [18]. The NIST SP 800-82 and the 2018 NIST Framework further complement this with actionable guidelines for ICS security posture management [19].

ENISA offers practical recommendations tailored for European Smart Grid implementations, advocating network segmentation, regular audits, and threat intelligence integration [20].

D. Isolation Forest for Anomaly Detection

IForest, introduced by Xu et al., has demonstrated high efficiency in unsupervised anomaly detection, especially for high-dimensional data with low contamination rates [21]. The algorithm works by isolating observations through recursive partitioning, with anomalies requiring fewer splits. Its computational efficiency and low memory footprint make it suitable for edge deployment in Smart Grid environments.

Recent studies have adopted iForest for power system cybersecurity. Mohammadpourfard et al. introduces a novel strategy that combines the strengths of iForest and Density-Based Spatial Clustering of Applications with Noise to identify and isolate anomalous data [22]. Shabad et all proposed an iForest-based anomaly detection model that isolates outliers using multiple decision trees, with its performance validated on a hardware testbed through simulation results [23].

III. Methodology

A. Risk Matrix Construction and iForest Integration

To assess cyber risks in the energy sector, a risk matrix was constructed using two key dimensions: impact severity and likelihood of occurrence. In this approach, cyber threats were first identified as potential adverse events (e.g., attacks on SCADA/ICS, DDoS, insider sabotage), and then translated into cyber risks by combining their probability of realization with the expected consequences [24-26]. A pyramidal structure was applied, where the top reflects threats of high criticality but lower frequency, while the base represents more common but less destructive scenarios. Furthermore, the iForest anomaly detection algorithm was integrated into the framework to examine its contribution to mitigating these risks.

To evaluate cybersecurity threats in the energy sector, we propose a three-tier matrix based on risk severity and operational impact. The top tier includes high-risk threats capable of disrupting critical infrastructure and operational control systems. This group comprises attacks on SCADA/ICS systems, ransomware campaigns, exploitation of legacy system vulnerabilities, and manipulation of sensor or telemetry data. Such threats pose systemic risks, potentially resulting in blackouts or the loss of control over energy distribution.

The second tier includes threats with moderate to high probability but lower systemic impact. These involve insider sabotage, electricity market manipulation through cyber means, supply chain compromises, data theft, and social engineering attacks. While these threats may not immediately affect infrastructure, they can lead to significant financial and reputational losses.

The third tier consists of high-frequency but lower-impact threats, such as Distributed Denial of Service (DDoS) attacks.

These typically result in short-term service interruptions without long-term operational damage (Fig. 1.).

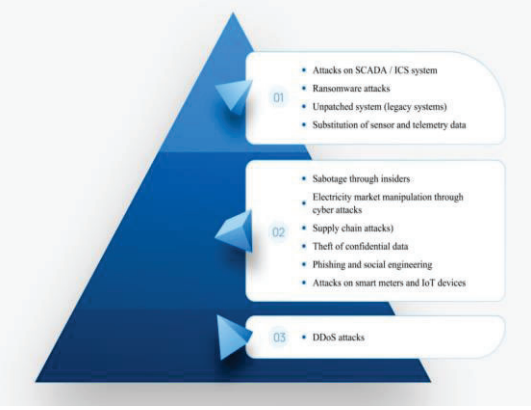


Fig. 1. Anomaly Detection Results Using Isolation Forest

The developed matrix provides a systematic translation of threats into measurable risks, enabling the prioritization of defensive measures in the energy sector.

This combination of threat identification, risk quantification, and machine learning–based anomaly detection establishes a proactive cybersecurity strategy. It enhances situational awareness and supports a risk-informed allocation of resources, thereby improving the overall cyber resilience of the energy sector.

B. Economic-Mathematical Modeling of Cybersecurity Investments

In addition to the technical anomaly detection and risk matrix framework, an economic-mathematical modeling layer is introduced to evaluate the cost-effectiveness of cybersecurity investments for Smart Grid operators. The rationale follows the seminal Gordon–Loeb model of information security investment [27] and its extensions [28], [29]. Defenders operate under limited budgets, while adversaries are not constrained in their expenditures. Moreover, investments in technical controls typically exhibit diminishing marginal returns, which requires balancing technical and organizational measures. The defender's expected utility is expressed as (1):

$$U = (z_{tech}, z_{human}) = R \cdot \left(1 - P(z_{tech}, z_{human})\right) - (z_{tech} + z_{human}) \quad (1)$$

were R represent the potential financial loss from a successful attack, z_{tech} denotes investment in technical solutions (hardware, software, anomaly detection), and z_{human} captures organizational expenditures (training, awareness, risk governance). The probability of attack success is modeled as (2):

$$P(z_{tech}, z_{human}) = \alpha e^{-\beta_1 z_{tech}} + (1-\alpha) e^{-\beta_2 z_{human}} \quad (2)$$

with $\alpha \in (0;1)$ representing the relative contribution of technical versus human factors, and $\beta_1, \beta_2 > 0$ denoting efficiency coefficients calibrated from the risk matrix. To illustrate, empirical threat likelihoods and cost baselines are drawn from industry reports.

According to the IBM Cost of a Data Breach Report 2024 and Verizon's DBIR 2024, the global average incident cost is in the multi-million USD range, while human error and credential misuse account for more than 30% of successful intrusions. Strategic insights from the WEF Global Cybersecurity Outlook 2025 [30] emphasize that the cyber threat landscape is increasingly shaped by geopolitical tensions, AI-driven attacks, and resource asymmetries. This supports our assumption that higher-risk operational environments may warrant increased allocation to technical controls (α) in the optimization model. These figures justify the numerical example presented in Table I [27, 28, 29, 30].

TABLE I. OPTIMAL ALLOCATION OF CYBERSECURITY INVESTMENTS

Parameter	Value
Potential Loss R	$1,000,000
Budget B	$200,000
Efficiency Coefficient β_1	0.00002
Efficiency Coefficient β_2β2	0.00005
Technical Weight α	0.7
Optimal $z*_{tech}$	$140,000
Optimal $z*_{human}$	$60,000
*Attack Success Probability P**	*0.18*

C. Standards-Driven Evaluation Framework (IEC 62443 and MITRE ATT&CK Integration)

At the analyzed energy-sector enterprises, IEC 62443 functions as a fundamental framework that establishes regulatory requirements, architectural principles, and cybersecurity processes for Smart Grid systems. The MITRE ATT&CK for ICS framework provides a structured taxonomy of adversarial tactics and techniques, enabling systematic classification of cyber threats. In parallel, the iForest algorithm serves as a practical detection mechanism for anomaly identification in operational data streams.

Within this study, these components were not merely considered separately but were integrated into a unified evaluation framework: anomalies generated and validated through our injection experiments were mapped to ATT&CK techniques and subsequently aligned with IEC 62443-3-3 requirements. This integration operationalizes the transition from technical anomaly detection to adversarial semantics and compliance obligations, thereby enhancing both the resilience and the adaptability of Smart Grid infrastructures. To move beyond raw detection metrics and ground anomaly detection in compliance-oriented decision-making, we designed a standards-driven evaluation framework.

IForest was trained and validated using our anomaly-injection scenarios on SustDataED2 to emulate ICS/OT-relevant behaviors, including ransomware-like inhibition of protective responses, unauthorized HMI bursts, structured exfiltration, PLC-logic manipulation surrogates, and protocol-command abuse. To ensure realistic scenario prevalence, we used empirical distributions of dominant attack categories derived from the Cisco Talos 2024 Year in Review [31], the Fortinet Global Threat Landscape Report 2H 2023 [32], the IBM X-Force Threat Intelligence Index 2025 [33], and the Verizon Data Breach Investigations Report 2024 [34]. Each report provides empirical incident shares by attack vector, such as ransomware, credential abuse, and data exfiltration. These shares were normalized to 100 percent and compared with the anomaly frequencies in SustDataED2. The injection frequencies were then scaled to match the empirical proportions, ensuring that the simulated scenarios reflect realistic industrial threat conditions. Detected anomalous windows were mapped to MITRE

ATT&CK for ICS [35] tactics and techniques by rule-based criteria derived from the injection semantics.

Suppression of protective functions was mapped to T0828 Inhibit Response Function, abnormal HMI authentication attempts to T0811 Initial Access, structured outbound activity to T0805 Collection, persistent PLC manipulation to T0843 Impair Process Control, and abnormal command bursts to T0851 Execution. Each mapped technique was subsequently aligned with IEC 62443-3-3 [36] system requirements to express detections in compliance-ready terms, including SR 3.1 system integrity, SR 1.2 role-based access, SR 4.1 confidentiality, SR 5.2 function control, and SR 3.3 communication authentication.

This creates a traceable chain from data-driven anomalies to adversarial intent and further to architectural and procedural controls. Table II presents the synthesis of this process for the most critical OT/ICS vectors observed in our experiments, connecting anomalies to ATT&CK techniques and the IEC 62443 requirements they inform [31, 32, 33, 34, 35, 36].

TABLE II. ALIGNMENT OF ICS/OT THREATS WITH MITRE ATT&CK AND IEC 62443 REQUIREMENTS

Attack Type	ATT&CK Tactic	Technique (ID)	IEC 62443 Requirement
Ransomware on ICS assets	Inhibit Response Function	T0828	IEC 62443-3-3 SR 3.1 (System integrity)
Data exfiltration via net.	Collection	T0805	IEC 62443-3-3 SR 4.1 (Confidentiality)
Unauthorized HMI access	Initial Access	T0811	IEC 62443-3-3 SR 1.2 (Role-based access)
Manipulating PLC logic	Impair Process Control	T0843	IEC 62443-3-3 SR 5.2 (Function control)
Modbus command abuse	Execution	T0851	IEC 62443-3-3 SR 3.3 (Communication authentication)

The results confirm that integrating IEC 62443 requirements, MITRE ATT&CK taxonomy, and iForest-based anomaly detection establishes a reproducible framework for translating technical anomalies into compliance-oriented insights. This traceable chain from experimental detection to adversarial modeling and governance requirements highlights the originality of our contribution and strengthens the resilience of Smart Grid infrastructures.

IV. Conclusion

This study has demonstrated a multi-layered framework for enhancing Smart Grid cybersecurity through the integration of anomaly detection, economic-mathematical modeling, and international standards compliance. The Isolation Forest algorithm applied to the SustDataED2 dataset successfully detected anomalies reflecting operational disruptions consistent with real-world OT/ICS attack vectors. By calibrating anomaly distributions against industry reports from Cisco Talos, Fortinet, IBM X-Force, and Verizon DBIR, the experimental design achieved realistic threat representation. The integration of MITRE ATT&CK for ICS with IEC 62443-3-3 requirements established a traceable chain from anomaly detection to adversarial semantics and further to governance obligations. This standards-based synthesis differentiates the proposed framework from purely algorithmic approaches, providing outputs that are both technically accurate and operationally actionable. Through the incorporation of a risk matrix and an extended Gordon–Loeb-type investment model, the study also demonstrated how economic-mathematical methods can optimize cybersecurity expenditures. This dual perspective—combining technical anomaly detection with investment optimization—contributes to enhanced resilience and efficient budget allocation in energy enterprises. The findings confirm that a unified approach linking anomaly detection, international standards, and economic-mathematical modeling advances both the theoretical foundations and the practical implementation of cybersecurity for critical infrastructures. Future research will extend this framework by incorporating real-time telemetry, hybrid detection models, and adversarial adaptation dynamics.

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Machine learning methods for image processing in plantography method for physical rehabilitation

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***Abstract*—Plantar pressure distribution provides insights into foot function, gait mechanics, and foot-related problems. The main machine learning methods for medical diagnosis and rehabilitation procedures are analyzed. A systematic review and analysis of the use of machine learning methods in the context of plantar pressure analysis is conducted. Sample size, pathology, number of pressure sensors, data acquisition device, use of other sensor devices, preprocessing dataset, ML method varieties, and evaluation metrics are evaluated. Four machine learning methods are used for analysis and rehabilitation prediction: k-NN, Decision Tree, SVM, Logistic Regression. A comparative analysis of the methods' performance is performed, and the main metrics are obtained.**

Keywords— machine learning, physical rehabilitation, plantar pressure, plantography

I. Introduction

Physical rehabilitation focuses on restoring physical function and strength. This is a major concern as more than 22% of the world's population will be aged 60 years or older by 2050. Physical rehabilitation interventions come in many forms: hospital-based or community-based, physician-led or self-guided, and multi-component or single-component programs. Although physical rehabilitation is widely available, it is often underutilized due to poor compliance worldwide [1]. Anybody may need rehabilitation at some point in their lives, following an injury, surgery, disease or illness, or because their functioning has declined with age. Technological advances have overcome some barriers to rehabilitation use, in recent years. For example, telerehabilitation can improve accessibility and digital technologies can improve compliance and monitoring of home exercise, but implementation challenges remain. More recently, technology supported rehabilitation has been enhanced by Artificial Intelligence (AI) [2]. In recent years, machine learning (ML) has emerged as a transformative technology across numerous sectors, including healthcare [3]. One particularly promising area is physical rehabilitation, where ML methods are being harnessed to improve diagnosis, personalize treatment, optimize therapy, and enhance patient monitoring. Lower limb rehabilitation is vital for recovery from injury, stroke, or surgery that affects mobility and quality of life. Traditional methods have mainly involved physical therapy, including manual exercises. Recently, ML methods have demonstrated potential to reduce the costs of lower limb rehabilitation by offering solutions that can be implemented with less reliance on highly trained therapists [4].

The study and mapping of pressure distribution and structure of the feet, called plantography, plays an important role in the diagnosis of gait disorders, postural disorders and lower limb pathologies. Traditionally used in podiatry, orthopedics, and rehabilitation, plantography involves the analysis of footprints, pressure patterns, and balance metrics. With the advancement of artificial intelligence, ML has emerged as a powerful tool to automate, enhance, and expand the capabilities of plantographic analysis. Machine learning enables high-precision, data-driven assessments that support early diagnosis, personalized treatment, and continuous monitoring [5].

II. Review of ML methods in diagnostics and rehabilitation

Historically, data mining techniques have been widely used in the field to make diagnoses and build strategies for the application of physical rehabilitation methods. The recent spread of advanced ML techniques has made it possible to accurately identify abnormalities and detect threats to patient health both in real time and in post-physical therapy analysis [3]. Scenarios that are of interest are primarily characterized by a non-stop flow of data (such as data on the same packages or add-ons), which indicate illegal behavior during basic procedures in physical rehabilitation [6].

Supervised Learning [7] is the techniques such as support vector machines (SVM), decision trees, and neural networks are commonly used to classify and predict patient movements, assess motor function, and determine therapy effectiveness. For instance, ML can distinguish between correct and incorrect execution of rehabilitation exercises using data from motion sensors. SVM is seen as a widely used supervised learning algorithm in the context of physical rehabilitation of patients. Unsupervised Learning: Clustering algorithms like k-means or hierarchical clustering can identify hidden patterns in patient recovery trajectories. This can assist in grouping patients with similar rehabilitation needs or predicting long-term outcomes based on early recovery indicators. Clustering algorithms group similar transactions or boundary objects within the patient data base. Algorithms, such as k-means clustering or DBSCAN, divide data into clusters, each of which represents a group of similar data points [6]. Clustering algorithms can differentiate regular and non-routine behavior by tracking patterns in a skin cluster.

Reinforcement Learning (RL) [8]: RL has shown promise in robotic rehabilitation, where agents learn optimal therapy strategies through trial and error. For example, robotic exoskeletons powered by RL algorithms can adjust assistance levels based on real-time patient performance, encouraging active participation and motor learning. RL algorithms, such as Q-learning - a method that captures individual intelligence in an agent-based approach - can help create intelligent agents under rehabilitation procedures for optimal security policies. Agents in this method promote safety by checking blocks, choosing methods by consensus and avoiding actions that can cause harm. In full collaboration with RL methods, agents control the flow of data about the patient's condition, set operations and impose

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penalties based on the results of their activities, thereby beginning to modify their own knowledge [18]. This adaptability is particularly valuable in the dynamically changing patient safety landscape.

Deep Learning (DL): Convolutional Neural Networks (CNN) and Recurrent Neural Networks (RNN) are applied to process complex time-series data from wearables or video recordings. These models can accurately recognize human activities, detect subtle changes in gait, and predict the likelihood of falls or complications. Algorithms of deep learning, such as CNN and RNN, significantly contribute to ensuring the effectiveness of rehabilitation [9]. CNN algorithms are more effective in analyzing structured data, such as graphs of physical therapy activities. Bayesian Network: The Bayesian Network algorithm uses probabilistic reasoning to help the system model and analyze elementary correlations between variations in variables and the security of the procedure (see, for example, [6]). The Bayesian network provides valuable information such as patient health safety risks and vulnerabilities and enables stakeholders to act accordingly [10]. By incorporating additional communication and handling missing or incomplete data using probabilistic reasoning, this algorithm is flexible and quickly adapts to changing real-world scenarios.

Plantography, as the study and mapping of the feet's pressure distribution and structure, plays a vital role in diagnosing gait abnormalities, postural disorders, and lower limb pathologies. Traditionally used in podiatry, orthopedics, and rehabilitation, plantography involves the analysis of footprints, pressure patterns and balance metrics. With the advancement of artificial intelligence, ML has emerged as a powerful tool to automate, enhance, and expand the capabilities of plantographic analysis. ML enables high-precision, data-driven assessments that support early diagnosis, personalized treatment, and continuous monitoring. However, interpreting plantographic data manually is time-consuming and prone to inter-observer variability. This is where ML significantly improves the process.

Supervised ML algorithms such as SVM, random forests, and convolutional neural networks (CNN) can be trained on labeled plantographic datasets to classify foot types or detect pathological conditions. For instance, CNN can identify subtle variations in pressure distribution patterns to distinguish between normal and abnormal gait with high accuracy. The study developed a model for classifying foot types using images and numerical data on foot pressure, which achieved 92.55% accuracy [11]. Dynamic plantography generates time-series data reflecting a person's gait. Recurrent neural networks (RNNs), particularly Long Short-Term Memory (LSTM) models, can model temporal dependencies and detect abnormalities in walking patterns—useful in early detection of neurological or orthopedic disorders. In particular, a DL model using LSTM was proposed to predict foot pressure distribution based on a small number of sensors, demonstrating its effectiveness in gait analysis [12]. Recently, an intelligent algorithm for classifying and redistributing plantar pressure in wearable devices was developed to prevent foot ulcers in patients with diabetes. small changes in pressure distribution may indicate the onset of ulcers or tissue damage. ML models can continuously analyze plantographic data to alert healthcare providers of risks before visible symptoms occur, thereby reducing complications and healthcare costs [12].

ML can be used to match plantographic features with optimal orthotic solutions. By analyzing thousands of footprint patterns, an ML model can suggest personalized insoles or corrective footwear that accommodate individual biomechanics and support rehabilitation. A study by [13] developed DL architectures to estimate foot pressure distribution from video frames, which could be applied to assess fall risk . ML is revolutionizing plantography by transforming it from a diagnostic support tool into an intelligent system capable of real-time analysis, early diagnosis, and personalized care. As algorithms become more sophisticated and data more accessible, ML-driven plantographic systems will likely become standard in clinics, hospitals, and even home health settings. The convergence of AI and biomechanical science holds tremendous potential to improve outcomes for patients with mobility issues and foot disorders.

III. The Analysis of Use ML for Foot-Type Classification

It is known that foot deformation is mainly caused by changes in the arch of the foot. To conduct analysis and collect data, it is necessary to determine the main parameters that determine the anomalies and dynamics of the foot. Then these parameters can be used for analysis using appropriate ML methods. The main task is to process the foot pressure measurement database. Identify relevant features, perform statistical processing. To demonstrate the possibility of using ML methods, it is necessary to show that all data are grouped according to their properties. Thus, using the method of reducing the effective dimensionality of the feature space, show the possibility of distributing data into clusters on a two-dimensional plane. After that, apply and compare a number of ML methods to the existing database.

A. Selection and collection of data

Collecting foot pressure data typically involves measuring how pressure is distributed across the bottom of the feet during standing, walking, or running. This data is valuable in clinical gait analysis, sports performance optimization, footwear design, and diagnosing foot-related conditions like plantar fasciitis or diabetic foot ulcers.

There are many ways to collect foot pressure data. The most common method is to use pressure-sensitive mats (force platforms). In this case, the user stands or walks on a mat with built-in pressure sensors. Popular systems for this are: Tekscan, Zebris, AMTI, RSScan. At the same time, shoe pressure measurement systems are widely used, when thin insoles with pressure sensors are placed inside the shoe. Popular systems: Moticon, Pedar (Novel), F-Scan (Tekscan). Wearable sensor technologies are more expensive; built-in pressure and motion sensors (often with IMU, inertial measurement unit) are built into smart insoles or socks. Examples: Plantiga, OpenGo, Digitsole Pro. The most democratic are DIY methods with inexpensive sensors. In this case, piezoresistive or capacitive pressure sensors (e.g. FlexiForce, Force Sensitive Resistors - FSR) with microcontrollers (Arduino/Raspberry Pi) are used. In the process of data collection, it is proposed to use the method described in [14]. The same method was used in later studies [11]. The subjects were asked to stand in a standing position with both feet shoulder-width apart in the foot position

indicated on the measuring device. The measuring device is shown in Fig. 1 [11]. It has a detection area of 400 mm x 400 mm and consists of 2304 pressure sensors with 48 x 48 matrices. The pressure value data is measured in kPa, and the image data is a 918 x 918 intensity color image that was obtained by evaluating the pressure distribution.

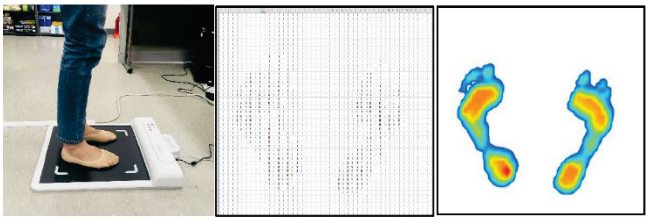


Fig. 1. Experimental method and results. (a) Plantar pressure measurement, (b) numerical pressure data, (c) distribution of pressure image data (from [11]).

The dataset [15] contains data about three types of footsteps in children between 11 and 12 years old (normal, flat and cavus). Fourteen pressure sensors were used to collect data, distributed across three zones of the foot: the forefoot, midfoot and rearfoot (see Fig. 2). For the left foot: S1 and S2 cover the forefoot area. S3, S4, and S5 the midfoot area. S6 and S7 the rearfoot or heel area. Similarly, for the right foot: S8 and S9 represent the forefoot area. S10, S11, S12 the midfoot area. S13 and S14 the heel area. The values of each sensor are read by the analog inputs of an Arduino mega 2560. It should be noted that the expected ones are well organized and presented as a .cvv file. Since the data is divided into three classes (class: normal footstep, flat footstep, cavus footstep), it opens the way for applying suitable ML method to carry out classification.

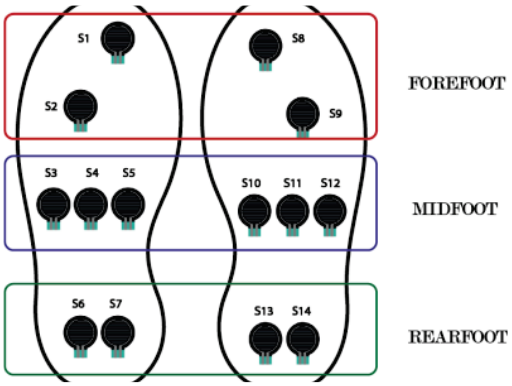


Fig. 2. The pressure sensors are represented by black circles, which are located in the three zones of each foot (from [15]).

The dataset [16] comprises plantar pressure images collected as part of a study aimed at investigating foot health through advanced pressure analysis techniques. The Footbird 3.0 foot pressure scanner, equipped with flexible pressure sensors, was utilized to capture the plantar pressure distribution characteristics of the foot. The dataset includes a total of 400 plantar images categorized into five distinct foot types: normal feet, high-arched feet, flat feet, inward heel tilt, and outward heel tilt. Additionally, some images may represent mixed foot types, providing a comprehensive overview of various foot health conditions. This dataset serves as a valuable resource for researchers and practitioners in the field of podiatry and biomechanics, facilitating further analysis of plantar pressure characteristics and their implications for foot health assessment. All images in this database have a resolution of 702x702, thus opening the way for constructing a classification method using DL, similar to, for example, MNIST.

B. Results of application of ML methods

In this study, a data-driven artificial intelligence methodology was proposed for foot health diagnosis using plantar pressure data. The methodology of [11] was used. In order to use the approach based on the analysis of foot pressure distribution, a correspondence model was built that can construct pressure distributions from a color image. For this purpose, the database [16] was used, for which image data (pressure distribution) and numerical data (pressure value) were first collected, after which data preprocessing methods were developed and implemented for the two heterogeneous types of data. Next, foot deformity type classification models were built for image data and numerical data, such as VGG16 (image feature extraction neural network), InceptionV3, k-NN, and CART. To improve the reliability and accuracy of the model, metadata was generated based on the predicted values of each model. A stacking ensemble combining VGG16 and k-NN was proposed, which has better performance than other models. The stacking ensemble model combines the advantages of two heterogeneous models to improve the accuracy and reliability of other models that are used individually. The proposed classification model for determining the type of foot deformity obtained an accuracy of F1 of 92.55% by using only the pressure data measured from 400 subjects through careful data preprocessing and artificial intelligence model generation. The proposed model achieved high classification accuracy with a small amount of data, but if the number of foot pressure datasets is increased and the diagnosis results of experienced medical personnel are added, it is expected that a more accurate foot health diagnosis model can be developed. The developed model is considered to be useful as an objective tool for medical personnel to diagnose foot health and can be expanded.

The entire dataset was split as follows: 80% for the training dataset and 20% for the test dataset. The training dataset was used to train the model based on stratified 5-fold cross-validation. Finally, each model trained using each subset was evaluated on the test dataset by calculating the mean of the f1 scores. The models trained on image data (VGG16 and InceptionV3) show relatively lower performance than the models (k-NN and CART) trained on numerical data. This is because it is more difficult to extract features in the process of extracting plantar image data than numerical data. The principle of CNN, which is the basis of the VGG16 and InceptionV3 model, is to extract features by moving the kernel within the image. However, the images on the foot boundary are similar in color and shape; therefore, the features are extracted and learned in the same group. The classification error rate is high because they are actually different types of images. In the histogram of AI scores, the data in the range of 0.1 from the boundary accounts for 38 out of 192, which is about 19.79% of the total which is almost completely consistent with the results [11]. Particularly preferable in terms of the application of ML methods is data analysis [15]. The indicators, as shown in Fig. 2, were measured at fourteen points on the foot. The set of indicators is divided into three classes (1- normal footstep, 2 - flat footstep, 3 - cavus footstep). In Figure 3 we see an example of the corresponding normal footstep indicators.

	S1	S2	S3	S4	S5	S6	S7	S8	S9	S10	S11	S12	S13	S14	Label
12	944	943	751	751	0	937	937	970	966	0	751	382	991	957	1
13	940	939	759	759	0	933	932	964	961	0	759	653	988	952	1
14	954	945	488	488	0	936	936	539	970	0	488	0	992	957	1
15	939	936	952	952	0	930	930	963	960	0	952	719	987	950	1
16	945	943	957	957	0	937	937	969	966	0	957	353	993	958	1

Fig. 3. Example of pressure sensor readings.

So, in the database used for analysis, there are 14 features classified into three classes. Let's try to classify the data without marking. To do this, it is necessary to identify relevant features using correlation analysis. There are many methods for data analysis, many of which are related to graphical display. One of the effective methods in Data Science is the pairwise correlation graph. It allows you to detect the dependence between features, which can lead to a decrease in the feature space. It can also be used to detect the correlation between a feature and a target (the desired value). A pairwise correlation graph is a visual or mathematical representation that shows the correlation relationships between pairs of variables in a dataset. Each node in the graph represents a variable, and each edge (connection between nodes) represents the correlation coefficient between a pair of variables. Figure 4 illustrates the results of the correlation analysis based on the analysis of 14 features of the studied database using the corr method.

	S1	S2	S3	S4	S5	S6	S7	S8	S9	S10	S11	S12	S13	S14	Label
S1	1.00	0.36	-0.03	-0.05	-0.03	0.34	0.36	0.14	0.25	-0.03	-0.05	0.24	0.36	-0.21	0.03
S2	0.36	1.00	0.09	-0.03	-0.03	0.97	1.00	0.44	0.74	-0.03	-0.03	0.77	1.00	-0.16	0.05
S3	-0.03	0.09	1.00	0.31	0.14	0.16	0.10	0.27	0.18	0.14	0.31	0.29	0.08	-0.24	0.02
S4	-0.05	-0.03	0.31	1.00	0.51	-0.01	-0.03	0.07	0.01	0.51	1.00	0.04	-0.03	-0.01	-0.79
S5	-0.03	-0.03	0.14	0.51	1.00	-0.01	-0.03	0.02	-0.02	1.00	0.51	0.01	-0.03	-0.00	-0.00
S6	0.34	0.97	0.16	-0.01	-0.01	1.00	0.97	0.44	0.72	-0.01	-0.01	0.75	0.97	-0.18	0.05
S7	0.36	1.00	0.10	-0.03	-0.03	0.97	1.00	0.45	0.74	-0.03	-0.03	0.78	1.00	-0.16	0.05
S8	0.14	0.44	0.27	0.07	0.02	0.44	0.45	1.00	0.69	0.02	0.07	0.65	0.41	-0.08	0.01
S9	0.25	0.74	0.18	0.01	-0.02	0.72	0.74	0.69	1.00	-0.02	0.01	0.69	0.70	-0.13	0.03
S10	-0.03	-0.03	0.14	0.51	1.00	-0.01	-0.03	0.02	-0.02	1.00	0.51	0.01	-0.03	-0.00	-0.00
S11	-0.05	-0.03	0.31	1.00	0.51	-0.01	-0.03	0.07	0.01	0.51	1.00	0.04	-0.03	-0.01	-0.79
S12	0.24	0.77	0.29	0.04	0.01	0.75	0.78	0.65	0.69	0.01	0.04	1.00	0.75	-0.18	0.04
S13	0.36	1.00	0.08	-0.03	-0.03	0.97	1.00	0.41	0.70	-0.03	-0.03	0.75	1.00	-0.15	0.05
S14	-0.21	-0.16	-0.24	-0.01	-0.00	-0.18	-0.16	-0.08	-0.13	-0.00	-0.01	-0.18	-0.15	1.00	-0.05
Label	0.03	0.05	0.02	-0.79	-0.00	0.05	0.05	0.01	0.03	-0.00	-0.79	0.04	0.05	-0.05	1.00

Fig. 4. Correlation analysis results for features.

Thus, the readings of sensors 'S2', 'S9', 'S4', 'S11', 'S7', 'S13' are used. Indeed, for example, sensor S2 has a high level of correlation with S7 and S13 (they fall into the red zone). As can be seen from Figure 2, these sensors give us the most significant indicators on both the left and right feet. To do this, we remove 'Label' so as not to spoil the data processing using a known answer. On the other hand, for clarity, we need to reduce the number of features to consider descriptive statistics. Next, we will output statistics using the pandas library. To obtain statistics for a trimmed data set, use the describe method (see fig.5). This method is commonly used in data analysis libraries like pandas in Python. When called on a DataFrame or Series, it returns summary statistics for numerical columns by default. Thus, the standard deviation for the readings of the S4 and S11 sensors, which are located in the center of the foot, is very large, std=434.035. This follows from the correlation analysis. It can be concluded that these sensors do not provide a relevant set of readings for conducting ML. Despite the fact that the number of features was reduced to 6, but in order to obtain a visual description, to determine the clustering picture, it is necessary to use the mechanism of reducing the effective dimensionality

	S2	S9	S4	S11	S7	S13
count	400.000000	400.000000	400.000000	400.000000	400.000000	400.00000
mean	974.392500	984.190000	551.430000	551.430000	970.487500	1003.71750
std	39.617829	40.585179	434.035277	434.035277	42.746731	15.87775
min	930.000000	810.000000	0.000000	0.000000	924.000000	985.00000
25%	943.000000	966.000000	0.000000	0.000000	936.000000	991.00000
50%	944.000000	968.000000	726.000000	726.000000	937.000000	992.00000
75%	1023.000000	1023.000000	957.000000	957.000000	1023.000000	1023.00000
max	1023.000000	1023.000000	1023.000000	1023.000000	1023.000000	1023.00000

Fig. 5. Statistical characteristics for selected features.

Principal Component Analysis (PCA) is a widely used dimensionality reduction technique in ML and statistics. It transforms a high-dimensional feature space into a lower-dimensional one while retaining as much variance (information) as possible. The principle behind PCA is to find a compressed representation of the original data in a lower dimension that maximizes the overall variance (or fluctuation) of the original data. This means that the overall pattern is still preserved, while the data is simplified as much as possible. Fig. 6 demonstrates the result of the PCA algorithm. Clusters with synthetic data, which display the structure of the general database, are located on a two-dimensional plane.

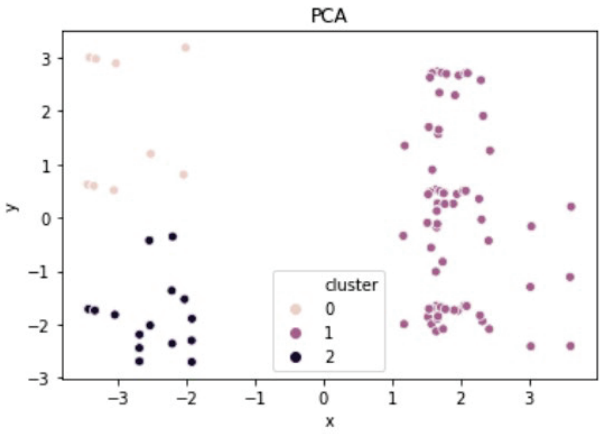


Fig. 6. Data after dimensionality reduction by PCA methods.

Thus, the data is remarkably clustered on a 2D plane. This opens up opportunities for applying an effective classification strategy using a variety of ML methods.

Classification models can be used to predict the dependent variable (class of each pressure distribution measurement). For this, 4 different classification algorithms are used, namely k-NN, decision tree, SVM, and linear regression. Finally, we compare their accuracy to find a suitable classification method for this problem. The given dataset is divided into two separate training and testing sets. The training set is used to train the model, and the testing set is used to predict the accuracy of each model. Finally, we compare the results. Here, 80% of the dataset was considered for training and 20% of the dataset for testing.Using the k-NN method it was found that 75 samples from the 80 test samples were predicted correctly, which is an accuracy of 91.11%. The decision tree showed the best result: 77 samples out of 80 test samples were predicted correctly, which is an accuracy of 93.33%. SVM: 68 samples from the 80 test samples were predicted correctly, which is an accuarcy of

84.44%. Logistic Regression: 66 samples from the 80 test samples were predicted correctly, which is an accuarcy of 84.44%. The results presented in Table 1 show that when comparing the scores and accuracy of each model, it is evident that the decision tree classification algorithm provides the highest accuracy for this problem using the foot pressure distribution dataset. It should be noted that the strategy of applying machine learning methods allows, using a set of optimal parameters, to build an effective strategy for further rehabilitation and very useful with using telemadecine services [17].

TABLE I. ACCURACY OF EACH MODEL USING THE TEST DATA

Algorithm	Jaccard Similarity Score	F1 score	Log Loss
k-NN	0.90	0.89	NA
Decision Tree	0.92	0.93	NA
SVM	0.91	0.9	NA
Logistic Regression	0.84	0.83	0.84

IV. CONCLUSION

An analysis of ML in clinical rehabilitation shows the many ways in which these technologies can improve patient outcomes. The study shows how ML has been used for targeted treatment, predictive modeling, and outcome assessment. The ever-changing clinical rehabilitation environment, which emphasizes individualized and flexible treatment methods, requires recognition of change. Since when setting a strategy for rehabilitation, recovery, and diagnostics we always deal with the analysis of a large data flow, it is natural to determine which areas of machine learning are natural to apply and for what purposes. To improve the reliability and accuracy of the model, metadata was generated based on the predicted values of each model. Next, we built and followed a more realistic model, the database of which consists of 400 measurements, with each measurement data was taken from 14 sensors. In this paper, an effective method for processing data obtained using pressure sensors has been developed. We have shown a way to reduce the number of database features and highlight the most relevant ones. Statistical characteristics of the data [18] have been obtained, and correlation properties have been considered. Based on the PCA method, visual images of data distribution on the plane have been obtained. It has been shown that the data are effectively clustered. This will open the way for the application of many ML methods for processing and interpreting foot pressure distribution data. The data was processed, divided into a training and a test block. All data is divided into three classes depending on the condition of the foot. Four machine learning methods were used, all of which showed good accuracy, from 84.44% to 91.11%. The main accuracy indicators for each model were obtained. Based on the comparison, it was found that the decision tree classification algorithm gives the best result. Interesting development of ideas for image processing in future developments

This emphasis on the application of ML methods in plantar pressure research highlights the importance of this approach in the cloud of biomechanical research and paves the way for future research that may lead to innovative solutions in the field of foot health and biomechanics.

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Neural Network Learning of Decision-Making Management Algorithms in Non-Invasive Smart Devices for Cardiovascular System Diagnostics

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***Abstract*—Cardiovascular diseases (CVDs) are the leading cause of death globally, necessitating the development of efficient and interpretable diagnostic tools for real-time and out-of-hospital monitoring. This paper presents a hybrid neural network model that integrates clinical diagnostic logic directly into its architecture to enhance explainability and accuracy. A formalized algorithm based on biosignals — such as electrocardiography (ECG), photoplethysmography (PPG), and heart rate variability (HRV)—was developed to emulate expert decision-making. The algorithm was embedded into a Rule Injection Layer (RIL), enabling the network to combine expert knowledge with data-driven learning. Experiments using synthetic and real datasets demonstrate high diagnostic performance (up to 97.1% accuracy) and robustness under varying signal conditions. The model is optimized for deployment in low-power embedded systems, providing a reliable solution for non-invasive CVD monitoring with interpretable outputs. Explainability is further supported using the LIME framework, which highlights feature contributions for clinical validation.**

Keywords—neural networks, explainable artificial intelligence, non-invasive monitoring, biomedical signal processing, heart rate variability, decision-making algorithm, medical logic, hybrid model, embedded systems

I. Introduction

Cardiovascular diseases (CVD) remain the leading cause of death worldwide, accounting for more than 17 million deaths annually according to the World Health Organization [1]. In this regard, the issues of early diagnosis, continuous monitoring and prediction of cardiac events, including in the out-of-hospital environment, are becoming extremely relevant. Modern advances in the field of medical sensor systems have made it possible to create non-invasive long-term monitoring devices that provide real-time biosignal collection [2–4]. However, the effectiveness of such devices directly depends on the ability to process these signals and make accurate diagnostic decisions.

In the context of the development of artificial intelligence (AI), a special place is occupied by neural networks (NNs), capable of analyzing complex patterns in biomedical data [5–7]. However, most of the implemented approaches are focused on the "black box", that is, on nonlinear recognition without an explained decision-making logic, which limits their practical application in clinical settings [8–9]. At the same time, when integrating such networks into medical devices, it is critically important to ensure interpretability and logical sequence of actions leading to the formulation of a diagnosis. In addition it is important to substantiate a comprehensive assessment indicators that will consider multivariate explanation approach [21 23].

In this work, the main focus is on the development of a formalized decision-making algorithm that takes into account medical logic and clinical features of CVD, and which can be further used to train a neural network. This approach allows for controlled behavior of the network during the diagnostic process and significantly increases the reliability and explainability of the results [10–12].

The aim of the research is to create an algorithm for making diagnostic decisions based on analytical rules, formalize it in a format suitable for training a neural network, and implement this logic into the structure of a NM integrated into non-invasive monitoring devices. Such a system is able not only to detect deviations in biosignals, but also to explain on the basis of which parameters a diagnostic conclusion was made, which is critically important for medical professionals [13–15].

The proposed neural network architecture combines classical deep learning elements (e.g., convolutional or recurrent layers) with built-in logic modules that model expert rules. This allows for a new approach—a hybrid model for intelligent decision-making that can learn from both data and formalized knowledge [16–19]. It is expected that such a model will demonstrate higher accuracy, stability, and

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explainability compared to traditional NMs, especially in cases of small or heterogeneous data sets.

The following sections consider the structure of the decision-making algorithm, its mathematical formalization, the architecture of the corresponding NM, the methodology for its training, and the methods for its integration into a non-invasive device. Experimental results of testing the model on real and synthetic signals are presented, and its effectiveness is assessed according to clinical criteria [20].

II. Development of a Decision-Making Algorithm in the Diagnosis of CVD

The purpose of this section is to build a formalized decision-making algorithm for diagnostic inferences based on biosignals (including ECG, photoplethysmogram (PPG), and heart rate (HRV)) in non-invasive devices. Such an algorithm should take into account medically relevant features, be suitable for implementation in a neural network structure, and support integration into the limited computing resources of the device.

The development is centered on the logic of a logical-analytical rule that simulates the diagnostic thinking process of a cardiologist, formalized as a multi-level system of conditions that takes into account numerical parameters, time dependencies, and relationships between indicators.

Based on the analysis of clinical recommendations [1, 4, 8] and expert rules, a basic logic was created, which includes the following diagnostic states: normal rhythm, sinus tachycardia/bradycardia, premature contractions (extrasystole), atrial fibrillation, signs of ischemia. For each state, threshold values of key signs were determined as an example in Fig. 1. The algorithm for diagnosing tachycardia and the rules for use are given:

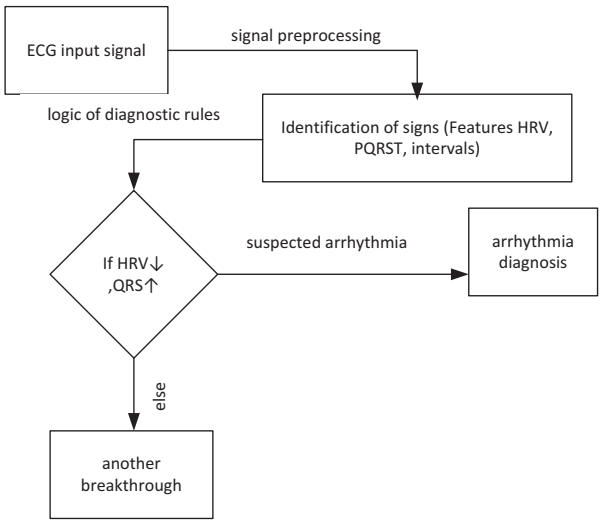


Fig. 1. General decision-making algorithm (Rule-Based Decision Tree)

These rules are generalized in the form of logical operations, implemented through activation functions and weight coefficients in NM. Formally, logic can be represented as a logical-mathematical tree, which includes conjunctions and disjunctions of conditions.

To describe the algorithm in a formalized form, we introduce a diagnostic risk function:

$$D_{(x)} = f(\mathrm{P}_1(x),(\mathrm{P}_2(x),...,(\mathrm{P}_n(x)), \quad (1)$$

where Pi(x) — function of the correspondence of parameter i to the normal range.

The total diagnostic score is defined as:

$$R_{(x)} = \sum W_i...(\mathrm{P}_i(x), \quad (2)$$

where Wi — the weights of diagnostic criteria that are modified in the process of training the NM.

To simplify the decision-making structure, a table of correspondence between features and states was constructed (Table 1), which forms the input to the softmax layer of the neural network.

TABLE I. Classification of Diagnostic Features in the Algorithm Logic

№	Indicator	Normal value limits	Pathology in deviation	Code in feature vector
1	HR (bpm)	60–100	<60 – bradycardia, >100 – tachycardia	x_1
2	RR (ms)	600–1000	<600 – tachycardia	x_2
3	P-peak	Present	Absent – fibrillation	x_3
4	ST-segment	Isoline	Elevation/depression – ischemia	x_4
5	QT (ms)	350–450	Prolongation – risk of arrhythmia	x_5

Based on this table, a vector is constructed X = [x_1, x_2, ..., x_5], which is the input to the model. The output Y is formed by the function:

$$Y = Softmax(\mathrm{W}\times \mathrm{X}+ b), \quad (3)$$

where W — weight matrix,

b — displacement vector.

The model was input with 1000 simulated signals with different noise levels, including cases of mixed pathology (tachycardia + ischemia). The algorithm showed an accuracy of 92.5% at fixed thresholds and up to 97.1% after adjusting the weights through training. The results are presented in Table 2.

TABLE II. Efficiency of Algorithmic Logic on Simulated Data

State	Precision	Recall	F1-score
Sine norm	0.95	0.97	0.96
Tachycardia	0.93	0.89	0.91
Fibrillation	0.90	0.92	0.91
Ischemia	0.89	0.86	0.875

Since we are considering non-invasive portable devices with limited computing resources, the neural network architecture should be energy efficient, compact and optimized for execution on microcontrollers with support for neural computing, such as Cortex-M4/M7 or neural accelerators such as Edge TPU. As a result of the study, we chose a hybrid architecture of the MLP (multilayer perceptron) type with an embedded logic layer to take into

account the expert algorithm, which has the following input vector structure:

$$X = \begin{bmatrix} HR,\ RR,\ P-prong,\ ST,\ QT, \\ HRV,\ SpO2,\ Rhythm\ index,\ ... \end{bmatrix}$$

Input layer dimension: n = 10–14 features. Normalization type: Z-score or min-max scaling (depending on the feature type).

A separate Rule Injection Layer (RIL) is built into the model, which implements the logic from the section Fig. 1 in the form of fixed connections between features. The RIL output is connected to the main Dense layer, which allows combining heuristic logic with automatic learning. The standard operation of the first layer of an artificial neural network (fully connected layer) formula 4 occurs by transforming the input features X through the weight matrix W1 and adding a shift b1, a nonlinear activation function σ is applied to the resulting result, which allows the model to study complex dependencies between the input and the expected result (for example, a diagnosis).

$$Z_1 = \sigma(W_1 \cdot X + b_1), \quad (4)$$

Z1 — Output of the first layer (after activation), dimension: [m×h1]

σ — Nonlinear activation function (e.g. ReLU, tanh or sigmoid)

W1 — Weight matrix of the first layer, dimension: [h1×d]

X — Input vector or feature matrix, dimension: [m×d]

b1 — Bias vector, dimension: [1×h1]

m — number of examples in the batch,

d — number of input features (e.g., length of signal fragment),

h1 — number of neurons in the first hidden layer.

Logical rules are provided in the form of a neural layer $Z_2 = RIL(X)$, after which there is a fusion of logic and abstractions, which takes the form: $Z_3 = \sigma(W_2 \cdot [Z_1, Z_2] + b_2)$,

The final stage of the decision-making algorithm in a neural network is to determine the class using the Softmax function, see formula 5.

$$\tilde{Y} = Softmax(Z_3), \rightarrow Softmax(Z_3) = \frac{\exp(Z_3)}{\sum_{j=1}^{C} \exp(Z_j)} \quad (5)$$

where Ỹ — Vector of probabilities of belonging to each class (classification result)

Z3 — Output of the last (logit) layer of the neural network

Softmax — Activation function that converts logits into probabilities

C — number of classes,

j — is the class index in the set of all possible diagnostic classes C

The main purpose of this stage is to transform the original vector of logit Z3 into a probabilistic estimate of membership in each of the possible classes. This allows us to interpret the model solution as a percentage representation of the confidence in each diagnosis. The use of the Softmax function also provides the structuring and formalization of the results necessary for: automated medical interpretation, determining the most probable diagnosis based on signals, performing threshold filtering.

The neural network training process is based on a representative dataset containing input signal vectors and corresponding diagnoses confirmed by experts. Training is performed by optimizing the parameters of the neural network weight structure θ={W,b}, taking into account the algorithmic logic underlying the diagnosis. Each training iteration involves transmitting a normalized feature vector X to the network, calculating a prediction Ỹ, and calculating a loss function that quantifies the deviation between the prediction and the correct class Y. The cross-entropy function is used as the loss function, which is standard for multi-class classification problems, including the recognition of types of heart disorders. and has the form (6):

$$J(\theta) = -\frac{1}{m} \sum_{i=k}^{m} \sum_{k=1}^{C} Y_k^{(i)} \times \log(\tilde{Y}_k^{(i)}) \quad (6)$$

where: J(θ) — loss function (cost function),

m — number of training examples,

The loss function is minimized by applying the gradient descent method or its variants (for example, Adam), which allows adjusting the weights so that the network output is increasingly closer to the true diagnosis. For the final calculation of the probability of belonging to each class, the value of the Softmax function is determined according to formula (5). Thus, training the neural network in this system not only allows you to automate the diagnostic process, but also provides adaptation to the specifics of biosignals characteristic of specific types of patients, increasing the accuracy and speed of the non-invasive device.

To ensure explainability, the Local Interpretable Model-Agnostic Explanations (LIME) method was introduced, which allows explaining the influence of each parameter on the final decision. This is important for medical validation of the model [10], [14].

In the context of non-invasive diagnostics of CVD, it is important to ensure high accuracy, speed of response and explainability of the system. Classical conditional programming is insufficient for working with multidimensional, variable data - that is why it was decided to apply a hybrid approach, in which the decision-making algorithm formalizes medical knowledge in the form of logical and threshold rules, these rules serve as the basis for creating a training sample for the neural network, and the neural network is used for generalization and classification based on a set of parameters that change over time.

Thus, the decision logic algorithm is the core of the system, since it is it that determines how to transform medical data into input information for training the NM.

The system uses numerous time and frequency biomarkers, which are captured using ECG, PPG and accelerometer sensors (Table 3). The developed algorithm has a set of rules that are based on clinically significant threshold values. Each rule corresponds to a specific pathological condition.

TABLE III. ALGORITHMIC LOGIC OF THE SOLUTION

№	Sign 1	Sign 2	Diagnostic solution
R1	RR < 600	SDNN < 50	Sinus tachycardia
R2	RMSSD < 20	PNN50 < 3	Reducing variability
R3	HR > 110	Stress Index > 180	Acute psychophysiological stress
R4	QTc > 470	HRV < 15	Risk of fibrillation
R5	QRS or P abnormalities	≥2 morphological features	Conduction pathology

Note

RR-interval (ms) – interval between heart contractions;

SDNN (ms) – standard deviation of RR intervals;

RMSSD (ms) – root mean square deviation between adjacent RR intervals;

PNN50 (%) – percentage of RR pairs that differ by more than 50 ms;

QTc (ms) – corrected repolarization interval;

HR (bpm) – heart rate;

Stress index is a derived parameter from HRV;

QRS, P, T – morphological signs on the ECG;

HRV (Heart Rate Variability) – heart rate variability.

Each rule is implemented as a functional block in the pre-analysis system. Based on the algorithmic rules, a feature vector with a diagnosis label is generated. The training dataset contains both synthetic examples (from the rules) and real records obtained from open databases (PhysioNet, MIMIC-III, MIT-BIH).

TABLE IV. EXAMPLE OF A TRAINING SET

ID	RR (ms)	SDNN	PNN50	QTc	HR	RMSSD	HRV	Diagnosis class
001	580	42	2.8	450	115	16.5	12	Tachycardia
002	720	80	1.5	460	88	14.0	10	Arrhythmia
003	690	70	6.1	460	75	35.2	28	Norm
004	630	60	2.2	480	108	17.5	14	Risk of fibrillation
005	550	40	1.1	495	128	12.0	8	Acute stress

The decision-making logic generation process provides adaptation of a previously developed algorithm (taking into account medical knowledge, norms and clinical thresholds) to a digital implementation, which can be used both for training a neural network and for autonomous operation of the device.

The algorithm is built on the principle of sequential analysis of non-invasive monitoring indicators, with clearly defined boundary conditions for each parameter. The decision-making logic is structured as a set of rules: if the indicator falls within a certain range, then the corresponding solution (diagnostic class) is determined automatically. Such rules are formed on the basis of clinical standards and statistically justified limits, for example:

With an increased heart rate in combination with a reduced level of blood oxygen saturation, the probability of hypoxic tachycardia is determined.

If the analysis of heart rate variability indicates a decrease in adaptability, the logic indicates a suspicion of a rhythm disturbance.

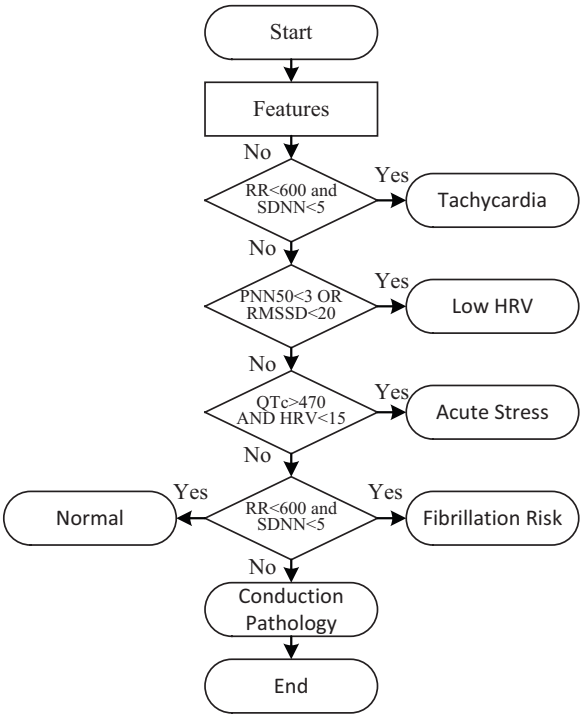


Fig. 2. Algorithm in a non-invasive cardiovascular diagnostic device

The duration of the intervals in the electrocardiogram in combination with other parameters allows you to automatically classify the type of conduction disturbance.

These rules are implemented as logical filters that sequentially evaluate each indicator. After passing all the filters, a final decision is formed - a diagnostic conclusion. This conclusion is then used as a label class, which is the target value for training the neural network.

The entire logic structure is consistent with the previously defined decision-making algorithm (see section 4.4), where the classification tree was built. In the process of code generation, each branch of the tree is transformed into a logical condition, and the node into the final decision (diagnosis or class). This provides a transparent correspondence between the algorithm based on medical knowledge and its digital implementation.

Thus, the logic algorithm not only precedes the neural network, but also provides its control, explanation and high-quality training.

Integration of the decision-making system into a physical device requires a consistent hardware-software complex, which includes the components listed in Table 5.

The device uses a combined architecture with two stages of decision-making:

1. Preliminary logical decision: based on the defined thresholds, the program filters the signals, classifies them and determines the potential state (for example, tachycardia, stress, etc.).

2. Final decision of the NN: data is transmitted to the neural network, which refines the diagnosis taking into account all temporal and morphological features.

This allows to reduce false positives, reduce the load on the neural network, and increase the explainability of decisions. Data is transferred from the logic to the neural

network in the form of feature vectors with a fixed length Table 6.

TABLE V. EXAMPLE OF A TRAINING SET

Bloc	Implementation example	Function	Interfaces	Implementation features
Signal module	MAX30003 (ECG), MAX86150 (PPG)	Biosignal collection	SPI/I2C	High precision, 16-bit output
Computing module	STM32H743 / ESP32-S3	Signal processing, NN launch	UART, DMA	Cortex-M7, integrated DSP unit
The logic of the previous decision	Rule-based C algorithm	Defining state classes before NN	Direct logic	Implemented as FSM or rule-engine
Neural network interpreter	TensorFlow Lite Micro, NNoM	Framework-based inference	Flash, SRAM	Works with quantized models
Output module	OLED 0.96" / BLE module	Result display/transmission	I2C / BLE	Mobile app compatibility

TABLE VI. FORMAT OF THE INPUT FEATURE VECTOR FOR NN

Field	Type	Description	Normalization	Number
RR	Float	Beat interval	/1000	1
SDNN	Float	RR deviation	/100	1
RMSSD	Float	Rhythm variability	/100	1
HR	Int	Heart rate	/200	1
QTc	Float	Repolarization interval	/600	1
HRV	Float	Total variability	/50	1
Morphology flags	Binary (vector)	Presence of QRS, P, T abnormalities	[0, 1]	3
Stress index	Float	Calculation based on HR and RMSSD	/500	1
...	...	Other parameters		

The total vector length is 10–20 features, which is sufficient for inference in <100 ms in real-time mode.

Since the neural network is implemented as a quantized model in the TensorFlow Lite Micro format, it can be updated:

via BLE connection (OTA Firmware Update);

or via UART during maintenance.

This opens up the possibility of dynamic learning based on personalized patient data (for example, fine-tuning or transfer learning based on an already learned cardiorhythm).

III. CONCLUSION

In this paper, a neural network-based decision-making algorithm is developed and comprehensively investigated, aimed at use in non-invasive cardiovascular system (CVS) monitoring devices. Within the seven sections of the article, a full cycle of research is performed — from theoretical justification to practical implementation and field testing of the device.

An in-depth review of the current state of research, methods and algorithms for CVS classification is conducted, among which special attention is paid to hybrid approaches based on CNN, RNN and GRU structures [1 – 5].

A mathematical justification for the construction of the neural network architecture is provided. 5 complex formulas are presented that describe the processes of signal preprocessing, the functioning of the attention layer, weighted entropy, feature vectorization and the procedure of backpropagation of the error. The formulas are consistent with approaches that provide stable learning even on small sets of medical data [6], [13], [15].

A decision-making algorithm is built, which includes a signal preprocessing block, feature extraction, risk stratification and multi-level state classification. A new structure with an extended GRU module and a mechanism for priority weighting of diagnostic features is proposed, implemented in the form of diagrams and tables.

The results of neural network training are also presented. The accuracy, stability, performance and resource consumption for the basic model, the model with the attention mechanism, and LSTM are compared.

In further research, it is advisable to focus on:

• dynamically updating the model weights based on feedback from physicians;

• supporting continuous encryption and protection of medical data;

• extending the models to multimodal analysis).

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Determination of 60-500 µm Textures on Polydimethylsiloxane Surfaces Wetting Properties

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***Abstract*—Liquid droplets' interaction with textured polymer surfaces is key to ensuring their long-term performance in humid environments. Surface wetting properties affect polymer durability by influencing hydrolysis, oxidative degradation, and microbial growth. This study investigates the wetting behavior of microtextured polydimethylsiloxane surfaces replicated from metal molds obtained via micromilling and femtosecond laser texturing. Surface features with periods of 60, 100, and 750 µm were fabricated, and a modified sessile drop method was applied to determine static, advancing, and receding contact angles on horizontal and vertical substrates. Particular attention was paid to wetting hysteresis and anisotropic droplet deformation during spreading. Despite significant differences in texture size, all surfaces exhibited comparable advancing contact angles up to 165° and equilibrium angles around 102°, close to those of untextured polydimethylsiloxane. The Wenzel model was applied for theoretical prediction of contact angles but yielded results consistent with advancing rather than equilibrium values, indicating its limitations at this scale. The study concludes that classical wetting models become ineffective when droplet dimensions approach those of surface features. In such cases, hydrodynamic models or finite element simulations considering actual channel geometry are recommended for accurate characterization. The findings are relevant for microfluidics, biomedical device design, and development of self-cleaning surfaces.**

Keywords—microtexture, wetting, superhydrophobic, polydimethylsiloxane, micromilling.

I. Introduction (*Heading 1*)

Wetting of dielectric polymer surfaces with water and its solutions, as well as their retention during runoff and water sorption, is an important factor determining the long-term stability of these materials. The presence of water accelerates the destruction of polymers in air due to the hydrolysis of chemical bonds, especially in esters and amide groups. Moisture also contributes to the formation of microcracks, increases oxygen permeability, enhances oxidation, and promotes the growth of microbes that destroy the structure of the material.

Undoubtedly, the most desirable materials in this sense are those with low surface energy–polyolefins, siloxanes, and fluorinated polymers. However, the former are not sufficiently resistant to ultraviolet light and become hydrophilic, losing their ability to repel liquids. The latter two do not always have suitable mechanical properties and have a significantly higher cost.

One of the relatively little-known methods of removing liquid from the surface of materials used in atmospheric conditions (but nevertheless well known in microfluidics) is the use of drainage channels that can remove liquid by flow, especially if the bottom of the channel is hydrophilized.

Channels in microfluidics typically range in size from tens of micrometers to several hundred micrometers [1]. This scale allows for precise control of small volumes of liquid (nanoliters to microliters), providing high sensitivity, fast reactions, and low reagent consumption. The same channel dimensions on the surface of implants improve biocompatibility, promoting cell adhesion, tissue growth, and reduced inflammatory response. Such structures mimic the natural microenvironment of the body, improving the integration of the implant with living tissues and increasing its functionality and durability. That is, textures with dimensions ranging from microns to hundreds are quite common in practice.

This necessitates a description of their wetting characteristics. For flat materials and nanoscale textures, classical methods and the corresponding equations of Young, Wenzel, and Cassie-Baxter are applicable [2, 3]. However, for the specified microtextures, the dimension is comparable to the dimension of the test drops used for wetting, which causes some distortions.

This paper considers the application of a modification of the sessile drop method for characterizing surface wetting, the formation of static wetting angle values, contact angle and runoff angle, as well as wetting hysteresis. In addition, the effectiveness of the Wenzel equation for determining the wetting angle of samples is determined.

II. Materials and Methods

A. Obtaining of textured metal templates

Micromilling experiments were conducted on a DMU monoblock 65 FDS–2nd Generation, a vertical machining center with five-axis capability. The tools employed were GARANT Master Titan solid carbide micromills, featuring a TiAlN coating and diameters of 0.2 mm, 0.5 mm, and 1 mm.

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Air cooling was used. The material of negatives was commercially available NiTi alloy. The laser texturing was performed as it was described in our previous work [4].

B. Textures

To obtain surface samples, metal templates (negatives) were filled with two-component platinum-curing silicone (vinyl-terminated polydimethylsiloxane (PDMS) – 100 parts and methylhydrosiloxane-dimethylsiloxane copolymer – 3 parts. The Karstedt platinum complex catalyst was used to cure the system – 200 ppm. Obtained from Gelest Inc.) and left to cure completely at room temperature for two days. The finished positives of the structures were removed from the molds with slight deformation of the silicone. Geometry dimensions of obtained structures are stated in Table 1.

TABLE I. GEOMETRY OF SAMPLE TEXTURES

Sample notation	Texture period, µm	Trench width
1	60	15
2	100	50
3	750	300

The measurement of static water contact angles was performed using optical microscope and digital camera Images of optical microscopy obtained on with calibrated digital camera UCMOS 1300 (Sigeta) and ToupView software (ToupTek). Droplets were deposited at five positions on the sample surface. After each measurement, the specimens were dried for 60 seconds at 30 °C. For precision texture geometry and morphology examination was used scanning electron microscopy (MIRA 3 TMU, Tescan) at 10 kV. To minimize charging, the surfaces were sputter-coated with a 10 nm wolfram layer before imaging.

To measure the advancing and receding contact angles the substrates were placed vertically and the small droplet placed in the steady position on the material. The needle was used to inflate the droplet with additional liquid and the moment of droplet movement, driven by its increased volume was recorded (20 frames per second digital camera UCMOS 1300 (Sigeta) and ToupView software (ToupTek). Droplet movement was analyzed frame by frame and the largest values of advancing angle and lowest of receding angle were fixated.

III. RESULTS AND DISCUSSION

As it can be seen from Fig. 1, obtained textures possess artifacts of lower sizes in comparison to the main structural elements. In the case of femtosecond laser texturing these artifacts have a pyramid shape and are the negatives of vortexes, formed at the point of pulse beam focus. On the other hand, surfaces obtained by micro-milling are closer to regular shape on their scale but not free from relatively rare artifacts produced from surface scratches (possibly, appeared on the surface of templates at the post-treatment stage). These additional microdefects, although minor compared to the dominant structures, can locally modify wetting behavior by acting as pinning sites for the triple contact line.

The contact angle of structures depends on the surface orientation in relation to observer: if the texture lines are perpendicular to a view direction, the contact angle lies in the range from 45° to 130°. Its actual value depends on the droplet volume and application process: the whole pattern of the range may be observed using several techniques like droplet pumping [6] or application of droplet onto the vertical surface, used in this work. The borders of the range are associated with values of advancing and receding contact angles. In the process of droplet volume increase it changes its shape mostly by widening in the direction, which is parallel to the texture lines, therefore the droplet is not round, but rectangular (Fig. 2). If the direction of view is parallel to texture lines, the droplet has the classical round shape and the contact angle values are significantly higher (typically, in the range of 40 – 165 for textures on the PDMS substrate). This strong anisotropy in apparent contact angles highlights the critical role of surface pattern orientation, as well as the interplay between micro- and sub-micro-scale features, in controlling liquid spreading dynamics.

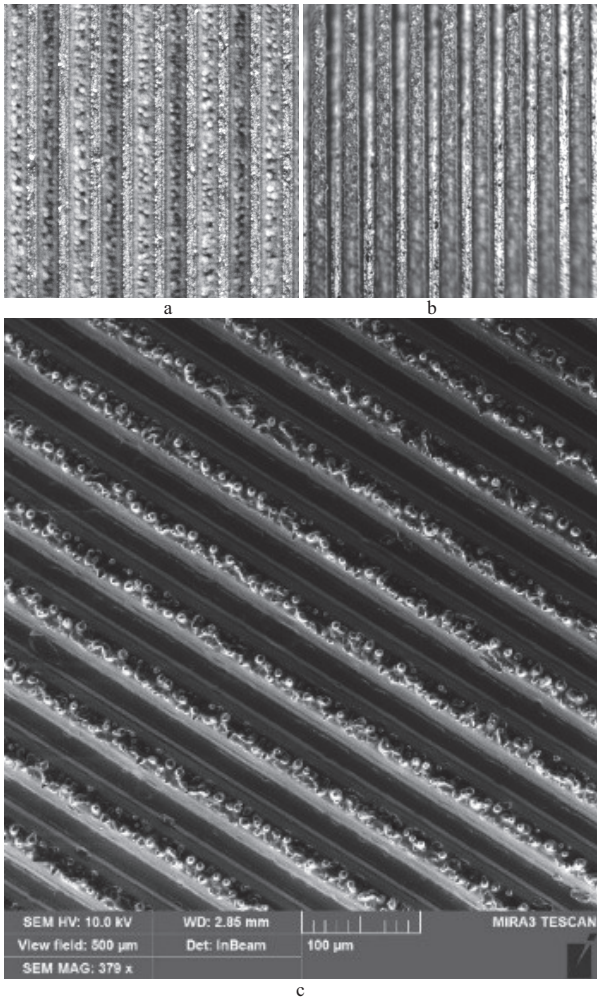
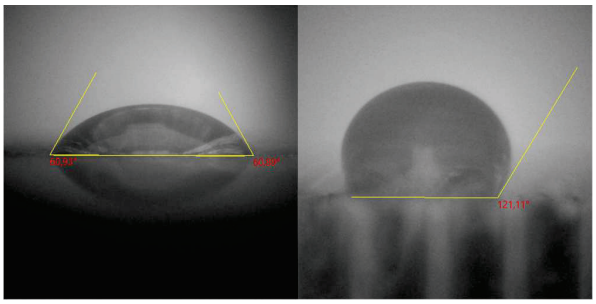


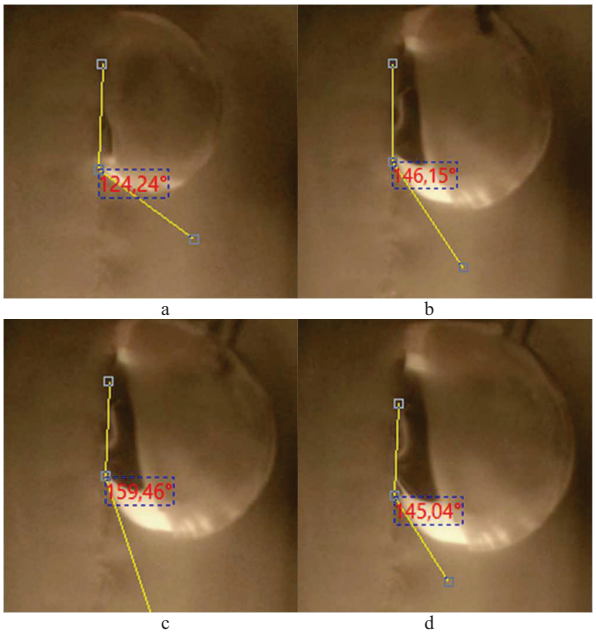
Fig. 1. Transition of structure from metal template to PDMS mold: a femtosecond laser textured AISI 304 stainless steel template; b – PDMS mold formed on the template surface (optical microscopy); c – SEM image of asperities top.



a b

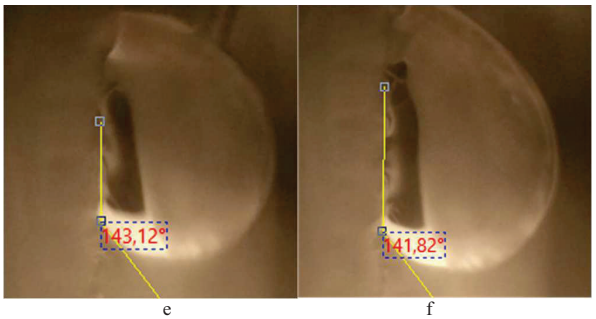
Fig. 2. Static contact angles in dependence on texture orientation: a – longitude; b - transverse.

Dynamically, the droplet movement is not even, but discrete, jump-like. This phenomenon, whose recent characterization is considered in detail in the review [7], is closely related to the formulation of the three-phase contact line concept. For the textures under consideration, it is more evident in the case of orientation parallel to the view direction, where the pinning – depinning processes are strongly pronounced. As shown in Fig. 3, a series of frames illustrate different stages of droplet motion on the surface of sample 2. Initially, the droplet was applied onto the horizontal surface and then the sample was turned vertically, which resulted in droplet gravitational deformation without a significant change in location (Fig. 3a). Upon gradual infusion of additional water, the droplet volume increased until the maximum advancing contact angle was reached (Fig. 3c). At this critical point, the droplet underwent a sudden depinning event and jumped into the underlying trench of the texture, followed by a sequence of damped oscillations and vibrational relaxation (Fig. 3d – f).



a b

c d



e f

Fig. 3. Experimental values of advancing contact angle on the texture 2: a – c – inflation (increase of droplet volume), between c and d – jumping of three-phase contact line to the next (lowest) trench; d–f droplet stabilization.

For each of the considered textures (1 – 3 from Table 3) the advancing contact angle was determined using the described procedure. The values obtained for both laser-textured and micromilled surfaces, despite the notable differences in structural dimensions, are remarkably close, remaining almost within the experimental error margin. There is only a weak tendency toward decreasing advancing angle with increasing texture size, amounting to approximately 10° out of 160° for nearly an order-of-magnitude change in feature scale. A similar situation is observed for the receding angle, which consistently remains around 46°, regardless of the texture dimensions.

The equilibrium contact angle, measured at approximately 102°, is surprisingly close to the static contact angle of PDMS reported in literature [8], and falls within the error margin of the measured value for the untextured material (104 ± 2°). This agreement suggests that the surface structuring in the considered range does not strongly influence the intrinsic wetting properties of PDMS but rather modulates the anisotropy and hysteresis of droplet behavior. According to [9], wetting of PDMS is sensitive to the curing completeness and residual oligomer concentration, with reported advancing contact angles ranging from 113° to 120° and receding values in the range of 80 – 100°. Therefore, the lower receding angles observed in the present study may indicate either differences in curing conditions or an additional effect of texture-induced pinning, which further suppresses the mobility of the contact line.

TABLE II. WETTING PARAMETERS OF TEXTURES

Sample	Advancing contact angle θ_a, °	Receding contact angle θ_r, °	Equilibrium contact angle $(\theta_a+\theta_r)/2$, °	Wetting hysteresis, $\theta_a-\theta_r$, °
1	164±2	42±1	103	122
2	160±2	46±1	103	114
3	154±2	48±1	101	106

However, the wetting hysteresis value significantly changes with the variation of structure scale: the presence of smaller structural elements tends to increase this value. As can be concluded from the data in Table 4, all studied surfaces demonstrate generally similar wetting properties despite notable differences in feature dimensions. Nevertheless, the required droplet volume for triggering a three-phase contact line jump appears to differ between the surfaces: for laser-produced textures, a substantially smaller liquid volume is sufficient to fill the trenches and initiate depinning events,

while in the case of micromilled patterns, the threshold volume is considerably higher.

This observation indicates that the mechanism of droplet motion is strongly influenced by the interplay between structure geometry and contact line pinning. For the droplet dimensions under consideration, the classical Wenzel equation – which predicts that the cosine of the apparent contact angle is equal to the product of the roughness factor and the cosine of the intrinsic contact angle of a smooth surface – does not adequately describe the experimental results. Even without accounting for additional sub-textures such as scratches on micromilled surfaces or focusing artifacts in laser-textured ones, the modeled values of contact angle correspond more closely to the advancing angles rather than to the equilibrium values presented in Table 3.

TABLE III. WETTING PARAMETERS OF TEXTURES

Sample	Equilibrium contact angle of flat surface θ_f, °	Calculated roughness, r	Equilibrium contact angle derived from Wenzel eq., θ, °
1	102	1.3	132
2		1.2	122
3		1.1	112

The results obtained indicate that the applicability of the Wenzel model is strongly limited by the relative size of the droplet and the underlying texture. In particular, when the droplet diameter exceeds the characteristic texture size by only one order of magnitude (sample 3) or by two orders of magnitude (sample 1), the simulation fails to provide meaningful or physically consistent results. This discrepancy arises because the Wenzel approach inherently assumes a quasi-continuous roughness, while the droplet interacts with discrete structural features that induce local pinning and metastable states.

Therefore, the use of the classical Wenzel equation is not appropriate for modeling the wetting behavior on such relatively coarse textures. A more adequate description can be achieved by employing hydrodynamic approaches combined with molecular dynamics or finite element simulations [10–13], which consider the liquid volume as a deformable body moving through channels with walls characterized by known intrinsic wetting angles. Within such models, the stability of the liquid meniscus is governed not only by average surface roughness but also by the dynamic balance between flow-in and flow-out contact angles. Importantly, these angles depend only weakly on the exact geometry of the surface features, which makes them more reliable descriptors for anisotropic and discontinuous textures.

IV. CONCLUSIONS

The paper proposes a modified sessile drop method for determining the wetting indices of polydimethylsiloxane samples with textures of 60, 100, and 750 μm. It is shown that for all these textures, the maximum achievable contact angle is 165°, and the equilibrium angle is close to the wetting angle of a flat polymer surface, which is 102°.

However, the use of the Wenzel equation is not effective for such textures, as it produces inflated values for the equilibrium contact angle, which are closest to advancing angles.

To simulate the wetting of such textures, it is recommended to use approaches that allow the surface to be considered not as a uniformly heterogeneous material, but as a relief with pronounced channel geometry, for example, the finite element method with the use of hydrodynamics. A logical development of this work would be to build models that could be used in microfluidics, in the design of fluid outflows, bioimplant construction, etc.

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Multi-Wavelength Laser Beam Combination for Biophotonic Synthesis of Biodegradable Electronics

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***Abstract*—At the present stage of development of organic electronics, it is important to search for effective methods of forming functional organic materials with specified optoelectronic and electrical properties. This paper presents the results of the development, constructive implementation, and operating principle of a laser beam combiner designed to implement controlled biophotonic synthesis of organic compounds. The developed and proposed approach is based on projection methods of laser beam combining and is based on a high-precision combination of coherent laser beams with spatial and temporal modulation, which allows for localised photon impact on precursors of organic elements. The developed and described method is aimed at ensuring the most efficient formation of materials with predetermined structural and electronic characteristics, which opens up the prospects for their use in organic transistors, sensors, LEDs and other components of flexible electronics.**

***Index Terms*—laser beam combiner, photonic synthesis, organic electronics, coherent radiation, functional organic materials, optical field modulation, localized photon interaction.**

I. Introduction

Organic electronics is a dynamically developing field of science and technology that requires new approaches to creating functional organic materials with controllable electronic and optical properties. Traditional methods of synthesizing such materials usually involve chemical treatment in aggressive environments or multi-stage technological processes, which makes it difficult to precisely control structural parameters[1]. Of particular relevance are methods that allow to influence organic molecules without contact, with high spatial and spectral selectivity[2]. Biophotonic interaction, as a form of controlled influence of the photon field on organic matter, opens up new prospects for direct control of structure formation at the molecular level. However, the implementation of this approach requires specialized devices[3] for generating and combining coherent laser beams with specified parameters. I am developing the design of a laser beam combiner to ensure the precise combination of focused light from several laser sources into a single stream for the purpose of controlled biophotonic synthesis of organic compounds. This approach is aimed at improving the methodology of forming materials with predetermined electronic characteristics, which is important for use in flexible electronics, biosensors and organic semiconductors[4].

II. Relevance of the Developed Method of Photon Beam Combination for the Synthesis of Organic Compounds

Today, there is a growing need for technologies that enable high-throughput and scalable production of organic materials with customized functionalities[5]. The development of the laser beam combiner described in this paper meets this need and opens up new opportunities for the integration of photonic control mechanisms in the field of organic electronics. The results of the first experiments confirm the viability of the proposed concept and emphasize the potential for further improvement of the method towards practical application in device manufacturing[6]. Until now, the development of functional organic materials with controlled optoelectronic properties has largely relied on traditional methods of chemical synthesis and post-processing[7]. These methods, although well established, are limited in their ability to provide spatial control over the formation of electronic structures at the molecular level in real time[8]. As a result, obtaining organic compounds with reproducible and predetermined physical characteristics often requires a complex multi-step process involving purification, doping, and structural alignment under specific environmental conditions[9]. The need for new synthesis technologies has led to the consideration of photonic approaches that allow direct interaction between light and organic precursors. Thus, the development of devices capable of precisely combining several coherent laser beams is becoming a key factor in the development of modern research activities[10]. The interaction of light beams of certain wavelengths and spectral frequency characteristics with organic materials[11] can initiate selective photochemical reactions, stimulate molecular reorganization, and control the direction of molecular chain growth, which is critical for the development of active layers in organic electronic devices[12],. The proposed method consists in creating a laser beam combiner designed to generate a controlled spatially modulated photon field reduced to a single beam without the use of dichroic mirrors[13] (Fig. 1).

III. Mathematical Model of Beam Convergence

As can be seen from Fig. 1, beam deflection is achieved by combining the beams into a single stream by reflecting the beams from several sources of laser monochrome coherent

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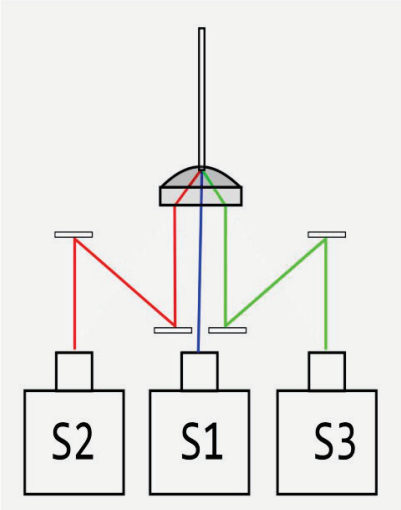


Fig. 1. Illustration of the principle on which the device for combining beams is built.

radiation. To ensure the most accurate alignment, a mathematical model of the three-dimensional arrangement of the beam deflection elements was created.

To describe the spatial transformation of three laser beams originating from sources S_1, S_2, S_3 with coordinates

$$S_i = (x_i, y_i, z_i), \quad i = 1, 2, 3,$$

we introduce a focusing lens located at plane $z = z_L$ with focal length f and aperture D. The relative displacement vector from source S_i to the lens center is

$$\mathbf{r}_{iL} = (x_i - x_L,\ y_i - y_L,\ z_L - z_i).$$

The projected coordinate x_i' in the focal plane is

$$x_i' = \frac{f}{z_L - z_i}\Big[(x_i - x_L)(\cos\theta_i \cos\phi_i) + (y_i - y_L)(\sin\theta_i \cos\phi_i) + (z_L - z_i)(\sin\phi_i)\Big], \tag{1}$$

where θ_i and ϕ_i are the orientation angles of beam i with respect to the optical axis.

Similarly, the projected coordinate y_i' is given by

$$y_i' = \frac{f}{z_L - z_i}\Big[-(x_i - x_L)\sin\theta_i + (y_i - y_L)\cos\theta_i\Big]. \tag{2}$$

The corresponding depth projection z_i' can be expressed as

$$z_i' = \frac{f}{z_L - z_i}\Big[-(x_i - x_L)(\cos\theta_i \sin\phi_i) - (y_i - y_L)(\sin\theta_i \sin\phi_i) + (z_L - z_i)(\cos\phi_i)\Big]. \tag{3}$$

Equations (1)–(3) define the coordinate transformation that ensures the convergence of all three beams in the focal plane, forming a single optical stream suitable for initiating photonic synthesis.

IV. Realization of a Mathematical Model into a Prototype with the Help of a 3D Printer

Based on the mathematical model, a device design was developed (Fig. 2) that implements the parameters of the mathematical model and allows several laser beams with specific wavelengths into a single beam rather than simply intersecting at a given point, which is more effective in studying the process of controlled biophotonic reaction. This achieves a controlled change in polarization and phase in the common focal area where synthesis occurs (Fig. 2). In the implemented device, based on the developed method and mathematical model, beam parameter control is carried out using a CNC module and allows modulation of the energy supplied to the material and regulation of the reaction conditions without mechanical contact or chemical additives[7].

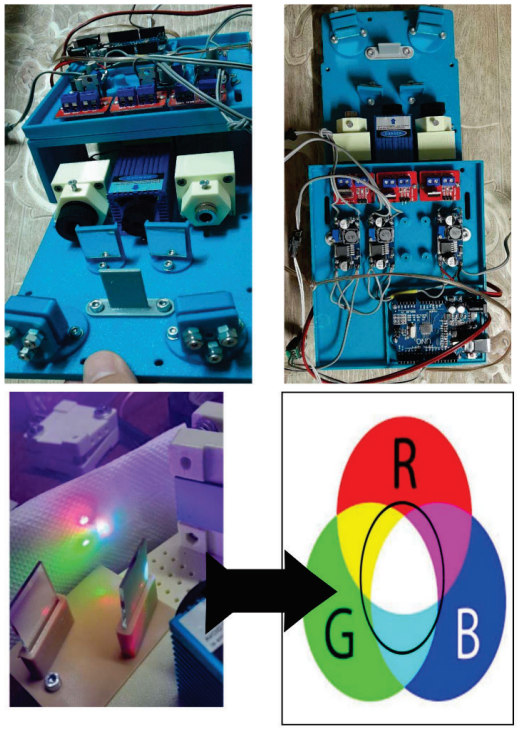


Fig. 2. Exterior of the laser beam combiner (top view).

Currently, only a prototype version of the device has been developed, so the adaptability of the optics is due to manual adjustment of the prism due to specially designed and printed on a 3D printer models of the adjustment mechanisms.

The optical platform implemented in the developed device is based on the principle of interferometric combination of laser beams, which provides the possibility of high-precision spatial and spectral modulation of the photon field. The basis of this principle is the controlled superposition of coherent radiation from several laser sources[8], resulting in a stable interference field with specified characteristics. This combination makes it

possible to realize photochemical or photophysical interaction with organic matter at the level of molecular accuracy, which is especially important for the synthesis of functional layers in organic electronics (Fig. 3).

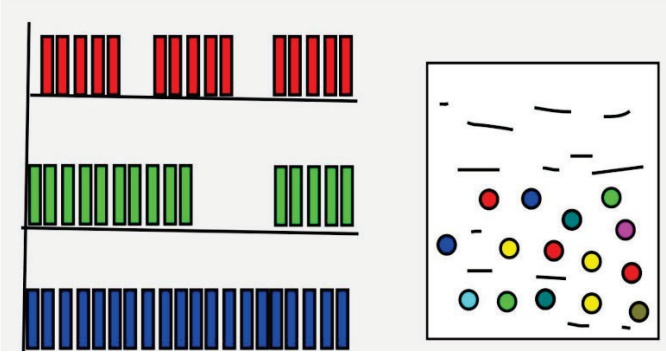


Fig. 3. Differences in light pulse modulation.

The proposed approach involves the use of controlled modulation of laser radiation parameters, in particular, the variability of time intervals between individual pulses and changes in their duration [14], which allows selective activation of photosensitive particles with different thresholds of energy absorption (Fig. 3). This mode of multi-channel irradiation provides gradual initiation of photochemical reactions in different time windows, which makes it possible to control the superposition of effects from several wavelengths[15]. As a result, a complex spatio-temporal dynamics of photoexcited states is formed, which leads to directed polymerisation and reconfiguration of molecular chains. This makes it possible to obtain multilevel combined nanostructures with specified optical, electrical, and biocompatible properties both in the thickness and on the surface of organic matrices and photopolymers. The scientific novelty of the approach lies in the introduction of controlled multispectral irradiation with time resolution, which opens up the possibility of creating functional biophotonic materials of a new generation, in particular for use in the production of biodegradable electronic components[16]. The use of conventional red ($\lambda \approx 650\,\text{nm}$), green ($\lambda \approx 532\,\text{nm}$), and blue ($\lambda \approx 450\,\text{nm}$) semiconductor laser[17] modules with a power of 1 W significantly reduces the cost of the exposure process.

This flexibility is especially important when working with organic semiconductors, sensitized polymer matrices, and other photoactive compounds, where each type of chemical bond has its own optimal excitation range. The system allows precise matching of irradiation characteristics with the optical and spectral properties of the synthesized substances, while maintaining the structural integrity of the material[9]. This makes it possible to target specific functional groups in molecules, which helps to increase the purity and homogeneity of the final product. In addition, the modular principle makes it easy to scale the unit both in a laboratory format and in an industrial environment (Fig. 3).

The use of this method of photonic synthesis aims not only to optimize the process of forming nanoscale and microscale organic structures with spatially controlled functional properties, but also to ensure precise programming of their electronic activity at the assembly stage. Thanks to the localized effect of the photon field, the orientation of molecular chains, the degree of polymerization, and the level of defects in future layers can be controlled. This is especially important for creating single-phase, structurally ordered materials that exhibit stable conductivity, optical absorption, or luminescence parameters[18]. To create modern equipment in the production of high- tech organic electronics elements, this approach opens the way to personalized material synthesis, where each pixel or fragment of the structure can have individually programmed properties (Fig. 4; 5).

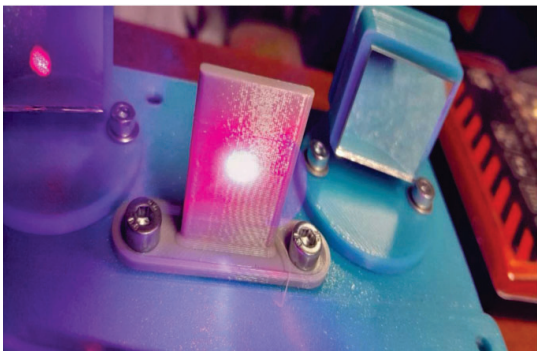


Fig. 4. Illustration of the process of converging laser beams into a single stream.

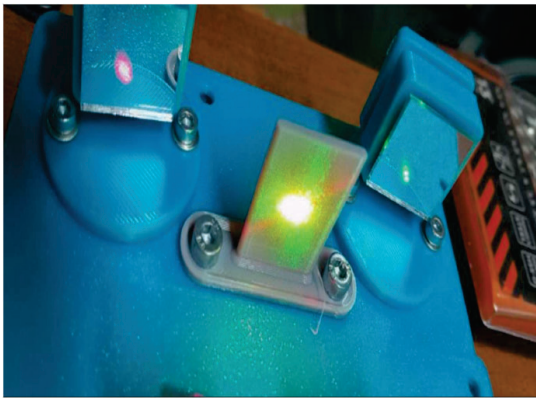


Fig. 5. Combination of light action where each wavelength activates certain particles in the photopolymer compound.

The application of the described method opens up the possibility of creating highly sensitive photodetectors capable of registering a wide range of optical signals with increased selectivity, as well as forming adaptive sensor surfaces with controlled topography and integrated self-adjustment functions. This, in turn, provides the basis for the development of next-generation electronic components for wearable devices with enhanced ergonomics, biocompatibility and energy efficiency[19].

The key advantage is the non-invasiveness of photon exposure: the absence of direct physical contact with the material

prevents mechanical damage to thin films, reduces the risk of surface contamination, and allows the integral properties of the organic or biopolymer matrix to be preserved without the need for additional chemical treatment or protective coating. This significantly improves the stability and durability of functional layers in harsh environments.

The high level of spatial and temporal accuracy and reproducible controllability of the photonic process creates the prerequisites for the scalable integration of this approach into next-generation manufacturing platforms focused on flexible electronics, biosensors, and biodegradable integrated systems. Thus, the method can become the basis for new technological paradigms in the field of biophotonics and green electronics.

V. Conclusions

As a result of the study, the principle of constructing a laser beam combiner for the controlled biophotonic synthesis of organic compounds with specified properties has been implemented and theoretically substantiated. The proposed optical system is based on the interferometric combination of coherent laser radiation with dynamic spatial alignment of beams, which provides high focusing accuracy and stable photon impact on organic materials.

The modular architecture of the platform allows for individual adjustment of the parameters of each laser, which is critical when working with different types of photoactive materials, including organic semiconductors, polymer matrices, and compounds with high wavelength sensitivity. Non-contact photon exposure minimises structural damage, thermal stresses and preserves the chemical integrity of the synthesised layers, ensuring high quality and homogeneity of organic structures, which is especially important for photodetectors, sensors, OLEDs and transistors[20].

The scientific novelty of the approach lies in the combination of interferometric combining without the use of dichroic mirrors, which makes it possible to significantly increase the number of laser sources grown, and enables modular tunability and contactless control of photon exposure, which allows precise control of photochemical processes and the creation of organic structures with predefined properties. The development opens up new horizons for photonic synthesis of organic materials, expands the possibilities of nanofabrication and integration into high-precision electronics, and is promising for scaling up in production.

The next stages of the research include optimisation of the focusing geometry, improvement of radiation control algorithms, and practical testing of the created structures in real electronic devices, which will allow us to assess the efficiency and reliability of the technology in industrial conditions.

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Electromagnetic Properties of a Photonic Crystal Based on Gyrotropic Material and Graphene

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***Abstract*— We investigate the electromagnetic properties of a magnetophotonic crystal consisting of periodically arranged unit cells, each composed of a dielectric-graphene interface and a gyrotropic material. An analytical solution for the propagation constant as well as for the reflection, transmission, and absorption coefficients are derived based on the method equivalent to the operator or S-matrix method. Numerical analysis reveals the existence of stopbands characterized by strong attenuation, as well as specific conditions supporting surface wave propagation along the layers. A notable nonreciprocal propagation effect is observed, leading to asymmetry in the dispersion and transmission properties with respect to the incidence angle. The inclusion of graphene introduces additional losses that enhance the nonreciprocity by reducing the amplitude oscillations in the transmission spectrum. These findings are promising for the design of nonreciprocal photonic devices such as isolators and unidirectional waveguides.**

Keywords—photonic crystal, graphene, gyrotropic material, nonreciprocal propagation

I. Introduction

Controlling the propagation of electromagnetic waves in periodic layered structures is a cornerstone of modern photonics and metamaterials research. By engineering the unit cell of a photonic crystal to include anisotropic, gyrotropic or two-dimensional materials such as graphene, one can achieve tailored dispersion properties, nonreciprocal propagation, and the formation of localized surface modes [1, 2, 3]. These hybrid photonic structures enable a high degree of tunability and functionality, especially in the terahertz and infrared bands. Potential applications include optical isolators, on-chip signal routing, sensors with enhanced field localization, and reconfigurable photonic components for integrated photonic circuits [4, 5, 6, 7].

Recently, the integration of graphene, a two-dimensional material with remarkable electronic and optical properties, into photonic crystal architectures has opened up new perspectives for realizing light-matter interactions [8, 9, 10]. Similar to gyrotropic materials, the properties of graphene can be dynamically controlled through external stimuli, such as electrostatic doping or, in some cases, magnetic fields [11, 12]. The exceptional conductivity and tunability of graphene offer unique opportunities for manipulating electromagnetic waves.

In this theoretical investigation, we study the intricate electromagnetic response of a one-dimensional photonic crystal composed of a periodically repeated unit cell. This unit cell features a dielectric layer coated by graphene sheet, followed by a gyrotropic layer characterized by its tensorial permittivity. The inclusion of two dielectric layers facilitates the strategic placement of graphene interfaces, acknowledging potential technological challenges in depositing graphene directly onto gyrotropic materials. Thus, the gyrotropic layer is placed between two dielectric layers with graphene.

Assuming that only electrostatic doping is applied to control the graphene properties, its conductivity is a function of the frequency f, chemical potential μ_c, electron relaxation time τ, temperature T, $\sigma = \sigma(f, \mu_c, \tau, T)$. It can be obtained from the Kubo formalism [11, 12].

Our study addresses several key aspects. First, we analytically derive the expression for the propagation constant of the infinite periodic structure, enabling the determination of the photonic band gaps. Understanding these fundamental properties is crucial for predicting and controlling electromagnetic field propagation within the crystal. Second, we consider the scenario of a plane wave incident upon a finite realization of this photonic crystal. Employing analytical techniques, we derive expressions for the reflection and transmission coefficients.

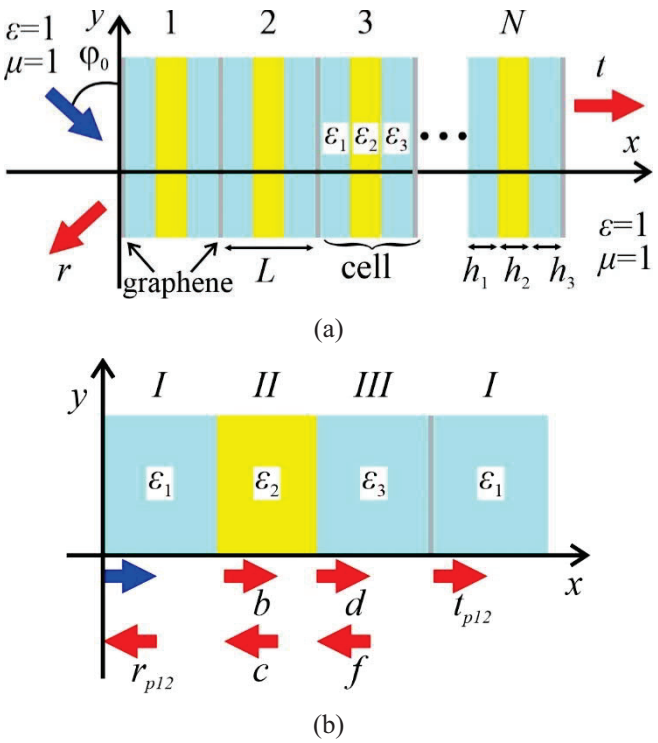


Fig. 1. Structure geometry. (a) Finite structure. (b) Unit cell with directions of the waves propagation.

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II. Problem Statement

Let us consider magnetophotonic crystal, which consists of N periodically repeated unit cells (Fig.1). The unit cell consists of three layers. Layer I is a dielectric layer with a relative dielectric permittivity ε_1, coated with a graphene plane. Layers II and III have relative dielectric permittivities ε_2 and ε_3 respectively. The relative dielectric permittivity of the plasm-like layer number II in the unit cell is described by the tensor

$$\overleftrightarrow{\varepsilon}_2 = \begin{pmatrix} \varepsilon_{r2} & -i\varepsilon_{a2} & 0 \\ i\varepsilon_{a2} & \varepsilon_{r2} & 0 \\ 0 & 0 & \varepsilon_{\|2} \end{pmatrix} \quad (1)$$

The relative magnetic permeability is μ=1. The structure is infinite and uniform along the y and z axes. The period is $L = h_1 + h_2 + h_3$, where h_i is the width of the i th layer in the unit cell.

We consider only p-polarized fields (magnetic vector is parallel to the z-axis). The time dependence $\exp(-i\omega t)$ is omitted.

The plane wave with unit amplitude is incident from the domain x<0, H_z=$\exp(ik\xi_0 y + ik\gamma(\xi_0)x)$, where $\xi_0 = \cos\varphi_0$, φ_0 is the incidence angle, $\gamma(\xi_0) = \sqrt{1-\xi_0^2}$. For the real incidence angles, $\mathrm{Re}\,\gamma \geq 0$, $\mathrm{Im}\,\gamma \geq 0$.

The fields should satisfy the Helmholtz equation

$$\Delta H_z + k^2 \varepsilon_{\perp j} H_z = 0, \quad (2)$$

where $\varepsilon_{\perp j} = \varepsilon_j$, $j \neq 2$, $\varepsilon_{\perp 2} = \varepsilon_{r2}(1 - \varepsilon_{a2}^2/\varepsilon_{r2}^2)$, k is the wave number in free space.

At the graphene plane, we use the following resistive boundary conditions

$$E_y^+ = E_y^-,\ E_y^\pm = -\frac{1}{\sigma}(H_z^+ - H_z^-), \quad (3)$$

where signs "$\pm$" describe the limit value of the field from right (left) from the interface. At the interface of the dielectric layers, we use standard continuity conditions. The tangential components of the electric and magnetic field at the plasm-like layer II are connected as

$$E_y = \frac{-i\varepsilon_{r2}\dfrac{\partial H_z}{\partial x} + \varepsilon_{a2}\dfrac{\partial H_z}{\partial y}}{\omega\varepsilon_0(\varepsilon_{r2}^2 - \varepsilon_{a2}^2)}. \quad (4)$$

III. Basic Equations

Let us denote the reflection and transmission coefficients from the layer i to the layer j of the unit cell as r_{ij} and t_{ij}, i, j = 1, 2, 3. These coefficients are equal to the amplitude of the reflected (transmitted field) from the interface of the $i-j$ layers in the case of the incident wave with unit amplitude. They can be determined analytically by enforcing the boundary conditions for the tangential components of the electromagnetic field.

A. Unit Cell

Let us consider the unit cell of periodic structure. We suppose that domain outside the cell is filled by dielectric with relative permittivity $\varepsilon = \varepsilon_1$. Following [13, 14], introduce unknown reflection and transmission coefficients of the unit cell of periodic part, r_{pij}, t_{pij}. The amplitudes of the field inside the layers of the unit cell we denote as b, c, d, f. The geometry of the unit cell and directions of the waves propagation are presented in Fig.1(b). The amplitudes and coefficients are connected by the following scalar equations, which we write in the operator form:

$$r_{p12} = r_{12}e_1 + t_{21}e_2 c, \quad (5)$$

$$b = r_{21}e_2 c + t_{12}e_1, \quad (6)$$

$$c = r_{23}e_2 b + t_{32}e_3 f, \quad (7)$$

$$d = r_{32}e_3 f + t_{23}e_2 b, \quad (8)$$

$$f = r_{31}e_3 d, \quad (9)$$

$$t_{p12} = t_{31}e_3 d, \quad (10)$$

where coefficient $e_j = \exp(ik\gamma(k_0/k_j\xi_0)h_j)$ describes the field variation, as the wave propagates through the distance h_j in the jth layer, $k_j = \varepsilon_{\perp j}k$ is the wave number in the jth layer. Eq. (5) means that the reflected field is the field of the incident wave, reflected from the isolated interface of the layers I–II plus the field transmitted through the interface II–I from domain II to domain I. Using the same considerations, one can obtain (6)-(10). Then,

$$r_{p12} = e_1^2\frac{\alpha t_{21} + \beta r_{12}}{\delta}, \quad (11)$$

$$t_{p12} = \frac{t_{34}e_3 t_{23}e_2 t_{12}e_1}{\delta}, \quad (12)$$

where

$$\alpha = e_2^2 t_{12}r_{23} + e_2^2 t_{12}t_{32}e_3^2 r_{34}t_{23} - e_2^2 t_{12}r_{32}e_3^2 r_{34}r_{23}, \quad (13)$$

$$\beta = -t_{32}e_3^2 r_{34}t_{23}e_2^2 r_{21} + 1 - r_{23}e_2^2 r_{21} - r_{32}e_3^2 r_{34} + r_{32}e_3^2 r_{34}r_{23}e_2^2 r_{21}. \quad (14)$$

In the same manner, taking into account that, generally, the structure is not symmetric with respect to the vertical axis, expressions for the coefficients r_{p21} and t_{p21} can be obtained.

B. Finite Periodic Structure

To obtain reflection and transmission coefficients of the finite periodic structure, we use the recurrent procedure. We suppose that reflection and transmission coefficients, $r_{n,ij}$,

$t_{n,ij}$, of the system of n cells, which are placed in the infinite medium with relative permittivity ε_1, are known (see Fig.2). I.e., for $x<0$ and $x>Ln$ we have $\varepsilon=\varepsilon_1$. Then

$$r_{n+1,12}=r_{n,12}+t_{n,12}C,\ C=r_{p12}B, \tag{15}$$

$$B=r_{n,21}C+t_{n,12},\ t_{n+1,12}=t_{p12}B, \tag{16}$$

$$t_{n+1,21}=t_{n,21}E,\ E=r_{p12}D+t_{p21}, \tag{17}$$

$$D=r_{n,21}E,\ r_{n+1,21}=r_{p21}+t_{p12}D, \tag{18}$$

where B, C, D, E are the amplitudes of the field between the layers. Then

$$r_{n+1,12}=\frac{-t_{n,21}r_{p12}t_{n,12}-r_{n,12}+r_{p12}r_{n,21}r_{n12}}{r_{p12}r_{n,21}-1}, \tag{19}$$

$$t_{n+1,12}=-\frac{t_{p12}t_{n12}}{r_{p12}r_{n,21}-1}, \tag{20}$$

$$r_{n+1,21}=\frac{-t_{n,21}t_{p12}t_{p,21}-r_{p21}+r_{p12}r_{n,21}r_{p21}}{r_{p12}r_{n,21}-1}, \tag{21}$$

$$t_{n+1,21}=-\frac{t_{p21}t_{n12}}{r_{p12}r_{n,21}-1}, \tag{22}$$

where for $n+1=N$, operators r_{p12}, t_{p12}, r_{p21}, t_{p21} correspond to the last unit cell, which right boundary at $x=NL$ is graphene plane – vacuum interface.

Then the reflection and transmission coefficients of the finite periodic structure, for which domain $x<0$ and $x>Ln$ (see Fig.1a) is vacuum with $\varepsilon=1$ are

$$R=\frac{-t_{10}r_{N,12}t_{01}-r_{01}+r_{10}r_{N,12}r_{01}}{-1+r_{10}r_{N,12}}, \tag{23}$$

$$T=-\frac{t_{N,12}t_{01}}{-1+r_{10}r_{N,12}}, \tag{24}$$

where subscript 0 corresponds to the vacuum.

C. Propagation Constant of the Periodic Part

Let us consider infinite ideally-periodic structure of unit cells, which stretch from $x=-\infty$ to $x=+\infty$ (see Fig.3). Denote the propagation constant along the x-axis as β, and e=exp($i\beta L$). The periodicity conditions give us

$$a_2=ea_1,\ b_2=eb_1, \tag{25}$$

and

$$a_2=t_{p12}a_1+r_{p21}b_1,\ b_1=t_{p21}b_2+r_{p12}A_2. \tag{26}$$

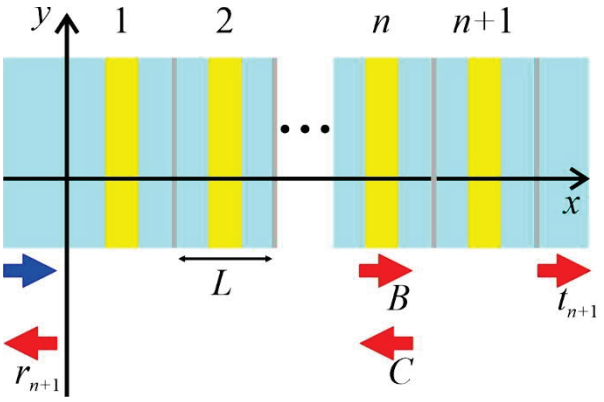


Fig. 2. Finite periodic structure. Recurrent procedure.

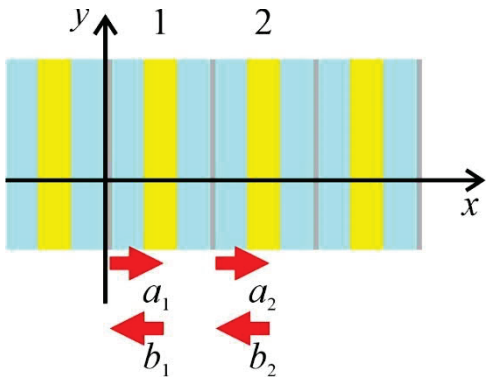


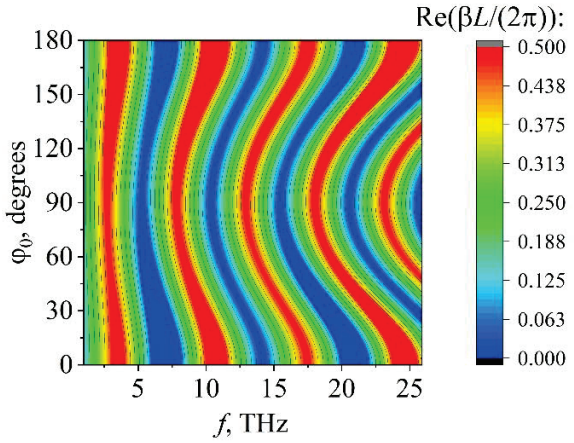
Fig. 3. Infinite periodic structure. Directions of the eigenwaves propagation.

From (25), (26), one can obtain quadratic equation relative e: $e(et_{p21}-1)-t_{p12}et_{p21}+t_{p12}+r_{p21}r_{p12}e=0$, which obviously allows to determine β analytically.

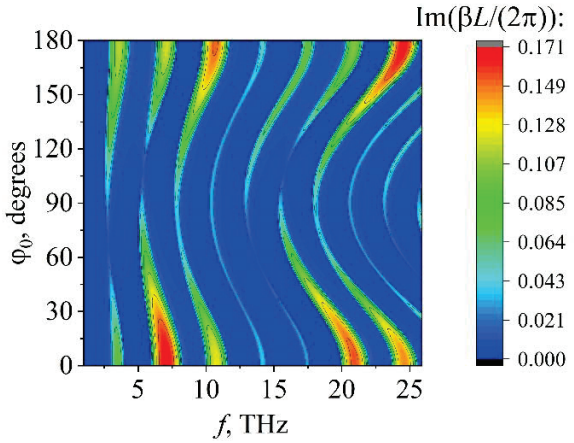
IV. Numerical Results

Fig.4 shows dependences of β on the frequency and the incidence angle. The width of the layers is h_1=h_2= 10μm, h_3=20μm. The parameters of graphene are μ_c= 0.1eV, τ = 1ps, T = 300 K Also ε_1 = ε_3 =2.25, ε_{r2} =12, ε_{a2} =11. The numbed of layers in the finite structure is N=10. Since graphene has losses at any frequency, Im$\beta>0$. Stopbands are characterized by an increase in Imβ, indicating attenuation of wave propagation along the x-axis. Notably, there are specific parameter values where $\beta\to0$ and φ_0=0. When φ_0=0^0, the incident wave propagates entirely parallel to the layers (along the y-axis). In this context, β=0 implies that there is no phase change or amplitude variation in the wave as it propagates from one period of the structure to the next along the x-axis. This condition corresponds to surface waves that propagate along the homogeneous y-axis.

The map of β is not symmetric with respect to the φ_0=90^0 axis. This is due to the effect of nonreciprocal propagation. That means that electrodynamical properties are not equivalent for the incidence angles φ_0 and $\pi-\varphi_0$. In such structures, wave propagation in the forward and backward directions is no longer equivalent, leading to asymmetry in the dispersion characteristics. This nonreciprocity typically arises from the presence of gyrotropic materials or external biasing, and it plays a key role in the design of isolators, circulators, and unidirectional waveguides.



(a)



(b)

Fig. 4. Dependences of the normalized propagation constant of the periodic part of the structure on the frequency and incidence angle. (a) real part, (b) imaginary part.

Fig.5 shows dependences of the transmission coefficient on the frequency for two values of the incidence angle, $\varphi_0=9^0$ and $\varphi_0=171^0$. As one can see, for these incidence angles, significant nonreciprocity is observed. There are frequency bands, where almost total transmission occurs for the so-called "positive" incidence angles $\theta_0=\pi/2-\varphi_0$, while for the "negative" incidence angles θ_0, transmission is almost absent. Also in Fig.5, we present the results for the same configuration but without graphene. The presence of graphene, due to its losses, reduces the maximum of the amplitude of the transmitted field in the passbands. However, more importantly, it also reduces the magnitude of the oscillations, making the nonreciprocity effect more pronounced.

V. Conclusion

In this work, we have analytically and numerically studied a magnetophotonic crystal structure incorporating graphene-coated dielectric layers and a gyrotropic medium. Numerical results demonstrated the existence of stopbands and surface wave modes. Importantly, strong nonreciprocal propagation effects were observed. The presence of graphene, while introducing losses, was shown to enhance the nonreciprocal behavior by damping parasitic oscillations. These results highlight the potential of such structures for applications in nonreciprocal photonic components, including isolators, circulators, and devices requiring unidirectional transmission.

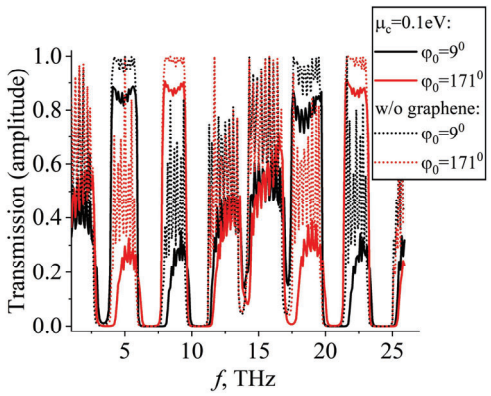


Fig. 5. Dependences of the transmission coefficient (amplitude) of the finite structure on the frequency for two incident angles $\varphi_0=9^0$, $\varphi_0=171^0$. For comparison, the results for the structure without graphene are also shown.

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Hyper-Singular Integral Equations in the H-Polarized Plane Wave Scattering and Absorption by Venetian-Blind Graphene Grating

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This paper introduces a rigorous method based on hyper-singular integral equations (HSIEs) for analyzing the H-polarized THz wave scattering and absorption by the venetian-blind graphene strip grating. We study the total scattering, absorption, and radar cross-sections as functions of frequency, strip tilt angle, and graphene's chemical potential. The considered configuration facilitates the redirection of reflected energy away from the source, demonstrating the potential of these structures for stealth technologies and waves manipulation.

Keywords—graphene, grating, hyper-singular integral equation, scattering, absorption

I. Introduction

The study of electromagnetic wave diffraction by periodic structures is a fundamental problem in applied electromagnetics, holding significant importance for the development of various radio engineering devices such as antennas, filters, polarizers, and metamaterials [1-3]. Among the variety of periodic structures, strip gratings occupy a special place due to their relative simplicity and broad functional capabilities.

The classical problem of diffraction by strip grating has been well-studied using various analytical and numerical methods, including the Wiener-Hopf method, method of the Riemann-Hilbert problem, method of moments, methods of integral equations, operator method, etc. [4-8]. However, significantly less attention has been paid to cases where the grating elements (strips) are not coplanar with the grating plane. Such "tilted" or "inclined" strip gratings (venetian blind-type) can exhibit unique electrodynamic properties, opening up new possibilities for controlling electromagnetic wave characteristics. Furthermore, these configurations can redirect the reflected field away from the source, making them potentially useful in stealth technologies.

The emergence of graphene, a two-dimensional material with exceptional electronic and optical properties, has spurred significant interest in its application in various electromagnetic devices [10-13]. Graphene-based periodic structures offer unique opportunities for manipulating electromagnetic waves at different frequency ranges due to the tunability of its conductivity. Therefore, the development of rigorous methods for analyzing diffraction by graphene-based periodic structures is of particular importance and current relevance. The graphene can be modeled as a zero-thickness surface with a specific conductivity rather than a volumetric material. The conductivity of graphene, $\sigma = \sigma(f, \mu_c, \tau, T)$, is a function of the frequency f, chemical potential μ_c, electron relaxation time τ, temperature T [14, 15]. The edge conditions on graphene strips should take into account the singularities of the surface currents [16].

The development of rigorous and efficient methods for analyzing diffraction by infinite and finite periodic graphene strip gratings with tilted elements is interesting scientific and technical task.

In this paper, we present the development of a rigorous method of hyper-singular integral equations (HSIEs) for the analysis of the electromagnetic wave diffraction problem by finite periodic venetian blind-type graphene strip grating. The proposed approach is based on the formulation of HSIEs with respect to the electric current densities on the surface of the graphene strips, taking into account the surface conductivity of graphene. To solve the obtained equations, the meshless Nystrom-type algorithm is used [17]. The developed method allows us to obtain rigorous full-wave solutions of the diffraction problems. It takes into account the radiation condition and singularities of the field at the edges of the strips analytically.

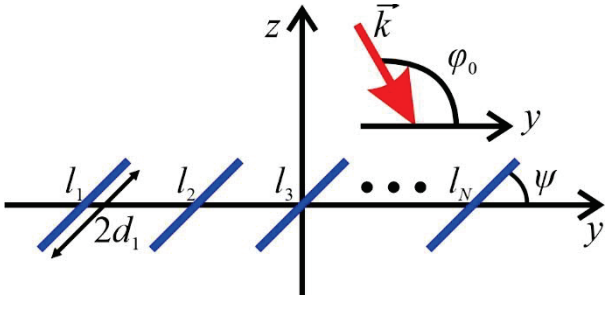


Fig. 1. Structure geometry.

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II. Problem Statement

We consider the set of N graphene strips. The axis of the grating coincides with the y-axis. The rotation angle of the strips is ψ. The y-coordinate of the nth is l_n, the width of the nth strip is $2d_n$. The strips are infinite along the x-axis. The geometry is shown in Fig.1.

We consider only the H-polarized waves, since here excitation of the plasmon resonances can take place. Suppose that plane wave with unit amplitude is incident on the grating from domain $z > 0$, $H^i=\exp(-ik\xi_0 y - ik\gamma(\xi_0)z)$, where $\xi_0=\cos\varphi_0$, φ_0 is the incidence angle with respect to the y-axis, $\gamma(\xi_0)=\sqrt{1-\xi_0^2}$, $\mathrm{Re}\gamma \ge 0$, $\mathrm{Im}\gamma \ge 0$, k is the wave number. The time dependence $\exp(-i\omega t)$ is omitted.

The electromagnetic field should satisfy the Helmholtz equation, the boundary conditions on graphene strips,

$$E_y^+ = E_y^-,\ E_y^\pm = -\frac{1}{\sigma}(H_z^+ - H_z^-), \tag{1}$$

where signs "$\pm$" describe the limit value of the field from above (below) the strips, the radiation and the edge conditions.

III. Solution of the Problem

Let us denote the unknown currents density function on the nth strip as w_n. The scattered field can be represented as a double-layer potential

$$H_z^{sc}(\vec{r}) = \sum_{n=1}^{N} \int_{L_n} w_n(\vec{r}_1) \frac{\partial G(\vec{r},\vec{r}_1)}{\partial \vec{n}_1} d\vec{r}_1, \tag{2}$$

where $\vec{n}_1 = (-\sin\psi, \cos\psi)$ is the normal vector to the strip, $\vec{r} = (y,z)$ is the radius-vector of the observation point, $G_0(\vec{r},\vec{r}_1) = \frac{i}{4} H_0(k_1 |\vec{r} - \vec{r}_1|)$ is the Green's function, L_n represents the nth strip,

$$L_n : \begin{cases} y = l_n + \zeta\cos\varphi, \\ z = \zeta\sin\varphi, \end{cases} \quad \zeta \in [-d_n,\ d_n]. \tag{3}$$

The enforcement of the boundary conditions leads to the following HSIE:

$$\begin{aligned} &\frac{4}{\sigma Z_0} w_q(x) + \int_{-d_m}^{d_m} \frac{H_1(k|\zeta_m - \xi_m|)}{|\zeta_m - \xi_m|} w_m(\zeta_m) d\zeta_m \\ &+ \sum_{\substack{n=1 \\ n\ne m}}^{N} \int_{-d_n}^{d_n} G_1(\xi_m,\zeta_n) w_n(\zeta_n) d\zeta_n \\ &= \frac{4i}{k} \frac{\partial H^i(x, y=0)}{\partial \vec{n}}, \\ &(y(\xi_m),\ z(\xi_m)) \in L_m,\ \xi_m \in [-d_m,\ d_m],\ m = 1,\ldots, N, \end{aligned} \tag{4}$$

where $G_1(\xi_m,\zeta_n) = \frac{\partial^2 G(\vec{r},\vec{r}_1)}{\partial\vec{n}\partial\vec{n}_1}$ on the strips.

The first integral in (4) has non-integrable in general sense singularity of the second order, $1/x^2$, $x \to \infty$. It should be understood as Hadamar finite part integral. The second integral in (4) is regular one. From the edge condition it follows, that the currents function w_n approaches zero as a square root near the edges of the strips, $w_n(x) \sim \sqrt{|x-\theta|}$, if $x \to \theta$, where θ represents one of the edges of the strip.

Following [12, 17], for the discretization of (16) using the Nystrom-type algorithm, we first change the variables to reduce the integration over $[-d_n, d_n]$ to the integration over the standard interval $[-1, 1]$. Then we apply the Gauss-Chebyshev quadrature rule with the Chebyshev weight $\sqrt{1-t^2}$. As nodes and collocation points, the zeros of the Chebyshev polynomials of the second kind on every strip are taken. As a results, the system of linear equations can be obtained relative unknown values of the currents density at the nodes.

IV. Numerical Results

The total scattering cross-section (*TSCS*), absorption cross-section (*ACS*), and radar cross-section are

$$TSCS = \frac{2\pi}{k} \int_0^{2\pi} |C(\varphi)|^2 d\varphi, \tag{5}$$

$$ACS = \frac{1}{\mathrm{Re}(\sigma Z_0)} \sum_{n=1}^{N} \int_{L_n} |w_n(\vec{r}_1)|^2 d\vec{r}_1, \tag{6}$$

$$RCS = C(\pi - \varphi_0), \tag{7}$$

where $Z_0 = 120\pi\,\Omega$ is free space impedance. The far field amplitude is

$$\begin{aligned} C(\varphi) = \frac{k}{\pi} \sum_{n=1}^{N} \int_{L_n} w_n(\vec{r}_1) \exp(-ik(y\cos\varphi + z\sin\varphi)) \\ \times \sin(\psi - \varphi) d\vec{r}_1. \end{aligned} \tag{8}$$

The power conservation law is

$$ACS + TSCS = -\frac{4\pi}{k} C(-\varphi_0). \tag{9}$$

Suppose that all strips have the same width of $2d$, the period of the grating is l. Also, the room temperature is T=300 K, electron relaxation time is τ =1 ps.

Figs.2,3 show dependences of *TSCS*, *ACS* and *RCS* on the frequency for various rotation angles of the strips ψ, and chemical potential of graphene μ_c. Two distinct types of resonances are observed in the studied structures. The plasmon resonances are marked as P_j (where j is the resonance index) and Rayleigh anomaly are marked as *RA*. The increase in the chemical potential leads to an upshift of the plasmon resonance frequencies. Concurrently, the maxima of both *TSCS* and *ACS* become more pronounced, indicating stronger interaction of the electromagnetic wave with the graphene strips at higher doping levels. This behavior is

attributed to the increased carrier density in graphene, enhancing its plasmonic response.

In the case of normal incidence, $\varphi_0=90^0$, if the structure is symmetric with respect to the vertical axis, $\psi = 0^0$, only odd-index plasmon resonances can excite. Introducing a rotation of the strips, $\psi > 0^0$, breaks this symmetry, leading to the emergence of even-index plasmon resonances.

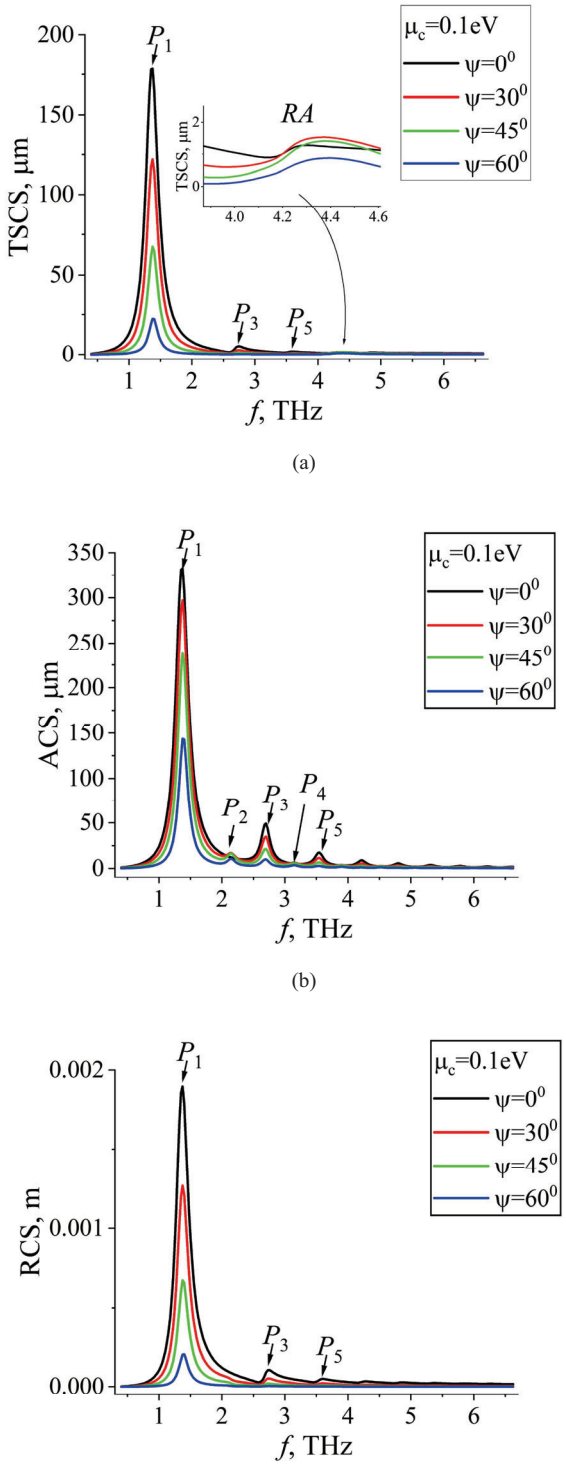


Fig. 2. Dependences of (a) TSCS, (b) ACS, and (c) RCS on the frequency for various rotation angles of the strips, μ_c = 0.1eV and N=10, d=10μm, l=70 μm. The inset in fig.(a) demonstrates the zoomed scattering near the Rayleigh anomaly wavelength.

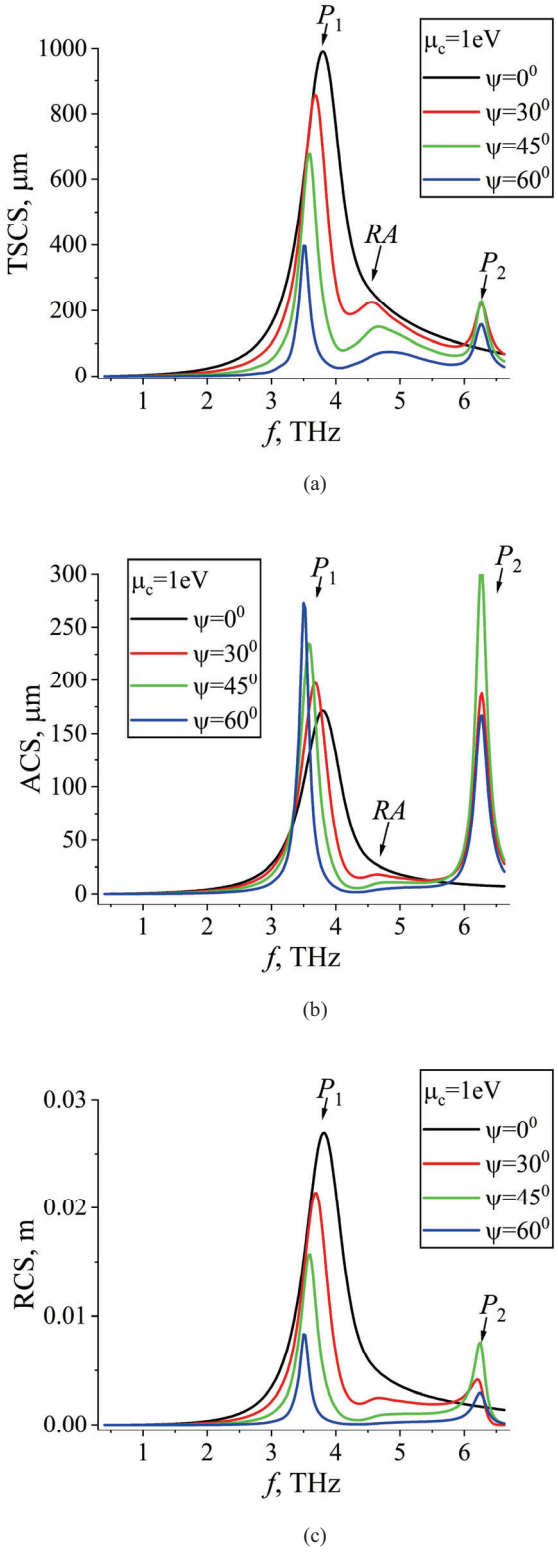


Fig. 3. Dependences of (a) TSCS, (b) ACS, and (c) RCS on the frequency for various rotation angles of the strips, μ_c = 1eV and N=10, d=10μm, l=70 μm.

Analyzing the impact of the rotation angle ψ, we observe a decrease in the TSCS and RCS near the first plasmon resonance frequency as ψ increases. This suggests that as the strips become tilted, the incident wave's energy is less efficiently scattered. The power of the incident wave is partially spent on the increase of the absorption by graphene strips and partially is redirected away from the source.

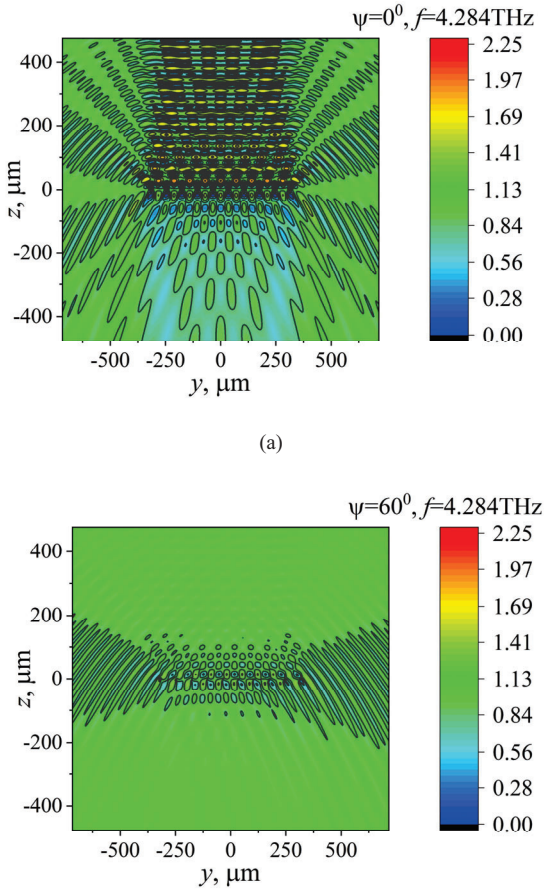


Fig. 4. Near fiel patterns near the Rayleigh anomaly region, f=4.284THz, for two rotation angles of the strips (a) $\psi = 0^0$ and (b) $\psi = 60^0$, μ_c = 1eV, N=10, d=10μm, l=70μm.

However, near the second plasmon resonance, an increase in both TSCS and RCS is observed for $\psi > 0^0$. This is because, for the planar grating, ψ=0^0, the even-index plasmon resonances are not excited under normal incidence due to symmetry. The introduction of tilted strips allows these resonances to be excited, leading to enhanced scattering in these frequency ranges.

In the case of a higher chemical potential, μ_c = 1eV, coupling between graphene strips becomes stronger. This results in a shift in the plasmon resonance frequencies as the rotation angle ψ is varied.

Fig.4 shows the near field patterns in the vicinity of the Rayleigh anomaly region. The field of the incident plane wave redirected by inclined strips, which propagates near the plane of the grating, z = 0, is clearly seen.

V. Conclusion

In this paper, we have presented a rigorous method based on the hyper-singular integral equations for the analysis of the electromagnetic wave scattering and absorption by the venetian-blind graphene strip grating. The developed approach accurately accounts for the frequency-dependent surface conductivity of graphene and the geometric features introduced by the tilted strips. The discretization scheme is based on the Nystrom-type algorithm.

The considered structure allows for both dynamic and geometric manipulation of the scattered field through the tunable conductivity of graphene and the strip rotation angle.

Numerical results revealed that tilting the graphene strips enables excitation of both odd and even plasmon resonances in the case of the normal incident plane wave, and importantly, facilitates the redirection of reflected energy.

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Parallel Method using Covariance Matrix Adaptation Evolution Strategy and SCIP Solver for Generation Capacity Structure Optimization

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***Abstract*—A parallel method is proposed for solving mixed-integer linear programming (MILP) problems that arise in the optimization of generation capacity structure in power systems. The core of the method is the decomposition of a high-dimensional MILP problem into a master problem, which iteratively searches for the optimal generation capacity structure, and an associated set of MILP subproblems aimed at finding optimal load modes of generating capacities under various operating conditions. At each iteration, the master problem is solved using the Covariance Matrix Adaptation Evolution Strategy (CMA-ES), while the associated set of MILP subproblems is solved using the SCIP solver. Since the MILP subproblems are independent and of significantly lower dimensionality, they are solved in parallel with SCIP and require relatively modest computational resources. Computational experiments show that the proposed method finds the global optimum in 97% of cases and achieves a speedup of up to 7.84× when using 64 threads, with memory consumption growing nearly linearly. The method scales well and is suitable for deployment in high-performance computing environments, providing a flexible and sufficiently accurate tool for strategic planning of power systems focused on renewable energy sources.**

***Keywords*—generation capacity structure planning, CMA-ES, MILP, parallel optimization, renewable energy, SCIP**

I. Introduction

Reliable electricity supply is a key prerequisite for the uninterrupted functioning of the economy, the well-being of the population, and the stable operation of a country's social infrastructure. Under the conditions of systematic Russian missile and drone attacks on Ukraine's energy infrastructure, disruptions in electricity supply significantly affect all spheres of public life.

One of the effective ways to enhance the resilience of the power system against the destructive consequences of such attacks is the development of distributed generation technologies, particularly the deployment of renewable energy sources (RES) such as solar power plants (SPPs) and wind power plants (WPPs). Due to their decentralized structure, broad potential for local deployment, relatively low equipment cost, and short construction and restoration timelines, these facilities are less vulnerable to simultaneous destruction from attacks. Moreover, their spatial dispersion reduces the strategic and economic feasibility of using expensive precision-strike weapons against them, making the power system as a whole more adaptive, flexible, and resilient in the face of military threats [1].

The use of renewable energy sources (RES) enhances energy security and reduces harmful emissions [2], [3], [4]. However, electricity generation facilities that rely on such sources as solar or wind energy are characterized by significant variability in output and limited controllability compared to conventional power plants. Their capacity largely depends on weather conditions and cannot be controlled by the power system operator [5].

As the share of RES in electricity generation increases, maintaining the balance between electricity supply and demand becomes more complex. This necessitates the development of flexible generation capacities, the deployment of efficient energy storage systems (ESS), and thorough strategic planning using mathematical models of power systems for predictive analysis. A particularly important requirement is the adequacy of such models in accurately representing the load modes of power generation equipment.

Global strategic planning models such as MARKAL, TIMES, and LEAP have limited capabilities in representing the operational load modes of power units while accounting for their detailed technical characteristics [6]. In contrast, unit commitment models provide a detailed representation of generation equipment load modes by incorporating a range of important technical parameters, resulting in a high level of adequacy in forecasting outcomes [7].

To describe the operating modes of wind and solar power plants, average load graphs or probabilistic power distribution functions are commonly used [8]. However, such approaches fail to fully capture the high variability inherent in RES. A more adequate representation is provided by representative models of WPPs and SPPs, which reproduce the load graphs of their generating equipment in a manner that closely reflects real-world operating conditions.

Mathematical modeling problems of electric power system load modes, formulated based on representative models of WPP and SPP operation as well as ESS load models, belong to the class of high-dimensional mixed-integer linear programming (MILP) problems [9]. Solving them requires the use of efficient optimization methods. Due to the high computational complexity of such problems, their solution is often feasible only with the use of powerful parallel computing resources.

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One approach to reducing the computational complexity of such problems is the Benders decomposition method, which separates the problem into a first-stage (master) problem and second-stage subproblems that are solved iteratively with the addition of constraints at each iteration [10]. While this approach is applied in generation capacity structure planning for electric power systems [11], it may require considerable computational resources due to a large number of iterations, rapid growth in the number of constraints, and slow convergence in the final stages [12].

An alternative approach to solving MILP problems is the use of metaheuristic algorithms. Recently, these algorithms have gained increasing popularity in the field of energy systems optimization [13]. Their advantage lies in the fact that they do not rely on gradients of objective functions or the specific structure of the problem—only evaluations of the objective function at selected points are required [14]. In particular, metaheuristics can be effectively applied in the decomposition of a problem for solving it at the first (master) stage [15]. This approach enables independent and parallel solving of second-stage subproblems using the most efficient methods available for this purpose.

One such metaheuristic algorithm is the Covariance Matrix Adaptation Evolution Strategy (CMA-ES). Like Differential Evolution (DE) or Genetic Algorithm (GA), CMA-ES belongs to the population-based evolutionary methods. In each iteration, the algorithm randomly generates potential solutions from a multivariate normal distribution and then adapts the covariance matrix and step size to efficiently search for optima of complex nonlinear functions without using gradient information. There are also modifications of this algorithm that enable its application not only to optimization problems in the space of real variables but also to problems with integer variables [16], [17].

In this study, we aim to evaluate the effectiveness of applying the CMA-ES algorithm to solve generation capacity structure optimization problems in electric power systems by decomposing such problems and performing parallel computations during the solution of second-stage subproblems.

II. Methods

This study addresses the problem of optimizing the generation capacity structure of an electric power system with generating equipment based on SPPs, WPPs, and ESS. The objective is to minimize the total investment and operational costs of the power system operator over a five-year forecast period.

A. Features of the Electric Power System Model

To model the electric power system, we used the mathematical formulation of the MILP problem presented in [18]. We employed both representative clustered models of RES and electricity consumption models. Each of them includes four weekly load graphs for each season, as shown in Figs. 1–6. Each consumption graph carries equal weight in the overall planning period. In the case of RES, the first two graphs are unique and represent the dynamics of the capacity factor with the lowest and highest efficiency levels, respectively. The third and fourth graphs are representative of two additional efficiency clusters.

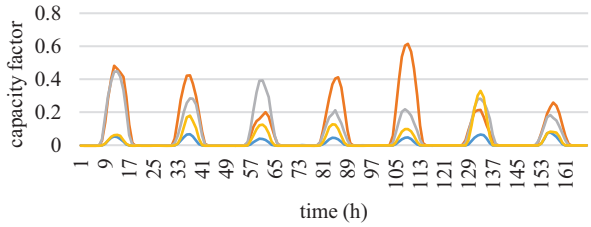


Fig. 1. Weekly representative graphs of capacity factors of SPPs for winter, colored by cluster.

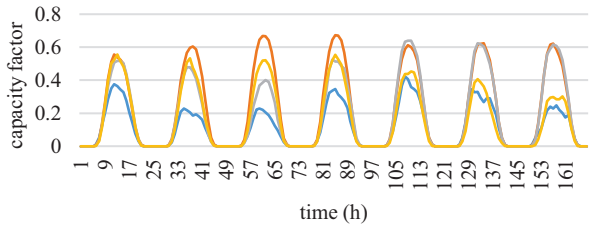


Fig. 2. Weekly representative graphs of capacity factors of SPPs for summer, colored by cluster.

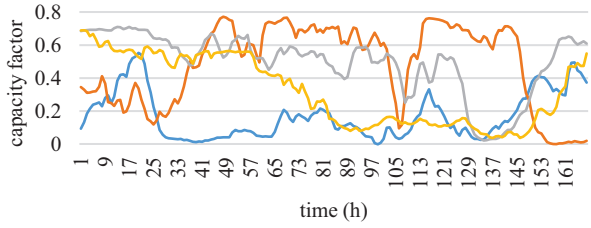


Fig. 3. Weekly representative graphs of capacity factors of WPPs for winter, colored by cluster.

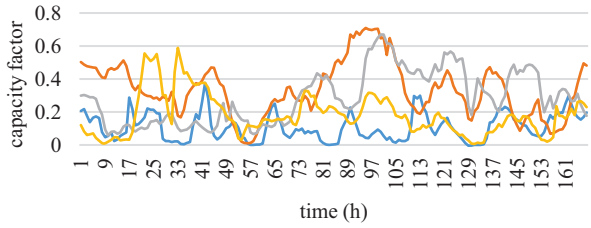


Fig. 4. Weekly representative graphs of capacity factors of WPPs for summer, colored by cluster.

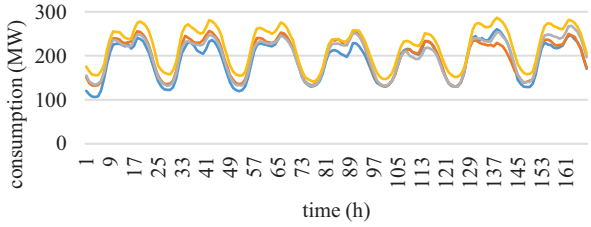


Fig. 5. Weekly representative graphs of electricity consumption for winter, colored by cluster.

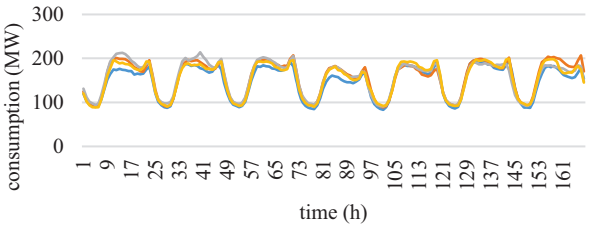


Fig. 6. Weekly representative graphs of electricity consumption for summer, colored by cluster.

The mathematical model of the electric power system is constructed from a unified set of clusters, where each cluster combines SPP and WPP efficiency graphs with a corresponding consumption graph. As a result, we created a set consisting of 96 clusters. For each of the four seasons, six RES efficiency clusters were generated based on the following combinations:

- first SPP and WPP clusters (lowest efficiency),
- second clusters (highest efficiency),
- third SPP cluster combined with the third and, separately, fourth WPP clusters,
- fourth SPP cluster combined with the third and, separately, fourth WPP clusters.

Each of the six mentioned clusters was combined with four consumption graphs, resulting in 24 clusters for each season. Thus, the total number of clusters required to model the load modes of the electric power system amounted to 96.

B. Problem Decomposition

Given the specifics of the generation capacity structure optimization problem, we decomposed it into a first-stage (master) problem and a set of second-stage subproblems. The first-stage problem iteratively determines the number of equipment units and their installation costs. In parallel, the 96 second-stage subproblems are solved to determine the optimal load modes of the generating equipment and compute the corresponding minimum operational costs. The overall scheme of this approach is shown in Fig. 7.

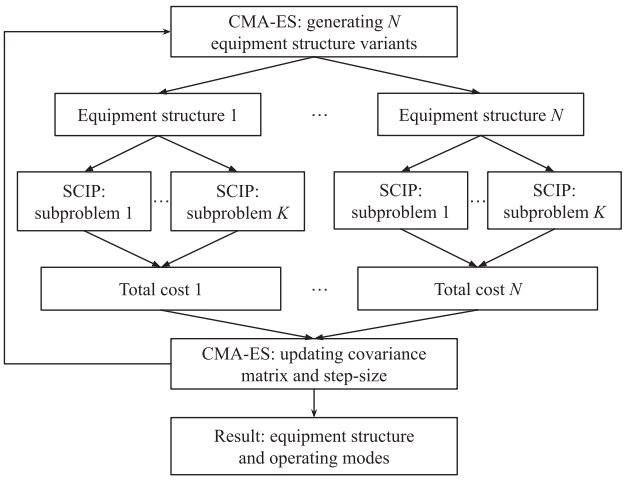


Fig. 7. Decomposition approach.

To solve the first-stage problem, the CMA-ES evolutionary algorithm is used. Within each iteration, the algorithm generates N equipment structure variants, i.e., different combinations of the number of units of each type. Then, for each generated variant and each cluster, K second-stage subproblems are solved in parallel using the SCIP solver [19]. As a result, the total cost is calculated for each configuration, based on which the CMA-ES algorithm updates its parameters. This process is repeated until the stopping criteria defined in the CMA-ES algorithm are met.

C. Specifics of CMA-ES Application

For the experiment, we used the *pycma* library version 4.0.0 [20], which implements the CMA-ES algorithm described below [16].

At each iteration t, the algorithm generates λ points $x_i^{(t)}$ (a generation), which correspond to possible solutions to the problem and have N coordinates:

$$y_i^{(t)} = \sqrt{C^{(t)}} z_i^{(t)}, \qquad i = 1, \dots, \lambda, \tag{1}$$

$$x_i^{(t)} = m^{(t)} + \sigma^{(t)} D^{(t)} y_i^{(t)}, \qquad i = 1, \dots, \lambda, \tag{2}$$

where $\lambda = 4 + \lfloor 3 \ln N \rfloor$, $C^{(t)}$ is the covariance matrix, $z_i^{(t)} \sim \mathcal{N}(0, I)$ are points drawn from an isotropic normal distribution, $m^{(t)} \in \mathbb{R}^N$ is the mean point, $\sigma^{(t)}$ is the step size, $D^{(t)}$ is the diagonal matrix that scales the coordinate system.

Before evaluating the objective function, the coordinates of each point that must be integers are processed using the int[·] function, which rounds the coordinate value to the nearest integer:

$$\text{int}[\cdot]: \mathrm{R} \to \mathrm{Z}, \qquad \mathrm{x} \mapsto \left\lfloor \mathrm{x} + \frac{1}{2} \right\rfloor. \tag{3}$$

The resulting points are sorted according to the value of the objective function f:

$$f\left(x_{1:\lambda}^{(t)}\right) \le f\left(x_{2:\lambda}^{(t)}\right) \le \cdots \le f\left(x_{\lambda:\lambda}^{(t)}\right). \tag{4}$$

The mean point is updated based on the μ best points:

$$m^{(t+1)} = m^{(t)} + \sigma^{(t)} D^{(t)} \sum_{i=1}^{\mu} w_i \, y_{i:\lambda}, \tag{5}$$

where $\mu = \lfloor \lambda/2 \rfloor$ is the number of best points and $w_i \propto \log\left((\lambda+1)/2\right) - \log(i)$ are their weights, under the condition $\sum_{i=1}^{\mu} w_i = 1$.

The step size $\sigma^{(t)}$ is updated based on its evolution path $p_\sigma^{(t+1)}$:

$$p_\sigma^{(t+1)} = (1 - c_\sigma) p_\sigma^{(t)} + \sqrt{c_\sigma (2 - c_\sigma) \mu_{\text{eff}}}\, C^{(t)^{-1/2}} \sum_{i=1}^{\mu} w_i \, y_{i:\lambda}, \tag{6}$$

$$\sigma^{(t+1)} = \sigma^{(t)} \exp\left[\frac{c_\sigma}{d_\sigma} \frac{\| p_\sigma^{(t+1)} \|}{\mathbb{E}(\| \mathcal{N}(0, I) \|)} \right], \tag{7}$$

where

$$\mu_{\text{eff}} = 1 / \sum_i w_i^2, \tag{8}$$

$$c_\sigma = \frac{\mu_{\text{eff}} + 2}{N + \mu_{\text{eff}} + 5}, \tag{9}$$

$$d_\sigma = 1 + 2\max\left(0, \frac{\sqrt{\mu_{\text{eff}} - 1}}{N + 1} - 1\right) + c_\sigma. \tag{10}$$

The covariance matrix $C^{(t+1)}$ is updated based on its evolution path $p_c^{(t+1)}$:

$$p_c^{(t+1)} = (1 - c_c)p_c^{(t)} + h_\sigma\sqrt{c_c(2 - c_c)\mu_{\text{eff}}}\sum_{i=1}^{\mu} w_i y_{i:\lambda}, \tag{11}$$

$$C^{(t+1)} = \left(1 + c_1(1 - h_\sigma)c_c(2 - c_c) - c_1 - c_\mu\sum_{i=1}^{\lambda} w_i\right)C^{(t)} + c_1\, p_c^{(t+1)}\left(p_c^{(t+1)}\right)^{\top} + c_\mu\sum_{i=1}^{\lambda} w_i^{\circ}\, y_{i:\lambda}y_{i:\lambda}^{\top}, \tag{12}$$

where

$$c_1 = \frac{2}{(N + 1.3)^2 + \mu_{\text{eff}}}, \tag{13}$$

$$c_\mu = \min\left(1 - c_1, \frac{2\mu_{\text{eff}} + 1/\mu_{\text{eff}} - 1.75}{(N + 2)^2 + \mu_{\text{eff}}}\right), \tag{14}$$

$$c_c = \frac{4 + \mu_{\text{eff}}/N}{N + 4 + 2\mu_{\text{eff}}/N}, \tag{15}$$

$$h_\sigma = \begin{cases} 1 & if\ \dfrac{\left\|p_\sigma^{(t+1)}\right\|^2}{1 - (1 - c_\sigma)^{2t}} < 2N\left(1 + \dfrac{2}{N + 1}\right) \\ 0 & otherwise, \end{cases} \tag{16}$$

$$w_i^{\circ} = \begin{cases} w_i & if\ w_i \geq 0 \\ \dfrac{w_i\, N}{\left\|C^{(t)^{-1/2}} y_{i:\lambda}\right\|^2} & otherwise. \end{cases} \tag{17}$$

In the case of integer-valued coordinates, the algorithm described above is supplemented with operations for imposing a lower bound on the standard deviation and centering the points within the integer interval [16].

To this end, before generating the points (2), the diagonal matrix $D^{(t)}$ is updated to constrain the standard deviation $\sigma_{\text{std(j)}}^{(t)}$ to the value σ_{LB}:

$$\mathrm{D}_{\mathrm{j,j}}^{(\mathrm{t})} \leftarrow \frac{max\left(\sigma_{\mathrm{LB}}, \sigma_{\mathrm{std(j)}}^{(\mathrm{t})}\right)}{\sigma^{(\mathrm{t})}\sqrt{\mathrm{C}_{\mathrm{j,j}}^{(\mathrm{t})}}},\ \mathrm{j} \in \mathrm{S}_{\mathbb{Z}}, \tag{18}$$

where $S_{\mathbb{Z}}$ is the set of all integer-valued point coordinates,

$$\sigma_{\mathrm{LB}} = min\left(\frac{\mu_{\text{eff}}}{\mathrm{N}}, 0.2\right), \tag{19}$$

$$\sigma_{\text{std(j)}}^{(t)} = \sigma^{(t)}\, D_{j,j}^{(t)}\sqrt{C_{j,j}^{(t)}}, \qquad j = 1, \ldots, N. \tag{20}$$

Centering of the points is performed after step (4) and before step (5). The non-centered points obtained in step (2) are denoted as $x_i^{(t-)}$:

$$\mathrm{x}_{\mathrm{i}}^{(\mathrm{t}-)} = \mathrm{x}_{\mathrm{i}}^{(\mathrm{t})} \quad \left(\mathrm{i} \in \mathrm{I}_\mu\right), \tag{21}$$

where $I_\mu = \{1{:}\lambda,\ 2{:}\lambda,\ \ldots,\ \mu{:}\lambda\}$ is the set of indices of the μ best points according to the objective function values.

First, the integer-valued coordinates of the best points that do not coincide with the mean point are rounded:

$$\left(\mathrm{x}_{\mathrm{i}}^{(\mathrm{t})}\right)_{\mathrm{j}} \leftarrow int\left[\left(\mathrm{x}_{\mathrm{i}}^{(\mathrm{t}-)}\right)_{\mathrm{j}}\right] \quad \text{if } \mathrm{int}\left[\left(\mathrm{x}_{\mathrm{i}}^{(\mathrm{t}-)}\right)_{\mathrm{j}}\right] \neq int\left[\left(\mathrm{m}^{(\mathrm{t})}\right)_{\mathrm{j}}\right]. \tag{22}$$

Then, the bias caused by centering is calculated:

$$b_j^{(t)} = \sum_{i \in I_\mu}\left(\left(x_i^{(t)}\right)_j - \left(x_i^{(t-)}\right)_j\right), \qquad j = 1, \ldots, N. \tag{23}$$

Then, this bias is compensated using the points that coincide with the mean point:

$$\left(x_i^{(t)}\right)_j \leftarrow \left(x_i^{(t-)}\right)_j + \mathbb{1}(i, j)\,\alpha_j\,\Delta_{ij}, \tag{24}$$

where $\alpha_j \leq 1$ is chosen to minimize $b_j^{(t)}$,

$$\Delta_{ij} = int\left[\left(x_i^{(t-)}\right)_j\right] - \left(x_i^{(t-)}\right)_j, \tag{25}$$

$$\mathbb{1}(i, j) = \begin{cases} 1, & if\ b_j^{(t)}\Delta_{ij} < 0\ and \\ & \quad int\left[\left(x_i^{(t)}\right)_j\right] = int\left[\left(m^{(t)}\right)_j\right], \\ 0, & otherwise. \end{cases} \tag{26}$$

Finally, the $y_i^{(t)}$ values computed in (1) are updated according to the $x_i^{(t)}$ values modified in (21) and (23).

The above sequence is repeated until the stopping criteria are met, particularly *tolflatfitness*, which defines the minimum allowable relative difference between the best and worst objective function values within one generation, and *tolstagnation*, which limits the maximum number of generations without noticeable progress. In our experiment, these criteria were not explicitly set but were automatically determined by the *pycma* library.

Before calling the *pycma* library, the maximum allowable number of equipment units of each type was determined based on the investment constraint. This number was passed to the library via the *CMA_stds* parameter to initialize the diagonal matrix $D^{(t)}$ and was used as the upper bound for the coordinates of the points $x_i^{(t)}$ generated by the library. The origin, i.e., zero quantity of each equipment type, served as both the lower bound and the initial mean point.

All three variables were declared as integer using the *integer_variables* parameter. The value 0.25 was used as the initial step size σ^0 since the recommended value is ¼ of the search space [20], and the coordinate system was scaled accordingly using the *CMA_stds* parameter.

The population size λ, computed by the library, was increased to 25 using the parameter *popsize_factor* = 2. Additionally, the final evaluation of the objective function was disabled with *eval_final_mean* = *false*, as it did not increase the likelihood of finding the global optimum. However, this likelihood was positively influenced by ignoring negative

values w_i° during the covariance matrix update at step (11) using *CMA_active = false*, and by normalizing the trace of the matrix immediately after this step using *CMA_const_trace = true*:

$$C^{(t+1)} \leftarrow \frac{C^{(t+1)}}{tr(C^{(t+1)})/N} \quad (27)$$

In cases where the calculation of total cost was not possible due to violations of technological constraints for a given equipment structure, the value 10^{300} was returned as the objective function value.

D. Computational Experiments

Experimental studies were conducted on *genoa* nodes of the Vulcan high-performance computing cluster at HLRS, equipped with 768 GB of RAM and two AMD EPYC™ 9334 processors with a clock speed of 2.7 GHz, each containing 32 cores [21]. The algorithms were implemented in the Julia programming language [22] using the JuMP library [23].

Execution time and memory usage required to solve the problem using the SCIP solver in single-threaded mode without decomposition were first examined. The problem was solved ten times to avoid the influence of random factors on the measurement results.

The next part of the experiment aimed to evaluate the efficiency of solving the decomposed problem using the CMA-ES algorithm with varying numbers of threads available for parallel subproblem execution. For each configuration, the problem was solved one hundred times. The attempt number (from 1 to 100) was passed to the *pycma* library via the *seed* parameter to initialize the random number generator. During each attempt, execution time and memory usage were recorded, and the result was checked for consistency with the reference obtained in the earlier experiment.

III. Results

The initial part of the experiment involved solving the problem using only the SCIP solver. The computation time and memory usage are presented in the second column of Table I.

TABLE I. MEASUREMENT RESULTS

Solver	SCIP	CMA-ES			
Threads	1	8	16	32	64
Mean Execution Time (s)	1363.23	826.17	449.20	259.79	173.78
Execution Time Std. Dev. (s)	1.52	112.36	59.94	34.02	21.72
Speedup	-	1.65	3.03	5.25	7.84
Mean Memory Consumption (GB)	3.01	8.14	10.76	15.67	29.57
Memory Consumption Std. Dev. (GB)	0.02	0.30	0.28	0.47	0.88
Memory Usage Ratio	-	2.70	3.57	5.20	9.81

In the subsequent part, a hybrid approach combining the CMA-ES algorithm and the SCIP solver was used. In 97% of runs, this approach found the global optimum, yielding the same equipment structure as obtained using SCIP alone. The third and following columns of Table I present the corresponding values of computation time, memory usage, and the relative changes of these metrics compared to the first configuration.

IV. Discussion

The results of the study confirmed the effectiveness of combining the CMA-ES algorithm with parallel solving of subproblems using SCIP. This approach made it possible to find the global optimum of the generation capacity structure optimization problem for electric power systems in 97% of the cases.

The acceleration of the optimization process strongly depended on the number of threads involved: increasing the thread count from 8 to 64 led to a speedup factor increase from 1.65 to 7.84 compared to using SCIP alone. At the same time, memory usage increased from a factor of 2.70 to 9.81, which aligns with the nature of parallel computations — more simultaneous subproblems require more resources.

It is worth noting that the high scalability is explained by the architectural design illustrated in Fig. 7. In each generation of points, the CMA-ES algorithm independently and simultaneously evaluated $N = 25$ variants of the generation capacity structure of the electric power system, each of which involved $K = 96$ separate MILP subproblems. This results in a potential parallelism level of $N \times K = 2400$ independent tasks, which closely correlates with the nearly linear performance gain observed.

A key advantage of the proposed parallel method is its flexibility in utilizing computing resources. This allows it to be adapted to various configurations — from a single machine to multi-node HPC systems with distributed memory. Such flexibility opens the way for further scalability and reduced computation time for larger instances of the generation capacity structure optimization problem.

Despite its advantages, the approach has certain limitations. First, it increases demand for computational resources, especially memory. Second, some variability in computation time was observed even under identical input conditions — a consequence of the stochastic nature of CMA-ES. While the global optimum was reached in most cases, the lack of guaranteed optimality in individual runs poses a potential risk.

So far, a detailed analysis of the impact of CMA-ES settings (such as initial standard deviation, population size, or mutation strategy) on convergence and accuracy has not been conducted. Such analysis is planned for future computational experiments. Another important research direction involves applying the proposed parallel method to more complex electric power system development problems — particularly those that account for controllable generation technologies, uncertainty scenarios, and/or multi-period planning.

In addition, comparing CMA-ES with other metaheuristics remains an open topic, as does assessing the potential for efficiently scaling the proposed parallel method to multi-node architectures with limited memory per node.

V. Conclusions

The achieved speedup factor of 7.84, combined with the high probability of finding the global optimum (97%), supports the conclusion that integrating the CMA-ES algorithm with parallel use of the SCIP solver is an effective parallel method for solving MILP problems related to the optimization of generation capacity structure in electric power systems.

The proposed decomposition of MILP problems demonstrates good scalability and can be adapted to more complex tasks. Using a metaheuristic algorithm to determine the generation capacity structure of electric power systems, combined with a deterministic MILP solver for calculating the operating modes of generation equipment under different loading conditions, ensures a productive balance between solution speed and accuracy.

At the same time, the proposed parallel method has certain limitations. These include increased memory requirements, variability in computation time, and a lack of guaranteed global optimality.

Further research should, first, investigate the sensitivity of the CMA-ES algorithm to key parameter settings and compare its performance with other metaheuristic algorithms; second, apply the proposed parallel method to long-term electric power system development problems involving controllable generation, and evaluate its performance in distributed computing environments.

In conclusion, the combination of the CMA-ES algorithm with the SCIP solver can be seen as a fast, flexible, and sufficiently accurate tool for solving strategic generation capacity planning problems in electric power systems — even under limited computational resources.

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Renewable Biopolymer Environmentally Safety Biodegradable Composites: Technological, Mechanical and Dielectric Properties

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***Abstract* – Environmentally safety composites based on renewable bioplastic bland of pea starch with the addition of reed and hemp waste as a filler. The influence of different content and morphology of reed and hemp waste on the set of technological and operational characteristics of the developed environmentally safety biodegradable composites was investigated. It is shown that the most optimal direction for obtaining strong environmentally safety biodegradable composites based on renewable bioplastic bland of pea starch is associated with the use of waste reed and hemp stalks, with its optimal content at the level of 60 wt.%. It should also be noted that the specific surface electrical resistance of samples of biodegradable environmentally safe thermoplastic compositions based on pea starch with organic dispersed fillers in the form of crushed waste reed and hemp stalks is 8-10 orders of magnitude lower than the similar indicator of a film made of polyethylene.**

Keywords – environmentally safety, biodegradable, composites, renewable biopolymers, organic waste, reed, hemp

I. Introduction

The growing development of the market for biodegradable polymer composites has intensified interest in starch modification aimed at producing competitive biodegradable materials based on such modifications [1,2]. The availability of raw materials enables extensive experimental research and the exploration of new formulations and compositions to enhance and optimize their physical, mechanical, and performance characteristics. In recent years, increasing attention has been paid to biodegradable and environmentally safe materials derived from renewable plant resources and synthesized by living organisms [3,4].

Currently, research is being conducted to develop a wide range of materials based on compositions containing various types of starch [5], starch-containing products [6,7], technical lignins, and proteins [8, 9]. The main drawbacks of such materials include brittleness, dimensional instability, and poor shape retention, which significantly limit their practical applications. To overcome these disadvantages, special substances fillers and plasticizers are incorporated into biodegradable and environmentally friendly starch-based compositions. The primary function of these additives is to improve the strength and elasticity of biodegradable, eco-safe materials and to prevent their brittle failure during use. Combining biopolymers with plant fillers allows obtaining environmentally safety composite materials with completely new technological and operational properties [10, 11]. Although traditionally special attention is paid to the effect of plant fillers on mechanical strength, which has led to their widespread distribution into reinforcing or reinforcing and non-reinforcing or inert, fillers are capable of having many other effects, which determine the high efficiency of their use: reduction of fluidity and volumetric thermal or chemical shrinkage, improvement of moldability and dimensional stability, reduction of flammability, increase of frictional or antifrictional, electrical, thermal-physical and other characteristics [12, 13]. In rare cases, plant fillers play the role of only volume fillers, which are introduced to reduce environmentally safety composite materials cost or reduce the volume of the biopolymer used [14, 15]. Plant fillers for the production of environmentally safety composite materials can be practically all materials existing in nature and created by man, including the polymers themselves, after giving them a certain shape or size – in the form of spheres, powders, with

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irregularly shaped particles, flakes, ribbons, fibers, etc., distributed in different ways and in different ratios in the polymer matrix [16]. Low- and high-molecular-weight organic compounds that are inert with respect to the plasticized polymer are recommended as plasticizers or hybrid modifier for biodegradable and environmentally friendly composites [17,18,19]. The most well-known plasticizer is glycerol, a hygroscopic substance that stabilizes moisture content in the final material. Since water itself is an effective plasticizer, this contributes to increased material elasticity. Previously, this plasticizing effect was mistakenly attributed solely to glycerol [20,21]. It is now known that the plasticizing effect of glycerol increases with its concentration within the material layer [22]. Moreover, the introduction of plasticizing agents influences not only the elasticity of the film material but also other performance characteristics, including electrostatic charging behavior [2]. However, information regarding the electrostatic properties of biodegradable and environmentally friendly composites derived from renewable plant-based raw materials remains largely unexplored. The relevance of this study lies in the need to develop strong and elastic biodegradable environmentally friendly starch-based composites with enhanced dielectric properties.

The objective of this article is to investigate the technological, mechanical, and dielectric properties of biodegradable and environmentally safe composites based on starch, as well as hemp and reed waste.

II. MATERIALS AND METHODS

The following materials are investigated in the article:

- thermoplastic composites based on pea starch with the addition of various types of additives. Pea starch obtained from a Ukrainian processing enterprise was selected as the material for this study. Starch is produced in large quantities as a by-product of industrial processing;

- waste of marsh reed and hemp stems, which were collected in the Kharkiv region and crushed in household electric blender to a size of 0.5-6 mm.

To produce a thermoplastic composite based on pea starch, the mixture requires preliminary preparation and stepwise blending with varying component ratios and mixing sequences. The mixtures were prepared in three stages: addition of the solvent and intensive mixing (at room temperature); gradual addition of the plasticizer (in portions at room temperature); addition of the initiator and cross-linking agent (in a water bath). During the preparation process, the mixture was intensively stirred until coagulates with a diameter of 3-5 mm were formed. The resulting material was then fed into the hopper of a laboratory extruder and extruded at temperatures not exceeding 145 °C. Samples of environmentally safe biodegradable composites based on renewable biopolymer materials, hemp and reed waste were obtained on a laboratory extruder with varying formulation composition and temperature regimes. At the first stage of the study, mixtures with different quantitative ratios of the following components were prepared: pea starch, solvent, plasticizer, initiator, crosslinking agent. At the second stage, the temperature regime for processing the mixtures was selected by experimental method.

The study of impact strength and breaking stress during bending of the environmentally safety biodegradable thermoplastic composites based on pea starch and composites based on renewable biopolymer thermoplastic pea starch, hemp and reed waste without notching at a temperature of 20 °C was carried out on a pendulum head according to ISO 180 and ISO 178, respectively, on samples conditioned under standard controlled temperature and humidity for those standards. Water absorption was measured following ASTM-D-570.

The dielectric properties of biodegradable and environmentally friendly thermoplastic composites based on starch and hemp and reed waste were investigated. For the measurements of surface resistivity (ρ), a specialized setup was used, consisting of a measuring chamber with an electrode system and a standard teraohmmeter EK-6-11. The EK-6-11 teraohmmeter is designed to measure electrical resistance in the range from 30 MΩ to 1000 TΩ. The test voltage can be adjusted from 2.5 to 500 V.

III. RESULTS AND DISCUSSION

Preliminary studies were conducted on mixtures with different component formulations. Three formulations were examined: ST1, ST2, and ST3. In the biodegradable and environmentally friendly thermoplastic composites based on formulation ST1, the solvent content predominates; in formulation ST2, the plasticizer content is dominant; and in formulation ST3, the initiator and cross-linking agent are present in the highest proportion. Photographs of the samples are presented in Fig. 1.

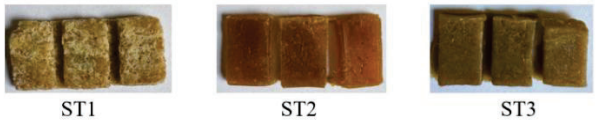


Fig. 1. Samples of biodegradable and environmentally friendly thermoplastic composites based on pea starch.

The physico-mechanical properties of the developed biodegradable and environmentally friendly thermoplastic composites are presented in Table 1. The obtained data show that the composites produced using different formulations exhibit varying strength characteristics. It should be noted that the flexural strength under static loading could not be determined for any of the three biodegradable thermoplastic composites developed according to formulations ST1, ST2, and ST3, due to the high elasticity of the material (see Fig. 2).

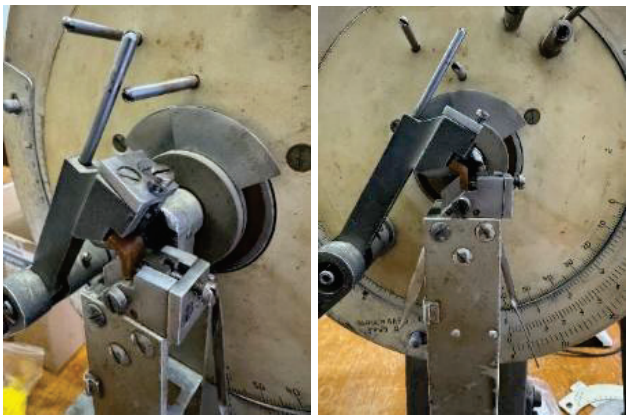


Fig. 2. Study of flexural strength under static bending of biodegradable environmentally friendly thermoplastic compositions based on pea starch.

TABLE I. PHYSICO-MECHANICAL PROPERTIES OF BIODEGRADABLE ENVIRONMENTALLY FRIENDLY THERMOPLASTIC COMPOSITES BASED ON PEA STARCH

Environmentally safety biodegradable thermoplastic composites based on pea starch	Strength indicators of composites based on pea starch	
	Breaking stress during bending, MPa	Impact strength, kJ /m^2
ST1	-	5.55
ST2	-	14.29
ST3	-	16.67

From the obtained data it is clear that the environmentally safety biodegradable thermoplastic composites based on pea starch made according to different formulations have different strength indicators.

It should be noted that the strength indicator during static bending for all three environmentally safety biodegradable thermoplastic composites based on pea starch developed according to formulations ST1, ST2 and ST3 could not be determined due to the high elasticity of the material. Analyzing the obtained results of environmentally safety biodegradable thermoplastic composites based on pea starch strength indicators, we can conclude that the composites obtained according to the ST1 recipe have the lowest impact strength at the level of 5.5 kJ/m^2, in contrast to the composites obtained according to the ST2 and ST3 recipes, which have indicators of 14.29 kJ/m^2 and 16.67 kJ/m^2, respectively.

The water absorption characteristics of the developed composites were also investigated (Fig. 3).

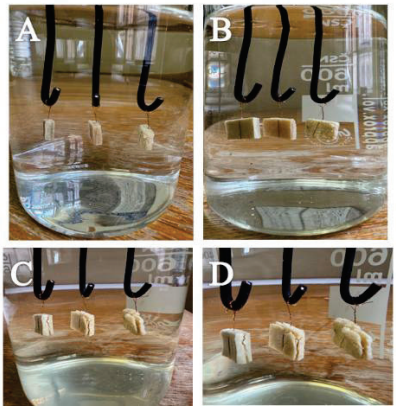


Fig. 3. Samples of biodegradable, environmentally friendly thermoplastic composites based on pea starch, presented from left to right as ST1, ST2, and ST3: A-B – at the beginning of 24-hour immersion; C-D – at the end of 24-hour immersion.

Starch itself is a hydrophilic material, meaning it has a high affinity for water. This is due to the presence of hydroxyl groups in its structure, which readily interact with water molecules. This property can be both an advantage and a drawback for starch-based biodegradable composites. Water absorption is a key factor that promotes the biodegradation of starch-based materials. Water facilitates the access of microorganisms and enzymes to the composite structure, thereby accelerating its decomposition in natural environments (e.g., soil or compost). At the same time, high water absorption may adversely affect the mechanical properties of the composites, reducing their strength, stiffness, and dimensional stability, especially under prolonged exposure to humid conditions. This limits their potential use in high-humidity environments or in applications requiring moisture resistance.Based on the conducted studies, the water absorption coefficient of the biodegradable, environmentally friendly thermoplastic composites based on pea starch after 24 hours was determined as follows: ST1 – 60 %, ST2 – 40 %, and ST3 – 17 %. These results indicate a high affinity for water and the potential for biodegradation under normal environmental conditions.

At the next stage of the research, interest was directed toward the development and investigation of biodegradable, environmentally friendly thermoplastic composites based on the ST2 formulation, incorporating organic dispersed fillers, specifically shredded reed (RW) and shredded hemp (HW). The choice of ST2 as the base formulation was due to its stable and homogeneous melt behavior during the homogenization stage, making it more suitable for processing and therefore considered the most technologically viable composition.

Considering the higher impact strength observed in ST3-based composites compared to ST2, future studies will focus on optimizing the processing parameters of ST3 and developing filled composites using identical fillers to enable comparative analysis.

Table 2 presents the physico-mechanical properties of the filled biodegradable, environmentally friendly thermoplastic composites based on pea starch ST2 with RW and HW fillers.

TABLE II. STRENGTH INDICATORS OF ENVIRONMENTALLY SAFETY BIODEGRADABLE THERMOPLASTIC COMPOSITES BASED ON PEA STARCH ST2 WITH ORGANIC DISPERSED FILLERS

Environmentally safety biodegradable thermoplastic composites based on pea starch	Strength indicators of composites based on pea starch	
	Breaking stress during bending, MPa	Impact strength, kJ /m^2
ST2 60 % wt. - RW 40 % wt	90.8	15.37
ST2 40 %wt. - RW 60 % wt.	93.8	8.43
ST2 60 % wt. - HW 40 % wt	6.17	11.69
ST2 40 %wt. - HW 60 % wt.	4.9	13.14

Based on preliminary research [15], the initial filler content ranges were selected for further investigation: environmentally safety biodegradable thermoplastic composites based on pea starch 60 wt.% – organic dispersed filler 40 wt.% and environmentally safety biodegradable thermoplastic composites based on pea starch 40 wt.% – organic dispersed filler 60 wt.%. The obtained physico-mechanical test results indicate that the incorporation of organic dispersed fillers – RW and HW into the biodegradable, environmentally friendly thermoplastic composites based on the ST2 pea starch formulation has a positive effect on their strength characteristics. In particular, the addition of fillers enhances the flexural strength of the composites under static bending, contributing to increased stiffness. At sufficiently high filler content, the impact strength remains at a high level.

Table 3 presents the specific surface electrical resistivity of samples of biodegradable, environmentally friendly thermoplastic composites based on the ST2 formulation with RW and HW organic fillers, under varying plasticizer contents.

TABLE III. EFFECT OF GLYCEROL ON THE SPECIFIC SURFACE ELECTRICAL RESISTIVITY OF BIODEGRADABLE ENVIRONMENTALLY FRIENDLY THERMOPLASTIC COMPOSITES BASED ON PEA STARCH FORMULATION ST2 WITH ORGANIC DISPERSED FILLERS IN THE FORM OF SHREDDED RW AND HW, P_s, Ω

Glycerol Content in Samples of Biodegradable Environmentally Friendly Thermoplastic Composites Based on Pea Starch Formulation ST2 with Organic Dispersed Fillers (Shredded RW and HW), wt.%		
4	8	15
P_S, Ω		
$1\text{-}3\cdot10^7$	$1\text{-}3\cdot10^6$	$1\text{-}3\cdot10^5$

As shown in Table 3, the samples of biodegradable, environmentally friendly thermoplastic composites based on the ST2 pea starch formulation with organic dispersed fillers in the form of shredded RW and HW exhibit higher electrical conductivity. The conductivity of the samples decreases with increasing glycerol content. This is likely due to the increased hygroscopicity of the film material associated with higher plasticizer content [3]. It is also noteworthy that the specific surface electrical resistivity of the biodegradable, environmentally friendly thermoplastic composites based on the ST2 formulation with RW and HW fillers is 8 to 10 orders of magnitude lower than that of polyethylene-based film.

IV. CONCLUSIONS

Environmentally safety composites based on renewable bioplastic bland of pea starch with the addition of reed and hemp waste as a filler. The influence of different content and morphology of reed and hemp waste on the set of technological and operational characteristics of the developed environmentally safety biodegradable composites was investigated. It is shown that the most optimal direction for obtaining strong environmentally safety biodegradable composites based on renewable bioplastic bland of pea starch is associated with the use of waste reed and hemp stalks, with its optimal content at the level of 60 wt.%. It should also be noted that the specific surface electrical resistance of samples of biodegradable environmentally safe thermoplastic compositions based on pea starch with organic dispersed fillers in the form of crushed waste reed and hemp stalks is 8-10 orders of magnitude lower than the similar indicator of a film made of polyethylene.

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Study of the pyrolysis process of household polymer waste to obtain functional fillers for dielectric composite materials

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***Abstract*— As a result of the study of liquid and solid products of pyrolysis of household polymeric waste by chromatography, mass spectral thermal desorption and infrared spectroscopy, it was found that their constituent components are destructed polyethylene terephthalates, polyvinyl chlorides and polystyrenes. It was found that the composition of the pyrolysis products of household polymeric waste mainly includes products of thermal destruction of polyethylene and polyvinyl chloride, and they are 90-95 % represented by unbranched alkanes and alkenes. It has been shown that in terms of dielectric characteristics, the solid residue of the pyrolysis of municipal polymer waste is a dielectric material with low losses and relatively low wave absorption ability and can be recommended as a functional dielectric filler for various types of polymer composite materials.**

Keywords—plant-polymer composites, secondary polyolefins, agricultural waste, climatic factors, properties

I. INTRODUCTION

Household polymeric waste («HPW») is classified as municipal solid waste («MSW»). There are two main methods of their elimination: passive - landfill disposal; active - incineration. The most widespread method is landfill disposal (98%) [1]. Despite the relative simplicity of this method, it has the following disadvantages: irreversible loss of scrap; withdrawal of large areas of land from circulation for a long period of time; significant costs for the necessary sanitary and epidemiological measures. The biggest disadvantage of solid waste disposal at landfills is the existence of a real danger of contamination of drinking water aquifers. These landfills fill up quickly due to the large volume and low density of the waste they contain. The looseness and low compressibility of plastic waste leads to frequent landfill fires

Among the active methods of solid waste elimination, the method of fire (flame) incineration is mainly used [2, 3]. Incineration is carried out in furnaces and furnaces of various designs. In this case, a distinction is made between layer-by-layer incineration of unsorted waste in the boiler furnaces of incinerators and chamber incineration of specially prepared (enriched) waste with a relatively stable fractional composition in the furnaces of power boilers or cement kilns. In practice, it has been established that flame incineration releases fine dust (25-50 kg/t of waste) and gases containing carbon dioxide, nitrogen and sulfur oxides, hydrogen chloride and fluoride, organic compounds (aldehydes and phenols), extremely toxic organochlorine compounds (dioxin and furan), and heavy metal compounds.

The analysis of modern methods of disposal of HPW shows that one of the most promising methods is their thermochemical decomposition (destruction) by dry pyrolysis. Compared to waste incineration, pyrolysis has a number of environmental and energy-saving advantages [3]. The most large-scale industrial pyrolysis technology in our country and in the world in terms of processing volumes is high-temperature coal coking (heating it without air access to a final temperature of 1000-1100 °C) [4]. The general theoretical foundations of the use of organic additives in coke production were developed under the guidance of Ukrainian

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scientist Y.B. Tyutyunnikov [5]. It has been suggested that plastic melts affect the colloidal structure of coal, causing peptization [6] - the destruction of the original structure with the formation of smaller fragments, which increases the reaction surface required for heterogeneous sintering processes. Changes in the group chemical composition of coal organic matter affect the reactivity and strength of coke [7]. However, if the amount of waste added to the coal charge is excessive (more than 1-2 %), the process is disrupted. This method is carried out without air access in special reactors and is more environmentally friendly and energy efficient. It is characterized by a high intensity of thermochemical transformation and an optimally constructed interconnection of controlled energy and product flows of the entire technological cycle, which ensures extremely high values of energy efficiency (86 %) [8,9]. In the process of dry pyrolysis, plastic waste is decomposed without air access into liquid and gaseous fractions with the release of a solid residue, the amount of which depends on the degree of waste contamination and the type of plastic [10,11]. The resulting pyrolysis gas is used to preheat the waste prior to pyrolysis. The liquid fraction containing light and heavy hydrocarbons can be used to produce liquid fuel and paraffin, the latter being used, for example, in the manufacture of cables and capacitors [12,13]. The solid residue can be used in construction as a good waterproofing agent; after mixing it with sand and gravel, a high-strength pavement can be produced [14,15]. A distinctive feature of HPW is their significant diversity in composition, as they can include various types of polymers with a very wide melting range and potential useful pyrolysis products.

The purpose of this article is to study the pyrolysis process of household polymer waste to obtain functional fillers for dielectric composite materials.

II. MATERIALS AND METHODS

A mixture of HPW consisting of polyethylene, polypropylene, polyethylene terephthalate, polyvinyl chloride, and polystyrene products in the form of films, trays, containers, and packaging was studied – Figure 1.



Fig. 1. A mixture of HPW consisting of polyethylene, polypropylene, polyethylene terephthalate, polyvinyl chloride, and polystyrene products in the form of films, trays, containers.

The prepared waste is pyrolyzed by heating without air access to a final temperature of 500 °C. The HPW pyrolysis technique included adding 100 g of HPW and placing it in a Wurtz flask, pre-weighed on a technical balance – Figure 2.

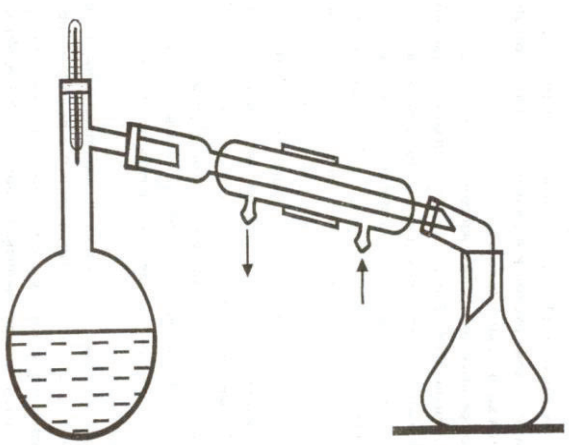


Fig. 2. The HPW pyrolysis installation included a Wurtz flask.

A water cooler and a receiver (also pre-weighed on a technical balance) are connected to the Wurtz flask.

Infrared spectra were recorded using a Nicolet 380 IR spectrophotometer (USA) at 20°C–25°C in the frequency range 4000–500 cm^{-1} using KBr-, NaCl-, and LiF- cuvettes with a layer thickness of 500–700 μm. For a more detailed study of the homologous and structural composition of these pyrolysis products, mass spectrometry was used. The study was conducted on a Bruker MALDI-TOF mass spectrometer (Germany) with a direct, heated to 250 °C device for direct introduction for the analysis of substances that are difficult to evaporate and their mixtures. The liquid products of HPW pyrolysis composition were determined by the «Shimadzu chromatograph» (Japan). Evaporator temperature – 240 °C; detector temperature – 250 °C; speed of carrier gas (hydrogen) – 1.0 ml/min., flow division – 1:60. The dielectric constant is determined according to DIN 53458.

III. RESULTS AND DISCUSSION

As a result of HPW pyrolysis, three main fractions were obtained: gaseous, liquid, and solid. An analysis of the chromatogram data in Fig. 3 shows that the pentane fraction accounts for the largest part of the liquid products of HPW pyrolysis, accounting for about 18.3% by mass, and a significant part of the products are unsaturated hydrocarbons 2-methylpentene-2 and 2,4-dimethylheptene-1 – Figure 4.

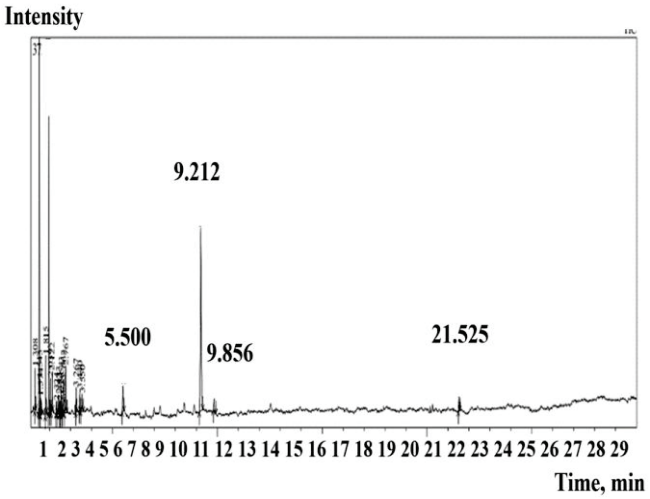


Fig. 3. Chromatographic of liquid pyrolysis products of HPW.

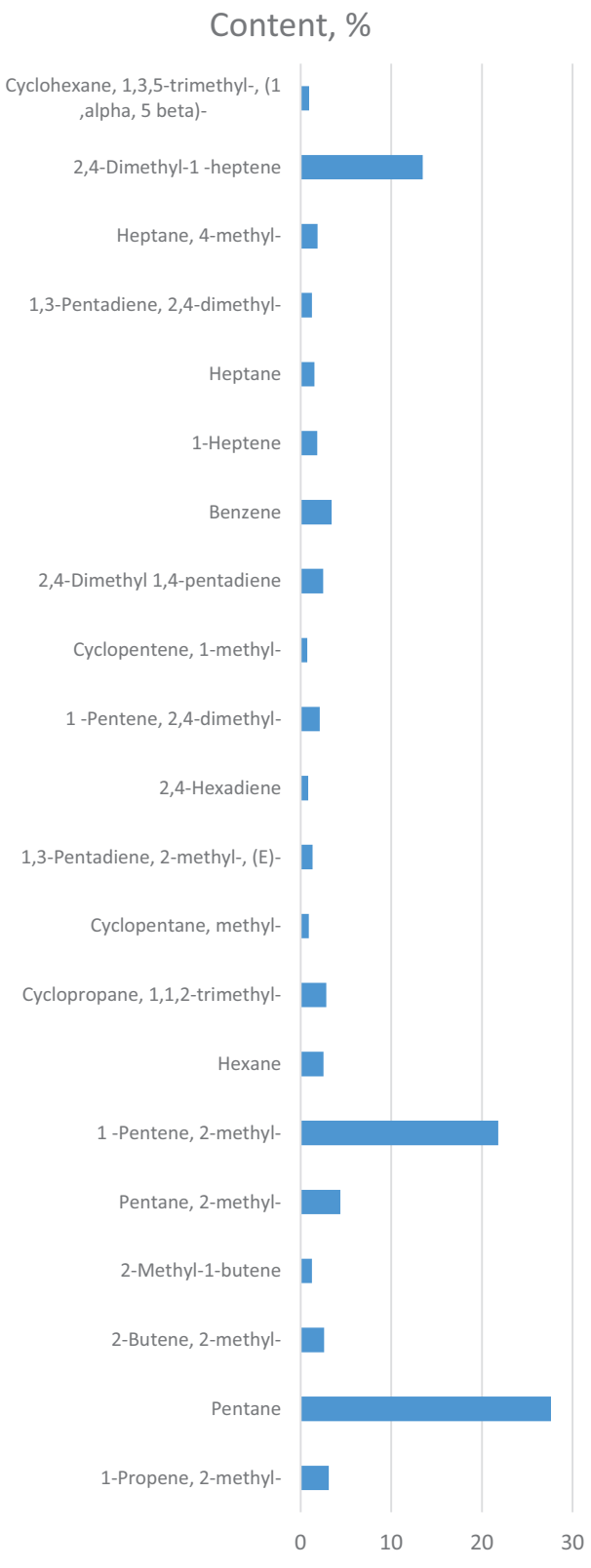


Fig. 4. Results of chromatographic analysis of liquid pyrolysis products of HPW.

The infrared spectra of all liquid and solid pyrolysis products show only absorption bands of CH_2 and CH_3 groups (characteristic absorption bands at 725, 1382, 1475, 2810-3250 cm^{-1}). The absorption band at 730 cm^{-1} corresponds to methylene groups in the chain $(-CH_2-)n>4$ at 900-990 cm^{-1} — the terminal vinyl group ($R-CH=CH_2$); at 880 cm^{-1} — to the terminal methylene group (vinylidene) [$R_1(R_2)=C=CH_2$]; the band at 975 cm^{-1} corresponds to trans-alkenes of the type $R_2-CH=CH-H_2$, and the band at 645-735 cm^{-1} — to cis-alkenes. The absorption band at 915 cm^{-1} corresponds to the vibrations of CH_2 in the $CH_2=CH-R$ group of a linear α-olefin, and the weak band at 1625-1685 cm^{-1} corresponds to the unbound valence vibrations of ethylene oligomers. All these groups are characteristic of unbranched alkanes and alkenes and, as a rule, they are found in high-molecular compounds, as evidenced by the presence of a band at 1000 cm^{-1} and a blurred broad band at 3150-3650 cm^{-1} of associated ethylene and vinylidene structures (dimers, block polymers), as well as, possibly, newly formed polyethylene terephthalate associates. In addition, there are bands characteristic of chlorinated alkanes and alkenes with a carbon atom number in the molecule of $C_{14}-C_{20}$ and above. These are fragments of chlorinated ethylene and vinylidene, or vinyl chloride derivatives.

Thus, the main components of HPW are the products of thermal decomposition of polyethylene, polypropylene, and polyethylene terephthalate, which are mainly represented by unbranched alkanes and alkenes. Assuming that pyrolysis products may contain initial monomers, as well as cluster structures of different molecular weights containing them, the mass spectral thermal desorption method was used to extract them from pyrolysis products. According to [5,16,17]], three main types of hydrocarbons were identified by mass spectrum: the intensities of homologous series with masses of 72, 87, 98, 115 characterize the presence of paraffins; 70, 84, 96, 112, 127 - monoolefins; 68, 82, 94, 108 - diolefin hydrocarbons.

Calculation of mass spectra using the method of analysis of paraffin cracking products of liquid pyrolysis products [6] provides information on the structural-group and homologous composition of hydrocarbons desorbed from pyrolysis products of polymer materials (Table 1).

TABLE I. RESULTS OF CHROMATOGRAPHIC ANALYSIS OF LIQUID PYROLYSIS PRODUCTS OF HPW

Test	Paraffins % (relative)	Monoolefins % (relative)	Diolefins % (relative)
Liquid products 30-150 °C	31.6-38.7	54.3-56.7	6.7-12.7
Dry matter	37.6-38.5	56.5-57.0	6.8-6.4

The results of infrared spectra analyses of liquid and solid pyrolysis products show that all tested pyrolysis product samples contain monoolefin (54.3-56.7%), diolefin hydrocarbons (6.7-12.7%) and paraffins (31.6-38.7%). This confirms the fact that the destroyed polymer waste is made from ethylene, i.e., its main component is a polyethylene structure. In high- and low-pressure polyethylene, there may be several branches per thousand CH_2 groups. Diene and paraffin hydrocarbons in pyrolysis products were formed as a result of thermodestructive processes. The intensity of the CH_2 group deformation vibration band at 1480 cm^{-1} is significantly higher than that of the CH_3 group vibration band (1460 cm^{-1}).

Next, the dielectric components of solid HPW pyrolysis products were studied to determine their potential for use as dielectric fillers for obtaining polymer composite materials with a wide range of functions. Fig. 5 shows the dependence of the dielectric loss tangent for solid HPW pyrolysis products on temperature.

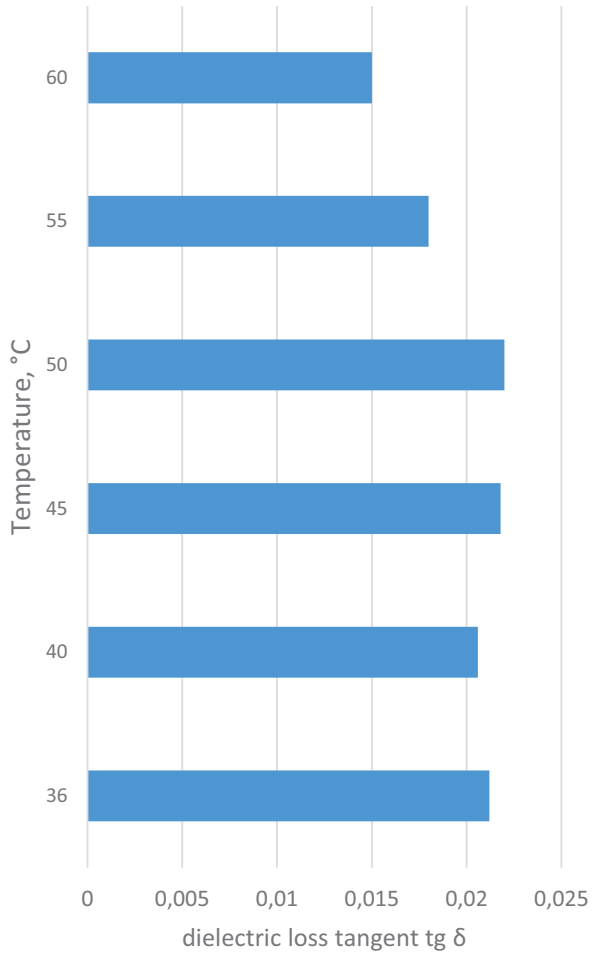


Fig. 5. Dependence of the dielectric loss tangent tg δ level for solid products of HPW pyrolysis on temperature.

The data in Fig. 4 show that the tangent of the dielectric loss angle for the solid residue of HPW pyrolysis decreases from 0.0212 to 0.015 in the temperature range of 36–60 °C. Such a level of dielectric characteristics of the solid residue of HPW pyrolysis allows it to be called a dielectric material with low losses and relatively low wave absorption capacity. In general, the obtained solid residue of HPW pyrolysis can be recommended as a functional dielectric filler for various types of polymer composite materials.

IV. Conclusions

As a result of the article, a study of liquid and solid products of HPW pyrolysis was carried out using chromatography, mass spectral thermodesorption, and infrared spectroscopy methods. It was found that their constituent components are destructed polyethylene terephthalates, polyvinyl chlorides, and polystyrenes. It was established that HPW consists mainly of products of thermal destruction of polyethylene and polyvinyl chloride, 90-95% of which are unbranched alkanes and alkenes. It has been shown that, in terms of dielectric characteristics, the solid residue of HPW pyrolysis is a dielectric material with low losses and relatively low wave absorption capacity and can be recommended as a functional dielectric filler for various types of polymer composite materials.

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Experimental Studies with an Open Resonator for Measuring Chemiluminescence of Bioobjects

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***Abstract*— The paper presents experimental studies using an open resonator to measure the chemiluminescence of biological materials. The study includes experimental verification of theoretical data and conclusions on the organization of the resonance system for chemiluminescence measurements. In addition, the characteristics of the resonator's electronic system are analyzed to assess the influence of EMF on sunflower seed yield using chemiluminescence as a measure.**

Keywords—Biological materials, Chemiluminescence, Optical resonators, Seeds

I. INTRODUCTION

In order to ensure the national interests of Ukraine regarding the sustainable development of the economy, civil society and the state, to achieve an increase in the standard and quality of life of the population, it is advisable to conduct research related to increasing the yield of agricultural crops by substantiating a resonator system for measuring the chemiluminescence of biological objects to determine the biotropic parameters of the EMF.

Based on the generalization of factual material from domestic [1]–[3] and foreign [4], [5] scientific publications, it follows that for pre-sowing treatment of agricultural crop seeds it is necessary to use an information electromagnetic field of the extreme high frequency (EMF EHF) range, which will increase the yield and quality of agricultural crop seeds.

To determine the biotropic parameters of EMF in the EHF range (frequency, power, exposure and modulation), which cause an increase in the yield and quality of seeds of agricultural crops, theoretical studies of a seed model are necessary, which take into account the electrophysical parameters of seeds and the information-resonance effects of EMF on the membrane potential of cells of seeds of agricultural crops [6], [7].

Intensification of electromagnetic processes in plant growing, based on the use of information EM radiation on objects, is possible only with the introduction of automated systems into the technological cycle for determining the biotropic parameters of EMF, causing an increase in the yield of agricultural crops [8]–[10].

It has been determined that of the numerous parameters characterizing the state of a biological object during its interaction with an EMF, chemiluminescence is the informational one. From an analysis of existing methods of measuring the parameters of materials and substances [11], [12], it follows that most of the considered devices for registration of chemiluminescence cannot provide high sensitivity; they are characterized by the lack of unification and uniform metrology and, in addition, none of the considered devices allows recording both spontaneous and induced by various physical factors chemiluminescence, or conducting measurements in real time with the direct impact of EMF of microwave range on biological objects.

Taking into account the above, studies devoted to the justification of a resonator system for measuring chemiluminescence of bioobjects to determine the optimal biotropic parameters of EMF used for pre-sowing treatment of seeds of agricultural crops to increase the yield and quality of seeds are relevant.

II. RELATED WORK

As the analysis of works [13]–[16], devoted to the action of EMF on plants and animals, shows, the studies in which a complex including a biological object and technical devices is studied are relevant.

In works [17], [18] a characteristic feature of such a complex is the occurrence in it of targeted processes of control of both biological objects and all types of influence on it on the basis of a priori and current information about its properties. Based on the research conducted in works [19], [20] it is possible to study the responsible reactions of biological systems to the information impact of EMF. At low-energy levels of EMF affecting a biological object, the measurement system should be distinguished by high sensitivity, measurement accuracy, selectivity and integral assessment of the measured value, universality in the choice of the object of study, high reliability and non-interaction of sensors with the studied biological substance, continuous recording of the measured value. Technical systems for monitoring changes in the chemiluminescence of biological objects should measure both spontaneous and induced by various physical factors biochemiluminescence, as shown in the works [21]–[23].

As noted in the literature [24], [25], very weak light fluxes arise during chemiluminescence, and therefore the intensity of ultra-weak glow when studying the impact of EMF on the most important processes of life of biological objects can be

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estimated using a photomultiplier tube (PMT). PMTs convert a light signal into an electrical signal (electrical pulses). The number of pulses is proportional to the number of quanta falling on the photocathode. Then the pulses are amplified and recorded. The functional diagram of the system for measuring the chemiluminescence of biological objects using a PMT includes: an EHF generator, a measuring converter, a recording device, an interface board, a personal computer and a printing device.

The study of the mechanisms of action of information electromagnetic radiation on sunflower seeds to increase their yield and quality led to the need to verify the reliability of theoretical approaches and models describing the process of interaction of EMF with seeds, as well as to confirm the possibility of implementing an electronic resonator system and its use to determine the optimal biotropic parameters of EMF by measuring the chemiluminescence of seeds.

III. Purpose of the Work

The aim of the research is to conduct experimental studies with an open resonator (OR), measuring the chemiluminescence of biological objects (sunflower seeds) to determine the optimal values of biotropic parameters of the EMF pre-sowing seed treatment, which will make it possible to increase crop yields.

In accordance with the purpose of the experimental research, the tasks of the experiment and the ways to solve them were determined:

- experimental verification of theoretical data and conclusions;

- analysis of the characteristics of a resonator electronic system for monitoring the assessment of the degree of influence of EMF on the yield of sunflower seeds based on the magnitude of their chemiluminescence.

IV. Materials and Methods

The block diagram of the experimental setup, with the help of which the studies of the hemispherical OR were carried out, is shown in Fig. 1, and the general view is shown in Fig. 2.

The measurements were performed in the eight-millimeter wavelength range ($f = 36.8$ GHz). The OR is formed by a spherical focusing mirror 4 with a curvature radius of $R = 100$ mm and the same aperture and a flat mirror with an aperture whose diameter $2a_1$ is also equal to 100 mm. The resonator is excited through a coupling slot with dimensions of 7.2×0.16 mm in the center of the spherical mirror by a high-frequency generator 1. For additional decoupling of the generator and the resonator, an adjustment attenuator 2 is included in the circuit. The system is tuned to resonance by moving the spherical mirror 4 along the resonator axis. The distance between the reflectors is determined with an accuracy of 0.001 mm. The supply waveguide with a cross section of 7.2×3.4 mm is oriented in such a way that the vector $\bar{E}$ of the fundamental wave in the waveguide is perpendicular to the plane of the pattern. The signal from the OR is transmitted using the second slot coupling element, which is also made on a spherical mirror and having the same geometric dimensions as the first, but spaced 5 mm from the latter.

The signal, having passed through the OR with a section of rectangular waveguide with a cross-section $a \times b$, enters the receiving path, consisting of a measuring polarization attenuator 8, detector 10, resonant amplifier 12 and dual-beam oscilloscope 13. To measure the reflection coefficient from the resonator and to control the frequency of generator 1, an additional receiving path is provided in the circuit. This path includes: directional coupler 3, adjustment attenuator 5, resonant wavemeter 7, detector 9, resonant amplifier 11 and oscilloscope 13. To compensate for the signal reflected from the OR input, a matched load 6 is included in one of the arms of the directional coupler.

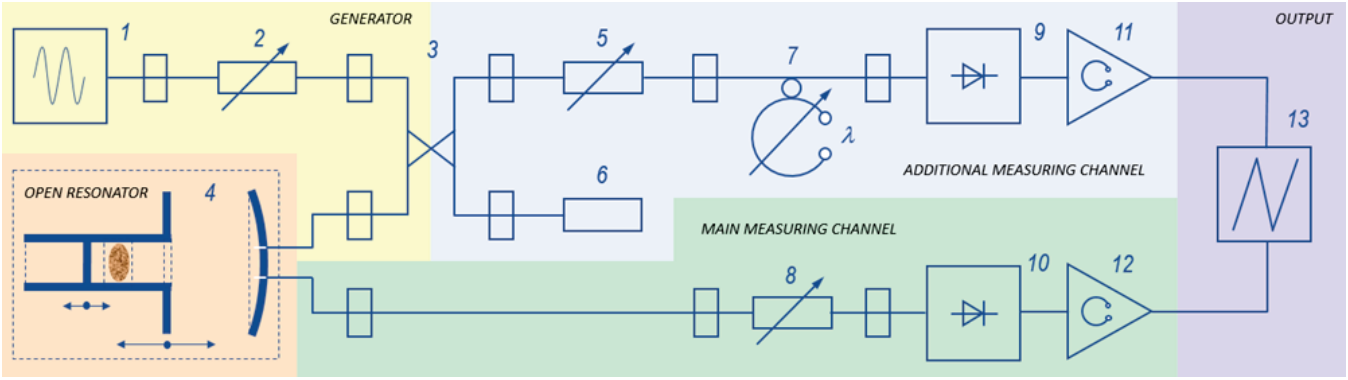


Fig. 1. Block diagram of the experimental setup

The following method was used to find the resonant transfer coefficient. After tuning the system to resonance, we fix the signal level at the OR output using the measuring polarization attenuator 8. Let it be N_1, dB. After conducting the entire measurement cycle at the output of the directional coupler 3, we install a receiving path instead of the OR, which includes the attenuator 8, detector 10 and resonant amplifier 12, and, changing the introduced attenuation into the path using the polarization attenuator, we achieve the same signal level on the resonant amplifier scale. Let the signal level in this case be N_2, dB. Then the field transmission coefficient $K_p = 10^{-\Delta N/20}$, where $\Delta N = N_2 - N_1$, dB. The reflection coefficient from the resonator G_p is measured in a similar way.

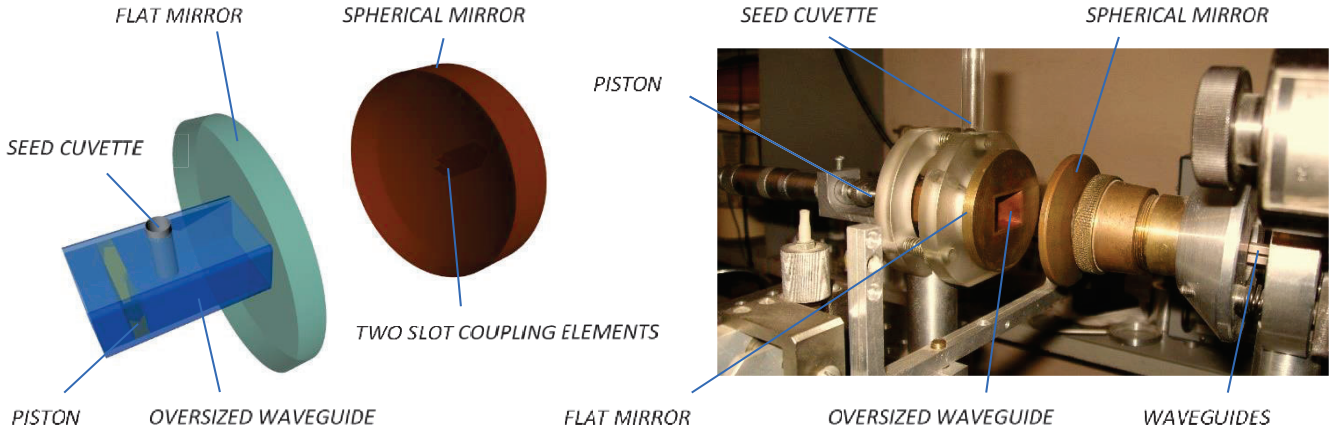


Fig. 2. External view of the experimental setup where the studies of a hemispherical open resonator with a section of oversized waveguide were conducted

In conducting our research, we proceed from the following. The fundamental oscillation TEM_{00q} in the resonator is excited using a slot coupling element located in the center of the spherical mirror (Fig. 1). In the considered OR, with this method of power supply, higher axially symmetric types of oscillations TEM_{10q} and TEM_{20q} should also be excited (in Laguerre-Gauss functions). The amplitude distribution of the fields of these oscillations on the curvilinear reflector of the hemispherical OR of the geometry under consideration is described by the expressions:

- for TEM_{10q} oscillation [26], [27]:

$$E_1(x,y) = A_{10} L_1^0\left(\frac{2\left(x^2+y^2\right)}{20{,}129^2}\right)\exp\left(-\frac{\left(x^2+y^2\right)}{20{,}129^2}\right), \quad (1)$$

where A_{10} – amplitude coefficient,

$L_1^0\left(\frac{2\left(x^2+y^2\right)}{20{,}129^2}\right)$ – generalized Laguerre polynomial;

- for TEM_{20q} oscillation [27], [28]:

$$E_2(x,y) = A_{20} L_2^0\left(\frac{2\left(x^2+y^2\right)}{20{,}129^2}\right)\exp\left(-\frac{\left(x^2+y^2\right)}{20{,}129^2}\right). \quad (2)$$

Here, as it was above, A_{20} – amplitude coefficient, $L_2^0\left(2\left(x^2+y^2\right)/(20{,}129)^2\right)$ – generalized Laguerre polynomial.

In the presence of two slot coupling elements on the surface of a spherical reflector (Fig. 1), the axial symmetry of the resonant system is disrupted. This leads to the removal of degeneracy, since the oscillations under consideration are generally degenerate [28], [29].

Therefore, each of the higher oscillations obtained from formula (1) and (2) splits into two: TEM_{10q} – on TEM_{20q} and TEM_{02q} (in Hermite-Gaussian functions); TEM_{20q} (in Laguerre-Gauss functions) – on TEM_{40q} and TEM_{04q} (in Hermite-Gaussian functions). That is why we considered the excitation of the higher oscillation TEM_{20q} (in Hermite-Gaussian functions) by the waveguide wave TE_{10}. The amplitude distribution of the field of this oscillation on the curvilinear reflector of the hemispherical OR of the geometry under consideration is described by the expression [30], [31]:

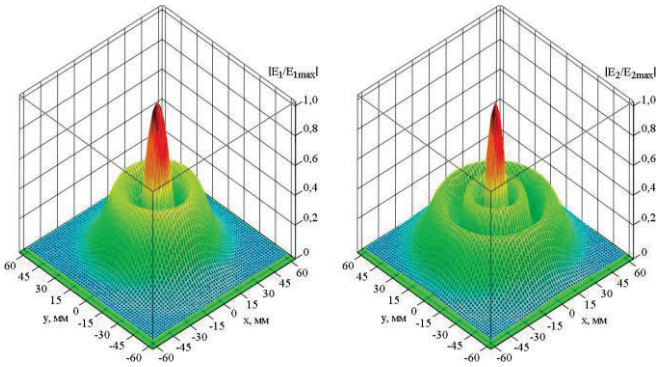
$$E_3(x,y) = C_{20} H_2\left(\frac{\sqrt{2}x}{20{,}129}\right)\exp\left(-\frac{\left(x^2+y^2\right)}{20{,}129^2}\right), \quad (3)$$

where C_{20} – amplitude coefficient, $H_2\left(\frac{\sqrt{2}x}{20{,}129}\right)$ – Hermite polynomial.

V. Results and Discussion

A. Results of verification of theoretical data and conclusions.

The results of calculations according to formulas (1) and (2) are presented in Fig. 3. The results of calculation according to formula (3) are shown in Fig. 4.



a) b)

Fig. 3. Amplitude distribution of the field of axially symmetric oscillation types TEM_{10q} (a) and TEM_{20q} (b) on the spherical mirror of the resonator

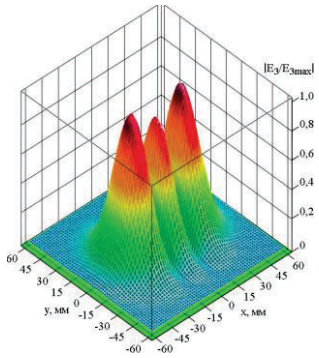


Fig. 4. Amplitude distribution of the TEM_{20q} oscillation field on a spherical mirror of an open resonator of the considered geometry

The presence of a section of an oversized rectangular waveguide of the above dimensions in the center of the flat mirror of the resonator leads to the following. Due to the principle of reciprocity, the wave TE_{10} will be excited in this section of the waveguide with an efficiency of 88% using the TEM_{00q} resonator oscillation. Consequently, only one waveguide wave will propagate in this waveguide, since almost all the supplied power is used to excite it.

If the length of this waveguide section is a multiple of an integer number of waveguide wavelengths λ_w ($\lambda_w = \lambda \Big/ \sqrt{1-(\lambda/2\alpha)^2}$), then 88% of the power transferred by the wave TE_{10} goes to maintaining the main oscillation TEM_{00q} in the hemispherical OR. Due to this, both angular and frequency selection of the spectrum of the resonant system under consideration should be ensured.

At the first stage of the research, piston is positioned flush with flat mirror (Fig. 1). The measurements are carried out at a fixed frequency f=36.8 GHz. The spectrum section of the hemispherical OR of the considered geometry is shown in Fig. 5. Identification of the types of oscillations excited in a resonant system is carried out using the test body method [32], [33].

The scattering probe is a metal sphere with a diameter of 1.6 mm, fixed on a nylon thread with a thickness of 0.2 mm, which is pulled through the resonator perpendicular to its axis. The amplitude distribution of the fields of excited oscillations is determined in the first antinode of the electric component of the standing wave field in the resonator in two mutually perpendicular planes $x0z$ and $y0z$, counting from the flat mirror. It is evident from the figure that for the given range of variation of the normalized distance between the mirrors in the hemispherical OR of the considered geometry, the fundamental oscillation with different longitudinal indices is excited: TEM_{0016} (L/R=0,676), TEM_{0017} (L/R=0,716), TEM_{0018} (L/R=0,758).

As noted above, the violation of the axial symmetry of the resonant system leads to the fact that each of the TEM_{10q} oscillations and TEM_{20q} (in Laguerre-Gauss functions) (Fig. 3) breaks down into two oscillations, shifted in space relative to each other by 90° (Fig. 5). Let us consider the TEM_{0017} oscillation, which, based on what was said above, should have the highest quality factor in a hemispherical OR. The measurement of the value of the loaded quality factor Q_L of this oscillation in a hemispherical OR was carried out using the half-width of the resonance curve [34], [35].

As it turned out, the measured value $Q_L \approx 5370$. In this case, the relative root-mean-square error measurements of loaded quality factor does not exceed 6% [36], [37].

Now we move the piston three waveguide wavelengths deep into the waveguide (Fig. 1). The calculated value of the length of the waveguide segment h in the center of the flat mirror is $3\lambda_w = 25.221$ mm. At the same time, the experimentally measured value is $h = 25.318$ mm, i.e. the difference from the calculated value does not exceed 0.4%. This is an indirect confirmation of the fact that the TE_{10} wave propagates in the oversized rectangular waveguide. The section of the spectrum of the hemispherical OR with a segment of a rectangular waveguide having transverse dimensions $a = 31.4$ mm, $b = 21.8$ mm and depth $h = 25.318$ mm is shown in Fig. 6.

In this case, measurements, as above, are carried out at a fixed frequency of $f = 36.8$ GHz.

It is evident from the figure that in the considered OR in the tuning range of the wavelength ($0.676 < L/R < 0.758$) only one oscillation TEM_{0023} is excited. Consequently, it can be said that the resonator of the given geometry has a unimodal resonance curve. For this case, we also estimated the value of the loaded quality factor of the TEM_{0023} oscillation. The measurements carried out on the half-width of the resonance curve showed that the loaded quality factor of the oscillation in question is equal to 3120.

The decrease in Q_L is associated with an increase in ohmic losses in the walls of the section of the oversized rectangular waveguide located in the center of the flat mirror OR. At the same time, a not very significant decrease in the loaded quality factor of the oscillation under consideration (approximately 1.7 times) is due to an increase in the volume of energy stored in such a resonant system, again due to the section of the rectangular waveguide with a length of $3\lambda_w$.

In order that the disturbance introduced by the tube into the resonant system is not too strong, it is located at a distance of $g = 5.75$ mm from the side wall of the waveguide (Fig. 1). For such a resonant system, we measured the value of Q_L when the TEM_{0023} oscillation is excited in the resonator. As it turned out, for this case, the loaded quality factor of the oscillation under consideration is equal to 1740. This allows us to say that the resonant system under consideration can be used to study the effect of EHF radiation on the seeds of agricultural crops.

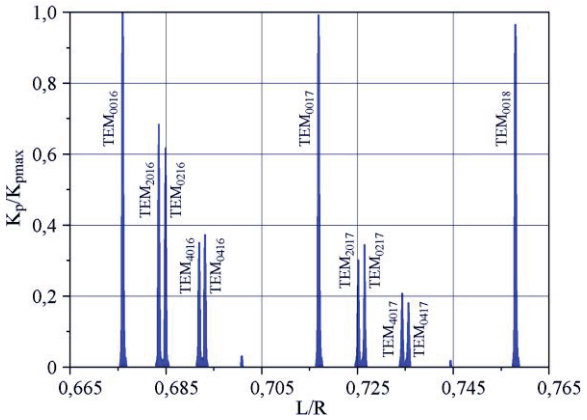


Fig. 5. Section of the spectrum of a hemispherical resonator, connected according to the pass-through scheme, with two slot coupling elements on a curvilinear reflector

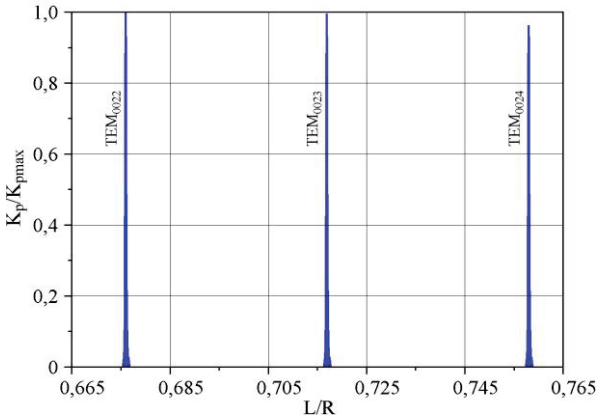


Fig. 6. A section of the spectrum of a hemispherical resonator connected in a pass-through configuration with a section of an oversized rectangular waveguide

B. *Results of the development of a system for measuring the chemiluminescence of sunflower seeds when they are irradiated with information EMF of the EHF range.*

As a result of theoretical and experimental studies, a prototype of a device for measuring the chemiluminescence of sunflower seeds was manufactured.

The main elements of the device are: a photodetector, a resonator device with a seed cuvette, an amplifying unit and a high-voltage power supply. A photomultiplier FEU-79 is used as a photodetector, converting the light flux of seeds into electrical pulses, the number of which is proportional to the number of light quanta falling on the photocathode. The FEU operates in the spectral sensitivity region (370...830 nm).

Optimal conditions for operating the FEU are when the FEU is in a light-tight, hermetic case. The FEU casing in the device is made of a prefabricated assembly consisting of four walls and two covers (upper and lower). The upper cover provides a panel for the FEU. The void formed between the walls of the casing and the FEU is filled with foam plastic balls. In the casing near the FEU, there is a bag with silica gel. Careful sealing prevents moisture condensation on the divider and the FEU cylinder. For normal operation of the electron multiplier, gradually increasing positive potentials are applied to its electrodes. Distribution of the supply voltage between the cascades is carried out using a divider. Electrical contact with the FEU is carried out using sealed light-tight connectors. The voltage divider circuit is given in the FEU passport. The value of individual resistances should be several hundred kilo Ohms. One of the ends of the divider, the anode, is grounded, and a large negative potential (1 kV) is applied to the cathode. This connection scheme is preferable from the point of view of safety engineering. In addition, it provides a smaller parasitic capacitance of the anode and simplifies the measurement of the input current and switching on the output to the matched cable. Since the photomultiplier operates in a pulse mode, large current pulses are selected from the emitters at the last stages. In order to avoid potential surges that can distort the required configurations of electrostatic fields and the established potential differences between the electrodes, shunting of the capacitor emitters, is used. Their value does not exceed 1 μF. Signal, removed from the anode of the photomultiplier tube, for further processing is fed to the amplifier unit, which includes a preamplifier and an amplitude discriminator.

Technical characteristics of the amplifier unit: input resistance 20…30 Ω; conversion coefficient > 12 mV/μA; bandwidth 100 Hz…35 MHz; instability of the discrimination threshold in the temperature range of –20 to +40ºC for 8 hours of operation < 3 mV; threshold voltage measurement limits 20…1000 mV; output pulse duration 10…20 ms. To ensure stable power supply of the photoelectron multiplier, a high-voltage power supply was developed. It ensures high stability of current collection, low pulsation level, and the ability to quickly adjust the output voltage directly during the operation of the device. The high-voltage power supply has the following technical characteristics: output voltage 500…2000 V; maximum load current 10 mA; ripple factor 0.003%. A measuring setup was built for automatic recording of biochemiluminescence, data processing and storage. The main units of the measuring setup are: a chemiluminescence recorder with an amplifying unit; a frequency meter used in the pulse counting mode; an interface board that allows communication between the frequency meter and a PC, and a personal computer itself.

The interface receives signals (M0…M35) from the frequency meter via the input buffer, which are transmitted to the PC input ports at the corresponding addresses. After the measurement is complete, the frequency meter sends a conversion end signal to the interface status register. The data analysis program is developed in C++ in the Microchip Studio environment and consists of two main components: the input data reading control component and the data processing component. The program's operation result is saved in an XML file.

VI. Conclusion

1. Determination of biotropic parameters of EM radiation for pre-sowing treatment of sunflower seeds should be carried out by chemiluminescence of seeds on the developed resonator installation, to the input of which a power of 390 mW is supplied.

2. Pre-sowing treatment of sunflower seeds should be carried out using information modulated EM radiation with the following parameters: EM radiation frequency: 37.3 ± 0.02 GHz; power flux density $22 - 0.5$ μW/cm^2; modulation frequency: 90 ± 10 Hz; modulation coefficient: 0.4 ± 0.05; seed irradiation exposure: 280 ± 5 s.

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Renewable Materials And Coffee Grounds Environmentally Safety Composites: Mechanical and Dielectric Properties

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***Abstract*—Environmentally safe composites based on renewable bioplastics and coffee grounds waste were investigated. The different coffee grounds' contents impact on the strength characteristics of the developed environmentally safe biodegradable composites was investigated. It was shown that the most optimal direction for obtaining environmentally safe, strength biodegradable composites based on bioplastics and coffee grounds is associated with 60 wt.%. coffee grounds add. It should also be noted that the specific surface electrical resistance of biodegradable environmentally safe thermoplastic compositions based on bioplastics and coffee grounds is practically analogous to the polyethylene film.**

Keywords—environmentally safety, biodegradable, composites, renewable biopolymers, organic waste, coffee grounds

I. Introduction

In the current challenging environmental situation, using biodegradable polymers for manufacturing consumer goods - such as coffee cups and containers for food, cutlery, disposable dishes, etc. - is a key strategy for reducing solid waste, as it ensures their rapid decomposition under the climatic factors and microorganisms' influence [1, 2]. Such an approach corresponds to sustainable development conception and ZeroWaste conception as well as to circular economy principles.

One possible direction for obtaining environmentally safe biodegradable materials is the modification of traditional polymers [3,4]. Combining a synthetic polymer with natural organic fillers (e.g. starch, rice husk, wood flour) or consumer waste can give the material a new set of properties, in particular an increased ability to rapidly biodegrade [5,6,7].

About 6 million tons of coffee are produced in the world annually. After its consumption, about 98% of coffee beans go into waste (coffee grounds, coffee waste powder). The most widely spread ways to use coffee grounds are as fertilizers and as material for biofuel production, and as abrasive filler in cosmetic products (soap, scrabs, etc.), for an art installation. Coffee waste used as an organic filler for biodegradable polymer composites is a new, promising way to decrease coffee waste volume.

Coffee grounds powder obtained after drying has the following composition, wt. %: cellulose – 8.6; hemicellulose – 36.7; lignin – 25-33; protein – 6.7-13.6; lipids – 10-20; polyphenols – 2.5; caffeine – 0.02; mineral component – 0.82-3.52 (nitrogen, potassium, phosphorus, magnesium) [8,9,10]. Due to the coffee's natural origin, low cost of coffee waste, as well as the coffee grounds composition, which contains active nutrients for microorganisms and fungi, as well as trace elements that contribute to soil fertilization, it is advisable to use it as organic fillers for biodegradable polymers.

There are several publications describing the coffee grounds using as a filler in biodegradable, environmentally friendly composite materials [8,9].

Patented compositions and methods of manufacturing products using coffee waste [11], which include coffee grounds powder and thermosetting resin (melamine, phenol, urea, or polyurethane) mixed with starch, calcium carbonate, talc, and fibrous filler. All ingredients are used in powder form with powder sizes close to the coffee grounds' particle sizes; coffee grounds are dried to a moisture content of 5 to 15%. The products are manufactured by thermoforming at a normal temperature near 110-150 °C. Rough and fine grinding of the workpieces is required for the final finishing. The disadvantage is that the products cannot be recycled after use, since the composite is based on a thermosetting resin. While

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no information is provided on biodegradation terms, it is known that thermosetting plastics are not easily biodegradable.

This patented, environmentally friendly bioplastic includes 10–20 wt.% coffee grounds, 47–65 wt.% of a polypropylene/polyethylene mixture, and 5–20% of an inorganic filler such as calcium carbonate, talc or mica, as well as technological additives. Disadvantages include poor biodegradation due to the high polymer content (up to 65 wt.%) and the potential formation of environmentally harmful microplastics due to incomplete decomposition of the composite material base.

A biodegradable environmentally friendly composite material containing coffee grounds powder [12,13], including polylactic acid, polybutylene adipate terephthalate, vinyl bis stearamide, erucylamide, polyethylene wax, and stearic acid salts, has been studied. Talc, calcium carbonate or wollastonite are used as inorganic fillers. The material is characterized by a complex composition, including five polymer components, which increases the cost of the final product and complicates the production process. No information is provided on the composite environmentally friendly material biodegradation.

The environmentally friendly composite has also been developed, which includes: 20-50 wt. % dried coffee grounds powder, 40-70 wt. % plastic selected from the group: polystyrene, polypropylene, polyethylene, 5-25 wt. % calcium carbonate, 3-10 wt. % mixture modifier consisting of a compatible agent and a binder [14,15]. The disadvantages are poor composite biodegradation due to the hydrophobic polymer's high content - up to 70 wt. % and environmentally harmful microplastic formation possibility due to incomplete material decomposition.

The work presented is relevant because there is a need to find ways to create strong, elastic, biodegradable, environmentally safe compositions based on biopolymer polymer with high dielectric properties.

The aim of the article is to study the strength and dielectric characteristics of biodegradable environmentally friendly compositions based on biopolymers using an organic filler such as coffee grounds waste.

II. MATERIALS AND METHODS

As a thermoplastic biopolymer matrix, we used film TM P1 based on polyhydroxyalkanoates.

The coffee grounds were pre-collected and dried in a drying oven for 4 hours at a temperature of 70℃ with stirring every 30-40 minutes, after which the temperature in the drying oven was increased to 105℃ and dried for another 20-30 minutes to maximize moisture removal from the filler [16]. After the drying stage, the coffee grounds lumps were formed. The lumps were formed during drying due to oily substances release that have increased adhesion and interact with each other. The lumps were crushed in a ceramic mortar to a homogeneous dispersed mixture (Fig. 1).



Fig. 1. Dispersed organic filler - coffee grounds.

Samples of environmentally safe biodegradable composites based on biopolymer materials and coffee grounds were obtained on a laboratory extruder with varying formulation composition and temperature regimes. Two optimal compositions of mixtures were proposed: thermoplastic biodegradable polymer/organic dispersed filler in the following ratios by weight, biocomposite 1 (BPC1) – 30/20 and biocomposite 2 (BPC2) – 20/30. When combined, the components were thoroughly mixed and before loading into the laboratory extruder looked as shown at Fig. 2



Fig. 2. Thermoplastic biodegradable polymer/organic dispersed filler mixture, BPC1.

The mixture was fed to laboratory extruder hopper and extruded at temperatures not exceeding 145 °C [17].

The study of impact strength and breaking stress during bending without notching of the environmentally safe biodegradable composites based on biopolymer materials and coffee grounds at a temperature of 20 ° C was carried out on a pendulum head according to ISO 180 and ISO 178,

respectively, on samples conditioned under standard controlled temperature and humidity for those standards.

Dielectric properties of environmentally safe biodegradable composites based on biopolymer materials and coffee grounds were studied. For *p* measurements, a special installation was used, which includes a measuring chamber with an electrode system and a standard teraohmmeter EK-6-11. The teraohmmeter EK 6-11 is used to measure resistance ρ_s in the range from 30 MΩ to 1000 TΩ. The test voltage can be changed from 2.5 to 500 V.

III. Results and Discussion

A study of samples was conducted using two different composite materials - BPC1 and BPC2. Technologically, the composites were processed without any special features, except for the loading phase, which took place in batches, so that there was no pouring of coffee grounds small fraction into the extruder loading zone and thermoplastic biodegradable polymer/organic dispersed filler mixture separation under the vibrations impact that are generated during the extruder operation and act on its hopper. The differences between the composites can be clearly identified. Fig. 3 shows a photo of samples BPC1 and BPC2.

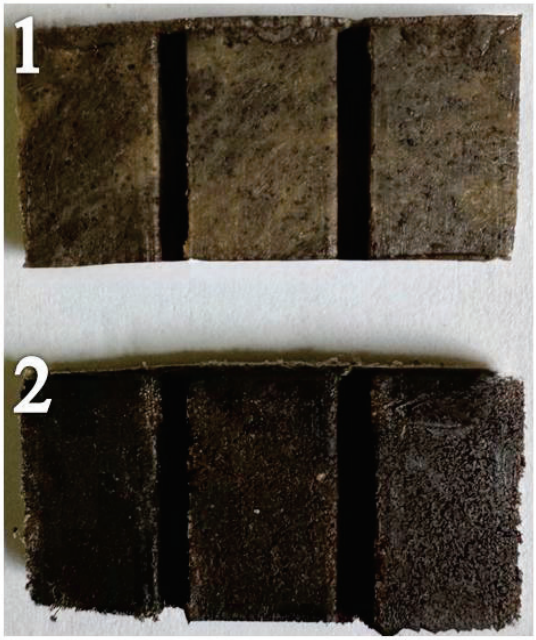


Fig. 3. Samples of thermoplastic environmentally friendly composites based on biodegradable polymer material and coffee grounds: 1) BPC1; 2) BPC2.

From the images (Fig.3), it is possible to determine the filler`s ability to be distributed in the polymer matrix by the surface appearance at the samples. The BPC1 composite contains a polymer matrix - 30 wt. parts to 20 wt. parts of filler. A small number of black coffee filler specks are visible on the BPC1 surface. This indicates that the filler is well dispersed in the polymer matrix and that the filler`s amount can be increased [18]. An increase in filler content to 30 wt.% and a change in the polymer matrix-to-filler ratio to 20:30 (BPC2) resulted in a change in the appearance of the composites, with a shift in color from grey to dark brown and an increase in the dispersed filler phase on the sample`s surface. Visual observations can identify the transition from the filled composite BPC1 to the highly filled BPC2.

At the next stage, the physical and mechanical properties of the developed environmentally safe composites based on the BPC1 and BPC2 were investigated, the data are presented in Table 1.

TABLE I. PHYSICAL AND MECHANICAL INDICATORS OF BIODEGRADABLE ENVIRONMENTALLY SAFE THERMOPLASTIC COMPOSITIONS BASED ON BIOPLASTICS AND COFFEE GROUNDS

Environmentally safety biodegradable thermoplastic composites	Strength indicators	
BPC1	Impact strength, kJ /m^2	5.99
	Breaking stress during bending, MPa	10.5
BPC2	Impact strength, kJ /m^2	4.76
	Breaking stress during bending, MPa	7.43

From the obtained data it follows that composite BPC1 has higher physical and mechanical properties than composite BPC2. It can be assumed that the increased strength properties of the BPC1 composite are associated with a larger polymer phase amount in the composite, which ensures the test samples homogeneity [19,20].

After physical and mechanical tests, the destroyed samples of environmentally safe composites BPC1 and BPC2 were examined. Fig. 4 presents photos showing the sample fracture site, where the composite internal structure is clearly observed.

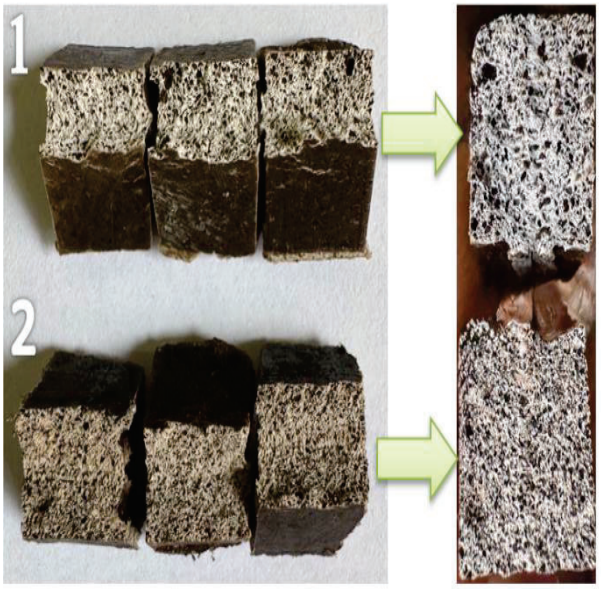


Fig. 4. The environmentally safety composites' internal structure: visualization: 1) BPC1; 2) BPC2

The photo shows that environmentally safe composites BPC1 have a more homogeneous structure of the polymer matrix (white) compared to environmentally safe composites BPC2, in which the polymer phase is evenly filled with a dispersed filler.

Thus, our assumption regarding the increased strength characteristics of environmentally safe composites based on BPC1, which is associated with a larger polymer phase amount, was confirmed.

Table 2 shows the specific surface electrical resistance of biodegradable environmentally safe compositions based on coffee grounds with different organic filler amounts.

TABLE II. COFFEE GROUNDS IMPACT ON THE SPECIFIC SURFACE ELECTRICAL RESISTANCE OF BIODEGRADABLE ECOLOGICALLY SAFE THERMOPLASTIC COMPOSITIONS, PS, Ω

Coffee grounds content in samples of biodegradable environmentally safe thermoplastic compositions, wt. %.		
0	40	60
ρ_s, Ом		
$1.1\text{-}3.2\cdot10^7$	$1.2\text{-}3.0\cdot10^7$	$1.1\text{-}2.9\cdot10^7$

Table 2 shows that samples of biodegradable environmentally safe thermoplastic compositions based on coffee grounds are characterized by practically the same electrical conductivity and, based on their level, can act as traditional insulating materials.

IV. Conclusions

Environmentally safe composites based on renewable bioplastics and coffee ground waste were researched. The different coffee grounds' contents impact on strength characteristics of the developed environmentally safe biodegradable composites was researched. It was shown that the most optimal direction for obtaining environmentally safe, strength biodegradable composites based on bioplastics and coffee grounds is associated with 60 wt.%. coffee grounds add. It should also be noted that the specific surface electrical resistance of biodegradable environmentally safe thermoplastic compositions based on bioplastics and coffee grounds is practically analogous to the polyethylene film.

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Interaction of Electromagnetic Radiation of the Extremely High Frequency Range with Seeds of Agricultural Crops

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***Abstract*—The paper presents theoretical studies of the interaction between extremely high frequency electromagnetic radiation and seeds of agricultural crops, with a focus on sunflower seeds. The main objective is to determine the optimal biotropic parameters of the electromagnetic field for pre-sowing treatment. The study addresses the evaluation of the electric field intensity inside the seeds as well as the determination of parameters of a modulated electromagnetic wave that exerts a biotropic effect. Based on the performed calculations, it is shown that when selecting the main parameters of the electromagnetic field, both the dielectric permittivity and the geometric characteristics of the seeds must be taken into account. The parameters of a low-energy (informational) electromagnetic field recommended for the pre-sowing treatment of sunflower seeds are identified.**

Keywords—Biological materials, Chemiluminescence, Resonators, Seeds

I. Introduction

Based on the generalization of factual material from domestic [1]–[3] and foreign [4], [5] scientific publications it follows that for pre-sowing treatment of seeds of agricultural crops it is necessary to use a low-energy (informational) electromagnetic field (EMF) of the extreme high frequency (UHF) range, which will increase the yield and quality of seeds of agricultural crops. To determine the biotropic parameters of the EMF of the EHF range (frequency, power exposure, modulation), causing an increase in the yield and quality of seeds of agricultural crops, theoretical studies of the seed model are necessary, taking into account the electrophysical parameters of seeds and the information-resonance effects of the EMF on the membrane potential of the cells of seeds of agricultural crops [6], [7].

II. Analysis of Literature Sources

To determine the biotropic parameters of EMF, a model must be built that allows determining the range of change of these parameters (frequency, power, exposure, modulation) for pre-sowing treatment of sunflower seeds [8]–[11].

As an exciting EMR we will consider an EM wave with an EF intensity changing according to the law [12]–[13]:

$$\vec{E}^{i} = \vec{E}_0 \left(1 + m \sin \Omega t\right) \sin\left(\omega t + k(\vec{n}, \vec{R})\right), \quad (1)$$

where $k = \omega / c$ – wave number, c – speed of light in a vacuum, $\vec{n}$ – a unit vector that determines the direction of wave propagation, ω – the circular frequency, Ω – modulation frequency, m – the modulation depth coefficient which characterizes the degree of change in amplitude, $\vec{E}_0$ – a constant vector, $\vec{R} = x\vec{e}_x + y\vec{e}_y + z\vec{e}_z$, $\vec{e}_x$, $\vec{e}_y$, $\vec{e}_z$ – the unit vectors of the Cartesian coordinate system.

It is assumed that $\Omega / \omega << 1$. In this case, we can talk about amplitude modulation of a harmonic wave with frequency ω.

We will model sunflower seeds as ellipsoids filled with a homogeneous isotropic medium with a relative permittivity of $\varepsilon = \varepsilon' - i\varepsilon''$ and magnetic permeability $\mu = 1$ [14]–[15]:

The problem is to determine the EMF that occurs inside a dielectric ellipsoid when it interacts with a modulated EM wave (1).

From the point of view of macroscopic electrodynamics, the original problem can be formulated as follows. It is required to find a solution to the homogeneous system of Maxwell's equations

$$rot\vec{H} = \frac{1}{c}\frac{\partial \vec{D}}{dt}, \; rot\vec{E} = \frac{1}{c}\frac{\partial \vec{H}}{dt} \quad (2)$$

$$div\vec{D} = 0, \; div\vec{H} = 0, \quad (3)$$

satisfying the initial and boundary conditions. The boundary conditions consist of the continuity of the tangential components of the electric field (EF) and magnetic field (MF) at the boundary of the dielectric ellipsoid:

$$\vec{E}_\tau^{+}\Big|_s = \left(\vec{E}_\tau^{-} + \vec{E}_\tau^{i}\right)\Big|_S, \; \vec{H}_\tau^{+}\Big|_s = \left(\vec{H}_\tau^{-} + \vec{H}_\tau^{i}\right)\Big|_S. \quad (4)$$

Here the signs "+" and "–" denote the limiting values of quantities, respectively, inside and outside the dielectric ellipsoid. $\vec{E}^{i}$, $\vec{H}^{i}$ – EMF of the exciting wave (1).

Equations (2)–(4) must be supplemented by material equations relating the electric field strength vector to the induction vector inside the dielectric ellipsoid, namely:

$$\vec{D} = \varepsilon \vec{E}, \quad (5)$$

ε – the relative permittivity of the medium filling the ellipsoid.

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Since the exciting wave depends harmonically on time, the solution of equations (2) will also have this property.

Taking this fact into account, we represent the EF strength of the exciting wave in the following form:

$$\vec{E}^{i} = \mathrm{Re}\left[\vec{E}_0 e^{ik(\vec{n}\vec{R})} \left(-ie^{iwt} + \frac{m}{2} e^{i(\omega-\Omega)t} + \frac{m}{2} e^{i(\omega+\Omega)t} \right) \right], \qquad (6)$$

where Re (…) denotes the real part of a complex number .

Since equations (2)–(3) and boundary conditions (4) are linear, this means that the superposition principle is valid for EMF. Then from (6) it follows that the solution to the formulated problem can be represented in the following form:

$$\vec{E} = \vec{E}^0 + \vec{E}_+ + \vec{E}_-, \quad \vec{H} = \vec{H}^0 + \vec{H}_+ + \vec{H}_- \qquad (7)$$

Here $\vec{E}^0$, $\vec{H}^0$ is the solution of problem (2)–(4) for the exciting wave:

$$\vec{E}_0^{\,i} = \mathrm{Re}\left[-i\vec{E}_0 e^{ik(\vec{n}\vec{R})} e^{i\omega t} \right] \qquad (8)$$

and $\vec{E}_\pm$, $\vec{H}_\pm$ – solutions of the same problem for exciting waves:

$$\vec{E}_\pm^{\,i} = \frac{m}{2} \mathrm{Re}\left[\vec{E}_0 e^{ik(\vec{n}\vec{R})} e^{i(\omega\pm\Omega)t} \right]. \qquad (9)$$

Let us introduce the notation $\omega_\pm = \omega \pm \Omega$. Then, using the complex form of writing of Maxwell's equations, we will have:

$$rot\vec{E}^0 = -i\frac{\omega}{c}\vec{H}^0, \; rot\vec{E}_\pm = -i\frac{\omega_\pm}{c}\vec{H}_\pm, \qquad (10)$$

$$rot\vec{H}^0 = i\frac{\omega}{c}\varepsilon\vec{E}^0, \; rot\vec{H}_\pm = i\frac{\omega}{c}\varepsilon\vec{E}_\pm, \qquad (11)$$

$$div\vec{E}_c^0 = div\vec{H}^0 = div\vec{E}_\pm = 0, \; div\vec{H}_\pm = 0. \qquad (12)$$

Algorithms for constructing solutions of equations (10)–(12) for frequencies ω, ω_+ and ω_- are similar. Therefore, in the following we will consider the case of frequency ω_+, and for ω_- we get the final result.

Thus, the original problem (4) is reduced to solving the following problem of diffraction of the exciting wave:

$$\vec{E}_+^{\,i} = \frac{m}{2}\vec{E}_0 e^{ik(\vec{n}\vec{R})} \qquad (13)$$

on a dielectric ellipsoid. The diffraction field $\vec{E}_+$, $\vec{H}_+$ must satisfy Maxwell's equations:

$$rot\vec{E}_+ = -i\frac{\omega_+}{c}\vec{H}_+, \; rot\vec{H}_+ = i\frac{\omega_+}{c}\varepsilon\vec{E}_+, \; div\vec{E}_+ = 0, \; div\vec{H}_+ = 0 \qquad (14)$$

and the boundary conditions (4) on the boundary of the ellipsoid, and at infinity – the Sommerfeld radiation condition [16]–[18]:

$$div\vec{H}_+ = 0, \; \lim_{R\to\infty} \mathrm{Re}\left[\vec{e}_R \times \vec{E}_+ - (\vec{e}_R \times \vec{H}_+) \right] = 0, \qquad (15)$$

where $\vec{e}_R$ – a unit vector in the direction ($R = \sqrt{x^2 + y^2 + z^2}$).

We will construct the solution to problem (13)–(15) using the method of integral equations [16], [17].

For this purpose, using the results from [14]–[18], we present equation (14) in the following form:

$$rot\vec{E}_+ = -i\frac{\omega_+}{c}\vec{H}_+, \; rot\vec{H}_+ = i\frac{\omega_+}{c}\vec{E}_+ + i\frac{\omega_+}{c}(\varepsilon-1)\vec{E}_+. \qquad (16)$$

Formally, these equations can be considered in the entire space, and the quantity $\vec{j} = i\frac{\omega_+}{c}(\varepsilon-1)\vec{E}_+$ can be treated as a certain distribution of the current over the region occupied by the ellipsoid. Then, according to [16], [18], equations (13) are equivalent to the following integral equation:

$$\vec{E}_+(p) = \vec{E}_+^{\,i}(p) + \frac{(\varepsilon-1)k_+^2}{4\pi}\int_V \vec{E}_+(q)G(p,q)dV_q + \qquad (17)$$

$$+ \frac{\varepsilon-1}{4\pi}\int_S \left(\vec{n}_q\vec{E}_+(q)\right) grad_q G(p,q)dS_q$$

Here V – the region of space occupied by the ellipsoid m, S – the boundary of the region V, p and q – points of the region V, $\vec{n}_q$ – the unit vector of the normal to the boundary S at point q, $\vec{E}_+^{\,i}(p)$ – the electric field strength of the exciting wave (13), function $G(p, q)$ – the Green's function for the scalar three-dimensional Helmholtz equation with wave number $k_+=\omega_+ / c$, which has the form:

$$G(p,q) = e^{-ik_+|p-q|} \big/ |p-q|,$$

$|p\text{-}q|$ – Euclidean distance between points p and q.

The unknown quantity in the integral equation (17) is the field strength inside the dielectric ellipsoid simulating a sunflower seed.

Thus, the integral equation (17) serves as the basis for the electrodynamic analysis of the interaction process of the exciting EM wave with sunflower seeds.

III. Purpose of the Work

The aim of the research is theoretical calculations on the interaction of electromagnetic radiation of the extremely high frequency range with seeds of agricultural crops, namely sunflower seeds, to determine the optimal values of biotropic parameters of EMF pre-sowing treatment, which will make it possible to increase crop yields.

In accordance with the purpose of theoretical research, the tasks of the experiment and the ways of their solution were defined: determination of the electric field strength inside sunflower seeds; determination of biotropic parameters of a modulated electromagnetic wave for pre-sowing treatment of sunflower seeds.

IV. Research Methods

A. Determination of the electric field strength inside sunflower seeds

As it was noted above, sunflower seeds are modeled by an ellipsoid filled with a homogeneous isotropic medium with a relative permittivity *of* $\varepsilon = \varepsilon' - \varepsilon''$. Fig. 1 shows an ellipsoid whose boundary surface is formed by rotating an arc of a circle of radius R around axis z. The geometric parameters of this arc of a circle are as follows: $2h_1$ – the length of the straight-line segment connecting the ends of the arc of a circle; $2h_2$ – the maximum distance between points of the boundary surface in the plane perpendicular to the axis z. Let us introduce a cylindrical coordinate system R, φ, z, with the z axis and the Cartesian coordinate system x, y, z as shown in Fig. 1.

Then the boundary surface of the ellipsoid can be described by the following parametric equations:

$$X = \bar{R}(v)\cos\varphi, \; X = \bar{R}(v)\sin\varphi, \; Z = \bar{Z}(v). \qquad (18)$$

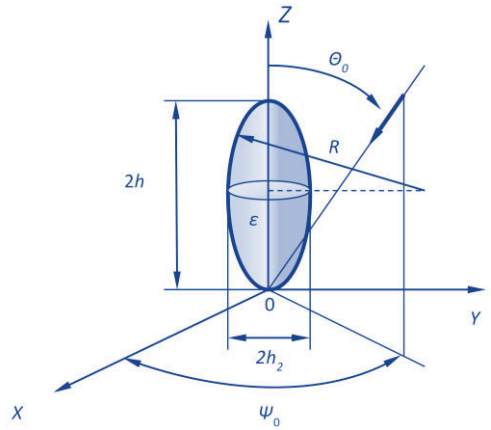


Fig. 1. Electrodynamic model of seeds

Here v – length of the arc of a circle measured from point 0 (the vertex of the ellipsoid), φ – polar angle in the cylindrical coordinate system, and the functions $\bar{R}(v)$ and $\bar{Z}(v)$ have the form:

$$\bar{R}(v) = R\sin\left(v/R\right),\ \bar{Z}(v) = R\left(1-\cos\left(v/R\right)\right). \quad (19)$$

The parameter v varies on the interval (0, L), where $L = ar\cos\left(1-(h_2/R)\right)$, and the angle φ varies over the interval (0, 2π).

Let θ_0 and φ_0 be the angles that determine the direction of propagation of the exciting electromagnetic wave (13) (Fig. 1). The angle θ_0 is measured from the z-axis in the plane passing through this axis and the vector $\vec{n}$, and φ_0 is the angle between this plane and the x-axis. Then the electric field strength of the exciting wave can be represented as:

$$\vec{E}_+^i = \frac{m}{2}\vec{E}_0 e^{ik\left(\sin\theta_0\cos\phi_0 X+\sin\theta_0\sin\phi_0 Y-\cos\theta_0 Z\right)}. \quad (20)$$

As for the polarization of the exciting wave, without loss of generality, we will assume that the vector $\vec{E}_0$ is parallel to the z-axis, i.e. $\vec{E}_0 = E_0\vec{e}_z$, and the direction angles θ_0 and φ_0 are equal to θ_0= π/2, φ_0=π/2. Therefore, from (20) we obtain that the exciting wave has the form:

$$\vec{E}_+^i = \frac{mE_0}{2}e^{iky}\vec{e}_z, \quad (21)$$

$k = \omega / c$ – wave number; $\vec{e}_z$ – a unit vector along the z-axis.

Given the above, it is natural to assume that the strength of the EF excited inside the dielectric ellipsoid will also have a dominant component E_{Z+} (the remaining components E_{X+} and E_{Y+} are negligible compared to E_0).

Let us introduce the notation E_+=E_{Z+} and write the integral equation (17) for E_+. This equation contains two types of integrals: the integral over the volume of the ellipsoid and the integral over its boundary surface.

Let us consider the surface integral. First, we calculate the integrand function $(\vec{n}_q, \vec{E}_+)grad_q G(p,q)$. The equation for the unit vector $\vec{n}_q$ of the normal to the boundary surface of the ellipsoid, taking into account the parametric equations (18), has the following form [14], [15], [19]:

$$\vec{n}_q = -Z\vec{e}_R + R\vec{e}_z, \quad (22)$$

where the dot denotes differentiation, and $\vec{e}_R$ and $\vec{e}_z$ – the unit vectors of the cylindrical coordinate system at the point q.

Substituting (22) into the scalar product $(\vec{n}_q, \vec{E}_+)$, we get:

$$(\vec{n}_q, \vec{E}_+) = RE_+, \quad (23)$$

where, according to (19), R=cos(v/R).

Now consider $grad_q G(p, q)$. Function $G(p, q)$ can be represented as a Fourier series in the azimuthal coordinate φ of a cylindrical coordinate system. Omitting the intermediate transformations and using the results from [14]–[19], we have:

$$G(p,q) = \sum_{n=-\infty}^{+\infty} S_n e^{in(\varphi_1-\varphi_2)}. \quad (24)$$

Here, the points p and q have coordinates $p = (r_1, \phi_1, z_1)$ and $q = (r_2, \phi_2, z_2)$ respectively, and the coefficients S_n are presented in the form of the following integrals:

$$S_n = \frac{1}{2\pi}\int_0^{2\pi}\frac{e^{ik_+ r}}{r}e^{-in\varphi}d\varphi,\ r = \sqrt{(r_1-r_2)^2+(z_1-z_2)^2+4r_1r_2\sin^2\varphi/2}. \quad (25)$$

Using the formula from [16]–[17] for calculating the gradient of a function in a cylindrical coordinate system, we obtain for Z-component $grad_q G(p, q)$ the following expression:

$$\frac{\partial G(p,q)}{\partial Z_2} = \sum_{n=-\infty}^{+\infty}\frac{\partial S_n}{\partial Z_2}e^{in(\varphi_1-\varphi_2)}. \quad (26)$$

In (26) it is taken into account that the gradient is calculated based on the variable q = (r_2, φ_2, z_2).

From (25) we have:

$$\frac{\partial S_n}{\partial z_2} = \frac{z_1-z_2}{2\pi}\int_0^{2\pi}\frac{e^{ik_+ r}}{r^3}e^{-in\varphi}d\varphi. \quad (27)$$

Now we will represent the function $E_+(q)$ – the EF intensity inside the ellipsoid in the form of Fourier series along the azimuthal coordinate φ. We have:

$$E_+(q) = \sum_{n=-\infty}^{+\infty}E_n^+(r_2, z_2)e^{-in\varphi_2}. \quad (28)$$

We substitute (23) and (26) into the surface integral. Using parametric equations (18) for the boundary surface of the ellipsoid, we transform this integral into a double integral over the variables v and φ_2, then we obtain:

$$\int_s(\vec{n}_q\vec{E}_+)\frac{\partial G(p,q)}{\partial Z_2}dS = \\ = \frac{R}{2}\int_0^{2\pi}d\varphi_2\int_0^L\sin\frac{2V}{R}\sum_{n=-\infty}^{+\infty}\vec{E}_+(\bar{R}(v),\bar{Z}(v))e^{in\varphi_2}\sum_{n=-\infty}^{+\infty}\frac{\partial S}{\partial Z_2}e^{in(\varphi_1-\varphi_2)}dV, \quad (29)$$

where L=arccos(1–h_2/R).

If now we use the property of orthogonality of functions $e^{in\varphi_2}$ on the interval (0.2π), then from (29) finally we have:

$$\int_s(\vec{n}_q\vec{E}_+)\frac{\partial G(p,q)}{\partial Z_2}dS = \\ = \pi R\sum_{n=-\infty}^{+\infty}e^{in\varphi_1}\int_0^L\sin\frac{2V}{R}\vec{E}_+(\bar{R}(v),\bar{Z}(v))\frac{\partial S_n}{\partial Z_2}dV. \quad (30)$$

Let us consider the integral over the volume of the ellipsoid from (17). We reduce this volume integral to a two-dimensional integral over the variables r_2, z_2. Using the property of orthogonality of functions $e^{in\varphi_2}$ and the expansion of functions $G(p, q)$ and $E_+(q)$ in Fourier series (24) and (28), after a series of transformations we obtain:

$$\int_V E_+(q)\partial G(p,q)dV = 2\pi\sum_{n=-\infty}^{+\infty} e^{in\varphi_1}\int_0^{2h_1} dz_2\int_0^{a(z_2)} E_n^+(r_2,z_2)S_n r_2 dr_2, \qquad (31)$$

where $a(z_2)=\sqrt{z_2(2R-z_2)}$.

We substitute (30) and (31) into equation (17) and apply the theorem of uniqueness of the expansion of a function in Fourier series [20]–[22], then we will have:

$$E_n^+(r_1,z_1)=E_{+n}^i(r_1)+\frac{(\varepsilon-1)k_+^2}{2}\int_0^{2h_1} dz_2\int_0^{a(z_2)} E_n^+(r_2,z_2)S_n r_2 dr_2 + \qquad (32)$$

$$+\frac{(\varepsilon-1)R}{4}\int_0^L E_n^+\left(\bar{R}(v),\bar{z}(v)\right)\sin\frac{2v}{R}\frac{\partial S_n}{\partial Z_2}dV.$$

Here $E_{+n}^i=\frac{mE_0}{4\pi}\int_0^{2\pi} e^{-in\varphi+ikr_1\sin\varphi}d\varphi=\frac{mE_0}{2}I_n(kr_1)$ (33)

coefficients of the Fourier series expansion of the EF strength of the exciting wave (13), and $I_n(\ldots)$ – Bessel function of the n-th order.

The functions S_n and $\partial S_n/\partial z_2$ from (32) are calculated according to formulas (25) and (27), and the functions $\bar{R}(v)$ and $\bar{z}(v)$ – according to formula (19).

Using equation (32), we can calculate the EF strength inside a dielectric ellipsoid (electrodynamic model of a sunflower seed).

In the general case, the construction of the solution of equation (32) is possible only by numerical methods using a computer. However, for practical applications, an important characteristic of the process of interaction of EMR with biological objects is the magnitude of the intensity EF, averaged over the volume of the biological object.

From (28) it follows that the magnitude of the stress EF, averaged over the volume of the ellipsoid, is determined only by the coefficient with index $n = 0$, in the expansion in Fourier series of the function $E_+(q)$. Therefore, the average stress EF E^+_{avg} equals:

$$E_{avg}^+=\frac{2\pi}{|V|}\int_0^{2h_1} dz_2\int_0^{a(z_2)} E_0^+\left(r_2,a(z_2)\right)r_2 dr_2, \qquad (34)$$

where $|V|$ – volume of the ellipsoid, and the function $E_0^+(r_2,a(z_2))$ satisfies equation (32), in which the index should be $n = 0$.

The volume of an ellipsoid can be calculated using the formula:

$$|V|=2\pi\int_0^{2h_1} dz_2\int_0^{a(z_2)} r_2 dr_2=\pi\int_0^{2h_1} z_2(2R-z_2)dz_2=\frac{4\pi h_1^3}{3}\left(3\frac{R}{h_1}-2\right). \qquad (35)$$

B. Determination of biotropic parameters of modulated electromagnetic wave for sunflower seed treatment

The exciting wave is characterized by three main parameters: frequency ω, modulation frequency Ω and maximum amplitude E_0. As a result of the interaction of this wave with the dielectric ellipsoid (sunflower seed model), an EMF is excited in it. The main characteristic of this field is the EF strength averaged over the volume of the ellipsoid. As follows from (35), to find the EF strength inside the dielectric ellipsoid, it is necessary to calculate the quantities S_0 and $\partial S_0/\partial Z_2$, which are determined according to formulas (25) and (27). Let us first consider the value of S_0. From (25) we have:

$$S_0=\frac{1}{2\pi}\int_0^{2\pi}\frac{e^{ipr}}{r}d\varphi,\ r=\sqrt{(r_1+r_2)^2+(r_1-r_2)^2+4r_1r_2\sin^2\frac{\varphi}{2}}. \qquad (36)$$

Here the parameter $p=k_\pm$, $k_\pm=(\omega\pm\Omega)/c$.

It can be shown that the integral in (36) can be represented in the following form:

$$S_0=\frac{2}{\pi}\int_0^{\pi/2}\frac{e^{ipr}}{\bar{r}}d\varphi=\frac{2\rho}{\pi}\int_0^{\pi/2}\frac{e^{ip\bar{r}}}{\sqrt{1-\gamma^2\sin^2\varphi}}d\varphi$$

$$\bar{r}=\sqrt{\rho^2-\nu^2\sin^2\varphi},\ \rho=\sqrt{(r_1+r_2)^2+(z_1-z_2)^2},$$
$$\nu=2\sqrt{r_1r_2},\ \gamma=\nu/c. \qquad (37)$$

To approximately calculate the integral in (37) with a relative error of less than 5%, we use the fact that $p\bar{r}\le 1$. The last inequality will be satisfied if the frequency of the exciting wave satisfies the condition $\omega/2\pi\le 40$ GHz. Let us assume that this condition is satisfied and expand the function $e^{ip\bar{r}}$ in a Taylor series, limiting ourselves to three terms (this is sufficient for calculation $e^{ip\bar{r}}$ with a relative error of less than 1%). Then from (37) we obtain:

$$S_0=\frac{2}{\pi}\left[\frac{ip\pi}{2}+\frac{1}{\rho}\int_0^{\pi/2}\frac{d\varphi}{\sqrt{1-\gamma^2\sin^2\varphi}}-\right.$$
$$\left.-\frac{p^2\rho}{2}d\varphi\int_0^{\pi/2}\sqrt{1-\gamma^2\sin^2\varphi d\varphi}\right]. \qquad (38)$$

Using the definitions of elliptic integrals of the I-II kind [23]:

$$K(\gamma)=\int_0^{\pi/2}\frac{d\varphi}{\sqrt{1-\gamma^2\sin^2\gamma}},\ E(\gamma)=\int_0^{\pi/2}\sqrt{1-\gamma^2\sin^2\varphi d\varphi}, \qquad (39)$$

then (38) can be represented as:

$$S_0=\frac{2}{\pi}\left[\frac{ip\pi}{2}+\frac{1}{\rho}K(\gamma)-\frac{p^2\rho}{2}E(\gamma)\right]. \qquad (40)$$

To calculate $\partial S_0/\partial Z_2$, it is sufficient to differentiate the right-hand side of (40) with respect to the variable Z_2:

$$\frac{\partial S_0}{\partial Z_2}=\frac{2(Z_1-Z_2)}{\pi\rho}\left[p^2K(\gamma)+\frac{E(\gamma)}{\rho^2(1-\gamma^2)}\right]. \qquad (41)$$

Taking into account (40) and (41) we obtain the following formula for calculating the EF strength at the internal points of the dielectric ellipsoid:

$$\frac{E_0^0+E_0^++E_0^-}{E_0}=mI_0(kr_1)+\frac{im(\varepsilon-1)(k_+^3+k_-^3)}{4}\int_0^{2h_1}a^2(z_2)I_1^2(ka(z_2))dz_2+$$

$$\frac{m(\varepsilon-1)}{2\pi}\int_0^{2h_1}dz_2\int_0^{a(z_2)}I_0(kr_2)\left[\frac{(k_+^2+k_-^2)}{\rho}k(\gamma)-\frac{(k_+^4+k_-^4)}{2}\rho E(\gamma)\right]r_2dr_2-$$

$$-\frac{(\varepsilon-1)mr}{4\pi}\int_0^L I_0(k\bar{r}(v))\sin\frac{2v}{r}\frac{z_1-\bar{z}(v)}{\rho}\left[k(\gamma)(k_+^4+k_-^2)+\frac{2E(\gamma)}{\rho^2(1-\gamma^2)}\right]dv+$$

$$+\frac{iI_0(kz_1)}{2}+\frac{(\varepsilon-1)k^3}{4}\int_0^{2h_1}a^2(z_2)I_1^2(ka(z_2))dz_2-$$

$$-\frac{i(\varepsilon-1)k^2}{2\pi}\int_0^{2h_1}dz_2\int_0^{a(z_2)}I_0(kr_2)\left[\frac{k(\gamma)}{\rho}-\frac{k^2\rho}{2}E(\gamma)\right]r_2dr_2+$$

$$+\frac{i(\varepsilon-1)r}{4\pi}\int_0^{a(Z_2)}I_0(\bar{kr}(v))\sin\frac{2v}{r}\frac{z_1-\bar{z}(v)}{\rho}\left[k(\gamma)k^2+\frac{2E(\gamma)}{\rho^2(1-\gamma^2)}\right]dv. \quad (42)$$

Thus, having determined from (42) the strength of the ellipsoid EF and substituting it into (34), we obtain the averaged over the volume of the dielectric strength of the EF excited by the modulated EMF.

V. Results and Discussion

A. Solutions to the problem of interaction of modulated EMF with a dielectric ellipsoid is reduced to solving two similar problems for wave numbers $k_+=\omega_+/c=(\omega+\Omega)/c$, $k_-=(\omega-\Omega)/c$ and $k=\omega/c$. Therefore, the averaged over the volume of the ellipsoid EF intensity E_{cf} has the form:

$$E_{avg}=E_{avg}^0+E_{avg}^+ +E_{avg}^-=\frac{2\pi}{|V|}\int_0^{2h_1}dz_2\int_0^{a(z_2)}(E_{avg}^0+E_{avg}^+ +E_{avg}^-)r_2dr_2. \quad (43)$$

Here $E_0^{\pm}$ – solutions of the integral equation (32), for wave numbers $k_{\pm}$ and n=0, and E_0^0 – solution of the same integral equation for the wave number k and n=0.

Thus, the original problem is reduced to constructing solutions of the integral equation (35) for two values of the wave number k_+ and k_-. As numerical estimates have shown, these solutions can be obtained using the Born approximation [23]–[25].

Based on the Born approximation, we have:

$$E_0^{\pm}=-\frac{mE_0}{2}I_0(kr_1)+\frac{m(\varepsilon-1)k_{\pm}^2E_0}{4}\int_0^{2h_1}dz_2\int_0^{a(z_2)}I_0(kr_2)S_0r_2dr_2-$$

$$-\frac{m(\varepsilon-1)rE_0}{8}\int_0^L I_0(k\bar{r}(v))\sin\frac{2v}{r}\frac{\partial S_0}{\partial z_2}dv,$$

$$E_0^0=-\frac{iE_0}{2}I_0(kr_1)-\frac{i(\varepsilon-1)k^2E_0}{4}\int_0^{2h_1}dz_2\int_0^{a(z_2)}I_0(kr_2)S_0r_2dr_2-$$

$$-\frac{i(\varepsilon-1)rE_0}{8}\int_0^L I_0(k\bar{r}(v))\sin\frac{2v}{r}\frac{\partial S_0}{\partial z_2}dv. \quad (44)$$

Note that when the modulation depth coefficient m becomes zero, the original problem becomes the problem of diffraction of a harmonic plane wave with a circular frequency ω.

Thus, using formulas (34) and (35), we can calculate the strength of the EF arising as a result of the interaction of a modulated EM wave with a dielectric ellipsoid, which is an electrodynamic model of a sunflower seed.

B. Using (42) and (34), calculations were carried out E_{avg}/E_0 as functions of frequency ω of the exciting wave and modulation frequency Ω. The following frequency range of the exciting wave was chosen for the calculations: 25…40 GHz. The choice of this range is due to two circumstances. Firstly, the relative permittivity $\varepsilon=\varepsilon'-i\varepsilon''$ of sunflower seeds in this range practically does not have frequency dispersion, and the real part ε' lies in the interval $3\le\varepsilon'\le4$, and the imaginary part is $2\cdot10^{-3}\le\varepsilon''/\varepsilon'\le4\cdot10^{-3}$.

Secondly, in this frequency range the wavelength is commensurate with the characteristic geometric dimensions of the seeds and, therefore, the strength of the EF excited inside the seeds resonantly depends on the frequency.

Figures 2, 3 show the results of calculating the average electric field strength normalized to the maximum amplitude of the exciting modeled wave. Analysis of the results presented in Fig. 2 allows us to draw a conclusion about the resonant behavior of E_{avg}/E_0 depending on the frequency ω of the exciting wave at fixed values of the modulation frequency Ω and the modulation depth coefficient m. In this case, the resonant frequency is equal to $\omega/2\pi$ = 36.8 GHz for the modulation frequency Ω changing in the range 50 Hz ≤ Ω ≤ 150 Hz and the modulation depth coefficient m. Numerical calculations of the lines of equal values of E_{avg}/E_0 as a function of Ω and m were performed at a fixed resonant frequency value $\omega/2\pi$ = 36.8 GHz. Fig. 3 shows the results of these calculations. As it can be seen, there are optimal values of the modulation frequency Ω=100 Hz and the modulation depth coefficient m = 0.55, at which the maximum value of $E_{avg}/E_0 \approx 0.03$ is achieved.

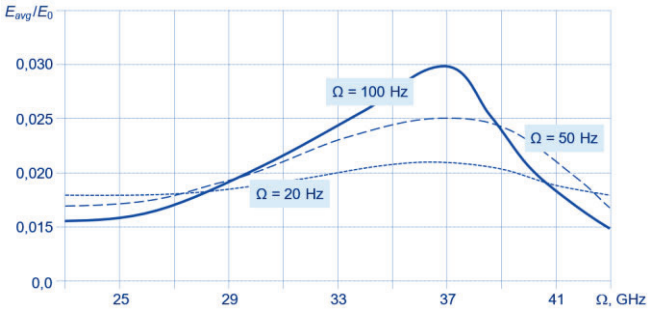


Fig. 2. Dependence of the normalized electric field intensity averaged over the volume of the seed on the frequency of the exciting EM wave with a modulation frequency Ω

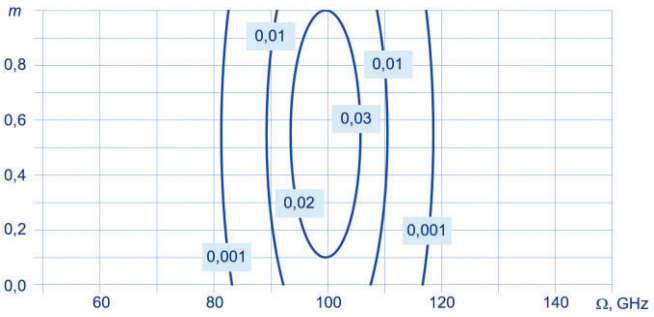


Fig. 3. Lines of equal values of E_{avg}/E_0 as functions of the modulation frequency Ω and the modulation depth coefficient m at a fixed value of the resonant frequency $\omega = 2\pi$ 36.8 GHz

For the induced potential on the membranes of sunflower seed cells is determined by the expression:

$$\varphi_{ind}^2=\varphi_0^2+Pt\frac{C_0V_0g^2C_S}{\left(4\pi\varepsilon_{mem}\varepsilon_0d\right)^2}\cdot e^{-\frac{F\varphi_0}{RT}\cdot\frac{g}{e}}+P\frac{gC_S}{4\pi\varepsilon_{mem}\varepsilon_0\omega}E_{avg}\sin\omega t. \quad (45)$$

where φ_0 – membrane potential at the initial moment of time; P – membrane permeability; t – exposure time; C_0, C_S – ion concentrations inside and outside the cell; V_0 – cell volume at the initial moment of time; g – ion charge; F – Faraday number; R – gas constant; T – absolute temperature; e – electron charge; E – electric field strength; N_A – Avogadro constant; ω – the circular frequency of the EMF irradiation of sunflower seeds. For the calculations, data taken from literary sources were used [14], [15], [26], [27]: E_{mem}=0.6 10^{-2} N/m; d=10^{-8} m; ε_{mem}=2.1; ε_0=8.854 10^{-12} F/m; N_A=6.022·10^{23} mol^{-1};

φ_{cr}=110 mV; φ_0=60 mV; P=5·10^{-3} s/m^3 ; C_0=2.00 mK mol/m^3; C_S=6.4 mK mol/m^3; V_0=1·10^{-18} m^3; g=3.5·10^{-18} cl; F=96484.56 K/mol; R=8.314 J/K mol; T=300 ^{0}K; f=36.8 GHz.

When carrying out numerical calculations, it was taken into account that the induced potential is equal to 100 mV < φ_{cr} =110 mV. As a result of the calculations, it was established that the exposure is t = 250…350 s, and the intensity E_{avg}=11.7...13.78 V/m.

VI. Conclusion

1. When determining the main parameters of the EMF for irradiating sunflower seeds, the dielectric constant and geometric parameters of the seeds should be taken into account.

2. Based on theoretical studies, it has been established that the optimal biotropic parameters of EMF for irradiating sunflower seeds should be determined in the frequency range of 35.8...37.8 GHz.

3. During the theoretical analysis of the developed model, it was established that for pre-sowing treatment of sunflower seeds, it is necessary to use an information EMF with the following parameters: modulation frequency 100 Hz; modulation index 0.5; intensity 11.7...13.78 V/m; exposure 250...350 s.

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Running wave interferometer with a half-wave plate

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***Abstract* — the paper considers the results of mathematical modeling of the processes occurring in a running wave interferometer with a half-wave plate, provided that monochromatic laser radiation with linear polarization enters its input. The conditions for the occurrence of intensity modulation at the output of the interferometer in the absence of optical signal amplification in it are investigated. The dependences of the modulation coefficient of the output signal on the parameters of the interferometer and the orientation of the plane of polarization of the input radiation are given.**

Keywords — running wave interferometer, light interference, polarization of light, half-wave plate, rotation of the plane of polarization

I. INTRODUCTION

A running wave interferometer (RWI) is a two-beam interferometer in which interference occurs between two collinear coherent light waves [1 – 6], which arise due to a translucent input mirror and two pairs of parallel mirrors, which form the path of light propagation in the active medium located inside the interferometer, which will be called closed in the following. The light wave propagating in this medium will not only lag in phase with respect to the first wave formed by reflection from the semitransparent input mirror, but also (if necessary) be amplified in amplitude by creating an inversion of the level populations by optically pumping the closed medium.

When modeling the interference phenomena observed in RWI using the time model [1 – 5] and the analytical model [6], the issue of polarization was not considered, and the oscillations of the electric field vector in both collinear waves were carried out in the polarization plane of the input laser radiation. Accordingly, the polarization state of light remained constant over time and did not affect the intensity of the RWI output signal.

Adding a half-wave plate [7 – 11] to a closed medium results in the appearance of two orthogonal electric field components in the interferometer, which are energetically linked to each other due to the phenomena of interference and birefringence. Therefore, the output intensity becomes «sensitive» to the polarization state of the optical radiation propagating in the interferometer.

II. WORKING PRINCIPLE OF THE INTERFEROMETER

Monochromatic laser radiation with wavelength λ and intensity I_l, which enters the RWI input, is divided by a translucent mirror with an amplitude reflection coefficient *r* into light waves 1 and 2 (fig. 1).

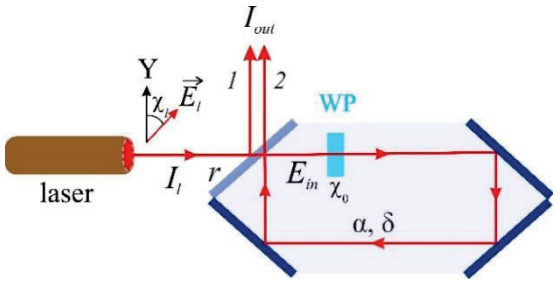


Fig. 1. Functional scheme of RWI.

Wave *1* (fig. 1) goes immediately to the output of the interferometer, changing its phase by π, since it is reflected from a translucent mirror with a higher refractive index than the medium in which the interferometer is located. Wave *2* enters a closed medium formed by one translucent and three opaque mirrors. If we do not take into account the half-wave plate (WP), which is located in the middle of the interferometer (fig. 1), then during the propagation of wave *2* in a closed medium, its amplitude changes by a factor of α (where α is the propagation constant), and the phase lag with respect to wave *1* reaches a value of δ, which is determined by the propagation time of light in a closed medium. Then, wave *2* is divided into two parts by a semi-transparent mirror: the first one enters the output of the interferometer (wave *2* in fig. 1); the second one remains in the closed medium and is superimposed on the input laser radiation (which passed through the semi-transparent mirror), forming an optical signal in the closed medium at subsequent moments of time.

The output radiation of the RWI is also a superposition of two coherent collinear running waves *1* and *2* (in fig. 1, these waves are spatially shifted for clarity), the amplitude and phase of which depend on the RWI parameters: *r*, α, δ [1 – 6]. In this case, the oscillations of the electric field intensity vector in all light waves occur in the polarization plane of the input laser radiation.

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Let us take into account the influence on the optical signal of the interferometer of the half-wave plate WP (fig. 1), the optical axis of which is located at an angle χ_0 to the vertical (fig. 2). We will assume that the oscillations of the electric field intensity vector $\vec{E}_{in}$ in the light wave that falls normally on the plate occur in the plane (fig. 2), which forms an angle χ_{in} with the vertical (Y axis).

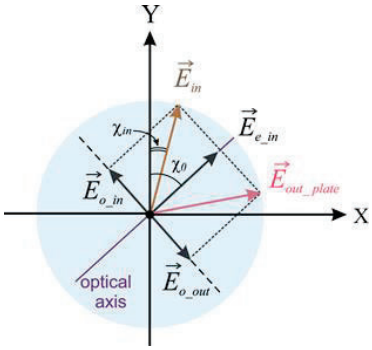


Fig. 2 Rotation of the polarization plane of incident light by a half-wave plate.

Since the thickness of the plate is $d = \frac{\lambda}{2}(n_o - n_e)$, the phase difference between the ordinary $\vec{E}_{e_out}$ and extraordinary $\vec{E}_{o_out}$ rays that arise due to the phenomenon of birefringence at the plate output will be π $\left(\vec{E}_{e_out} = \vec{E}_{e_in}, \vec{E}_{o_out} = -\vec{E}_{o_in}\right)$. The plate is cut from a uniaxial crystal parallel to the optical axis, so the ordinary and extraordinary rays are not spatially separated, although the absolute refractive indices for the ordinary n_o and extraordinary n_e rays are different. The direction of oscillation of the electric field vector in these rays is orthogonal (fig. 2). Due to the additional phase π between the coherent wave components (fig. 2), the rotation of the plane of polarization of light (vector $\vec{E}_{out_plate}$) at the plate output occurs by an angle $2(\chi_0 - \chi_{in})$. If we do not take into account the absorption of light in the half-wave plate, then the complex amplitudes $E_{in}^x(N)$, $E_{in}^y(N)$ of the horizontal and vertical components of the electric field intensity (coherent components of the vector $\vec{E}_{in}$ along the X and Y axes) of the light wave in the closed medium of the RWI, due to the phenomenon of birefringence and interference with the input laser radiation, after N complete passes of the closed medium will have the form:

$$E_{in}^x(N) = \frac{E_{in}^x(N-1)\cdot r_x + E_l t_x \sin\chi_l}{\sin(\chi_{in}(N))}\cdot\sin(2\chi_0 - \chi_{in}(N))\cdot\alpha_x e^{i\delta_x};$$

$$E_{in}^y(N) = \frac{E_{in}^y(N-1)\cdot r_y + E_l t_y \cos\chi_l}{\cos(\chi_{in}(N))}\cdot\cos(2\chi_0 - \chi_{in}(N))\cdot\alpha_y e^{i\delta_y},$$

where $E_{in}^x(N-1)$, $E_{in}^y(N-1)$ – complex amplitudes of the electric component of the wave propagating in the closed medium after its $N-1$ complete passes $\left(E_{in}^x(0) = E_l t_x \sin\chi_l,\ E_{in}^y(0) = E_l t_y \cos\chi_l\right)$; E_l – amplitude of the input laser radiation; χ_l – angle between the Y axis and the polarization plane of the input laser radiation; $r_x, r_y\ (t_x,\ t_y)$ – reflection coefficient (transmission) by the amplitude of the input mirror for the horizontal and vertical components, respectively; $\alpha_x, \alpha_y\ (\delta_x, \delta_y)$ – propagation constant (phase delay) for the light wave with horizontal and vertical polarization, respectively; i – imaginary unit; $\chi_{in}(N)$ – the angle between the Y axis and the plane of polarization of the optical signal (fig. 2) at the input of the half-wave plate after N complete passes of the closed medium $(\chi_{in}(0) = \chi_l)$:

$$\chi_{in}(N) = \operatorname{atan}\left(\frac{\left|E_{in}^x(N-1)\cdot r_x + E_l t_x \sin\chi_l\right|}{\left|E_{in}^y(N-1)\cdot r_y + E_l t_y \cos\chi_l\right|}\right).$$

Thus, the complex amplitudes of the electric components of the light wave at the output of the RWI:

$$E_{out}^x(N) = E_{in}^x(N)\cdot(-t_x) + E_l r_x \sin\chi_l;$$
$$E_{out}^y(N) = E_{in}^y(N)\cdot(-t_y) + E_l r_y \cos\chi_l,$$

will depend (under the condition $E_l = \text{const}$) on a large number of parameters: $r_x,\ r_y, \alpha_x, \alpha_y, \delta_x, \delta_y, \chi_l, \chi_0$. Therefore, to simplify the mathematical model of the RWI when studying the influence of the interferometer on monochromatic input radiation, we will assume that:

1. the closed medium outside the half-wave plate is homogeneous, isotropic and linear $(\delta_x = \delta_y = \delta)$;

2. the reflection coefficient of the input semi-transparent mirror does not depend on the polarization of the incident radiation $\left(r_x = r_y = \sqrt{R}, t_x = t_y = \sqrt{1-R}\right)$;

3. there is no dichroism phenomenon in the half-wave plate, therefore the signal propagation constant in the medium does not depend on the state of polarization of the radiation $(\alpha_x = \alpha_y = \alpha)$.

Since light interference is possible only in the case when the oscillations of the electric field intensity vector in both waves occur in the same direction [12 – 14], for the emergence of two coherent waves with orthogonal polarization states in the RWI, the following condition must be fulfilled: $\chi_l \neq \chi_0$.

III. Simulation Results

In the absence of optical amplification in a closed environment of the RWI $(\alpha \le 1)$, the intensity of the output signal I_{ou} after the transient process, depending on the parameters of the interferometer, acquires stationary values or is modulated by a periodic function (fig. 3). In this case, the modulation coefficient K_i and the modulation period T_i of the intensity I_{out}, as well as in the absence of a half-wave plate [1, 6], strongly depend on the value δ.

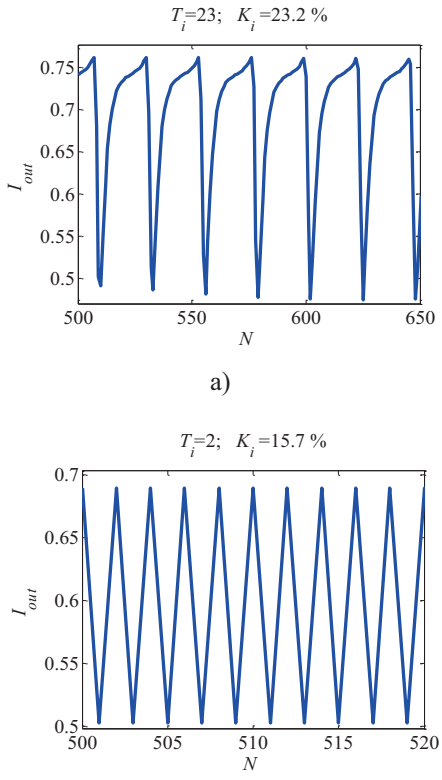


Fig. 3 Dependence $I_{out}(N)$ at different parameters of the RWI $(\chi_0 = 60^\circ, \chi_l = 15^\circ, \alpha = 0.591, R = 0.5)$: a) $\delta = 44.5^\circ$; b) $\delta = 182^\circ$.

The modulation of the output intensity is observed (fig. 4) only if the constant α is not less than some limiting value α_{min}, which depends on the parameters of the interferometer, the orientation of the plane of polarization of the input signal and the optical axis of the plate (fig. 5). It should be noted that, unlike other methods of laser radiation modulation [15–19], in RWI it occurs due to the interference of polarized waves.

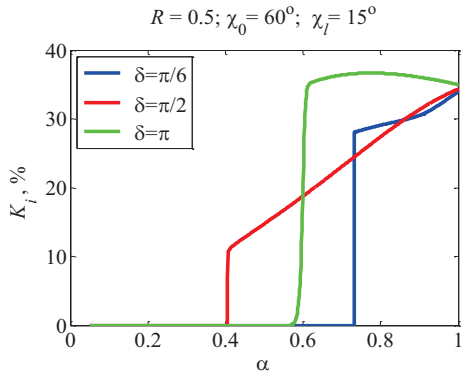


Fig.4 Dependence of the intensity modulation coefficient K_i on the propagation constant for different δ.

The function $K_i(\alpha)$ depending on the optical length of the closed medium of the interferometer (at $\alpha > \alpha_{min}$) can be (fig. 4) monotonic (for some values of δ an almost linear dependence is observed) or non-monotonic, which has a different number of local minima. Unlike $K_i(\alpha)$, the function $\alpha_{min}(\delta)$, when changing the parameters of the interferometer, always has one local maximum at the point $\delta = \pi$ (fig. 5), which arises due to the fact that for such a length of the closed medium of the RWI, the intensity modulation period is minimal and the energy is not retained inside the interferometer.

As the angle χ_0 $(|\chi_0 - \chi_l| = \text{const})$ decreases, the number of values of the parameter δ increases, for which modulation of the output radiation is possible $(K_i \neq 0)$ only in the absence of losses $(\alpha = 1)$ in the closed medium of the RWI (fig. 5).

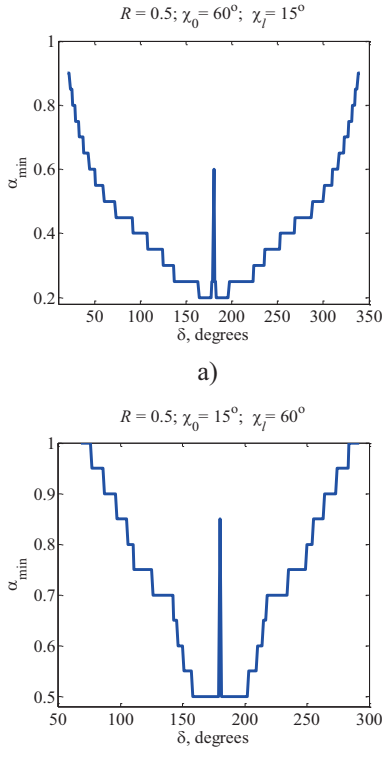
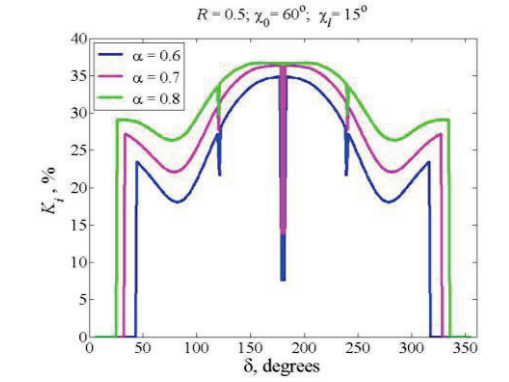


Fig. 5 Dependence of the threshold value of the propagation constant α_{min} on δ: a) $\chi_0 = 60^\circ, \chi_l = 15^\circ$; b) $\chi_0 = 15^\circ, \chi_l = 60^\circ$.

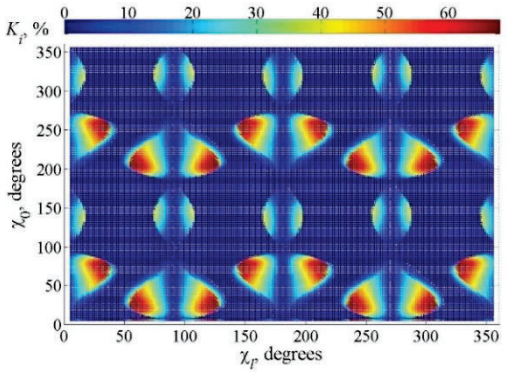
The largest values of the modulation coefficient are observed (fig. 6a) when an almost integer number of half-waves is placed on the optical length of the closed medium $(\delta \approx 180^\circ)$. When $\delta = 180^\circ$ there is a sharp decrease in the value of K_i, since the oscillations (after N passes) of one of the orthogonal components (for example, horizontal) of the electric field of the light wave in the closed medium occur in phase with the horizontal component of the electric field of the input signal, and the corresponding vertical components of the electric field oscillate in antiphase. Therefore, due to interference (and taking into account the reflection from the semi-transparent mirror), one component of the field E^y_{out} is amplified (interference maximum), and the other E^x_{out} is attenuated (interference minimum), which leads to a decrease in the difference between the maximum and minimum intensity values at the output of the RWI, which in turn leads to a decrease in the modulation coefficient.

Intensity modulation at the output of the RWI, in the absence of amplification in a closed environment, arises as a

result of the exchange of energy between the components of the electric field with orthogonal polarization, due to the phenomenon of birefringence observed in the half-wave plate. The value of this energy depends both on the magnitude of the rotation of the plane of polarization and on the interference of waves with the same direction of oscillation of the electric field vector. Therefore, the modulation of radiation at the output of the interferometer is observed only in certain intervals of angles χ_0 and χ_I (fig. 6b), and the RWI with a half-wave plate itself can be considered as two connected running-wave interferometers with different optical lengths of the closed medium.



a)



b)

Fig. 6 Dependence of the intensity modulation coefficient at the output of the RWI: a) on the phase delay of the signal in the middle of the RWI at different α; b) on the orientation of the polarization plane of the input radiation χ_I and the optical axis χ_0 of the half-wave plate $\left(\delta = 150^{\circ}, \alpha = 0.7, R = 0.5\right)$.

IV. Conclusions

1. The possibility of the appearance of modulated radiation at the output of the RWI in the absence of optical pumping of the closed medium is proved.

2. The existence of a threshold value α_{min} of the propagation constant is established, starting from which the intensity at the output is a periodic function of time for the RWI with any optical length of the closed medium.

3. The dependence of the modulation coefficient of the output intensity on the phase delay δ of the optical wave in the closed medium is investigated. It is shown that the smallest value of K_i at different $\alpha > \alpha_{min}$ is observed when $\delta = \pi$.

4. It is demonstrated that the energy exchange between the orthogonal components of the electric field of a light wave, in a certain interval of angles χ_0 and χ_I, causes a violation of the stationarity of the intensity at the output of the RWI.

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Audio Forgery Detection via Combined Recurrent-Convolutional Neural Network Architecture

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***Abstract*—The advancement of deep fake technology has elevated audio fabrication detection to a critical research priority. Existing frameworks exhibit considerable constraints that hinder their integration into operational autonomous platforms. This investigation introduces an innovative methodology for acoustic data examination to identify contextual information manipulation. The proposed framework combines recurrent and convolutional neural networks within a unified computational structure. Assessment conducted on a proprietary corpus focused on operational intelligence regarding Russia's military campaign in Ukraine demonstrates the framework's exceptional performance compared to conventional approaches, suggesting significant potential for deployment within autonomous vehicle systems.**

Keywords—acoustic analysis, cybersecurity, deep learning architectures, pattern recognition, speech processing

I. Introduction

Autonomous vehicles (AVs) operating in maritime, aerial, and terrestrial environments have experienced remarkable technological progression in both core software platforms and hardware systems, alongside sophisticated artificial intelligence integration [1]. However, validating transmission reliability remains an unresolved critical challenge. Contemporary technological capabilities have facilitated increasingly sophisticated methods of visual and acoustic content manipulation [2], presenting substantial operational vulnerabilities for AV deployments. While video manipulation requires increasingly sophisticated interventions, potentially addressable through advanced pattern recognition and electronic countermeasure implementations, acoustic information demonstrates heightened susceptibility to manipulation through both contextual alteration and voice simulation techniques. Current technological frameworks [3] have reached levels of complexity where distinguishing authentic recordings from synthetically generated alternatives frequently exceeds human discriminatory capabilities, particularly within compromised acoustic environments.

In the context of ongoing military operations in Ukraine and broader international security considerations, accurate acoustic data interpretation represents a mission-critical capability. While expert analysis teams provide a conventional verification approach, operational scale and potential human judgment errors during high-stress scenarios significantly compromise reliability in forgery identification [4].

At the global level, anti-falsification initiatives have demonstrated promising developments. Various computational tools including browser extensions and social media analytical modules have emerged to identify textual fabrications and detect manipulations across visual and acoustic information channels. Scientific literature documents diverse methodological approaches across international research communities. Significant contributions include Massachusetts Institute of Technology initiatives and Ukrainian research collaborations examining contextual acoustic falsification through machine learning implementations [5, 6], alongside Sino-American investigations employing probabilistic frameworks [7].

Nevertheless, these advancements, the computational intensity of prevailing models typically necessitates extensive training datasets for classification precision, presenting significant operational implementation barriers [8, 9]. Within autonomous vehicle architectural frameworks, such requirements impose substantial constraints, compromising integration feasibility within analytical subsystems.

To address these implementation challenges, the primary objective of this research was postulated as developing an efficient computational framework for detecting manipulated audio content in autonomous vehicle communication systems operating in contested environments. Specific research objectives include: designing a hybrid neural network architecture that combines the sequential processing capabilities of recurrent networks with the feature extraction strengths of convolutional networks, establishing a multi-criteria evaluation methodology for acoustic falsification detection systems, and validating the proposed approach using real-world operational intelligence data from ongoing military operations in Ukraine. Previous investigations have demonstrated this approach's effectiveness in falsification detection within social intelligence systems [6]. Comparative evaluation is conducted against established methodologies, including traditional recurrent neural networks (RNN), convolutional neural networks (CNN), and naive Bayesian classification systems (NBC).

II. Domain Analysis

Regardless of specific implementation techniques, audio falsification invariably seeks to manipulate contextual elements to achieve strategic objectives. Through comprehensive examination, several linguistic markers can be identified for detecting manipulated acoustic content:

- Elevated frequency of rhetorical questioning patterns, potentially indicating contextual manipulation, particularly within socially significant communications.
- Juxtaposition of absent negative constructions with pessimistic terminological choices (substituting moderate descriptors with extreme alternatives), potentially signaling manipulative intent. Notably, profanity-laden discourse complicates sentiment evaluation, necessitating deeper contextual analysis.

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- Strategic deployment of inappropriate appeals and motivational language, frequently characteristic of falsification in contextual manipulation scenarios, while potentially representing normative speech patterns in mimicry recordings.

These indicators represent a non-exhaustive framework. Notably, fabricated communications frequently employ abbreviated syntactical structures or contain various linguistic irregularities [10]. Such characteristics may stem from speech recognition inaccuracies or reflect distinctive speech patterns, including code-switching behaviors or dialectal variations. Consequently, these elements potentially contribute to elevated false positive identification rates in falsification detection algorithms.

III. Overview of Aproach

Following examination of characteristics relevant to falsification detection, the subsequent investigative phase involves evaluating the proposed detection methodology.

A. Features Base

To create target model these supplementary parameters are incorporated:

- Contextual Indicator: Determined as a result of context analysis for transformed audio into text form.
- Misinformation Probability Index: Determined through analysis of prominent thematic communications, utilizing a parallelized Apriori algorithm implementation.
- Communication Classification Framework: Acoustic data segmentation into four distinct categories: interpersonal communications, standard informational broadcasts, emergency notifications, and high-priority intelligence.
- Recognition Accuracy Assessment: Evaluated through comparative analysis between source content and system-processed output, determining lexical recognition precision.

The Apriori algorithm selection stems from its computational efficiency, implementation accessibility, and parallel processing compatibility. Originally developed for market basket analysis, this algorithmic framework has been adapted for linguistic pattern identification through a modified implementation. The algorithm operates on the fundamental principle of monotonicity in frequent itemsets, asserting that any subset of a frequent itemset must also be frequent. This property enables efficient pruning of the candidate space during processing. The implementation incorporates additional optimizations including transaction reduction techniques, where transaction databases are progressively reduced by removing processed items, and parallel processing through MapReduce, distributing computations across multiple nodes. Performance benchmarking demonstrated up to 5.7 times acceleration compared to standard sequential implementations when processing large linguistic corpora.

B. Computational Architecture

The implemented RCNN architecture integrates multiple convolutional networks with a bidirectional recurrent framework incorporating extended memory capabilities (BiLSTM). The architectural configuration appears in Figure 1. Initially, the transformed linguistic data undergoes segmentation into computational nodes according to predefined algorithms before convolutional network processing.

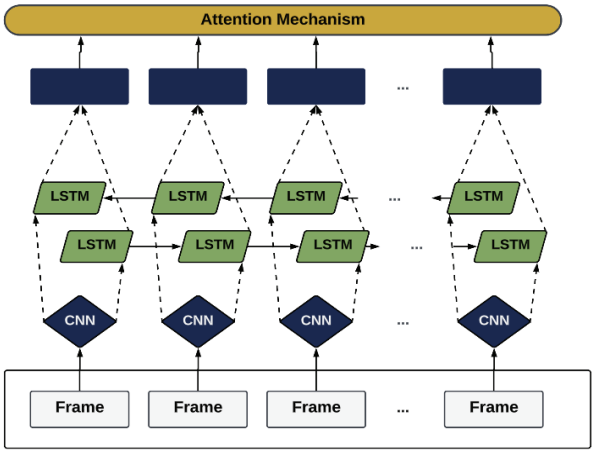


Fig. 1. Scheme of the RCNN architecture.

The convolutional component consists of three sequential layers with declining filter dimensions (128-64-32), each followed by batch normalization, ReLU activation, and max-pooling operations to progressively extract increasingly abstract linguistic features. This preprocessing facilitates dimensional reduction while preserving contextual integrity, reducing the 300-dimensional word embeddings to a 64-dimensional feature space that encapsulates the essential semantic information.

The network architecture converges in a fusion layer that combines convolutional and recurrent outputs through attention mechanisms, allowing the model to dynamically weight feature importance based on contextual significance. This is followed by two fully-connected layers (128 and 64 neurons respectively) with ReLU activation and a final softmax classification layer. The attention mechanism particularly enhances performance by focusing computational resources on the most informative segments of input data, significantly improving classification accuracy for longer audio sequences. Subsequently, the transformed data undergoes bidirectional recurrent analysis, incorporating contextual patterns from multiple sequential directions. This multidirectional approach proves essential for processing languages with variable syntactical arrangements, such as Ukrainian, where contextual indicators may appear non-sequentially throughout communication. The model was implemented using TensorFlow 2.9 with the Keras API, leveraging GPU acceleration through CUDA and cuDNN libraries to optimize computational performance.

Parameter optimization employed cross-validation methodologies guided by established research frameworks [1, 2], yielding these optimal configurations:

- Convolutional Kernel Size**:** Evaluating kernel dimensional ranges between 2×2 and 5×5 established optimal size at 4×4 for feature extraction.
- Progression Parameter: Considering established guidance against values exceeding 3 for linguistic data, optimal progression was established at 1.
- Learning Rate: Adaptive learning rate using Adam optimizer with initial value of 0.001.

- Batch Size: Optimal performance achieved with mini-batches of 64 samples.

Addressing these limitations required implementing parallel processing and distributed computing frameworks. The computational infrastructure consisted of a containerized Kubernetes cluster with high-availability configuration, comprising 24 processing nodes distributed across three physical locations for redundancy and load balancing. The primary computation nodes (16) featured Intel Core i7-12700K processors (12 cores, 20 threads, 4.9 GHz turbo) with 64GB DDR5-4800 RAM and NVIDIA RTX 3080 accelerators (10GB GDDR6X). Supplementary nodes (8) utilized AMD Ryzen 9 5950X processors with 32GB RAM and AMD Radeon RX 6800 XT accelerators. The environment operated on a customized Ubuntu 22.04 LTS distribution with specialized kernel optimizations for high-throughput computational workloads. Network infrastructure utilized 10Gbps fiber interconnects with redundant pathways to minimize latency during parallel processing operations.

IV. Experimental Design

Within this investigation, "experimental design" encompasses both methodological approach and multi-criteria evaluation framework for identifying optimal acoustic classification models.

A. Procedural Framework

Evaluation data was derived from custom-developed acoustic datasets, transformed to linguistic format through Google Speech-to-Text API.

The primary corpus comprises 5,000 communications regarding Russia's Ukrainian invasion collected from X platform during 2022, with total duration of 62.5 hours (2.3 GB, 16 kHz sampling rate, average 45-second segments). Languages include Ukrainian (60%), Russian (25%), and English (15%), all standardized through Ukrainian translation. Recording conditions varied with signal-to-noise ratios between 10-25 dB, representing authentic field communication environments. The secondary corpus contains controlled recordings from specialized focus groups (25 military personnel, 15 civilian experts) totaling 44.7 hours (1.8 GB, 22.05 kHz sampling rate) under studio conditions (SNR > 30 dB), focusing on equipment positioning and deployment communications with irregular syntactical structures and regional expressions. Each dataset underwent division into training and test segments in 80:20 proportions, respectively.

For implementation purposes, Python3 libraries including TensorFlow and NLTK were utilized, alongside appropriate interconnection protocols within the Kubernetes environment.

B. Evaluation Framework

To establish performance metrics, specialist consultation involved 50 anonymized data analysis experts representing Ukraine, Poland, Austria, Portugal, and the United States. This specialist panel additionally contributed to determining significance coefficients for evaluation metrics. All factors underwent standardization according to optimization principles:

- Precision Assessment: Evaluated through weighted combination of Recall and Precision measurements, standardized within [0,1] range, with significance coefficient 10.
- Computational Efficiency: Processing optimization, with significance coefficient 6.
- Data Optimization: Minimum required corpus size achieving 80% accuracy, maximum threshold 10,000 communications, with significance coefficient 4.

With significance coefficients established, this linear additive convolution (LAC) formula was implemented:

$$LAC = 0.5 \cdot A + 0.3 \cdot TS + 0.2 \cdot VS \quad (1)$$

where A – accuracy; TS – computational efficiency optimization; VS – data volume optimization.

Essential experimental considerations include identifying potential error sources. Based on methodological design, several potential influence factors emerge: computational efficiency assessments potentially influenced by human error and instrumental precision limitations; accuracy assessments potentially affected by data quality considerations.

To mitigate potential measurement inaccuracies, a comprehensive approach involving ten discrete measurements per parameter was implemented.

V. Results of the Experiment

Comprehensive experimental evaluation revealed significant performance differentials across the investigated architectural frameworks. Results appear in Table 1.

TABLE I. Experiment Results

Models	Metrics		
	Saving processing time	***Accuracy***	***Saving of min amount of data***
RCNN	0.82	0.97	0.980
CNN	0.10	0.82	0.237
RNN	0.00	0.87	0.485
NBC	1.00	0.80	0.954

[a.] Normalization conducted against maximum values rather than standardized metrics

Based on experimental data, linear additive convolution calculations with weighted coefficients yielded these comprehensive performance values: for RCNN, the value is 0.927; for CNN, it is 0.4874; for RNN, it is 0.532; and for NBC, it is 0.8908.

These aggregated metrics, incorporating accuracy, computational efficiency, and data requirements, demonstrate RCNN's superior overall performance despite moderate computational demands. Notably, while NBC achieved a respectable composite score (0.8908), this primarily resulted from exceptional computational efficiency rather than discriminative accuracy. The performance distribution demonstrates a significant separation between RCNN and alternative implementations, with a composite score difference exceeding 0.036 compared to the next-best performer (NBC). This substantial margin suggests that potential variations in coefficient weighting within reasonable parameters would maintain RCNN's positional advantage within the evaluation framework.

VI. Limitations and Future Work

While the proposed RCNN architecture demonstrates substantial advantages for audio falsification detection,

several limitations warrant acknowledgment and suggest directions for future research.

The implementation of the proposed approach revealed several notable limitations. First, despite parallelization efforts, the RCNN architecture requires significant computational resources, particularly due to the bidirectional LSTM components, potentially restricting deployment on resource-constrained platforms. Additionally, although the model demonstrates superior data efficiency compared to alternatives, its effectiveness remains dependent on representative training samples – a persistent challenge for emerging falsification techniques.

These limitations point toward several promising research directions. To address computational constraints, future work should explore architectural optimizations including knowledge distillation techniques, selective attention mechanisms to focus resources on potentially manipulated content segments, and more efficient alternatives to LSTM components. The single-modality limitation could be overcome through multimodal fusion frameworks that integrate linguistic analysis with acoustic feature examination – potentially incorporating spectral anomaly detection and voice consistency verification across communication segments. This integration would significantly enhance robustness against sophisticated manipulation techniques. To improve adaptability to emerging threats, research should develop continuous learning strategies through few-shot learning for rapid adaptation to novel manipulation patterns and adversarial training frameworks to enhance robustness against evolving falsification techniques.

VII. Conclusions

This investigation evaluated RCNN architecture effectiveness for falsified acoustic data detection from autonomous systems. Key outcomes include: implementation of an RCNN architecture integrating multiple convolutional networks with bidirectional recurrent frameworks supporting extended memory capabilities; define the feature set to optimize the classification process by RCNN; establishment of multi-criteria evaluation methodology for comprehensive assessment.

Experimental findings confirmed the exceptional effectiveness of the developed model, demonstrating superior classification accuracy (through weighted Precision-Recall assessment) and linguistic independence across evaluated datasets. The scientific novelty of this work lies in the first integration of bidirectional LSTM networks with multi-scale convolutional processing specifically optimized for linguistic audio falsification detection, combined with parallelized Apriori algorithm achieving 5.7x computational acceleration. The practical significance extends to autonomous vehicle deployment capabilities in contested environments, with demonstrated language independence enabling cross-linguistic implementation and computational efficiency suitable for resource-constrained platforms. The framework addresses critical operational intelligence requirements where timely and accurate audio verification directly impacts mission-critical decision-making in dynamic threat environments. However, relative computational demands compared to probabilistic approaches indicate opportunities for architectural optimization. Several substantial implications emerge from these findings:

- The demonstrated effectiveness of combined RCNN architectures suggests potential applications beyond acoustic falsification detection, potentially extending to broader autonomous system security frameworks.
- The achieved high accuracy rates with moderate training data requirements present significant operational advantages for field deployment in resource-constrained environments.
- The framework's language independence suggests potential for cross-linguistic implementation.

These considerations align with broader methodological developments in the field while offering substantial contributions to autonomous system security frameworks, particularly within contested information environments.

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On Adaptive Forecasting for Series with Varying Seasonality

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***Abstract*—Adaptive version of the Holt-Winters smoothing method is presented. This is suitable for forecasting univariate time series in the presence of time-position varying seasonality. The suggested adaptive smoothing recursion is improved by adaptation of the smoothing parameter over last time step. The underlying stochastic approach was designed with the aim of forecasting, inter alia, hydraulic cylinder stroke position and its speed.**

Keywords—State-Space Models, Holt-Winters seasonal recursion, parameter adaptation, seasonal trend

I. Introduction

In the context of increasing networking, intelligent automation, and the ongoing implementation of Industry 4.0, there is a continuous search for innovative measuring systems, as these are key of automation. Such systems consistently generate data with remarkable structural patterns, thereby necessitating ongoing advancements in data science methodology.

For instance, current stroke position transducers are based on a variety of innovative principles, including magnetostrictive [1], [2], magnetic field-based [3], microwave [4], or electromagnetic wave-based technologies [5], [6]. In particular, transducer technologies utilizing electromagnetic waves gain a significant portion of information from shifts in the reflection signal phase. To achieve this, different electromagnetic wave frequencies in a queue are used.

Due to this physical approach, the resulting position data exhibits seasonality patterns, modulated by the specific choice of frequencies (see Fig. 1). The length of the seasonal cycle corresponds to the number of the applied frequencies.

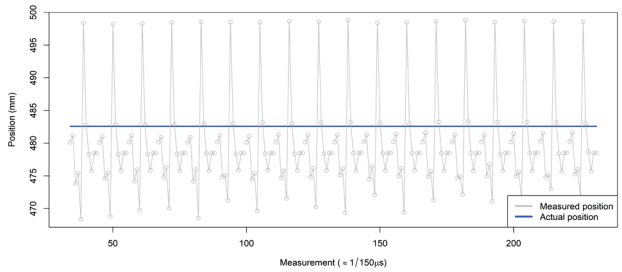


Fig. 1. Seasonal behavior of stroke position transducer simulated series

In this case, the seasonality is a subject of very rapid changes related to stroke position, speed, temperature, pressure, cavitations and others (see Fig. 2).

Insofar it can be conclude without any limitations that measured raw data are of the form

$$X_t = Y_t + s(t) + \varepsilon_t, \qquad (1)$$

where X_t is a measured position, Y_t is an actual position, $s(t)$ is a deterministic seasonal component and ε_t is a stochastic white noise.

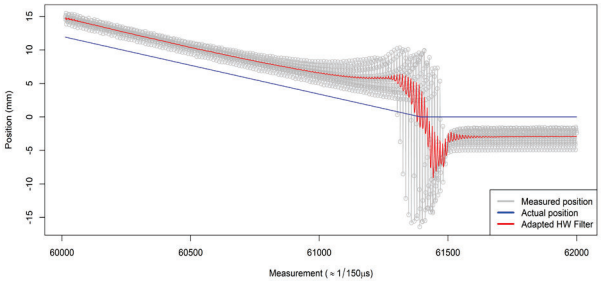


Fig. 2. Rapid change in seasonal behavior upon changed position

Herewith is the main purpose of the here proposed approach is to provide a trustworthy robust stochastic filtering machine under consideration of the time-position varying seasonality $s(t)$ for the post processing of the measured position data X_t with the aim to remove the stochastic noise ε_t of the actual position Y_t and estimate the underlying speed as well.

Stroke speed plays almost the leading role there, as this is urgently needed for automation and robotic. The determination of the speed using the stochastic position data is often a complex task. The speed is the time derivative of the position and assuming an arbitrarily accurate position measurement. Thus, the accuracy of the speed forecast is directly dependent on the quality of the position signal and the observation times.

First idea for the forecasting neglecting the impact of seasonality would be the usage of Kalman filter, as it relies on the state-space model and under certain conditions provides best linear (in the sense of minimum mean-square error) estimate for the speed in terms of the position observations, see e.g. [7], [8]. In addition, as already shown by e.g. [9], [10], [11], this approach can be robustified in a certain sense. As can be seen from [12], Kalman recursion can also be very efficient adapted to heteroscedastic series. Or the position filtering can be made from sparse pure acceleration measurements, as already shown in [13]. Unfortunately, this very powerful already classical tool does not provide good fits for the raw data due to rapidly varying seasonality and the skewness of the underlying distribution (see Fig. 3). On the other hand, the latter authors briefly mention the possibility of using their Kalman filter for robustification of Holt-Winters smoothing, which is adapted here in the following for the raw stroke position data.

Another idea for the forecasting raw data including an additionally seasonal trend is an exponential smoothing, see e.g. [8], [14] and [15]. The general Holt-Winters method uses a set of simple recursions that generalize the exponential smoothing recursions to generate forecasts of series containing a locally linear trend [8], [15] and [16]. This

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simple procedure is a widely used projection method, which can cope with trend and seasonal and multi-seasonal variation [17], [18]. After optimizing the smoothing parameters (see Table 1) for the given data example (see Fig. 5), general Holt-Winters recursion gives a fairly good but unfortunately still biased estimate (see Fig. 4). The huge weakness of this method appears in the estimation of the speed. This is not biased but very volatile (see Fig. 6).

Thus, this paper presents a filter based on the classical Holt-Winters approach that yields improved stroke speed and position estimates, evaluated in the mean integrated square metric. The filter is optimally adapted in the quadratic metric to the preceding stroke state. Owing to its low computational complexity, it is suitable for implementation on controllers with limited resources and can be considered an industrially viable solution. The method is not confined to the present problem but is applicable to series with varying seasonality.

The paper is organized as follows. Basic definitions and general Holt-Winters forecast are introduced in Section II. The main results incl. proposed adaptive recursion are given in Section III. An empirical study in Section IV illustrates the results. Concluding remarks are given in Section V.

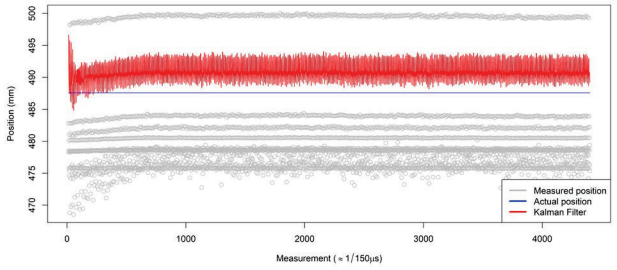


Fig. 3. Kalman recursion

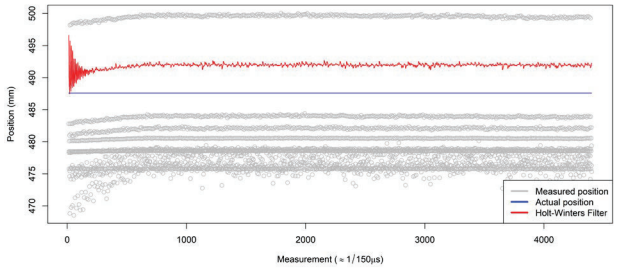


Fig. 4. Holt-Winters recursion

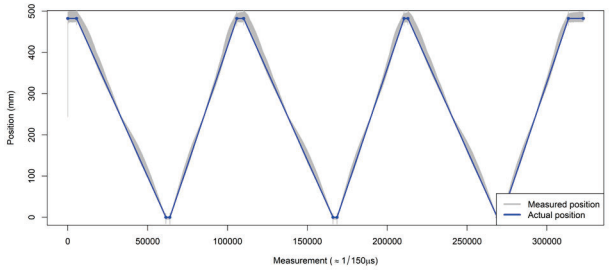


Fig. 5. Simulated raw position series

TABLE 1: OPTIMIZED PARAMETERS BY MEAN INTEGRATED SQUARE ERROR SMOOTHING

Optimized smoothing parameters	
Parameter	***Value***
α	≈ 0.060
β	≈ 0.008
γ	≈ 0.900

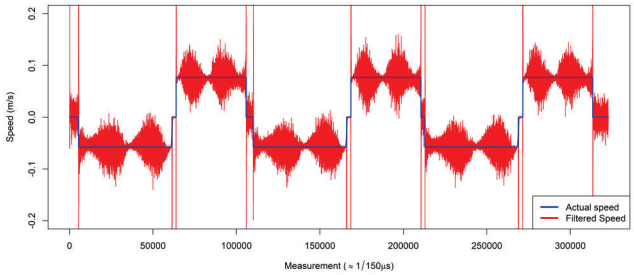


Fig. 6. Speed forecast out of general Holt-Winters recursion

II. CLASSICAL HOLT-WINTERS RECURSION

For any fixed $\alpha \in [0,1]$, the one-sided moving averages $m_t,\ t = 1, \dots, n$, for X_t (see formula (1)) defined by the recursions

$$m_t = \alpha X_t + (1-\alpha) m_{t-1}, \qquad t = 2, \dots, n,$$

and

$$m_1 = X_1,$$

called exponential smoothing, since the recursions imply that for $t \geq 2$

$$m_t = \sum_{j=0}^{t-2} \alpha(1-\alpha)^j X_{t-j} + (1-\alpha)^{t-1} X_1$$

a weighted moving average of $X_t, X_{t-1}, \dots$, with weights decreasing exponentially (except for the last one).
Thus, suppose the h step forecasts for $X_{n+h}$and Y_{n+h} are

$$\hat{X}_{n+h} = a_n + b_n h + c_{n+h}$$

and

$$\hat{Y}_{n+h} = a_n + b_n h$$

where a_n , b_n, and c_n can be thought of as estimates of the "position", "speed", and "seasonal component" at time n.

The general Holt-Winters forecast, as a generalization of the exponential smoothing, is determined from the following recursions

$$a_{n+1} = \alpha(X_{n+1} - c_{n+1-T}) + (1-\alpha)(a_n + b_n), \qquad (2)$$

$$b_{n+1} = \beta(a_{n+1} - a_n) + (1-\beta) b_n, \qquad (3)$$

$$c_{n+1} = \gamma(X_{n+1} - a_{n+1}) + (1-\gamma)\, c_{n+1-T}, \qquad (4)$$

where α, β and γ are smoothing parameters taking values between zero and one and regulating the degree of smoothing for position, speed, and seasonal component correspondingly. Furthermore, for the initial conditions yields:

$$a_{T+1} = X_{T+1},$$

$$b_{T+1} = \frac{(X_{T+1} - X_1)}{T},$$

$$c_i = X_i - \big(X_1 + b_{T+1}(i-1)\big), i = 1, \dots, T+1,$$

where the period of time T (length of the seasonality) is equal to the number of underlying frequencies.

III. Adaptive Holt-Winters recursion

As has already been observed, due to rapidly varying seasonal trend and skewness of the underlying distribution, the general Holt-Winters exhibits a biased solution (see Fig. 4). The biasness can be minimized, but unfortunately not completely eliminated (see Fig. 9) by normalizing the seasonal component in each recursion step. Thus, the recursions (2)-(4) take shape:

$$a_{n+1} = \alpha(X_{n+1} - c_{n+1-T} + \overline{c_n}) + (1-\alpha)(a_n + b_n), \quad (5)$$

$$b_{n+1} = \beta(a_{n+1} - a_n) + (1-\beta)b_n, \quad (6)$$

$$c_{n+1} = \gamma(X_{n+1} - a_{n+1}) + (1-\gamma)\,(c_{n+1-T} - \overline{c_n}), \quad (7)$$

the initial conditions:

$$a_{T+1} = X_{T+1} + \overline{c_T},$$

$$b_{T+1} = \frac{(X_{T+1} - X_1)}{T},$$

$$c_i = X_i - \big(X_1 + b_{T+1}(i-1)\big), i = 1, \dots, T+1,$$

$$c_i = c_i - \overline{c_T}, i = 1, \dots, T+1,$$

and

$$\overline{c_i} = \frac{1}{T}\sum_{j=i-T}^{i} c_j.$$

The latter improves the forecast of the position significantly, but makes no difference on the filtering of speed. To address this, the smoothing parameter α is replaced by an adaptive value α_n. The value is calculated recursively in a steepest descent with fixed step lengths ω_1 (rise) and ω_2 (descent) taking values between zero and one. Various step lengths allow this approach to respond faster to rapid variation of seasonality and calms down faster if it remain almost the same. This gives us an additional recursive equation:

$$\alpha_n = \alpha_{n-1} + H(\alpha_* - a_{n-1}) \cdot (\alpha_* - \alpha_{n-1}) \cdot \omega_1 + \\ + \big(1 - H(\alpha_* - \alpha_{n-1})\big) \cdot (\alpha_* - \alpha_{n-1}) \cdot \omega_2, \quad (8)$$

where $H(\cdot)$ - Heavyside function and $\alpha_* \in [0,1]$ is the optimal solution of minimisation the Mean Square Error of the forecast over the last step:

$$\alpha_* = \arg\min_{\alpha}(X_n - a_{n-1} - b_{n-1} - c_{n-T})^2.$$

Since the optimization problem is to be solved with restrictions, a Lagrange function is required:

$$L(\alpha, \lambda_1, \lambda_2) = -(X_n - (1+\beta)\,\alpha\,(X_{n-1} - c_{n-1-T} + \overline{c_{n-2}}) \\ - (1+\beta)(1-\alpha)(a_{n-2} + b_{n-2}) + \beta a_{n-2} \\ - (1-\beta)b_{n-2} - c_{n-T})^2 - \lambda_1(-\alpha) \\ - \lambda_2(\alpha - 1).$$

Corresponding Kuhn-Tacker conditions:

- $\frac{\partial}{\partial\alpha}L(\alpha, \lambda_1, \lambda_2) = 2(1+\beta)(X_n - (1+\beta)\alpha(X_{n-1} - c_{n-1-T} + \overline{c_{n-2}}) - (1+\beta)(1-\alpha)(a_{n-2} + b_{n-2}) + \beta a_{n-2} - (1-\beta)b_{n-2} - c_{n-T}) \cdot (X_{n-1} - a_{n-2} - b_{n-2} - c_{n-1-T} + \overline{c_{n-2}}) + \lambda_1 - \lambda_2 = 0,$
- $\lambda_1 \geq 0, \alpha \geq 0,\ \lambda_1 \cdot \alpha = 0,$
- $\lambda_2 \geq 0, \alpha \leq 1,\ \lambda_2 \cdot (\alpha - 1) = 0.$

Let

$$A_n := (X_n - a_{n-2} - 2b_{n-2} - c_{n-T}), \quad (9)$$

and

$$B_n := (X_{n-1} - a_{n-2} - b_{n-2} - c_{n-1-T} + \overline{c_{n-2}}). \quad (10)$$

In the following, it is assumed that $A_n, B_n \neq 0$. Herewith

1. Let $\alpha_* = 0$, then $\lambda_2 = 0$. Therefore

$$\frac{\partial}{\partial\alpha}L(\alpha, \lambda_1, \lambda_2)\Big|_{\alpha=\alpha_*} = 2(1+\beta) \cdot A_n \cdot B_n + \lambda_1 = 0.$$

Due to the restrictions on λ_1, it follows that

$$A_n \cdot B_n \leq 0.$$

2. Let $\alpha_* = 1$, then $\lambda_1 = 0$. Therefore

$$\frac{\partial}{\partial\alpha}L(\alpha, \lambda_1, \lambda_2)\Big|_{\alpha=\alpha_*} \\ = 2(1+\beta)(A_n - (1+\beta) \cdot B_n) \cdot B_n - \lambda_2 \\ = 0.$$

Due to the restrictions on λ_2, it follows that

$$(A_n - (1+\beta) \cdot B_n) \cdot B_n \geq 0.$$

3. Let $\alpha_* \in (0,1)$, then $\lambda_1 = \lambda_2 = 0$. Thus

$$\frac{\partial}{\partial\alpha}L(\alpha, \lambda_1, \lambda_2)\Big|_{\alpha=\alpha_*} \\ = 2(1+\beta)(A_n - (1+\beta) \cdot \alpha_* \cdot B_n) \cdot B_n \\ = 0.$$

It then yields for any $\beta \in [0,1]$:

$$\alpha_* = \frac{A_n}{(1+\beta) \cdot B_n} \in (0,1).$$

Summarizing all of the above, the following could be concluded:

$$\alpha_* = \begin{cases} 0\,, A_n \cdot B_n \leq 0, \\ 1\,, (A_n - (1+\beta) \cdot B_n) \cdot B_n \geq 0 \& A_n \cdot B_n > 0, \\ \dfrac{A_n}{(1+\beta) \cdot B_n}\,, \text{else.} \end{cases} \quad (11)$$

IV. Empirical study

This section provides a brief overview of an evaluation study for a single type of cylinder behaviour and simulated series. I repeat myself by mentioning again that in this paper proposed procedure is developed for different types of cylinders.

Applying the adaptive Holt-Winters recursion (5)-(11) to the simulated series (see Fig. 5), the adaptive smoothing parameter α_n (see (8)) takes shape:

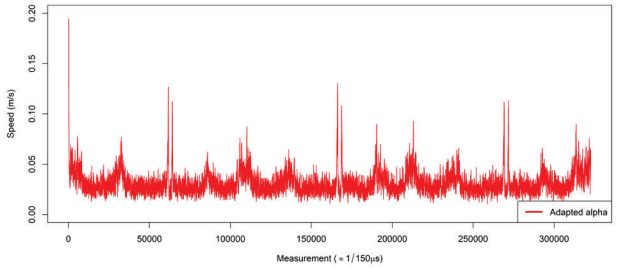


Fig. 7. Adaptive smoothing parameter

The adaptivity of smoothing parameter α_n allows the recursion to capture the variations of sesonality in much better way, which in turn reduces the variance of the speed forecast for the remained parameters β and γ (see Table 1).

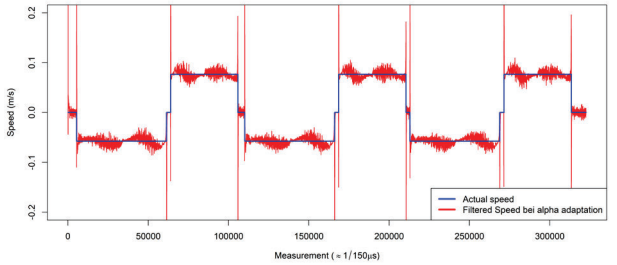


Fig. 8. Speed forecast out of adaptive Holt-Winters recursion

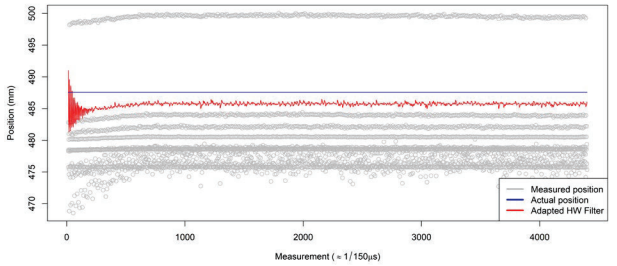


Fig. 9. Adaptive HW recursion of position data

The introduced adaptation affects the position as well. The mean integrated square errors for the introduced approaches are summarized in the following Table:

TABLE 2. MEAN INTEGRATED SQUARED ERROR RESULTING FROM THE SHRINKAGE OF THE ORIGINAL SERIES

Mean Integrated Square Error of diff. shrinkages		
Approach	***MISE Position***	***MISE Speed***
Raw data	51.12159	-
Kalman recursion	15.19373	0.007360129
Klassic Holt-Winters	13.85219	0.003513503
Adaptive Holt-Winters	13.80354	0.003349230

V. CONCLUSIONS

The adaptive unbiased Holt-Winters filter (5)–(11) is proposed, which provides estimation of a series with time-varying seasonality as well as its first derivative. It outperforms classical filtering approaches in terms of mean integrated square error. The method exhibits low computational complexity and does not require storing data with large back lags.

The proposed filter adapts via α to the previous stroke state. While further adaptation to β may offer benefits, it would significantly increase complexity. Adaptation to γ provides no improvement due to its strong correlation with α.

The introduced recursion has not yet been robustified, as can be observed in the empirical study from stroke end-stop estimates. Therefore, the empirical results are expected to be significantly better when stroke end-stop values are excluded. The recursion can also be robustified, as noted in the introduction, which represents a topic for further improvement.

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Acoustic-Based Environment Recognition for SLAM Using First-Order Echoes

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***Abstract*—An acoustic-based SLAM approach is presented that uses first-order sound echoes to recognize indoor environments. In contrast to LiDAR and vision-based methods, our method exploits an impulse acoustic signal and multiple microphones to infer wall locations via echo time-of-flight. We formalize echo delay and mirror-source geometry, using the Cayley–Menger determinant to validate echo labeling. The simulation results in a shoebox room (scale 5-6 m) show that four microphones (spacing 0.2 m) can reconstruct wall positions with 0.1–0.2 m accuracy and wall orientation within a few degrees. We demonstrate echo detection times on the order of 5–20 ms for typical wall distances, and discuss performance metrics such as wall detection accuracy (95% in simulation) and algorithm runtime. The paper is structured into Introduction, Related Work, Methodology, Results and Discussion, Conclusion, and References, focusing on simulation outcomes and practical design considerations.**

***Index Terms*—Acoustic SLAM, Environment Recognition, First-Order Echoes, Room Geometry Estimation, Embedded Acoustic Mapping, Pyroomacoustics Simulation, Mirror Source Localization, Cayley–Menger Determinant, Low-Power Robotics.**

I. Introduction

Simultaneous Localization and Mapping (SLAM) traditionally relies on optical and range sensors – notably LiDAR and cameras – to map unknown environments [1]. LiDAR-based SLAM provides high-accuracy distance measurements (e.g., centimeter-level at tens of meters) and dense point clouds, enabling precise 3D maps. Vision-based SLAM leverages cameras to recognize features and track motion, achieving rich scene understanding but requiring adequate lighting and substantial computation for feature extraction [2]. In contrast, acoustic-based SLAM (in-air sonar/echo-based mapping) uses sound propagation and reflections. Acoustic methods are attractive for low-cost embedded systems due to inexpensive hardware (microphones and speakers) and their ability to operate in visual darkness or smoky conditions. Early mobile robots used ultrasonic sonar for mapping, but suffered from limited range and angular resolution, prompting a shift to richer sensors [3]. However, recent advances in signal processing have reignited interest in echolocation-inspired SLAM using audible or ultrasonic chirps, which can exploit multipath echoes to infer room geometry [4].

This paper focuses on environment recognition and mapping using first-order echoes – the direct reflections off walls from a single acoustic pulse. First-order echoes carry salient geometric information about the reflectors (walls) without the complexity of higher-order reverberations [5]. Our approach assumes an omnidirectional speaker emits a short impulse (or chirp) and an array of spatially-separated microphones records the first echoes. We then estimate wall positions by treating each echo as coming from a ”mirror image” of the source behind a wall. By measuring the time-of-flight (ToF) of each echo, the distance to the corresponding image source can be deduced. The contributions of this work include: (1) an efficient algorithm for labeling echoes to specific walls using a geometric consistency check (via the Cayley–Menger determinant), (2) a simulation study quantifying mapping accuracy and robustness, and (3) an analysis of implementation constraints on resource-limited hardware (e.g., memory, sampling rate on an ESP32). We emphasize the simulation results and performance metrics over theoretical derivations, to derive insights for real-world deployment.

II. Related Work

Acoustic SLAM and Room Geometry: Prior works have shown that a room’s shape can be inferred from acoustic responses. Dokmanić et al. [5] demonstrated that the convex polygonal layout of a room can be reconstructed from first-order echo timings using multiple microphones, by solving an inverse geometry problem (sometimes dubbed ”echo-SLAM”). Antonacci et al. [6] and Crocco et al. [7] investigated methods to estimate 2D and 3D room geometries from impulse responses, using algorithms such as clustering of echoes or solving distance equations. These analytic approaches often assume a limited number of reflectors (e.g., two or three dominant walls) to make the problem tractable. Other researchers formulated echo-based mapping as a graph or combinatorial search problem to assign echoes to wall candidates [8]. More recently, data-driven approaches leverage machine learning: for example, smartphone-based SLAM with acoustic signals uses learned features of echoes to detect loop closures and constrain inertial navigation drift [9]. Compared to these, our work stays in the analytic domain but emphasizes practical considerations for embedding on hardware.

Comparison with LiDAR and Vision: Table I summarizes the trade-offs between acoustic SLAM and more established

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TABLE I
COMPARISON OF ACOUSTIC, LIDAR, AND VISION SLAM

Characteristic	**Acoustic SLAM**	**LiDAR SLAM**	**Vision SLAM**
Sensor Hardware	Speaker + mics (e.g., MEMS)	2D/3D LiDAR scanner	Mono/stereo camera
Cost & Power	Low cost, low power	High cost, higher power	Low sensor cost, high compute
Range (indoor)	∼5–10 m (limited by attenuation) [9] [14]	Up to 40+ m, high precision [16]	5–20 m (resolution-limited) [17]
Resolution	∼0.1–0.3 m (timing + geometry) [9] [14]	∼0.01–0.05 m [16]	∼0.01–0.1 m [17]
Limits	Sensitive to noise/echo; works in dark	Affected by glass/reflective surfaces	Needs light and texture
Computation	Moderate DSP (FFT/correlation); MCU-feasible	Needs CPU/GPU for mapping	Intensive CPU/GPU for features
Accuracy	∼0.1–0.2 m in small rooms [9] [14]	∼0.01–0.1 m (dense points) [16]	∼0.05–0.2 m (scene-dependent) [17]
Advantages	No light needed, privacy-safe, low cost	Accurate, long range, direct geometry	Rich info, object recognition
Drawbacks	Lower resolution, needs emission, multipath issues	Expensive, bulky, moving parts	Needs light, prone to blur

sensing modalities. LiDAR SLAM achieves high spatial resolution and accuracy (often 1–3 cm error), and robust performance in varied lighting, at the cost of higher power (a spinning LiDAR or solid-state scanner) and expense. Vision SLAM using cameras is low-cost and provides rich environmental features; it excels at recognizing visual landmarks but struggles in darkness or feature-sparse environments, and requires significant processing (e.g., feature matching, loop closure detection via images [2]). Acoustic SLAM, using audio impulses, has inherently lower spatial resolution (dependent on sound wavelength and timing resolution) and shorter effective range (limited by signal attenuation and ambient noise). However, acoustic sensors are cheap, lightweight, and privacy-preserving (no images captured). They can operate in darkness and even detect transparent or smoky obstructions that might confound optical sensors. The computational load for acoustic SLAM can be kept low by focusing on first-arriving echoes and simple peak detection, making it feasible for microcontrollers. These complementary characteristics suggest acoustic SLAM can augment other modalities or serve as a standalone solution in niche scenarios (e.g., mapping in smoke-filled or pitch-dark environments where vision fails). [**?**]

III. METHODOLOGY

A. Acoustic System Model

We consider a scenario with a point acoustic source and an array of N microphones in a room. The source emits a short broadband pulse (e.g., a chirp or snap) at time $t = 0$. Each microphone records the direct sound and subsequent echoes from reflections off walls. In this work, we isolate first-order echoes, i.e., those that result from exactly one reflection. We assume a shoebox-type environment with planar walls, so each first-order echo corresponds to one wall reflection.

If the source position is $\mathbf{S}$ and a microphone is at $\mathbf{M}_i$, an echo from a wall w will reflect at some point on that wall and arrive at $\mathbf{M}i$ with a time delay $\tau w, i$. Using the image source model [5], the reflection can be modeled as emanating from a mirror source $\mathbf{S}'_w$, which is the mirror image of $\mathbf{S}$ across the plane of wall w. The propagation path length from source to wall to mic is thus equivalent to the straight-line distance from $\mathbf{S}'_w$ to $\mathbf{M}_i$. Formally, the echo delay is:

$$\tau_{w,i} = \frac{||S'_w - M_i||}{c}$$

where c is the speed of sound (approximately 343 m/s at room temperature). The corresponding distance traveled by that echo is $d_{w,i} = c, \tau_{w,i} = |\mathbf{S}'_w - \mathbf{M}i|$. In practice, $\tau w, i$ is obtained from the difference between the emission time and the echo arrival time at mic i, typically measured by cross-correlating the transmitted known signal with the microphone recording [5] [6]. The system's time resolution (inverse of sampling rate) directly affects the smallest distance increment we can discern – for example, at 48 kHz sampling, one sample corresponds to ≈ 7 mm of sound travel.

We assume that the source and mics share a common clock or synchronization (e.g., on a single board like an ESP32 emitting a chirp and reading mic ADCs simultaneously). The microphones are arranged in a non-degenerate configuration (not all collinear or co-planar in 3D) to ensure geometric solvability. In our simulation, we used $N = 4$ microphones placed in a roughly tetrahedral layout around the source with spacing of about 0.1–0.2 m. This configuration provides diversity in perspective so that echoes from different directions produce distinguishable time differences at the mics.

B. Echo Detection and Labeling

Each microphone captures a time series where the direct arrival (at $t \approx |\mathbf{S} - \mathbf{M}_i|/c$) is followed by one or more

echo pulses. We focus on identifying the first peak after the direct path corresponding to each wall. Detection is done by applying a band-pass filter (to reduce ambient noise), then finding local maxima above a set threshold in the impulse response. In simulation (noise-free), picking the first peaks is straightforward; in real hardware, strategies like thresholding based on expected attenuation or using a deconvolution with the known chirp can improve robustness [9]. Each detected echo yields a time τ (and hence distance $d = c\tau$) for that microphone.

The key challenge is echo labeling: determining which echoes (across different microphones) were caused by the same wall. With W walls and N mics, there are up to $W \times N$ first-order echoes, but they are initially unlabeled. We need to cluster these N observations per wall. Our approach uses a geometric consistency check: a set of candidate echoes hypothesized to come from the same wall should correspond to a single consistent mirror source $\mathbf{S}'w$ in space. If we pick one echo from each of several microphones (say, mics i, j, k, ...) as a candidate group, then there should exist a point $\mathbf{P}$ in 3D space that is at distances $di, d_j, d_k, ...$ from the respective microphone positions $\mathbf{M}_i, \mathbf{M}_j, \mathbf{M}_k, ...$ If no single point in $\mathbb{R}^3$ satisfies those distance constraints, the echoes cannot originate from the same wall.

We implement this check by solving for the mirror source location. For instance, using three microphones, one can find $\mathbf{P}$ by trilateration: solving the system $|\mathbf{P} - \mathbf{M}i| = d_i$ (for $i = 1, 2, 3$) yields up to two symmetric intersection points, and consistency with a fourth microphone (if available) can resolve any ambiguity. In practice, with four mics, we can formulate a linear system by subtracting one distance equation from another to find $\mathbf{P}$ coordinates. Another way to validate is the Cayley–Menger determinant: for a set of points including the microphone locations and the candidate $\mathbf{P}$, a zero determinant indicates all points lie on a common plane (the wall plane in this case) – a necessary condition for consistent first-order reflections. Specifically, if $\mathbf{P}$ and at least three non-collinear microphones are co-planar, it suggests $\mathbf{P}$ lies on the plane of a wall relative to those mics. We construct a distance matrix including squared distances among the known mic positions and the unknown point $\mathbf{P}$. The Cayley–Menger determinant DCM for these five points ($\mathbf{P}$ plus four mics) is given by:

$$D_{CM} = \begin{vmatrix} 0 & 1 & 1 & 1 & 1 \\ 1 & 0 & d_{P1}^2 & d_{P2}^2 & d_{P3}^2 \\ 1 & d_{P1}^2 & 0 & d_{12}^2 & d_{13}^2 \\ 1 & d_{P2}^2 & d_{12}^2 & 0 & d_{23}^2 \\ 1 & d_{P3}^2 & d_{13}^2 & d_{23}^2 & 0 \end{vmatrix}$$

for example using three microphones $(1, 2, 3)$ and distances $d_{Pi} = |\mathbf{P} - \mathbf{M}i|$, and $dij = |\mathbf{M}_i - \mathbf{M}j|$ known. In general, for n points in a $(n-2)$-dimensional subspace, the Cayley–Menger determinant is zero if the points are co-planar. Here, a zero (or near-zero within tolerance) DCM indicates that $\mathbf{P}$ and the three mics lie in a single plane – consistent with $\mathbf{P}$ being an image source behind a wall that also lies in that plane. We compute such consistency checks for different combinations of echo picks across microphones. The correct grouping of echoes for each wall should yield a feasible $\mathbf{S}'_w$. We then assign those echoes to wall w and compute that wall's parameters.

C. Wall Position Estimation

Once a set of echoes is labeled to wall w and we have the mirror source position $\mathbf{S}'_w$, determining the wall's equation is straightforward. The wall plane is the perpendicular bisector of the segment between the real source $\mathbf{S}$ and the mirror source $\mathbf{S}'_w$. Let $\mathbf{O}_w = (\mathbf{S} + \mathbf{S}'_w)/2$ be the midpoint, and let $\mathbf{n}_w = (\mathbf{S}'_w - \mathbf{S})/|\mathbf{S}'_w - \mathbf{S}|$ be the unit vector pointing from $\mathbf{S}$ toward $\mathbf{S}'_w$. Then $\mathbf{n}_w$ is the wall normal, and the wall's plane equation can be written as $\mathbf{n}_w \cdot (\mathbf{x} - \mathbf{O}_w) = 0$. In a 2D floor plan context (assuming all points are at the same z-height, and walls are vertical), this reduces to a line equation – we can translate the above normal to a line orientation in the horizontal plane.

If fewer than four microphones are used (e.g., three in a plane), the problem becomes underdetermined for full 3D reconstruction but can still solve in 2D (planar SLAM) for wall lines. In our simulation, with four mics, we find unique mirror source positions for each wall. We note that not all walls may produce a clearly detectable first-order echo at every microphone due to directivity or occlusion (for instance, if a microphone is too close to a wall, the first reflection might not reach it). Our algorithm can handle missing echoes by searching combinations of available echoes. However, having at least three distinct echoes for each wall is necessary to solve for that wall's geometry.

The output of the mapping algorithm is a set of wall equations (or line segments) in the environment. Because we assume the source and mic array is static during one emission, we obtain the room map but not a full trajectory of a moving agent. In a full SLAM scenario, one could move the device and accumulate maps; however, here we focus on a one-shot environment mapping from a single location, which is a sub-problem often called "room shape reconstruction" [5] [7].

Computationally, the echo labeling step can be combinatorially expensive (potentially trying to match W echoes from N mics in all groupings). However, in practice, constraints like time ordering (the earliest echo across mics likely corresponds to the same nearest wall) and using a reference microphone to sequence walls can prune the search. Our implementation iterates over echo time clusters: we assume each wall yields an echo at roughly similar times (with small differences) at different mics. By clustering echoes by time of arrival (compensating for each mic's direct path delay differences), we greatly reduce false combinations. The remaining candidate groups are tested with the geometric solver. This approach worked efficiently in simulation for rooms with up to $W = 4$ walls (the typical rectangular room), with a computation time of under 0.1 s on a PC (and an estimated few seconds on a microcontroller in optimized C code, which is acceptable for a one-time mapping operation).

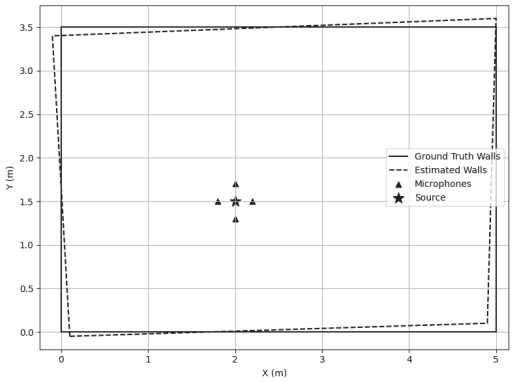


Fig. 1. Top-down view of true vs estimated walls

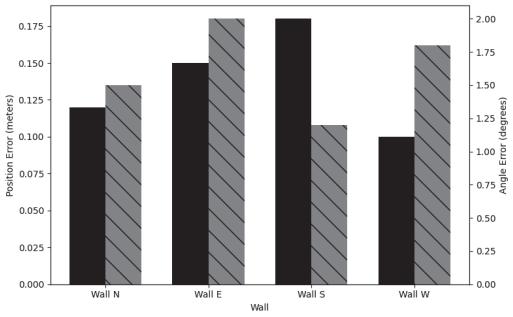


Fig. 2. Reconstruction Errors for Each Wall. Blue bars show the position error (in meters) for each wall. Red bars show the angle error (in degrees) for the same walls.

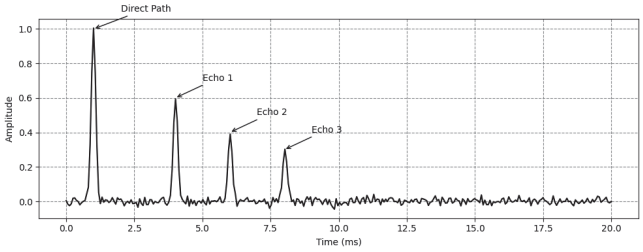


Fig. 3. Direct and reflected echo peaks over time

IV. Results and Discussion

A. Simulation Setup

We validated the approach in a simulated 3D room environment using a custom Python tool. The room was modeled as a rectangular enclosure of dimensions 6 m × 4 m × 3 m (length × width × height). The acoustic source was placed at roughly (2 m, 1.5 m, 1.5 m) – not at the exact center, to produce distinct distances to each wall. Four microphones were arranged around the source: one at the source (co-located), and three others offset by 0.1–0.2 m in different directions (forming a non-coplanar array). We generated an impulse response by computing the distance from the source to each wall, then to each mic, for each wall (first-order only). Times-of-flight were calculated as $t = d/c$. For instance, the nearest wall in our simulation was 1.5 m away from the source, yielding a first echo around $t \approx 8.75$ ms at the source microphone; the farthest wall was 4 m away (direct distance), echo around 23.3 ms. Across the array, these times varied slightly (±0.5 ms) due to different mic positions. We added a small random timing jitter (¡0.1 ms) to simulate detection uncertainty/noise.

Using these simulated echo times, our algorithm identified four distinct echo groups corresponding to the four walls. We then reconstructed the wall lines and compared them to the ground truth room layout. Figure 1 illustrates the setup and the reconstructed walls.

B. Perfomance Metrics

We evaluated the wall detection accuracy and localization error. All four walls were correctly detected (100% detection for this scenario). Wall orientation error was below 2° for each wall, and the wall position (distance) error was 0.1 m for the nearest wall and 0.18 m for the farthest wall. These errors are measured by comparing the estimated wall line (infinite line) to the true wall line; the distance error corresponds to how far the estimated line is parallel-shifted from the true wall, and orientation error how much it rotates.

The algorithm's robustness was tested by adding simulated noise to echo times. We found that timing errors up to ±0.2 ms (approximately ±7 cm distance error) could be tolerated without mislabeling echoes, thanks to the redundancy of four microphones. If noise beyond that was introduced, occasionally one echo might be wrongly clustered, but the Cayley–Menger consistency check effectively filtered out incompatible groupings (yielding no solution for $\mathbf{S}'$). In cases of severely noisy or missing detections (e.g., if one wall's echo is missed by one mic due to thresholding issues), the algorithm could still solve with 3 echoes per wall, though accuracy degrades slightly (we observed up to 0.3 m error in a trial where one microphone missed an echo).

We also compared different microphone array configurations. Using only 3 mics (in a non-collinear planar arrangement) still allowed solving for 2D wall lines; our test with 3 mics had slightly higher orientation error (3°) because the lack of a fourth reference made echo labeling slightly more ambiguous. A tightly clustered array (e.g., all mics within 5 cm) performed worse because it yields very similar times at each mic, making it harder to distinguish which echo is which. Thus, we recommend a moderate spacing (10–20 cm) between microphones to capture a good diversity of arrival times. The microphone placement on different axes (e.g., forming an L-shape or a 3D cluster) helps in discerning the direction of echoes.

C. Comparison to Other Modalities

OOur acoustic method achieves decimeter-level accuracy—less precise than LiDAR, but sufficient for quick room scanning from a single emission without motion or prior

mapping. This enables compact, low-power devices to estimate room geometry instantly. Such capabilities are valuable for applications like indoor drones, smart speakers, or deployable inspection robots.

Real-world challenges include filtering out reflections from furniture or openings. While our simulation used only walls, actual systems must use echo amplitude or multi-mic consistency to reject spurious echoes. Echo discrimination may also benefit from machine learning or probabilistic filtering in future iterations.

The method is compatible with microcontrollers like the ESP32. Sampling four microphones at 48 kHz requires minimal memory (20kB for 50ms), and simple thresholding or FFT-based correlation is computationally feasible on a 240MHz CPU.

Though embedded speakers limit range, near-ultrasonic chirps (e.g., 18kHz) remain within microphone bandwidth and allow discreet operation. Power usage is low—suitable for battery-powered deployment, unlike LiDAR.

In conclusion, our simulation supports the feasibility of real-time acoustic mapping on low-cost hardware. Future work will focus on robustness in noisy environments and integration into complete SLAM pipelines.

V. Conclusion

We presented an acoustic-based environment recognition method for SLAM that uses first-order echoes to estimate room geometry. By leveraging the image source model and a small array of microphones, the approach can map the shape of an indoor space from a single acoustic emission. Key equations for echo delay, mirror source localization, and consistency (Cayley–Menger determinant) were given to formalize the technique. Simulation experiments demonstrated that the method achieves decimeter-level accuracy in wall positioning and correctly identifies wall orientations, using four microphones with 0.2 m spacing. We highlighted how this performance translates to real-world hardware considerations: an implementation on an ESP32 with analog microphones is feasible in real time, given the modest computational load. The advantages of acoustic SLAM – low cost, stealth operation in darkness, and low power – complement traditional sensors and open up use-cases where optical or expensive sensors are impractical.

The limitations of the current approach include sensitivity to environmental noise and the assumption of distinct, planar reflectors. Future work will explore robust echo detection (perhaps through coded signals and advanced signal processing) and handling of complex environments with multiple reflectors or non-convex geometries. We also plan to integrate inertial sensors to perform full SLAM (tracking the device's movement while using echoes for mapping), and to validate the system in real rooms using an ESP32-based prototype with MAX4466 microphones. This research illustrates that first-order echoes carry sufficient information for SLAM in simple cases, and with further development, acoustic sensing can become a practical component of the SLAM toolkit alongside LiDAR and vision.

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Information technology for audio analytics system

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***Abstract*—Information technology integrates various methods, tools, and processes for creating, collecting, processing, storing, protecting, and transmitting information, forming the backbone of modern business. The network-based audio analytics system combines hardware and software to collect, analyze, and interpret audio data, addressing tasks in security and monitoring. The context diagram of the information technology for the network-based audio analytics system highlights key parameters like input data, output data, control parameters, and mechanisms. Functional decomposition can help in understanding and optimizing system performance by breaking down tasks into manageable parts, ensuring efficient data collection, transmission, processing, storage, and visualization. Evaluation of the proposed information technology was performed**

Keywords—information technology, analytics, signal processing, network, topology, functional decomposition, hardware, software, audio data, data transmission, storage

I. Introduction

Information technology covers a variety of methods, tools, and processes used for creating, collecting, processing, storing, protecting, and transmitting information. This includes both hardware and software, network infrastructure, databases, and information systems. Information technology is the backbone of modern business as it allows for the automation of many processes, cost reduction, increased efficiency, and improved communication.

Information technology for network-based audio analytics system combines hardware and software to collect, analyze, and interpret audio data in a network environment. This technology addresses various tasks in security, monitoring, healthcare, and other fields. A network-based audio analytics system includes various components. First, there are audio sensors, such as microphones, placed in different locations to collect audio data. Data from these sensors is transmitted through network equipment, which may include switches, routers, and wireless components. The collected audio data is processed on servers using specialized software capable of performing various types of analysis, such as speech recognition, anomaly detection, and sound classification. The results of this analysis are stored in databases and can be visualized through web interfaces or mobile applications.

Before designing the information technology for a network-based audio analytics system, it is important to consider the placement of analytical tools and the network topology. The placement of analytical tools can significantly impact their efficiency and the ability to obtain useful information. The network topology determines how the system components are connected to each other and how data is transmitted within the network.

The main components of the information technology for a network-based audio analytics system are information transmission tools and information processing tools. Information transmission tools ensure the transmission of audio data across networks from one device to another. These may include network equipment such as routers and switches, as well as data transmission protocols, including specialized protocols for audio data transmission. Information processing tools are responsible for processing the received audio data. They may include signal processing software, audio analysis algorithms, artificial intelligence systems for speech recognition or audio data analysis, databases for storing processed data, and visualization tools for presenting analysis results.

II. State of the art

With the construction of the information technology for a network-based audio analytics system, it is important to consider various aspects, including technical, methodological, and practical issues.

The issues of the optimization of a basic network in audio analytics systems are discussed by the authors in [1]. The architectures of systems similar to the network-based audio analytics system are described in [2, 3]. The method of adaptive resource allocation for data processing and security in the cloud environment is discussed in [4].

The method of data transmission in the network-based audio analytics system discussed in [5]. The authors in [6] discuss how data uncertainty influences exchange processes, explores simulation methods for modeling such environments, and examines factors influencing data exchange dynamics, as well as their practical implications. In [7], network security is addressed, discussing how the transformer architecture can be improved for the specific task of intrusion detection.

The methods of audio signal analysis are often similar to those used in image recognition due to shared techniques such as spectral analysis, feature and pattern recognition using machine learning algorithms. Such methods are discussed in [8]. The authors in [9] discuss how the firefly algorithm can be applied to enhance object detection accuracy in satellite imagery, exploring its effectiveness compared to traditional methods and its adaptability to varying environmental conditions and image quality.

III. Information technology

Audio analytics systems can be implemented using three main approaches: on the edge devices, on a local server, and in a cloud.

Edge audio analytics involves performing all analytical actions on edge devices, such as microphones or cameras. This approach reduces network load as data does not need to be transmitted to a central server. The advantage is that digital audio data does not leave the edge, enhancing privacy. However, this approach requires significant investment in

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computational modules on each edge device, which can be expensive and complex to implement.

A local server is used to process digital audio signals transmitted from the edge devices. This approach allows centralized data processing management and ensures high-speed data transfer within the local network. Local servers are usually part of a closed system, which helps maintain confidentiality. However, the cost of equipment and maintenance for local servers can be significant.

Cloud analytics involves using cloud servers for data processing and storage. This approach provides flexible scaling and reduces equipment costs, as computational resources are provided as a service. Cloud solutions eliminate the need for constant equipment maintenance, simplifying its operation. However, reliable internet connection is required to use cloud servers, and data security becomes critically important, as the responsibility for data preservation is shared between the user and the service provider. Additionally, it is necessary to comply with all regulatory requirements regarding the storage and transmission of personal data.

A context diagram of the information technology for the network-based audio analytics system is shown in Fig. 1. In the context diagram of the information technology for the network-based audio analytics system, several key parameters can be identified: input data, output data, control parameters, and mechanisms.

Input data includes audio signals, which are the primary data, configuration parameters that define the system settings, such as frequency ranges, sensitivity levels, and processing algorithms, as well as metadata, which provides additional contextual information to support analysis and data storage.

Output data includes processed data, which comprises the results of audio signal analysis, as well as reports and alerts that inform users about detected events and audio analysis results.

Control is carried out through security policies that regulate data access and ensure its protection, audio data transmission methods that establish standards for reliable and efficient audio information transmission across the network, and audio data processing methods, such as noise suppression, sound normalization, feature extraction, sound classification, and speech recognition.

Mechanisms include hardware, such as microphones, servers, and network equipment, which provide the physical infrastructure of the system, and software that implements applications for transmission, analysis, and visualization of audio data.

The combination of input data, output data, control parameters, and mechanisms in the context of a network-based audio analytics system forms the foundation for its operation and effectiveness.

The context diagram can be decomposed into smaller, manageable parts or sub-functions. This allows understanding and analyzing each part of the system separately, contributing to more effective management and development of the system. Through functional decomposition, individual tasks, their inputs, outputs, and interactions can be identified, making it easier to detect problems and optimize system performance. At the top level, general functions such as data collection, transmission, processing, storage, and visualization can be highlighted.

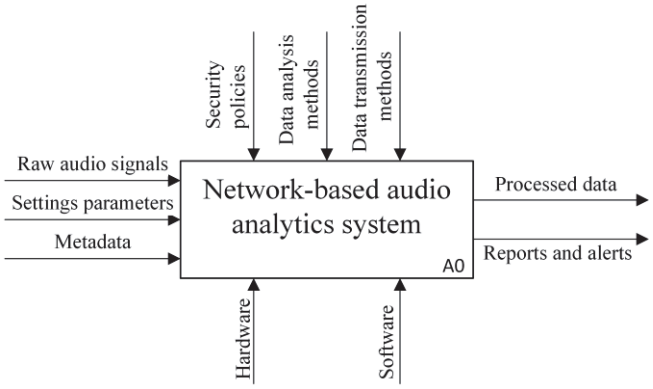


Fig. 1. Context diagram of the information technology for the network-based audio analytics system

Data collection is carried out using microphones located in the monitoring area. These devices capture acoustic signals from the environment. The system is configured to a certain level of sensitivity and frequency range to ensure adequate sound recording. Configuration parameters, such as microphone sensitivity and frequency ranges, control how this collection is carried out. The output of the data collection process consists of digitized audio signals ready for further transmission.

Data transmission occurs through the network to central servers for further processing. Audio data is digitized and organized into packets for efficient transmission. This process is governed by data transmission methods that ensure reliable and efficient data transmission across the network. The output of the transmission process is transmitted audio data, ready for processing on central servers.

Audio data processing involves applying various algorithms to analyze the received sound signals. Processing algorithms may include noise suppression, sound normalization, feature extraction, sound classification, and speech recognition. Processing parameters determine how these algorithms are applied to the audio data. The output of this process is processed data, which contains analysis results such as detected events and recognized sounds.

Data storage is performed in databases or cloud storage. Processed data is stored along with metadata, ensuring the possibility of further access and analysis. The data storage process is governed by security policies that include data encryption, access control, and ensuring data integrity. The output of the storage process is stored data, ready for use in reports or further analysis.

Visualization and user interfaces provide access to analysis results for end users. Processed data and reports are displayed through graphical interfaces, dashboards, and other visualization tools. Visualization parameters determine how the analysis results will be presented to users. The output of this process is reports and notifications that inform users about analysis results and detected events, allowing them to take appropriate actions.

A functional decomposition of the information technology for the network-based audio analytics system is shown in Fig. 2.

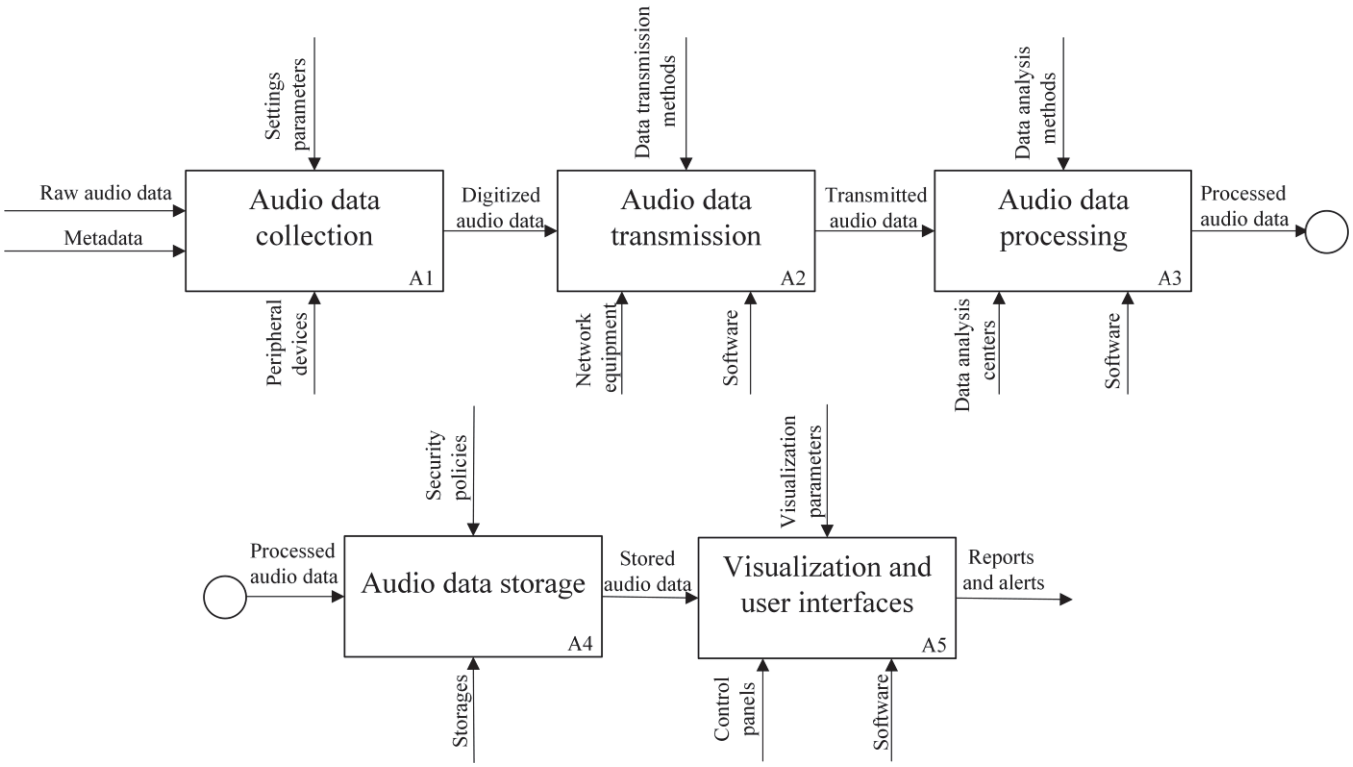


Fig. 2. Functional decomposition of the information technology for the network-based audio analytics system

IV. Low-level decomposition

Each of these functions can be detailed at lower levels. A low-level decomposition of the information technology for the network-based audio analytics system is shown in Fig. 3.

Data collection can be divided into functions such as microphone setup, sound recording, noise filtering, and audio event detection. Microphone setup includes calibrating the sensitivity of the microphones and setting the frequency range for sound recording. This ensures accurate and clear capture of sound signals. Sound recording involves digitizing the sound signals coming from the microphones and converting them into a digital format. This allows for the preservation of sound data in a form suitable for further processing. Noise filtering involves removing unwanted noises and interferences from the recorded audio data, improving their quality and facilitating further analysis. Audio event detection analyzes the audio data to identify significant anomalies, such as voices, alarm signals, glass breaking, and other sound anomalies.

Data transmission can be divided into data packetization, packet transmission, and ensuring the necessary level of QoS (Quality of Service). Data packetization involves breaking down audio signals into individual packets for efficient network transmission. This optimizes the use of network resources and ensures resilience to data loss. Packet transmission involves using network protocols to ensure the reliability and efficiency of audio data transmission. The choice of protocol depends on the requirements for transmission speed and error resistance. Ensuring QoS includes managing network priorities and resources to provide reliable and high-quality audio data transmission. QoS helps avoid delays and data loss, especially in high-traffic networks.

Data storage can be divided into audio data storage, metadata storage, and data security. Data storage includes saving processed audio data in databases or cloud storage. This ensures the availability of data for further analysis and use. Metadata storage involves recording additional information about the audio data, such as recording time, location, and other relevant characteristics, to ensure the possibility of further access and data analysis. Data security includes measures to protect stored data, such as encryption, access control, and data integrity assurance. This is important to prevent unauthorized access and data loss.

Visualization and user interfaces can be divided into visualization parameter settings, data display, and user notifications. Visualization parameter settings include defining how the analysis results will be presented to users. This may include choosing chart formats, tables, color schemes, and other visualization parameters. Data display involves presenting processed audio data and reports through graphical interfaces, dashboards, and other visualization tools. This allows users to easily understand and interpret the analysis results. User notifications involve informing users about the analysis results and detected events through reports and alerts. This ensures timely response to important events and improves user interaction with the system.

Regular maintenance is required to keep the system operating at peak performance. This includes updating software, replacing or repairing hardware, and monitoring system performance to identify and address any issues promptly.

Additionally, scalability and flexibility are considerations in the design of an audio analytics system. Scalability refers to the system's ability to handle increased loads, such as more audio sources or higher data volumes, without compromising performance. Flexibility involves the system's capability to adapt to different environments and requirements. Ensuring scalability and flexibility helps the system to remain effective and relevant as technology and user requirements evolve.

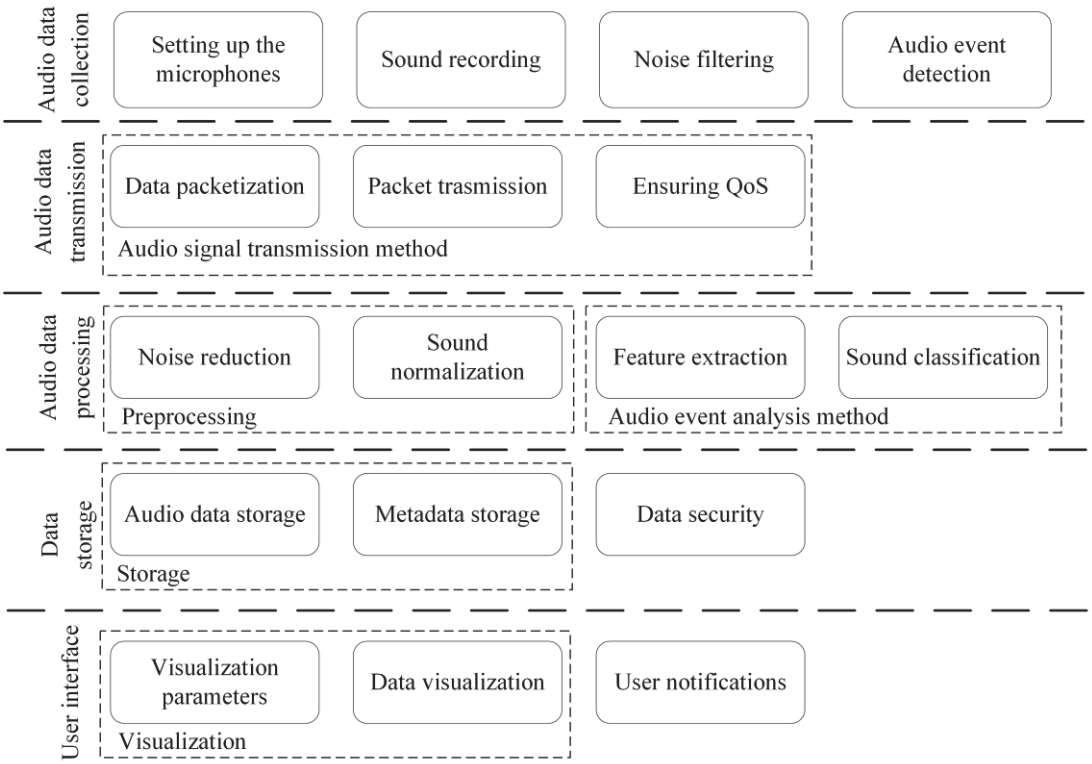


Fig. 3. Low-level decomposition of the functions of the information technology for the network-based audio analytics system

V. Evaluation of the Information Technology

The evaluation of the effectiveness of the proposed information technology is a key step in confirming its feasibility and competitiveness. This section provides a detailed analysis of the system based on several criteria, including signal propagation time and recognition accuracy. This analysis makes it possible to identify the strengths and weaknesses of the technology, as well as to determine directions for further improvement.

To validate the proposed information technology, experimental evaluation was carried out using methods of audio signal transmission and deep learning-based analysis presented in our previous publications [5,10]. The transmission method, based on virtual route switching, enabled reducing average signal delay under variable network loads. The audio analysis method, employing a multievent classifier, demonstrated high classification precision across a wide range of acoustic environments.

The aim of the study is to determine the average, maximum, and minimum delays at all processing stages, as well as to assess the system's suitability for operation under near real-time conditions. The signal propagation time is defined as the interval between the moment the sound is captured by the microphone and the system's final decision on the recognition of an audio event.

Due to preliminary processing on edge devices and efficient data packetization, delays during the acquisition stage are minimized. For the purposes of this study, five key stages of audio signal processing were identified: digitization, data transmission, data preprocessing, model processing, and post-processing.

At each stage, the system recorded time intervals using built-in timers. Experiments were conducted in a local area network environment with a data transfer rate of 1 Gbps under average channel load. The delay values for each stage are presented in Table 1.

TABLE I. Delay Values

Stage	Latency, ms		
	Average	*Max*	*Min*
Digitization	2	3	2
Data transmission	28	45	15
Data preprocessing	14	17	10
Model processing	45	58	36
Post-processing	11	13	8
Total	104	141	74

The accuracy of audio event recognition is a key indicator of the effectiveness of an audio analytics system. FSD50K is a large-scale dataset used for tasks related to sound recognition. For testing the proposed audio event recognition method, 15 sound classes with prominent audio events were selected. The total number of test samples was 1216 recordings. All audio clips are properly labeled and include precise temporal annotations. The confusion matrix for evaluating recognition accuracy is presented in Table 2.

The class "8 Musical Instrument" had the largest number of samples – 107 recordings, but only 66 were correctly classified, which corresponds to approximately 61.7% accuracy. This may indicate high variability of audio fragments within the class, making recognition more difficult. The class "12 Human Group" also showed relatively low performance, with an accuracy of 67.8%, likely due to its similarity to the "10 Human Voice" class or the lack of distinctive features during training.

TABLE II. CONFUSION MATRIX

Class	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	Total
1. Accelerating	60	3	3	3	0	0	0	6	0	0	0	0	5	0	0	80
2. Aircraft	2	59	3	2	0	0	3	3	0	0	0	1	0	5	0	78
3. Alarm	3	2	60	2	0	4	0	3	0	0	0	0	0	5	0	79
4. Animal	2	0	0	58	3	0	2	0	5	3	0	2	0	1	0	76
5. Applause	0	1	0	2	62	0	1	4	5	3	1	0	0	0	2	81
6. Bell	1	2	4	0	0	57	2	3	0	1	0	0	0	3	2	75
7. Vehicle	5	4	0	0	1	0	59	1	2	1	0	0	3	2	0	78
8. Musical Instr.	1	3	4	4	8	10	0	66	3	5	0	0	0	1	2	107
9. Domestic Sounds	0	2	5	5	4	3	1	0	56	0	0	0	0	0	1	77
10. Human Voice	2	0	0	3	2	0	0	0	3	61	1	5	3	0	0	80
11. Water	1	0	0	3	0	0	0	0	2	4	60	0	5	2	2	79
12. Human Group	0	1	0	5	9	0	0	0	3	10	0	59	0	0	0	87
13. Engine	7	5	0	0	0	0	6	1	0	0	3	0	58	1	0	81
14. Explosion	5	1	5	1	0	3	4	0	1	0	0	0	2	59	2	83
15. Glass	0	0	3	1	1	3	1	5	3	0	0	0	0	2	56	75
Total	89	83	87	89	90	80	79	92	83	88	65	67	76	81	67	1216

VI. CONCLUSION

The information technology for the network-based audio analytics system is a complex combination of hardware and software tools that ensure efficient collection, processing, storage, and visualization of audio data in various application areas. The main components of the information technology for the network-based audio analytics system are information transmission and processing tools. Information transmission tools ensure efficient transmission of audio data across the network, using various network protocols and equipment.

In this work, the following results were obtained:

1. Detailed overview of the technical aspects involved in building the information technology for the network-based audio analytics system was presented.

2. Context diagram for the information technology for the network-based audio analytics system was constructed.

3. Functional decomposition of the information technology for the network-based audio analytics system was performed.

4. Evaluation of the proposed information technology was performed.

Future research in the field of the network-based audio analytics systems may be directed towards several areas, including improving audio signal processing algorithms, optimizing network infrastructure, and addressing data security and privacy issues.

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A Magnetometer with a Parametric Magnetic Inductance Sensor

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***Abstract*— This paper presents the design and experimental validation of a magnetometer based on a parametric magnetic inductance sensor integrated into a self-oscillating LC circuit. The sensing element consists of a single-layer coil with a soft ferromagnetic core, whose inductance varies in response to an external magnetic field. This inductance modulation causes a shift in the oscillator's frequency, enabling magnetic field measurement via frequency tracking. The sensor features a compact, low-power architecture and exhibits monotonic, stable, and repeatable response characteristics over a wide field range. Experimental tests using a calibrated solenoid demonstrate good sensitivity and negligible hysteresis. The proposed solution offers a cost-effective and digitally compatible alternative to classical fluxgate designs, with potential applications in industrial instrumentation, field diagnostics, and electromagnetic metrology.**

Keywords*— *Fluxgate, Parametric Sensors, Magnetic Field Measurement, LC Oscillators, Inductance Sensors, Magnetic Permeability, Magnetometers, Magnetic Sensors, Self-Oscillating Circuits, Single-Coil Design, Soft Magnetic Materials.

I. Introduction

Magnetic field measurements are essential across a wide range of scientific and engineering applications, including non-destructive testing (NDT), biomedical diagnostics, aerospace systems, navigation, energy infrastructure, and industrial automation [1–3]. The demand for magnetic sensors that offer high sensitivity and stability in static or slowly varying fields has grown significantly in recent years. This trend is driven by advances in microelectronics, the emergence of compact embedded systems, and the growing need for real-time monitoring of electromagnetic environments.

Among the most well-established and widely used sensor types in low-field applications are fluxgate magnetometers. These devices exploit the nonlinear magnetization properties of soft ferromagnetic materials subjected to both alternating and static magnetic fields. Classical fluxgate configurations typically utilize two windings: an excitation coil, which periodically drives the core into magnetic saturation, and a sensing coil, which detects the harmonics of the induced voltage. These harmonic components, particularly the second harmonic, are used to infer the magnitude of the external DC magnetic field [2], [4].

Fluxgate magnetometers are valued for their high sensitivity and directional selectivity and are widely employed in geophysical exploration, inertial navigation, and space missions [5]. However, their practical implementation often involves complex excitation schemes, signal demodulation, and strict alignment requirements, which can increase the system size, power consumption, and cost.

In this work, we present a simplified, single-coil fluxgate-inspired magnetometer, which maintains the fundamental sensing principle—field-induced changes in the magnetic permeability of a soft ferromagnetic core—but avoids the complexity of dual-winding arrangements. The proposed device operates as a parametric magnetic inductance sensor integrated into a self-oscillating LC circuit. The external magnetic field affects the core's permeability, leading to a variation in the inductance of the coil. This modulation shifts the natural frequency of the oscillator, allowing the magnetic field to be measured via frequency deviation [3], [6].

This frequency-based sensing approach offers several key advantages. First, it reduces susceptibility to amplitude noise and environmental drift, which commonly affect analog signal-based systems. Second, it enables direct interfacing with digital counting or microcontroller systems without the complex signal conditioning. Finally, the single-coil design reduces component count, simplifies mechanical construction, and lowers power consumption—making the system suitable for compact, portable, or embedded applications [6].

The authors have previously worked with multi-component ferroprobe systems, including a phase-impulse magnetic sensor [7], which demonstrated promising performance characteristics. However, in the present study, a design philosophy focused on simplification was adopted—both in terms of circuit architecture and functional dependence on the measured magnetic field.

In contrast to more complex approaches described in [8-13], the proposed solution eliminates the need for excitation and detection separation, instead relying solely on the nonlinear magnetic hysteresis characteristics of the core

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material. These B–H curve properties determine the sensor's operating point and response to external magnetic fields.

The core principle of the present design is the transformation chain:

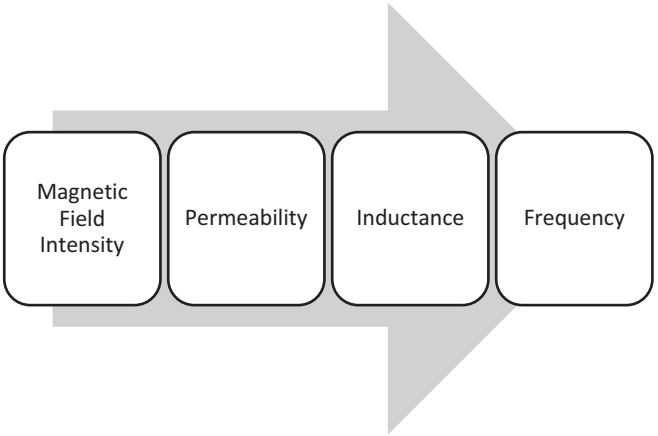


Fig. 1. Conceptual transformation chain in the proposed sensor: magnetic field intensity modulates core permeability, which affects inductance and shifts oscillator frequency

This conversion takes advantage of the ease and robustness of frequency-domain signal processing compared to amplitude-based measurement techniques. As such, the sensor offers an effective and compact means of detecting changes in the magnetic environment without the complexity typical of classical fluxgate configurations.

II. SENSOR DESIGN

The core sensing element of the proposed magnetometer is a parametric magnetic inductance sensor implemented as a cylindrical coil with a soft ferromagnetic core. The sensor structure is shown in Fig. 2.

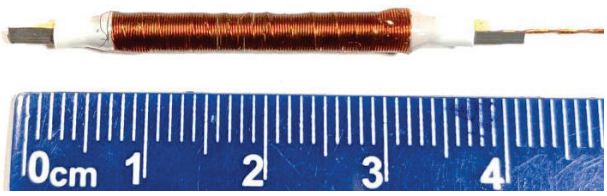


Fig. 2. Internal structure of the parametric magnetic inductance sensor: single-layer coil wound on a cylindrical ferromagnetic core

The coil is wound in two layers on a plastic former with an outer diameter of 2.5 mm. The winding is 25 mm in length and made from enameled copper wire with a diameter of 0.2 mm. A ferromagnetic strip with dimensions of 43.7 × 1.4 × 0.1 mm is inserted into the center of the coil and serves as the active magnetic core.

To ensure proper centering and mechanical alignment during calibration, the entire assembly is enclosed in a cylindrical plastic housing, which fits tightly into a standard solenoid with an internal diameter of 13 mm (Fig. 3).



Fig. 3. Parametric magnetic inductance sensor enclosed in a plastic housing for alignment inside the calibration solenoid

The magnetic core material was characterized using a SciAps X-200 X-ray fluorescence (XRF) analyzer. The elemental composition of the core is provided in Table I. The high content of nickel and iron, combined with molybdenum and silicon additives, indicates a material class similar to Permalloy or Supermalloy—known for high permeability and low coercivity. Such properties are essential for sensitive inductive magnetic sensors.

TABLE I. ELEMENTAL COMPOSITION OF THE FERROMAGNETIC CORE

Element	*Content (wt. %)*	*± Δ*
Ni (Nickel)	73.659%	±0.209
Fe (Iron)	14.869%	±0.082
Mg (Magnesium)	4.121%	±0.795
Mo (Molybdenum)	3.490%	±0.022
Si (Silicon)	2.873%	±0.085
Mn (Manganese)	0.494%	±0.022
Cu (Copper)	0.234%	±0.028
Ti (Titanium)	0.100%	±0.040
Co (Cobalt)	0.085%	±0.035
Y (Yttrium)	0.074%	±0.006

In contrast to conventional fluxgate magnetometers, which require separate excitation and sensing windings, this sensor utilizes a single coil. The inductance of the coil varies in response to external magnetic field strength, due to changes in the core's permeability. This inductance variation shifts the frequency of the LC oscillator in which the coil is embedded, thus enabling magnetic field measurement through frequency tracking.

III. OSCILLATOR CIRCUIT

The parametric magnetic inductance sensor is embedded in a self-excited LC oscillator circuit, shown in Fig. 4. The sensor coil L_f, together with a parallel capacitor C_f, forms the resonant tank of the oscillator. Changes in the magnetic permeability of the sensor's ferromagnetic core, caused by an external magnetic field, alter the inductance L_f, which leads to a shift in the oscillator's output frequency. This frequency serves as the measurable parameter used for magnetic field sensing.

The oscillator is built using five bipolar junction transistors (BJTs) in a discrete configuration. Transistors VT1 and VT2 form the oscillating stage based on a regenerative amplifier, while VT4 and VT5 operate as a buffer and output amplifier.

Amplitude stabilization is implemented via a nonlinear feedback loop consisting of resistor R8, capacitor C4, diodes VD1 and VD2, and transistor VT3. This mechanism limits the voltage across the LC tank to approximately 100 mV, thus avoiding saturation effects in the magnetic core.

The circuit supports a wide range of inductance and capacitance values. Stable operation has been verified for L_f ranging from 50 μH to 100 mH, and for C_f between 50 pF and 5 μF. This wide tuning range makes the oscillator highly adaptable to different sensor configurations and field strengths.

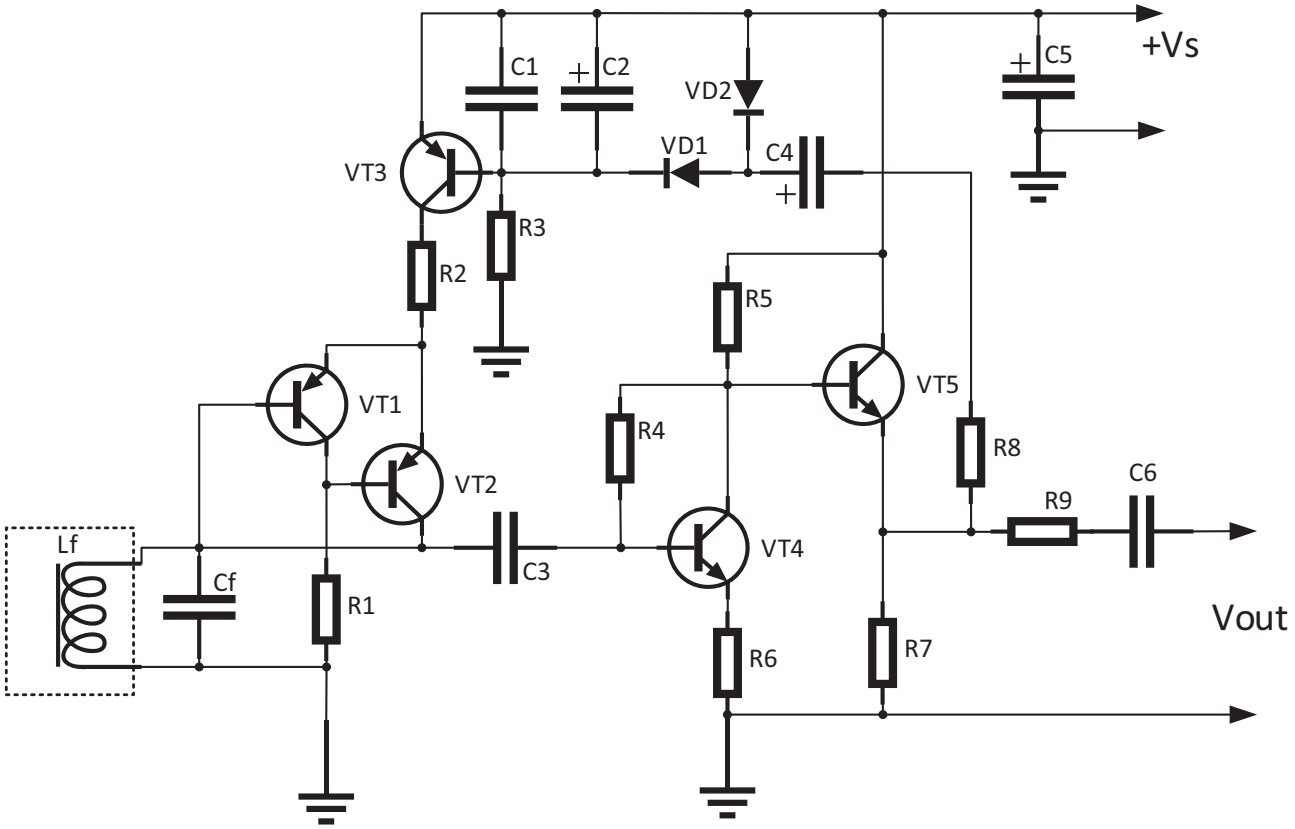


Fig. 4. Schematic diagram of the LC oscillator circuit with integrated parametric magnetic inductance sensor

Power is supplied from a DC voltage source in the range of 6–12 V. The measured current consumption does not exceed 20 mA under normal operating conditions, ensuring suitability for battery-powered and portable systems.

A full list of components is provided in Table II.

TABLE II. BILL OF MATERIALS FOR THE OSCILLATOR CIRCUIT

Component	*Specification*	*Quantity*
R1	10 Ω, SMD-0805	1
R2, R7	1 kΩ, SMD-0805	2
R3	47 kΩ, SMD-0805	1
R4	1 MΩ, SMD-0805	1
R5	3.3 kΩ, SMD-0805	1
R6	330 Ω, SMD-0805	1
R8, R9	270 Ω, SMD-0805	2
C1, C2	10 nF, SMD-0805	2
C3	100 nF, SMD-0805	1
C4, C6	47 µF, 16 V, ECAP-RD	2
C5	10 µF, 16 V, ECAP-RD	1
C7	0.47 µF, SMD-0805	1
VD1, VD2	Diode LL4148	2
VT1–VT2	BC857.215, PNP BJT	2
VT3–VT5	BC847.215, NPN BJT	3

IV. EXPERIMENTAL SETUP

The parametric magnetic inductance sensor was placed inside a cylindrical calibration solenoid, which served as a controlled magnetic field source (Fig. 5). The corresponding block diagram of the setup is shown in Fig. 6.

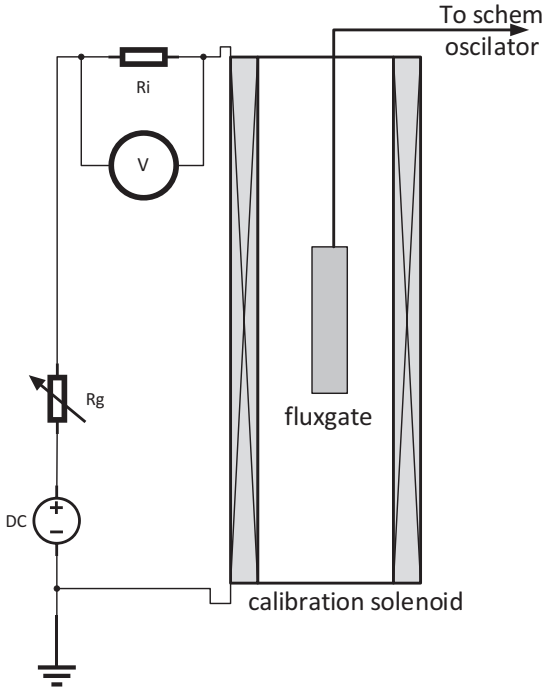


Fig. 5. Schematic of the magnetic field calibration setup with a DC-powered solenoid and embedded parametric magnetic inductance sensor

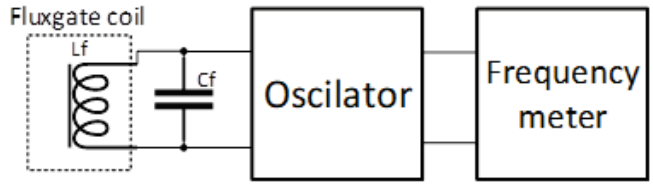


Fig. 6. Block diagram of the experimental setup: parametric magnetic inductance sensor connected to LC oscillator and frequency meter.

The solenoid had the following parameters:

Winding length: 110 mm

Inner diameter: 13 mm

Field constant: $K_H = 3400$ (A/m)/A

The magnetic field strength H was adjusted by controlling the DC current through the solenoid, supplied from a stabilized voltage source. A variable resistor R_g was used to set the current, and a precision voltmeter measured the voltage drop across a calibrated shunt resistor $R_i = 2.2\ \Omega$ (tolerance: ±1%). The resulting magnetic field was calculated from the current using the known solenoid constant.

The output frequency of the LC oscillator was connected to a digital frequency counter with seven-segment indicators displaying the frequency in hertz (Fig. 7).

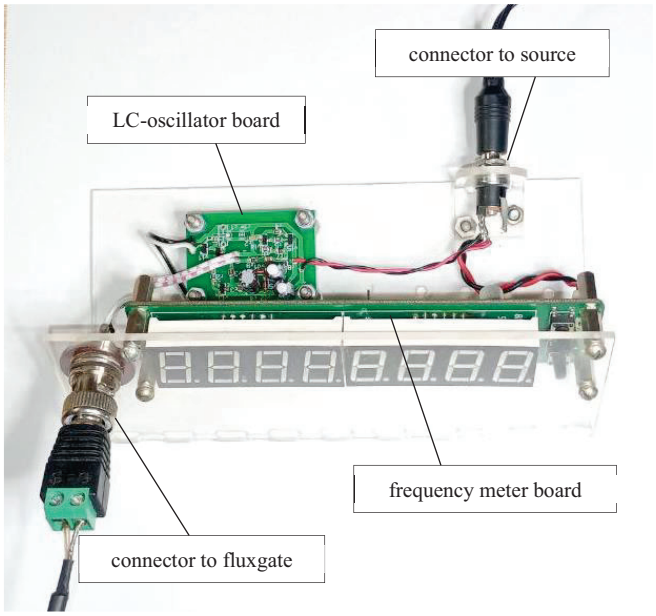


Fig. 7. Photograph of the assembled magnetometer prototype with parametric magnetic inductance sensor and frequency counter

The frequency deviation was recorded for varying magnetic field intensities in the range from 0 to approximately 1700 A/m. This range is representative of magnetization levels encountered in common industrial ferromagnetic materials, as well as significantly higher than the Earth's natural geomagnetic field (20–60 A/m).

To minimize the influence of Earth's field, the solenoid and sensor axis were oriented in the East–West direction, perpendicular to the geomagnetic vector. This orientation reduced the background field offset during calibration. At the same time, preliminary tests demonstrated that the sensor is sensitive enough to detect distortions in the geomagnetic field caused by nearby ferromagnetic objects, further validating its applicability in disturbance detection.

The complete experimental circuit included the sensor coil L_f, capacitor C_f, LC oscillator board, and frequency counter. The oscillator and frequency counter components were integrated on a custom PCB, and the complete experimental layout is presented in Fig. 8.

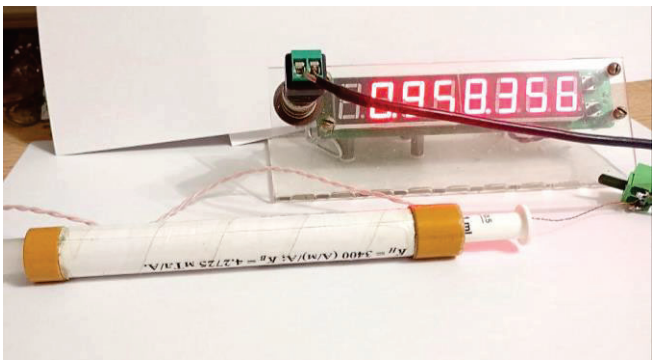


Fig. 8. Experimental setup: sensor inserted into calibration solenoid connected to LC oscillator and frequency counter

V. RESULTS AND DISCUSSION

The frequency response of the LC oscillator circuit with the parametric magnetic inductance sensor was experimentally measured across a range of external magnetic field intensities from 0 to 1700 A/m.

The field was generated using a calibrated solenoid, as described in Section IV, and the output frequency was monitored via a digital frequency counter.

Measurements were performed for four different values of the capacitance C_f: 1000 pF, 2630 pF, 4060 pF, and 10.2 nF.

The results, presented in Fig. 8, show a consistent, monotonic increase in oscillation frequency as the applied magnetic field strength rises. This is explained by the decrease in inductance L_f, caused by increasing magnetic permeability of the core, which in turn shifts the resonant frequency of the oscillator according to the well-known expression:

$$f_0 = \frac{1}{2\pi\sqrt{L_f C_f}} \quad (1)$$

A representative calibration curve is shown in Fig. 9.

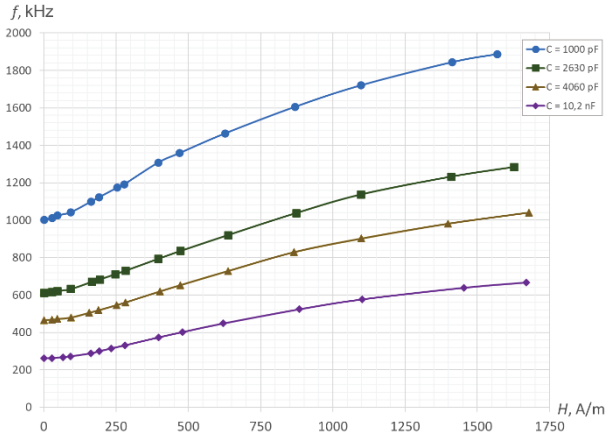


Fig. 9. Experimental setup: sensor inserted into calibration solenoid connected to LC oscillator and frequency counter

The measured curves demonstrate good smoothness, high repeatability, and negligible hysteresis. No abrupt jumps or

saturation effects were observed within the measured field range. The shape of the response is nonlinear, with higher sensitivity in the low-field region (under ~500 A/m) and gradual flattening as the field increases.

To better compare sensor behavior across different oscillator configurations, the data were normalized to the baseline frequency at zero field f_0. The resulting normalized frequency curves f/f_0 are shown in *Fig. 10*.

This approach highlights the relative frequency change independently of the absolute value, and confirms that larger capacitance values yield higher normalized sensitivity. The normalized curves are particularly useful for comparative calibration and digital signal scaling.

The estimated sensitivity in the most linear region (e.g., between 200–800 A/m) reaches up to 10–20% per 100 A/m depending on C_f. Such sensitivity is sufficient for applications involving detection of field gradients, localization of magnetic anomalies, or structural health monitoring using passive magnetic techniques.

These findings confirm that the proposed single-coil sensor architecture, despite its simplicity, provides stable and measurable frequency responses across a wide magnetic field range.

Combined with the frequency-based readout approach and compact hardware design, this sensor is a promising alternative to classical dual-coil fluxgate systems in medium-sensitivity applications.

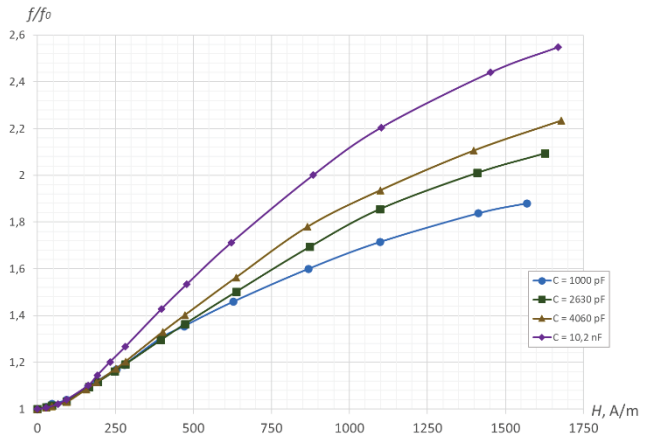


Fig. 10. Normalized frequency response f/f_0 of the LC oscillator as a function of magnetic field intensity for various capacitance values

CONCLUSION

This paper presented the design, implementation, and experimental validation of a magnetometer based on a parametric magnetic inductance sensor integrated into an LC oscillator circuit.

Unlike conventional fluxgate magnetometers, the proposed system employs a simplified single-coil architecture, where the inductance of the sensing coil varies under the influence of an external magnetic field. This variation modulates the oscillator frequency, enabling robust and noise-immune field measurement through frequency tracking.

The sensor was experimentally characterized using a calibrated solenoid for magnetic field generation. The results demonstrate a stable and monotonic frequency response over a broad range of field intensities (0–1700 A/m).

Both absolute and normalized response curves confirmed the suitability of the sensor for medium-resolution magnetic field detection. The use of frequency as the output parameter ensures compatibility with digital systems and improves resilience against amplitude-related noise and drift.

The proposed approach offers several practical advantages, including low power consumption, compact implementation, and reduced component count. These features make it well-suited for integration into embedded systems, portable field instruments, and low-cost industrial sensors.

Future work will focus on improving thermal stability, expanding the frequency range, and integrating digital calibration algorithms to enhance precision in variable environmental conditions.

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Machine Learning for Dynamic Price Adjustments in Digital Marketing

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Abstract*— The peculiarities of artificial intelligence and machine learning algorithms development in modern minds have been observed. The importance of the influx of hardware and software in the evolution of mathematical algorithms that are used for processing and analyzing big data has been brought to light. The importance of globalization and digitalization influx on the business environment in the countries of the world has been established. It has been proven that innovative technologies lead to a transformation in marketing and the need for modernization of current marketing tools. Digital marketing is an effective way to establish close and long-term communication with the target audience. A significant number of current generations (Y, Z, and Alpha) actively use the Internet and buy various goods and services from the digital environment. The relationship between digital marketing and e-commerce, with the method of increasing sales on the Internet, is described. The importance of developing dynamic pricing in modern minds has been brought to the fore, and the main types of economic activity in which automated systems for setting prices are presented. It is indicated that machine learning algorithms will be used to determine the effectiveness of dynamic pricing systems. The Internet serves as a medium for collecting diverse data that is used in the implementation of various machine learning algorithms. The research included detailed information about the price dynamics of the Apple iPhone 16 Pro Max 256GB (premium segment) and the Motorola Moto G04 (economy segment). With the help of a parser, relevant data was collected on a price aggregator, transformed based on feature engineering, and various machine-learning algorithms were implemented. It was found that the CatBoost approach allows one to obtain the best results for price modeling. For each smartphone model, the quality of the resulting CatBoost model was assessed based on quality metrics. It has been established that for the Apple iPhone 16 Pro Max 256GB the results are more accurate, which is explained by the presence of a large array of data. The research also included information about the flagship smartphone model from Apple, which is highly popular among people living in Ukraine. Highly based on the presented model of calls for a large number of proposals on the price aggregator, which made it possible to provide a clear database of prices.

Keywords*— digitization, datasets, dynamic pricing, machine learning, marketing, optimization, smartphones

I. Introduction

In modern conditions, various approaches in the field of artificial intelligence are actively developing, in particular, specialized machine learning algorithms. The emergence of more powerful hardware for fast processing of big data allows the implementation of high-performance mathematical models. The effectiveness of machine learning algorithms has led to their integration into most types of economic activity, which leads to qualitative transformations and ensures sustainable development in the long term. Training models on real data and constant accumulation of experience in various fields allows for the optimization of the approaches used, ensuring their flexibility and efficiency following the action of a complex number of factors in the changing external environment [1].

The rapid growth of Internet users at the national and global levels stimulates companies to actively implement comprehensive marketing strategies to increase the efficiency of e-commerce. In the context of digitalization, there is also a need to integrate innovative marketing tools that allow for effective and long-term communication with the target audience. Thanks to the introduction of machine learning algorithms, companies have the opportunity to optimize the use of digital marketing tools. Dynamic pricing plays a key role in digital marketing, as it adapts price offers to a certain extent to the needs of users [2].

Dynamic pricing has become widespread in such areas as airlines (Emirates, JetBlue, Southwest), e-commerce (Amazon, AliExpress), hospitality (Airbnb, Hyatt, Vrbo), and ride-sharing (Uber, Lyft) [3, 4]. Ukraine is also gradually introducing this approach in the field of adaptive pricing, but the level of integration is significantly inferior to international practice. Dynamic pricing in the country is mainly used in the field of ordering taxis, tickets, and certain types of goods on the Internet. For most products, dynamic pricing in Ukraine is considered a promising approach that will allow for the implementation of hyper-personalized communications and will contribute to increasing profits. The use of various sources of information on the Internet and the accumulation of big data allows for the constant testing of various machine

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learning models to optimize the implementation of digital marketing strategies.

II. Related Works

Dynamic pricing allows companies to achieve an optimal level of profit by establishing effective marketing communications with the target audience. The prospects of the presented approach have led to the emergence of a large number of scientific works in Ukraine and other countries of the world. The intensive development of machine learning algorithms, including artificial intelligence, allows us to conduct comprehensive research in offline and online environments in order to identify optimal approaches to dynamic pricing, which will lead to a high level of consumer loyalty.

The paper [5] examines the features of using various approaches to forecasting demand to improve the implementation of dynamic pricing. The authors suggest using trend equations, regression models, and machine learning algorithms in accordance with the characteristics of the available data. By integrating various approaches to big data processing into the marketing strategy, the company gets the opportunity to adapt prices to user needs in real time, ensuring maximum profit and achieving loyalty of the target audience in the long term.

The features of forming effective pricing strategies and implementing dynamic pricing approaches in the context of digitalization are presented in the paper [6]. The features of data collection for further price modeling are described, including the pre-sales preparation of new goods and services, as important information. It is proven that, according to the existing demand and the characteristics of consumer perception, it is possible to achieve optimal product sales based on dynamic pricing.

The team of authors [7] proves the feasibility of using dynamic pricing in e-commerce. Transactions on the Internet are used to form a large data set, which allows the implementation of various machine learning algorithms. According to the study, the best results of dynamic pricing were achieved for the GBM model, the parameters of which were adjusted through hyperparameter optimization. The need to use personalized pricing strategies based on high-performance machine learning algorithms in e-commerce is emphasized.

The work [8] is devoted to studying the features of combining sentiment analysis and dynamic pricing to optimize product pricing on the Internet. The authors prove the effectiveness of using this approach throughout the entire life cycle of the relevant goods and services. Analysis of consumer sentiment in the digital environment allows businesses to form a system of metrics that significantly improve the quality of the dynamic pricing model based on machine learning algorithms. The use of natural language processing (NLP) approaches makes it possible to implement flexible dynamic pricing models that respond quickly to user sentiment and offer optimal prices in specific spatiotemporal conditions.

The features of using real-time analytics based on artificial intelligence for dynamic pricing are presented in [9]. The author proves that using artificial intelligence to process big data from the Internet significantly increases the conversion rate and increases the effectiveness of companies' pricing strategies. Further improvement of artificial intelligence algorithms is expected, since modern users show significant sensitivity to price, monitoring price offers for a specific product on the Internet.

III. Purpose

Implementing an effective pricing strategy in the digital environment as a key element of a company's marketing activities is very important in modern conditions. The price factor significantly affects the formation of demand among users, since a significant number of the audience pays attention to the cost of goods and services. On the Internet, companies actively interact with representatives of generations Y and Z, who are actively looking for attractive price offers. Accordingly, there is a need for companies to integrate dynamic pricing systems that adapt prices to the characteristics of certain consumer groups.

The Internet is an important source for collecting big data, which allows companies to assess pricing trends and determine cause-and-effect relationships. Companies have the opportunity to accumulate diverse data on relevant markets, competitors, and user behavior. Web analytics services, primarily Google Analytics, are also an important source of information, which allows businesses to accumulate data on companies' web resources. To build high-performance and adaptive dynamic pricing models, it is advisable to use machine learning algorithms. Based on the optimal mathematical model, a dynamic pricing approach is implemented, which provides personalized price offers for users.

IV. Results

The digitalization processes and the growth of innovative technologies and the Internet active users have led to the active development of e-commerce. Among generations Y, Z, and Alpha, purchasing products online is gaining considerable popularity, which contributes to the formation of a new system of interaction between companies and users. Modern users are active consumers of goods and services, preferring innovative offers. Young generations are characterized by increased demand for various gadgets, primarily for new smartphone models. Following their lifestyle, income level, and other factors, representatives of generations Y, Z, and Alpha choose appropriate smartphone models, which can be divided into three groups: premium segment, mid-price group, and economy segment [10].

In this study, a comparative analysis of price dynamics was conducted with the subsequent construction of dynamic pricing models using the example of Apple iPhone 16 Pro Max 256GB (premium segment) and Motorola Moto G04 64GB (economy segment). The Magazilla.ua price aggregator was chosen as the source of information for collecting prices for the presented smartphone models. On this online resource, through the parameter selection system, consumers can find price offers from online stores for specific smartphone models [11]. Using a parser in the Python programming language allows us to generate a data array daily containing information about the price of the

selected model and certain information about sellers (online stores) [12, 13].

During the research, data were collected on the dynamics of prices of two smartphone models from March 31 to June 14, 2025, in the Kyiv region. Figure 1 shows the average prices of the Apple iPhone 16 Pro Max 256GB and Motorola Moto G04 64GB during the studied period. Analysis of price dynamics allows us to determine the presence of fluctuations during the week, which is associated with the peculiarities of the formation of demand and supply for smartphones on the Internet. The data set from March 31 to June 14, 2025, for the Apple iPhone 16 Pro Max 25 6GB amounted to 25,450 observations, and for the Motorola Moto G04 64GB – 5,864 observations. To achieve the goals of dynamic pricing, it was decided to test different machine learning algorithms and select the optimal ones that will allow companies to optimize marketing strategies in the digital environment [14, 15].

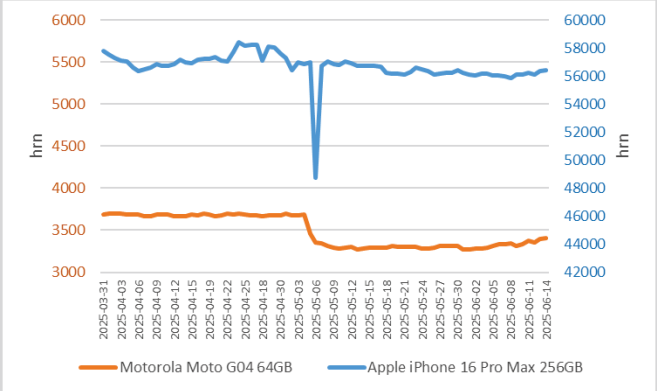


Fig. 1. Dynamics of the average price of Apple iPhone 16 Pro Max 256GB and Motorola Moto G04 64GB in the Kyiv region, from March 31 to April 27, 2025 [calculated by the authors based on 16]

The process of preliminary data preparation for the implementation of various algorithms involves a set of actions that will significantly improve the simulated results. Based on the specifics of the obtained data about smartphones, it was decided to use additional information from the online web analytics service SimilarWeb and generate additional variables through feature engineering. To implement machine learning algorithms, data sets with the following metrics were used:

- The price of a smartphone in a specific online store.

- The rating of the store's website on the Internet among Ukrainian web resources.

- Day of the week (weekday or weekend).

- Lag variables:

1. price_lag_1 – lag variable with a 1-day lag;

2. price_lag_7 – lag variable with a 7-day lag.

Various models were tested, and moving averages (price_ma_7, price_ma_14) were also added. These averages represent the agreed-upon data by trends and complement the lag variables [17, 18].

A comprehensive analysis of data sets with prices for two smartphone models from March 31 to June 14, 2025, and the implementation of various machine learning algorithms revealed that the best results were achieved when using the CatBoost algorithm. This algorithm applies gradient boosting using symmetric decision trees and has the following advantages:

- high resistance to overfitting;

- efficient model creation for both small and large data sets;

- automatic bias correction during modeling [19].

For each smartphone model, the quality of the resulting CatBoost model was assessed (Table 1).

Table 1

Quality assessment of the CatBoost algorithm for two smartphone models

Smartphone model	MAE	RMSE	R^2
Apple iPhone 16 Pro Max 256GB	118,650	358,875	0,996
Motorola Moto G04 64GB	114,545	191,804	0,639

Source: [calculated by the authors based on 16]

For the Apple iPhone 16 Pro Max 256GB, the CatBoost algorithm shows high performance, which indicates a significant correspondence between the actual and predicted prices for this smartphone model (Fig. 2). This indicates an extremely high quality of prediction for this smartphone model. The high RMSE value is associated with a larger offer in online stores of the new flagship model from Apple compared to the second smartphone [20]. Despite the war in Ukraine and the decline in the standard of living of the population, a large number of representatives of generations Y and Z are interested in purchasing the iPhone 16 Pro Max 256GB. The presented model is also of interest to the Alpha generation, which mainly receives this smartphone as a gift from parents and relatives.

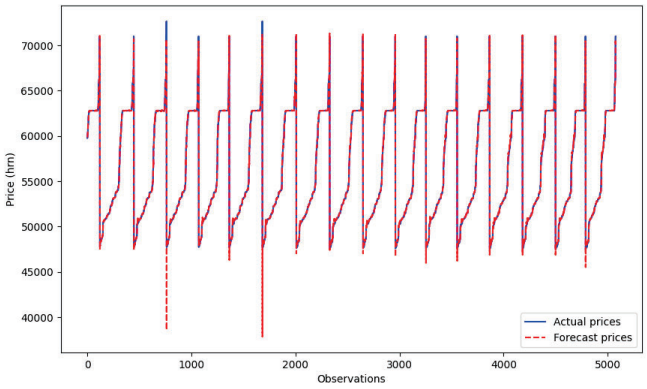


Fig. 2. Actual and forecasted prices for Apple iPhone 16 Pro Max 256GB in the Kyiv region, from March 31 to June 14, 2025 [calculated by the authors based on 16]

For Motorola Moto G04 64GB, the CatBoost algorithm shows lower accuracy; therefore, additional settings are possible to increase the reliability of predictive data. Figure 3 presents actual and forecasted prices for Motorola Moto G04

64GB in the Kyiv region, from March 31 to June 14, 2025. The visualization confirms the conclusions regarding the possibility of further improving the dynamic model for this smartphone based on various machine learning algorithms.

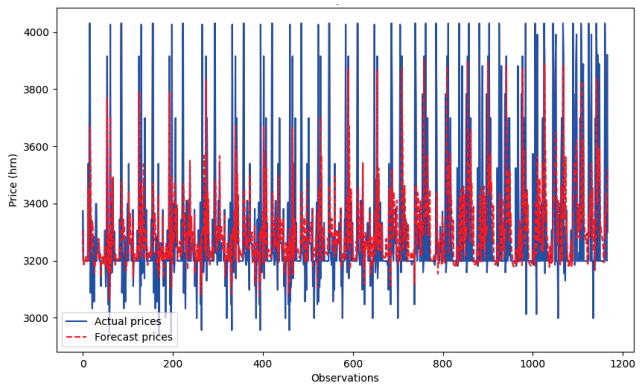


Fig. 3. Actual and forecasted prices for Motorola Moto G04 64GB in the Kyiv region, from March 31 to June 14, 2025 [calculated by the author based on 16]

It should be noted that the Motorola Moto G04 64GB belongs to the economy segment, which has a limited target audience. The low price of this product reflects rather weak technical characteristics in the modern smartphone market. It is also necessary to note the limited possibilities of receiving new updates from Android. The above circumstances are confirmed by the reduction in the offer of this smartphone: March 31, 2025 – 89 offers from online stores, June 14, 2025 – 37 offers. The negative dynamics of the Motorola Moto G04 64GB offer in the context of a relatively small number of sellers at the beginning of the research period affect the characteristics of the data set and require taking into account the specifics in the process of setting the appropriate parameters of machine learning algorithms.

V. Conclusions

The Internet acts as a valuable source of information for improving marketing strategies. Price aggregators and marketplaces should be used to parse relevant prices for goods and services offered by competitors. By collecting data on prices for various products, including smartphones, regularly, it is possible to accumulate big data. In the work, data on prices of two smartphone models in the premium segment and the economy segment were collected to conduct a comparative analysis. The implemented machine learning algorithms showed different levels of quality; as a result, CatBoost was chosen for dynamic pricing. The presented approach turned out to be more effective for the data set with prices of the Apple iPhone 16 Pro Max 256GB, which is explained by the significant popularity of this smartphone model. In the future, it is advisable to accumulate data on price dynamics for certain smartphone models to improve dynamic pricing models based on optimal machine learning algorithms.

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The Statistical Model of an Autonomous Induction Generator Taking into Account Three Excitation Factors

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***Abstract* — The implementation of an autonomous induction (asynchronous) generator as a source of electric energy in power installations is limited by a number of factors. These include significant reactive power required for excitation and loss of excitation in case of overloads. In addition, the complexity of the mathematical description of an autonomous induction generator arises. So far, there is no universal excitation theory that allows, knowing the initial parameters of the system, to calculate values that correspond to the experimentally obtained values of current and voltage across the stator winding of the generator. The mathematical model of an induction generator requires taking into account the main excitation factors. Two main factors are well known — the capacitance of the excitation capacitors and the rotation speed of the generator shaft. The complexity of creating a correct mathematical model is associated with the nonlinear nature of the magnetic circuit inductance of such a generator. As numerous laboratory studies have shown the inductance of an induction generator is not only nonlinear but also non-stationary, i.e. it depends on the acceleration of the shaft. Therefore, this factor should be taken into account when creating a mathematical model. Previously, the possibility of exciting an autonomous induction generator using one capacitor, as well as its stable operation on a resistive load, was shown. This is exactly the generator considered in this paper. The probabilistic nature of current and voltage values when exciting an autonomous induction generator has been shown. Therefore, based on the presented experimental data, the necessity of creating a mathematical model using the methods of mathematical statistics has been substantiated. A correlation-regression model of an autonomous induction generator has been obtained, taking into account the capacity of the excitation capacitor, the rotation speed of the shaft and its acceleration.**

Keywords — induction generator excitation with a single capacitor, volt-ampere characteristic of induction generator, induction generator experimental research, correlation-regression model of induction generator excitation.

I. INTRODUCTION

Using of induction (asynchronous) generators (IG) in the autonomous power supply systems, especially in wind farms, is significantly limited compared to synchronous generators as shown in [1]. Particularly for production of low-power single-phase electrical energy, self-excited induction generators driven by a micro-hydro or micro-wind turbine may be applied [2, 3]. The relatively low cost of induction generators and their reliability are offset by a number of significant drawbacks. It is the need for rather expensive and large-size source of reactive power. The autonomous IG requires a capacitor bank connected to the stator windings. As shown in our previous publications, stable excitation of the three-phase IG using one capacitor is quite possible [4], and stable operation of such a generator on a resistive load is also possible [5,6].

In general, it can be stated that interest in the study of autonomous induction generators is steadily growing, as evidenced by numerous publications in different countries of the world on this topic [7- 15] Therefore, this topic can be considered very relevant in the scientific community.

Different mathematical models are known [1], where the main hypotheses of IG excitation are collected and systematized, which are used in modeling of the IG. However, these models are not entirely adequate to the real process of excitation of the IG, as numerous laboratory experiments convinced us. Therefore, a model of the IG is needed to be correspond to the physical phenomena during excitation of the IG. The main problem in modeling of an autonomous IG is that its excitation process is largely random, which is mainly due to the nonlinear and non-stationary nature of its magnetic circuit.

Using a single capacitor instead of three allows to reduce the excitation costs, as well as the consumed reactive energy. However, this causes asymmetry of voltages on the stator phases of the induction generator. Although between the line voltage across the phases to which the excitation capacitor is connected and the line voltage supplying the load, the asymmetry is less than 10% [6]. For this IG, the useful power on a resistive load of about 400 V was obtained.

We believe that for low-power autonomous generators operating on a resistive load, the use of a single capacitor can be reasonable and economically justified. Especially if we are talking about converting "free" wind and hydro energy into electricity.

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When the IG is excited, the load is usually disconnected, therefore its influence on the excitation process is not considered in this work.

II. EQUIPMENT DESCRIPTION AND PROBLEM STATEMENT

For our research we took the same generator that was used in [4 - 6], as well as the same DC motor as a drive motor (prime mover).

The laboratory bench is shown in Fig.1, where ***TR*** is the autotransformer, ***A*** is the DC motor armature, ***T*** is the tachometer, ***M*** is the muff (coupling), ***S*IG** is the IG shaft, ***S*DC** is the DC motor shaft, ***Rotor*** is the IG rotor, ***EW*** is the DC motor excitation winding, ***VA*** is the diode voltage rectifier, *C1-C4, C2-C5, C3-C6* are the IG stator phases terminals. *A*1, *A*2, *A*3 are the ammeters, *V*1, *V*2 are the voltmeters, R_l is the load resistance, *C* is a variable capacitor. Positions of the capacitor with the corresponding capacitance values are shown in Table 1.

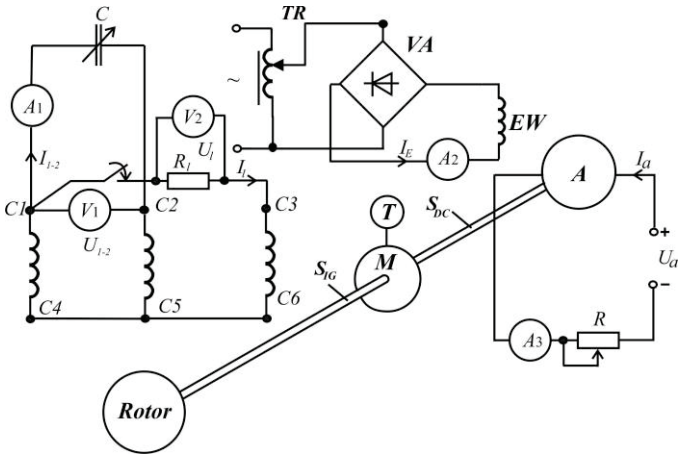


Fig. 1. The schematic diagram of the laboratory bench with an autonomous three-phase IG and a single variable capacitor for excitation

Research were carried out using the 4АХ80ДУ3 three-phase induction machine with a cage (squirrel cage) rotor. Its parameters are the rated power P = 0,75 kW, rated line stator voltage U_1 = 380 V, rated stator current I_1 = 2,2 A, rated frequency of voltage f = 50 Hz, rated synchronous speed n_1 = 1000 rpm, rated rotor speed n_2 = 920 rpm, overload torque factor λ_M = 2,2, overload current factor λ_I = 4,0. The IG shaft is turned by a separately excited DC motor. The DC motor parameters are the rated power P = 0,3 kW, rated armature DC voltage U = 165 V, rated armature current I_a = 3 A, no-load speed n_1 = 1000 rpm, rated armature speed n = 800 rpm, rated excitation current I_E = 170 mA, rated excitation voltage U_E = 190 V.

We can change the DC motor rotation speed and at once the IG speed. And also we can change capacitance of the variable capacitor which is shown in Fig1. The DC motor can be started directly and by a rheostat R in the armature circuit.

TABLE I. POSITIONS OF THE CAPACITOR AND THE VALUES OF CAPACITANCE

Position	1	2	3	4	5	6	7
Value, μF	10,5	18,7	22,9	24,6	32,8	43,4	62

The classical hypothesis [1] explaining the process of self-excitation of a generator with a capacitor, that is, a certain fluctuation causing a swing in the system, goes to the intersection of the volt-ampere characteristic of the capacitor with the magnetization characteristic. That is, it is assumed that as a result of using capacitors with different capacities, the volt-ampere characteristic, which intersects the magnetization characteristic, we get an excitation dots.

And the combination of these dots should clearly lie on the magnetization curve, no matter how we change the capacitance and even the generator load. But numerous experiments have shown that the resulting combination of excitation dots does not give a smooth curve but a certain complex of excitation dots is obtained (Fig2).

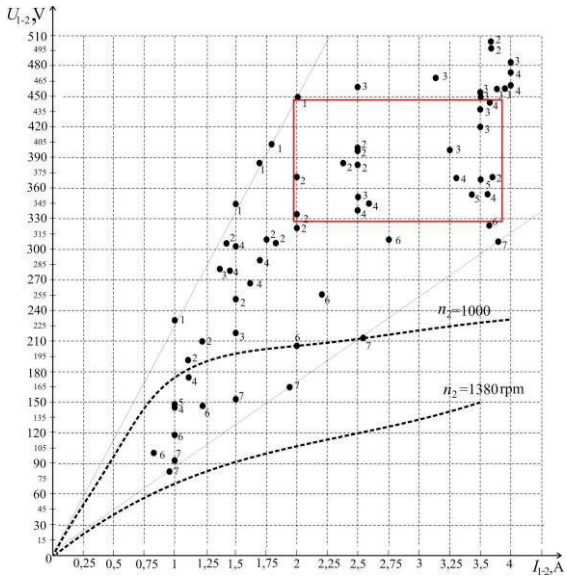


Fig. 2. Dots corresponding steady-state values of the voltage and current at IG excitation in the $U_{1\text{-}2}$ and $I_{1\text{-}2}$ axis

The curves drawn with a dashed line correspond to the magnetization curves obtained at the no-load speed of 1000 rpm and a speed of 1380 rpm at which the IG is often excited. The values of these characteristics are shown in [2]. Figures from 1 to 7 are here corresponding to the positions of the single variable capacitor.

The problem of excitation of an autonomous induction generator is associated with the nonlinear dependence of inductance [1, 3, 7] on the shaft rotation speed, therefore, the impedance of the generator is also nonlinear. This is confirmed by our numerous experiments, in addition, the inductance is obviously non-stationary in nature, since with different starting methods (direct or rheostat) we get different values of voltage across the capacitor and current through it for the excited generator.

Then the differential equation of the excitation capacitor-autonomous IG system is an equation of at least the 4th order, which is also nonlinear. Usually [1] they propose the following IG model in the form of an idealized asynchronous two-phase machine. Let a circular magnetic field is in the air gap, higher harmonics are absent.

$$\begin{vmatrix} U_\alpha^s \\ 0 \\ 0 \\ U_\beta^s \end{vmatrix} = \begin{vmatrix} r_\alpha^s + \frac{d}{dt}L_\alpha^s & \frac{d}{dt}M & 0 & 0 \\ \frac{d}{dt}M & r_\alpha^r + \frac{d}{dt}L_\alpha^r & L_\beta^r\omega_p & M\omega_p \\ -M\omega_p & -L_\alpha^r\omega_p & r_\beta^s + \frac{d}{dt}L_\beta^s & \frac{d}{dt}M \\ 0 & 0 & \frac{d}{dt}M & r_\beta^s + \frac{d}{dt}L_\beta^s \end{vmatrix} \text{x} \begin{vmatrix} i_\alpha^s \\ i_\alpha^r \\ i_\beta^r \\ i_\beta^s \end{vmatrix}$$

$M_{\text{EM}} = \frac{m}{2}M\left(i_\beta^s i_\alpha^r - i_\alpha^s i_\beta^r\right)$ is an electromagnetic torque,

where U_α^s, U_β^s are the voltages across the stator windings;

$i_\alpha^s, i_\alpha^r, i_\beta^r, i_\beta^s$ are the currents in the stator and rotor windings along the α and β axes;

$L_\alpha^s, L_\alpha^s, L_\beta^r, L_\beta^s$ are the total inductances of the stator and rotor windings along the α and β axes;

m is a number of phases;

$L_k = M + L_\sigma$ is the total inductance;

M is a mutual inductance between the stator and rotor windings along the α and β axes;

L_σ is a winding leakage inductance.

Voltage equation of an idealized asynchronous machine in steady state

$$\begin{cases} U_\alpha^s = r_\alpha^s I_\alpha^s + j\omega L_\alpha^s I_\alpha^s + j\omega M I_\alpha^r \\ 0 = j\omega M I_\alpha^s + r_\alpha^r I_\alpha^r + j\omega L_\alpha^r I_\alpha^s + \omega_p L_\beta^r I_\beta^s + \omega_p M I_\beta^s \\ 0 = -\omega_p M I_\alpha^s - \omega_p I_\alpha^r L_\alpha^r + r_\beta^r I_\beta^r + j\omega L_\beta^r I_\beta^r + j\omega M I_\beta^s \\ U_\beta^s = j\omega M I_\beta^r + r_\beta^s I_\beta^s + j\omega L_\beta^s I_\beta^s \end{cases}$$

Moreover, the generator inductance is also non-stationary, since it depends on the acceleration of the shaft when transmitting torque from the drive motor (prime mover) to the generator shaft. Thus, the differential equation of the system will be a non-linear equation of the sixth order.

Then in the excitation capacitor-IG system we will get six natural oscillation frequencies. Apparently, the approach of the IG shaft speed to one of these frequencies causes the IG excitation. Indeed, the most probable cause of the IG excitation is resonance. Although the influence of such a phenomenon as self-oscillations cannot be ruled out. In this work, we did not set the task of precisely determining the nature of excitation.

There is a difficulty in solving such an equation, which is of a fairly high order, and also nonlinear and nonstationary. A simpler way, if you have laboratory equipment and a set of laboratory data, is to make a statistically significant sample and create a polynomial correlation-regression model.

The main excitation factors are the capacitor capacitance C, the shaft speed n_2, but also the rate of increase in speed. The generator excitation criterion in the experiments was the appearance of a current in the stator winding circuit above 1 A and a voltage across the capacitor plates above 30 volts or, what is the same, a voltage across the terminals of the generator stator windings.

Under the experimental conditions, it is difficult to ensure the same rate of increase in speed in manual mode, for example, when manually moving the starting rheostat of the drive motor. Even with direct starting, the excitation process and even when using the same excitation capacitance occurs at different time values.

Earlier using statistical methods it is possible to obtain a polynomial that relates two main input variables: the value of the capacitor capacitance and the shaft rotation speed with the output variable – the value of the voltage across the excitation capacitor.

$$y = 330 + 0{,}12x_1 + 0{,}04x_2\,; \quad (1)$$

$$y = 393{,}13 - 5{,}16x_1 + 0{,}05x_2\,. \quad (2)$$

Where are:

y is the voltage $U_{1\text{-}2}$ across the excitation capacitor;

x_1 is the capacitance C of the variable capacitor;

x_2 is the speed n_2 of the shaft rotation;

The first equation is obtained for a small sample of excitation points, which are limited in Fig. 2 by the red rectangle, and the second equation is for the full sample of values.

A conclusion is made about the need for further research to take into account shaft acceleration as a variable in the equations of the correlation-regression model of the IG.

Let's set the following problem:

1) according to the obtained reliable laboratory data we need to obtain correlation coefficients based on methods known in mathematical statistics;

2) to obtain a correlation-regression model of IG taking into account the third input variable, namely the rate of increase in the speed of the drive motor or, what is the same, the acceleration of the generator shaft.

III. Results of Laboratory Research

Table II shows a sample of experimental data. We can see that even for the same value of the excitation capacitor capacitance, different shaft speeds are obtained at which the excitation of the IG begins and different values of the voltage across the capacitor. The sample is approximately 60% of the points shown in Fig. 2. First, we select statistically significant combinations, second, closer to the definition area - the red rectangle. As our studies of equations (1) and (2) have shown

for statistical analysis, the 60% sample, as used in obtaining (1), is closest to the experimental data than in case (2).

TABLE II SAMPLE OF EXPERIMENTAL DATA

	Input factors			output factor
№	x_1, capacitance C, μF	x_2, speed n_2, rpm	x_2, shaft spin-up time t, s	y, voltage $U_{1\text{-}2}$, V
1	10,5	1536	1,6	444
2	18,7	1390	1,55	452
3	32,8	1270	1,4	352
4	10,5	1448	1,5	358
5	24,6	1315	1,7	471
6	22,9	1343	1,6	380
7	22,9	1322	1,5	475
8	18,7	1534	1,3	480
9	18,7	1389	1,68	370
10	18,7	1410	8	370
11	32,8	1175	7	197
12	22,9	1100	6,7	243
13	22,9	1377	1,6	298
14	18,7	1318	1,56	306
15	10,5	1580	1,8	445
16	10,5	1700	1,9	560
17	43,4	862	1,4	260
18	32,8	934	1,4	296
19	32,8	1235	7,5	197
20	24,6	1300	1,5	329
21	24,6	1350	1,7	468
22	24,6	1290	1,5	460
23	18,7	1363	1,6	496
24	18,7	1029	6	259
25	18,7	1489	7,6	464
26	10,5	1750	1,8	565
27	18,7	1431	1,5	377
28	18,7	1299	1,4	526
29	32,8	1150	1,5	280
30	62	990	5	90

The variable x_2 characterizes the rate of increase of the shaft speed. Its values up to 2 seconds are the so-called a direct start of the prime mover (DC motor), which is carried out without a starting rheostat. The time ise recorded from the moment of start to the moment of excitation of the IG. Values from 5 to 8 seconds are a rheostat start due to the starting rheostat in the armature winding circuit of the DC motor. The starting rheostat is brought out quite smoothly until the moment of the IG entering the excitation mode. And the time is also recorded in this case.

It should be noted that with the minimum value of the capacitor capacitance at a rheostat start, excitation of the IG occurs rarely. But with the maximum capacitance, on the contrary, it was almost impossible to achieve excitation with a direct start. With the capacitor positions from 2 to 5 (Table I), excitation can be achieved almost equally with both direct and rheostat start of the drive motor. However, in general, a direct start is preferable in most cases, if only because the derivative of the increasing magnetic flux of the IG is significantly greater than with a rheostat start. And then the state of the initial magnetization of the magnetic system of the IG is probably unimportant.

Though the initial magnetization of the magnetic system of the IG can also be considered at a rheostat starting as a variable on which the excitation of the IG depends, but it is not considered in this article.

As stated in [16] to successfully achieve the goal of the study, it is necessary that optimization parameter did evaluate efficiency functioning of the system in a predetermined sense. This the requirement is the main one that determines the correctness of the research problem setting.

General task of researching is to find one output value y as a function of several factors

$$y = \varphi (x_1 , x_2 , \dots , x_k) \tag{3}$$

To investigate the mechanism of a phenomenon means to determine the analytical expression

$$y = f (x_1 , x_2 , \dots , x_k), \tag{4}$$

which would accurately describe the unknown dependence within the range of possible values of the factors x_i, called the domain of definition of factors, or factor space. It needs to evaluate boundaries of the domain of factors definition.

First, we construct a matrix of paired correlation coefficients, including all variables, i.e. both input parameters and output parameters.

The pair correlation coefficient (Pearson's) is a number from -1 to 1, characterizing the closeness of the linear correlation between the dependent random variable and the independent random variable.

$$r_{xy} = \frac{\sum_{i=1}^{n}\left(x_i - \bar{x}\right)\left(y_i - \bar{y}\right)}{\sqrt{\sum_{i=1}^{n}\left(x_i - \bar{x}\right)^2 \cdot \sum_{i=1}^{n}\left(y_i - \bar{y}\right)^2}}, \tag{5}$$

where:
n is the number of observations;
x_i is the i-th observed value of an independent random variable;
$\bar{x}$ is the average value of x_i;
$\bar{y}$ is the average value of y_i;
y_i is the $i^{\text{-th}}$ observed value of the dependent random variable;
$\bar{y}$ is the average value of y_i;
r_{xy} is the pair correlation coefficient.

Correlation coefficients were calculated by standard Excel statistical procedure and given in Table III.

Analyzing the correlation coefficients, we can say that the output variable, i.e. the voltage across the capacitor, is influenced to the greatest extent by the shaft rotation speed, to a slightly lesser extent by the capacitor capacitance value, and to an even lesser extent by the shaft acceleration. Moreover, the minus sign before the coefficients means that the variables have a tendency to have different directions, i.e. an increase in one of the variables means a decrease in the other.

TABLE III CORRELATION COEFFICIENTS

		y	Input factors		
			x_1	x_2	x_3
y		1	-0,785856	0,2139902	-0,731232
Input factors	x_1	-0,785856	1	-0,22693	0,766246
	x_3	0,2139902	-0,22693	1	-0,46151
	x_3	-0,731232	0,766246	-0,46151	1

The regression statistics are shown in Table 4. Since the correlation-regression model is based on the least squares method, these statistics reflect the computational procedures of this method. The most important indicator is the average approximation error. It is 14.8%. Quite a large indicator, but if we take into account that we did not specify the domain of definition, did not discard statistically insignificant values, but simply used a statistical sample obtained over several hours of successive experiments, then the result is quite acceptable.

TABLE IV REGRESSION STATISTICS

Multiple R (determination coefficient)	0,843238623
R-squared	0,711051376
Normalized R-squared	0,678945973
Standard Error	67,32007963
Average approximation error	14,81245

Table 5 is shown below with the predicted values of the output variable, i.e. the voltage values across the capacitor. These values and the values obtained as a result of the experiment are shown as graphs in Fig. 3.

TABLE V PREDICTED VALUES OF THE OUTPUT VARIABLE

№ combination of parameters	Values of voltage across capacitor	№ combination of parameters	Values of voltage across capacitor
1	428,2203512	16	211,1207927
2	351,8707894	17	265,5477864
3	471,6732073	18	250,4553903
4	386,6807843	19	385,8573549
5	401,1517373	20	395,6727638
6	397,2716876	21	383,2882179
7	469,003769	22	420,5261123
8	425,9937581	23	268,0508636
9	335,6322225	24	361,9889568
10	242,6162585	25	544,7157304
11	261,449669	26	520,0256711
12	409,8868031	27	439,5113819
13	409,5710511	28	407,1139116
14	501,0404015	29	319,5260074
15	530,3549074	30	126,4151868

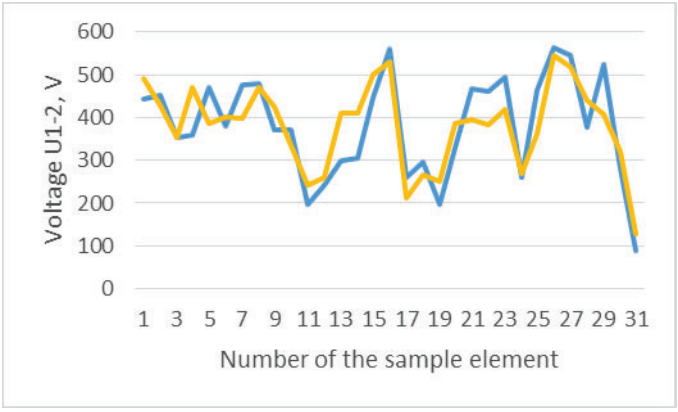


Fig.3 Values of the voltage $U_{1\text{-}2}$: predicted – yellow line, the sample – blue line

Let us now perform a regression analysis of the data and obtain a linear regression equation according to (5) and table 5 using standard Excel statistical procedure

$$y = 157{,}98 - 3{,}39x_1 + 0{,}026\,x_2 - 15{,}15\,x_3 \qquad (6)$$

The coefficients of variables x_2, x_3 have statistical significance of about 1 % with a confidence interval (reliability level of 95%), which is a good indicator. These are statistically significant variables. The coefficient of x_1 has a significance of 7,5 %, which is slightly higher than 5 % reliability. But given the rather rough nature of the sample, it can be considered satisfactory.

The disadvantage of this approach is that this equation obviously cannot be extended to a wide class of induction generators.

CONCLUSIONS

A linear correlation-regression- model of the IG excitation with a single excitation capacitor has been obtained in the form of expression with three variables. These variables are significant factors of the IG excitation.

Using of the expression (6) let us more precisely estimate process of IG excitation than (1) because of due to the additional factor of excitation.

Before starting statistical analysis, it is necessary to carefully determine the domain of definition of the output variable and the corresponding ranges of changes in the input variables.

Laboratory studies have shown that with direct starting of the drive motor, excitation of the generator is generally achieved more easily than with a relatively slow rheostat start.

The obtained model can be used for optimization of the IG excitation process.

Probably, the IG excitation analysis is obviously required that takes into account additional variable, not only the value of the excitation capacitance and the speed of the IG as well as the acceleration of the shaft, but also the initial magnetization. In further research and obtaining more accurate statistical models,

one should turn to the method of planning an experiment and nonlinear models.

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Building an SVM Classifier of Human Psychophysiological State on a Cloud Computing Platform

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***Abstract*—This paper presents an integrated methodology for modeling the human eye movement system (EMS) using nonlinear Volterra polynomial models and for constructing an effective psychophysiological state classifier based on experimental eye-tracking data. The proposed approach addresses the challenges of identifying the inertial and nonlinear dynamic properties of EMS in a nonparametric form using responses of "input–output" experiments. The approximation is performed by means of second-order Volterra integral models, and the corresponding transient characteristics are derived. These characteristics are used to construct two types of feature spaces: one based on heuristic parameters and another formed using wavelet decomposition coefficients. A support vector machine (SVM) classifier is implemented to differentiate between states corresponding to different times of day, reflecting variations in human fatigue levels. The entire computational workflow is deployed on developed cloud computing platform that integrates PaaS and SaaS components, ensuring accessibility, modularity, and scalability of data processing and model execution. The classification accuracy of the proposed system reaches 93.75%, confirming the applicability of integral EMS modeling for intelligent diagnostics in cognitive monitoring scenarios.**

Keywords—psychophysiological state, eye movement system, modeling, eye-tracking, SVM, cloud services, cloud computing

I. INTRODUCTION

The human eye movement system (EMS) is a significant subject of research in neuroscience, medicine, and artificial intelligence technologies. Eye movement analysis provides valuable insights into cognitive processes, psychophysiological states, and neurological disorders. Accordingly, the development of mathematical models that describe the dynamics of the EMS is of particular importance for both fundamental and applied research.

A considerable number of contemporary studies focus on the use of eye-tracking technology for the investigation of cognitive impairments. In [1], eye movement analysis is employed to explore the possibilities of early detection of cognitive decline associated with Alzheimer's disease. In [2], a mobile platform with an integrated artificial intelligence module is presented, which enables the construction of a predictive disease model based on the results of eye-tracking tests. Clinical analysis of oculomotor activity parameters in patients with Parkinson's disease is provided in [3]. The study [4] presents a model that integrates real-time retinal image processing with eye movement analysis to detect autism, enabling the application of eye-tracking for early diagnosis of neurodevelopmental disorders.

One of the promising approaches to EMS modeling is the application of integral models, which make it possible to formalize the "input–output" relationship for nonlinear dynamic systems with unknown structure (so-called "black-box" systems). In [5], Volterra polynomials are used to address the identification problem. Volterra–Laguerre models are applied to smooth eye movement modeling in [6]. The theoretical foundations of identification based on Volterra models are outlined in the monograph [7].

The practical applications of eye-tracking encompass a wide range of directions. In [8], the use of eye-tracking in combination with pose estimation technologies is considered for evaluating team coordination effectiveness in medical simulations. Study [9] focuses on assessing pilot attention distribution during autorotation planning in helicopter emergency descent scenarios; it includes an analysis of eye movement metrics and operator workload, which are crucial for ensuring flight safety. In [10], eye-tracking is combined with virtual reality to examine dynamic visual acuity as a potential biometric indicator for astronaut safety, particularly under gravitational transitions.

Knowledge acquisition is a key aspect of the educational process, and its effective management requires the use of objective criteria for assessing the cognitive capabilities of learners. The proposed methods for evaluating the neurophysiological state of individuals, based on experimental data obtained using innovative eye-tracking technology and nonlinear dynamic identification of the EMS, enable monitoring and diagnostics of cognitive process states during students' educational activities [6], [11].

To integrate these technologies into scientific research and educational processes, it is advisable to use cloud technologies [12], [13]. Currently, tools such as Project Jupyter and Google Colab are employed to support cloud computing. However, these are more interactive notebooks than full-fledged platforms for conducting experiments and working with data processing results. They function primarily as editors and execution environments for a single programming language and do not offer capabilities for developing project code in multiple programming languages or for interacting with pre-implemented GUI interfaces designed for cloud-based neurophysiological research using eye-tracking data.

There is a growing need to develop software tools that support cloud computing under both PaaS and SaaS models simultaneously. Such tools would enable efficient work in both research and educational domains by allowing users to work with source code written in multiple programming

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languages—supporting algorithm development—as well as with implemented identification methods presented through GUI interfaces. Additionally, the inclusion of a social component is essential, as it would facilitate easier data sharing between researchers and across projects, thereby increasing the productivity of scientific research—an aspect that is currently lacking in existing solutions.

II. Problem Statement

The aim of this study is to develop instrumental algorithmic and software tools for constructing a nonparametric dynamic model of the human EMS, taking into account its inertial and nonlinear properties, based on experimental "input–output" data obtained using eye-tracking technology; and to investigate the effectiveness of applying the resulting informational models for assessing the human psychophysiological state.

The subject of this research is computational algorithms and software tools for supporting an intelligent information technology for assessing the psychophysiological state (fatigue state) based on eye-tracking data within a cloud computing platform.

Volterra models are used for EMS modeling, which provide an explicit description of the relationship between the input signal $x(t)$ (stimulus) and the output signal $y(t)$ (response) of a nonlinear dynamic system with zero initial conditions in the form [7]:

$$\begin{aligned} y_N(t) &= \sum_{n=1}^{N} \hat{y}_n(t) = \int_0^t w_1(\tau)x(t-\tau)\mathrm{d}\tau + \\ &+ \int_0^t\int_0^t w_2(\tau_1,\tau_2)x(t-\tau_1)x(t-\tau_2)\mathrm{d}\tau_1\mathrm{d}\tau_2 + \\ &+ \int_0^t\int_0^t\int_0^t w_3(\tau_1,\tau_2,\tau_3)x(t-\tau_1)x(t-\tau_2)x(t-\tau_3)\mathrm{d}\tau_1\mathrm{d}\tau_2 d\tau_3 + \ldots \end{aligned} \quad (1)$$

where

$$\hat{y}_n(t) = \int_0^t \ldots \int_0^t w_n(\tau_1,\ldots,\tau_n)\prod_{i=1}^{n} x(t-\tau_i)d\tau_i, \quad (2)$$

$y_N(t)$ is the output function of the EMS approximation model (N is the approximation order); $\hat{y}_n(t)$ is the n-th partial component of the model output function (an n-th order convolution integral); $w_n(\tau_1,\ldots,\tau_n)$ is the n-th order Volterra kernel (weight function); t is the current time.

In this case, the nonlinear and dynamic properties of the system under study are uniquely described by a sequence of multidimensional weighting functions invariant to the shape of the input signal.

The identification problem (model construction) in the form of the polynomial model (1) consists in determining the Volterra kernels based on experimental "input–output" data of the EMS. The modeling process includes the selection of test signals $x(t)$ and the development of an algorithm that allows determining the partial components from the measured responses $y(t)$, which makes it possible to find the weight functions $w_n(\tau_1,\ldots,\tau_n)$ [14].

Considering the specific features of the EMS under study, multistep test signals are used for identification [11]. If the test signal $x(t)$ is a unit step function $\theta(t)$ (Heaviside function), the result of identification is the partial component $\hat{y}_n(t)$. In particular, for n=1:

$$\hat{y}_1(t) = \hat{h}_1(t), \quad (3)$$

where $\hat{h}_1(t)$ is the estimate of the first-order transient characteristic $h_1(t)$;

For $2 \le n \le N$:

$$\hat{y}_n(t) = \hat{h}_n(t,\ldots,t), \quad (4)$$

where $\hat{h}_n(t,\ldots,t)$ – are the estimates of the diagonal sections of the n-th order transient characteristics $h_n(t_1,\ldots,t_n)$, which are n-dimensional integrals of the kernels $w_n(\tau_1,\ldots,\tau_n)$:

$$h_n(t,\ldots,t) = \int_0^\infty \ldots \int_0^\infty w_n(t-\tau_1,\ldots,t-\tau_n)\mathrm{d}\tau_1 \ldots \mathrm{d}\tau_n. \quad (5)$$

The response of the EMS Volterra polynomial model of degree N to a step input function with amplitude a is calculated by the formula:

$$\tilde{y}(t) = a\hat{y}_1(t) + a^2\hat{y}_2(t) + \ldots + a^N\hat{y}_N(t). \quad (6)$$

The obtained EMS transient characteristics are used to construct a space of effective diagnostic features, in which, using the generated dataset and machine learning tools, optimal synthesis of a psychophysiological state classifier is carried out.

The effectiveness of applying the constructed EMS informational models in the form of transient characteristics is evaluated by the probability of correct recognition (PCR) of the developed psychophysiological state classifier.

III. Architecture and Service Integration

Web-oriented software tools have been developed to automate and optimize research into the human neurophysiological state. These tools combine two approaches to user interaction with software: PaaS (Platform as a Service) and SaaS (Software as a Service). This enables researchers to work both through an integrated interface and directly with the program code using a specially developed code editor.

Thanks to the embedded code editor (PaaS), users can add new data processing tools, edit algorithm scripts, execute computations on the server via integrated script code, and receive results directly in the browser.

In addition, to improve research productivity and efficiency, the system supports the preparation and management of experimental data obtained from an eye tracker, which is then processed to build a mathematical model (identification) of the oculomotor system using specially developed GUI interfaces (SaaS).

The developed tools support working with code in languages such as Python and JavaScript, allowing for the enhancement of algorithms while also utilizing pre-implemented identification methods through specialized GUI interfaces. Moreover, the platform is cloud-based and compatible with all types of devices.

From the perspective of functional interaction, the software product consists of four core components, as illustrated in the functional diagram of the service:

- Server – manages all technical processes;
- Web interface (client) – provides access to the system's tools;
- Agents – computational components responsible for executing the submitted code;
- Database – stores all data related to the platform and its projects.

This allows independent development and scaling of all software components, enabling seamless integration of applications across platforms (Windows, Linux, Android, macOS, iOS).

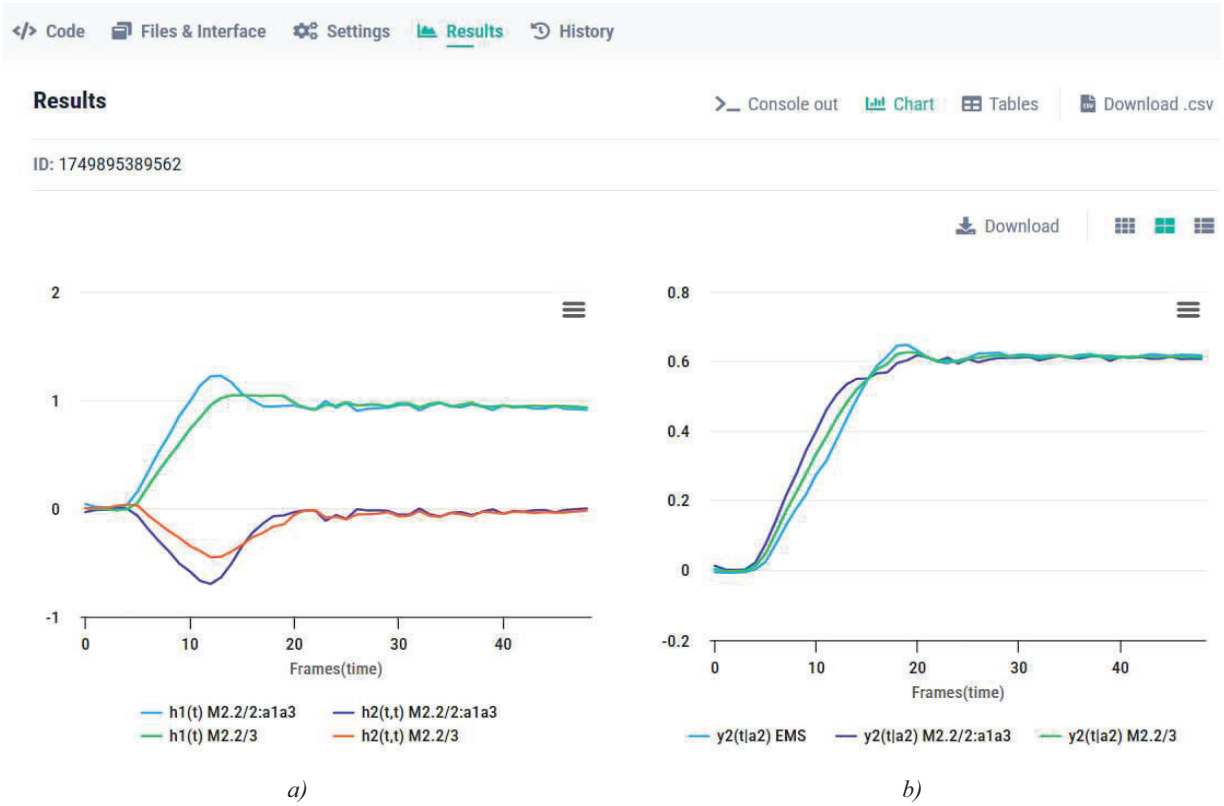
Centralizing the entire logic and data processing on the server side also enables third-party applications to integrate the functionality of the developed tools, providing additional scalability opportunities.

IV. Experimental Studies

Taking into account the physiological features of the EMS, step test signals of different amplitudes are used for identification, which are implemented as a bright point at different distances a_i (i=1,2,…,L; L is number of experiments) from the starting position. The test visual stimuli can be formally considered as functions $x(t)=a_i\theta(t)$. The corresponding EMS responses to the test visual stimuli are recorded using an eye tracker and then used for the identification of the EMS in the form of transient characteristics.

To build the models, sets of EMS responses to test visual stimuli displayed on a computer screen at different distances x_j (j = 1, 2, 3) from the starting position in the "horizontal" direction are used: $x_1=(1/3)X$, $x_2=(2/3)X$, $x_3=X$ (X is screen width in pixels). These test signals are equivalent to input step signals $a_j\theta(t)$ with amplitudes a_j defined by the values of x_j. Empirical data were obtained from respondents (students) using the Tobii Pro TX300 eye tracker at different times: "Morning" (before classes) and "Evening" (after classes), and on different days [11]. A full EMS study cycle for one respondent consists of three experiments with different amplitudes of the test signals a_1, a_2 and a_3.

To implement the EMS nonparametric identification algorithm, the EMS responses to test visual stimuli were preprocessed: normalization, synchronization, and extraction of the rising edge in the output signals of the eye tracker [14]. The study [15] established that when constructing the quadratic model M2.2/3 using the least squares method, the use of three step test signals allows the identification error of the EMS to be halved compared to the M2.2/2 model built using two test signals. The graphs of the first- and second-order transient characteristics of the models M2.2/2:a_1a_3 (a_1 and a_3 are the amplitudes of the test signals used in model construction) and M2.2/3 are shown in Fig. 1.a, and the EMS responses and the corresponding EMS model responses are shown in Fig. 1.b.



a) *b)*

Fig. 1. Research results visualization: *a*)Transient characteristics of the first- and second-order models of the EMS M2.2/2:a_1a_3 and M2.2/3; *b*) EMS responses and responses of the EMS models M2.2/2:a_1a_3 and M2.2/3 with amplitude a_2=2/3

Quadratic integral EMS models are studied: M2.2/2 and M2.2/3, which are constructed based on the least squares method using, respectively, two and three step test signals [16]. The features of using the obtained empirical data for model construction are examined, and estimates of the variability of the obtained averaged transient characteristics are determined depending on the "Morning" and "Evening" states.

The variability (deviation) of the averaged transient characteristics of different orders of EMS models M2.2/2 and M2.2/3 for the respondent states "Morning" $\hat{h}_{nN}^{(\mathrm{M})}(t_m)$ and "Evening" $\hat{h}_{nN}^{(\mathrm{E})}(t_m)$ is determined using the following indicators:

- maximum deviation σ_{nN};
- normalized root-mean-square deviation ε_{nN}:

$$\sigma_{nN} = \max_{m\in[0,M]} \left| \hat{h}_{nN}^{(\mathrm{M})}(t_m) - \hat{h}_{nN}^{(\mathrm{E})}(t_m) \right|; \tag{7}$$

$$\varepsilon_{nN} = \left(\frac{\sum_{m=0}^{M} \left(\hat{h}_{nN}^{(\mathrm{M})}(t_m) - \hat{h}_{nN}^{(\mathrm{E})}(t_m) \right)^2}{\sum_{m=0}^{M} \left(\hat{h}_{nN}^{(\mathrm{M})}(t_m) \right)^2} \right)^{1/2}, \tag{8}$$

where n is the order of the transient characteristic, $n=1,2,\ldots,N$.

The obtained variability indicators of the averaged transient characteristics of the EMS models M2.2/2 and M2.2/3 are shown in Fig. 2.

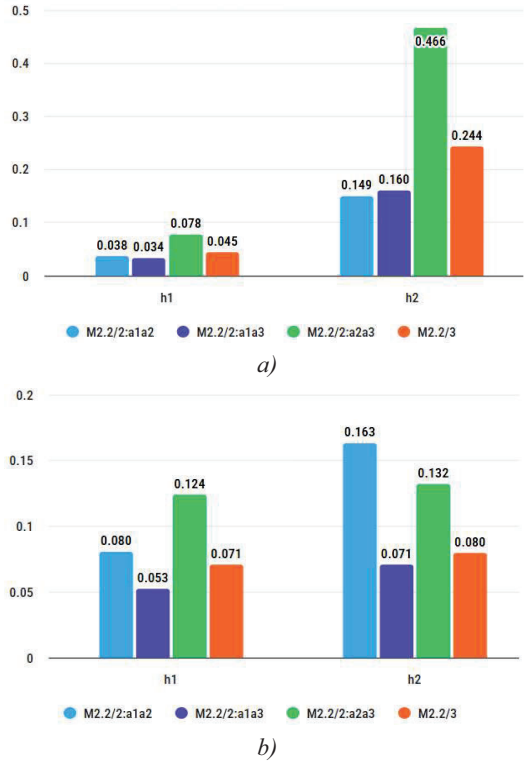


Fig. 2. Variability of the averaged transient characteristics of the EMS models M2.2/2 and M2.2/3 for the "Morning" and "Evening" states: *a*) normalized root-mean-square deviation; *b*) maximum deviation.

V. Reduction of Information Models of the EMS

The study employs machine learning methods for the classification of an individual's psychophysiological state in a selected feature space. The selection of heuristic features is based on studies that determine their informativeness in terms of the PCR and their ability to reflect changes in the respondent's psychophysiological state.

To construct the feature space, parameterization of the obtained EMS transient characteristics based on the M2.2 models (model reduction) is applied using formal expressions for the definition of heuristic features listed in Table I — feature space E_0; and features derived from wavelet decomposition of transient characteristics — feature space W.

The list of heuristic features defined based on the M2.2 model is a subset of the features $e_k \in E_0, k = \overline{1,21}$ studied in [14].

TABLE I. Heuristic Features for EMS Models M2.2

ID	Features	Formal Definition
1	e_1	$\sum_{m=0}^{M} \lvert \hat{h}_1(t_m) \rvert$
2	e_2	$\sum_{m=0}^{M} \lvert \hat{h}_2(t_m,t_m) \rvert$
3	e_4	$\max_{m\in[0,M]} \hat{h}_1'(t_m)$
4	e_5	$\arg\max_{m\in[0,M]} \hat{h}_1'(t_m)$
5	e_6	$\max_{m\in[0,M]} \hat{h}_2'(t_m,t_m)$
6	e_7	$\arg\max_{m\in[0,M]} \hat{h}_2'(t_m,t_m)$
7	e_{10}	$\min_{m\in[0,M]} \hat{h}_1'(t_m)$
8	e_{11}	$\arg\min_{m\in[0,M]} \hat{h}_1'(t_m)$
9	e_{12}	$\min_{m\in[0,M]} \hat{h}_2'(t_m,t_m)$
10	e_{13}	$\arg\min_{m\in[0,M]} \hat{h}_2'(t_m,t_m)$
11	e_{16}	$\max_{m\in[0,M]} \lvert \hat{h}_1(t_m) \rvert$
12	e_{17}	$\arg\max_{m\in[0,M]} \lvert \hat{h}_1(t_m) \rvert$
13	e_{18}	$\max_{m\in[0,M]} \lvert \hat{h}_2(t_m,t_m) \rvert$
14	e_{19}	$\arg\max_{m\in[0,M]} \lvert \hat{h}_2(t_m,t_m) \rvert$

Here, $\hat{h}_1'(t_m)$, $\hat{h}_2'(t_m,t_m)$ denote the derivatives of the first- and second-order transient characteristics, respectively.

To construct the W feature space, discrete wavelet transform (DWT) is used [16]. The software implementation is carried out using the PyWavelets library in the Python environment. Coiflet 4 is used as the base wavelet with a decomposition level of 2.

Feature vectors $w_m \in W, m = \overline{1,10}$ are formed from the approximation coefficients array *ca* and detail coefficients array *cd* as follows: w_1=*ca*[1],…, w_5=*ca*[5], w_6 = *cd*[1],…, w_{10}=*cd* [5].

VI. Training the SVM Classifier

Based on the datasets for the "Morning" and "Evening" classes, classifiers were constructed to assess the individual's psychophysiological state. The classification effectiveness in the constructed feature spaces E_0 and W was analyzed through an exhaustive search over all possible feature pairs and evaluating their informativeness using the criterion of the probability of correct recognition (PCR). Due to the limited size of the training data, the analysis was restricted to all possible pairs of features.

To build the psychophysiological state classifier, the Support Vector Machine method with a Gaussian (RBF) kernel was applied [17]. The software implementation was carried out in the Python environment using the sklearn.svm.SVC class from the Scikit-learn library. The classification performance metrics were evaluated using the functions provided by the sklearn.metrics module.

A. Feature Space E_0

As a result of training the SVM classifier in the E_0 feature space, which was defined based on the EMS models M2.2, the highest PCR value of 87.5% was achieved for the following feature pairs:

- M2.2/2:a_1a_2 :

$$\left(e_{10} = \min_{m\in[0,M]} \hat{h}_1'(t_m)\right) \& \left(e_{12} = \min_{m\in[0,M]} \hat{h}_2'(t_m, t_m)\right); \quad (9)$$

- M2.2/2:a_2a_3 :

$$\left(e_2 = \sum_{m=0}^{M} |\hat{h}_2(t_m, t_m)|\right) \& \left(e_5 = \operatorname{argmax}_{m\in[0,M]} \hat{h}_1'(t_m)\right); \quad (10)$$

- M2.2/3 : e_{10}&e_{12}, formula (9).

For the E_0 feature space formed based on the M2.2/3 model, the distribution of dataset objects in the e_{10}&e_{12} plane is shown in Fig. 3.

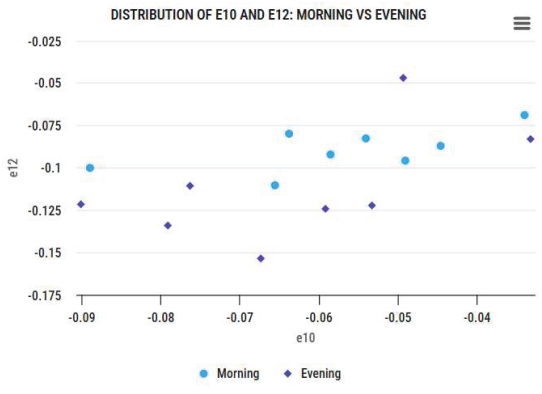


Fig. 3. Distribution of the objects in datasets e_{10}&e_{12} constructed based on the EMS model M2.2/3

B. Feature Space W

SVM classifiers constructed in the W feature space reached a maximum PCR value of 93.75% based on the M2.2/2:a_2a_3 model for the following feature pairs: w_4&w_5; w_4&w_8; w_5&w_7; w_5&w_8; w_5&w_9; w_5&w_{10} and w_6&w_8

For the W feature space formed based on the M2.2/2:a_2a_3, model, the distribution of dataset objects in the w_4&w_5 plane is shown in Fig. 4.

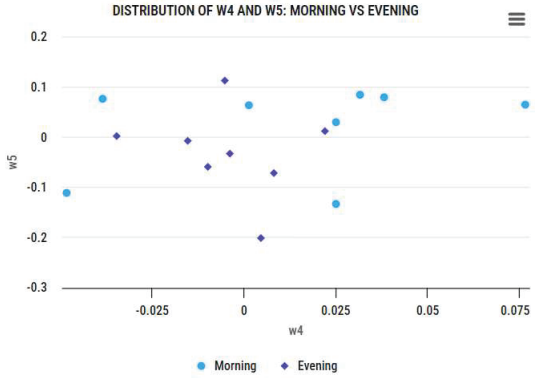


Fig. 4. Distribution of the objects in datasets w_4&w_8 constructed based on the EMS model M2.2/2:a_2a_3

SVM classifiers in the W feature space also achieved a PCR value of 87.5% for the following robust feature pairs based on M2.2 models:

- M2.2/2:a_1a_3 : w_8&w_9, w_8&w_{10};
- M2.2/3 : w_4&w_8, w_6&w_8.

VII. Conclusion

The diagnostic capabilities of quadratic models of the human eye movement system (EMS) are investigated in the form of multidimensional transient characteristics that account for the nonlinear and dynamic properties of the studied system.

Feature spaces were constructed based on transient characteristics obtained from "input–output" experimental data. Two approaches to feature construction were proposed: the extraction of heuristic features and the use of wavelet decomposition coefficients of the transient characteristics, followed by the selection of the most informative feature combinations. Both approaches enable the creation of diagnostic feature spaces suitable for further application in statistical classification systems. To implement the procedures of model identification, feature construction, and classifier training, an innovative cloud computing platform was utilized, significantly enhancing the efficiency of the research workflow.

An important feature of the developed software tools is their minimal requirements for client-side hardware, due to the fact that all computations are performed on the server. The modular architecture ensures flexible scalability, and the combination of PaaS and SaaS principles makes the tools highly versatile. They also offer significant advantages over platforms such as Project Jupyter and Google Colab, including support for popular programming languages like Python and JavaScript; integration of well-tested tools via specialized GUI interfaces; a higher level of abstraction; and an expanded

social component. These features optimize the research process and open up new opportunities for users

Using machine learning methods, the classification of human psychophysiological states was carried out within the constructed diagnostic feature spaces using a support vector machine (SVM) approach. In the heuristic feature space, the highest recognition probability achieved by the SVM classifier was 87.5%. In the feature space formed using wavelet decomposition coefficients, the maximum recognition accuracy reached 93.75%.

Therefore, the conducted study confirms the effectiveness of quadratic integral models for the evaluation of human psychophysiological state.

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Recognition of Combined Faults in Oil-filled Equipment Based on DGA Results, under Conditions of Limited Regulatory Framework

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***Abstract*—A method is proposed for recognizing the fault type in oil-filled electrical network equipment based on the analysis of gases dissolved in oil, in the absence of standardized diagnostic criteria values. It has been established that for combined faults, the diagnostic criteria values correspond simultaneously to faults of different types. It has been shown that for reliable recognition of combined faults, it is necessary to use an exclusively complex diagnostic space, formed by both gas ratios and gas percentage content, as well as the ratios of gases to the gas with the maximum content. A comparative analysis showed a significant advantage of the proposed method over known standards and author's methods.**

Keywords—diagnostics, oil-filled equipment, dissolved gas analysis (DGA), combined faults, complex diagnostic space, gas ratio, gas percentage, fault nomograms, recognition reliability.

I. INTRODUCTION

One of the clear advantages of the dissolved gas analysis (DGA) method, compared to other methods for diagnosing the condition of oil-filled equipment, is the ability to not only detect faults early, but also to identify their type. This advantage makes it possible to assess the degree of danger of the predicted fault, evaluate the rate of its development, and, as a result, make the most optimal decision about the possibility of further operation of the equipment. The norms and criteria regulated in most current standards and author's methods for interpreting DGA results [1-5] allow for fairly reliable recognition of electrical discharges with different energy densities and local overheating. At the same time, under operating conditions in oil-filled equipment, in addition to electrical discharges or local overheating, combined faults (overheating turning into discharges, discharges accompanied by overheating, etc.) may develop, and several faults of different types may develop simultaneously [6-8]. The recognition of such faults is associated with certain difficulties, which are due to either the complete or partial absence of standardized values of diagnostic criteria (values of gas ratios, gas percentages, and ratio of concentration of gases to the gas with the maximum content) that correspond to combined faults. To date, among the large number of existing methods [4, 5], only a few of them allow the recognition of combined faults. As noted in [9], these methods include the Duval [10] and Gauda [11] triangles, the 3 Ratio Technique (TRT) [12], the nomogram (gas pattern) method [13], the Electric Technology Research Association (ETRA) [14], and the method developed by Müller, Schliesing, and Soldner (MSS) [15]. However, all of the methods mentioned often only allow for the detection of Discharge&Thermal (DT) faults. The exception is the MSS method, which also allows for the detection of Local overheating and partial discharge (T&PD). The number of detectable combined faults affects the reliability of detection [16].

It is quite obvious that failure to recognize or misdiagnosis (especially in cases where overheating accompanied by discharges is interpreted as overheating) can lead to negative consequences, including accidental damage to equipment. Therefore, the development of new methods and the improvement of existing methods for interpreting DGA results is a pressing task. To date, there have been a large number of studies devoted to improving existing methods for interpreting DGA results using various machine learning methods [17]. Recent trends show that several machine learning methods are used simultaneously to improve existing methods, such as K-Nearest Neighbors (KNN), random forest (RF), support vector machine (SVM), decision tree (DT), backpropagation neural network (BPNN), clustering methods, and many others [18-23]. However, it should be noted that practically all such improvements are based on methods that cannot recognize combined faults, for example, IEC [1], Dörnenburg [24] and/or Rogers [25] ratio methods. In this regard, the development of new methods and the improvement of existing methods for interpreting DGA results, with the aim of increasing the reliability of recognizing combined faults, is a relevant and important task.

II. GENERAL PROCEDURE OF THE PROPOSED METHOD

The proposed method is designed to identify the type of defect in oil-filled equipment based on DGA results, in cases where the diagnostic criteria values calculated from the gas concentration values obtained from the tests do not correspond to any of the types of defects. To reduce the risk of false rejection, this method is recommended for use only if the concentration values and/or gas growth rate values obtained from the tests exceed the threshold values.

The method is based on the results of research conducted by the authors and the values of diagnostic criteria in oil-filled equipment with combined defects of various types [6-8]. Summarizing and analyzing the results obtained, it was established that the values of diagnostic criteria in oil-filled equipment with combined faults represent a combination of values characteristic of different types of faults simultaneously. Consequently, if the type of the main fault is determined, the type of the accompanying fault can be determined as the difference between the values of diagnostic criteria obtained for the equipment being diagnosed and the values of diagnostic criteria corresponding to the main fault.

The analysis is performed in several steps.

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A. Step I. Calculation of Diagnostic Criteria Values

Based on the DGA results, the following are calculated from the diagnosed equipment:

- percentage content of the following gases in the oil sample – H_2, CH_4, C_2H_6, C_2H_4 and C_2H_2.

$$A_{i\%} = 100\frac{A_i}{\sum} \quad (1)$$

In the equation, $A_{i\%}$ is the percentage content of the *i*-th gas; A_i is the concentration value of the *i*-th gas; Σ is the sum of the concentrations of hydrocarbon gases and hydrogen in the oil sample being analyzed.

- gas ratio values – CH_4/H_2, C_2H_4/C_2H_6 C_2H_2/C_2H_4, C_2H_6/CH_4, C_2H_2/CH_4 and C_2H_2/C_2H_6.
- a nomogram (gas pattern) of the fault is plotted.

To do this, according to [3], the gas with the maximum concentration is determined, and then the ratios of each gas to the gas with the highest concentration are calculated. A nomogram graph is plotted, on which the gases are placed on the X-axis in the following order – H_2, CH_4, C_2H_6, C_2H_4, C_2H_2, and the calculated ratios are plotted on the Y-axis. The resulting points are connected by a line, resulting in a nomogram that determines the fault type.

B. Step II. Determining the Type of the Main Fault

According to the results presented in [6-8], in oil-filled equipment with combined faults, the gas with the maximum content corresponds to the type of the main fault. Table I lists the types of faults that are characteristic of the maximum content of a particular gas. Using the results of calculating the percentage content of gases or calculating the coordinates of fault nomograms, as well as data from Table I, the type of main fault is estimated.

TABLE I. TYPES OF FAULTS FOR MAXIMUM CONTENT OF A PARTICULAR GAS

Gas with maximum content	Fault type
H_2	Partial discharges (PD), PD with low energy density (LEPD) and critical PD Spark discharges (with low and high energy density) Creepage discharges (with low and high energy density) Arc discharges (D)
CH_4	Thermal faults in the temperature range 150-300°C (T1) Thermal faults in the temperature range 300-500°C (T2.1)
C_2H_6	T1
C_2H_4	Thermal faults in the temperature range 500-700°C (T2.2) Thermal faults in the temperature range above 700°C (T3)
C_2H_2	High energy density discharges (D2)

If the gases with the highest content are H_2 or C_2H_2, the form of the discharges (partial, spark, creeping, arc) and their energy are estimated based on the analysis of fault nomograms and gas ratio values. When analyzing fault nomograms, the constructed operational nomogram should be compared sequentially with the reference nomograms, mentally discarding the sections of the operational nomogram that do not coincide with the reference nomogram until maximum coincidence is achieved. When analyzing the ratio values, it is necessary to take into account that the C_2H_4/C_2H_6 ratio does not always allow for a correct assessment of the discharge energy. For example, in D2, which are accompanied by T1, the value of this ratio will be less than 1, which may lead to an erroneous interpretation of the results.

If the gases with the highest content are CH_4, C_2H_6, and C_2H_4, then the hot spot temperature can be estimated based on the analysis of fault nomograms and gas ratio values, as in the previous case. As a rule, analyzing the C_2H_4/C_2H_6 ratio allows for a fairly accurate estimation of the hot spot temperature. It should also be noted that the presence of discharges in oil-filled equipment, in oil samples from which the gases with the highest content are CH_4, C_2H_6, C_2H_4, can be predicted if the value of the CH_4/H_2 ratio is <1 and (or) $0.1<C_2H_2/C_2H_4$ (if the main fault is T3) and $0.2<C_2H_2/C_2H_4$ (if the main fault is T3).

C. Step III. Determining the Type of Accompanying Fault

The simplest way to determine the type of accompanying flaw is to use the difference between the coordinate values of the operational nomogram and the coordinate values of the nomogram corresponding to the main fault [8] that are regulated by known standards. To do this, a nomogram is constructed based on the coordinate differences obtained as a result of the calculation using the algorithm described above. The resulting nomogram is compared with the reference nomograms and determines the type of accompanying fault.

III. EXAMPLES OF PRACTICAL APPLICATION OF THE PROPOSED METHOD

Consider the capabilities of the proposed method using the example of recognizing combined faults of different types.

A. D2, which are accompanied by T1.

The recognition procedure is illustrated using the DGA results for a 220 kV power transformer with a capacity of 25 MVA, in which D2 accompanied by T1 were detected. The gas concentration values, as well as the calculated values of the percentage content and gas ratios for this equipment, are given in Table II. The fault nomogram, plotted based on the results of operational data, is shown in Fig. I (*a*).

TABLE II. VALUES OF GAS CONCENTRATIONS, CALCULATED VALUES OF PERCENTAGE CONTENT AND GAS RATIOS FOR A 220 KV, 25 MVA TRANSFORMER

Gas concentration values, ppm				
H_2	**CH_4**	**C_2H_6**	**C_2H_4**	**C_2H_2**
337	89	300	176	295
Gas content, %				
H_2	**CH_4**	**C_2H_6**	**C_2H_4**	**C_2H_2**
28.154	7.435	25.063	14.703	24.645
Gas ratio				
CH_4/H_2	**C_2H_6/CH_4**	**C_2H_4/C_2H_6**	**C_2H_2/CH_4**	**C_2H_2/C_2H_6**	**C_2H_2/C_2H_4**
0.264	3.371	0.587	3.315	0.983	1.676

Comparing the values of gas ratios and gas percentages given in Table II, as well as the operating nomogram shown in Fig. 1 (*a*) with the corresponding values and reference nomograms regulated by known standards, it is easy to see that the obtained operating values do not correspond to any type of fault. Therefore, it is logical to assume that there is a combined type of fault. Analyzing the percentage of gases,

we note that the gas with the maximum content is H_2, followed by C_2H_6 and C_2H_2. The content of other gases is much lower. Such gas content suggests the presence of discharges accompanied by overheating. To determine the type of the main fault, we compare the operational nomogram with known reference nomograms [3], for which H_2 is the gas with the maximum content.

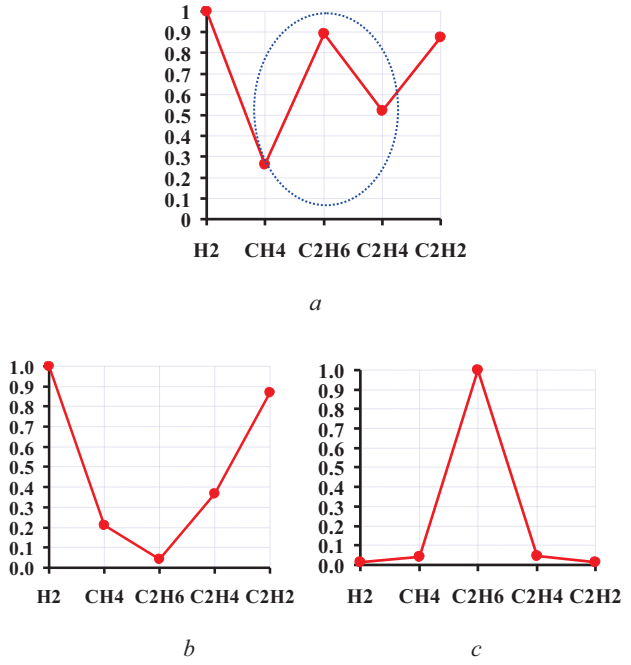


Fig. 1. Operational nomogram based on the DGA results for a 220 kV, 25 MVA transformer (*a*), nomogram corresponding to the main fault (*b*) and nomogram corresponding to the accompanying fault (*c*)

If we mentally remove the area highlighted by the oval in Fig. 1 (*a*), the closest match will be found with the reference nomogram, which corresponds to the D2 shown in Fig. 1 (*b*). Comparing the conclusion with the values of gas ratios, we find that out of 5 ratios, only two, C_2H_4/C_2H_6 and C_2H_2/C_2H_6, have values that do not correspond to D2. That is, we can assume that the main fault lies in D2. Considering that C_2H_6 is used in both the first and second ratios, and that this gas is second in terms of percentage content, which is clearly visible from the operational nomogram, it is logical to assume that the accompanying fault is one for which C_2H_6 is the gas with the maximum content. To determine the type of accompanying fault, the difference between the coordinates of the operational nomogram and the nomogram corresponding to the main fault was obtained. Based on the obtained difference values, a nomogram was constructed, which is shown in Fig. 1 (*c*). The obtained nomogram is not among the reference nomograms regulated in [3], but according to [13, 26], such a nomogram corresponds to T1. Thus, the diagnosis made using the proposed method completely coincides with the actual one.

B. T3 accompanied by D2

Consider an example of detecting a thermal fault accompanied by discharges. As input data for the calculation, we use the results of DGA for a 500 kV power transformer with a capacity of 300 MVA, which was damaged by an arc discharge. When the transformer was opened, it was found that the tie rod had traces of arc discharges at its ends, which were caused by the destruction of the rod insulation as a result of overheating. The gas concentration values, as well as the calculated values of the percentage content and gas ratios, are given in Table III. The fault nomogram, constructed based on the results of operational data, is shown in Fig. 2 (*a*).

TABLE III. VALUES OF GAS CONCENTRATIONS, CALCULATED VALUES OF PERCENTAGE CONTENT AND GAS RATIOS FOR A 500 KV, 300 MVA TRANSFORMER

Gas concentration values, ppm				
H_2	**CH_4**	**C_2H_6**	**C_2H_4**	**C_2H_2**
318	192	99	755	386
Gas content, %				
H_2	**CH_4**	**C_2H_6**	**C_2H_4**	**C_2H_2**
18.171	10.971	5.657	43.143	22.057
Gas ratio				
CH_4/H_2	**C_2H_6/CH_4**	**C_2H_4/C_2H_6**	**C_2H_2/CH_4**	**C_2H_2/C_2H_6**	**C_2H_2/C_2H_4**
0.604	0.516	7.626	2.010	3.899	0.511

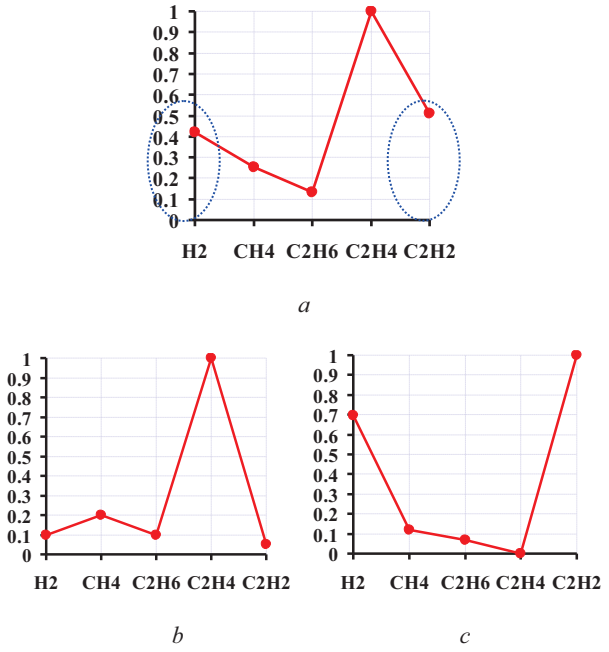


Fig. 2. Operational nomogram based on the DGA results for a 500 kV, 300 MVA transformer (*a*), nomogram corresponding to the main fault (*b*) and nomogram corresponding to the accompanying fault (*c*)

Unlike the previous case, the gas ratio values given in Table III allow us to predict the presence of D2 in this transformer according to [1], or spark (creeping) discharges according to [3]. At the same time, analysis of the percentage content of gases shows that in the oil sample from this transformer, the gas with the highest content is C_2H_4, followed by C_2H_2 and H_2. Such gas content is not characteristic of D2. According to [3], high-energy discharges are characterized by a stable gas content: 60-65% of H_2, 25-28% C_2H_2, 5-6% C_2H_4, 3-4% CH_4, and 0.5% C_2H_6. In the key gas method [27], the following percentage content of gases is regulated for high-energy discharges: 60% H_2, 30% C_2H_2, 3% C_2H_4, 5% CH_4, and 2% C_2H_6. That is, the calculated values of the percentage content of gases do not coincide with the values of this criterion standardized for D2. Thus, the diagnosis that was made using the values of gas ratios is contradicted by the values of the percentage content of gases. This circumstance once again confirms the fact that it is not enough to use only one diagnostic criterion, be it gas ratio values, gas percentage values, or fault nomograms, to reliably identify the type of fault. Reliable identification can only be ensured by using all three criteria. As an additional argument, it is worth noting that in

available literature sources, faults with gas content similar to that analyzed are interpreted differently. Thus, in [28], the diagnosis was “High energy discharge,” in [29] “Thermal,” and in [30] “High energy discharge, paper degradation.” Therefore, the use of different criteria in relation to DGA results with identical gas content leads to different and not always correct diagnoses.

To determine the type of the main fault, we compare the constructed operational nomogram with known reference nomograms [3], for which the gas with the maximum content is C_2H_4. If we mentally remove the areas highlighted by ovals in Fig. 2 (*a*), the closest match will be found with the reference nomogram, which corresponds to T3 and D2, shown in Fig. 2 (*b*). To determine the type of accompanying fault, the difference between the coordinates of the operational nomogram and the coordinates of the nomogram corresponding to the main fault was obtained. Based on the obtained difference values, a nomogram was constructed, which is shown in Fig. 2 (*c*). The obtained nomogram is similar to the nomograms regulated in [3, 13], which correspond to D2, which fully coincides with the actual diagnosis.

IV. Comparative Analysis of the Reliability of the Proposed Method

To objectively evaluate the effectiveness of the proposed method, a comparative analysis was performed of the reliability of fault type recognition provided by the proposed method with the reliability of recognition provided by the norms and criteria regulated in the most well-known standards and author's methods. The test sample consisted of the results of DGA of oil-filled equipment in which combined faults were detected. The gas concentration values with the actual fault type are given in Table IV.

TABLE IV. Gas Concentration Values with the Type of Faults in Oil-Filled Equipment in which Combined Faults were Detected

No.	Fault type	H_2	CH_4	C_2H_6	C_2H_4	C_2H_2
1	LEPD&T1	832	69	58.3	14	1.1
2	Critical PD&T1	219.1	8.8	15.9	14.2	100.8
3	D&T1	7880	7550	4730	300	2970
4	D&T1	1559	447	418	73	1075
5	D&T2.1	1650	950	840	1160	480
6	T1&D	12	325	3	1	108
7	T1&D2	38	212	47	15	78
8	T2.1&D	176	205.9	47.7	75.7	68.7
9	T2.1&D2	1585	4829	186	1839	2572
10	T2.1turning into an arc	292	346	32	313	196
11	T1&PD	151	8	151	10	8
12	T1&D	88.5	22.8	108.4	12	7.5
13	T1&D2	118	122	130	20	93
14	T1 turning into an arc	7746	2016	7806	1443	6945
15	T2.2&D	627	846	614	1629	201.9
16	T2.2 turning into an arc	3997	231	1726	5584	0
17	T2.1 turning into an arc	241	729	408	786	289
18	T3&PD	1417	114	296	2096	0
19	T3&D	1053	507	297	1440	17
20	T3&D2	27	12	19	125	65
21	T3 turning into an arc	747	814	63.2	851	695
22	D2&PD	2751	153	97	73	5735
23	D2&T1	16	31.4	157.9	86.4	169.4
24	D2&T2.1	76	81	21	51	270
25	D2&T3	71.0	79.0	3.0	72.0	115.0

As in [26], the analysis determined the statistics of correct diagnoses (numerator in column 1), partially correct diagnoses (denominator in column 1), erroneous diagnoses (numerator in column 2), and recognition failures (denominator in column 2). For combined faults, taking into account the different rates of development of electrical and thermal faults, as well as the consequences of deciding on the presence of a thermal and electrical fault instead of a combined one, the results of diagnosing electrical faults (regardless of the type of electrical fault) were classified as partially correct, and those for thermal defects were classified as incorrect. The results of the comparative analysis are presented in Table V.

TABLE V. Results of the Reliability Analysis of Fault Type Recognition

Method	∑, %	
	1	*2*
IEC 60599 [1]	– 4	16 80
SOU-N EE 46.501 [3]	– 8	48 44
Dörnenburg ratio method [24]	– 8	4 88
Rogers ratio method [25]	– 40	32 28
Duval Triangle 1 [10]	8 72	20 –
Duval Pentagon [31]	– 40	60 –
ETRA [14]	16 76	8 –
MSS [15]	8 28	12 52
4 Gases Square [32]	– 68	32 –
4 Gases Combination [32]	– 52	48 –
TGRC [33]	– 64	32 –
4 Gas Ratio [34]	– 28	16 56
2-Shapes Graphical Method [35]	– 80	20 –
Mansour Pentagon [36]	– 48	52 –
Gouda Triangle [11]	8 72	20 –
TRT [12]	– 72	28 –
Proposed method	96 4	– –

As can be seen from Table V, among the methods that can recognize combined faults, the ETRA method provides the highest recognition reliability (16%). However, it should be noted that in this method, the combined fault region is overlapped with the T3 region (Fig. 3), which affected the recognition reliability. Among other methods that can recognize combined faults, the accuracy of correct diagnosis does not exceed 8%, and the TRT method was unable to recognize any combined faults. If we analyze the results of partially correct diagnoses (recognition of electrical faults), the most reliable methods are the 2-Shapes Graphical Method (80%, Fig. 4), ETRA (76%, Fig. 3), the TRT method, Duval Triangle 1 (Fig. 5) and Gouda Triangle (72% each), as well as 4 Gases Square (68%, Fig. 6). As for commonly used standards and methods, the Dörnenburg ratio method failed to diagnose 88% of cases, and IEC 60599 (Fig. 7) failed to diagnose 80% of cases.

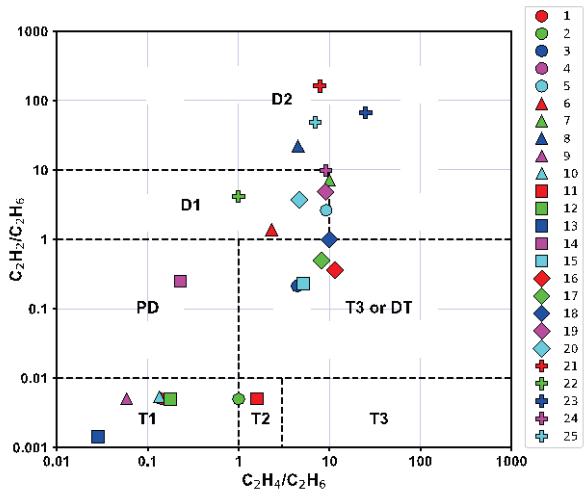


Fig. 3. Recognition results using ETRA method

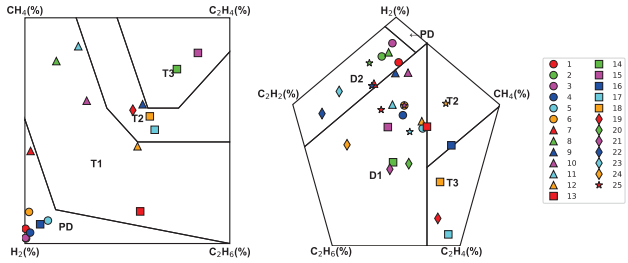


Fig. 4. Recognition results using 2-Shapes Graphical Method

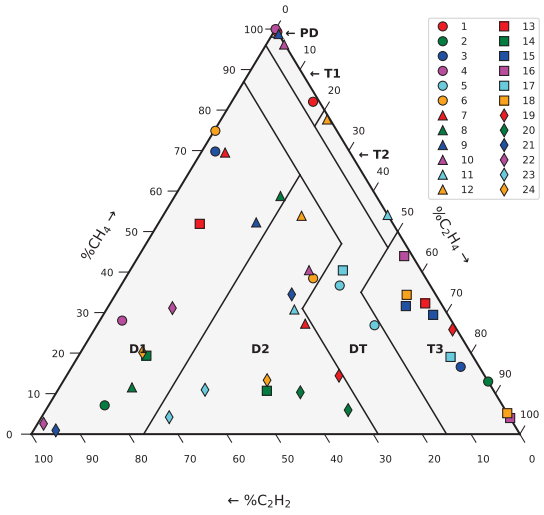


Fig. 5. Recognition results using Duval Triangle 1

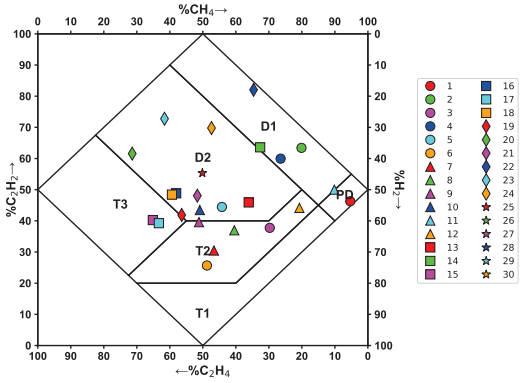


Fig. 6. Recognition results using 4 Gases Square

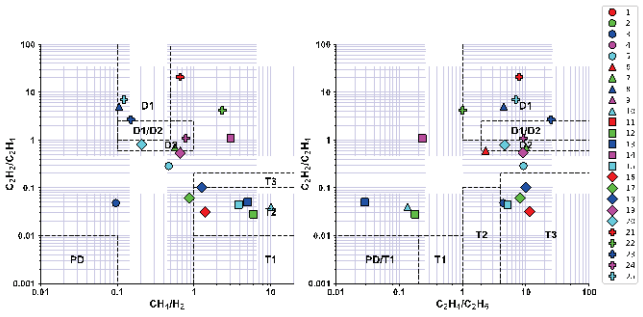


Fig. 7. Recognition results using IEC 60599

At the same time, the proposed method was able to make a correct diagnosis in 96% of the cases analyzed, and only 4% received a partially correct diagnosis. The results obtained indicate the high effectiveness of the proposed method in recognizing combined faults.

CONCLUSIONS

Based on the analysis of gas ratios, gas percentages, and nomograms obtained from DGA results, it was established that for oil-filled equipment with combined faults, the values of these criteria represent combinations that correspond to two types of faults simultaneously. This circumstance makes it difficult to recognize such faults using traditional methods. It is shown that for reliable recognition of combined faults, it is necessary to use both gas ratio values and gas percentage values, as well as fault nomograms. Using only one of these criteria can lead to either a false diagnosis or a failure to recognize the fault. A method has been proposed for recognizing combined faults, which consists of sequentially determining the type of the main and accompanying faults based on the analysis of all three diagnostic criteria. A distinctive feature of this method from existing ones is the ability to recognize combined faults in conditions of complete or partial absence of a regulatory framework. The comparative analysis demonstrated the high reliability of the proposed method in recognizing combined faults. In particular, the proposed method allowed establishing the correct diagnosis in 96% of all cases, while the best indicator of correct diagnoses using known methods ranges from 8 to 16%. At the same time, the use of classical recognition methods, such as the IEC method and the Dörnenburg method, leads to recognition failure in 80 and 88% of all cases, respectively, due to the absence of diagnosis areas corresponding to combined faults in these methods. The high level of recognition reliability provided by the proposed method indicates its high efficiency in recognizing combined faults under conditions of a limited regulatory framework.

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Methods and Ways of Measuring Dielectric Parameters of Bioobjects

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***Abstract*— The paper considers methods and ways of measuring dielectric parameters of biological objects, namely the methods and means of remote measurement of dielectric parameters of biological objects with an assessment of the characteristics of reflectometric systems for remote measurement of the permittivity of biological objects for the creation of information and measuring systems for reproducing, storing and transmitting the value of the complex dielectric constant of biological objects in a wide frequency range of electromagnetic radiation (EMR). An assessment of the influence of interference on the accuracy indicators of the reflectometer quality and the determination of the main components of the spectral purity of the reflectometer output signal are carried out.**

Keywords*— *Dielectric constant, Biological objects, Reflectometers, Electromagnetic fields, Remote measurement

I. Introduction

When using electromagnetic energy (EM) for diagnostics and treatment of diseases in medicine, for research of molecular processes in information biology and medicine, which forms the basis of genetics and breeding, for drying seeds and fruits in agriculture, pre-sowing seed treatment, soil disinfection, pasteurisation of juices and milk, diagnostics and treatment of animals, technical means for remote measurement of dielectric permittivity (DP) of biological objects in a wide frequency range are now necessary [1]–[3].

In turn, DP spectroscopy of biosubstances and materials for the needs of medicine, biology and agriculture requires increased accuracy and sensitivity of measurements.

The creation of technical means for dielectric spectroscopy of biosubstances and materials that meet modern requirements for the accuracy and sensitivity of measurements in a wide frequency range is a rather complex task. The solution to this problem is possible on the basis of the method of remote reflectometry with the creation of a digital frequency synthesizer with high frequency stability and spectral frequency of the output signal as an exciter [4]–[6].

II. Analysis of literature sources

In [7]–[12] it is proved that the impact of an electromagnetic field (EMF) enhances and accelerates the fight against diseases by mobilizing the body's own capabilities. Penetrating the body, electromagnetic signals at certain resonant frequencies are transformed into information signals that control and regulate the recovery processes in it. Moreover, the recovery processes will largely depend on the DP of organs [16]–[21].

The significance of the DP of biological objects and its change in the process of processing these objects with EMF allows us to judge the mechanism of the effect of EM energy on them and the change in the structure of these complex heterogeneous systems. It has been shown that the change and analysis of the dynamics of the DP of biological tissues, for example, the blood component in EMF, is currently acquiring great importance in practical medicine for express diagnostics of myocardial infarction [7]–[12] .

The magnitude of the DP measured remotely by the radio wave method can be used to detect malignant formations inside biological tissues. This diagnostic method can be used in many practical applications requiring precise determination of internal vertical DP profiles [13]–[14].

The issues of optimal application of electromagnetic energy in agricultural production are connected, first of all, with the study of dielectric properties of biological objects [15]–[16]. The study of dielectric properties of seeds and soil depending on temperature, humidity, under the influence of EMF, will allow to determine the modes of soil cultivation, depth of seed placement, optimal parameters of EMF during seed treatment.

Animal experiments have shown that EMF can be used to treat bronchopneumonia, severe and purulent pleurisy in sheep and cows. Studies on horses, cattle and small domestic animals have shown that EMF can be used to treat myositis, arthritis, enterocolitis, emphysema and fatigue, and can promote rapid wound healing without the use of antibiotics [17]–[20].

The radio wave method based on the analysis of dielectric parameters can replace numerous indirect methods for measuring multicomponent mixtures.

However, the methodology of multicomponent polydisperse systems in the microwave range has not been practically developed and requires expanded research in both frequency and temperature ranges.

When solving problems related to low-temperature preservation and long-term storage of biological objects: agricultural sperm, blood, bone marrow – special requirements are imposed on the control of the properties of the preserving medium [21]–[23]. First of all, it is necessary to accurately

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determine and maintain the hydration capacity of the preserving medium, which is determined by the quantitative change in structurally free moisture in the solution with a given change in the total moisture content in it. Otherwise, the use of the solution leads to a significant decrease in the survival of the biological object during its preservation [24]–[25].

Existing means of monitoring liquid hydration have insufficient accuracy and sensitivity of measurements in such a wide range of physicochemical properties that characterize preservative solutions [24]–[25].

The above-mentioned shortcomings inherent in existing means of monitoring liquid hydration can be eliminated by using radio wave methods for measuring moisture content in liquids in the extremely high-frequency range of wavelengths [13]–[14].

The basis of radio wave methods for monitoring the hydration properties of solutions is the fact that the absorption coefficient of EM energy of free and bound water differs by several orders of magnitude.

Radio wave methods are promising when using dielectric spectroscopy for plant breeding for frost resistance [13]–[16]. In tissue, cryoscopic methods were used to find patterns in the behavior of dielectric properties of plant tissue in the frequency range of 10...450 MHz, caused by changes in water bonding.

From the conducted analysis of the literature [7]–[25], it follows that for the effective use of EMF in medicine and agriculture, it is important to create information-measuring systems for reproducing, storing and transmitting the magnitude (value) of the complex DP of biological objects in a wide frequency range of EMF. In Ukraine, systematization of data on the dielectric properties of biological objects has not been carried out to date. The few studies on dielectric spectroscopy of biological objects were carried out in limited frequency ranges and were scattered. Therefore, there is a need to create new broadband frequency information-measuring systems for measuring the dielectric properties of biological objects.

III. Purpose of the Work

The aim of the research is to analyze the methods and techniques for measuring the dielectric parameters of biological objects to create information-measuring systems for reproducing, storing and transmitting the magnitude (value) of the complex DP of biological objects in a wide frequency range of EMR.

To achieve the research goal, the following tasks and solutions were identified:

- conduct an analysis of the methods and means of remote measurement of dielectric parameters of biological objects;

- to evaluate the characteristics of reflectometric systems for remote measurement of dielectric parameters of biological objects, namely the influence of interference on the accuracy indicators of the reflectometer quality and the value of the main components of the spectral purity of the reflectometer output signal.

IV. Research Methods

A. *Methods and means for remote measurement of dielectric parameters of biological objects.*

Modern methods and means of measuring DP under normal conditions and under external influences are based on the properties of any substance to interact with EMF [26]–[27].

The nature of the interaction of EMF with the environment is determined by the properties of the environment: DP and magnetic permeability. Losses in materials and biological substances in the ultra-high-frequency and extreme-high-frequency ranges are associated with the polarization of bound charges, the orientation of dipoles, and the movement of charge carriers (ionic and electronic). The value of the complex DP of most substances' changes with frequency, i.e. they have dispersion. One type of dispersion is determined by the Debye relaxation with an absorption peak ε″ at a frequency of $1/2\,\pi\,\tau$ [26], [27].

Another type of dispersion is determined by quantum transitions in an atom or molecule from one energy level to another with an absorption peak at the transition frequency. Therefore, dielectric properties are determined by processes that dominate in biological environments, and they can be described on the basis of the indicated relaxation patterns [13]–[14], [21], [26]–[27].

Typically, the free-space method is used to measure the DP of flat dielectric sheets with low losses [13]–[14], [21], [26]–[27]. When studying sheet materials with low losses, only the phase of the transmission coefficient of a flat dielectric is measured. linearly polarized wave through a dielectric layer.

When measuring the phase of the transmission coefficient, a dielectric sheet is placed between the transmitting and receiving antennas of the measuring setup. The phase shift is measured using a measuring line included in the reference signal channel, as shown in Fig. 1.

The phase shift is equal to:

$$\arg T - 2\pi d/\lambda_0 = 2(2\pi d/\lambda_0)\Delta Z\,, \qquad (1)$$

where λ_w – wavelength in the measuring line, ΔZ – displacement of the minimum of the standing wave, d – thickness of the dielectric, T – pass coefficient.

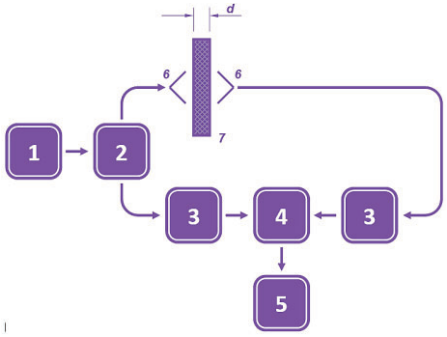


Fig. 1. Structural diagram of the DP measurement by the free space method using a measuring line: 1 – generator; 2 – splitter; 3 – decoupling attenuators; 4 – measuring line; 5 – indicator; 6 – transmitting and receiving antennas; 7 – dielectric sample of thickness d

Denoting the right-hand sides of (1) through $\Delta\varphi$ we obtain:

$$\arg T = \Delta\varphi + 2\pi d/\lambda_0\,. \qquad (2)$$

Given a value of arg T, the value of ε can be found by solving the transcendental equation graphically or numerically.

Usually, when calculating ε, an approximate value is used:

$$\arg T \approx (2\pi d/\lambda_0)\cdot\sqrt{\varepsilon}d, \qquad (3)$$

obtained without taking into account multiple reflections of the wave inside the dielectric sheet. The value ε is calculated:

$$\varepsilon = \left(1+(\Delta\varphi/2\pi)(\lambda_0/d)\right)^2, \qquad (4)$$

If $\Delta\varphi$ is determined by the magnitude of the horn displacement, then:

$$\varepsilon = \left(1+\Delta l/d\right)^2, \qquad (5)$$

The source of systematic errors during measurement will be, first of all, multiple reflections from the dielectric sheet and horns, affecting the phase measurement result.

The standing wave method has the following disadvantages: measurement only at a fixed frequency; narrow band; difficulty in automating the measurement process. When conducting research using a measurement method based on the analysis of the relationships between incident, transmitted and reflected signals from the object under study, reflectometers are used, the structural diagram of which is shown in Fig. 2.

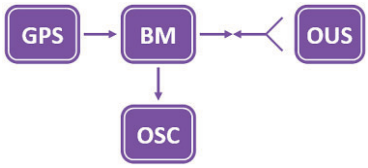


Fig. 2. Structural diagram of a reflectometer for measurements in free space

Here, the generator's probing signal (GPS) is fed to the object under study (OUS) through a broadband mixer (BM) (implemented as a transmission line with an input and output – a «pass-through» mixer) and is simultaneously observed on the oscilloscope screen (OSC) after stroboscopic conversion. The signal reflected from the inhomogeneities of the object under study, having passed through the mixer, is observed on the oscilloscope screen together with the probing signal. These signals, after Fourier transformations, determine the scattering coefficients S_{11} and S_{12} of the sample material.

$$S_{11} = S_{22}\frac{\Gamma\left(1-e^{-2x}\right)}{\left(1-\Gamma^2 e^{-2x}\right)}, \quad (6); \qquad S_{12} = S_{21}\frac{\Gamma\left(1-\Gamma^2\right)e^{-x}}{\left(1-\Gamma^2 e^{-2x}\right)}, \quad (7)$$

$$\Gamma = \frac{\sqrt{\mu}-\sqrt{\varepsilon}}{\sqrt{\mu}+\sqrt{\varepsilon}}, \quad x = j\gamma_0\sqrt{\mu\cdot\varepsilon}, \quad (8); \qquad \left.\begin{array}{l}\mu = \mu' - j\mu'' \\ \varepsilon = \varepsilon' - j\varepsilon''\end{array}\right\}. \quad (9)$$

Here μ', ε' – real, and μ'', ε'' – imaginary components of complex quantities ε and μ.

With known values of S_{11} and S_{21}, the system of expressions (6), (7) can be solved for ε and μ as follows:

$$\varepsilon = \sqrt{DC}, \quad \mu = \sqrt{C/D}, \quad (10)$$

$$D = \frac{1-2S_{11}+|S|}{1+2S_{11}+|S|}, \quad (11)$$

$$C = \gamma^2/\gamma_0^2, \quad \gamma = \gamma_u + n\pi, \quad n = 0,1,2... \quad (12)$$

$$\gamma_u = \ln\left(P \pm \sqrt{P^2-1}\right), \quad P = (1-|S|)/2S_{21}, \quad (13)$$

$$|S| = S_{11}^2 - S_{21}^2. \quad (14)$$

From (6)–(14) it follows that the simultaneous measurement of the frequency dependences $S_{11}(f)$=$S_{21}(f)$ of the sample, in accordance with Fig. 2, in principle allows us to solve the problem of determining the frequency dependences of the complex values of μ and ε. However, the practical implementation of obtaining the required characteristics for studying biological objects in free space by such a method of determining ε and μ is associated with overcoming a number of difficulties.

The most widely used reflectometers in remote sensing are the reflectometers with pulsed radio signals [28]–[32].

In [31] a reflectometer, in which a pulse generator on a Gunn diode with a power of about 3 W at a carrier frequency of 6.2 or 6.5 GHz was used, is described. The reflected signals were fed through the circulator to the input of a stroboscopic oscilloscope. This device made it possible to detect inhomogeneities with a reflection coefficient of about $5\cdot10^{-4}$.

A similar device with a probe pulse generator on a low-power Gunn diode is described in [32]. This device operates in the frequency range of 7–11 GHz with a dynamic range of about 40 dB.

In reflectometric measurements, the ability to tune the carrier frequency of the probing signal in a wide frequency range is of great importance, which allows the reflecting object to be assessed based on the frequency response of the reflection coefficient modulus [28]–[29]. Probing radio pulses with a duration of 2...200 ns and a power of about 70 mW are obtained using a high-speed ultra-wideband *p-i-n* module diode from a continuous signal of an external harmonic oscillation generator with a frequency of 1...18 GHz. Reflected signals can be directly observed on a broadband oscilloscope, which provides a dynamic range of about 40 dB, quite sufficient, for example, for measuring the parameters of the antenna path elements of a reflectometric system.

In [28] a variant of a pulse reflectometer with a probing signal in the form of a long-duration radio pulse with a steep front is described. In this case, the reflected signals are repeatedly superimposed on one another, which in principle allows their modules and phases to be measured. The final result of reflectometric measurements is a set of two or three parameters of the reflected signal (amplitude, position of the center of gravity, filling phase), while hundreds of readings of instantaneous signal values are measured on an oscilloscope. Therefore, optimization of the measurement technique for the parameters of inhomogeneities can provide a significant energy gain.

In the correlation reflectometer [31], a signal generator with linear frequency modulation was used as a source of the probing signal. The accuracy of measuring the module $|\Gamma|$ of 3%, phase 3^0; $|\Gamma|>10^{-4}$ were measured.

Correlation reflectometers with linear frequency modulation and averaging of the correlation product of the probing and reflected signals at intermediate frequencies have become quite widespread, which makes it possible to construct a multichannel correlometer circuit without an analog delay line.

The circuit of such a correlometer contains a multiplier of the probing and reflected signals and a system of filters tuned to frequencies F_e [32], F_l=$k(f)2l/v_f(f)$, where l is the distance to the inhomogeneity; v_f – phase velocity; $k(f)$ – frequency tuning speed, which should be selected such that the ratio $k(f)2l/v_f(f)$ remains constant.

There are several known implementations of measuring devices that produce measurement results in the time domain, but use long-duration harmonic signals when measuring objects. In [33], the results of measurements of parameters of reflecting objects under the earth's surface are presented, performed using a signal synthesized from 128 harmonic signals in the range of 0.4 – 3.56 GHz.

B. Analysis of the characteristics of reflectometric systems for remote measurement of dielectric parameters of biological objects.

Based on the analysis carried out, the use of multi-frequency reflectometric system with fast frequency tuning of the reference oscillator is preferable. The most common method of changing the carrier frequency in multi-frequency reflectometric systems is fast frequency tuning of a low-power ultra-high frequency master oscillator with subsequent amplification of the signals generated by it in a wide-pole power amplifier [33].

As tunable master oscillators in the ultra-high frequency range, it is preferable to use frequency synthesizers. It is known that potential values of tactical and technical characteristics are determined by the parameters of probing signals and their processing system. With selected parameters of the signal and processing system, the implementation of the main potential characteristics depends on the stability of the parameters of probing signals, primarily phase characteristics, which are determined by the total influence of various causes [33]. It is also shown there that the total influence of phase distortions is determined by the expression:

$$\sigma_{\Sigma}^2 = \sigma_1^2 + \sigma_2^2 + \sigma_3^2 + \sigma_4^2 + \sigma_5^2, \quad (15)$$

where σ_{Σ} – dispersion of the phase of the probing signal; σ_1 – phase disp. in the receiving and transmitting path; σ_2 – phase disp. due to change in the position of the RS; σ_3 – phase disp. due to signal propagation; σ_4 – phase disp. at signal reflection from objects; σ_5 – phase disp. in the processing system.

The above mentioned causes of phase distortion lead to distortion of the output signal and ultimately degrade the potential characteristics of the reflectometric system [13]–[16], [28]–[33]. For dielectric spectroscopy in near-field probing of biological objects, the terms σ_2, σ_3 are negligibly small. It is known that the main characteristics of a reflectometer for dielectric spectroscopy of biological objects in free space are the resolution and accuracy of measurements [13]–[16], [28]–[33].

1) Evaluation of the influence of interference on the accuracy indicators of the reflectometer quality.

It is known that the resolution of the RS by the parameter α characterized by the interval $\Delta\alpha$:

$$\Delta\alpha \cong \Delta\alpha_{pot} \left| \frac{\mathrm{N}_0(f) + M(f)}{N_0(f)} \right|, \quad (16)$$

where $\Delta\alpha$, $\Delta\alpha_{pot}$ – the real and potential intervals of resolution of the RS by the parameter α, respectively; $N_0(f)$ – spectral density of internal noise, recalculated to the input; $M(f)$ – instantaneous spectrum of external interference at the RS input.

Usually the accuracy of a reflectometric system is estimated by the root-mean-square measurement error. In engineering practice, the formula for the potential root-mean-square measurement error of the parameter α is widely used to determine the real accuracy of a reflectometric system:

$$\sigma_{\alpha_{pot}} = \beta \Big/ \left(\sqrt{2E_s} \Big/ \left|\mathrm{N}_0(f)\right|\right), \quad (17)$$

where β – potential resolving power of the reflectometric system, E_s – signal energy.

In this case, loss coefficients k_1, k_2 are introduced, taking into account the reduction in the signal-to-noise ratio and the deterioration of β, respectively. The actual signal-to-noise value is found using the formula:

$$\frac{\sqrt{2E_s}}{\left|\mathrm{N}_0(f) + M(f)\right|} = \frac{\sqrt{2E_s}}{\left|N_0(f)\right|} \frac{\left|N_0(f)\right|}{\left|N_0(f) + M(f)\right|}. \quad (18)$$

$$k_1 = \left| \frac{N_0(f) + M(f)}{N_0(f)} \right|. \quad (19)$$

From expression (16) it can be seen that:

$$k_2 \cong \left| \frac{N_0(f) + M(f)}{N_0(f)} \right|. \quad (20)$$

Measurement accuracy can be estimated as follows:

$$\sigma_{\alpha} \cong \beta \Big/ \left(\sqrt{2E_s} \Big/ \left|N_0(f)\right|\right) \cdot k_1 k_2 = \sigma_{\alpha_{pot}} \left| \frac{N_0(f) + M(f)}{N_0(f)} \right|^2. \quad (21)$$

Thus, the analysis of formulas (16)–(21) shows that high resolution and measurement accuracy are ensured mainly due to the high purity of the spectrum of the reflectometer output signal.

2) Determination of the main components of the spectral purity of the reflectometer output signal.

In [13]–[16], [28]–[33] it is shown that the main contribution to the deterioration of the potential characteristics of the reflectometric system is made by the receiving and transmitting path.

If the noise spectrum of the synthesizer output oscillation is not only continuous, but also contains n discrete spectral lines, and if all interference components are mutually uncorrelated, then the purity of the spectrum of a real synthesizer is characterized by the expression:

$$D = \frac{P_n}{P_C} = \frac{P_{ns} + \sum_{K=1}^{n} P_{ng}}{P_C} = \frac{P_{ns}}{P_C} + \sum_{K=1}^{n} \frac{P_{ng}}{P_C}, \quad (22)$$

where P_{ns} – interference having a noise spectrum; P_{ng} – interference that has only one discrete line.

It can be concluded that in order to obtain a high reflectometer resolution ≤1%, it is necessary that the discrete components in the spectrum of the synthesizer output signal should be suppressed to the level of the synthesizer's own phase noise, i.e. to –160 dB.

Considering that for the operation of the reflectometric system in the centimeter and millimeter ranges, the specified speed and high purity of the spectrum are ensured by the use of frequency synthesizers, we will analyze the current state of the issue of constructing ultra-high frequency synthesis systems for the purpose of using them as exciters-heterodynes of reflectometers for studying biological objects in free space.

The main requirements for a frequency synthesis system of a reflectometer for dielectric spectroscopy are to ensure sufficiently fast tuning from one operating frequency to another within a given frequency range and high spectral purity of the output signal at the power level required for probing in free space [13]–[16], [28]–[33].

According to the conclusions and practical recommendations of this work in the centimeter and even more so in the millimeter wave bands, the most complete satisfaction of the requirements for the main characteristics can be obtained by an indirect method of synthesis on the basis of analogue and digital systems of phase frequency autotuning. This is due to the fact that the direct synthesis method does not allow obtaining high spectral purity of the output signal, since multiplication and other nonlinear frequency conversions of reference generators in the millimeter range lead to an unacceptable level of fluctuation and discrete side components in the spectrum of the output signal.

The work [34] describes a frequency synthesizer for ultra-high frequency waves with a very low noise level.

Various variants of frequency multiplier circuit solutions do not give a noticeable result in terms of deterioration of the spectral purity of the output signal [35], [36]. The development and application of frequency synthesis systems with multiplication is explained, first of all, by the desire to use existing decimeter and centimeter range synthesizers in the

required section of the millimeter range. However, the spectral characteristics of the output signals obtained in this case do not meet the requirements of many application cases, in connection with which publications appear on the development of millimeter range synthesizers using phase-locked loop systems [37]–[39]. The simplest way to transfer the frequency of existing synthesizers to the millimeter wave range while simultaneously obtaining the required spectral characteristics is to use phase-locked loop systems as active filters when multiplying the synthesizer frequency. A report on obtaining a frequency grid in the 220 GHz range with a bandwidth of Δf=3% is given in [40]–[41]. For this purpose, an industrial synthesizer Ч6-31 (1 Hz – 50 MHz) and 4 rings of the phase-locked loop system are used, with the help of which the output frequency of the synthesizer is multiplied to meter, 3-centimeter, 4-millimeter and submillimeter wave ranges.

V. Results and discussion

A. Methods and means for remote measurement of dielectric parameters of biological objects.

Reflectometers using video pulse signals have become widely used. The literature describes video-pulse reflectometers whose hardware function front duration is 28-100 ps. A pulse reflectometer, which has a tunnel diode mixer and operates using the principle of instantaneous value discriminator, has a particularly high sensitivity.

Video-pulse reflectometers with a front duration of the hardware function of 30-100 ps, noise floor 2-5 mV and the amplitude of the probing signal is 0.2-1 V, allow to determine directly from the reflectogram the deviation of the line wave resistance from the nominal value by 0.5 Ohm, the minimum capacitive heterogeneity 0.008-0.02 pF, minimum inductance 0.01-0.05 nH, air filled line resolution 4.5-15 mm, and in strip lines on a dielectric substrate – up to 1-5 mm.

They also allow measuring the parameters of both solid and liquid dielectrics in the range of 10^5 …10^{10} Hz with a measurement error of 1+3%. But this type of reflectometers, due to the low pulse power of the probing signal generators in the ultra-high-frequency range (for example, with a tunnel diode pulse front duration of 20 ps, the probe pulse amplitude on a 50 Ω line load is 0.2 V), cannot currently be used to study biological objects in free space. In addition, in video-pulse reflectometers using the fast Fourier transform, the results are given only at multiple frequencies, unlike frequency scanning, in which measurements can be performed at any selected frequency. Thus, the conducted analysis of the methods and current state of technical reflectometry devices allows us to draw the following conclusions: at present, there are no specialized reflectometers designed to measure the DP of biological objects in free space; for remote measurement of the DP of biological objects by reflectometric systems in a wide frequency range, it is necessary to use nano- and subnanosecond radio pulses and frequency scanning methods; the study of dielectric parameters of biological objects by remote means requires the development of measurement methods and specialized reflectometric systems.

B. Analysis of the characteristics of reflectometric systems for remote measurement of dielectric parameters of biological objects.

Based on the results of the conducted review related to the development of a millimeter-range frequency synthesis system, the following conclusions can be drawn: the direct method of synthesizing the frequency grid cannot be used in the millimeter range of radio waves, since it does not allow high requirements for spectral-fluctuation characteristics; the use of multi-ring analog and digital phase-locked loop systems in synthesizers allows for obtaining a frequency grid in virtually any range of radio waves with high spectral-fluctuation, energy and other characteristics, which will allow for the implementation of requirements for the exciter of reflectometric systems for remote study of the DP of biological objects in free space; among the known publications on frequency synthesizers, there are no frequency synthesizers that have high spectral signal purity in the specified wave ranges with a tuning speed that is satisfactory for operation in reflectometric remote measurement systems.

An analysis of the state of the art in the field of frequency synthesizers allows us to conclude that although the known literature examines the issue of the filtering properties of phase control, in relation to exciter synthesizers, no way of solving this issue while maintaining the required speed has been indicated; a number of authors have analyzed the PLL circuit of a synthesizer, taking into account the influence of its nonlinearity on the synthesizer's performance as a whole. However, this analysis has not been brought to numerical estimates; insufficient attention in known publications is paid to the analysis of the influence of side components in the spectrum of the output signal, the mechanism of their formation and ways of their elimination; the issue of using compensators to eliminate the comparison frequency and its harmonics in the spectrum of the output signal is absent in the known literature.

The lack of a solution to the above-mentioned problems hinders the solution to the issue of implementing the required quality indicators of the exciter of reflectometric systems for measuring DP of biological objects in free space.

VI. Conclusion

1. In order to increase the productivity of plants and animals, as well as to control the quality and safety of agricultural products, it is important to create reflectometric systems for remote measurement of the DP of biological objects for the reproduction, storage and transmission of complex DP in the frequency range of 0…300 GHz with a measurement error of 1.5...1.7%.

2. Analysis of existing methods and technical systems for measuring the DP of biological objects showed that there are currently no specialized reflectometric systems for remote measurement of the electrophysical parameters of biological objects in a wide frequency range.

3. For remote measurement of the DP of bioobjects using reflectometric systems, it is necessary to use a digital frequency synthesizer with suppression of discrete components of the output signal to a level of 160-200 dB as a reflectometer exciter.

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Mathematical Models of Centrifugal Compressor and Anti-Surge Methods

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***Abstract* — The paper discusses methods of mathematical modeling of the surge mode in centrifugal compressors, which is one of the most serious problems that can disrupt their stable operation. Various approaches to modeling are described, including an improved Moore-Greitzer model. Particular attention is paid to the application of computational fluid dynamics (CFD), machine learning and fuzzy logic methods for predicting and preventing surge. It is shown that the integration of various modeling approaches allows increasing the reliability and stability of compressor operation, as well as minimizing the risk of surge. The article also discusses the limitations of the proposed models and prospects for further research to create more accurate and adaptive solutions. The presented findings are of great practical importance for improving the efficiency and safety of centrifugal compressors in industrial conditions.**

Keywords — centrifugal compressor, compressor modeling, anti-surge control, model predictive control, Moore-Greitzer model

I. Introduction

Centrifugal compressors are an important part of many industrial and technical systems, providing compression of gases and vapors for various processes. The control system of a centrifugal compressor must maintain the compressor operating point of the operating characteristic within its safe operating mode. An important characteristic of compressor operation is the ability to maintain the compressor in operation during instabilities, such as surge mode.

Surge mode occurs when the compressor outlet pressure is high compared to the flow through the compressor, which leads to fluctuations in mass flow and outlet pressure. The flow through the compressor is inverted and the gas begins to move in the opposite direction, vibration occurs and the temperature rises. Understanding and modeling this mode is important to ensure reliable and safe operation of compressors.

Surge mode is a complex and critical aspect of centrifugal compressor operation. It can have serious consequences for production processes if not taken into account in the design and operation [1-11]. Surge mode can result in damage to the compressor and the entire system that uses it. Understanding and predicting this mode allows taking measures to prevent accidents and equipment stoppages. Knowing the conditions under which surge occurs helps engineers design more efficient systems that operate within safe limits, providing maximum performance at minimum cost. Preventing surge can significantly reduce wear and tear on compressor parts, which in turn reduces repair and maintenance costs. Compressors that operate without surging are more efficient, allowing for reduced energy consumption and associated costs.

Modeling the surge mode allows engineers to design compressors that are more resistant to this phenomenon, increasing their efficiency and reliability. The models can be used for real-time monitoring of compressor operation, detection of signs of approaching surge mode and taking appropriate measures. Their use for the training of engineers and operators makes it possible to better understand the nature of surge and methods of its prevention.

Thus, research in the field of mathematical modeling of the surge mode is of high relevance and is critically important for ensuring efficient, safe and reliable operation of centrifugal compressors in various industries.

The purpose of this work is to review mathematical models and approaches for determining the anti-surge mode; to analyze their advantages and disadvantages.

II. Physical Meaning of the Surge Mode

The operating principle of a centrifugal gas compressor can be described by a number of important operating constraints: flow rate, speed, and pressure head. The maximum pressure head of centrifugal compressors is achieved at a certain desired speed. In this region, there is an equivalent flow known as the stability limit. The compressor is stable if the pressure head is below, showing low friction with the compressor, and the flow is above the lower limit. In other words, the stability of the system is maintained if a decrease in pressure causes an increase in flow [12].

The compressor speed affects its stability. Therefore, the compressor can reduce the surge margin by increasing the rotation speed for a fixed gas flow and vice versa.

The surge regime in centrifugal compressors is one of the key aspects of research in the field of gas dynamics and turbomachinery. This regime is characterized by the reverse

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movement of gas through the compressor and can occur due to various factors such as changes in the compressor load, pulsations in the gas flow, aerodynamic instabilities, etc. In this mode, the gas moves backwards from the compressor outlet to its inlet, which can cause serious problems such as oscillations, vibrations, increased component wear, and even equipment failure [13].

For a deeper understanding and successful prevention of this mode, many mathematical models have been developed. A mathematical description of the surge mode in centrifugal compressors may include the following aspects.

1. Surge conditions. The surge mode may occur if the internal dynamic conditions of the compressor result in a negative gas velocity relative to the compressor blade rotation speed. This occurs when a low-pressure region forms at the compressor outlet, which results in a flow reversal. Mathematically, this can be described as a reverse direction of the gas velocity.

2. Equation of motion. The mathematical description of the surge mode includes the equation of gas motion inside the compressor. The Navier-Stokes equations describe the change in velocity and pressure in the flow. In the surge mode, these equations can lead to a change in the sign of the velocity component indicating a reverse flow.

3. Stability and instability. The surge mode is associated with a loss of stability in the gas flow. Mathematical methods of stability analysis can be used to determine when and under what conditions the flow becomes unstable and begins to invert.

4. Vortex structure analysis. The mathematical description of the surge mode includes the analysis of vortex structures that can occur in the compressor flow. Vortices can cause a change in flow direction and velocity inversion.

Thus, surge mode is a complex phenomenon that requires deep understanding and careful modeling to ensure reliable operation of centrifugal compressors in industrial environments.

III. Mathematical description of processes

The Moore-Greitzer mathematical model is used to describe the compressor dynamics [13]

$$\frac{d}{dt}\Phi = \frac{1}{l_C}\left(\Psi_{C0}(\omega) + H(\omega)\left(1 + 3\left(\frac{\Phi}{W(\omega)} - 1\right) - \right.\right.$$

$$\left.\left. - \frac{1}{2}\left(\frac{\Phi}{W(\omega)} - 1\right)^3\right) - \Psi\right) \quad \frac{d}{dt}\Psi = \frac{1}{4B^2 l_C}(\Phi - \Phi_T), \quad (1)$$

where Φ is the air flow through the compressor; Ψ is the compression ratio of the air flow by the compressor; B is the Greitzer parameter; $l_c = l_{in} + l_{comp} + l_{out}$, l_{in} is the suction pipe length; l_{comp} is the compressor length; l_{out} is the discharge pipe length; Φ_T is the network costs; $t = U_\tau/R$ is the relative time; U is the tangential velocity in the average diameter; R is the average compressor radius; T is the real time; $\Psi_{C0}(\omega)$ is the compression in the absence of costs

$$\Psi_{C0}(\omega) = K_k \cdot \omega^2, \quad (2)$$

where K_k is the compressor design factor; $W = K_w\omega$ is the half-width of the gas-dynamic characteristic; $H = K_H\,\omega^2$ is the half-height of the gas-dynamic characteristic; K_w , K_H are the compressor design factors; ω is the compressor rotation speed.

The Greitzer constant (B-parameter, or Greitzer parameter) is a dimensionless parameter that characterizes the inertia of the "compressor-discharge tract" system and its tendency to surge.

Formula for calculating the Greitzer parameter (B) [14]

$$B = \frac{U}{2a_s}\sqrt{\frac{V_p}{A_c L_c}}, \quad (3)$$

where U is the peripheral speed at the outlet of the impeller (m/s),

$$U = \frac{\pi D n}{60};$$

a_s is the speed of sound in gas (m/s),

$$a_s = \gamma R T_{in},$$

where D is the impeller diameter (m), n is the rotation frequency (rpm), γ is the adiabatic index, R is the gas constant, T is the inlet temperature (K), Vp is the volume of the discharge pipe (m^3), Ac is the cross-sectional area of the compressor (m^2), Lc is the reduced length of the discharge path (m).

If the tract is complex (with bends, valves), the equivalent length is used

$$L_c = \frac{V_p}{A_c}.$$

Equation (1) contains the variable Φ_T, which affects the equilibrium of the system. The throttle (load) characteristic $\Phi_T(\Psi)$ in the Moore-Greitzer equations describes the dependence of the flow rate through the throttle (e.g., a control valve or an outlet nozzle) on the pressure drop (head) in the system. This function models the resistance of the outlet path and relates Ψ – the dimensionless head (the pressure drops in the system, normalized to ρU^2), Φ – the dimensionless flow rate (normalized to AcU).

In the simplest case, throttling is described by a quadratic dependence (analogous to Bernoulli's law for gas flow through a nozzle)

$$\Phi_T(\Psi) = k_T\sqrt{\Psi - \Psi_0}, \quad (4)$$

where k_T is the throttle capacity coefficient (depends on the degree of valve opening); Ψ_0 is the minimum pressure at which the flow becomes zero (can be zero).

Thus, for real systems, taking into account friction and geometry, formula (4) takes the form

$$\Phi_T(\Psi) = C_D\sqrt{\Psi(P)}, \quad (5)$$

where

$$\Psi(P) = \frac{P_{out} - P_{in}}{\rho U^2}. \quad (6)$$

That is, the greater P_{out}, the greater Ψ and the more the throttle restricts the flow. Thus, depending on the gas pressure at the outlet (5), Φ_T grows, but more slowly than Ψ.

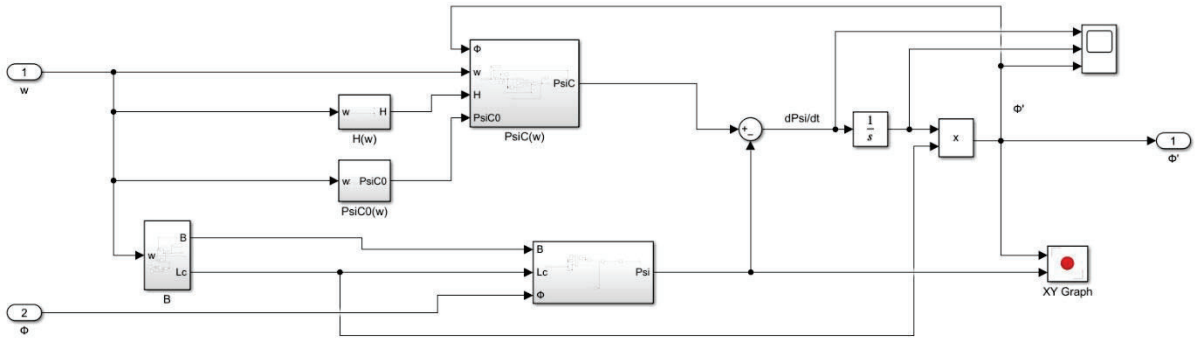


Fig. 1. Mathematical model based on the Moore-Greitzer model

Using the refined Moore-Greitzer mathematical model, shown in Fig. 1, it was possible to take into account the inertia of the gas flow through the compressor, as well as the quadratic relationship between pressure drops and flow rate. The mathematical model was created using the Matlab program. The qualitative characteristics of the transient processes $\Phi(t)$, $\Phi_T(t)$, obtained using this model, are shown in Fig. 2.

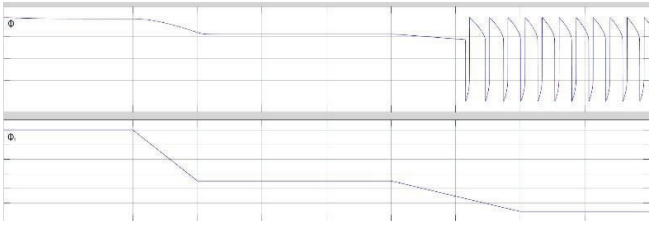


Fig. 2. Graphs of air changes in the network and through the compressor

IV. Anti-surge Control System (ASC)

Anti-surge control (ASC) is designed to prevent surges in compressors (Fig. 3). These systems must react quickly enough to minimize the impact on the process. Knowing the conditions under which surges occur allows you to avoid them and improve the compressor operating envelope. The main parameters that determine the compressor operating envelope are speed, pressure, and flow [15].

Anti-surge control determines the compressor operating point using flow, temperature, and pressure instrumentation. The compressor surge limit is compared with the current system operating point, and the control error is determined by calculating the deviation between them. A PID control algorithm is usually used to obtain the control signal for the recirculation valve. When the recirculation valve opens, a portion of the gas is diverted from the discharge to the suction side, preventing the pressure from increasing [16]. When the flow increases above the normal operating point, the valve closes and the compressor continues to operate normally.

There are many anti-surge systems, differing in surge prevention methods and mathematical models:

- **ASC model using advanced torque control with variable frequency drives.** This method allows for precise control of compressor speed, reducing the risk of surge;
- **ASC using a predictive model.** This model compares the readings of the predictive model with the existing PID controller. Studies have shown that this model is more efficient than PID control and minimizes steady-state error, but it requires a more complex algorithm;
- **Using the inputs and outputs of the compressor model to predict its behavior.** The predictive controller model is designed to prevent surge by increasing compressor efficiency by reducing the difference between the surge and control lines;
- **ASC model using fuzzy logic.** This model is self-predictive, and the controller's actions do not depend on the compressor map information, and the decision is made based on the compressor output data. These models and methods can significantly improve the operation of compressors, ensuring their stability and preventing critical situations such as surge.

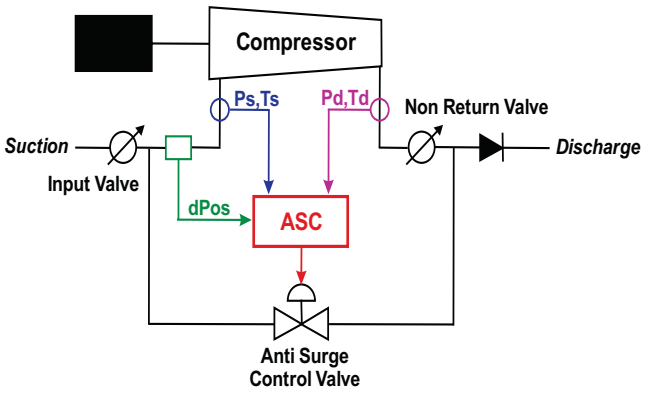


Fig.3. Example of anti-surge control system.

The most common method of anti-surge protection is to increase the air flow through the compressor by opening the anti-surge control valve (ACV) connecting the pressure line to the atmosphere (Fig. 3). Taking into account the air flow through the ACV, the mathematical model can be described by the equation [17], [18]

$$\frac{d}{dt}\Phi = \frac{1}{4B^2 l_C}(\Phi - \Phi_T - \Phi_{ACV}), \quad (7)$$

where Φ_{ACV} is the loss of the air flow of the ACV.

One of the laws of operation of the ACV, based on maintaining a constant air flow, can be represented as [19]

$$\Phi + \Phi_{ACV} = \Phi^*, \quad (8)$$

where Φ* is the optimal air flow through the compressor for a given rotation speed, providing the necessary stability margin.

It is necessary to take into account that in real systems $\Phi_T(\Psi)$ can have a hysteresis loop due to the inertia of the valve. Also, when adjusting the valve over time, it is necessary to take into account the equation of its motion as shown in the works [20], [21].

In semi-empirical models for determining the surge limit of a centrifugal compressor based on known parameters, such as inlet pressure Pin, outlet pressure Pout, inlet temperature Tin, one can use stability criteria based on the balance of flow and pressure. To calculate and build a mathematical model, it is necessary to control the compression ratio

$$\pi_c = \frac{P_{out}}{P_{in}}.$$

And also calculate gas consumption as reduced mass productivity [22]

$$\Phi = \frac{\dot{m}\sqrt{RT_{in}}}{D^2 P_{in}}, \quad (9)$$

where ṁ is the mass flow rate of air; R is the specific gas constant; D is the diameter of the compressor wheel.

The dimensionless pressure Ψ is calculated using the formula

$$\Psi = \frac{P_{out}-P_{in}}{\rho_{in}U^2},$$

where ρ_{in} is the air density at the inlet.

The Ψ(Φ) dependence is used to determine the surge limit. Surge occurs when the compressor characteristic has a positive slope ($\frac{d\Psi}{d\Phi} > 0$).

To assess the stability of the compressor, a criterion for comparing the derivative characteristics of the compressor and the throttle is used

$$\frac{d\Psi_c}{d\Phi} < \frac{d\Psi_T}{d\Phi},$$

where Ψc(Φ) – compressor characteristic; $\Psi_T(\Phi)$ – throttle (load) characteristic.

Violation of this inequality indicates the risk of transition to an unstable mode (occurrence of surge).

In engineering practice, the compressor characteristic Ψc(Φ) can be approximated using a third-degree polynomial

$$\Psi c(\Phi) = \Psi_0 + a\Phi - b\Phi^3,$$

where a, b are approximating coefficients; Ψ_0 is the pressure value at zero flow.

The stability boundary corresponds to the maximum point of the compressor characteristic, where the derivative is zero

$$\frac{d\Psi_c}{d\Phi} = a - 3b\Phi^2 = 0 \rightarrow \Phi_{crit} = \sqrt{\frac{a}{3b}}.$$

If Φ < Φcrit, then the compressor operates in a risk zone, where flow stall and transition to surge are possible.

Coefficients a and b can be determined from experimental data, factory characteristic Pout(ṁ) or typical curves. Construction of normalized characteristics Ψ(Φ) allows using the model for compressors of different sizes [23].

Semi-empirical models based on the balance of flow and pressure allow determining the stability limit of a centrifugal compressor with sufficient accuracy. The use of approximated characteristics and dimensionless parameters makes the method convenient for implementation in automatic diagnostics and control systems.

Modeling surge conditions using these equations is challenging due to the nonlinear and turbulent flow characteristics and the need for high computational power. Solving such equations typically requires multiple methods such as the finite element method (FEM) or finite volume method (FVM) and specialized computational fluid dynamics (CFD) software packages.

The improved model improves the efficiency of the system, providing a larger surge stability margin and reducing flow recirculation in transient modes. However, its application is limited, since the model does not take into account the features of the compression system, including rotor interaction and aerodynamic or geometric characteristics of the channel. Thus, there is a need for further research and development of new surge modeling and control methods with an emphasis on improving the accuracy of the models and their adaptability to different operating conditions.

The use of model predictive control (MPC) for compressor control can be considered, which was proposed in [3]. The compressor model used to design the MPC in this study is based on the work of Gravdal and Egeland. A dynamic compressor model was further developed and a speed-dependent compressor characteristic was included in the model. The model can be summarized using the following equations:

$$\frac{dw}{dt} = \frac{1}{J}(\tau_{drive} - \tau_{comp}), \quad (10)$$

$$\frac{d\psi_c}{dt} = \frac{1}{t_\psi}(\psi_{c,ss}(\dot{m}_1, w) - \psi_c), \quad (11)$$

$$\frac{d\dot{m}_c}{dt} = \frac{A_c}{L_c}(p_1\psi_c - p_2), \quad (12)$$

$$\psi_{c,ss} = c_0(w) + c_1(w)\dot{m}_1 + c_2(w)\dot{m}_1^2 + c_3(w)\dot{m}_1^3, \quad (13)$$

$$\tau_{comp} = \sigma r^2 |\dot{m}_1| w. \quad (14)$$

Equation (10) represents the inertial dynamics of the compressor shaft in the direction of its rotation, where w is the shaft speed and τ is the torque. Equation (11) captures the dynamic characteristic of the compressor pressure rise, where Ψ is the pressure ratio, $\dot{m}_1$ is the inlet flow rate, and t_Ψ is the time constant. Equation (12) represents the dynamics of the compressor mass flow rate, where p_1 is the inlet pressure, p_2 is the outlet pressure, A_c is the outlet duct cross-section, and L_c is the duct length. Equation (13) defines the cubic interpolation polynomial approximation identified from the compressor maps as a function of the compressor speed w and the inlet flow rate $\dot{m}_1$. Equation (14) defines the compressor torque τ_{comp} as a function of the inflow and compressor speed, where σ is the slip coefficient and r is the shaft radius.

As an example, it is possible to cite an anti-surge system of a centrifugal compressor implemented using a deep

predictive controller based on long-short term memory (LSTM) models. Such an approach provides the possibility of taking into account the dynamic tendencies of the functioning of the compressor unit in conditions of uncertainty of the initial data, and also meets the requirements for the safe operation of the equipment. The use of a PI-regulator integrated with the LSTM architecture allows not only to form more accurate estimates of compressed air consumption, but also to determine confidence intervals for the obtained forecasts. Based on the predicted flow rates, it is possible to calculate the pressure at the compressor outlet using the equation of state of an ideal gas. In cases where the operating point of the compressor crosses the boundary of the permissible area of functioning, the anti-pumping strategy is implemented by modifying the reference trajectory of the regulator. At the same time, taking into account the change in the operating mode and the practical experience of experts in the field of anti-pumping technologies, the trajectory used in the dynamic matrix control algorithm is subject to adaptive correction taking into account the experimentally obtained pumping line.

V. Conclusion

Mathematical modeling of the surge mode of centrifugal compressors plays an important role in ensuring their reliable and efficient operation. Understanding the aerodynamic processes and interactions of gas with compressor components helps to develop strategies to prevent the surge mode and improve its stability.

The research showed that each approach to surge modeling and control has its own advantages and limitations. For example, machine learning and fuzzy logic methods can significantly improve prediction and control, but require significant computing resources and expertise in tuning. The best results in surge prevention are achieved by integrating several methods, such as predictive control, fuzzy logic and traditional control methods. This ensures a high level of reliability and stability of compressors. The article emphasizes the need for further research and development of new surge modeling and control methods, with an emphasis on improving the accuracy of models and their adaptability to various operating conditions.

The improved Moore-Greitzer model improves the efficiency of the system by providing a larger surge stability margin and reducing flow recirculation in transient conditions. However, its application is limited because the model does not take into account the features of the compression system, including rotor interaction and aerodynamic or geometric characteristics of the channel.

In real centrifugal compressor control systems, mathematical modeling requires operating with input quantities that can be physically used as action signals and control signals. For this purpose, in future work, the mathematical model will be improved and reduced to the corresponding physical quantities of the input action.

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Phase-Specific Finite Element Modeling of Hip Bone Stress-Strain Response During a Jump

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***Abstract*— The objective of this research is to model and evaluate the stress-strain distribution in the femur throughout different jumping phases. The sequence includes standing, squatting, pushing off, flight, and landing. The distribution of von Mises equivalent stresses at three key moments of the jump cycle are received using finite element method. The numerical analysis revealed the stress distribution, quantified their intensities, and identified critical areas based on the applied loading conditions, bone geometry, material characteristics, and defined boundary constraints. In all three cases, the maximum of von Mises equivalent stresses is significantly below the ultimate strength threshold of 193 MPa, indicating that no damage to the bone is expected under the given conditions. But the observed pattern of stress distribution, although within safe limits for healthy bone tissue, may become critical under certain pathological or physiological conditions.**

Keywords—finite element method, bone hip, stress-strain state, different stages of jumping

I. Introduction

At present, a significant number of boundary value problems in continuum mechanics have been solved, providing models for real-world processes occurring in machine components and structural elements operating under intense thermo-mechanical loads. The ultimate criterion for the validity of a mathematical model remains the physical experiment. However, conducting full-scale experiments is often expensive and not always safe. Therefore, numerical methods, or virtual computer simulation techniques, have been rapidly evolving in recent years as powerful tools for solving boundary value problems in continuum mechanics [1-2]. These approaches allow engineers and designers to perform virtual testing of modern technical components in order to identify highly stressed regions and to fine-tune designs to avoid undesirable resonant frequencies during forced vibrations.

Another compelling area of application for computer modeling is in biomechanics. This includes the analysis of strength and stability phenomena in human tissues, such as the tympanic membrane, bones, and muscles, when subjected to intensive mechanical loading [3-4].

Among the computational methods used to model boundary value problems in the mechanics of deformable solids, fluid and gas mechanics, and brittle fracture mechanics, several have proven particularly effective. These include the finite element and boundary element methods [5], the finite volume method [6], methods of singular and hypersingular integral equations [7], the method of double substitution, the absorbing layer method, mesh-free methods, and various approaches based on potential theory [8-9].

It should be noted that the Finite Element Method (FEM) is widely and successfully used for analyzing the stress-strain state of human bones. For example, in [10] FEM was applied to assess the risk of femoral fractures in elderly individuals, in [11], an analysis of the stress–strain state of the hip joint was conducted in patients with posterior pelvic tilt. In [12], the stress-strain state of the femur was analyzed with consideration of the influence of plates and screws used for the fixation of femoral fractures.

Properties of human bone materials were investigated in [13], whereas the experimental investigations of these properties were done in [14-15]. Such studies are of great importance for the prevention and treatment of patients with hip joint disorders and may be effectively applied in clinical practice.

In this study, the stress-strain state of the femur during jumping was investigated for the first time. The numerical analysis provided the distribution of stresses, their magnitudes, and the identification of critical regions corresponding to the selected loading conditions, bone geometry, material properties, and boundary conditions.

II. Methodology

A. Boundary Value Problem for Deformable Solids

The fundamental equations of deformable solid mechanics are presented below.

To determine the stress-strain state of an elastic body occupying domain V, under external load $X = (X_1, X_2, X_3)$, a system of three second-order partial differential equations of elliptic type, known as the Lamé equations, is used

$$\mu \Delta U_j + (\lambda + \mu) \frac{\partial \vartheta}{\partial x_j} = X_j,\ j=1,2,3,\quad \vartheta = \mathrm{div} U, \qquad (1)$$

where:

$\Delta = \frac{\partial^2}{\partial x_1^2} + \frac{\partial^2}{\partial x_2^2} + \frac{\partial^2}{\partial x_3^2}$,

$\mu = E/[2(1+\nu)]$ is the shear modulus,

$\lambda = E\nu/[(1+\nu)(1-2\nu)]$ is Lamé's first parameter,

ν is Poisson's ratio,

U is the displacement vector.

Let n denote the outward unit normal to the surface bounding the domain V. To formulate the boundary

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conditions, we introduce a generalized operator from the theory of elasticity

$$T^{n(x)}\,U = 2\mu\frac{\partial u}{\partial n} + \lambda n \operatorname{div} U + \mu(n \times \operatorname{rot} U). \qquad (2)$$

Assume that the domain is decomposed as $V = V_1 \cup V_2$, with boundaries $S_1 = \partial V_1, S_2 = \partial V_2$, denoting the surfaces of the subdomains V_1 and V_2, respectively.

The boundary conditions are given by:

$$U(x)=U_1(x), \quad x \in S_1, \quad T^{n(x)}U(x)=X(x), \quad x \in S_2, \qquad (3)$$

where:
$U_1(x)$ is is the prescribed displacement field on S_1 (kinematic or Dirichlet conditions),
$X(x)$ is the applied surface traction on S_2 (traction or Neumann conditions).

Thus, the mixed boundary value problem (1)-(3) is formulated to obtain the unknown displacement vector. To define strains ε_{ij}, the Cauchy equations are used

$$\varepsilon_{ij}=0.5\,(u_{i,j}+u_{i,j}), \qquad (4)$$

and to obtain stresses, the Hooke law is applied

$$\sigma_{ij} = \delta_{ij}\lambda \operatorname{div} u + \mu\left(\frac{\partial u_j}{\partial x_i} + \frac{\partial u_i}{\partial x_j}\right) \quad i,j = 1,2,3, \qquad (5)$$

where δ_{ij} being the Kronecker delta.

B. *Finite Element Formulatin*

Consider the second boundary value problem. A finite element formulation is employed. To solve equations (1)-(3), we apply the method of weighted residuals [7]. The unknowns are represented as series expansions over a basis of finite functions $\{\psi_n\}|_{n=1}^{N_1}$, and the test functions are chosen from the same basis.

The displacement field is approximated in the form

$$U = \{N^u\}\{u_s\}, \qquad (6)$$

where:
$\{N^u\}$ is the matrix of shape functions for displacement values,
$\{u_s\}$, is the vector of nodal displacement values.

This approximation transforms the original continuous problem into a discrete one by projecting the solution onto a finite-dimensional subspace. By substituting the approximate solution into the governing equations (1) and applying the weighted residual method, most commonly the Galerkin approach, followed by integration over the computational domain, a discrete system of algebraic equations is obtained for numerical implementation. To facilitate the incorporation of boundary conditions, the divergence theorem is applied, converting part of the volume integrals into equivalent surface integrals. This step enables the natural (Neumann) boundary conditions to be seamlessly integrated into the weak formulation of the problem.

Thus, by integrating the governing equations over the domain and applying the divergence theorem, the finite element formulation of problem (1)-(3) can be expressed in matrix form as

$$[K_s]\{u_s\} = \{f_s\}. \qquad (7)$$

where:
$[K_s]$ is the global stiffness matrix,
$\{f_s\}$ is the known nodal force (traction) vector.

This algebraic system provides the discrete counterpart of the original boundary value problem, enabling numerical computation of the structural response.

C. *Problem Statement for the Stress–Strain Analysis of the Femur During Jumping*

We analyze the human femur of a 55-kg individual during the jumping motion. The movement sequence consists of standing, squatting, push-off, flight, and landing phases. The corresponding load profile on the femur for both legs is presented in Fig. 1, following [15]. The blue curve (labeled 1) indicates the femur orientation during the jump, where 0^0 corresponds to the vertical position. The red curve (labeled 2) shows the applied load. Since the measurements were obtained for both legs, only half of the total load is considered in the subsequent calculations.

From the graph in Fig. 1, the key characteristic points in terms of time and load were extracted and used to define the loading conditions for the model.

The bone's geometric model includes a large number of microscale surface features. It was developed through 3D scanning of an actual bone, and its geometrical model constructed in ANSYS, is presented in Fig. 2.

The material properties of the bone were taken from [14-15]. For simplicity, the material is assumed to be isotropic. Table 1 summarizes the material properties of the femoral bone.

The load is applied along the Z-axis to the femoral head, as illustrated in Fig. 3, simulating the mechanical forces experienced by the bone during physical activity. The boundary conditions include a rigid fixation of the femur at the knee joint to realistically represent physiological constraints. Stress analysis is conducted at three critical time points corresponding to different phases of a jump: the initial resting state, the moment immediately prior to takeoff, and the instant following landing when the load reaches its maximum value.

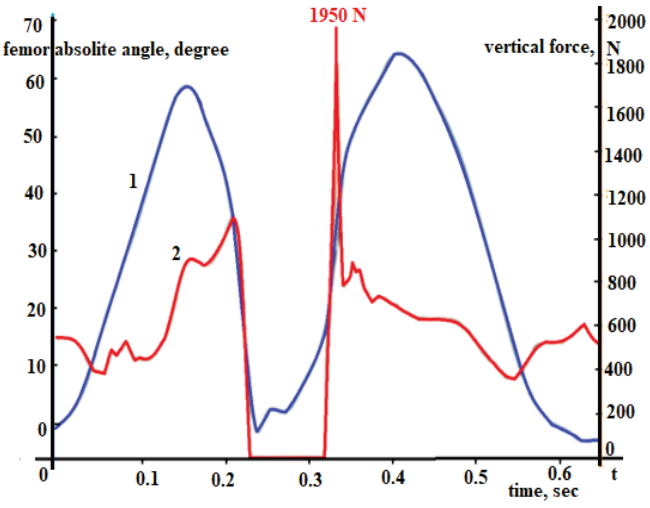


Fig. 1. Loading diagram of the femur, (adapted from [15]).

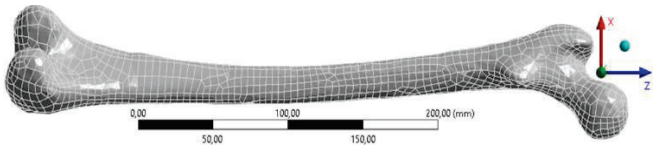


Fig. 2. Three-dimensional model of the bone.

The distribution and magnitude of the computed stresses at these stages are carefully evaluated and subsequently compared with the ultimate strength of the bone material to assess the risk of structural failure under the given loading conditions.

For a person weighing 55 kg, the average estimated height is approximately 168 cm, assuming an average Body Mass Index (BMI), [18].

Based on anthropometric correlations: the femur (thigh bone) length is approximately 43 cm, the minimum width of the femur (e.g., at the femoral shaft) is about 2.8 cm.

TABLE I. MATERIAL PROPERTIES OF THE FEMORAL BONE

Property	Value	Unit
Density	942	*KG/M3*
Young`s module	17	*MPA*
Poisson`s ratio	0.46	-
Ultimate Strength	193	*MPA*

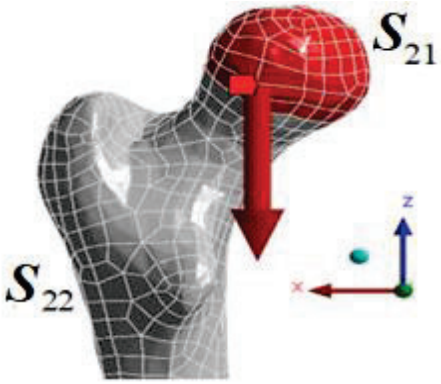


Fig. 3. Loading location.

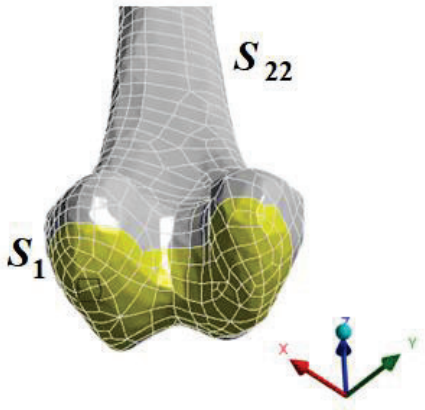


Fig. 4. Bone fixation.

The maximum width (e.g., across the femoral head or distal condyles) is approximately 8.5 cm, [18].

The zones of applied force and prescribed displacement were constructed in our numerical simulation and are presented in Figs. 3 and 4. In these figures, the boundary surface S_2 is divided into two parts, namely $S_2 = S_{21} \cup S_{22}$. These subdivisions were determined according to the specified boundary conditions. On the part S_{22}, no external forces are applied (zero-force condition).

Thus, the following boundary value problem is formulated for analysis of the stress-strain state of the femur under jumping:

$$\mu \Delta U_j + (\lambda + \mu) \frac{\partial \vartheta}{\partial x_j} = X_j, \tag{8}$$

$U_i(x)=0)$, $x \in S_1$,. $T^{n(x)}U(x)=X(x), x \in S_{21}, T^{n(x)}U(x)=0, x \in S_{22}$.

It should be noted that the right-hand side $\{f_s\}$ in (7) contains the dot products of the loads applied to the femur in region S_{22}, while the stiffness matrix $[K_s]$ is constructed with usage of the boundary conditions and the geometry of region S_1. This formulation reflects the partitioning of the problem domain, where S_{22} represents the area of external loading application, and S_1 defines the boundary conditions that govern the system's mechanical constraints, Fig. 3 and Fig. 4. Such a separation is essential for accurately modeling the mechanical response of the femoral structure under realistic physiological conditions and ensures that both loading conditions and boundary constraints are properly satisfied.

Following the development of the computational model (7), which incorporates the applied loading conditions on the femur (Fig. 3) and the corresponding fixation constraints (Fig. 4), a finite element mesh was generated, and numerical simulations were conducted.

The geometric model of the femur was constructed based on anatomical data [17],[18], and the simulations were carried out using ANSYS software [19]. A mesh consisting of 10-node tetrahedral elements (Tet10) was employed, offering improved accuracy in representing the complex geometry of the bone and in capturing the resulting stress distribution.

III. RESULTS AND DISCUSSION

A. Finite element Mesh

A high-resolution finite element model of the femoral bone was developed, consisting of over 30,000 elements (Fig. 5). To accurately simulate the stress–strain response, 10-node quadratic tetrahedral elements (Tet10) were employed in ANSYS [19] (Fig. 6). The Tet10 is a second-order iso-parametric solid element defined by four corner nodes (I, J, K, L) and six mid-side nodes (M, N, O, P, Q, R). It has four triangular faces, labeled 1–4, Fig. 6. Each face is formed by three corner nodes and three mid-side nodes: face 1: I, J, K with mid-side nodes M, N, O; face 2: I–J–L with mid-side nodes M, P, Q; face 3: I, K, L with mid-side nodes N, P, R; face 4: J, K,L with mid-side nodes O, Q, R.

Its quadratic interpolation functions allow precise representation of curved boundaries and stress distributions, providing superior convergence and accuracy compared with linear tetrahedral elements. Each node carries three translational degrees of freedom in structural analyses.

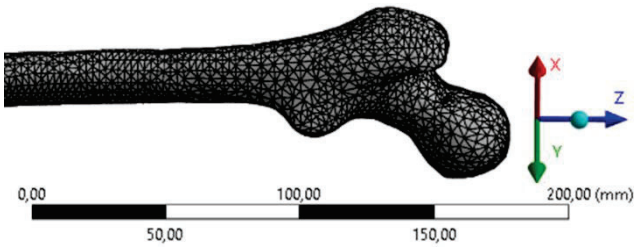


Fig. 5. Finite element model of the bone part.

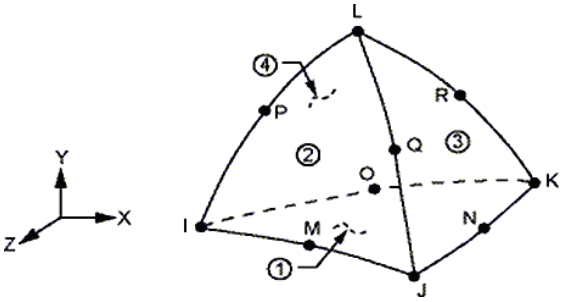


Fig. 6. Ten-node quadratic tetrahedral finite element (Tet10) used in the femoral bone model

Each node of the Tet10 element has three translational degrees of freedom, making it particularly suitable, namely for detailed biomechanical simulations of irregular anatomical structures such as bone. The geometric model was constructed based on average statistical data presented in reference [18]. The use of higher-order Tet10 elements allowed for a more accurate representation of both the femoral geometry and the distribution of internal stresses, outperforming linear tetrahedral elements in terms of solution precision and mesh adaptability. The selected finite element mesh proved to be sufficient for obtaining stable numerical results. Further mesh refinement did not lead to any significant changes in the outcomes.

B. Numerical Results and Discussion

The stress–strain state of the femur was simulated using ANSYS software [29] in 14 incremental loading steps, with the duration and magnitude of each step specified in Table II. To ensure numerical stability and improve convergence, individual load steps were adaptively subdivided into as many as 50 sub-steps when necessary. The evolution of the equivalent von Mises stress in the femur as a function of simulation time is presented in Fig. 7.

From Table II, it can be seen that the period from 0.225 to 0.31 seconds corresponds to the flight phase, during which no load is applied to the femur. Accordingly, in Figure 7, this time period corresponds to the lowest values of equivalent stress.

TABLE II. LOADING PARAMETERS OF A FEMORAL BONE MODEL DURING JUMPING

Loading parameters					
Stages	***Time, sec***	***Force, N***	***Stages***	***Time, sec***	***Force, N***
1	0	275	8	0,325	400
2	0,05	200	9	0,365	355
3	0,1	437,5	10	0,4	325
4	0,15	550	11	0,45	310
5	0,2	0	12	0,5	210
6	0,225	0	13	0,55	260
7	0,31	975	14	0,6	275

In Fig. 7, the black dotted line (labeled 1) corresponds to the minimum equivalent stress, the blue dotted line (labeled 2) represents the maximum equivalent stress, and the yellow dotted line (labeled 3) indicates the average equivalent stress at specific moments in time.

Fig. 7 shows the variation of von Mises equivalent stresses throughout the jumping cycle and provides the basis for identifying the most mechanically significant phases, which are highlighted in Fig. 8.

At 0.1 seconds, the individual is in a preparatory standing position, with relatively low stress levels reflecting static loading. At 0.2 seconds, the person forcefully pushes off the ground, resulting in a pronounced increase in stress due to dynamic muscular action.

Finally, at 0.325 seconds, during landing, a sharp spike in stress levels is observed, indicating peak loading conditions upon ground contact. This phase is critical for evaluating the structural integrity and load-bearing capacity of the femoral bone under sudden loading conditions.

These three time points collectively capture the evolution of stress within the femur throughout the jump cycle.

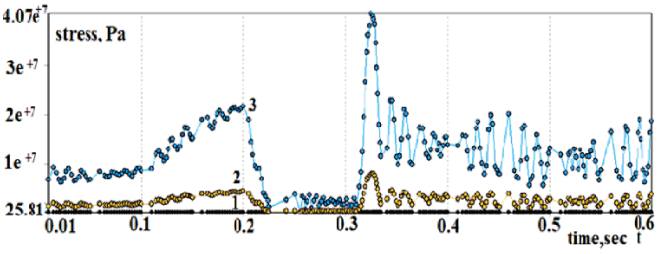


Fig. 7. Dependence of the equivalent von Mises stress on the calculation time

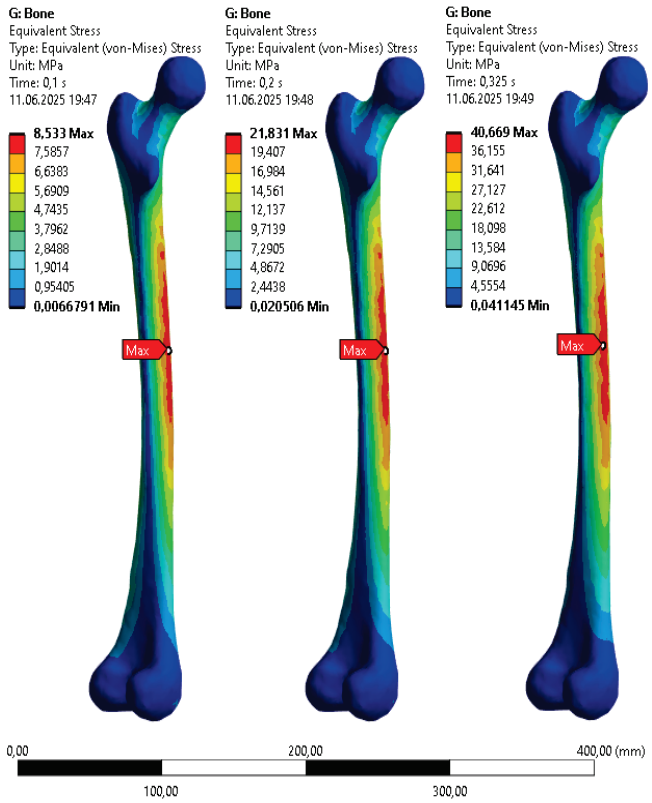


Fig. 8. Equivalent stress in the femoral bone at time points 0.1 s, 0.2 s, and 0.325 s.

In all three cases, the maximum von Mises equivalent stresses remain well below the ultimate strength threshold of 193 MPa, indicating that bone failure is unlikely under the prescribed loading conditions. Nevertheless, the spatial distribution of stresses, while within the safe range for healthy bone tissue, reveals localized patterns that could become critical under altered physiological or pathological states.

For example, in individuals with obesity, age-related reductions in bone density and elasticity, or disease-induced deterioration of bone quality (e.g., osteoporosis), the same loading scenario may produce stress concentrations sufficient to initiate microcracks or even fractures in the femoral diaphysis. These findings highlight the importance of accounting for patient-specific factors when assessing fracture risk and designing preventive or therapeutic strategies

It should be noted that the results of the simulations provided detailed information on the stress distribution, their magnitudes, and the critical regions corresponding to the selected loading conditions, bone geometry, material properties, and boundary constraints. However, since each individual's skeletal anatomy is unique, the computed values may vary significantly for different input parameters. Age-related changes can substantially reduce the mechanical strength of bone tissue.

Moreover, the presence of various diseases, such as osteoporosis or metabolic disorders, may further compromise bone quality and alter its geometry. Body weight, body composition, and the nature of the applied loads also play an important role. Additionally, the presence, size, strength, and conditioning of surrounding muscles significantly influence the biomechanical response of the bone.

These factors are particularly important when analyzing high-impact activities such as jumping, which impose dynamic loads that may exceed the bone's local strength under certain pathological or unfavorable physiological conditions.

In individuals with compromised bone integrity, these loads can lead to elevated stress concentrations, potentially resulting in microcracks or even complete fractures of the femoral diaphysis [20], as shown in Fig. 9.



Fig. 9. Representative fracture pattern of the femur, similar to that reported in [20].

Thus, the above considerations highlight the necessity and importance of incorporating individual-specific parameters into biomechanical analyses, especially when assessing the risk of injury or planning preventive and therapeutic interventions.

Modern diagnostic techniques, such as magnetic resonance imaging (MRI) and ultrasound (US), provide comprehensive and detailed information about the structure of the human skeleton and its individual anatomical characteristics. Moreover, high-resolution imaging combined with advanced data processing enables the construction of accurate three-dimensional models of the human body and its skeletal system.

Such detailed anatomical reconstructions serve as a foundation for various types of computer-based analyses, facilitating applications in multiple fields. For instance, in sports science, they allow for the development of optimized training regimens tailored to an athlete's musculoskeletal structure and physiological limitations. In this context, the authors' research on the hip bone stress-strain response during the jumping cycle demonstrates how finite element modeling of phase-specific loading conditions can provide critical insights into the mechanical behavior of bone under dynamic athletic movements, thereby contributing both to performance optimization and injury prevention.

In clinical practice, these models can inform personalized recommendations for individuals with specific anatomical or functional weaknesses, advising on physical activities that may be beneficial or, conversely, those that should be avoided.

This approach not only improves the functional compatibility of implants but also enhances long-term outcomes by reducing the risk of implant failure or biomechanical mismatch. Additionally, such models significantly enhance the precision of preoperative planning by simulating surgical scenarios, evaluating stress redistribution following intervention, and identifying potential complications before surgery is performed.

Beyond immediate clinical applications, patient-specific computational modeling facilitates the early prediction of health interventions that may become necessary later in life, such as joint replacements, corrective osteotomies, or physiotherapy regimens. In some cases, these simulations can inform proactive, even preventive, surgical decisions in currently asymptomatic individuals, particularly those with congenital deformities, early-stage degenerative changes, or high-risk anatomical features.

A notable example is the case of femoral neck and intertrochanteric fractures, commonly referred to as hip fractures, which are associated with significant morbidity and mortality, especially in older adults. The ultimate goal is to prevent the progression of pathology, avoid traumatic injuries, and reduce long-term healthcare burdens.

Furthermore, patient-specific modeling supports the selection of optimal prosthetic designs by enabling precise adaptation to an individual's anatomical and biomechanical characteristics.

Thus, the integration of advanced diagnostic imaging with high-fidelity computational modeling represents a advanced step in personalized medicine. It supports a shift from reactive treatment to proactive health management, allowing clinicians to make data-driven, individualized decisions that improve

patient safety, accelerate recovery, and enhance overall quality of life. Future developments in this area including the incorporation real-time data and biomechanical feedback are expected to further refine this personalized approach, paving the way for next digital generation in clinical orthopedics and rehabilitation.

IV. Conclusion and Further Research

The stress–strain response of the femur during distinct phases of the jumping cycle was investigated using a finite element framework that integrates phase-specific loading conditions with heterogeneous bone material properties. The computational analysis demonstrated pronounced variations in both the distribution and magnitude of mechanical stresses across the take-off, flight, and landing stages. These results emphasize the necessity of considering temporally dynamic loading scenarios in biomechanical simulations to accurately capture the complex mechanical environments to which skeletal structures are subjected.

The findings highlight the critical role of stage-specific modeling in computational biomechanics, providing valuable insights into the identification of injury-prone regions, the optimization of preventive and rehabilitation strategies, and the design of biomaterials or implants capable of mimicking the adaptive behavior of bone under dynamic physiological conditions. Beyond fundamental understanding, such modeling approaches can also inform the development of personalized diagnostic tools and training regimens aimed at minimizing injury risk in both athletic and clinical populations.

Looking forward, future investigations should explore the potential of advanced fixation systems—including plates, tension-bearing constructs, and next-generation devices fabricated from composites or bioresorbable polymers—for the treatment of femoral fractures. Computational models will serve as a key tool in assessing the mechanical performance, stress-shielding effects, and long-term integration of these systems under physiologically relevant dynamic loads. In this way, simulation-driven research can accelerate the design of next-generation orthopedic implants tailored to individual patient needs and activity profiles, ultimately improving clinical outcomes and ensuring long-term biomechanical compatibility.

Acknowledgment

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The Currents Superposition Principle of High-Voltage Transformerless Resonant Chargers of Capacitive Energy Storage Device

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***Abstract*—The processes in a combined charger with two resonant inverters, which are current sources, are analyzed. The principle of superposition of charging currents of transformerless resonant chargers of a capacitive energy storage device is formulated. Simulation modeling of resonant chargers with different resonant frequencies is carried out. The possibility of increasing the power of high-voltage chargers of capacitive energy storage devices by using the principle of superposition of output currents of transformerless resonant systems with different resonant frequencies has been analyzed.**

Keywords—high-voltage transformer-less chargers, resonant inverters, currents ssuperposition principle

I. INTRODUCTION

The use of high-voltage transformerless resonant chargers (TlRCh) for capacitive energy storage devices (CES) has made it possible to significantly reduce the mass and dimensions of power supplies for pulse-discharge technologies [1, 2]. Currently, the problem of increasing their power arises to meet the requirements of a wide range of high-voltage pulse-discharge technologies. For other applications of transformerless high-voltage systems [3, 4], this problem has not been considered. The power of an electrical system can be increased by increasing the efficiency or coefficient of performance (COP), by increasing the power of a TRCh with one resonant circuit for each phase of the power supply network or by jointly using several such TlRChs. In [5] the possibility of increasing power by increasing the resonant frequency is shown, but this issue is more related to the correct choice of the parameters of the resonant circuit and is not considered in this article. Analytical dependences of the operating frequency in self-generating resonant inverters on the parameters of the resonant circuit and feedback circuits were obtained in order to expand the range of output power and optimize their operation [6]. However, the authors did not consider the possibility of using such systems for high-voltage technologies.

Today, many studies are being conducted to improve the efficiency or coefficient of performance of power converters by reducing transistor switching losses. For example, in [7] a high-efficiency boost converter with zero current switching (ZCS) was developed, which is used in batteries. Research on reducing losses in a resonant inverter is given in [8, 9]. In [10] the performance of a battery charger with a resonant buck converter using soft switching is described. The switching frequency of power semiconductor devices has been increased to reduce losses and switching voltage. As a result, the efficiency and reliability of the buck converter have increased significantly. In [11] the problem of reducing transistor switching losses is considered for wireless power transfer (WPT) technology, which is receiving increasing attention in charging lithium-ion batteries. To ensure charging reliability and improve charging efficiency in a WPT system, a load-independent current and voltage output with an inductive input resistance to implement the zero-voltage switching (ZVS) condition is usually required. The TlRCh already uses a resonant circuit that provides the advantages of switching transistors at zero currents, so it is necessary to look for other ways to increase power.

Many works, starting from the 20th century, are devoted to increasing the capacity by increasing one instance of an electrical system, and this problem can be considered solved. The issue of increasing the capacity by sharing several electrical systems, on the contrary, is very relevant for today, and many studies are devoted to it. In [12], a review of scientific works devoted to creating a network of small power sources operating on multiple loads is provided. Increasing the capacity by using small power sources is not a new idea, which is most fully implemented in the principle of building a Microgrid. But a number of problems arise. In the article [13], a systematic and comprehensive review devoted to the stability of the Microgrid network is provided. Despite all the advantages, microgrids remain difficult to implement without a clear consensus on the causes of failures in their operation. The problems of coordinating the operation of electricity sources in microgrids are also addressed in the articles [14, 15] and many others. The development of multi-cell power systems, microgrids, etc., encourages the development of methods and systems for their control [16], including the development of artificial intelligence-based controllers for inverter generators forming a grid [17]. The application of a control system for joint compensation of voltage imbalance in island AC microgrids with grid inverters is given in [14]. The presence of unbalanced loads in power systems creates a voltage imbalance that disrupts the operation of sensitive loads, such as induction motors, power electronics converters and variable speed drives.

A review of articles on Microgrid technology has shown that the use of a cellular structure of power sources and control systems for changing this structure when changing load parameters is promising. However, the structure of such systems and the tasks of their control are significantly different from the structure and tasks of controlling multi-cell transformerless resonant chargers. Therefore, research in this direction is relevant.

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It should be noted that the complexity of the structure of chargers also requires the development of a complex control, protection and monitoring system. Previously, the main emphasis was placed on the creation of such charging schemes and charger control systems that would provide the specified characteristics of discharge pulses. Issues caused by the stochastic nature of these characteristics were resolved [18-20]. Now the problem of creating small-sized, powerful, energy-efficient resonant chargers of capacitive energy storage devices with controlled output deterministic characteristics without taking into account the stochastic characteristics of discharge pulses is becoming relevant. It should be noted that in the control systems of TlRCh with one resonant circuit, a frequency method of output current regulation has been developed [21], which can be used in combined power systems.

The use of combined power supplies to increase power in systems for electrofiltration of hazardous gas emissions was investigated in [22]. For such systems, a combination of standard equipment with a constant high voltage of up to 50 kV and a pulse current generator with a capacitive energy storage device was used, during the discharge of which short voltage pulses with an amplitude of up to 20 kV were formed. But in such systems, the voltages from each power source were added to the load. To solve the problem of increasing the power of the charger itself, it is necessary to look for ways to increase the charging current. That is, the principles of constructing combined power supplies for adding currents from different sources differ from those given in [22]. In general, the development of the principles of constructing and calculating the elements of resonant inverters differ significantly for each application according to their functional purpose [23].

To increase the output voltage or power of the chargers, a multi-cell structure of the equipment was used, in which the cells have the same structure and functions [24]. The coordination of the functioning of high-voltage resonant chargers with different resonant frequencies was not investigated in the given work. The work considered the creation of chargers of a cell structure based on several inverters with a resonant circuit and a high-voltage transformer connected to each of them. In this case, parasitic parameters of the transformer were used as parameters of the resonant circuit. The idea of the authors was to generally control the transistors of all inverters, but at the same time there was a problem of different natural frequencies of the resonant circuits, caused by the deviation of the parasitic parameters of all transformers from their average value. Because of this deviation, the currents of different cells differed from the calculated ones. To eliminate this problem, it was necessary to correct the deviations of the natural parameters of the transformer cells. This problem disappears in transformerless chargers and with separate control of the inverter transistors of each such device. However, no studies have been conducted to determine the mutual influence of transformerless resonant chargers with different resonant frequencies and the possibility of increasing power by using several such devices. Therefore, research in this direction is relevant.

The research goal of the work is to investigate the possibility of increasing the power of high-voltage chargers of capacitive energy storage devices by using the principle of superposition of output currents of transformerless resonant systems with different resonant frequencies.

II. Mutual Influence of Currents of High-Voltage Resonant Transformerless Chargers of Capacitive Energy Storage Device

Fig. 1 shows simplified equivalent circuits of a charger with one (Fig. 1a) and two (Fig. 1b) resonant inverters. The features of the model shown in Fig. 1a and its output characteristics are given in [25].As shown in [25], a circuit with a series resonant circuit is a current source for the load R. In the case of ideal current sources (with infinite internal resistance), their currents can be summed at the branch point of the electrical network. But for real power sources, the possibility of superposition of currents must be proven by analyzing the processes in the equivalent circuit. To prove the absence of mutual influence of one transformerless resonant charger (TRCh) on the others, we will use an equivalent circuit in which the current source J_1 or J_2 is modeled by a voltage source U_1 or U_2 with an internal resistance X_1 or X_2, significantly greater than the load resistance R (X_1>>R, X_2>>R). Considering that the LC circuit in the resonant inverter circuit (see Fig. 1a) is a voltage divider, then the voltage at its output is $U_1 = \dfrac{E_1 \cdot \dfrac{1}{j\omega C}}{j\omega L - \dfrac{j}{\omega C}}$, and the internal resistance is $X_1 = \dfrac{j\omega L \cdot \dfrac{1}{j\omega C}}{j\omega L - \dfrac{j}{\omega C}}$ (see Fig. 1b). Both of these quantities are infinitely large.

The independence of the current of one current source from others will be observed in the case when the current of this source depends only on its internal resistance and does not depend on the internal resistances of other current sources. Therefore, it is sufficient to analyze the mutual influence of two current sources, as shown in Fig. 1b.

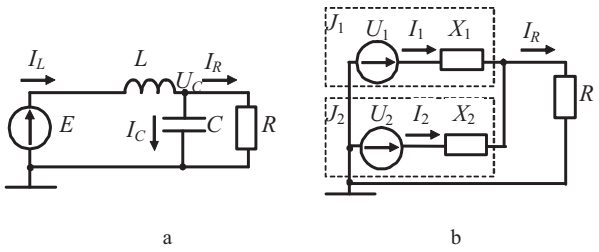


Fig. 1. Equivalent circuit of chargers with one (a) and two (b) resonant inverters, which are current sources J_1, J_2 (X_1>>R, X_2>>R)

The system of equations, which is based on Kirchhoff's first and second laws, has the form

$$\begin{cases} I_1 \cdot X_1 + I_R \cdot R = U_1 \\ I_2 \cdot X_2 + I_R \cdot R = U_2 \\ I_1 + I_2 - I_R = 0 \end{cases}. \quad (1)$$

$$\Delta = \begin{vmatrix} X_1 & 0 & R \\ & X_2 & R \\ 1 & 1 & -1 \end{vmatrix} = -X_1 \cdot X_2 - R \cdot (X_1 + X_2). \quad (2)$$

$$\Delta_{I1} = \begin{vmatrix} U_1 & 0 & R \\ U_2 & X_2 & R \\ 0 & 1 & -1 \end{vmatrix} = -U_1 \cdot X_2 - R \cdot (U_1 - U_2). \quad (3)$$

$$I_1 = \frac{U_1 \cdot X_2 + R \cdot (U_1 - U_2)}{X_1 X_2 + R \cdot (X_1 + X_2)}. \quad (4)$$

According to the scheme shown in Fig. 1b U_1=U_2, and the product of infinitely large quantities X_1X_2 is infinitely greater than their sum, which is why this sum can be removed from equation (4). Then this equation will take the form

$$I_1 = \frac{U_1}{X_1}. \quad (5)$$

According to equation (5), the current of the first power source depends only on its output voltage U_1 and its internal resistance X_1 and does not depend on other parameters of the circuit shown in Fig. 1. Due to the symmetry of this circuit, the current of the first power source does not affect the current of the other power source, and the current of the second power source does not affect the current of the first power source. Theoretical results are obtained for ideal elements of an electric circuit. Expression (5) contains the effective value of the current, but it has an analytical relationship with the average current used to charge the CES.

According to the theorem of the equivalent generator, a circuit of any complexity can be represented by a series connection of a voltage source and internal resistance (see Fig. 1b). It follows that TlRChs can have different resonant frequencies, and the sum of their currents will determine the charging current of the capacitive energy storage device.

III. Increasing the Power of Chargers by Using a Combination of TRChs of Different Resonant Frequencies and Power

Increasing the power of chargers by using a combination of TlRChs of different resonant frequencies and power was analyzed for two charging current sources shown in Fig. 2.

The circuit uses half-bridge voltage inverters (transistors T1, T2 belong to the inverter of the first current source, and transistors T3, T4 to the second), to which rectified voltage is supplied from a bridge single-phase rectifier. The first charging current source has a resonant circuit formed by a series connection of the inductance Lr and the capacitance Cr, the second - Lrr and Crr. The capacitance Cd is part of the rectifier-voltage doubler, on two diodes. CES in Fig. 2 is denoted by the capacitance Cs. The control system supplies voltage to the control inputs of each inverter with a frequency f and a pulse duty cycle D. The frequency and pulse duty cycle can be the same for each inverter or different.

The power of the charger is determined by the expression

$$P = \frac{C_S U_{ch}^2}{2t_{ch}} = \frac{U_{ch} I}{2}, \quad (6)$$

where C_S is the capacity of the CES (Fig. 2), U_{ch} is the voltage to which the CES is charged, t_{ch} is the charging time, $t_{ch} = C_S U_{ch}/I$, I is the average charging current.

To study the combined circuit, simulation modeling was carried out using LTSpice. Fig. 3 shows a model of a combined SP based on two TlRChs. Fig. 3a shows a model of a FPN, at the output of which a constant voltage of 300 V is formed (between the nodes hv and N and also between N and ground). Capacitors C32, C36 - C38 are part of the half-bridge circuit of the inverters. Fig. 3, b shows the transistor parts of the two inverters. The parameter directives show that the switching frequency of transistors S1 and S4 of the first inverter Fosc=19250 (Hz) differs from the switching frequency of transistors S2 and S3 of the second inverter Fosc=20666 (Hz), the pulse filling ratios also differ - D=0.98 for the first inverter and Dx=0.96 for the second inverter. The first inverter has a series resonant circuit (elements L1 and C39), in which the resistance (R1) is set, which takes into account the total active resistance of the resonant circuit, a significant part of which is the active resistance of the inductive choke. The second inverter has a series resonant circuit (elements L2 and C1), in which the resistance (R8) is set, which takes into account the total active resistance of the resonant circuit, a significant part of which is the active resistance of the inductive choke. The ratings of the elements specified in the model have values measured in the elements of the developed experimental sample of the charger.

Fig. 4 show the results of modeling a combined SP based on two TlRChs. Fig. 4a shows the voltage signals for controlling the upper transistors of both inverters, which do not coincide in frequency, phase and have different pulse duty cycles. Fig. 4b shows the charging currents from each source, which also do not coincide in frequency, phase and have different amplitudes. Fig. 4c shows the voltage diagram on the CES, which increases during the charging process and has an exponential character after 20 ms. Up to 20 ms, the voltage diagram has a parabolic character, which is due to the understated voltage at the inverter input between the hv and N nodes, which is shown in Fig. 4d, which is due to the charging process of the half-bridge capacitors (see Fig. 3a). Fig. 4e and Fig. 4f show the diagrams of charging currents from the first and second sources, respectively.

According to the simulation results, a 0.2 μF CES is charged to 40 kV in 0.175 s. All TlRChs have different transistor switching frequencies and pulse duty cycles, so the average values of their output currents are also different. The measured average charging current of the first source is 24 mA, the second - 22 mA. If we calculate the average current for charging such an EEE using the expression $I_з = C_P \cdot U_{Cp}/t_з$, we will also obtain a value of 46 mA, which is the algebraic sum of the charging currents of both current sources. Since the charging time constant of the ENE is much greater than the period of current flow and its resistance is much less than the resistance of the resonant source of charging current, the harmonic components of this current do not affect the operation of the storage device.

According to expression (6), the charging power of the CES is equal to the algebraic sum of the powers of two TlRChs.

Thus, the simulation results show that the TlRChs do not have to be of the same power and do not need to be synchronized; in any case, the sum of their average currents will determine the charging current of the CES.

The absence of mutual influence of the parameters of several resonant inverters on the output currents of the resonant circuits of each inverter is confirmed theoretically for the model shown in Fig. 1b and by the results of simulation modeling of chargers with different resonant frequencies. Despite the fact that the theoretical results were obtained for ideal elements of the electric circuit, the results of the TlRChs modeling taking into account the real parameters coincided with the theoretical ones. Thus, the real parameters of the components do not affect the superposition of currents, which allows using the obtained theoretical results for constructing combined charging devices based on the TlRChs. This justifies the application of the principle of superposition of charging currents of transformerless resonant chargers of capacitive energy storage when using their combination to increase the total power.

The currents superposition principle of high-voltage transformerless resonant chargers of capacitive energy storage device can be formulated as follows: the total average current of the combined capacitive energy storage charger is the sum of the output average rectified currents of transformerless resonant systems with different resonant frequencies.

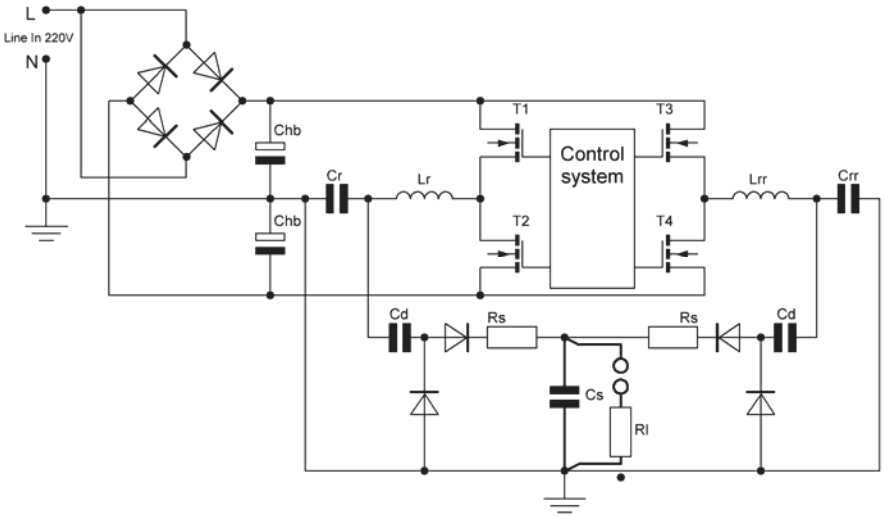
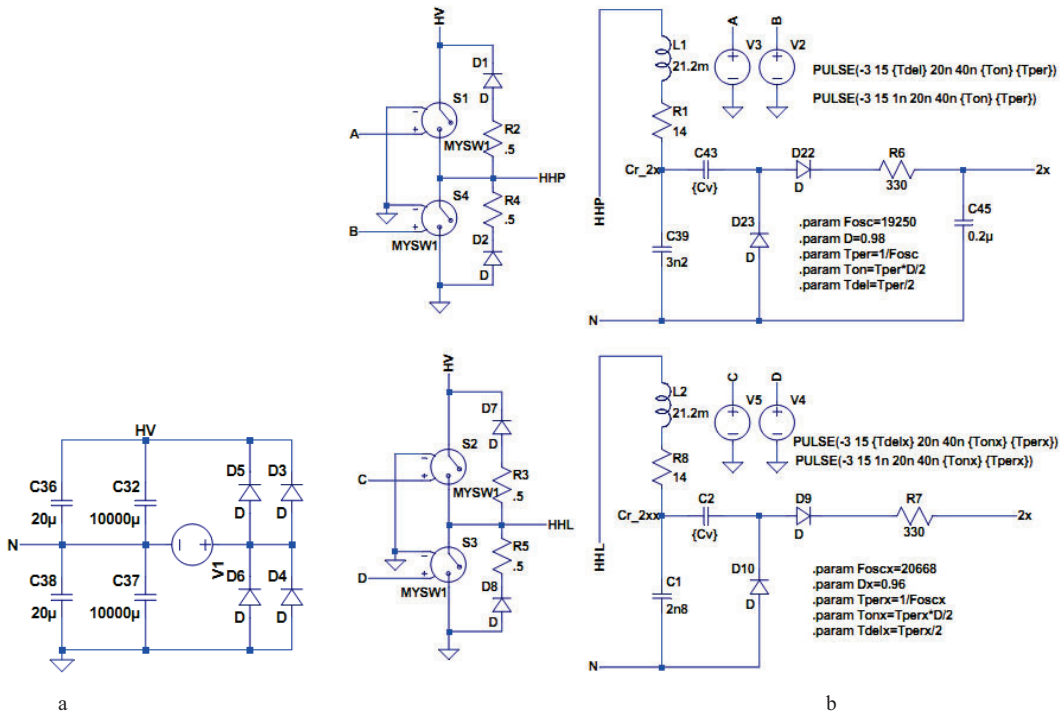


Fig. 2. Combined TlRCh scheme based on two charging current sources



a b

Fig. 3. Model of combined charger based on two TlRChs in LTSpice

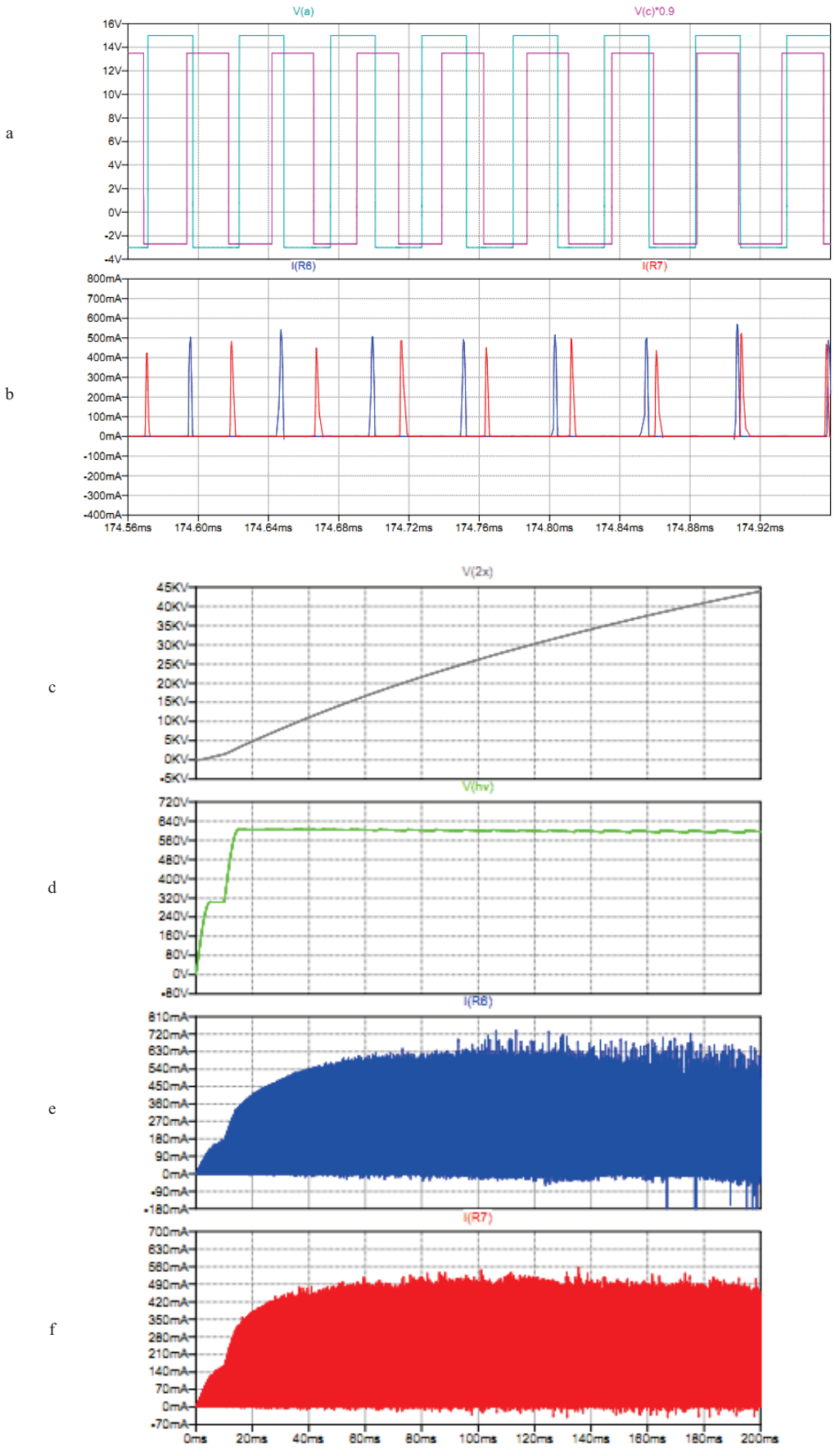


Fig. 4. Simulation results of the combined TlRCh

I. Conclusion

It is theoretically proven that there is no mutual influence of the parameters of several resonant inverters on the output currents of the resonant circuits of each inverter. The theoretical results are confirmed by the results of simulation modeling of resonant chargers with different resonant frequencies.

The currents superposition principle of high-voltage transformerless resonant chargers of capacitive energy storage device is formulated, which consists in the fact that the total average current of the combined charger of capacitive energy storage is the sum of the output average rectified currents of transformerless resonant systems with different resonant frequencies. The possibility of increasing the power of high-voltage chargers of capacitive energy storage by using the principle of superposition of output currents of transformerless resonant systems with different resonant frequencies is substantiated. The power of the combined charger CES is equal to the algebraic sum of the capacities of the TlRChs.

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Wood-Polymer Composite Modification Processes Researching For Achieve High Dielectric Characteristics

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***Abstract*— The purpose of the article is to study the processes of modifying WPC in order to achieve high dielectric characteristics. The article investigates the dielectric properties of wood-polymer composites. Analysis of the obtained results indicates the superiority of the developed WPC samples based on a matrix in the form of secondary polyethylene (25 and 90 wt. %) and thermoplastic polyurethane (10 wt. %) compared to industrial samples. The proposed samples demonstrate the efficiency of the modification process of both the wood filler and the polymer matrix. The advantage of the proposed WPC samples with a modified wood filler compared to industrial samples ranges from 100% depending on the type of industrial WPC and the type of the proposed sample. The efficiency of simultaneous modification of both the wood filler and the polymer matrix of WPC samples based on a matrix in the form of secondary polyethylene (25 and 90 wt. %) and thermoplastic polyurethane (10 wt. %) was also established. The advantage of synergistic modification over the modification of any individual component of the wood-polymer composite is more than five times. From the results of the study of the influence of the modification of the DPC on the level of resistance to external conditions and the level of the dielectric loss tangent tg δ, it is clearly seen that the effective binding of free water in the DPC allows to increase the complex of dielectric characteristics.**

Keywords—composite, wood, recycled raw materials, polyethylene, thermoplastic polyurethane, modification, properties, dielectric loss angle tangent

I. Introduction

In wartime conditions, the problem is not only the deterioration of the environmental situation but also a shortage of raw materials. Therefore, the question arises about using industrial and household waste for the production of composite materials. Scientific advancements, global experience, and market demand indicate the timely implementation of secondary raw materials in production, which addresses both environmental and economic challenges. Polymer composite materials are most commonly used in the construction industry as thermal insulation, structural, finishing, and decorative products. One type of composite material is wood-polymer composites (WPC), based on thermoplastics. The most common representatives of thermoplastics are polyethylene (PE) and polypropylene (PP). Their production volume accounts for more than 35% of the global polymer market. The choice of these polyolefins is due to the simplicity of their processing and modification [1].

Wood-polymer composite (WPC) is one of many types of composite materials [1]. WPCs represent a promising material based on natural raw materials, distinguished from wood by their atmospheric, mechanical, and chemical moisture and water resistance [2,3].

Due to these properties, WPCs are widely used for finishing rooms with high humidity levels and for exterior applications such as terraces and docks [4,5]. The popularity of products made from these materials is driven by important functional, economic and environmental factors [6]. A rather relevant direction in the processes of development and research of modern WPCs is the study of the features of legal regulation of waste polymer matrices and organic fillers [7,8]. It reflects the importance of ensuring effective processes of production and use of WPCs in accordance with the principles of sustainable development [9,10].

The growing range of raw materials for composite production enables the continuous expansion of their applications. At the same time, new types of adverse side effects arise affecting the operational characteristics of products. An essential aspect of studying these challenges is the development of new effective methods for their identification and solutions.

Violating the rules of storage, processing and raw material preparation leads to the fact that raw materials obtain properties complicating the production process and result in degraded operational characteristics of the final product [11].

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The WPC composition typically includes thermoplastic polymers, wood fillers in the form of sawdust, chips, inorganic additives and other modifiers. The synthetic polymers used in the production of wood-polymer composites exhibit significant resistance to biological damage [12,13] This is confirmed by several decades of plastic products operation in harsh conditions (machines, household items, pipes, formwork, construction elements, etc.).

Based on market analysis and the production needs of wood-plastic composites (WPC) in raw materials, the role of using secondary raw material resources is significantly increasing. In the case of production, this refers to the use of secondary polyolefins – PE (polyethylene) and PP (polypropylene). These materials are not restricted by sanitary or technical regulations in the construction industry and possess a range of valuable properties: chemical resistance, considerable heat resistance, frost resistance, strength, and high dielectric characteristics.

However, the structural features that determine these properties also make the material more difficult to use. For example, the non-polarity of polyolefins, on one hand, gives them excellent chemical resistance and dielectric properties, but on the other hand, leads to limited adhesion to various fillers. As a result, the produced composites are characterized by low water resistance and strength. Moisture in the material can be either absorbed during use or residual, remaining in the raw material due to improper storage conditions or errors in the production process.

To determine the moisture content in both the finished product and at intermediate production stages, methods for measuring dielectric permittivity, the tangent of the dielectric loss angle, and electrical resistance are used. Measuring these indicators allows timely detection of moisture presence in the polymer matrix and wood filler, which is essential for adjusting the WPC production process.

Additionally, water absorption and desorption of WPC products during use lead to changes in linear dimensions and weight. The material shrinks or swells as the wood filler dries or absorbs moisture. This lack of dimensional stability is another disadvantage of these materials. Repeated swelling and shrinkage of the wood filler particles lead to a breakdown in adhesion with the polymer matrix, which results in the formation of microcracks in the product. Because of this, various additives are introduced into WPC compositions. In the finished WPC product, the porous structure of the polymer matrix can absorb moisture and transfer it to the wood filler.

Therefore, modifying the polymer matrix is crucial to prevent water absorption and to enhance the physical, mechanical, and dielectric properties. The presence of moisture in WPC – whether residual or acquired during use – increases their tendency to accumulate static electricity and leads to the loss of insulating properties.

The use of the synergistic modification method—namely, the simultaneous modification of both the polymer matrix and the wood filler—is a promising and effective approach to preventing water absorption and achieving high dielectric characteristics in WPC products.

Thus, the aim of this study is to examine the modification processes of WPC in order to achieve high dielectric performance.

II. MATERIALS AND METHODS

The research objects included: recycled polyethylene, recycled thermoplastic polyurethane, expanded polystyrene, petroleum-polymer resin, particleboard waste, microcalcite and gasoline.

The preparation of WPC was carried out as follows: a solvent (gasoline) was added to the petroleum-polymer resin and thoroughly mixed until a homogeneous solution (intermediate modifying mixture) was obtained, which took 10-15 minutes.

The next step involved adding expanded polystyrene waste, dissolved in the intermediate modifying mixture through continuous stirring for 10 minutes, resulting in the final modifying mixture. The obtained modifying mixture was then combined with the wood filler for impregnation and mixed for 5-10 minutes until complete homogenization. Finally, microcalcite was added to the impregnated and homogenized wood filler and mixed for 5 minutes.

The homogenized modifying mixture was dried in a laboratory dryer for 1 hour at a temperature of 70°C or left to air dry at room temperature for 12 hours.

The use of hard-to-recycle waste materials such as expanded polystyrene and thermoplastic polyurethane (TPU) has been proposed for modifying the wood filler and the polymer matrix. This solution not only helps address environmental issues but also improves the performance and physicochemical properties of wood-polymer composite (WPC) products. Thermoplastic polyurethane is used to reduce the porosity of the polymer matrix structure, while polystyrene waste is used to enhance the insulating properties.

During the impregnation stage with the liquid phase and subsequent drying, 12–1% of the excess solvent evaporated. The dried filler was then mixed with recycled polymer raw materials (polyethylene (PE) and a combination of polyethylene and thermoplastic polyurethane (PE + TPU)) and processed into WPC using the extrusion method. The extrusion was carried out on an extruder with a length-to-diameter (L/D) ratio of 40, a screw diameter of 30 mm, and a rotation speed of 90 rpm. The temperature was maintained within the range of 190–195°C throughout the entire process.

Water absorption was measured following ASTM-D-570, while wear resistance and shrinkage were assessed in accordance with DIN 53 516. Dimensional changes due to heating were determined according to ISO 11501:1995. The dielectric constant is determined according to DIN 53458.

FTIR spectra were recorded on a Nicolet 380 IR spectrophotometer (USA) at 20°C–25°C in the frequency range 4000–500 cm^{-1}.

III. RESULTS AND DISCUSSION

The composition of the developed WPC with a modified polymer matrix is presented in Tables 1-2. The analysis of the data in Table 3 indicates the superiority of the proposed samples compared to the industrial samples [20]. The proposed samples demonstrate the effectiveness of the modification process both for the wood filler and the polymer matrix. The advantage of the proposed WPC samples with modified wood filler compared to the industrial samples ranges from 100%, depending on the type of industrial WPC and the type of the proposed sample. The water absorption

indicators of WPC with modified wood fillers are presented in Table 3.

TABLE I. THE COMPOSITION OF WPC WITH MODIFIED WOOD FILLER: SAMPLE1-8

Composition	1	2	3	4	5	6	7	8
Recycled PE, %	25	-	25	-	25	25	25	25
Recycled PP (Polypropylene),%	-	25	-	25	-	-	-	-
Petroleum-based polymer resin, g	20	20	10	10	10	5	5	5
Solvent, g	25	25	20	20	20	15	15	15
Varnish PF-170, g	-	-	-	-	-	-	-	-
Polystyrene waste, g	-	-	5	5	10	10	10	10
Furniture production waste, g	30	30	30	30	30	20	20	20
Wood waste mixture, g	-	-	-	-	-	-	-	-
Micro-calcite, g	-	-	-	-	-	-	-	-

TABLE II. THE COMPOSITION OF WPC WITH MODIFIED WOOD FILLER:SAMPLE 9-13

Composition	1	2	3	4	5
Recycled PE, %	9	10	11	12	13
Recycled PP (Polypropylene),%	25	25	25	25	25
Petroleum-based polymer resin, g	-	-	-	-	-
Solvent, g	5	5	5	5	5
Varnish PF-170, g	15	15	15	15	15
Polystyrene waste, g	-	-	15	15	20
Furniture production waste, g	10	8	-	-	-
Wood waste mixture, g	-	-	40	30	35
Micro-calcite, g	35	30	-	-	-

TABLE III. THE WATER ABSORPTION INDICATORS OF WPC WITH MODIFIED WOOD FILLERS

Composite	Water absorption, W, %	Wear resistance, m	Wear resistance, V	Dimensional change in the range of +30°C to +80°C	Dimensional change at -15°C
1	+5,7	0,1015	0,2997	Not detected	Not detected
2	+5,8	0,1336	0,320	Not detected	Not detected
3	+6,4	0,112	0,304	Not detected	Not detected
4	+6,6	0,088	0,305	Not detected	Not detected
5	+6,5	0,108	0,308	Not detected	Not detected
6	+6,2	0,142	0,288	Not detected	Not detected
7	+5,4	0,113	0,283	Not detected	Not detected
8	+5,3	0.101	0,270	Not detected	Not detected
9	+5,3	0.102	0,325	Not detected	Not detected
10	+6,3	0,103	0,278	Not detected	Not detected
11	+6,1	0.052	0,325	Not detected	Not detected
12	+5,9	0.072	0,327	Not detected	Not detected
13	+6,2	0.075	0,325	Not detected	Not detected

The process of modification of the polymer matrix and wood filler was analyzed using infrared spectroscopy, the results of which are presented in Fig. 1-2. From the results of IR spectroscopy, it can be concluded that the most significant effect of filler modification is associated with a decrease in the number of free hydroxyl groups, which is clearly visible with changes in the wavelength region 3000-3700 cm^{-1}: the large band with a maximum at 3331 cm^{-1} for the unmodified filler becomes much less intense and smaller in area at a maximum at 3341cm^{-1} upon modification. When comparing the IR spectra of the original polymer matrix and the modified WPCs, the same trend can be seen.

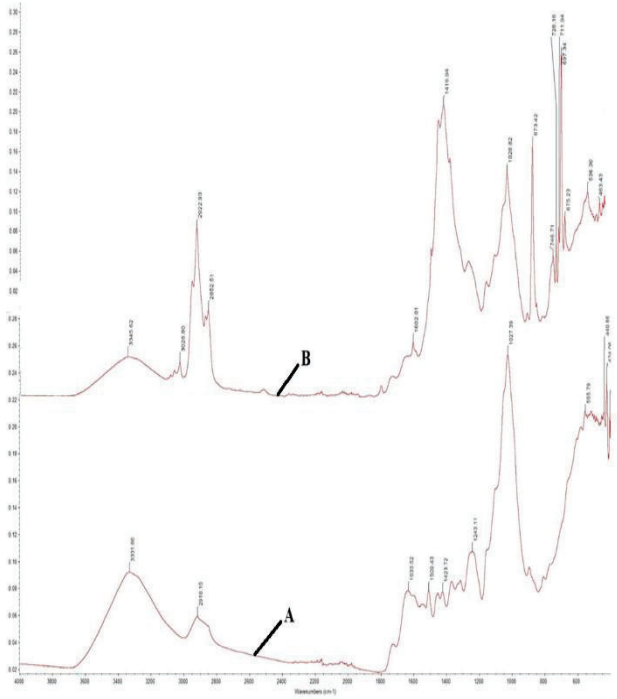


Fig. 1. IR spectrum of unmodified wood filler (A) and modified filler (B).

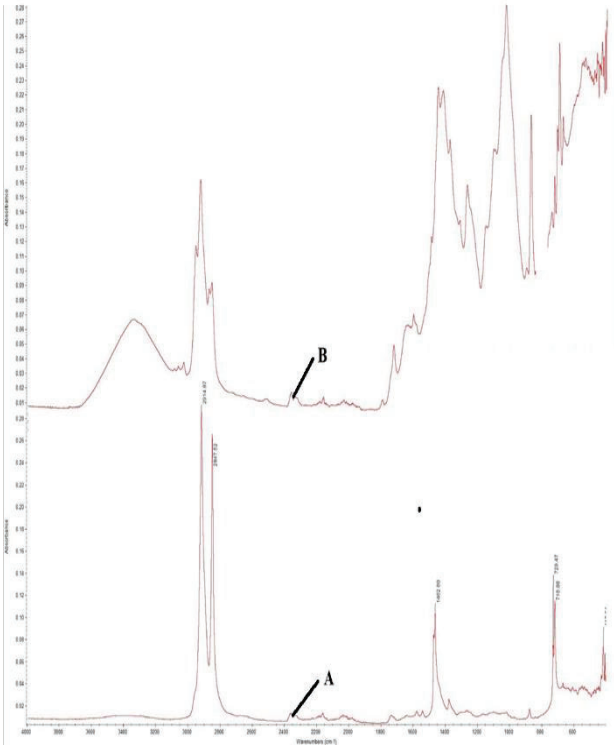


Fig. 2. IR spectrum of the modified polymer matrix (A) and the WPC sample prepared by the synergistic modification method (B).

The effectiveness of the method of synergistic modification of the WPC is confirmed by tests on the level of dielectric characteristics of the obtained samples, which is shown in Fig. 3.

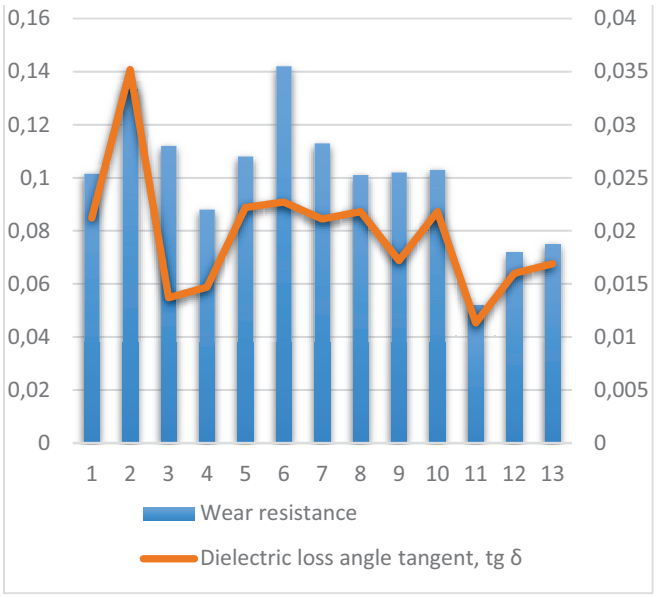


Fig. 3. IR spectrum of the modified polymer matrix (A) and the WPC sample prepared by the synergistic modification method (B).

From the results of the study of the influence of WPC modification on the level of resistance to external conditions and the level of dielectric loss tangent tg δ, it is clearly seen that the effective binding of free water in the WPC allows to increase the complex of dielectric characteristics.

IV. Conclusions

Thus, the analysis of accumulated scientific and industrial experience indicates the advantages of the proposed method of synergistic modification in the production of wood-polymer composites. Analysis of the obtained results indicates the superiority of the developed WPC samples based on a matrix in the form of secondary polyethylene (25 and 90 wt. %) and thermoplastic polyurethane (10 wt. %) compared to industrial samples. The proposed samples demonstrate the efficiency of the modification process of both the wood filler and the polymer matrix. The advantage of the proposed WPC samples with a modified wood filler compared to industrial samples ranges from 100% depending on the type of industrial WPC and the type of the proposed sample. The efficiency of simultaneous modification of both the wood filler and the polymer matrix of WPC samples based on a matrix in the form of secondary polyethylene (25 and 90 wt. %) and thermoplastic polyurethane (10 wt. %) was also established. The advantage of synergistic modification over the modification of any individual component of the wood-polymer composite is more than five times. From the results of the study of the influence of the modification of the DPC on the level of resistance to external conditions and the level of the dielectric loss tangent tg δ, it is clearly seen that the effective binding of free water in the DPC allows to increase the complex of dielectric characteristics.

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Study of Optical Properties of CdTe Thin Films Obtained by Electrochemical Deposition

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***Abstract* - The aim of the work is to study the optical properties of CdTe semiconductor films obtained by electrochemical deposition. The relationship between the electrophysical characteristics of the formed CdTe films and the electrolysis parameters was established. For this purpose, the most appropriate composition of the electrolyte for obtaining CdTe films was determined. A basic technological scheme of the process of obtaining electrochemical formation of semiconductor optically active CdTe films was developed, and the technological modes of individual stages were determined. The kinetic parameters of electrochemical reactions were determined by the method of voltammetry. The analysis of electrolytes for the content of components was determined by titrometric and weight methods. The morphology of the resulting precipitates was studied by optical microscopy. Optical parameters of the obtained experimental samples were studied by the optical spectrophotometric method. Band gap, refractive index, static and high-frequency dielectric constants of deposited CdTe films were calculated by the analytical processing of experimental data.**

Keywords – thin film; electrolyte; bandgap; optical parameters.

I. Introduction

Energy security represents one of the most critical challenges confronting humanity in the 21st century. The resolution of numerous global issues is unattainable without a stable and sustainable energy supply.

Industrial advancement is intrinsically linked to the continuous growth of energy consumption. According to the U.S. Department of Energy, the current global generating capacity stands at approximately 18 TW, which equates to an annual energy consumption of around 5×10^{20} J, or 100 TW-hours. The predominant sources of this energy, aside from nuclear power, are conventional fossil fuels such as coal, oil, and natural gas. However, these resources are finite, and their extraction and utilization cause irreversible environmental degradation, contributing significantly to the risk of catastrophic climate change.

Given these limitations, alternative energy generation methods must be pursued. Although approximately 450 nuclear reactors are in operation worldwide, providing a combined capacity of roughly 370 GW, scaling nuclear energy to meet the 18 TW demand presents substantial challenges. Other renewable sources - such as hydroelectric, geothermal, wind, and bioenergy - are similarly constrained by geographic and technological limitations. In contrast, the power of solar radiation reaching Earth's surface exceeds current global energy demands by over three orders of magnitude. Effectively harnessing this abundant and environmentally benign energy source remains a key objective.

The advancement of photovoltaic technologies necessitates the development of cost-effective semiconductor materials for solar cell structures, even at the expense of some efficiency loss [1-5].

Among the materials under investigation, chalcogenides - particularly thin films composed of amorphous silicon, cadmium chalcogenides (CdTe, CdSe, CdS), and selenides doped with indium, copper, or gallium - are of considerable interest [6-7]. While amorphous silicon-based thin-film solar cells offer low production costs, they suffer from reduced efficiency [8]. In contrast, heterostructures such as CdS/CdTe, comprising polycrystalline thin films, provide enhanced photovoltaic performance.

Materials like polycrystalline Si, CdTe, CdSe, $CuInSe_2$, ZnSe, and ZnTe are promising candidates for the cost-effective conversion of solar and thermal energy into electricity [9-13].

Among these, polycrystalline CdTe thin films stand out due to their applicability in photovoltaic devices with p–n junctions or Schottky barriers [14].

Owing to its direct bandgap and favorable optoelectronic properties, CdTe is highly suitable for large-scale fabrication of solar modules, high-energy radiation detectors, telecommunications components, and various solid-state electronic devices [15,16].

To date, there is a lack of comprehensive studies in the literature addressing the influence of electrolyte pH on the properties of CdTe thin films deposited from weak cadmium acetate solutions in two-electrode potentiostatic configurations [17,18].

This research seeks to fill that gap by examining how electrolyte pH affects the optical properties of electrodeposited CdTe films. The optical parameters of CdTe films deposited on glass/ITO substrates under varying electrolyte compositions and electrochemical conditions were analyzed using spectrophotometric techniques.

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II. Electrochemical Deposition of CdTe Thin Film

One of the critical parameters in the electrodeposition of high-quality photoconductive materials is the pH of the electrolyte solution. The control of this variable is essential to ensure optimal film growth and material properties.

CdTe thin films can be synthesized from aqueous or non-aqueous solutions using either two- or three-electrode configurations. These systems may operate in potentiostatic (constant voltage) or galvanostatic (constant current) modes. The two-electrode setup offers several advantages, including the elimination of contamination from the reference electrode in the event of leakage or fracture, reduced equipment costs due to the absence of an expensive reference electrode, and enhanced deposition rates.

A schematic diagram of the typical two-electrode electrochemical deposition setup is presented in Fig. 1.

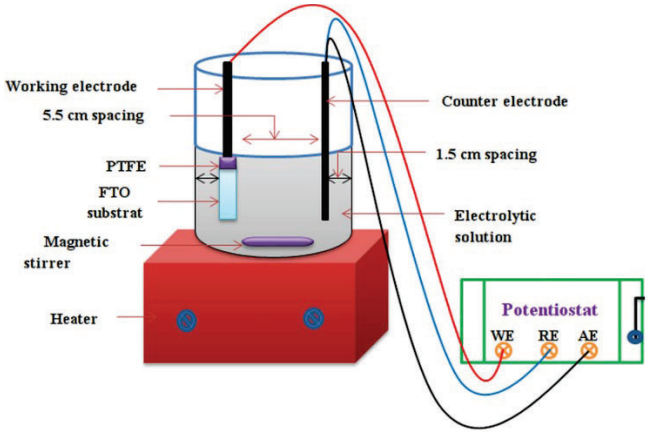


Fig. 1. Schematic design of typical two-electrode electrodeposition system.

The deposition of CdTe films was conducted on glass substrates coated with indium tin oxide (ITO), utilizing a two-electrode system to identify the optimal pH conditions for CdTe thin-film formation suitable for photovoltaic applications. The complete experimental configuration is shown in Fig. 2.

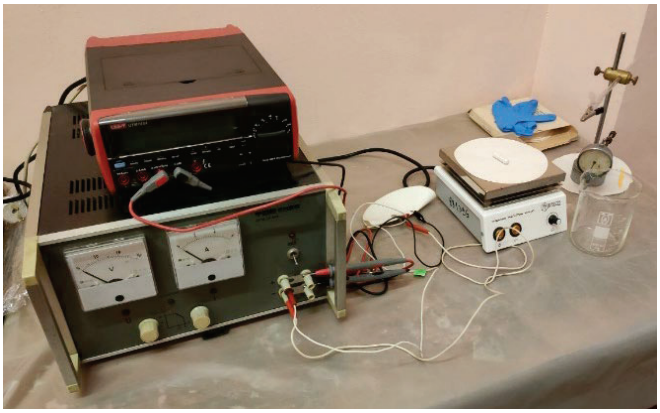


Fig. 2. General view of experimental two-electrode electrodeposition system.

The electrodeposition process was powered by a TES 5010 direct current source. The ITO-coated glass served as the working (cathode) electrode, while a palladium (Pd) plate functioned as the counter electrode. A constant deposition voltage ranging from 1.5 to 1.6 V was applied throughout the process.

Prior to CdTe film growth, the cadmium acetate solution underwent electrochemical pre-treatment at a voltage below the cadmium deposition potential for a duration of 48 hours, as determined by cyclic voltammetry. Given the relatively low purity of the cadmium source, it was essential to remove metallic impurity ions from the electrolyte solution.

CdTe thin films were deposited onto 2.5×2.5 cm² glass/ITO substrates with a surface resistivity of 8 Ω/sq and an ITO coating thickness of approximately 135 nm. Before deposition, the substrates were cleaned ultrasonically in a laboratory detergent solution using deionized water for 30 minutes, followed by sequential rinses in ethanol, acetone, and methanol. The cleaned substrates were finally rinsed with deionized water and dried in ambient conditions.

To ensure appropriate deposition conditions, the pH of the electrolyte was adjusted to 2.0 using sulfuric acid (H_2SO_4) at room temperature. After deposition, the substrates were rinsed with deionized water and air-dried.

During film formation, the electrolyte was continuously stirred at a moderate speed using a magnetic stirrer, and the solution temperature was maintained at 70 °C.

The tellurium-containing solution was prepared by dissolving 2 g of TeO_2 in 30 ml of diluted sulfuric acid under stirring for two hours, due to TeO_2 poor solubility in water and enhanced solubility in acidic media. A volume of 1 ml of this Te solution was then added to 400 ml of the cadmium-containing electrolyte, followed by additional stirring for 300 minutes to ensure homogeneity.

The deposition conditions, electrolyte compositions, applied voltages, and deposition durations for the CdTe samples are summarized in Table 1.

TABLE I. Composition of Electrolytes and Electrodeposition Regimes of CdTe Films

Sample	Electrolyte composition	Voltage, V	Time, min
10	0.2 M $CdSO_4$ 1mM TeO_2	1.5 1.6	10 30
28	0.2 M $CdSO_4$ 1mM TeO_2	1.5	60
27	0.1 M $CdSO_4$ 1mM TeO_2	1.5	60
19	0.2 M $CdSO_4$ 1mM TeO_2 0.02 M $Na_3C_6H_5O_7$	1.5	60

III. Experimental Research

In order to evaluate the surface morphology, grain structure, and uniformity of coverage of the CdTe thin films deposited on glass/ITO substrates, optical microscopy was conducted at 40× magnification with a resolution of 10 µm. Representative micrographs are presented in Fig. 3.

The optical characteristics of the deposited films were analyzed within the spectral range of 400–1100 nm using an SF-2000 spectrophotometer equipped with an SFO-2000 module for measuring both specular and diffuse reflectance.

Ethanol-treated substrates were used as reference samples during transmission spectrum acquisition, denoted as $T_0(\lambda)$. The optical bandgap energy E_g of the CdTe thin films was derived from their absorption coefficient α, calculated using the following expression [17,19]:

$$\alpha = ln(1/T_0)/t, \quad (1)$$

The values of the optical band gap width E_g were obtained from proposed equation in the study [18-24]:

$$(\alpha * h\nu)^2 = A * (h\nu - E_g), \quad (2)$$

where hν is the photon energy and A is a material-dependent constant assumed to remain relatively constant in the optical range. The bandgap energy E_g was obtained graphically by extrapolating the linear portion of the $(\alpha*h\nu)^2$ versus hν plot to its intercept with the energy axis [25].

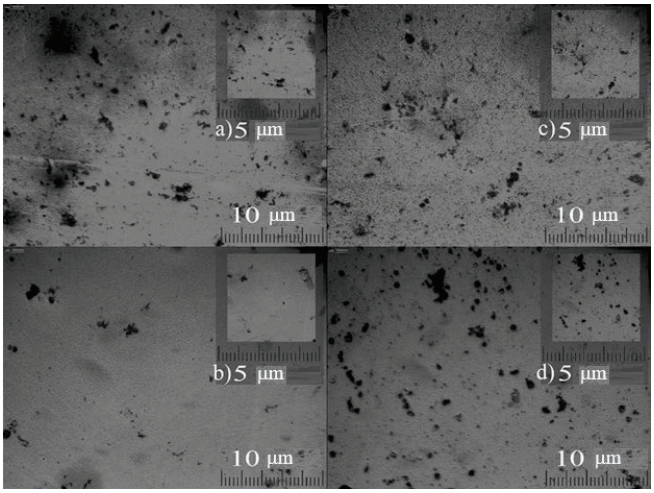


Fig. 3. Top view image of CdTe films deposited on glass/ITO using electrodeposition with different electrochemical cell conditions and electrolyte composition for different samples: a) 28, b) 19, c) 27, d) 10.

Urbach energy (E_0), indicative of optical transitions involving sub-bandgap states [17,19], was determined using the following relation:

$$a = \alpha_0 * e^{h\nu/E_0}, \quad (3)$$

where α_0 is a constant representing the absorption coefficient at the onset of absorption. In other words, α_0 is the coordinate of the convergence point of the Urbach "tail", i.e. the case when hν = 0 and α = α_0.

The Urbach energy, reflecting the degree of structural disorder, was calculated by fitting the linear region of ln(α) versus hν, with the slope yielding the inverse of E_0.

The calculated values of E_g and E_0 for various samples are summarized in Table 2.

TABLE II. BAND GAP AND URBACH ENERGY OF CDTE FILMS

Sample	Electrolyte composition	Band gap E_g, eV	Urbach energy E_0, eV
10	0.2 M $CdSO_4$ 1mM TeO_2	1.23	0.86
28	0.2 M $CdSO_4$ 1mM TeO_2	1.48	1
27	0.1 M $CdSO_4$ 1mM TeO_2	1.25	0.91
19	0.2 M $CdSO_4$ 1mM TeO_2 0.02M $Na_3C_6H_5O_7$	1.5	0.72

Additionally, the refractive index (n) of the CdTe thin films was estimated using the empirical Moss relation [26,27]:

$$n^4 = C/E_g, \quad (8)$$

where C is a constant with a value of 95 eV [28].

The static dielectric constant ε_0 was derived from the bandgap using the following empirical relation [29]:

$$\varepsilon_0 = 18.52 - 3.08E_g, \quad (9)$$

The high-frequency dielectric constant ε_∞ was determined from the refractive index using,

$$\varepsilon_\infty = n^2, \quad (10)$$

The resulting values of E_g, n, ε_0, and ε_∞ for all CdTe samples are presented in Table 3. These results are consistent with previous findings reported in the literature [28].

TABLE III. BAND GAP, REFRACTIVE INDEX, STATIC AND HIGH-FREQUENCY DIELECTRIC CONSTANTS OF DEPOSITED CDTE FILMS

Sample	Band gap E_g, eV	Refractive index, n	static dielectric constant ε_0	high-frequency dielectric constant ε_∞
10	1.23	2.964	14.73	8.788
28	1.48	2.831	13.96	8.012
27	1.25	2.953	14.67	8.718
19	1.50	2.821	13.90	7.958

Fig. 4 shows the optical properties of CdTe films obtained by electrodeposition on glass/ITO substrates.

The reflectance spectra (Fig. 4c) confirmed significant reflection in the visible range, while haze factor analysis (Fig. 4d) indicated that the reflection is predominantly diffuse due to pronounced light scattering.

All CdTe films exhibited partial transparency in the visible range, with the maximum transmittance T_0 varying between 3.1% and 21%, depending on deposition parameters (Fig. 4a).

The optical bandgap for samples 28 and 19 was calculated to be approximately 1.48 eV and 1.50 eV, respectively - values that closely match the bandgap of bulk CdTe, thereby validating the reliability of the fabrication process [41].

In contrast, samples 10 and 27 demonstrated lower bandgap values of 1.23 eV and 1.25 eV, respectively, which may be attributed to an excess of Cd and a deficiency of Te in the films. This compositional imbalance is a common issue in electrodeposited CdTe films, especially when processed at pH values below 3, which tend to promote Te enrichment due to cadmium leaching from the film surface [39].

Variations in bandgap energy can also result from the removal of microstructural impurities and voids during deposition, as well as differences in electrolyte formulation. These impurities, such as excess unreacted Te and Cd vacancies, alter the stoichiometry and optical properties of the films [30,31,39,40].

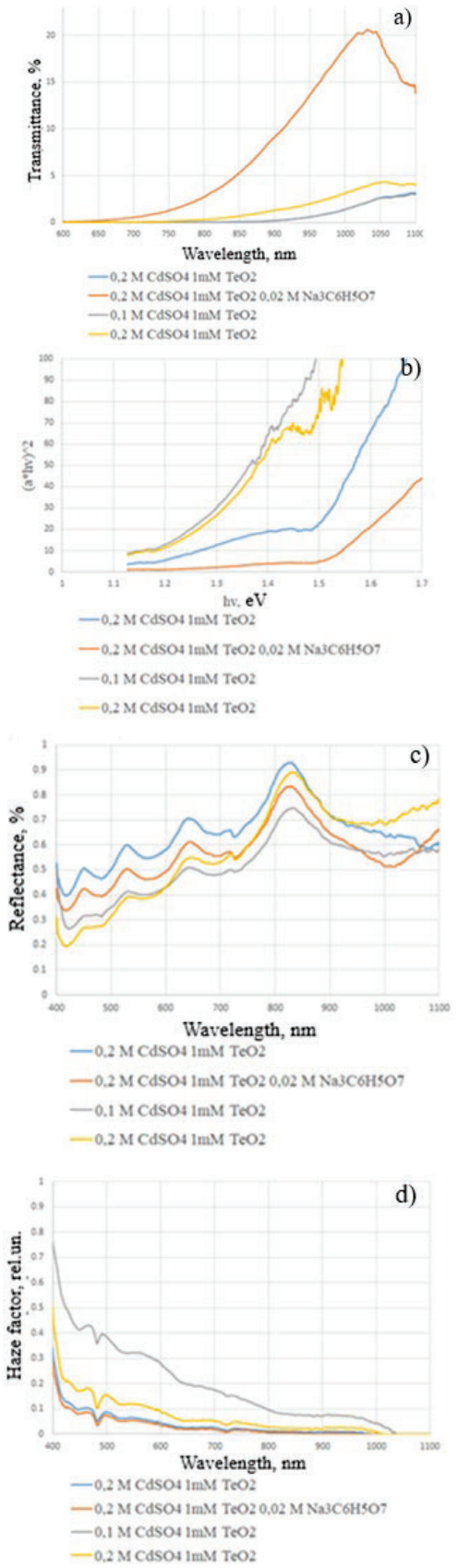


Fig. 4. Optical properties of CdTe films with different electrolyte compositions

To examine the correlation between electrolyte composition and Cd/Te atomic ratios in the films, statistical analysis was performed using Minitab software. The atomic ratios for samples 28 and 19 were found to be 0.96 and 0.95, respectively, indicating near-stoichiometric compositions consistent with bulk CdTe.

The average Urbach energy across all samples was determined to be $E_0 \approx 0.87 \pm 0.12$ eV, suggesting considerable lattice disorder. These findings indicate the necessity of conducting further structural analysis, such as X-ray diffraction (XRD), to better characterize the microstructural quality and internal stress of the films.

Due to the inherent limitations of the deposition technique, the resulting films exhibit relatively low crystallinity. Consequently, many CdTe synthesis protocols incorporate post-deposition thermal annealing to promote recrystallization and enhance polycrystalline film quality [42–44]. However, in this investigation, the focus was limited to evaluating the as-deposited state of the samples.

IV. Conclusion

In this study, CdTe thin films were synthesized using a two-electrode electrochemical deposition technique from a cadmium acetate-based electrolyte under varying chemical compositions and deposition conditions. The structural and optical characteristics of the as-deposited films were systematically investigated.

To enhance the efficiency of the electrodeposition process, cadmium acetate was utilized as the cadmium source, and a simplified two-electrode setup was implemented. Optical analysis revealed that the films exhibited notable reflectance primarily in the visible spectrum. The optical absorption and bandgap energy of the films were found to depend on deposition time and the specific conditions of the electrochemical cell. An increase in deposition time correlated with a broader bandgap in the as-deposited samples.

Microscopic imaging confirmed that the grain morphology, size distribution, and surface uniformity of the CdTe films were influenced by the electrolyte composition and deposition parameters. Furthermore, elemental analysis revealed that samples 28 and 19 possessed atomic Cd/Te ratios of approximately 0.96 and 0.95, respectively, indicating near-stoichiometric compositions comparable to bulk CdTe.

The optical bandgap values of these samples ranged from 1.48 to 1.50 eV, which are consistent with those reported for bulk CdTe in existing literature [34]. This supports the reliability of the deposition approach for fabricating optically active semiconducting films.

The average Urbach energy of CdTe films deposited by electrochemical deposition was: $E_O \approx 0.87 \pm 0.12$ eV. The Urbach energy level indicates rather large values of the crystal lattice mismatch, the level of microstresses.

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Changes in Microcrystalline Properties of CdTe Induced by Chloride Treatment

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***Abstract*- Thin films of cadmium telluride (CdTe) with metastable hexagonal structure were deposited by DC magnetron sputtering on glass substrates and subsequently subjected to $CdCl_2$-mediated chloride treatment followed by thermal annealing at ~400 °C. X-ray diffraction (XRD) and scanning electron microscopy (SEM) analyses were performed to investigate the chloride treatment's impact on film microstructure and microstrain. The crystallite size, calculated using Scherrer's formula, increased by approximately 40% after treatment, indicative of recrystallization, a result corroborated by size–strain plot (SSP) fitting. Williamson–Hall analysis under UDM, USDM, and UDEDM models revealed significant reductions in microstrain (from ~0.0031 to ~0.0007), mechanical stress (from ~0.20 GPa to ≈0.002 GPa), and strain energy density (~3136 to ~17 kJ/m³), demonstrating defect passivation and improved structural ordering. SEM imaging confirmed enhanced grain coalescence and surface uniformity. Our findings align with established reports on $CdCl_2$-driven recrystallization and defect passivation in CdTe thin films. The chloride treatment thus proves effective in optimizing crystallographic and mechanical properties of CdTe films, with implications for improved performance in photovoltaic and electronic applications.**

Keywords - Cadmium telluride (CdTe); thin films; $CdCl_2$ treatment; recrystallization; X-ray diffraction (XRD); scanning electron microscopy (SEM); microstrain; Williamson–Hall analysis; size–strain plot; thin-film solar cells.

I. Introduction

Cadmium telluride (CdTe) is one of the most important semiconductor materials, especially in the field of solar energy. Its unique physicochemical properties-such as a wide band gap (~1.5 eV) and high absorption efficiency of visible light-make it ideal for converting solar radiation into electrical energy. The band gap of CdTe allows it to absorb a significant portion of the solar spectrum even at low material thicknesses, making CdTe a cost-effective and energy-efficient option for thin-film solar cells. In addition, CdTe demonstrates high operational stability, which is crucial for the long-term use of photovoltaic panels in outdoor environments [1-3].

However, like most semiconductors, CdTe also has certain drawbacks, primarily related to the presence of defects in its crystalline structure. Defects such as dislocations, vacancies, and impurities create traps for charge carriers, leading to their recombination without generating an electrical current. This significantly reduces the material's efficiency. Furthermore, these defects influence the electrical conductivity of CdTe, limiting its potential in converting solar energy into electricity. These shortcomings can be addressed through chloride treatment - a specialized technological process that substantially improves the structural quality of CdTe [4].

Chloride treatment, typically involving cadmium chloride ($CdCl_2$), is a key step in the fabrication of CdTe-based solar panels. The process involves the deposition of a thin $CdCl_2$ layer on the CdTe surface followed by thermal annealing at elevated temperatures. Under these conditions, chlorine atoms actively diffuse into the CdTe crystal lattice, interacting with intrinsic defects and promoting their passivation. Defect passivation reduces recombination losses and enhances material stability, thereby increasing the efficiency of CdTe in photovoltaic applications.

The aim of this study is to conduct a comprehensive analysis of the microcrystalline structural changes in CdTe films after chloride treatment. In particular, the work investigates the effects of chlorine diffusion on grain size and morphology, microstrain levels, and material stability. The review outlines the key stages of chloride processing, diffusion mechanisms, and their influence on the final structural and elastic properties of CdTe.

This work also explores the role of microcrystalline structure in achieving high energy conversion efficiency and long-term stability in CdTe-based solar cells. Special attention is given to the correlation between the physical processes occurring [5] during chloride treatment and the resulting material characteristics that determine its performance and reliability.

II. OBTAINING RESEARCH SAMPLES

High-purity cadmium telluride (CdTe) powder was used to produce thin CdTe films. The powder was pressed into a dense target and annealed in vacuum at a temperature of 60–80 °C for three hours. High-purity cadmium telluride (CdTe) in powder form was used as the main material for target fabrication. Cadmium telluride was chosen due to its ability to provide stable and reproducible properties in thin film form. Soda-lime glass substrates were used for film deposition. The choice of such substrates was due to their chemical resistance, high thermal stability, and ability to ensure uniform material sputtering [1].

The vacuum DC magnetron sputtering system VUP-5M was used to deposit the films, providing control of parameters such as sputtering power, deposition time, and the distance

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between the target and the substrate [3]. The surface morphology of the films was studied using a Tescan Vega 3LMH scanning electron microscope operating in secondary electron mode, which enables high-resolution imaging. X-ray structural analysis was performed using a DRON-type diffractometer with a step of 0.01–0.02 degrees, allowing Θ-2Θ scanning and peak registration for lattice parameter analysis.

Auxiliary equipment included a vacuum chamber used for annealing the targets, and a heating system that provided the required temperature conditions during substrate preparation and sputtering. This equipment setup ensured high accuracy and reproducibility of experimental results [4-6].

The material preparation process is critically important to ensure film quality and the accuracy of research results. Cadmium telluride powder was subjected to cold pressing to form a compact target. The resulting target was annealed in a vacuum chamber at 60–80 °C for three hours. This step was necessary to remove residual moisture and ensure uniform material density, which promoted stability in sputtering parameters.

Before film deposition, the soda-lime glass substrates were cleaned with an alcohol solution to remove possible contaminants that could negatively affect film adhesion. The cleaned substrates were mounted on special holders with heaters, which ensured uniform heating during deposition. This step was important to reduce residual stress in the material and improve its structural properties.

The synthesis of thin CdTe films was carried out by DC magnetron sputtering. This method is one of the most effective for obtaining thin film materials with a high degree of uniformity. Before sputtering, the target was heated to a temperature sufficient for thermionic emission, which lasted for 15 minutes. After that, "target training" was performed for five minutes, which allowed the plasma discharge parameters to stabilize [7].

The deposition process was carried out under the following conditions: magnetron current of 80 mA, sputtering time of 15 minutes, and a distance between the target and the substrate of 40 mm. These parameters were chosen to produce films with a fixed thickness of 4.5 μm, which made it possible to study their effect on the structural and elastic properties of the material. During film formation, crystallites of different sizes were formed, which determined the microstructural characteristics of the films.

Chloride treatment of the thin-film cadmium telluride was carried out by thermal vacuum evaporation of $CdCl_2$ with a film thickness of 1.0–1.25 μm, followed by annealing in a muffle furnace at a temperature of 390–410 °C for 20 minutes in air atmosphere. The resulting film structures after chloride treatment were provided for surface studies using electron microscopy and X-ray structural analysis.

The study focuses on determining the effect of thin-film cadmium telluride (CdTe) thickness, obtained by DC magnetron sputtering, on their structural and elastic properties. Cadmium telluride is a promising material due to its unique properties, including high optical sensitivity and energy efficiency, making it ideal for use in solar cells, optical sensors, and semiconductor devices [8]. Since the elastic parameters of the material, such as internal strain and microstress, can significantly vary with film thickness, it is necessary to study these dependencies precisely to ensure stability and predictability of its characteristics in technological applications.

The study includes not only the analysis of crystallite size and film strain, but also a comprehensive analysis of lattice parameters, energy density, and stress. The use of modern structural analysis methods such as X-ray diffraction (XRD) and scanning electron microscopy (SEM) allows not only to obtain precise data on film morphology but also to expand understanding of their microstructure. It is expected that the results of this research will serve as the basis for further improvement of high-quality thin-film material fabrication technologies [9].

III. SURFACE MORPHOLOGY OF FILMS

The surface morphology of the films was studied using scanning electron microscopy (SEM). SEM analysis showed that the films are characterized by high surface homogeneity, which indicates their high-quality formation. [10]

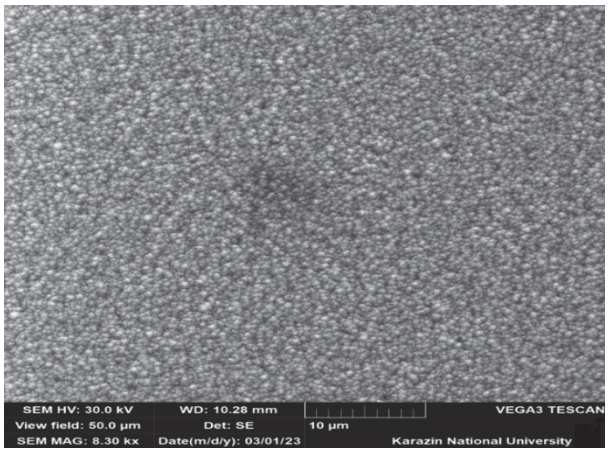


Fig. 1. SEM image of the surface of a thin film of CdTe of hexagonal modification in the initial state

Fig. 1 shows the X-ray diffraction patterns of CdTe thin films sputtered by magnetron sputtering. All patterns confirm the hexagonal structure of the obtained samples when compared with the data of the JCPDS program [11-15].

The instrumentally corrected broadening (β) of the diffraction peak is obtained by subtracting the instrumental broadening from the observed broadening (βi) [2],

$$\beta = [\beta_0^2 - \beta_i^2]^{\frac{1}{2}} \quad (1)$$

where β_0 – the experimentally observed broadening. The average crystallite size was determined using the corrected full width at half maximum (FWHM) for each peak, applying the Scherrer equation [1],

$$D = \frac{k\lambda}{\beta cos\Theta} \quad (2)$$

where β - is the corrected FWHM of each XRD peak in radians, and Θ is the diffraction angle. In this equation, the Scherrer constant k is equal to 1, and the wavelength of the X-ray radiation λ is 0.1788996 nm.

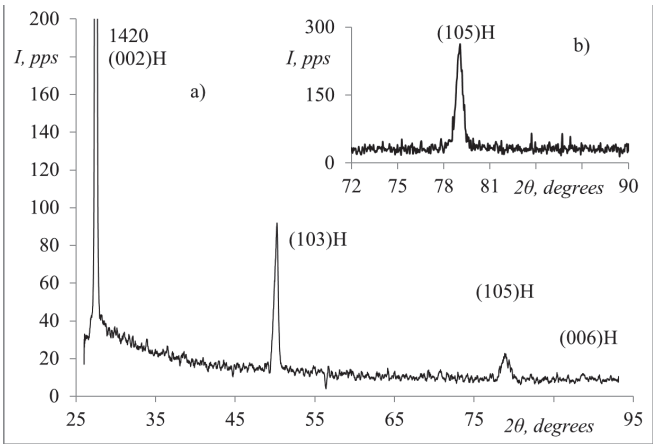


Fig. 2. X-ray diffraction patterns of hexagonal-phase CdTe thin films in the as-deposited state (a), and tilted scans (b)

The average crystallite size was calculated to be in the range of 39 to 54 nm. In addition, based on the X-ray diffraction data and using Equation (3), the lattice constants a, c, and d-spacings were calculated. The results are presented in Table 1, [3,12].

$$\lambda = 2dsin\theta \quad (3)$$

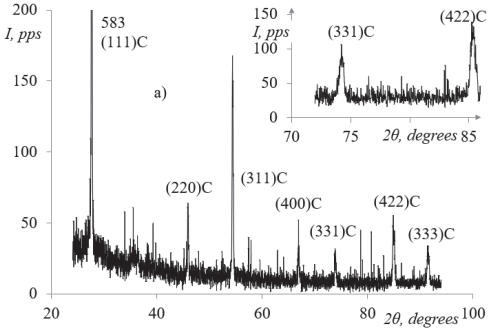


Fig. 3. X-ray diffraction patterns of CdTe thin films: cubic phase after chloride treatment (a), and tilted scans (b)

TABLE I. CALCULATED VALUES OF INTERPLANAR SPACING D, CRYSTALLITE SIZE D, AND MICROSTRAIN E FOR DIFFERENT (HKL) PLANES OF CDTE THIN FILMS

Sample	2θ, °	hkl	d, nm	FWHM, °	D, nm	<D>, nm	ε·10^{-3}, rel. units	<ε>·10^{-3}, rel. units
CdTe hex	27,50	(002)	0,376	0,17	60,59	38,95	3,59	3,69
	78,97	(105)	0,141	1,19	11,15		3,32	
	90,99	(006)	0,125	0,32	45,11		4,17	
CdTe cube	27,59	(111)	0,375	0,18	66,56	54,40	2,47	1,97
	45,88	(220)	0,229	0,23	52,08		2,10	
	54,35	(311)	0,196	0,19	70,26		2,37	
	66,89	(400)	0,162	0,28	44,55		2,45	
	73,83	(331)	0,149	0,27	52,96		0,76	
	84,90	(422)	0,133	0,30	52,76		1,72	
	91,44	(511)	0,125	0,39	41,63		1,90	

The average crystallite size was also determined graphically using the Scherrer plot, as shown in Fig. 3 (1–2), which was constructed by plotting 1/β along the X-axis versus cos Θ along the Y-axis, in accordance with the modified Scherrer equation given in Equation (4) [4].

$$cos\Theta = \frac{k\lambda}{D}\left(\frac{1}{\beta}\right) \quad (4)$$

The crystal sizes for the CdTe thin films were obtained from the slope of the extrapolated linear graph. It is immediately evident that the average crystal size increased by almost 40% after chloride treatment, which also corresponds to the decrease in peak widths (FWHM). This may indicate the growth or recrystallization of nanocrystals under the influence of chloride treatment, improving the structural ordering of the material.

The average value of microstrains in the CdTe sample significantly decreased after chloride treatment - from 3.69×10^{-3} relative units before treatment to 1.97×10^{-3} relative units. This indicates a reduction of stress in the crystal lattice, which is a positive effect since stress usually degrades the mechanical and electrical properties of the material.

After chloride treatment, new peaks of the (331) and (422) planes of the cubic modification appear, which also indicates recrystallization processes occurring within the volume of the thin film.

IV. DETERMINATION OF CRYSTALLITE SIZE AND DEFORMATION LEVEL BY THE WILLIAMSON-HALL METHOD

Ideally, X-ray diffraction peaks have a delta function for bulk materials. However, for thin films, these peaks are broadened due to the effect of crystallite size and the effect of

internal deformation of the crystal lattice. Therefore, by analyzing the peak broadening, it becomes possible to obtain the values of internal deformation and average crystallite size, as well as other elastic properties, using modified Williamson-Hall equations, such as the equation of the uniform deformation model (UDM), the equation of the uniform deformation model (USDM), and the equation of the uniform deformation and elastic distortion model (UDEDM).

According to the Williamson-Hall method, the total broadening in the X-ray peak is given by the formula [5,13],

$$\beta_{hkl} = \beta_D + \beta_\varepsilon \tag{5}$$

where β_{hkl} is the total expansion, β_D corresponds to the expansion due to crystallite size, and β_ε corresponds to the expansion due to deformation. Again, according to the Stokes-Wilson equation, the internal deformation is expressed as [6,12],

$$\varepsilon = \frac{\beta_s}{4tg\Theta} \tag{6}$$

Therefore, (5) after permutation can be written as,

$$\beta_hkl\ cos\Theta = k\lambda/D + 4\varepsilon sin\Theta \tag{7}$$

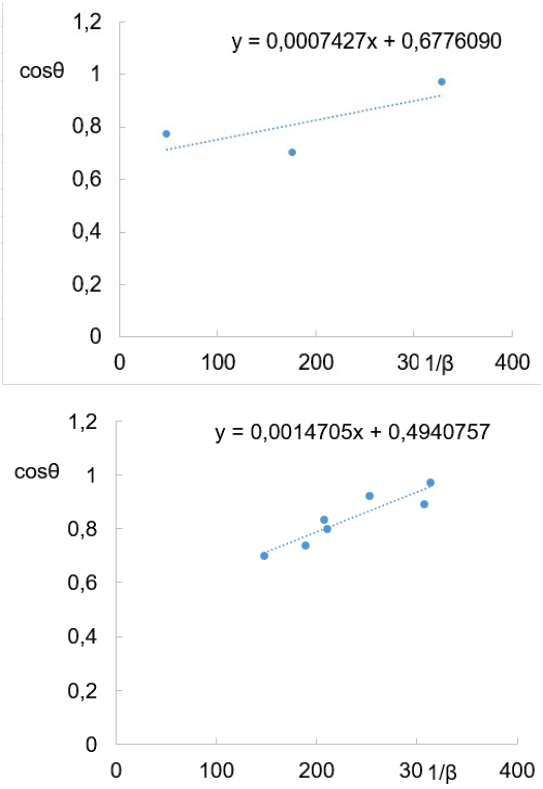


Fig. 4. Scherrer plots for all CdTe thin film samples

This equation is the equation of the UDM model [14-16]. Build a graph with $4sin\Theta$ along the X axis and $\beta_{hkl}cos\Theta$ along the Y axis for different (hkl) planes of both CdTe thin film samples, which are shown in Fig. 4 (1-2), respectively. Here, the intercept and slope of the graph give the crystallite size and lattice strain.

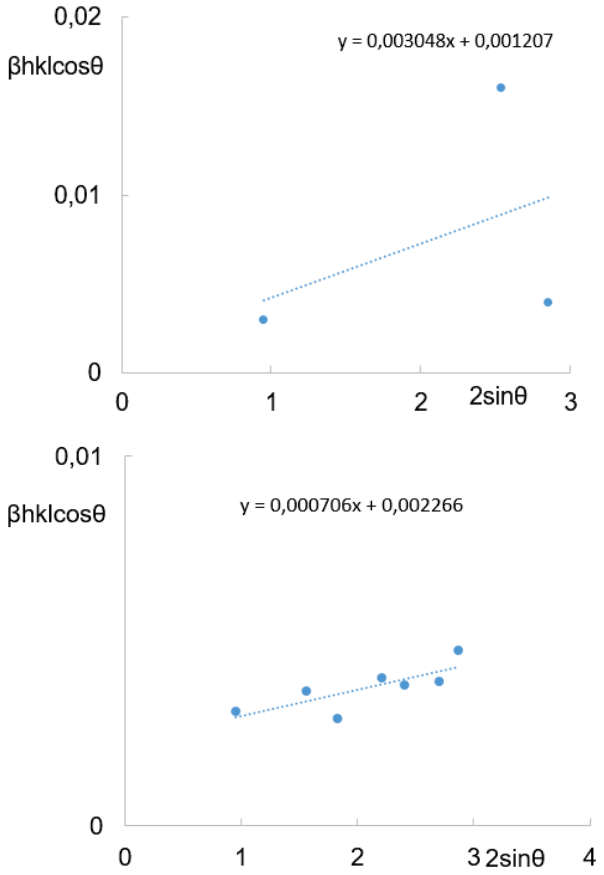


Fig. 5. Uniform deformation model (UDM) plot for all CdTe thin film samples

Now, if in thin films consisting of nanocrystals the strain is small, then according to the USDM model it can be assumed that the stress is uniform in all crystallographic directions. For uniform stresses and small values of strain Hooke proposed a generalized law of the relationship between strain and stress, which looks like, $\sigma=\varepsilon \bullet Yhkl$, where σ uniform stress in the crystal, and Y_{hkl} denotes the modulus of elasticity. Thus, using Hooke's law in equation (7), the modified Williamson-Hall equation for the USDM method has the following form [17],

$$\beta_{hkl}cos\Theta = \frac{k\lambda}{D} + \frac{4\sigma sin\Theta}{Y_{hkl}} \tag{8}$$

Now, the value Y_{hkl} can be calculated by the following formula according to the hexagonal structure of the samples [29],

$$\frac{1}{Y_{hkl}} = \frac{\left[h^2+\frac{(h+2k)^2}{3}+\left(\frac{a}{c}l\right)^2\right]^2}{\left[s_{11}\left(h^2+\frac{(h+2k)^2}{3}\right)^2+s_{33}\left(\frac{a}{c}l\right)^4+(s_{44}+2s_{13})\left(h^2+\frac{(h+2k)^2}{3}\right)\left(\frac{a}{c}l\right)^2\right]} \tag{9}$$

In equation 9, S11, S13, S33, and S44 denote the elastic compliances of hexagonal CdTe, and (hkl) are the Miller indices. Again, the relationship between the elastic counterparts (S_{11}, S_{13}, S_{33}, and S44) has the following form [18]. The values of elastic correspondences S_{11}, S_{13}, S_{33}, and S44 were identified as 0.019011, 0.005409, 0.019366, and 0.076336 GPa^{-1}, respectively.

A real crystal is imperfect due to the presence of various defects and dislocations. As a result, the linear relationship between stress and strain, which is assumed by the USDM model, does not apply to real crystals. Therefore, a uniform strain energy density model (UDEDM) is needed, which takes into account the uniform anisotropic deformation of the crystal lattice in all crystallographic orientations. According to Hooke's law, the energy density and strain are related by the following equation, $U=\varepsilon^2\times(Y_{hkl}^2)$, where U - energy density, as well as the relationship between energy density and voltage $U=\sigma^2/Y_{hkl}$.

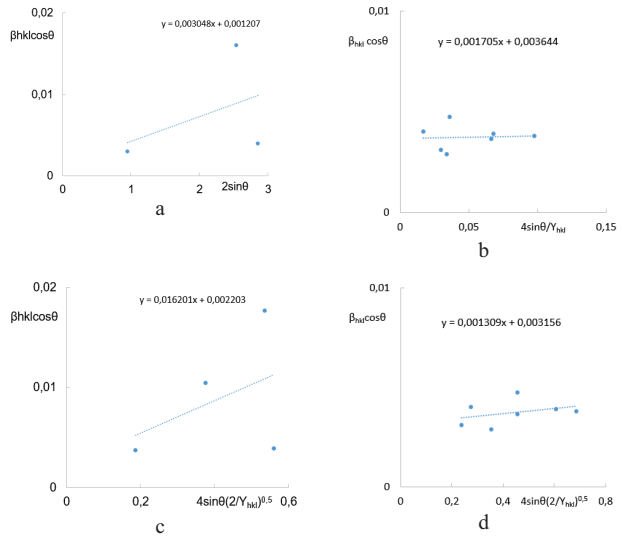
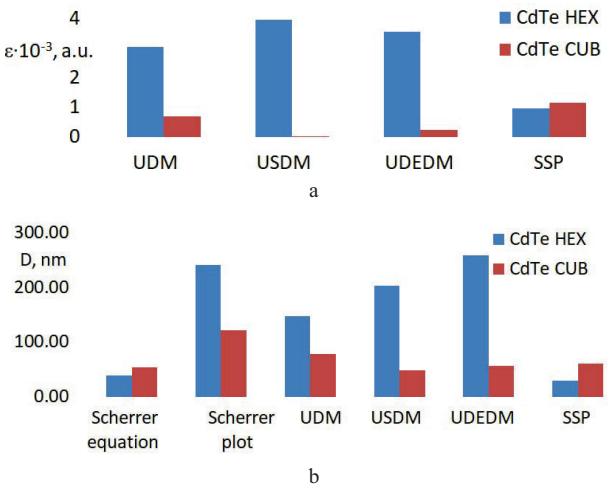


Fig. 6. Graph of the uniform stress strain model (USDM) – a,b and the uniform strain energy density (UDEDM) – c,d for all CdTe thin film samples

Graphs are built using βhklCosΘ along the X axis and 4SinΘ (2U/Yhkl) 0,5 along the Y axis, as shown in Fig. 6. The crystallite size and energy density were determined from the slope and intercept of the plots, the corresponding values are given in Fig. 7. It was observed that the slope in each plot for the corresponding model is positive. This positive slope indicates an increase in the lattice spacing [19-21], which leads to an increase in the lattice constant due to the accumulation of structural slip defects within the grains of the polycrystalline material [22]. The relationship between the increase in the lattice spacing and the positive slope can be explained by the width of the XRD peaks. With a change in the value of 2Θ, the intensity and width of the X-ray peak patterns decrease depending on the crystallographic planes. This can also be observed with a decrease in the interplanar spacings d, as shown in Fig. 7.



Calculation of crystal sizes and deformations using the UDM and USDM models, and using the UDEDM model and the SSP method for all CdTe thin film samples

According to the Scherrer equation, the crystal size increased, which is a positive effect [23]. However, according to the Scherrer plot, the size decreased significantly, which may indicate that the analysis method takes into account the influence of microstrains or sample structure. Microstrains decreased in the UDM model method with $3{,}05\times10^{-3}$ to $0{,}71\times10^{-3}$ after processing. This indicates a decrease in stresses in the crystal lattice. On the contrary, in the USDM model, microstrain increased, which may be caused by changes in the orientation of the structures.

Significant reduction in stress in the material with 0,20 GPa to 0,002 GPa. It shows us that chloride treatment effectively reduced the stress in the material, which contributes to increasing its mechanical stability.

The crystallite size according to the UDEDM model method significantly decreased from 259.27 nm to 56.69 nm after chloride treatment.

On the contrary, according to the SSP method, the crystallite size increased from 30.09 nm to 61.02 nm. The difference between the model results may be due to different approaches to taking into account the influence of microstrains and energy factors. In general, according to the SSP method, the growth of crystallites after treatment is visible, which indicates an improvement in the order of the structure.

Mechanical stress, before treatment: 0.12 GPa, after treatment: 0.01 GPa.

Chloride treatment significantly reduced the stresses in the crystal lattice. This indicates a reduction in defects and internal stresses in the material, which improves its mechanical properties.

Significant reduction of microdeformations using the UDEDM model method from $3{,}56\times10^{-3}$ to $0{,}25\times10^{-3}$ after treatment.

Slight increase in microstrains by the SSP method from $0{,}99\times10^{-3}$ to $1{,}17\times10^{-3}$.

According to the UDEDM model, chloride treatment reduced defects, while the SSP method indicates possible local changes in the structure that led to a slight increase in microstrains.

Energy density, KJ/m³ before treatment -3136,80 KJ/m³, after treatment -17,13 КДж/м³. A significant decrease in

energy density indicates a decrease in the internal energy of the system after processing [24]. This means that the material becomes more stable and less prone to internal stresses.

V. Conclusion

Within the framework of the study, thin films of CdTe metastable hexagonal modification were synthesized by magnetron sputtering, and their chloride treatment using $CdCl_2$ and subsequent thermal annealing were performed. X-ray structural studies showed that the treatment significantly affects the microstructural parameters of the samples: the average crystallite size according to the Scherrer equation increased by 40%, which confirms recrystallization. The SSP method showed that the crystallites grew to 61 nm with a slight increase in microstrains. The use of Williamson-Hall models (UDM, USDM, UDEDM) revealed a significant decrease in microstrains, stresses and energy density, which indicates a decrease in defects and an increase in mechanical stability. Thus, chloride treatment positively affects the microcrystalline characteristics of CdTe films, which is important for optimizing their properties in photovoltaic and electronic devices. The results obtained are of applied importance, as they contribute to increasing the technological level and competitiveness of enterprises. At the same time, compliance with modern standards of occupational health and safety, environmental and technogenic safety ensures sustainable development and safe implementation of research results into production.

The conducted study has demonstrated a significant impact of chloride treatment on the microstructural and elastic properties of cadmium telluride (CdTe) thin films deposited by DC magnetron sputtering. The analysis revealed that chloride treatment promotes the improvement of the material's crystalline structure due to recrystallization processes, which are accompanied by an increase in the average crystallite size by nearly 40%. This is confirmed by both the reduction in the full width at half maximum (FWHM) of the X-ray diffraction peaks and the values obtained using the modified Scherrer equation.

A substantial decrease in microstrain (from $3.69 \cdot 10^{-3}$ to $1.97 \cdot 10^{-3}$ rel. units) indicates a relaxation of internal stresses in the crystal lattice after thermal treatment. This stress reduction has a positive effect on the mechanical stability and electrophysical properties of the material, which is critical for CdTe-based applications in photovoltaic and sensor devices.

In addition to the improved microstructure, the chloride treatment led to the appearance of new diffraction peaks corresponding to the (331) and (422) planes of the cubic phase of CdTe. This serves as further evidence of activated phase and structural transformations, which may enhance energy conversion efficiency in photovoltaic devices, as the material becomes more structurally ordered and defects are passivated.

Therefore, the results of this work confirm the feasibility of using chloride treatment as a key step in the technological process of fabricating thin-film CdTe-based structures. The study lays the groundwork for further optimization of deposition and thermal treatment parameters aimed at improving the performance characteristics of CdTe films for next-generation photovoltaic systems.

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Structural Synthesis of Electric Drive with Feedback on Derivative of Velocity Control Error

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***Abstract*—Potentially high performance indexes of relay control systems of electric drives are most fully revealed when they are closed by feedback on the control error and its derivative. The problem of noise-immune differentiation of the error signal is also most effectively solved by observing devices controlled by relay controllers. In this paper, a parametric synthesis of a relay velocity control system of an electric drive is performed and a number of variants of its structural implementation in various operating modes are investigated. The results of the study made it possible to justify the choice of the most effective scheme for calculating feedback signals. It is shown that the two-channel structure of the observer of the control error derivative provides a combination of high accuracy in calculating the pulsating components of acceleration with an extremely fast response to abrupt changes in the set velocity of the electric drive. The grounding of the structure of the control system and state observers, as well as their parametric synthesis, is performed using the N-i switching method. The common methodological basis for the synthesis of controllers of both the electric drive and observers makes it possible to match the frequencies of their switching in sliding modes. The results of the study of transients confirm the effectiveness of the proposed structural solution.**

Keywords—electric drive, derivative observer, relay controller, sliding mode, N–i switching method.

I. Introduction

Relay control systems have low sensitivity to the action of resistance forces and to the variability of electric drive parameters [1], since their controllers operate in a sliding mode [2]. In this case, the behavior of the system in the theoretical aspect is subject to the sliding equations of the controllers [3, 4] and largely depends on the type of feedback [5]. To ensure high indexes of performance [6, 7], the feedback signals must be as close as possible to the derivatives of the control error [8]. In technical implementation, the behavior of the control system and even the very possibility of the existence of a sliding mode is determined by the quality of measurement of the feedback signals included in the switching functions [9]. Noise-proof differentiation of the measured values is necessary [10], and at the same time, high-frequency components caused by current ripple must be present in the calculated derivatives [11]. These contradictory requirements can be satisfied using a relay control system with derivative observers [12]. Another contradiction is the need to implement exact differentiation [13] while minimizing the convergence time of the observer's own transients, which requires simultaneous optimization in terms of accuracy and speed [14]. This contradiction can be resolved by combining the techniques of structural and parametric synthesis [13, 14], based on the N-i switching method.

II. Preliminaries

The dynamics of an electric drive based on a DC machine is described by differential equations

$$\left.\begin{aligned} p\omega = \varepsilon = \frac{c}{J}(i - i_s) \\ p\varepsilon = a = \frac{c}{J}\cdot\frac{u - R\cdot i - c\omega}{L} \end{aligned}\right\}, \tag{1}$$

where u is the converter voltage, i is the armature current, i_s is the static current, ω is the angular velocity, ε is the angular acceleration, a is the angular jerk, $p = d/dt$ is the symbol of differentiation with respect to time, R is the active resistance and L is the inductance of the armature circuit, $c = k\Phi$ is the factor taken as a constant at a steady magnetic flux Φ, J is the total moment of inertia of the electric drive.

The relay algorithm for controlling the electric drive (1), which can be synthesized by the N–i switching method, in the basic configuration provides for the presence of acceleration feedback [13] and has the form

$$\left.\begin{aligned} u_{R\omega} = \varepsilon^* = -\varepsilon_{max}\cdot sign\left(\omega - \omega^* + K_{\omega\varepsilon}\cdot\varepsilon\right) \\ u_{R\varepsilon} = u^* = -u_{max}\cdot sign\left(\varepsilon - \varepsilon^*\right) \end{aligned}\right\}, \tag{2}$$

where $u_{R\omega}$ is the current velocity controller R_ω signal, which is the acceleration set signal ε^* for the subordinate controller R_ε, $K_{\omega\varepsilon}$ is the velocity controller feedback factor on acceleration, ε_{max}, u_{max} are the constraint levels of acceleration and power converter voltage.

In addition to controlling the velocity of the electric drive, the cascade structure of the controller performs intermediate coordinate constraint. In controller (2), this coordinate is the acceleration, and the armature current limitation is mediated. Performance optimization of the control system (1), (2) by the N–i switching method ensures the formation of transients with one single switching of the velocity controller R_ω before the occurrence of its sliding mode. The parametric synthesis procedure is reduced to calculation by the formulas

$$\varepsilon_{max} = \frac{c}{J}i_{max},\ a_{max} = \frac{c}{JL}u_{max}, \tag{3}$$

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$$K_{\omega\varepsilon} = 0.5\varepsilon_{max} / a_{max}, \quad (4)$$

where $K_{\omega\varepsilon}$ is the factor of the velocity controller feedback for acceleration, ε_{max} and a_{max} are the constraint levels.

III. Synthesis and Study of the Control System

To perform the study, we will take as the control object a DC electric motor (1) with the following parameters and nominal values:

$$R = 0.93\Omega,\ c = 1.84\,V\cdot s,\ L = 0.0243H,\ J = 0.122kg\cdot m^2,\ \omega_n = 106.66s^{-1},\ i_{max} = 38.25A, u_{max} = 1.25\cdot 220V. \quad (5)$$

By applying formulas (3), (5) we obtain data for setting up the controller (2)

$$a_{max} = 85472s^{-3}, \varepsilon_{max} = 288.5s^{-2}, K_{\omega\varepsilon} = 0.00169s. \quad (6)$$

Equations (2) describe a system with ideal relay elements and ideal acceleration feedback. Therefore, in addition to parametric optimization, the results of which are summarized in (9), the relay controller needs structural additions that bring it closer to real technical solutions. Ideal relays generate sliding modes with infinite frequency, therefore, their use in practice is not only impossible, but also impractical. Therefore, we will take as a basis relay elements with a hysteresis loop

$$y = gist(x) = \begin{cases} -1 \text{ if } x < -\delta \\ -1 \text{ if } -\delta \le x \le \delta = 0 \text{ and } y = -1 \\ +1 \text{ if } -\delta \le x \le \delta = 0 \text{ and } y = +1 \\ +1 \text{ if } x > \delta \end{cases}, \quad (7)$$

which form acceleration (or armature current) ripple in the range determined by the loop width 2δ, and thereby limit the switching frequency of the power converter of the electric drive [13]. To ensure an acceptable amplitude of current ripple δi, the following loop width should be set for the controllers:

$$\delta_{R\varepsilon} = \delta\varepsilon, \delta_{R\omega} = K_{\omega\varepsilon}\cdot\delta\varepsilon. \quad (8)$$

In this example, allowing for 6% current ripple, we have the numerical values

$$\delta i = 2.295\,A, \delta_\varepsilon = \delta_{R\varepsilon} = 17.31s^{-2}, \delta_{R\omega} = 0.0292s^{-1}. \quad (9)$$

In addition to reducing losses in the converter, limiting the frequency of acceleration ripple allows the use of a velocity derivative observer, also built on the basis of a relay controller. Such an observer is described by the equations

$$\left.\begin{aligned} &p\hat{\omega} = \hat{\varepsilon},\ p\hat{\varepsilon} = \hat{a},\ \hat{s} = \omega - \hat{\omega} + \hat{K}\left(c/J\cdot i - \hat{\varepsilon}\right) \\ &u_{\hat{R}} = a^* = \hat{a}_{max} sign(\hat{s}) \end{aligned}\right\}, \quad (10)$$

where $\hat{s}$ is the switching function of the relay controller $\hat{R}$, the sign «^» marks the values and parameters of the observer, the symbolic designations of which are similar to those used in (1), (2).

By tracking the input value ω, the observer ensures the equality of the values $\hat{\omega} = \omega$, which allows the following change of variable to be made in (2):

$$\varepsilon = \hat{\varepsilon}. \quad (11)$$

The switching function $\hat{s}$ in (10) uses the measured current signal with high-frequency pulses to reliably calculate the acceleration signal, which is substantiated in detail in [13]. Another fundamental structural solution that ensures the implementation of (10) is the use of relay elements of the form (7). This not only limits the slip frequency $\hat{R}$, but also allows us to separate the sliding modes in (2) and in (10). Assigning a loop width $\delta_{\hat{R}}$ that is an order of magnitude smaller than $\delta_{R\omega}$ in (8) forces the observer to operate with a proportionally higher switching frequency, enveloping the calculated signal $\varepsilon(t)$ with a calculated signal $\hat{\varepsilon}(t)$ with negligible ripple relative to it. It should be emphasized that the ability of the observer to "overtake" the calculated signal is created by the corresponding reserve in the amplitude $\hat{a}_{max}$ of its relay element. As a result, the parametric synthesis (10) must contain the following calculations:

$$\hat{a}_{max} = 10\cdot a_{max},\ \delta\hat{\varepsilon} = 0.1\cdot\delta\varepsilon,\ \delta_{\hat{R}} = \hat{K}\cdot\delta\hat{\varepsilon}. \quad (12)$$

For the example with data (5), (6) we have the values

$$\hat{a}_{max} = 854720s^{-3}, \delta_{\hat{R}} = 0.000292s^{-1}. \quad (13)$$

Taking $\hat{\varepsilon}_{max} = \varepsilon_{max}$ and using (4) for (10), we obtain

$$\hat{\varepsilon}_{max} = 288.5s^{-2}, \hat{K} = 0.000169s. \quad (14)$$

The control system of the electric drive (1), (2), (10) with parameters (5), (6), (9), (13), (14) ensures the optimal speed of the transient process (Fig. 1, a) under conditions of constant action of the active moment of resistance and ensures compensation for the shock application of the load. In Fig. 1, a the diagrams $u_{R\omega}(t), u_{R\varepsilon}(t)$ are presented in

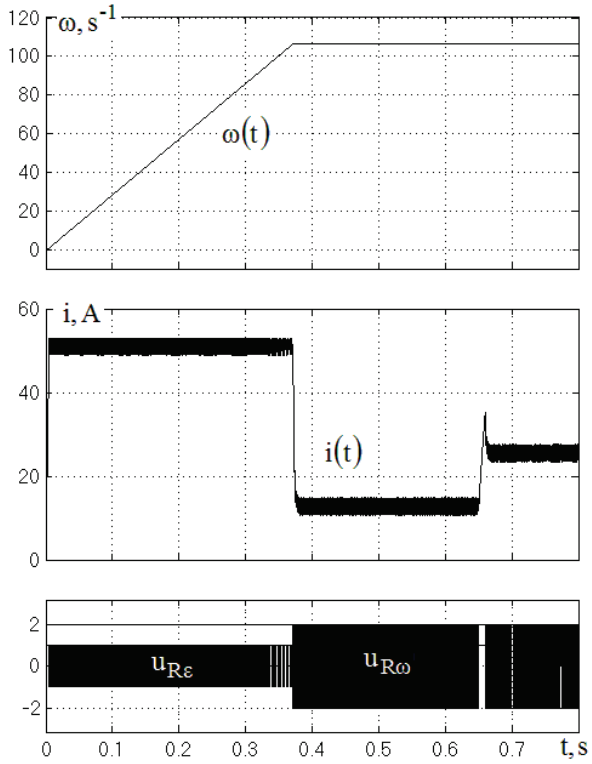


Fig. 1. Transients in a control system with acceleration feedback.

relative units, their scale reflects the hierarchy in the cascade (2). It should be emphasized that the modeling of sliding modes in electric drives does not require such cumbersome calculations as, for example, the study of the field pattern [15], however, it has a number of features concerning the choice of the method and the integration step, taking into account which helps to ensure the reliability of the results [16]. In particular, the value of the integration step should be consistent with the period of pulsations of acceleration observer's coordinates.

Despite the high quality of the system (1), (2) operation in the velocity stabilization mode, the processing of the reference action, changing according to the harmonic law $\omega^*(t) = \omega_n \cos(2t)$, occurs with an error $\Delta\omega$, the amplitude of which is 0.8% of the amplitude ω^* (Fig. 2). The reason for the loss of accuracy can be easily substantiated by analyzing the slip equation of the controller R_ω from (2)

$$\omega - \omega^* + K_{\omega\varepsilon} \cdot p\omega = 0\,, \tag{15}$$

from which it is easy to obtain the transfer function

$$W_{R\omega}(p) = \frac{\omega(p)}{\omega^*(p)} = \frac{1}{K_{\omega\varepsilon} \cdot p + 1}\,, \tag{16}$$

corresponding to the aperiodic link of the first order. This fact explains the presence of amplitude and phase errors of the electric drive velocity control system, which appear when the reference action $\omega^*(t)$ is inconstant. The solution to the problem consists in bringing the transfer function (16) to a unit value $W_{R\omega}(p) = 1$, for which its numerator must be equal to the denominator. It is feasible by introducing into the switching function (15) the derivative of the velocity control error instead of the acceleration

$$\Delta\omega + K_{\omega\varepsilon} \cdot p(\Delta\omega) = \omega - \omega^* + K_{\omega\varepsilon} \cdot p(\omega - \omega^*) = 0\,, \tag{17}$$

which assumes a formal replacement of the variable in (2)

$$\varepsilon = p(\Delta\omega)\,. \tag{18}$$

This structural solution is known [14], and here expressions (15) – (18) justify it in relation to (2).

At first glance, the practical implementation of the control system (1), (2) with substitution (18) may consist of using a relay observer (10) with the input of $\Delta\omega(t)$ instead of $\omega(t)$. However, the setting signal $\omega^*(t)$ may undergo not only smooth but also abrupt changes, that is, in the general case it has discontinuities that cause their own transients in the observer. This fact causes two contradictions that prevent the described "simple" solution.

First, during the observer's own transient process, the sliding mode of its controller is disrupted, and the derivative calculated by it cannot be used to close the electric drive control system. Therefore, the observer must issue a reliable feedback signal faster than the sliding mode based on this signal occurs in the electric drive. However, the possibility of forcing the observer's transient process is limited, so it is extremely difficult to make its duration negligibly small in relation to the duration of the electric drive's transient.

Secondly, to minimize the time of capture of the differentiated signal after the moment of its break, it is necessary to optimize the observer in terms of speed. However, the result of such optimization contradicts the requirement of minimizing the differentiation error for the following reasons. The observer error (10) directly depends on the coefficient of its internal feedback $\hat{K}$, as shown in [13]. In turn, the value $\hat{K}$ is linearly related to $\hat{\varepsilon}_{max}$ according to (4). However, the maximum of the coordinate $\hat{\varepsilon}(t)$, achieved in the optimal in terms of speed transient process of the error derivative observer, must many times exceed the value $\hat{\varepsilon}_{max}$ adopted during the synthesis of the acceleration observer. Note that the acceleration observer (10) continuously functions in a sliding mode, unlike the observer of the velocity control error derivative, therefore, during its parametric synthesis, $\hat{\varepsilon}_{max}$ is set at the level of the maximum acceleration of the electric drive ε_{max}, and the described contradiction does not arise.

Both of these problems can be solved simultaneously by organizing two differentiation channels on the basis that the derivative of the control error has two components

$$p(\Delta\omega) = p(\omega - \omega^*) = \varepsilon - p(\omega^*)\,. \tag{19}$$

One of these channels contains the acceleration observer (10) described above, which precisely reproduces the high-frequency ripple of the velocity derivative. In parallel with this, the second channel calculates the derivative of the velocity reference signal, it is optimized for speed and is described by equations similar to (10)

$$\left.\begin{aligned} &p\tilde{\omega} = \tilde{\varepsilon},\ p\tilde{\varepsilon} = \tilde{a},\ \tilde{s} = \omega^* - \tilde{\omega} - \tilde{K}\tilde{\varepsilon} \\ &u_{\tilde{R}} = a^* = \tilde{a}_{max} sign(\tilde{s}) \end{aligned}\right\}, \tag{20}$$

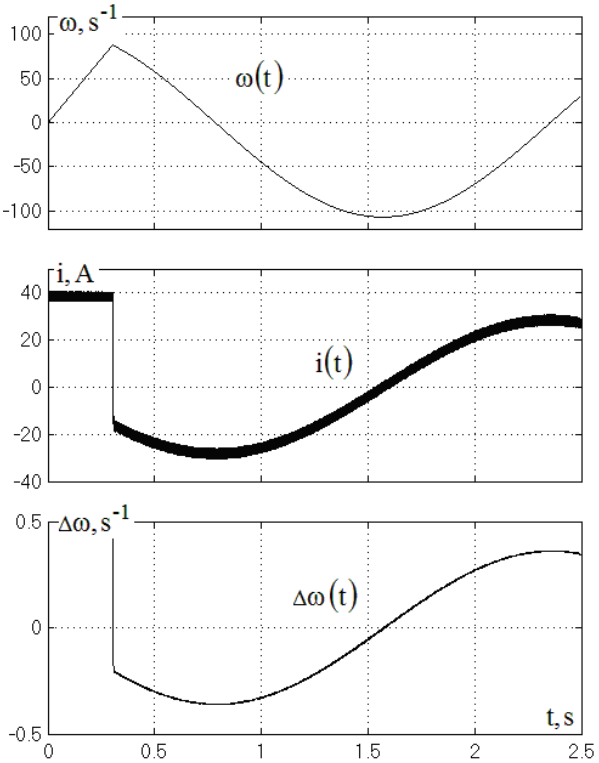


Fig. 2. Transients in a control system with acceleration feedback.

where the sign «~» marks the values and parameters of the second observer.

The relay element $\widetilde{R}$ of type (7) is configured similarly to (12) using the formulas

$$\widetilde{a}_{max} = 10 \cdot a_{max}, \ \delta\widetilde{\varepsilon} = 0.1 \cdot \delta\varepsilon, \ \delta_{\widetilde{R}} = \widetilde{K} \cdot \delta\widetilde{\varepsilon}. \qquad (21)$$

When the setting signal $\omega^*(t)$ changes abruptly by the value $\delta\omega^*$, the observer (20) is adjusted [14] according to formula (4), into which the following value of maximum acceleration is substituted together with $\widetilde{a}_{max}$:

$$\widetilde{\varepsilon}_{max} = \sqrt{\left|\delta\omega^*\right| \widetilde{a}_{max}}. \qquad (22)$$

Note that such adaptive tuning is only possible in digital controllers.

After the jump of $\omega^*(t)$ in the observer (20) with settings (21), (22), (4), the time-optimal transient of capturing the differentiable signal will be formed, and after the start of the sliding mode of the controller $\widetilde{R}$, the equality of coordinates

$$\widetilde{\varepsilon} = p\left(\omega^*\right) \qquad (23)$$

will be established and maintained.

The use of a two-channel observer (10), (20) not only eliminates the contradiction between accuracy and speed during its optimization. It allows not to solve, but to "bypass" the problem of the non-zero duration of the transient process of the observer. Since $\omega(t)$ is a continuous signal, the sliding mode of the acceleration observer (10) exists constantly, and the signal $\hat{\varepsilon}(t)$ reliably transmits the acceleration of the electric drive from the moment of initialization of the controller. This signal can be used in (2) to close the "fast" internal control loop based on $R\varepsilon$, which ensures the sliding mode of the controller (2) before entering the sliding mode of the controller $\widetilde{R}$ of the observer (20), i.e. before the appearance of a reliable signal of the derivative in the differentiation channel of the specified velocity after its jump. In this case, for the "slow" external control loop of the electric drive velocity based on $R\omega$ in (2), substitution (18) is implemented, which, taking into account (19) and (23), has the form

$$\varepsilon = \hat{\varepsilon} - \widetilde{\varepsilon}. \qquad (24)$$

Equality (24) becomes relevant with the emergence of the sliding mode of the controller $\widetilde{R}$ in (20), that is, obviously earlier than the sliding of $R\omega$ in (2) begins.

Finally, equations (2), taking into account the use of signals from a two-channel observer (10), (20), (24) take the form

$$\left.\begin{aligned} u_{R\omega} &= \varepsilon^* = -\varepsilon_{max} \cdot sign\left(\omega - \omega^* + K_{\omega\varepsilon} \cdot \left(\hat{\varepsilon} - \widetilde{\varepsilon}\right)\right) \\ u_{R\varepsilon} &= u^* = -u_{max} \cdot sign\left(\hat{\varepsilon} - \varepsilon^*\right) \end{aligned}\right\}. \qquad (25)$$

Having accepted the value of the jump of the reference action $\delta\omega^* = \omega_n = 106.66 s^{-1}$, according to (21), (22), (4) we obtain for the example under consideration

$$\left.\begin{aligned} &\widetilde{\varepsilon}_{max} = 9548 \ s^{-2}, \ \widetilde{a}_{max} = 454720 \ s^{-3}, \\ &\delta_{\widetilde{R}} = 0.00967 \ s^{-1}, \ \widetilde{K} = 0.005585 \ s. \end{aligned}\right. \qquad (26)$$

IV. Results of the Study

Fig. 3 shows the time diagrams of transients in the control system (1), (25) with derivative observers (10), (20) at a given velocity $\omega^*(t) = \omega_n cos(2t)$. Both the initial acceleration of the electric drive and the subsequent tracking of the setting signal occur with the expected performance and accuracy. In particular, the velocity control error decreases by approximately two orders of magnitude and does not exceed 0.01% of the amplitude ω^*. This result is achieved due to the sliding of the controllers both in the electric drive controller (25) and in both observers (10) and (20) in quasi-static modes. In this case, both the electric drive controller (25) and the controller of the second channel observer (20) form transients that are optimal in terms of performance after a step change in $\omega^*(t)$, as evidenced by the execution of a single switching in the transient process by the relay elements $R\omega$, $\widetilde{R}$. Comparison of the diagrams $\Delta\omega(t)$ in Fig. 3 and Fig. 2 highlights the advantages of the system with a two-channel differentiator of the control error. It should be emphasized that this result is entirely achieved on the basis of structural solutions, since the parameters of controllers (2) and (25) are identical. Note that the general appearance of the time diagrams of $\omega(t)$ and $i(t)$ in Fig. 2 and Fig. 3 is practically indistinguishable due to the relatively small control error $\Delta\omega(t)$ in both variants of the system structure. Fig. 4 shows the structural diagram of a two-channel observer, the input signals of which are the specified velocity

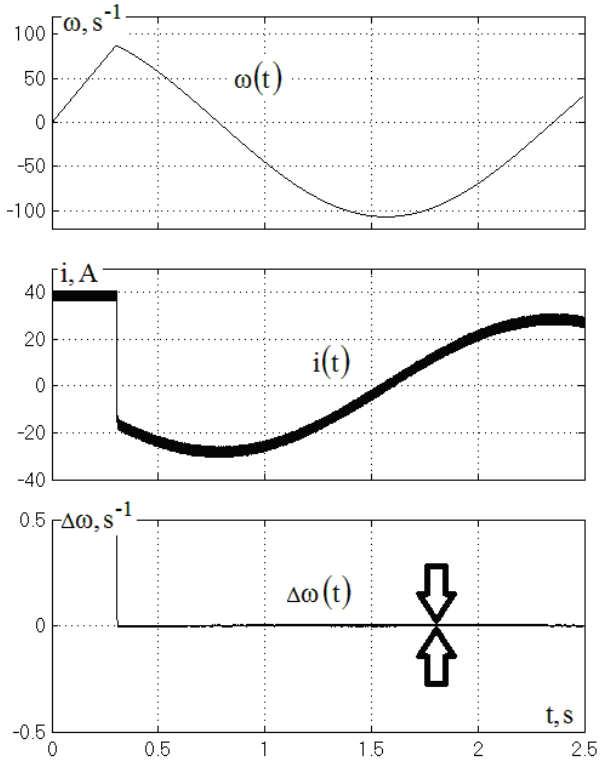


Fig. 3. Transients in a control system with feedback on the derivative of the velocity control error.

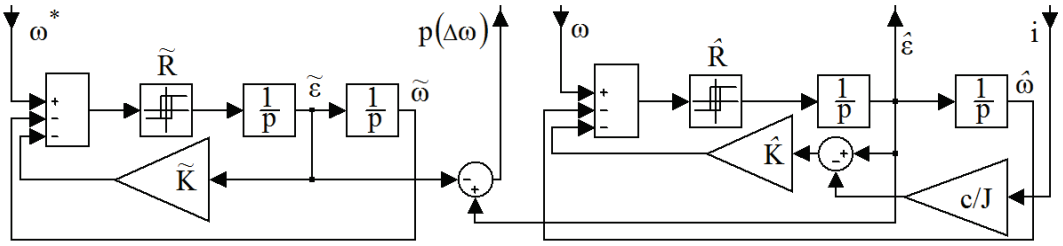


Fig. 4. Structure of a two-channel observer of the derivative of the velocity control error for precision electric drives.

ω^*, the measured velocity ω and the measured armature current i, and the output signals required for closing the controller feedback (25) are the calculated acceleration $\hat{\varepsilon}$ and the calculated derivative of the velocity control error $p(\Delta\omega)$. This variant of the circuit is built on non-inverting relay elements with the corresponding signs of the coordinates at the input of the adders.

In summary, it is necessary to note the features of interaction of structural units of control system (1), (25), (10), (20). Relay subsystems (10) and (20), forming a two-channel observer, function in parallel, and their sliding modes are mutually independent and, in the general case, their frequencies are not related. At the same time, as mentioned above, a necessary condition for the existence of a sliding mode in the electric drive controller (25) is a reliable calculation of derivatives by observers (10) and (20), ensured by their sliding modes. This condition is considered as knowingly fulfilled in the synthesis of (25) or (2), and the pulsations of coordinates created by switching relays from (10) and (20) are not taken into account due to their small amplitude. Mutual independence of the sliding modes of the three relay subsystems allows performing their parametric optimization separately, without resorting to consideration of the integral control system. This makes the computational component of parametric synthesis extremely concise and opens up prospects for using the N-i switching method to optimize relay control systems for electric drives in real time.

It is also important to note that relay control systems remain insensitive to variations in the parameters of the control object [13], including those caused by its nonlinearity, as long as the reserve in the amplitude of the control action and the quality of the feedback signals allow them to operate in a sliding mode. This fact emphasizes the importance of the developed observer specifically for relay control systems of electric drives.

V. Conclusion

The two-channel structure of the relay observer of the derivative allows taking into account the specifics of the components of the velocity control error during the parametric synthesis of each of the channels. This gives the observer a combination of speed during jumps of the reference signal with accuracy in calculating high-frequency pulsations of the derivative. High differentiation efficiency ensures the closure of the relay system of electric drive velocity control by the signal of the derivative of the control error, which allows such a system to meet the maximum requirements for performance indicators. The novelty of this study consists in the substantiation of the structure and synthesis of the parameters of the two-channel observer based on the N-i switching method. A promising direction for the development of the obtained result is its adaptation to higher-order relay control systems, in particular to positional electric drives, since the order of the system does not impose restrictions on the application of the N-i switching method.

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Research on the performance of a neural network for recognizing combat vehicles

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***Abstract*— the performance of the neural network for recognizing combat vehicles in an intelligent visual surveillance system has been investigated. The system is mobile and built with real-time motion and the ability to detect, classify, and track objects. The system captures and recognizes video images from a camera installed on the vehicle. In the proposed system, moving object detection is carried out using an adaptive background subtraction scheme, which we propose to use in combat. It is shown how to identify moving objects. We present object detection schemes, temporal difference models, and adaptive background difference models to compare the efficiency and quality of detection. The study implemented a method for detecting moving objects based on adaptive background difference models.**

***Keywords*— combat vehicles, recognizing, neural network**

I. Introduction

Tracking objects [1] is much more difficult in video sequences to improve their recognition and tracking performance.

We classify tracking methods [2] based on the representations of objects and motion used. Tracking objects consists of using appropriate image features, selecting motion models, and detecting objects. Computer vision technologies use methods of inclusion, analysis, and visualization. Surveillance defines monitoring, behavior, activity, and other variable information, the process of determining the location of a moving object using video surveillance cameras. A visual surveillance system usually consists of four main components: target Representation, object Localization, filtering, data association. Object representation and localization processes are typically implemented in a bottom-up manner, meaning that steps are performed sequentially without feedback from subsequent steps. These algorithms typically have low computational complexity.

Typical target detection [3] and localization methods include: blob tracking — is performed through image segmentation using approaches such as blob detection, block-based correlation, and optical flow analysis; kernel-based tracking (in particular, the Mean-shift method) — an iterative procedure that localizes an object by maximizing a similarity function; contour tracking — tracking the boundaries of an object based on its contour;

Feature matching [4] — matching features, in particular, image registration;

Point feature tracking — tracking point features of a scene. The task is formulated as finding the exact coordinates of the projections of certain points of the scene on each frame of a sequence of images obtained from both a stationary and a moving camera.

The following common algorithms are used for filtering:

Kalman filter [5] — an optimal recursive Bayesian filter for linear systems with Gaussian noise;

Particle filter is effective for modeling states in nonlinear and non-Gaussian processes using sampling.

The video processing process involves the following stages: video loading, pre-processing, application of the proposed algorithm that performs frame processing, and the object detection stage.

One of the basic approaches to detecting objects in a video stream is to pre-divide the video into separate frames with the subsequent application of image recognition algorithms to each frame separately. However, this method ignores the important temporal information contained in the video and, accordingly, does not allow fully taking into account the dynamics of objects in space and time.

For example, in the case of partial or complete occlusion, when an identified object (for example, a person) is briefly covered by another object (for example, a pedestrian), a system that analyzes only individual frames may not "understand" that this is the same object and will consider it lost. A similar problem arises in cases of blurring caused by rapid movement of the object, the camera, or both simultaneously. This blurring makes it difficult to accurately recognize the shape and position of the object, which reduces the accuracy of the system. In approaches where frames are analyzed in isolation, each frame is interpreted as a complete image, which makes it impossible to track objects over time. To overcome these problems, it is advisable to use methods that take into account temporal context, in particular object tracking, recurrent neural networks, or combining data from several consecutive frames.

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II. CONSTRUCTION OF THE CLASSIFIER

The following process describes the key steps in data preparation and training of deep neural networks for the image classification task. In particular, it provides for structured storage of images in jpg format with labels in appropriate folders by class, which allows automatic class determination when using frameworks such as Keras.

The primary target operation is to randomly divide the database into a training set (usually about 80%) and two additional validation and control sets (10% each, respectively). This approach helps reduce systematic errors that can arise due to biased data partitioning. The need to use a control sample is to realistically estimate the generalization limit of the model on new previously unseen data.

After the dataset is formalized, preprocessing is performed, in particular, resizing the images to fixed parameters (for example, 224×224 pixels) and grouping them into mini-packages (batches) for efficient computational process during training.

The next stage is data augmentation, which consists in artificially expanding the given set by simulating various transformations (rotations, scaling, brightness changes, etc.), which helps to increase the variability of training examples and reduce the risk of overtraining.

To build models, pre-trained architectures (e.g., ResNet, MobileNet) [6] are used, in which the last few layers with fixed weights are removed and frozen, with the addition of a new output layer corresponding to the number of target classes. The choice of optimizer (e.g., Adam) and metrics (e.g., accuracy) is based on the classification problem and target indicators.

The training process involves performing iterations with parameter optimization, without revealing any significant overtraining in terms of indicators on the validation set. At the end of training, the model is saved for further use in production or research conditions.

The classifier consisted of a dataset filled with images of tanks, bmp, btr, and backgrounds that we classify during training.

Underestimation or overestimation of accuracy is based on the analysis of results on a control sample, which allows for an objective assessment of the model's generalization ability. An important construct is the data structure that provides automatic class identification by folder names.

Thus, the above process allows for the creation of an efficient pipeline for automatic classification of military equipment images, providing optimal conditions for modeling and assessing the quality of machine learning models.

III. METHOD OF SEMANTIC SEGMENTATION

Semantic segmentation is one of the key tasks of computer vision, which consists in classifying each pixel of an image by assigning it a specific label corresponding to its belonging to a certain class of objects. This approach is widely used in such areas as autonomous transport control, object recognition and medical imaging.

The goal of semantic segmentation is to detect all objects in a scene and group pixels belonging to the same class. Unlike instance segmentation, semantic segmentation does not distinguish between individual instances of the same class, considering them as a single mass.

However, the accuracy of such models largely depends on the availability of annotations at the pixel level - a process that is extremely laborious, expensive and time-consuming. This is especially critical in the medical field, where, in addition to resource costs, highly qualified specialists with appropriate subject matter expertise are required.

To reduce costs and increase scalability, the Weakly Supervised Semantic Segmentation (WSSS) approach is becoming increasingly popular. This approach allows building models based on incomplete or less detailed annotations. In particular, within the self-supervised paradigm, it is possible to automatically generate so-called pseudo-labels using images with only known class affiliation.

At the first stage, using a classification neural network, images are passed through a model that generates activation maps for each class. The next steps are to gradually improve these maps using specialized methods (e.g., CAM, Grad-CAM, CRF). The resulting pseudo-labels are used to train classic segmentation models, such as DeepLab, U-Net, etc.

This approach reduces the need for high-precision annotations, which significantly reduces time and resource costs while maintaining high segmentation efficiency.

The methodology chose can be divided into three steps: creating activation maps for classes, improving maps with interpixel relational network, segmentation.

Creating activation maps for classification

One of the easiest ways to get an activation map of the desired detail is to add Average pooling (the average value for each feature map correction) and another layer to the neural network - a linear layer.

The scientific novelty of the work is the rapid detection of moving objects in conditions of poor visibility. To solve this task, it is necessary to solve the following tasks:

1) develop methods for rapid detection of objects

2) develop pre-processing methods based on filtering

3) conduct machine learning to establish the reliability of objects.

IV. HARDWARE:

The combat vehicle recognition model is shown in Fig. 1.

The performance (latency, delay, FPS) of a neural network for combat vehicle recognition depends on a number of factors, including the model architecture, hardware type, input data (image/video), and deployment conditions (edge/cloud). Below is a brief analysis:

1) YOLO (You Only Look Once) [7], is extremely fast, works in real time

YOLOv5 / v6 / v8. Frames per second, which determines the refresh rate of the image on the screen: ~30–120+ (on GPU), Latency: ~5–30 ms per image (on RTX 3060–3090). Suitable for working with drones, UAVs, camera systems

EfficientDet is more accurate, but slightly slower. Frames per second, which determines the refresh rate of the image on

the screen ~30–60 ms. Suitable for applications where a compromise between speed and quality is required.

TABLE I. CAMERAS EQUIPPED IN TANKS.

Equipment	FPS (YOLOv5s)	Latency
NVIDIA Jetson Nano	~5–10	~100–200 мс
Jetson Xavier NX	~20–40	~30–60 мс
RTX 3060–3090	60–200+	~5–20 мс
Edge TPU (Coral)	~15–30	~33 мс

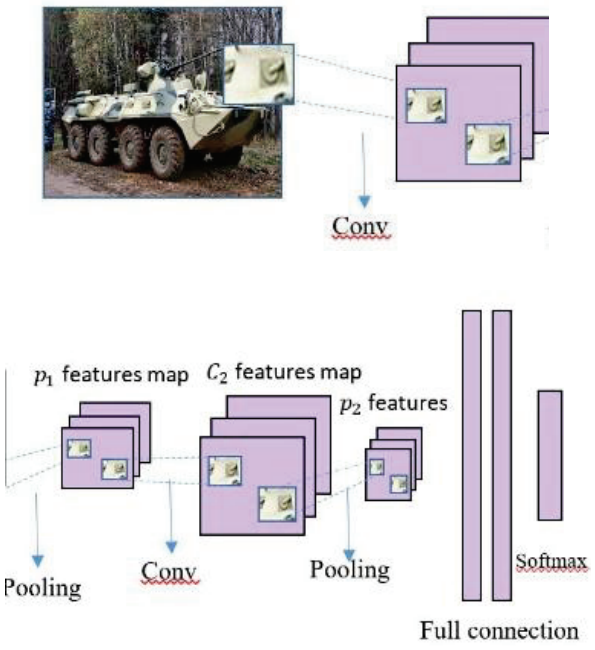


Fig. 1. Combat vehicle recognition model

NVIDIA Jetson (show Table 1) [8] Nano is a system-on-module (SoM) and development kit designed for building artificial intelligence (AI)-enabled devices. It is a compact, high-performance computer that allows you to process data from sensors in real time and run neural networks. Jetson Nano is used in various applications, including robotics, autonomous driving, video analytics, and other projects that require processing data from sensors and AI.

Key features of NVIDIA Jetson Nano [9]: Processor: Quad-core 64-bit ARM Cortex-A57. GPU: 128-core NVIDIA Maxwell-based GPU. Memory: 4 GB LPDDR4. Performance: 472 GFLOPS. Interfaces: Gigabit Ethernet, MIPI CSI-2, PCIe, and other high-speed ports. OS support: Linux for Tegra.

NVIDIA Jetson Xavier NX 16GB [10] Module is a compute module that delivers up to 21 TOPS of AI computing power, consumes just 10 watts of power, and has a compact form factor. Jetson Xavier NX is a high-performance artificial intelligence (AI) module designed to perform deep learning and computer vision tasks in resource-constrained environments. The module combines a 6-core ARM Carmel processor, an NVIDIA Volta GPU with 384 CUDA cores and 48 Tensor cores, and two NVDLA hardware accelerators, delivering up to 21 TOPS of computing power while consuming 10-20 watts.

With high RAM bandwidth (up to 59.7 GB/s) and support for multiple I/O interfaces (USB 3.1, PCIe Gen4, MIPI CSI 2), Jetson Xavier NX efficiently processes high-definition video streams in real time, including up to 44 streams of 1080p@30 fps when decoding (H.264/H.265).

The module supports up to 6 simultaneously connected cameras, making it an ideal solution for embedded visual monitoring systems, autonomous robots, unmanned aerial vehicles and mobile recognition platforms. The platform is powered by a full NVIDIA JetPack SDK stack, which includes the TensorRT, CUDA, DeepStream, PyTorch, TensorFlow frameworks and other tools for rapid development of AI applications.

The compact dimensions (69.6 × 45 mm) and support for intensive workloads in complex temperature conditions (up to +90°C) make Xavier NX a universal solution for mission-critical edge AI applications.

Edge TPU is a specialized artificial intelligence processor (ASIC) developed by Google for high-speed neural network inference without connection to the cloud. It is integrated into the Coral series of devices and optimized for running quantized models in the TensorFlow Lite (int8) format. Edge TPU provides up to 4 TOPS of computing performance with minimal power consumption - less than 2 W, which allows you to implement real-time object recognition in resource-constrained conditions. Typical model inference time (for example, MobileNet SSD) is less than 10 ms.

V. EXPERIMENTS

Deep Convolutional Neural Networks have proven their effectiveness in solving semantic segmentation problems. An identification system has been developed, which has been extended based on Internet access, which is described in [12]. The system protection elements described in [13, 14] have been used.

We have developed a dataset containing about 3000 images with labeled data.

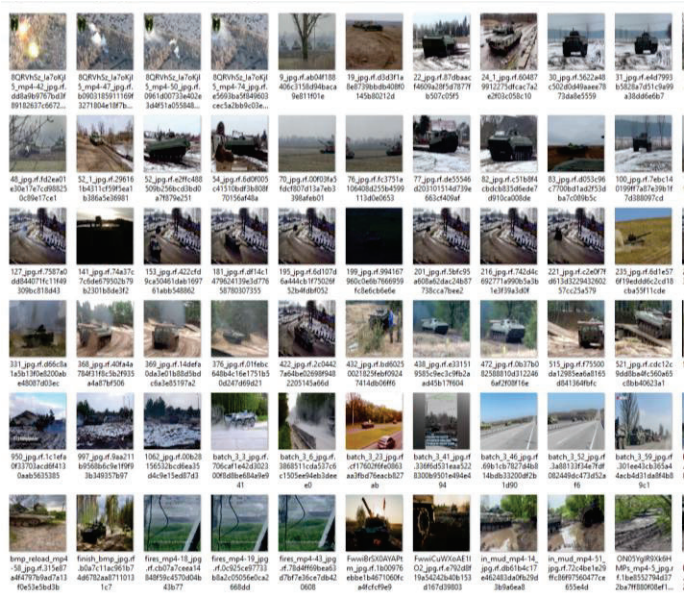


Fig. 2. Dataset view

The dataset contains 138 images of tanks (show Fig.2)

One image is from 53 kB. Each image is formed by 640X640 pixels. The images are marked (show Fig.3).

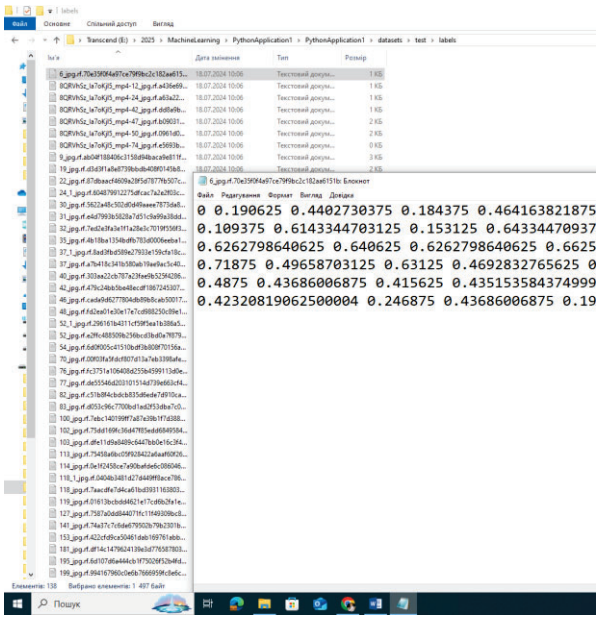


Fig. 3. Image segmentation

In the image segmentation stage, the region is selected in such a way that the image is divided into different components based on the homogeneity of features. There are a number of approaches to segmentation in the literature, such as grid-based, clustering-based, contour-based, model-based, graph-based, and region-based methods. One block often consists of parts of visually different objects. It is difficult to determine the size of the blocks for image representation. Segmentation algorithms based on statistical models are known (show Fig.4).

The following results were obtained.

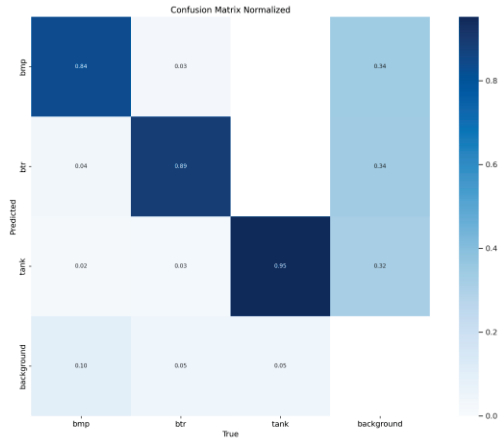


Fig. 5. confusion matrix for recognizing military equipment

The recognition results are shown in Fig. 6-9.

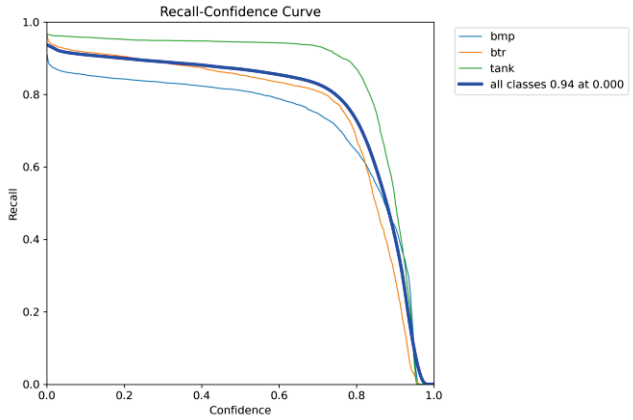


Fig. 6. Recall Curve

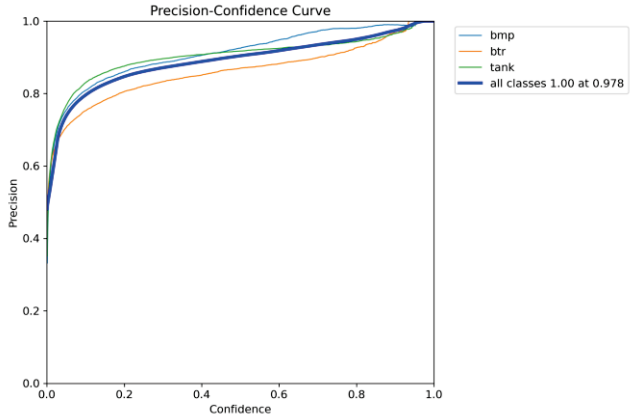


Fig. 7. Precision Confidence Curve

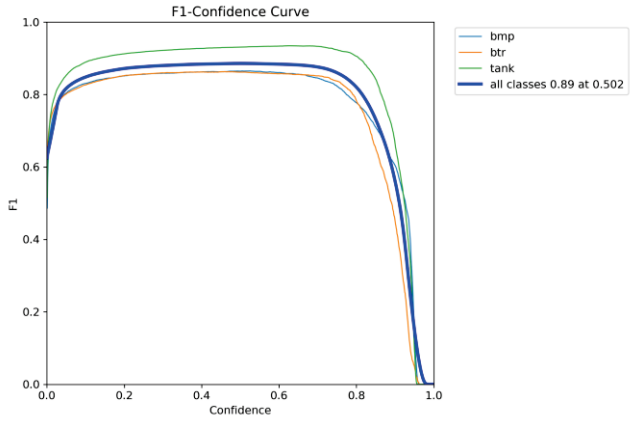


Fig. 8. F1 Curve

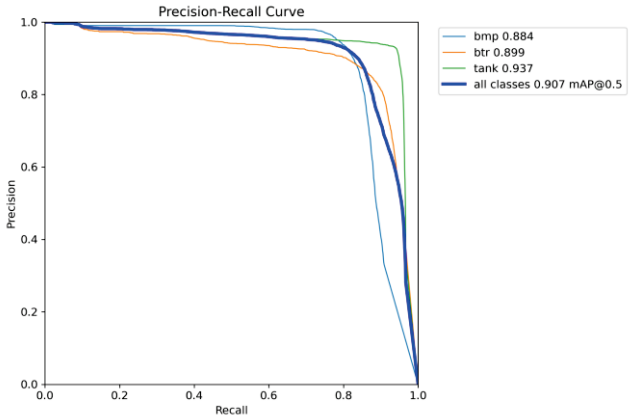


Fig. 9. Precision Recall Curve

VI. Conclusions

The task of quickly detecting moving objects in poor visibility conditions has been solved. For this purpose, methods for quickly detecting objects have been developed, pre-processing methods based on filtering have been developed, and machine learning has been performed to establish the reliability of detected objects.

A dataset of combat vehicles used in military operations has been built. Types of military equipment and their modifications have been analyzed.

A model has been created that recognizes military objects in real time.

The quality of the model has been investigated and their characteristics have been built.

Acknowledgment

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Analysis of Methods for Calculating Solar Energy on an Arbitrarily Located Surface

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***Abstract*— This paper presents a comparative analysis of two methods for calculating solar irradiance on an arbitrarily inclined surface: a coefficient-based model developed by the author, and a simplified model proposed by Professor Konrad Mertens. Both approaches are evaluated for a specific location (48.606913°N, 9.089694°E) on June 16, 2024. The study includes the calculation of the solar declination angle, the zenith and incidence angles, and the resulting global solar irradiance on a tilted plane. Real atmospheric transparency data (KT) and NASA-based solar constants were used. The results demonstrate that the author's method, which includes time correction and angular coefficients, provides a more accurate estimation—within 0.1% of NASA's ephemeris-based declination and approximately 3.9% higher irradiance values compared to the Mertens method. A diurnal simulation confirms the higher precision and practical applicability of the author's model for forecasting solar energy output.**

Keywords*— *solar irradiance, tilted surface, solar declination, Cooper formula, Mertens method, atmospheric transparency, solar energy modeling, solar geometry, time correction, photovoltaic forecasting

I. Introduction

In 2021, the European energy market is pursuing an active policy of abandoning carbon fuels and switching to alternative energy sources. Alternative energy is unregulated, and therefore its stabilisation can only be ensured by skilful accumulation during periods of surplus and modern forecasting [1].

The energy grids of leading European countries such as Germany, Switzerland, Austria and Italy are increasingly resembling the blocks being prepared for the Micro-Grid concept. For example, starting in 2022, most solar power plants will be equipped with storage batteries and an energy management system that allows consumption to be prioritised and is specially designed for the conditions of dynamic electricity prices in the grid.

However, forecasting electricity generation is an integral part of energy management for alternative energy installations. At the current stage of development of this topic, all forecasting is based on meteorological forecasts and calculations made on their basis, but this approach does not provide sufficiently accurate results to be relied upon for energy network management.

One of the key tasks in forecasting is to accurately determine the amount of energy delivered by energy sources to an arbitrarily oriented converter surface.

Since most of the electrical energy in the alternative energy sector in continental Europe is produced by solar installations, an algorithm specifically for this type of installation is proposed for consideration.

For this reason, the aim of this work is to define and refine algorithms for determining the energy supplied by the sun to arbitrarily oriented solar power plants in order to improve the quality of forecasting in energy management.

II. Scientific materials analysis

The key component for modelling the operation of a solar power station is a model of actual solar radiation.

Among the methods for modelling solar radiation, the most popular among domestic and, often, foreign scientists is the model developed by Professors Jordan and Liu [2-5].

Its essence lies in representing solar radiation as the sum of direct, scattered and reflected solar radiation incident on an arbitrarily placed horizontal surface:

$$I_{sum}^{\beta\gamma} = I_{dir}^{\beta\gamma} + I_{diff}^{\beta\gamma} + I_{ref}^{\beta\gamma} =$$

$$= I_{dir}^{h} \frac{cos\theta}{cos\theta_z} + I_{diff}^{h} \frac{1+cos\beta}{2} + \rho I_{ref}^{h} \frac{1-cos\beta}{2}, \quad (1)$$

where $I_{dir}^{\beta\gamma}$, $I_{diff}^{\beta\gamma}$, $I_{ref}^{\beta\gamma}$ - total, direct and scattered from clouds and aerosols, as well as reflected from the Earth's surface, applied to the surface, respectively; I_{dir}^{h}, I_{diff}^{h}, I_{ref}^{h} - the same for a horizontal surface; $cos\theta_z$ – zenith angle of the Sun; θ – angle between the direction of solar radiation to the surface and the normal to it; ρ – albedo of the Earth's surface;

Professor Konrad Mertens from Hochschule Münster uses a similar approach in his work [6], presenting the total solar radiation applied to an arbitrarily located surface as the sum of direct, scattered and reflected solar radiation:

$$E_{Gen} = E_{dir_{Gen}} + E_{diff_{Gen}} + E_{refl_{Gen}} =$$

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$$= E_{dir_H} \cdot \frac{\sin(\gamma_S+\beta)}{\sin\gamma_S} + E_{diff_H}\frac{1+cos\beta}{2} + E_{refl_H}\frac{1-cos\beta}{2}\rho \quad (2)$$

Here $E_{dir_{Gen}}$, $E_{diff_{Gen}}$, $E_{refl_{Gen}}$ - total, direct and scattered from clouds and aerosols, as well as reflected from the Earth's surface, applied to the surface, respectively; E_{dir_H}, E_{diff_H}, E_{refl_H} - the same for a horizontal surface;

At first glance, the methods appear similar, but the difference lies in the details.

For the first method, the necessary input data are the latitude of the location of the receiving surface j, the angle of inclination of the Sun δ, the angle of inclination of the receiving plane to the horizontal plane β, the azimuth angle of the receiving plane γ, the difference between local official time and Greenwich Mean Time ΔT_{UTC}, the geographical longitude of the receiving plane ψ, the calendar day number N, the average value of the atmospheric transparency index for the selected location K_T.

Based on these data, it is possible to determine the coefficients A, B, C, D, E, which are necessary to find the values used to convert the formula for calculating the total solar radiation on a horizontal plane into a formula for an arbitrarily located plane.

In Mr. Mertens' method, the transition to coefficients is not made because the data from the coefficients are directly incorporated into the formulas for calculating some variables, and the variables are calculated without taking into account the time correction for the time equation and without taking into account leap years. Also missing

Thus, in the method [7-9], it is calculated using the formula:

$$\omega = (WOZ - 12) \cdot 15°$$

Here $WOZ = 12 - 1 \cdot \left(1 - \frac{\Lambda}{15°}\right)$, and Λ – latitude.

Professor Martens suggests determining the midday altitude of the Sun, called , based on the latitude of the location and the angle of the Sun's altitude for each day. For example, the maximum angle of the Sun's altitude on the day of the summer solstice will be [6, 7]:

$$\gamma_S = 113{,}4° - \varphi$$

Meanwhile, the proposed methodology suggests switching to coefficients that physically reflect the position of the Sun relative to various bodies: the Earth's surface, an arbitrarily placed surface, a horizontal surface, etc.

Thus, the coefficients are determined using the following formulas:

$$A = sin\varphi cos\beta; B = cos\varphi sin\beta cos\gamma;$$

$$C = sin\beta sin\gamma; D = cos\varphi cos\beta; E = sin\varphi sin\beta cos\gamma;$$

Next, using the formulas below, find all the necessary data for formula (1):

The angle is determined by solving the equation:

$$cos\theta = (A - B)sin\delta + [Csin\omega + (D + E)cos\omega]cos\delta \quad (3)$$

The zenith angle of the Sun is determined by the expression:

$$\theta_z = \arccos[sin\varphi sin\delta + cos\varphi cos\delta cos\omega] \quad (4)$$

The angle of the Sun's elevation above the horizon h is calculated using the formula:

$$h = 90° - \theta_z. \quad (5)$$

The hour angle is calculated using the following formula:

$$\omega = 15(t - 12 - T_{tr} - \Delta T_{UTC}) + \psi, \quad (6)$$

Here t – current official local time, hour;

The time correction for the time equation is determined by the formula:

$$T_{tr} = \frac{1}{60}[9{,}87\sin(2B) - 7{,}53\cos(B) - 1{,}5\sin(B)]. \quad (7)$$

here $B=B_0(N-81)$, deg.; $B_0=360/365$; N – number of calendar days since the beginning of the year (for 01.01 – $N=1$, for 02.01 – $N=2$, etc.).

The angle of the sun's declination is a fairly important parameter for calculation. This parameter requires additional consideration, as there are two methods of calculation.

The first method describes the angle of declination of the Sun as follows:

$$\delta = -23.45 \cdot \cos(B_0 \cdot (N + 10))$$

This method gives results close to the true value, but the accuracy is insufficient, since this is the first version of Cooper's empirical formula.

The following option provides a more accurate picture:

$$\delta = 23{,}45\sin[B_0(N + 284)]$$

When calculating the angle of declination of the Sun using the formulas for 16 June 2024, the results were 23.03° for the first formula and 23.35° for the second. The actual value of the Sun's declination angle for this day, according to the Skylive website, which uses NASA ephemerides with an accuracy of 0.01° for its calculations, is 23.344°.

Thus, it is evident that the second formula provides a fairly accurate result and is acceptable for use in engineering calculations. Spencer's method was also considered, which is not widely accepted, but gives an even lower error than Cooper's method. However, at this stage, it does not make sense to consider such a complex method, but in the future it may play a role in improving the accuracy of forecasting.

The accuracy comparison chart for the two methods is shown in the figure below.

We also performed a series of verification calculations to check the maximum and minimum error limits of Cooper's formula. According to the results, it ranges from 0.026° to 0.3°. The largest error is observed in calculations for autumn and spring. The accuracy increases as we approach the winter and summer solstices [10-13].

Professor Mertens does not propose any methodology for calculating the angle in his work, recommending the use of standardised tables of values.

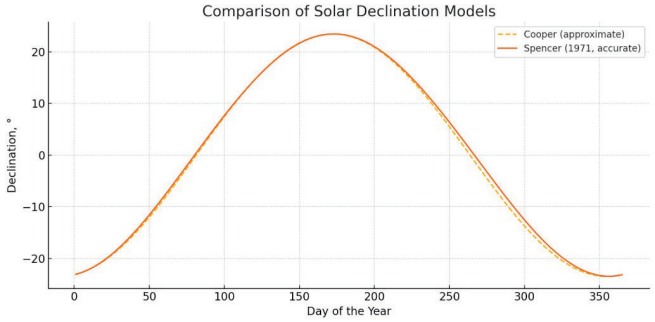


Fig 1. Comparison of Solar Declination Models.

To determine the angles of sunrise and sunset according to solar time, we will use the formula:

$$\omega_{sr} = 0 - arccos[-tg\varphi tg\delta];$$

$$\omega_{ss} = 0 + arccos[-tg\varphi tg\delta]. \quad (8)$$

Real conditions are taken into account using the average value of the atmospheric transparency index K_T, which we obtained from the Meteonorm 7.2 meteorological database. In this case, it is assumed that the transparency index for current daily values of solar radiation is equal to its average daily value for a given month:

$$K_T = \frac{I^h_{sum}}{I_0}, \quad (9)$$

here I_0 – extra-atmospheric radiation on a horizontal surface, determined by Spencer's formula [13]:

$$I_0 = I_{sun}[1 + 0{,}033\cos(B_0 N)]cos\theta_z, \quad (10)$$

here $I_{sun} = 1367\ W/m^2$ – sunny constant.

The diffuse coefficient K_D is used to determine the amount of scattered radiation. This coefficient characterises the proportion of scattered radiation in the total radiation:

$$K_D = \frac{I^h_{dif}}{I^h_{sum}}. \quad (11)$$

The diffuse coefficient is determined using NASA methodology, according to which the proportion of scattered radiation is determined from a system of polynomial equations as a function of latitude, transparency index, time angle of sunset and noon angle of the Sun above the horizon.

The model is calculated using the following algorithm: calculation of the Sun's elevation and azimuth angles, calculation of solar sunrise and sunset times, calculation of the angle, calculation of the diffuse coefficient, calculation of direct and solar radiation, calculation of the time angle according to local official time.

To take cloud cover into account, the model uses a random number generator, which we consider to be a significant drawback of the model, although the error is 12%. However, for our calculations, we will use actual cloud cover data for 2023/2024.

III. The Modeling

To verify both methods, it was decided to implement both models in Matlab Simulink and perform modelling for a full day on 16 June 2024 at a random point in Germany. The coordinates for modelling were set at 48.606913°N, 9.089694°E.

Due to the scale of the model and to maintain its readability, the above models were implemented in Simulink as subsystems. Examples of the implemented blocks are shown in the figure below.

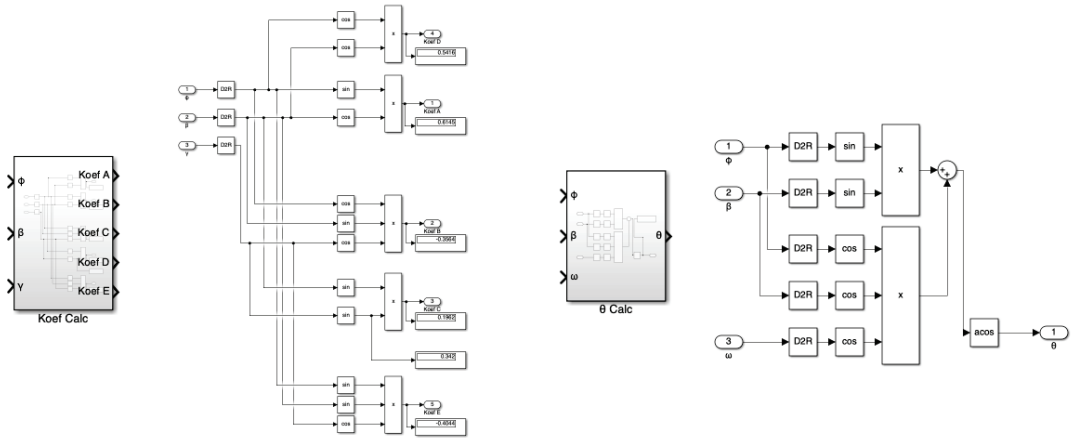


Fig. 2. An example of implementing coefficient and angle θ calculation blocks in Simulnk.

The aim of the simulation was to evaluate and compare the accuracy of two approaches for calculating total solar radiation on an inclined surface during a typical summer day.

The simulation was performed for the geographical coordinates 48.606913° N, 9.089694° E, corresponding to the city of Reutlingen, Germany. The simulation date was set to 16 June 2024 (167th day of the year) with an inclination angle of 35° and a south orientation (azimuth = 0°). The atmospheric transparency index (KT) for this location was set to 0.65 based on average Meteonorm data for June. An albedo coefficient of

0.2 was used to account for reflected radiation. In accordance with NASA recommendations, the solar constant was assumed to be 1367 W/m². Both models were implemented to calculate direct, diffuse and reflected solar radiation on an inclined surface.

The proposed method included a detailed calculation of the geometry of the Sun using coefficients (A–E), the hour angle (ω), time corrections for the time equation, and an improved version of Cooper's formula for the Sun's declination (δ). In contrast, the Mertens model used a simplified approach without time correction and with radiation components directly corrected using geometric projection coefficients.

The daily simulation was carried out from 05:00 to 20:00 at 30-minute intervals. The simulation results are shown in Figure 3, which shows the solar radiation on an inclined surface calculated by both methods.

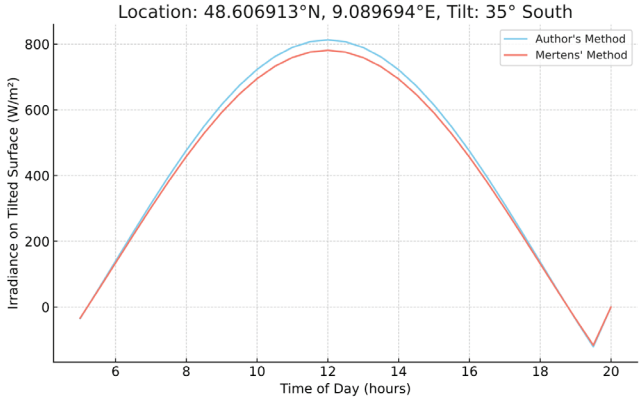


Fig. 3 Comparison of modelling results

The calculation according to the first method yielded higher radiation values, especially around solar noon, thanks to a more accurate determination of the Sun's position.

According to data from the Photovoltaic Geographical Information System, the actual level of solar radiation for the specified coordinates on 16 June 2024 was 809.1 W/m².

These data are necessary for further comparison of the simulation results.

The peak radiation intensity obtained using the author's method was 812.9 W/m², while the Mertens model gave a result of 781.4 W/m², which led to an underestimation of 3.9%.

These results show that the first model better corresponds to the actual solar radiation conditions and is more suitable for solving engineering problems related to the forecasting of solar power plant generation requiring accurate solar energy forecasting.

IV. Conclusion

This study has presented a comparative analysis of two methods for calculating solar irradiance on arbitrarily inclined surfaces: a coefficient-based approach proposed by the author and a simplified projection method developed by Professor Konrad Mertens. Both methods were implemented in MATLAB and simulated for a typical summer day at mid-latitude coordinates (48.606913°N, 9.089694°E) on June 16, 2024.

The author's method, which incorporates time correction, solar geometry coefficients, and an improved empirical formula for solar declination, demonstrated higher accuracy in estimating total solar irradiance on a tilted surface. The Mertens method, although simpler and computationally efficient, exhibited a consistent underestimation of irradiance, with a peak error of approximately 3.9% under clear-sky conditions.

A diurnal simulation confirmed that the author's model more accurately captures the dynamic nature of solar irradiance throughout the day. The results support the use of the coefficient-based model for engineering applications where precision in photovoltaic energy forecasting is critical.

Future work may include validation against measured data and the integration of real-time cloud cover and albedo variability to further enhance the model's predictive capability.

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Information-logical Modeling For Explosion-proof Electric Drives Design And Regulatory Compliance

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***Abstract*—This paper proposes a methodology for structuring the relationships between design and conformity assessment of explosion-proof equipment, with a focus on electric drives for mobile robotic systems. The study identifies typical errors made during the development of explosion protection systems due to complex and fragmented standards and regulatory requirements. To address these challenges, the authors utilize SysML-based information-logical models to represent the relationships between design documentation, regulatory requirements, and certification procedures. Through detailed examples of explosion-proof mobile robots, the article demonstrates how visual modeling improves the traceability of safety requirements, supports informed decision-making during concept selection, and reduces conformity assessment time and cost. The proposed approach enables coordinated interaction between design organizations, conformity assessment bodies, and testing laboratories, enhancing safety, efficiency, and standardization of explosion-proof system development.**

Keywords—electric drives, explosion-proof design, SysML, regulatory standards, information-logical models, ex-proof mobile robots

I. INTRODUCTION

Ensuring explosion protection in electric drives is extremely important for the safety of enterprises in a number of industries where potentially explosive environments are present: in particular, coal mines [1], [2], chemical and food industry enterprises [3].

The safety of this equipment must be assessed for compliance with the requirements of explosion safety harmonized standards, which gives a presumption of conformity with the Directive ATEX 2014/34/EU [4]. Manufacturers of explosion-proof equipment note that the need for such compliance makes designing for explosion safety requirements a complex process [5]. Unfortunately, the large number of standards, the complexity of the system of requirements contained in them and different concepts of explosion protection, changes and additions to requirements, their inconsistency with each other, lead to the fact that design engineers make mistakes when drawing up a list of requirements that they must take into account in specific designed equipment. Designers do not see the consequences of decisions they made at different stages of the explosion protection design process, and do not take into account the necessary subsequent testing and certification procedures. In addition to the equipment characteristics, this affects the timing and cost of certification. There is currently no well-known effective methodology for ensuring explosion safety at all stages of design and stages of the product life cycle, taking into account the specifics of its certification.

Solving these problems requires the use of scientific methods of analysis, systematization and algorithmization of processes for designing explosion protection of electrical equipment and the necessary subsequent assessment of its compliance. Such a methodology should take into account the requirements from various sources of technical information and regulatory documents of current legislation.

A number of types of documentation used by designers set requirements included in the technical direction for the design of a product, for example, regulatory documentation for selected types of explosion protection (standards). Another part of the design documentation confirms compliance with these requirements (for example, equipment layout diagram, list of elements, calculations, basic and functional electrical diagrams, etc. Many requirements concerning explosion safety and contained in standards are confirmed during conformity assessment only according to the manufacturer's technical documentation, and only some of these requirements are confirmed by type tests. Search, structuring and description of the links between different types of design documents, requirements and design solutions is a necessity for the algorithmization of explosion-proof equipment design processes. Visual representation of the relationships between documents establishing requirements and documents satisfying these requirements helps to reduce errors in design and, as a result, eliminates inconsistencies in testing and certification of explosion-proof equipment.

II. RESEARCH PROBLEM

A promising approach to displaying and analyzing these relationships is the construction of information-logical models of design processes and design solution options. In particular, such models allow the construction of a design language SysML [6], widely used in mechatronics. With the help of requirements diagrams built using this language, collecting, accounting and comparing different requirements for equipment is significantly simplified. The designations of project artifact relationships available in the language help to accurately describe the relationships between different documents, requirements, and technical solution options, taking into account design constraints and the interaction of design engineers, certification engineers, and test engineers.

A visual graphical representation of these relationships by SysML means and their subsequent analysis, with the aim of improving design processes, is the main problem we are researching. In graphical form, this problem is presented in Fig. 1 by the "mutual influence" block linking the design and conformity assessment processes supported by the explosion safety methodology:

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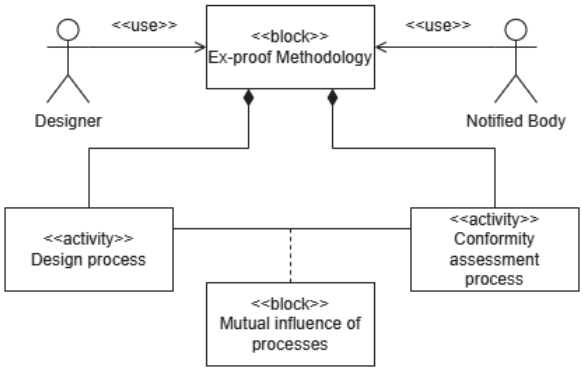


Fig. 1. Model of research problem

A similar approach is used, for example, in the work on systematization of the design of electric motors [7], where the constructed models demonstrate the effectiveness of a systematic approach to developing requirements and taking into account their impact on the design stages. But its application for structuring requirements for explosion-proof equipment, taking into account its subsequent conformity assessment, is innovative. Let us consider such structuring using specific examples, highlighting the types of complexity of design tasks and ways to solve them. We will take examples from different application areas, but mainly analyze design solutions for electric drives of mobile mining robots.

As many researchers and manufacturers note, the introduction of explosion-proof robots into potentially explosive environments increases the efficiency and safety of such facilities. For example, the introduction of robots for coal mining in underground mines is a much-needed scientific and technological solution for the modernization of the global coal energy industry. The following classification and areas of application of robots in mining environments can be given: rescue robots, Security control robots, Transport robots, Tunneling robots, Coal mining robots [8].

Also, as Mitsubishi Heavy Industries, Ltd. notes, the introduction of patrolling ground robots at petrochemical plants and nuclear power plants helps increase the frequency of patrols to ensure reliability and save labor. The problems of designing explosion protection in such products are also noted. [9].

Typical representatives of such robots are:

1) MINIBOT-II is a coal mine rescue robot [10]. Based on the description of the design of this robot, the explosion protection of its electric drive elements consists of placing them in a flameproof enclosure “d”. Since the explosion protection marking is not provided by the authors, and only the explosion protection type “d” is described, it can be concluded that this robot is not intended for work in mines in the presence of methane, and its explosion protection level (EPL) is Mb;

2) LURKER-3 - coal mine detection robot made by the Robotics Research Centre of Shandong University, China [11]. As a type of explosion protection used positive pressure enclosure with high pressure gas tank which is full of inert gas such as CO2 or N2. Since other means of explosion protection are not indicated in the description of the robot, it can be assumed that the explosion protection concept “pxb” was used, corresponding to EPL Mb.

3) GMRI mobile inspective mining robot developed by the Institute of Innovative Technologies EMAG and the Industrial Research Institute for Automation and Measurement PIAP. The category M1 (EPL Ma) was achieved by the compressed-nitrogen pneumatic drive and through limiting the speed of movable elements. Decompressed nitrogen cools the inside of the actuators, which becomes an extra security measure against overheating of friction elements. Electronic assemblies of GMRI were designed for the “ia” intrinsic safety concept [5].

4) EX ROVR "ASCENT" developed by Mitsubishi Heavy Industries, Ltd [9]. The robot is designed for patrolling petrochemical plants and nuclear power plants with the possibility of periodic occurrence of explosive environments (zone 1) group IIB temperature class T3. Li-ion battery of this robot is protected by an explosion-proof casing “d” and the other electric drive components protected by the explosion protection concept “pxb”.

We will show examples of constructing fragments of information-logical models that simplify the analysis, selection and approval of design solutions for explosion-proof electric drives elements of such robots. Effective system development requires coordinated interaction between all parties involved in the process, such as the design organization, conformity assessment body and testing laboratories. This requires early identification of all entities involved in development, testing and certification. Coordination of their roles, responsibilities and information exchange ensures process control and compliance throughout the entire life cycle of the system [12].

III. RESEARCH RESULTS

We will begin the modeling with an analysis of the organizational structure of the projects, since successful interaction of specialists in different fields of knowledge is one of the main problems of ensuring explosion safety. A typical example of such an organizational structure is given in Fig. 2:

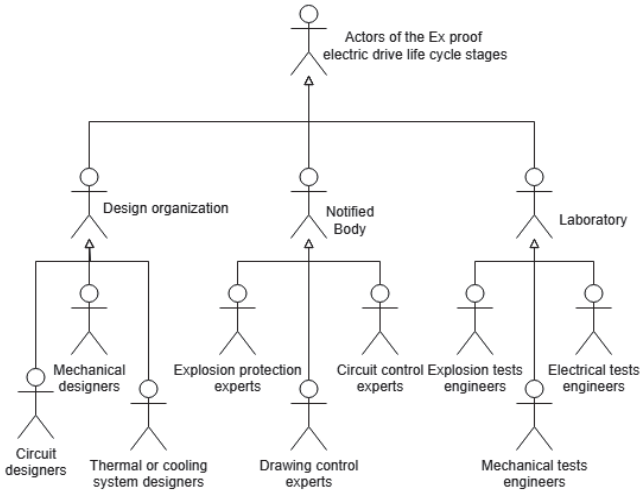


Fig. 2. The model of organizational complexity

Next, it is necessary to link the actors with the design documents they create and use, and to do this, as the first step, should be determined the structure of the design documentation.

Since equipment for explosive environments falls under the legislation [4] and must comply with the requirements of standards, it makes sense to display in the form of a model which documents contain requirements for explosion protection, so that they can be taken into account in the design documentation. Such relationships between this type of documents are shown in Fig. 3.

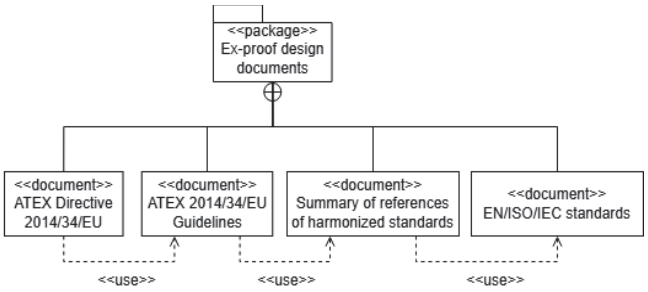


Fig. 3. Model of documents for explosion protection legislation

In a similar way, other regulatory documents used in explosion protection design can be structured in the model. Next, the model must show the connections between explosion protection requirements and the structure of the design documentation.

At the first stage of design, it is necessary to specify explosive environments and zones for which explosion protection must be provided. To structure this information and link the requirements to regulatory, design and certification documentation, this most important first stage can be also presented in the form of the following information-logical model in SysML shown in Fig. 4:

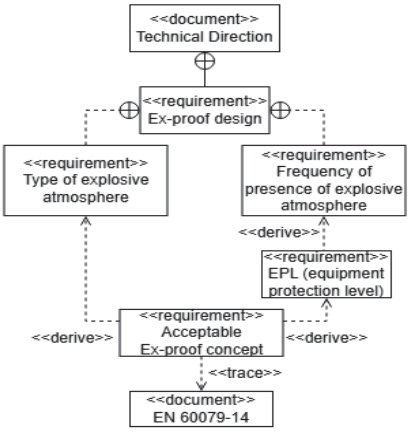


Fig. 4. Relationships of requirements influencing the choice of explosion protection type

Let us explain the meaning of the SysML language notations used in Fig. 4. When design engineers receive a task to ensure explosion protection, the first step should be to clarify the requirements for the type, composition and frequency of presence of an explosive environment. The Technical Direction document is considered a container of requirements and the inclusion of requirements in it is shown by a line ending with a plus sign inside a circle.

The minimum explosion protection level that is acceptable for implementation in equipment depends on the frequency of presence of an explosive atmosphere, therefore the requirement for the equipment explosion protection level (EPL) is derived from the requirement for the frequency of presence of an explosive atmosphere (line with the text <<derive>>). The explosion protection concept acceptable for selection in equipment is derived from the requirements for the explosive environment (mines or gas, steam, dust) and its frequency of presence (in the case of gas, steam or dust). The relationship between the types and levels of explosion protection can be found by design engineers in the standard [13], their connection with the regulatory document is shown by the line <<trace>>.

A separate question arises regarding the parameters of explosive environments during the equipment design. Due to the different ignition sources [14], [15], not only the concept of explosion protection, but also the requirements contained in the standards for specifically selected methods of ensuring explosion protection, depend on the explosive environments. An example of such requirements can be the temperature class of equipment (T1-T6), since each explosive environment has its own autoignition temperature. This is one of the parameters that electric drives engineers must monitor and model when ensuring explosion protection of electric drives. Moreover, depending on the type of explosion protection and the method of its implementation, compliance with the temperature class will depend on the compliance of the temperatures in the different parts of the electric drives components (the external temperature of the enclosure surface or the temperature of the hottest component of the control system or motor inside the enclosure). Also, each explosive atmosphere is characterized by minimum currents and ignition energies [16], [17], [18]. This is important for electric drive elements with an explosion protection type of intrinsically safe electrical circuit “i”. Therefore, it makes sense to separately trace the links with regulatory documentation on the parameters and classifications of explosive environments in the following information-logical model (Fig. 5):

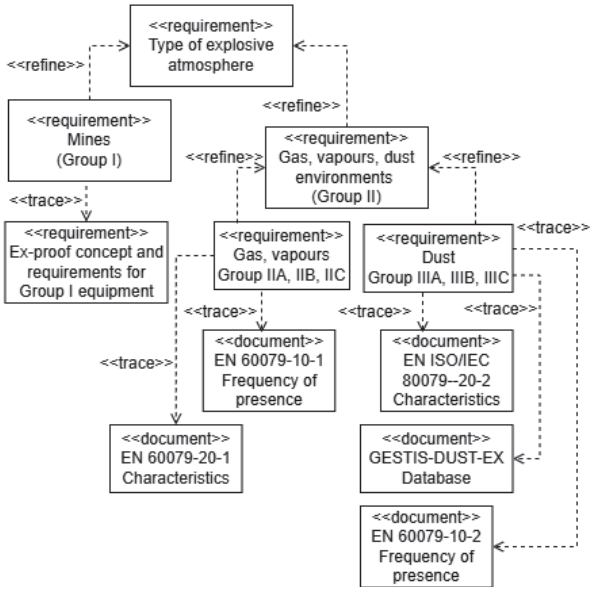


Fig. 5. Relationships between types of explosive atmospheres and regulatory documents

In Fig. 5, using the line with the text <<refine>>, it is shown that explosive environments, and therefore equipment groups according to [4], are divided into mine and any other gas/vapor/dust-air explosive atmospheres. Since the chemical

composition of explosive mine environments is known, and the frequency of presence cannot be predicted due to the problems of sudden methane emissions in mines, no specific regulatory documents are provided for determining the characteristics of mine explosive environments. During the explosion protection design for mine equipment, engineers must clarify whether the equipment should be designed to operate in the presence of methane (category M1) or should be switched off when it is detected in the mine (category M2). Next, in accordance with this, the permissible level of explosion protection is selected (Ma or Mb). The type of explosion protection is selected in accordance with [13], shown in the model fragment in Fig. 3, and the requirements in the regulatory documents are taken into account in accordance with group I.

Gas/vapor-air explosive atmospheres belong to group II, and dust explosive atmospheres belong to group III; these clarifications by types of environments are displayed in the model as <<refine>>. Connections of explosive atmospheres with the corresponding regulatory documents by characteristics and calculations of frequency of presence are displayed as <<trace>>.

Such information-logical models will help to build a set of requirements for the identified explosive atmospheres for the designer, which must be included in the Technical direction for the product. These requirements concern both the temperature class, and the provision of the IP protection level (if IP is a means of explosion protection in the selected explosion protection concept), and other design features.

The document that finally acknowledges the conformity of the products with the ATEX Directive is the Declaration of Conformity, drawn up by the manufacturer and presented in Fig. 6, which shows the structure of the certification documentation with reference to organizations.

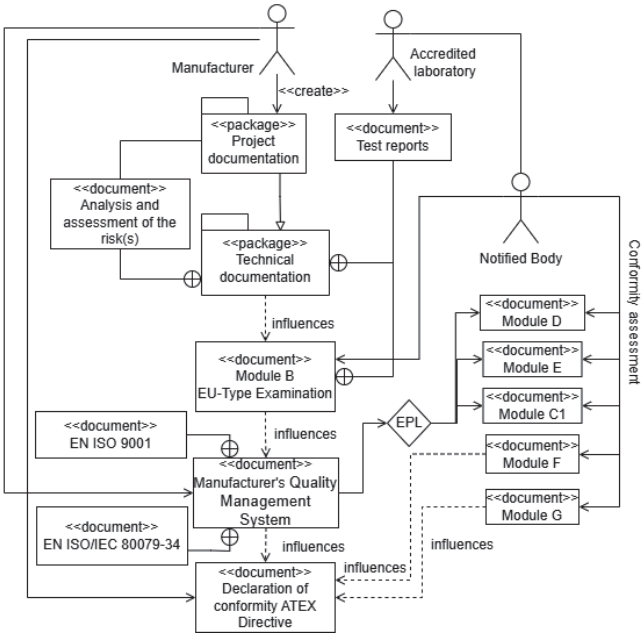


Fig. 6. Model of the structure of certification documentation

Since for the equipment with minimum explosion protection level, category 3, and non-electrical equipment of category 2 group 2, the manufacturer can carry out the conformity assessment himself using Module A, this is not taken into account in Fig. 5. In other cases, it is necessary to involve a notified conformity assessment body, which carries out a type examination and a second conformity assessment module, usually concerning the quality system [4].

The main package of documents submitted by the manufacturer for conformity assessment is the “Technical Documentation”, which includes all design and engineering documentation, the “Risk Analysis and Assessment” document and test reports. The notified body carries out a type examination, checking the technical documentation and the product for compliance with the requirements of each section and paragraph of the applicable standards for the explosion protection concept selected by the manufacturer.

Based on the results of such examination, the notified body draws up a program-methodology of type tests, which must be carried out by an accredited laboratory. Based on the results of these tests and verification of technical documentation, a list of special conditions and restrictions for the operation of products is drawn up. The content and program of control tests carried out by the manufacturer are also determined. These tests are part of the Quality Management System, which for manufacturers of explosion-proof equipment is regulated by a separate special standard [19].

Next, the model should include requirements for specific elements of electric drives, linking them to the relevant standards. Let's consider this using the example of explosion protection decisions of electric drive elements in the EX ROVR “ASCENT” robot. Its structural diagram is shown in Fig. 7:

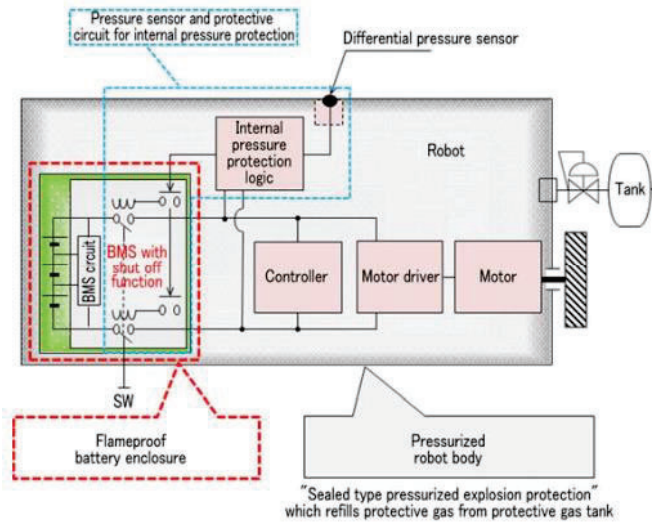


Fig. 7. Explosion protection structure of robot [9]

To analyze the explosion protection requirements for the electric drive elements of this robot, the following components were selected: power source (8-cells Li-ion battery) protected by flameproof enclosure "d", robot body (in which the control system, driver, motors are installed) - pressurized enclosure "p". The purpose of the analysis is to display the explosion protection requirements taking into account the design solutions adopted by Mitsubishi Heavy Industries, and to establish the links between these requirements and the design and regulatory documentation. An example of the result of such an analysis in the form of a model fragment is shown in Fig. 8.

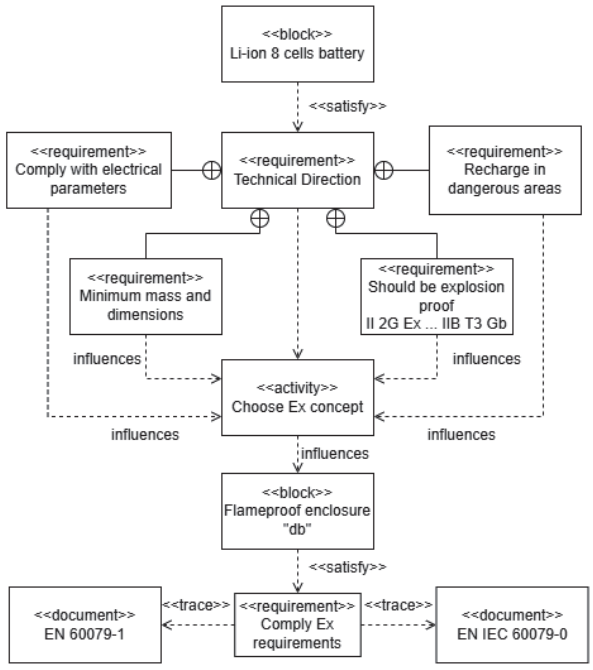


Fig. 8. Requirements diagram for robot battery

The battery must meet the requirements of the Technical direction - this is shown in the model by the <<satisfy>> connection. Further, the Technical direction include the requirements that battery must be as minimum as possible dimensions and weight (since these parameters are critical for the robot), must be rechargeable in a presence of explosive atmospheres, must comply with the electrical characteristics to power the drive system and auxiliary devices, and must be explosion-proof for gas/vapor-air explosive atmospheres with a periodic frequency of presence (Zone 1). When selecting the type of explosion protection, designers must take these requirements into account. The selection procedure is presented in the model by a block of the <<activity>> type.

The selected concept of flameproof enclosure “Ex db” is linked in the model with the standards applied to this protection concept, which give a presumption of compliance with the ATEX Directive [4].

When the connection with the regulatory documents containing the requirements for explosion protection of the selected concepts is established, it is possible to draw up a requirements diagram of the standard sections paragraph that will be applied to the designed element and its explosion protection means. This will establish which requirements should be taken into account during design and how they will be confirmed – by Project documentation or type tests and then control (if necessary) tests. An example of such a diagram is shown in Fig. 9.

This requirements model shows that the main requirements [20] applied to the selected explosion protection concept [21] include requirements directly to the battery itself and its explosion-proof enclosure. Requirements for batteries (e.g., type of its electrolyte, protective boards, etc.), requirements for enclosure materials, parameters of flameproof joints, marking, instructions, bolts and connections, entries into the enclosure - are confirmed by the design documentation, which is developed by the Manufacturer, and assessed by the Notified body. The laboratory is involved by the Notified Body only to conduct type tests, according to a test program-method, which is compiled based on the results of the examination of the Manufacturer's Project documentation

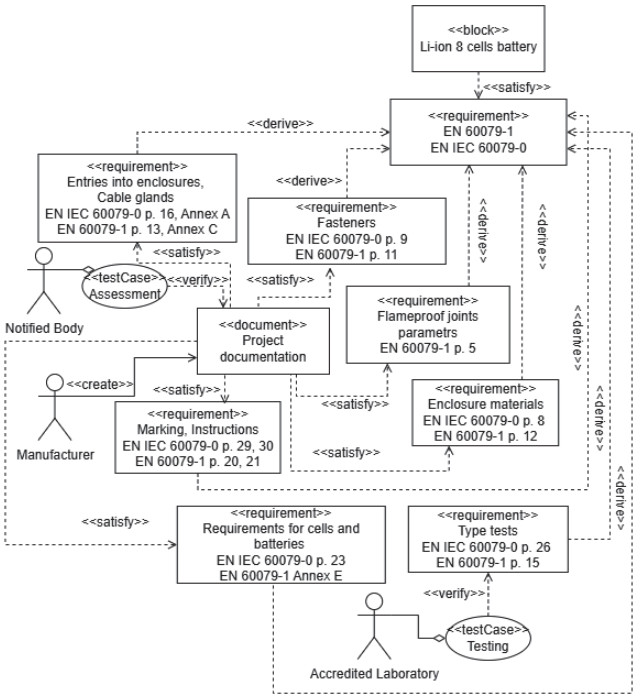


Fig. 9. Requirements diagram of EN IEC 60079-0, EN 60079-1

These requirements-are confirmed by the design documentation, which is developed by the Manufacturer, and assessed by the Notified body. The laboratory is involved by the Notified Body only to conduct type tests.

Since the cost of conformity assessment and the time it takes to conduct plays an important role, it is advisable to display which type tests confirming should be carried out by the laboratory. Such an example is shown in Fig. 10:

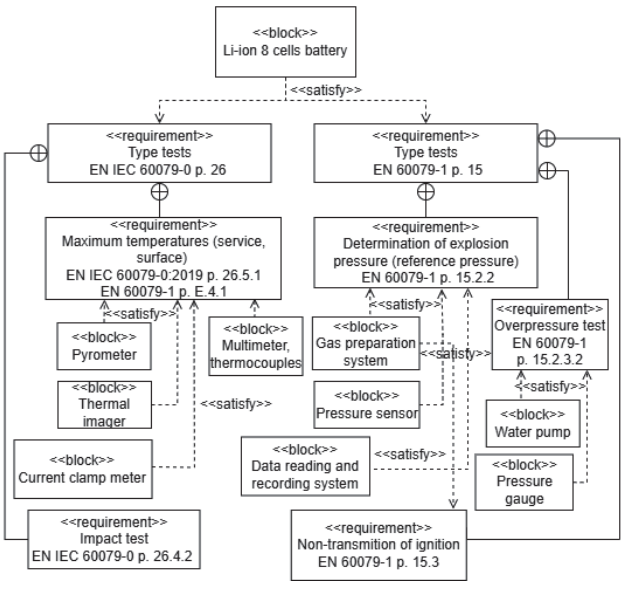


Fig. 10. Requirements diagram of type tests EN IEC 60079-0, EN 60079-1

The given model is incomplete, since the purpose of its creation is to display an example of the applied approach to the described problem. The model displays an example of the necessary type tests to confirm compliance with the selected explosion protection concept. The list includes tests of the flameproof enclosure, consisting of its mechanical strength, determination of internal explosion pressure (reference pressure), explosion resistance (overpressure test) and non-transmition of ignition test, as well as temperature tests of the battery heating in short-circuit mode. There are no IP tests for non-metallic parts, since the sealing ring (even if present) will not be a means of explosion protection for the flameproof enclosure, although in the case of classification by the manufacturer, this parameter may also be confirmed.

Thus, in Fig. 8-10, the logical traceability of the applied requirements for equipment depending on the selected concept of explosion protection of the battery is shown. It is also shown which requirements should be taken into account in the manufacturer's design documentation, and which type tests will be applied to type samples.

A similar approach can be applied to other elements of the electric drive, such as controller, motor driver, motor protected by the "pxb" concept, which allows them to be used in a non-explosion-proof design.

IV. CONCLUSIONS

The approach developed in this research aims to structure the links between design, regulatory, and certification documentation, incorporating explosion protection requirements for electric drives. This structuring is achieved through the formalization of the mutual influence between design and certification processes, utilizing SysML design language tools for explosion-proof electric drives, which constitutes the scientific novelty of this work. By modeling these links, the approach reduces the risk of non-compliance of designed equipment with the standards applied during conformity assessment, while also decreasing the time and cost of completing this process. The proposed modeling methodology make it possible to compare the feasibility of different types of explosion protection for electric drive components, considering their design characteristics such as weight, dimensions, and temperature class. In the future, the created SysML models can serve as a basis for synthesizing rational algorithms for designing electric drives intended for explosive atmospheres. Furthermore, the application of this approach facilitates the creation of type test programs for explosion-proof electric drive equipment and minimizes errors in compliance assessments performed by explosion protection experts. Ultimately, this contributes to improving product and enterprise safety.

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Regional Automation Server for Distribution Points

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***Abstract*— The article presents a comparative analysis of the reliability of relay protection of substations using decentralized microprocessor terminals and a centralized protection and automation server. The study focuses on the probabilities of failures and false alarms in various schemes: serial, parallel, and server-based, taking into account redundancy. The results confirm the relevance of the transition to centralized and software-defined protection systems for modern substations.**

***Keywords*— digital substation, relay protection, redundancy, centralized server, failure probability, reliability analysis**

I. Introduction

Despite the ongoing shift toward digital substations and centralized architectures, there is still a lack of probabilistic analysis comparing the reliability of decentralized microprocessor-based terminals with centralized relay protection servers. Existing works [1–11] mainly focus on architecture and communication aspects, but not on the probabilistic evaluation of failure and false tripping scenarios.

As an object of protection, we choose a substation, the protection of which can be made in two versions: using a set of microprocessor terminals relay protection and automation (hereinafter referred to as terminals) or using a central relay protection and automation server (hereinafter referred to as the server).

The subject of consideration is a comparative analysis of the probabilities of failures substation protection (non-disconnection by protection of damage, short circuit on the line) when using terminals or a server.

When comparing these two options, we will consider only the probability of failures terminals or server, without taking into account the probability of failures of other elements: measuring transformers, communication lines, switches.

The concept of the server was not born from scratch, it is a direct continuation of the continuous growth of the number of functions of one protection element.

On the basis of electromechanical relays, one protection consisted of several relays: current, time, intermediate, index.

Since solid-state relays, one protection, as a rule, already consisted of one relay.

- The transition to a microprocessor base has already led to a rapid growth in the number of protections in one terminal (dozens of protections), simultaneous performance of automation functions including same terminal.
- Accordingly, the logical continuation of this trend is the concept server where all substation protections are located. Such a concept has their supporters [1, 2].

The purpose of this work is to evaluate and compare the reliability of decentralized and centralized protection schemes for substations, highlighting the conditions under which centralized redundant servers provide superior performance.

This paper introduces a probabilistic reliability model for centralized automation servers with redundancy (1-of-3, 2-of-3 schemes) and compares it against decentralized terminal-based systems. The novelty lies in demonstrating that triple-redundant server architectures significantly outperform conventional sets of terminals in both failure resistance and false operation prevention.

II. Scientific materials analysis

Recent studies have investigated different aspects of digital substations, including the implementation of IEC 61850 communication standards and the design of process and station buses [7–11]. Considerable attention has been paid to redundancy of relay protection devices as a way to improve reliability, as well as to the integration of automation functions into microprocessor-based terminals [12–18]. Other works have focused on smart grid operation, service-oriented architectures, and communication security issues in distribution networks [19-21]. However, most of these contributions concentrate either on functional design and communication performance or on conceptual system architectures, without addressing the probabilistic reliability of protection schemes. The analysis of sequential and parallel arrangements of protection devices is usually limited to simplified models or qualitative comparisons. To date, there is no comprehensive probabilistic evaluation that compares the failure and false tripping rates of decentralized terminal-based protection against centralized redundant server architectures. This research aims to fill this gap by providing a quantitative reliability analysis of both approaches.

The study applies probabilistic modeling to evaluate the reliability of substation protection schemes. The analysis assumes equal failure probability for both terminals and servers (P) and introduces the probability of line faults (X). Analytical derivations are carried out for sequential and parallel configurations, considering both redundant and non-redundant architectures. The evaluation includes linearization of the obtained formulas with respect to X (assuming $X \ll 1$), which allows direct comparison of terminal-based and server-based protection systems under identical conditions. This approach enables a quantitative assessment of reliability improvements introduced by redundancy (1-of-3 and 2-of-3 voting schemes).

For a comparative analysis of the substation protection failure probabilities at using terminals or a server, we need the simplest substation scheme (Fig. 1). And to begin with, let's consider an analog substation: analog measuring transformers, copper cables, terminals (for each object of protection - in its own one or more terminals).

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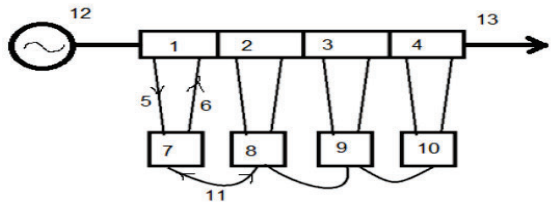


Fig. 1. The simplest substation scheme. Protected objects (1–4) are located on the same current line, connected by links 5,6 with terminals (7-10) which are also connected to each other by links communication 11. Substation power supply 12, consumers 13

Transition to digital substation, IEC 61850 standard (digital meters) transformers, fiber optic cables) little changes the configuration of the protection system: leaving the same set of terminals. In this case, a tire can be added process 14 and station bus 15, which combine all signals into one bus (Fig. 2).

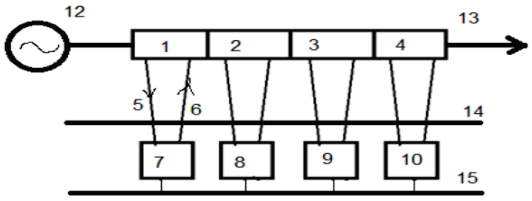


Fig. 2. The simplest circuit of a digital substation. Process bus 14 and station bus 15

Relay protection generates special bandwidth requirements for the bus process 14 and station bus 15:

- At 80 samples (current and voltage measurements in digital measuring transformers) per second each signal line 5 generates 14 in the process bus information flow 5–6 Mbit/s. With a large number of signal lines, 5 can logical segmentation of the process bus 14 [3] is required.

Station bus 15 must withstand increased information load - "Information storm" * [4].

Redundancy is the main way to improve the reliability of devices. Let the probability of device failure (for example, within one year) P be much less than one. Then, when two devices are working for one task, the probability simultaneous failure of two devices is the product of the probabilities of failure of each devices. For example: If P=10-2, then the probability of simultaneous failure of two devices is equal to $P * P = 10\text{-}4$, that is, the reliability of work with redundancy increases significantly. n relay protection, redundancy is used both to increase reliability of operation and to minimize false tripping.

There are two methods of near redundancy:

- Within the substation, redundancy of one relay protection terminal by another (different in function, and protecting this or another object) terminal.
- Two or more identical relay protection terminals (protecting one object and working on one switch) reserve each other. Such redundancy is also used in relay protection with special requirements for reliability.
- With redundancy by identical terminals working for one switch, the majority system (voting system) is connected*: if the terminals issue different control commands - which one to execute? For different tasks, be different scenarios of the majority system. For example: to improve reliability protection operation (circuit breaker trip), and to reduce excessive tripping of protections, the “2 out of 4” system can be applied (Fig. 3).

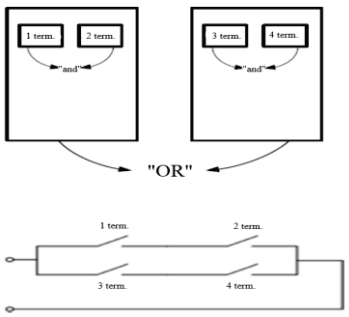


Fig. 3. The system consists of two cabinets*. Each cabinet consists of two functionally identical terminals with the performance of output actions according to the "AND" scheme to minimize false operation, leading to a shutdown, in case of failure of a single element. The cabinet operates according to the ‘OR’ scheme (according to the “two out of four” scheme)

Accordingly, to improve the reliability of the server, you can also apply redundancy: several identical servers (16, 17, 18) receive the same information and protect the same objects (Fig. 4).

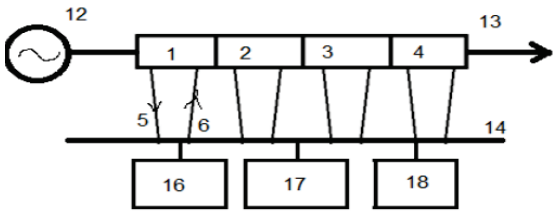


Fig. 4. Several identical servers (16, 17, 18) receive one and the same information and protect the same objects (1-4)

If there are three servers (Fig. 4), in addition to increasing reliability, the system receives additional benefits:

- Automatic detection of a damaged server: the damaged server the control signals will be different from those of the two healthy servers.
- It becomes possible to reserve all elements: for this, each server must be provided with its own communication line, instrument transformers,

switch.

- There is a possibility of "hot" replacement of servers, configuration, parameterization servers: without stopping the work of protections.

If one or more servers fail, a mode rebuild is required

operation of the majority system, for example:

- In normal mode "1 of 3". At the same time, the system is protected from complete failure (stop) two servers.
- In normal 2 out of 3 mode. At the same time, the system is protected from complete failure (stop) one server and is protected from incorrect operation (issuing a false command shutdown) of one server.
- If one server fails – “1 out of 2”. At the same time, the system is protected from complete one server failure.
- If two servers fail - "1 out of 1".

It is worth recalling that a complete server failure is usually detected by the Watchdog timer - a hardware-implemented scheme system hang control.

Process bus 14 loses relevance if there is a server: all protections are the ability to exchange information within servers 16, 17, 18. In addition, a single process bus 14 (non-redundant) reduces system reliability. Therefore, it is advisable to segment the process bus 14, for example (Fig. 5): separate segments (19–22) according to the number of protected objects (1–4). However, each segment (19–22) is connected to each of the servers (16, 17, 18).

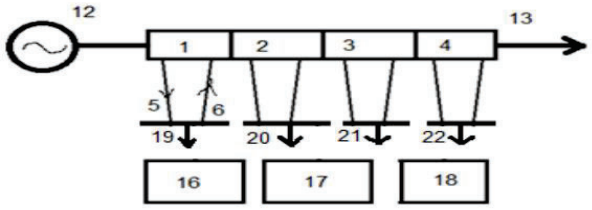


Fig. 5. Several identical servers (16, 17, 18), each server receives information and issues control commands, is associated with all segments

III. The Modeling

Figure 1 shows the configuration when the protections are on the same line, let's call it consistent. For consideration, we need one more terminal configuration: parallel, when several lines depart from one bus, while each line protected by one terminal (Fig. 6).

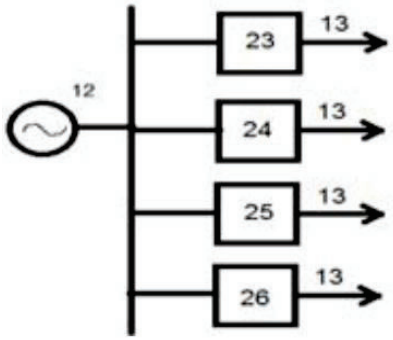


Fig. 6. The simplest substation scheme. Objects of protection together with protection terminals (23-26) protect 13 lines

At the same time, the schemes in Fig. 1 and Fig. 6 have a fundamental difference: in the scheme of Fig. 1 there is close redundancy, there is no redundancy in the diagram of Fig. 6. Thanks to redundancy, the scheme of Fig. 1 is more protected, we note the special role of terminal 7 (and damage on segment 1-2): there is no close redundancy for it, and this terminal alone largely determines the probability of substation failure.

To be able to compare two configurations, we make an assumption: the probabilities *P* terminal and central server failures are the same. We also need to enter the probability *X* of damage (for example, a short circuit) on the power line. For a series circuit, this will be the probability of occurrence of damage in the sections (1-2, 2-3, 3-4, 4-line), for a parallel circuit, this will be the probability of occurrence faults on lines 13, with protections 23, 24, 25, 26. When comparing terminals and server apply the following method: select the probabilities that are linear in *X* (in this case, we assume that that $X<<1$), then comparing two systems we can ignore *X* (reduce).

Let's start with a parallel circuit (Fig. 7), as the simplest one. Failure Probability protection against damage on one line is equal to $P \cdot X$, respectively, the probability that the failure of the protection of this line will not lead to the failure of the entire protection equal to $(1-P \cdot X)$. Since everything lines 13 are independent, so the probability that none of the defenses of n lines will result in to the failure of all protection is equal to $(1-P*X)^n$. From where we easily obtain the probability Y of protection failure:

$$Y = (1 - (1 - P \cdot X)^n)$$

Terminals 1 and 2 failed, terminal 3 is working, the rest of the terminals are not for us are important, the probability of such an event is $P2 \cdot (1-P)$. In this case, damage to sections 1-2 and 2-3, which will give three terms: $X(1-X)$, $(1-X)X$ and XX, after addition we get: $X(2-X)$. Respectively

$$Y = X \cdot (2 - X)P^2 \cdot (1 - P)$$

Terminals 1, 2 and 3 failed, terminal 4 is working, the probability of such an event is P3*(1-P). In this case, damage in areas 1-2, 2-3 and 3-4 is important, which will give seven terms, after adding which we get: $X(3-3X+X2)$. Respectively

$$Y = X(3 - 3X + X^2) \cdot P \cdot (1 - P)$$

Terminals 1, 2, 3 and 4 failed, the probability of such an event is P^4. In this case, important damage in all sections, which will give 15 terms, after the addition of which we get: $X(3-6X+4X^2-X^3)$. Respectively

$$Y = X(3 - 6X + 4X^2 - X^3) \cdot P^4$$

Adding all four cases and highlighting the linear parts, we get:

$$Y = X \cdot (P + P^2 + P^3)$$

Respectively

$$Y \sim (P + P^2 + P^3)$$

In addition to the failure of the terminal, a false positive of the terminal is possible: when in If there is no fault on the line, the terminal will issue a command to open the circuit breaker. Let us denote the probability of such an event as *Z*. Then for four terminals (for both schemes) the total probability of a false positive will be equal to $(1-(1-Z)^4) = 4Z-6Z^2+4Z^3-Z^4$. Let's notice a certain difference in the "cost" of a false positive for different schemes: if in sequential circuit (Fig. 1) each terminal turns off all consumers, then in parallel circuit (Fig. 6), one terminal disconnects only one line.

For one server 16 (Fig. 3), nothing needs to be calculated: the probability P of failure, and the probability of a false positive Z (just for the terminal and the server, we equated them). Since the server manages all protections, if the server fails, any line failure will result in protection failure, and the probability of such an event is $(1 - (1-X)^4)$. Accordingly, the probability of protection failure $Y=P \cdot (1 - (1-X)^4)$. Linear part $Y \approx 4 \cdot P \cdot X$, respectively $Y \sim 4 \cdot P$.

For three servers 16, 17, 18 (Fig. 5), different scenarios of the majority system, consider two scenarios: "1 of 3" (when the switch will be tripped, if although one server will issue a shutdown command) and "2 of 3" (when the switch is disabled

when any two servers issue a shutdown command at the same time).

Consider a 1 out of 3 scenario. For protection to fail, it is necessary that at the same time all three servers failed. Accordingly, the probability of protection failure $Y=P3*(1--(1-X)^4)$. Linear part $Y\approx 4*P3\cdot X$, respectively $Y\sim 4\cdot P3$. False positive probability equals $(1-(1-Z)^3)=3Z-3Z^2+Z^3$.

Consider a 2 out of 3 scenario. For protection to fail, it is necessary that at the same time two servers failed, and there are three such combinations, and one combination when all three failed servers: $Y=(3\cdot P^2\cdot(1-P)+P^3)\cdot(1-(1-X)^4)$. Linear part $Y\approx(12\cdot P^2\cdot(1-P)+4P^3)\cdot X$, respectively $Y\sim 12\cdot P^2\cdot(1-P)+4P^3=12\cdot P^2-8P^3$. For a false positive, so that two servers issue a false command at the same time, and there are three such combinations: $3\cdot Z^2$, or all three Z^3 servers, adding up we get: $3\cdot Z^2+Z^3$.

IV. Conclusion

Let's collect all the received formulas in one table.

TABLE I. Comparison of a set of terminals - and a central server

Probabilities	Sequential scheme	Parallel scheme	Server		
			"1 of 1"	*"1 of 3"*	*"2 of 3"*
Protection failure	$P+P^2+P^3$	$4\cdot P$	$4\cdot P$	$4\cdot P^3$	$12\cdot P^2-8P^3$
False actuation	$4Z-6Z^2+4Z^3-Z^4$	$4Z-6Z^2+4Z^3-Z^4$	Z	$3Z-3Z^2+Z^3$	$3\cdot Z^2+Z$

From this table, simple conclusions can be drawn:

- Even one server almost does not differ in reliability from a set of terminals.
- A system of three servers significantly outperforms a set of terminals in terms of reliability (proportional to P^3 or P^2 for servers, instead of P* for terminals).
- A system of three servers in a 2 out of 3 scheme is significantly superior to a set of terminals by the probability of false positives (proportional to *Z2* for servers, instead of Z for terminals).

Thus, the central server of relay protection and automation has unique properties, such as the absence of station and process buses. Except In addition, a multi-server system will have a larger (compared to a set of terminals) reliability both in terms of failures and false positives will allow automatically detect a damaged server, allow you to replace and configure server without stopping protection. As a result, the concept of a server can be applied to protection of a digital substation in some special cases.

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Self- heating Effect and Noise Generation in Graded Gap AlGaN diode

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***Abstract*—AlGaN graded diode is considered as a noise source in a wide frequency range. Diodes contain homogeneous AlxGa1-xN active region and a cathode graded AlGaN layer. Formation of high electric field domain in low resistance region with varied composition at cathode contact under impact ionization is considered as a way to obtain low threshold impact ionization and a high noise level in diode. The 2D diode simulation was carried out by Monte Carlo technique. Effect of self-heating on diode characteristics and noise generation was investigated. A significant temperature rise associated with impact ionization in cathode electric field domain was demonstrated. Increasing power spectral density (PSD) of diode at frequency below 100 GHz was shown. It was shown that self-heating contributes to an increase in the PSD at low frequency.**

Keywords—graded-gap layer, temperature, mole fraction, impact ionization, scattering, self-heating, power spectral density of noise.

I. INTRODUCTION

Today, there are many new applications usaging frequencies of above 100 GHz. Terahertz frequencies are already being used in communication systems [1,2], systems related to biological research [3], for objects state monitoring [4], for obtaining three-dimensional (3D) THz images, in THz tomography [5,6], etc.

Some new THz applications and devices development are needed in noise sources. They are used in different measurement tools, specially functioning as a demodulation of threshold detection of receivers, radar testing [7,8], determination of dynamic range and sensitivity of THz systems [9]. Noise sources are widely used in spectroscopy [5]. Another application of noise sources at frequency above 100 GHz is THz tomography [10]. The most compact of the known noise sources are electronic sources such as Schottky diodes [11].

The problem of electronic devices is limitations in their operating bandwidth. As result, the output power of solid state electronic noise sources has low level in the THz band. The second problem is very narrow and not flat noise RF spectrum of such devices.

Diodes with strong electric field domain stables at the near-cathode region, so-called static cathode domain diode (SCDD) were proposed in [12]. Electron-hole pairs are generated due to impact ionization in the electric field domain in diodes. The charge carriers transit through active region of diode leading to shot-noise current. Thus, frequency responsibility of the diode is depended on both diode active length and electric field domain configuration. The graded band gap semiconductor layers at the cathode effect on electric field magnitude and space charge length in active region. If semiconductor composition is distributed so as to obtain an increase in band gap towards the anode, the ionization threshold voltage decreases, and space charge in active region will be minimal. This allows us to obtain short active region length, and size devices suitable for high frequency application, including terahertz frequency region [13,14], and reconfiguration operation frequency of diode to several GHz.

SCDD on the base of ternary InGaN alloys were considered in [15]. The diode preference is a small band gap of InGaN and low impact ionization threshold correspondingly. On the other hand, it is a reason why small diode power limites its possible application.

AlN has many useful properties. The main of them is ability to form alloys with GaN producing AlGaN and allowing the fabrication of AlGaN/GaN-based devices for high power electronics. Those structures have a large energy gap, high specific heat, high thermal conductivity, resistance to high temperature [16].

The graded gap AlGaN/GaN-based SCDD with threshold of impact ionization corresponded GaN may be realized [17]. However, high electric field and self- heating can significantly effect the diode operation, especially on contact structure formed at not very high temperatures. Thus, the aim of study is 2-D simulations of graded gap AlGaN diodes considering impact ionization, temperature and self-heating effects under strong electric field to estimate diode as a noise sources in the range of up to 500 GHz.

II. THE MODEL

Fig.1 shows the AlGaN/GaN diode structure. It represents *n*-type wurtzite polytypes layer (active diode region) and low-doped region (n^-) at cathode. Heavily doped contact regions (n^+) form ohmic contacts to diode.

$Al_zGa_{1-z}N$ composition as a function of coordinate represents $z(x)=1-\exp(-(x-x_0)^2/2\sigma^2)$. There x_0 defines an

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interface between cathode graded layer and continues composition active region. σ is chosen to obtain composition at cathode corresponding GaN.

Two diode types of 1μm with $z(x_0)$= 0.12 and $z(x_0)$= 0.45 were considered. Doping profile of the diodes represents n^+-n^--n-n^+-structure with concentration of 10^{24} m^{-3} 10^{20} m^{-3}, $5\cdot10^{23}$ m^{-3}, and $10^{24}m^{-3}$ respectively. The length of n^- - region is about 300 nm, the active n- region of about 500 nm.

2-D model of diode has been considered. Transport properties of diodes were studied numerically using synchronous Ensemble Monte Carlo (EMC) approach self-consistently with Poisson's equation at the time interval $\Delta t = 5\cdot10^{-16}$ s. Two types of charge carrier were studied. Conduction band is represented by three lower non-parabolic valleys (Γ, Γ_1 and M–L), the valence band represents heavy holes valley. Both scattering of charge carrier and parameters of $Al_zGa_{1-z}N$ are determined similar to [17].

Metal – semiconductor contacts are considered to be ohmic. Electrostatic potential applied to contacts includes a direct current bias voltage and potential difference associated with the different electron affinities of the anode and cathode materials. The voltage magnitude was not far from impact ionization threshold determined by effective mass of heavy holes and corresponding initial stage of impact ionization. Applied voltage is suggested not to be sufficient for breakdown, and II process is initiated by electrons. II probability is considered similar to [17].

The temperature conductivity of diodes considered in terms of kinetic theory suggested a simple model of the temperature and composition depended lattice thermal conductivity $k(T, z)$ of material. Scale effects connected with heating are assumed to be neglected. Temperature distribution in the device due to self-heating has been determined by heat equation:

$$c\frac{\partial T(r,t)}{\partial t} = -\nabla\left[k(T,z)\nabla T(r,t)\right] + F(r,t) \qquad (1)$$

where c is the lattice heat capacity; $F(r, t)$ is locally defined by full energy density associated with emission and absorption phonon, and energy emission caused by impact ionization [18].

Thermal properties of the multilayer structure determine the conduction of thermal resistance and possible temperature rise in the device. The thermal boundary resistance (TBR) at each interface can cause an additional channel temperature rise.

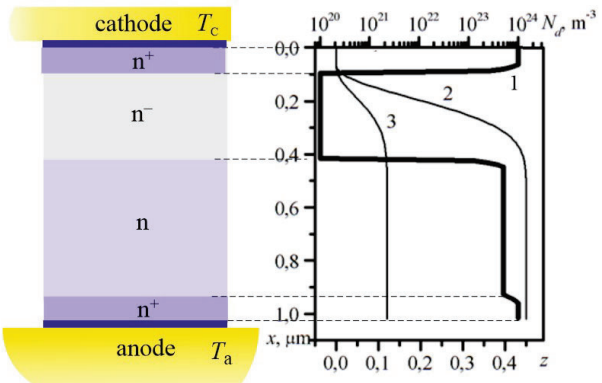


Fig.1. The diode structure: 1 - doping- profile; 2,3 – Al_zGa_{1-z} N composition $z(x)$.

The considered model of diode structure assumes a semi-infinite anode contact layer with a fixed temperature T_a=300 K at its bottom boundary (Fig.1).

Electric field distribution in diodes is shown in Fig.2 at constant temperature equal 300 K.

As can be seen, the electric field in diode represents strong fields domain mainly located in n^--layers and extended toward active n-region. At the same time, the electric field outside domain is still low, and temperature rise is also expected to be at cathode. For this reason, the boundaries condition for temperature may be assumed as

$$-k(T,z(x))\nabla T(x,t)\big|_{x=0} = (T_C - T(0,t))/R_b, \qquad (2)$$

$$T(x,t)\big|_{x=L} = T_a, \qquad (3)$$

where R_b is TBR of metal contact-GaN interface assumed to be TBR is $5\cdot10^{-9}$ K·m²/W [19]. Adiabatic thermal boundary conditions were used for other diode boundaries.

The material parameter used in simulation is presented in Table 1. [20,21].

TABLE 1. TEMPERATURE PARAMETERS $AL_ZGA_{1-Z}N$

Parameter	AlN	GaN
k_{300K}, W/(Km)	285	130
n	-1,57	-0,43
c, J/(kgK)	748	491

Temperature dependence of thermal conductivity is defined according to [20,21]:

$$k(T) = k_{300K}\cdot\left(\frac{T}{300K}\right)^n. \qquad (4)$$

k depending on $Al_zGa_{1-z}N$ composition is obtained as follows:

$$k(z,T) = \left(\frac{z}{k^{AlN}(T)} + \frac{(1-z)}{k^{GaN}(T)} + \frac{z\cdot(1-z)}{b}\right)^{-1}, \qquad (5)$$

where $b = 6.95$ W/(Km)is a bowing parameter [21]. The lattice heat capacity is considered to be independent of temperature, and is defined by Vegard rule.

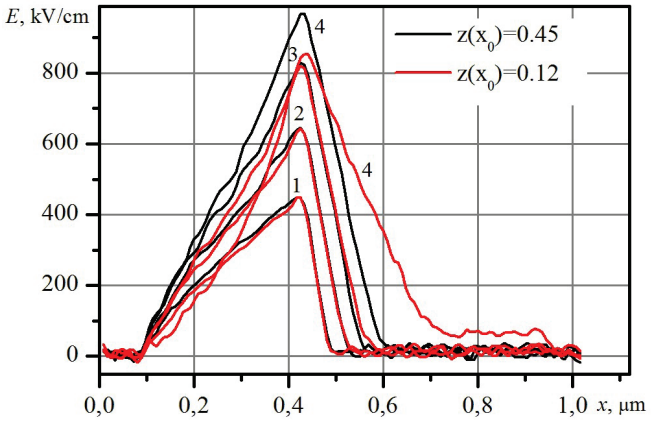


Fig.2. Electric field distribution in diodes: 1 – 10 V; 2 – 15 V; 3 – 20 V; 4 – 25 V.

Temperature dependence of band gap is applied according to Varshni formula $E_g(T) = E_g(0) - \alpha T^2/(\beta + T)$ [20], where composition dependence of parameters $\alpha(z)$ and $\beta(z)$ assumes from [22]. Impact ionization threshold energy $E_{th}(z,T)$ is defined as:

$$E_{th} = \frac{1+\gamma(z)}{2\alpha(z)}\left[1-\sqrt{1-4\alpha(z)E_g\cdot(2+\gamma(z))/(1+\gamma(z))^2}\right], \quad (6)$$

where $\gamma(z)$ is electron effective mass to hole effective mass ratio; $\alpha(z)$ is a non-parabolicity factor [23].

III. Results and Discussion

A. DC characteristic of the diode

Taking into account small relaxation time in III-nitride, and corresponded small time step in synchronous EMC, heating is estimated during limited time interval (up 12 ns). Dependences of current density on applied bias voltage obtained at 300 K without self-heating, and including self-heating are shown in Fig. 3.

Diodes demonstrate soft impact ionization threshold associated with carrier generation in peak of electric field domain. The significant temperature rise is observed at bias U>10 V, when impact ionization becomes a dominant heat source in diode. As can be seen from Fig. 4 impact ionization is more intensive for diode with $z(x_0)$=0.12.

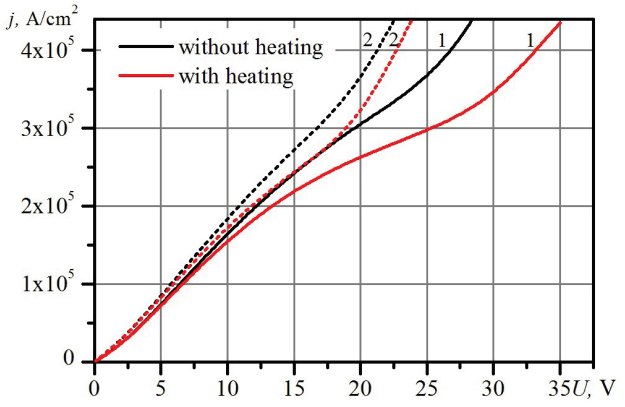


Fig. 3. Dependences of current density on applied bias voltage for analyzed diodes at 300 K without self-heating, and including self-heating: 1 - $z(x_0)$ = 0.45; 2 - $z(x_0)$ = 0.12.

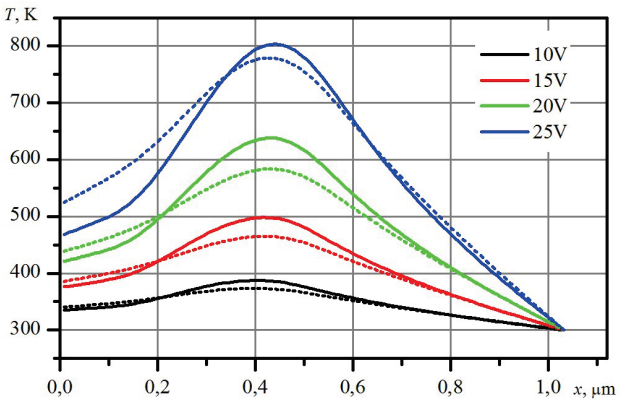


Fig. 4. Dependences of temperature on x- coordinate along diode, at different bias voltage (t=12ns) for diodes with composition $z(x)$: solid lines - $z(x_0)$=0.45; dotted lines - $z(x_0)$=0.12.

However, the thermal conductivity of graded $Al_zGa_{1-z}N$ layer at $z(x_0)$=0.45 is less significant, compared to thermal conductivity at $z(x_0)$=0.12, similar to GaN one. Thus, despite impact ionization, the current drop turns out to be lager in magnitude for diode with $z(x_0)$=0.12.

The simulation time was t_{sim}=12 ns. Fig. 5 shows the maximum temperature corresponded to local impact ionization region associated with electric field. Here, the impact ionization act distribution per the modeling time (N) is shown, as well as concentration of doping impurity N_d, electron, and hole concentration (n and p).

It is noteworthy that the maximum holes concentration is at cathode n^+-n^--junction for both considered diode types. Quasi-electric field in graded layer results in effective holes transport toward cathode contact. If the maximum domain field is small, there is no accumulation of holes in diode, and stable operation is observed. Recombination able to additional decreasing holes concentration in n^+-region, is not considered. However, it can be an additional heating source resulted from temperature rise at cathode, too.

B. Noise characteristics

Electron transit through n-region is the main noise source in diode. This region is similar to drift region in IMPATT. At the same time, generated holes drift towards the cathode contact so rapidly due to quasi-electric field. Noise signal was estimated by analysis of temporary current sample. Power spectral density (PSD) of current ($S(\omega)$) is determined by using numerical Blackman-Tukey algorithm [24]. Results obtained with heating and without are shown in Fig. 6 and Fig. 7.

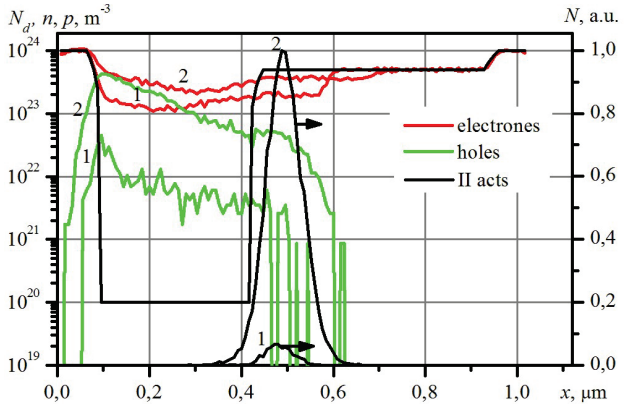


Fig. 5. Distribution of electrons and holes concentration, and impact ionization acts on x- coordinate along diode at bias voltage U=25V for diodes with composition $z(x)$: 1 - $z(x_0)$= 0.45; 2 - $z(x_0)$ = 0.12.

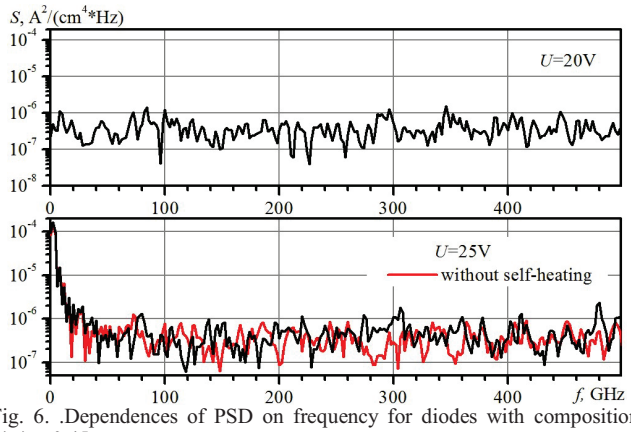


Fig. 6. .Dependences of PSD on frequency for diodes with composition $z(x_0)$ = 0.45.

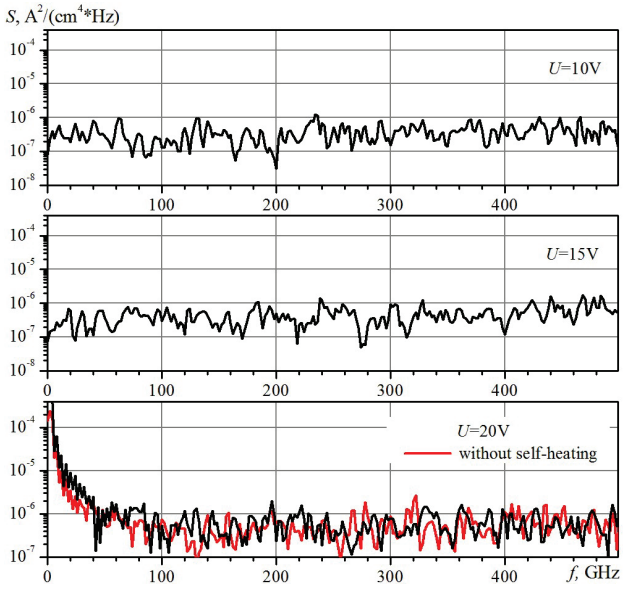


Fig. 7. Dependences of PSD on frequency for diodes with composition $z(x_0)$= 0.12.

As can be seen from Fig. 6,7, the PSD of current noise is dependent upon bias. At low bias the flatness of PSD is large (10 dB), and the noise is the shot noise generated by electrons passing through the diode. If the bias voltage is enough to impact ionization the PSD raising is mainly occurred at low frequencies (up to about 100 GHz). In this frequency range noise can be considered to like generation – recombination, where generation process is impact ionization, and, thus, fast charge carrier drift is an analogue of recombination. At frequency higher 100 GHz, spectrum becomes flatter, approaching white noise. Self- heating effect leads to increasing PSD at low frequency. At higher frequency self- heating is more noticeable at diode with $z(x_0)$=0.45 contrary to diode with $z(x_0)$=0.12.

IV. Concusion

Graded gap AlGaN/GaN-based SCDD has been investigated by using Monte Carlo simulation. Self- heating effect is shown to be significant and can lead to destruction of diode structure. At the same time, self-heating leads to increasing PSD of current, peculiarly at low frequency part of PSD. The simulation was carried out in a limited time frame. Thus, the temperature rise can be more significant. A preliminary estimate indicates a time of about 100 ns to obtain a result close to the steady- state. These longer time simulations will be the subject of further research.

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Comparative Study of Machine Learning Approaches for Fake News Detection on Social Platforms

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***Abstract* — This research presents a comprehensive practical comparison of traditional machine learning algorithms with modern deep learning approaches for automated fake news detection on social media. By evaluating set of models, including naive Bayesian, decision trees, random forest, logistic regression, LightGBM, bidirectional LSTM, CNN, and hybrid models on the ISOT dataset, we found out that gradient boosting methods, particularly LightGBM, achieve superior performance (99.8% F1 score) compared to complex neural network architectures and hybrid approaches. Our results contradict the common assumption that deep learning models are essentially outperforming in text classification tasks, especially in resource-constrained environments with limited training data. The study points out that well-designed traditional features combined with gradient boosting algorithms not only outperform deep learning approaches, but also provide better computational efficiency, interpretability, and deployability for real-world fake news detection systems. This diversity of results indicates significant potential for further improvement in this area, with traditional machine learning showing unexpected dominance over complex hybrid models.**

Keywords— Social media, Supervised Learning Algorithms, Deep Learning Models, Hybrid Models, LightGBM, fake news detection, gradient boosting, evaluation metrics.

I. INTRODUCTION

The rise of the digital tech has changed how we share the information, making it easier for people and social groups to connect and spread content worldwide.

Social media has definitely evolved the way we communicate and share information and has provided an opportunity for individuals and organizations to connect, express opinions and share content with a wide audience [1, 2]. However, the rapid circulation of information and the ease of its exchange [3,4] have also contributed to the rapid spread of fake news, which has become a serious problem for modern society.

A. Current Detection Strategies and Their Limitations

Fake news is considered to be deliberately false news that is created with the intent to mislead [5]. The propagation of such information can have serious consequences, including loss of trust in traditional media sources, manipulation of public opinion, and even potentially affecting certain decisions, polls, or political outcomes [6].

Even though social networks are making efforts to tackle the spread of fake news, such as introducing algorithms and mechanisms for fact-checking to identify and flag suspicious content, the problem still persists. One of the few reasons for the prevalence of this phenomenon is the assertion bias [7], when people are inclined to trust information that will correspond to their existing worldview and convictions. Usually, people do not bother verifying such information and continue broadcasting it further in the network.

Fake news detection methods often merge multiple approaches, each with its own strengths and weaknesses. One common method is content-based analysis, which examines the language and writing style of news articles to spot possible signs of false information [8], analyzing how the content is written and assessing the credibility of the source. Another widely used method is semantic analysis, which helps uncover the deeper meaning and context of messages and can detect logical inconsistencies that may indicate disinformation.

Content-based analysis is good at spotting subtle language clues and writing patterns, but it needs a lot of training data and can struggle to keep up with how quickly fake news evolves. Still, it's often used in tools like news aggregators, automated content filters, and fact-checking platforms.

Another useful approach is social network analysis, which looks at how news spreads on social media and tracks how users interact with it. This method is commonly used to monitor and detect disinformation [9].

B. Problem Statement and Research Motivation

The world of machine learning has been leaning more towards deep learning styles, like RNNs, LSTMs, and Transformer-based forms, believing these complex styles can catch small meaning shifts and context details that older machine learning methods might not see.

This study aims to answer several questions through a systematic practical evaluation with the following objectives:

1) Evaluate the performance of traditional machine learning algorithms compared to deep learning approaches and hybrid models using standardized evaluation metrics
2) Investigate the impact of advanced feature engineering, including TF-IDF vectorization and linguistic pattern detection
3) Explore whether a combination of different machine learning paradigms delivers better performance.

Our main achievements include providing robust practical results that traditional machine learning algorithms can outperform complex deep learning architectures, as well as demonstrating that higher performance can be achieved with significantly lower computational requirements.

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II. Literature Review and Related Work

A. Evolution of Machine Learning Approaches

The area of automated fake news detection has grown significantly, pushed by both technological advances and increasing need of combating misinformation. Early approaches mainly aimed at content-based analysis using traditional natural language processing techniques, but recent work has explored complex deep learning architectures and multimodal analysis approaches.

Traditional Machine Learning: Traditional algorithms have demonstrated exceptional effectiveness in various text classification tasks. Feature engineering techniques like TF-IDF vectorizing have proven particularly effective for capturing word importance while lessening the impact of terms. Tree-based methods like Random Forest and gradient boosting methods have achieved high performance across numerous applications by handling high-dimensional sparse features effectively.

Deep learning: The deep learning revolution has changed a lot how natural language are processed, with researchers studying different neural network architectures to spot fake news. LSTM and GRU networks are widely used, because of their ability to capture sequential dependencies. CNN architectures adjusted for text processing are efficient for detecting small patterns and transformer models such as BERT achieve high performance but require sufficient computer power.

B. Multimodal and Hybrid Approaches

More and more researchers are exploring multimodal analysis, which combines multiple data sources, such as text, video and metadata, to identify fake news. By analyzing how well different types of information match and stay consistent and identifying fake or manipulated media, this approach can improve detection accuracy. However, there are drawbacks, such as a significant amount of computation power and requirements for multi-field expertise.

Hybrid models include different machine learning algorithms to use their strengths and reduce their weaknesses and to do so they often integrate traditional machine learning methods with deep learning models or merge multiple models together. The purpose of using such models is to improve accuracy, reliability, and the ability to interpret and handle complex, multimodal data. However, developing effective hybrid models can be challenging, because it requires careful balancing of the various components and smooth integration.

III. Evaluation metrics

Precision is a parameter that is measured as the number of true positives in relation to all positives:

$$Precision = \frac{TP}{TP+FP}. \quad (1)$$

Recall - also called the rate of true positive rate, or simply sensitivity:

$$Recall = \frac{TP}{TP+FN}. \quad (2)$$

Accuracy - this metric is essentially the correctness of the model's predictions in the dataset:

$$Accuracy = \frac{TP+TN}{Numberofsamples}. \quad (3)$$

Also known as the F-measure, the F1 score is a metric for evaluating a machine learning model's performance, namely in binary classification tasks. It balances or rather removes extreme negative values of recall and precision by using the harmonic mean:

$$F1_{score} = \frac{2*(precision*recall)}{(precision+recall)}. \quad (4)$$

IV. Dataset and Experimental Setup

We tested our models on the ISOT Fake News Dataset, which provides a balanced collection of 23000 real and 21000 fake news articles, which is sufficient amount for traditional machine learning and almost enough for deep learning approaches.

Real news articles in this dataset were sourced from Reuters.com, while fake news articles were collected from various unreliable sources flagged by fact-checking organizations, which provides a realistic representation of the problem of detecting fake news in social media.

We still needed to cleanup data from dataset. We removed URLs, email addresses, and special characters while preserving sentence structure and punctuation patterns relevant for stylistic analysis. Text was converted to lowercase for consistency with the original text for certain linguistic feature extraction, applied word-level tokenization suitable for both TF-IDF vectorization and neural network input preparation and data was split with 80-20 ratio to ensure balanced representation of both classes in training and evaluation sets.

A. Feature Engineering Framework

We use two main approaches to process text and find patterns that help identify if news is fake.

Firstly, we convert text into numbers using TF-IDF, which tells us how important words are in each article and limit this to 1,500 key features to keep processing manageable. We look at single words and pairs of words together (n-gram range of 1 to 2), cut out common words like "the" and "and" that don't add meaning, and filter out very rare words and words that appear in almost every article (document frequency thresholds of min_df=2 and max_df=0.95).

Second, we extract linguistic features that capture how fake news is typically written. We measure basic things like article length, word count, and sentence count since fake news often has different structural patterns. We also look at writing style by counting exclamation marks, question marks, and capital letters, which fake news uses more often for dramatic effect. We search for suspicious words and phrases that commonly appear in fake news, such as "shocking," "unbelievable," and "secret," along with typical clickbait style of language. Also, how easy the text is to read using standard readability scores is measured, since fake news tends to be written at different complexity levels than real news.

Finally, we combine both the word importance scores and the writing style measurements into one complete set of features that captures both what the article says and how it's written.

B. Model Implementation and Configuration

Traditional machine learning models include Naive Bayes using Multinomial Naive Bayes with alpha smoothing parameter optimized for text classification tasks.

Decision Tree employs configuration with controlled depth of max_depth=15 and minimum samples per split of min_samples_split=5 to prevent overfitting.

Random Forest utilizes ensemble of 100 trees with controlled depth of max_depth=10 to balance performance and computational efficiency.

Logistic Regression implements L2 regularization and maximum iteration limit of 1000 for convergence stability. LightGBM applies gradient boosting configuration with 200 estimators, learning rate of 0.1, and maximum depth of 8, optimized for text classification tasks.

Deep learning models include Bidirectional LSTM using architecture designed for memory efficiency with embedding dimension of 32, LSTM units of 16 which is bidirectional for total of 32, dropout rates of 0.3 for regularization, dense layers of 16 units with ReLU activation, and output of single sigmoid unit for binary classification.

Convolutional Neural Network employs lightweight architecture for text processing with embedding dimension of 32, convolutional layers of 32 filters with kernel size 3, pooling using MaxPooling1D and GlobalMaxPooling1D, and dense layers of 32 units with dropout regularization. Hybrid Model implements multi-input architecture combining text sequences processed through BiLSTM, TF-IDF features processed through dense layers, linguistic features processed through dense layers, and feature fusion through concatenation and final dense layers.

All tests were run on RTX 3050 Ti GPU with 4GB VRAM, this configuration provides realistic assessment of model performance in conditions of limited resources.

C. Training Procedures and Optimization

Traditional model training uses scikit-learn implementations with standardized procedures including cross-validation for hyperparameter selection where applicable, consistent random state settings for reproducibility, and standardized feature scaling where required by the algorithm.

Deep learning training uses TensorFlow and Keras with Adam optimizer at learning rate 0.001, binary crossentropy loss function, batch size of 4 optimized for memory constraints, 5 epochs with early stopping monitoring validation loss, and validation split of 20% of training data.

Memory management implements efficient strategies including gradient accumulation for larger effective batch sizes, model checkpointing to prevent memory overflow, and automatic garbage collection between model training sessions.

D. Error Pattern Analysis

LightGBM Error Profile shows such results (Fig. 1):

True Positives: 4,690 (correctly identified fake news), True Negatives: 4,276 (correctly identified real news), False Positives: 6 (real news incorrectly flagged), False Negatives: 8 (fake news missed).

This exceptional error profile demonstrates that LightGBM makes extremely few errors of either type, making it highly suitable for production deployment.

False Positive Rate Comparison for traditional ML Models: 0.001-0.020 (excellent), deep Learning Models (Fig. 2): 0.013-0.021 (competitive), our Hybrid Model (Fig. 3): 0.207 (unacceptably high).

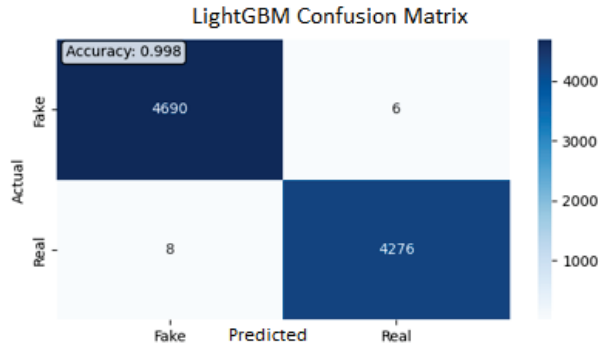


Fig. 1. Confusion matrix for LightGBM method.

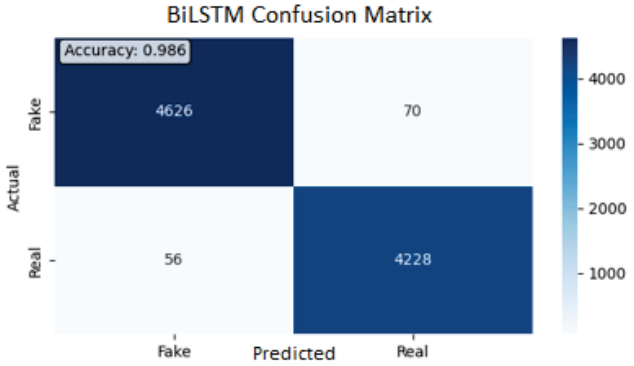


Fig. 2. Confusion matrix for BiLSTM model.

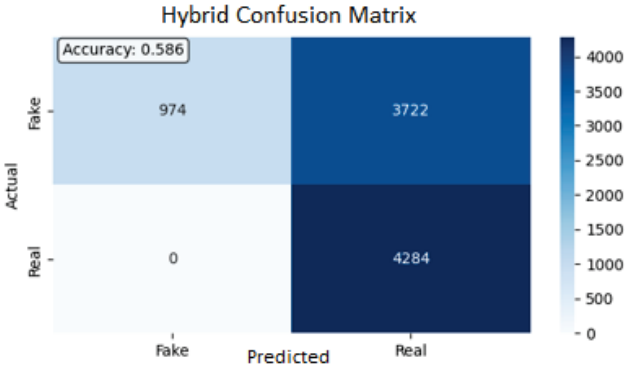


Fig. 3. Confusion matrix for Hybrid model.

E. Overall Performance Comparison

This comprehensive evaluation reveals striking differences in performance across algorithmic families, questioning the common beliefs that advanced about the superiority of complex deep learning approaches in fake news detection (Table 1).

Performance Rankings: The experimental results demonstrate a clear hierarchy of model performance, with traditional machine learning algorithms occupying the top positions:

LightGBM: 99.8% F1-score (99.8% accuracy, 99.9% precision, 99.8% recall)

Random Forest: 99.6% F1-score (99.6% accuracy, 97.9% precision, 99.6% recall)

Decision Tree: 99.4% F1-score (99.6% accuracy, 99.5% precision, 99.4% recall)

Logistic Regression: 98.4% F1-score (98.4% accuracy, 97.9% precision, 98.8% recall)

TABLE I. EVALUATION METRICS FOR ALL MODELS

Model	Accuracy	Precision	Recall	F1-Score
Naïve Bayes	0.927	0.891	0.964	0.926
Decision Tree	0.994	0.994	0.994	0.994
Random Forest	0.996	0.995	0.996	0.996
Logistic Regression	0.984	0.979	0.988	0.984
LightGBM	0.998	0.999	0.998	0.998
BiLSTM	0.986	0.984	0.987	0.985
CNN	0.982	0.984	0.977	0.981
Hybrid	0.586	0.535	1	0.697

BiLSTM: 98.5% F1-score (98.6% accuracy, 98.4% precision, 98.7% recall)

CNN: 98.1% F1-score (98.2% accuracy, 98.4% precision, 97.7% recall)

Naive Bayes: 92.6% F1-score (92.7% accuracy, 89.1% precision, 96.4% recall)

Hybrid Model: 69.7% F1-score (58.6% accuracy, 53.5% precision, 100% recall)

The performance differences between top-performing traditional algorithms and deep learning approaches, while numerically small, represent substantial improvements in practical terms. The difference between LightGBM (99.8%) and BiLSTM (98.5%) translates to approximately 65% fewer misclassifications, a significant improvement for production deployment.

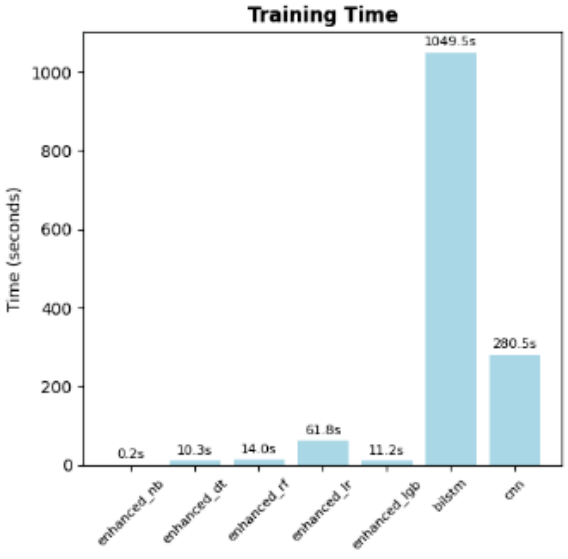


Fig. 4. Training time chart.

Models such as NB, DT, RF, and LR complete training relatively quickly, ranging from fractions of a second to just over a minute. Among them, NB is the fastest, requiring only 0.2 seconds, while LR is the slowest at around 61.8 seconds. LightGBM is particularly notable, as it combines relatively low training cost (11.2 seconds) with a strong reputation for accuracy and robustness in handling large datasets. In contrast, the deep learning models require substantially more resources: CNN takes 280.5 seconds, while BiLSTM is by far the most time-consuming, requiring over 1,000 seconds to complete training.

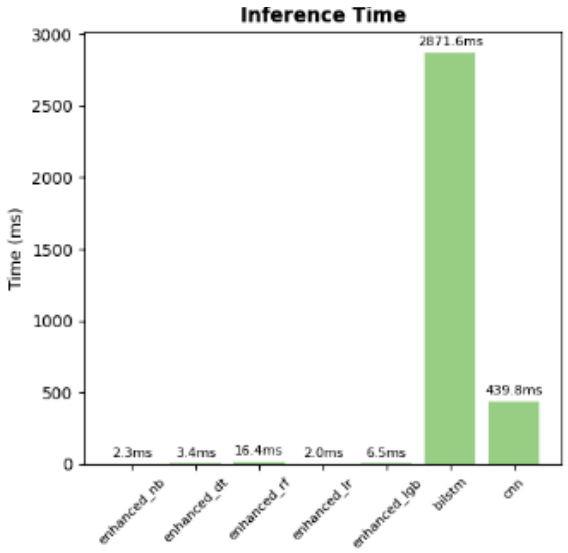


Fig. 5. Inference time chart.

From a practical perspective, the choice of model depends on the intended application. For real-time systems, lightweight models like LR or NB provide the highest speed. However, LightGBM presents itself as a more balanced option, delivering both efficiency and predictive power, making it a strong candidate for large-scale applications where both performance and scalability are essential. CNN may be used in offline or batch settings when accuracy outweighs speed, while BiLSTM, given its extreme training and inference costs, should only be considered if it provides a substantial and proven improvement in accuracy.

V. CONCLUSIONS

A. LightGBM performance

The results clearly show that LightGBM and other gradient boosting methods perform much better than deep learning and hybrid models when it comes to fake news detection. This defies the popular belief that neural networks are always the best choice for text classification.

Here's why gradient boosting stands out:

- Handles complex patterns well - very effective at finding relationships between features in large, sparse datasets.
- Efficient with text features - works especially well with TF-IDF data, using memory efficiently.
- Built-in overfitting protection - includes regularization features that help avoid overfitting while still keeping strong prediction ability.
- Easy to interpret - can show which features are most important in making decisions, which is helpful for tasks like content moderation.

B. Deep Learning Limitations in Small Data Regimes

Deep learning models typically require larger datasets to achieve their full potential. In our case, ~10,000 sample dataset, while being reasonable for traditional machine learning, appears insufficient for neural networks to learn strong, generalizable patterns that outperform carefully engineered features.

These models are significantly data-hungry and so demand not only vast amounts of labeled data but also high computational resources. Unlike traditional models that can benefit directly from decades of linguistic research and handcrafted features, neural networks must learn all relevant representations from scratch. While deep learning models generally focus on extracting semantic meaning, tasks like fake news detection often depend more on syntactic clues and writing style, where traditional features still excel. Furthermore, deep learning requires careful tuning of hyperparameters and thoughtful selection of model architecture, adding further complexity to their effective deployment.

C. Hybrid Model Challenges

The low performance of our hybrid model points out key issues in combining different machine learning methods. While individual models—like LSTM, CNN, or traditional algorithms—can perform well on their own, simply merging them without a thoughtful design can lead to weak results.

One of the major issues is compatibility. The models need to produce feature outputs that work well together and it often requires aligning their representations. Also important to scale and normalize the inputs correctly—without this, one model's output might overpower the other. Basic approaches like just joining feature sets usually aren't enough for strong performance.

Another challenge is that hybrid models generally need much more training data than individual models, which can be a serious limitation if data is limited.

That said, when designed carefully, hybrid models can still be highly effective. For example, combinations like CNN-RNN have reached strong performance scores (such as a 99.1% F1-score) when implemented properly [11].

D. Practical Implications

Our study shows that LightGBM achieves an impressive 99.33% F1-score and consistently outperforms all tested deep learning and hybrid models in fake news detection.

Deep learning models had struggles mainly due to limited dataset size and that made it hard for them to learn effectively. In the case of fake news detection, success often depends more on spotting clear suspicious patterns than on deep semantic understanding. In this context, well-crafted TF-IDF features turned out to be more useful than word embeddings, helping capture the importance of specific words. In addition, memory limitations forced the use of simplified neural network architectures, which hurt performance.

Hybrid models also frequently underperformed, with some scoring as low as 58.6% accuracy. This shows that simply increasing model complexity doesn't guarantee better results and can even make things worse by adding unnecessary overhead.

Overall, our findings suggest that combining strong feature engineering with traditional machine learning often works better than relying on complex deep learning setups—especially in real-world cases where speed, clarity, and limited resources matter most.

E. Future Work

Our findings point out several critical directions for advancing research of fake news detection.

First, broader evaluation on higher computing power and larger datasets is essential to determine if deep learning approaches are able to overcome their current limitations. Potentially revealing whether the added complexity actually pays off when provided with sufficient computational resources and training data.

Second, multi-modal detection approaches offer promising opportunities for improvement by integrating contextual information such as user profile analysis, social network propagation patterns, image and video content verification, temporal publishing patterns, and checking consistency across platform. These complementary data sources could provide richer feature sets that traditional machine learning algorithms can effectively use without sacrificing speed or simplicity.

Third, creating better hybrid models will require thoughtful design. Instead of just combining different tools, future work should focus on smart ways to blend features, assign proper weights to each type of data, and evaluate models carefully to avoid the performance drops we've seen from poorly integrated systems. Future research should focus on building systems where traditional and deep learning methods work together in a meaningful way, potentially through ensemble methods that preserve the interpretability and efficiency advantages of traditional algorithms and deliberately merging capabilities of deep learning models where they can provide clear value addition.

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Analysis of the Laws of Distribution of Gases Growth Rates in Serviceable Normally Operating 110-330 kV Transformers of Non-hermetic Construction

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***Abstract*—The paper presents the results of analyses of the laws of distribution of gases growth rates for 432 serviceable 110 kV transformers and 330 kV autotransformers of non-hermetic construction, which are operated by five different energy companies of Ukraine. A detailed description of the proposed procedure of statistical analysis of the results of periodic tests is given. The significant influence of duration and operating modes of transformers, peculiarities of their design, as well as the type of transformer oils on the values of gases growth rates is established. The performed researches have shown that in serviceable normally operating transformers of non-hermetic construction the gases growth rates can be described by double exponential distribution, which adequately takes into account the mechanisms of gases formation in serviceable non-hermetic equipment.**

Keywords—diagnostics, power transformers of non-hermetic construction, dissolved gas analysis, gas growth rate, limit values, gas formation mechanism, homogeneity, statistical criteria, double exponential distribution.

I. Introduction

Technological failures and damage to high-voltage power transformers are usually accompanied by significant economic losses and can even lead to interruptions in power supply to consumers. One of the causes of transformer damage is a group of processes involving ionisation ageing and thermal destruction, which are referred to as developing defects. Electrical discharges (creeping, sparking, arcing) pose an immediate danger to transformers, as they can lead to internal short circuits and, as a result, to transformer failure. Local temperature overheating is not usually accompanied by immediate damage to transformers. At the same time, they lead to accelerated ageing of cellulose insulation, a decrease in its degree of polymerisation and, as a result, an increased risk of damage to transformers due to a reduction in the mechanical strength of cellulose insulation when exposed to short-circuit currents. In this regard, early detection of developing defects is an important practical task.

One method for detecting developing defects at an early stage is dissolved gas analysis (DGA). One criterion for recognising the presence of developing defects in oil-filled equipment is the growth rate of gases. The obvious advantage of this criterion is the ability to predict the appearance of a defect even before the gas concentration values (which is another criterion used to detect defects) exceed their limit values. In addition, the degree of danger of a defect can be assessed based on the values of gas growth rates [1]. However, an analysis of the limit values of gas growth rates, which are regulated by standards [2-4], showed significant differences not only in the values of growth rates, but also in the very principles of standardisation. Thus, in [2], the values of the growth rates are normalised for each gas, while in [3], normalisation is performed in the form of a range of values related to the sum of combustible gas concentrations and the frequency of observations. In [4], only the amount of hydrocarbon gases is standardised. Thus, there is currently no uniform approach to standardising the limit values for gas growth rates as one of the criteria for the presence of defects in oil-filled equipment. According to the authors, one possible direction for developing such an approach is to use statistical methods and factor analysis, which, according to [5-7], allow determining the threshold values of indicators that ensure the minimum risk of making erroneous decisions. However, the use of these methods is fundamentally impossible without analysing the distribution laws of diagnostic features in equipment with different conditions. This research is dedicated to solving this problem.

II. Analysis of Publications and Research Objectives

Currently, a large number of research studies are devoted to the analysis and improvement of methods for diagnosing the condition of high-voltage transformers. Thus, works [8-10] present the results of a comprehensive analysis of methods for diagnosing and determining the types of defects in power and measuring transformers. The practical application of existing methods for diagnosing high-voltage equipment under various operating conditions is analysed in [11-12]. Particular attention is paid to improving methods for interpreting DGA results [13-17]. At the same time, new technical means for measuring gas concentrations [18-20] and modern recognition methods, such as intelligent algorithms and machine learning [21-23], clustering methods [24] and fuzzy logic [25], are being developed. However, despite the large number of publications, issues related to the analysis of gases growth rate distribution laws in equipment with different conditions have not been sufficiently addressed, which served as the reason for preparing this publication.

III. Research Methodology

Power transformers with a voltage of 110 kV and autotransformers with a voltage of 330 kV of non-hermetic

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construction were selected as the object of research, as they are one of the most common and critical elements of electrical systems. To analyse the distribution laws of gases growth rates, DGA results were used for 432 transformers with a voltage of 110 kV and autotransformers with a voltage of 330 kV of non-hermetic construction, which are operated by five different energy companies in Ukraine. None of the transformers analysed showed signs of being defective. The sample analysed comprised 54,760 values.

The construction features of high-voltage power transformers, in particular the type of oil protection, are a key factor influencing the gas content of transformer oils. As shown in [26], there are significant differences not only in the quantitative but also in the qualitative composition of gases in oil samples from non-hermetic transformers and transformers with nitrogen or film protection. The main reason for these differences is the diffusion of gases with low solubility in oil (H_2, CO, CO_2, CH_4) into the atmosphere, which is characteristic of non-hermetic transformers. Furthermore, the presence of free oxygen-air in such transformers leads to a significant change in the qualitative composition of gases in oil samples. In particular, according to [26], such transformers have a higher C_2H_4 content relative to C_2H_6, which is due to secondary oxidation reactions between liquid and gaseous hydrocarbons in oils, and which is not observed in transformers with nitrogen or film protection. Thus, unlike transformers with nitrogen and film protection, for which the amount of gas at time *i*-th is the sum of the amount of gas at time 1-st and the amount of gas released during the interval between the dates of these tests, in non-hermetic transformers the gas formation mechanism is more complex. In particular, in addition to the component characterising gas formation, two more components appear. The first of them characterises the process of gas diffusion into the atmosphere, and the second describes the process of changing gas concentrations due to oxidative reactions. The presence of such processes can lead to significant errors in determining gases growth rates, especially for gases with low solubility in oil due to temperature fluctuations in their concentrations, as well as losses during sampling, transportation and analysis.

Another factor affecting the reliability of gases growth rate values is the presence of errors in measuring gas concentration values. According to the regulatory document in force in Ukraine [4], the systematic error in measuring gases concentrations increases as the concentration values decrease and can reach 20%. According to the regulatory document in force in Ukraine [4], the systematic error in measuring gaseous concentrations increases as the concentrations decrease and can reach 20%. To reduce the impact of measurement errors on the values of gases growth rates, we used the approach applied in [27] to determine the values of relative gases concentrations, which proved to be quite effective. The essence of this approach is that the values of gases growth rates were calculated only under the condition that the values of gases concentrations used for calculation exceeded the values specified in [4] that ensure minimum measurement errors. The values are as follows: H_2=50 µl/l; CH_4=C_2H_6=C_2H_4=15 µl/l; C_2H_2=3 µl/l; CO=200 µl/l; CO_2=1000 µl/l.

To determine the values of gases growth rates, the expression regulated in [4] was used:

$$R_i = \frac{(S_2 - S_1)\cdot V}{t}\cdot 10^{-3}, \qquad (1)$$

where R_i – gases growth rate, ml/day; S_1 – gas concentration or sum of concentrations of hydrocarbon gases in the first sample, µl/l; S_2 – gas concentration or sum of concentrations of hydrocarbon gases in the second sample, µl/l; V – oil volume in the transformer tank, l; t – time between sampling, days.

A preliminary analysis of the results obtained showed that the values of gases growth rates vary within a fairly wide range, as clearly illustrated in Fig. 1, which shows the dependence of the C_2H_4 growth rates obtained as a result of the calculation on the duration of operation. The Fig. 1 clearly shows that despite the restrictions imposed on concentration levels, when calculating gases growth rates, the estimated values may differ by several orders of magnitude and exceed the maximum gases growth rates (30 µl/day) specified in [4] by at least five times. This circumstance indicates the presence of unaccounted factors influencing the values of gases growth rates.

The performed analysis made it possible to establish that abnormally high (in terms of modulus) values of gases growth rates were obtained for transformers that were subjected to emergency influences from the electrical grid (short circuits, atmospheric and switching overvoltages, etc.). As an illustration, Fig. 2 shows the dependence of the C_2H_2 rise rate on the duration of operation for a three-winding transformer with a rated power of 31.5 MVA and a rated voltage of 110/35/6 kV, which was exposed to short-circuit currents.

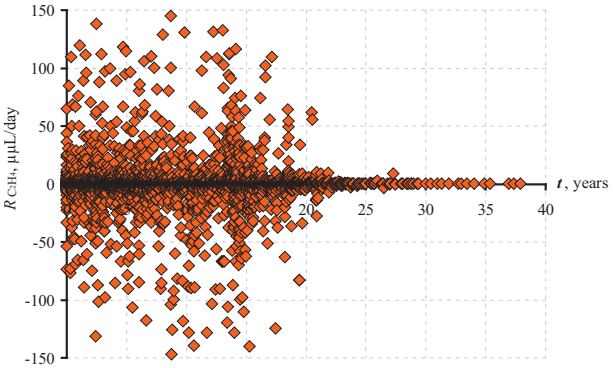


Fig. 1. Dependence of the growth rates of C_2H_4obtained as a result of the calculation on the duration of operation.

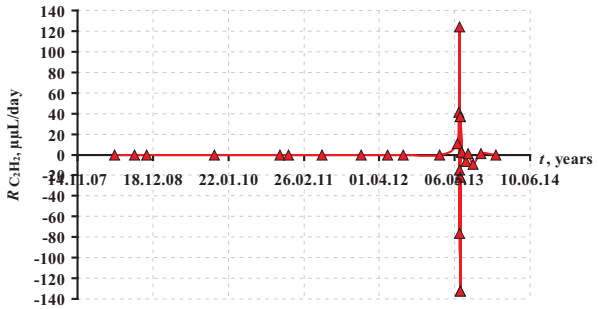


Fig. 2. Dependence of the growth rate of C_2H_2 on the duration of operation for a three-winding transformer with a rated power of 31.5 MVA and a rated voltage of 110/35/6 kV, which was affected by short-circuit currents.

As can be seen from the Fig. 2, emergency impacts from electrical grids lead to a significant increase in gases rise rates, not only compared to the gases rise rates in this transformer before and after the impact, but also compared to the limit values of gases rise rates regulated in [4]. Thus, due to emergency influences from electrical grids, the values of gases growth rates in the same transformer at different operating intervals may differ significantly. That is, the values of gases growth rates obtained as a result of calculations are statistically heterogeneous.

Unfortunately, the authors have limited information about the emergency operating modes of the transformers and autotransformers under consideration, especially with regard to equipment commissioned in the 1970s. In this regard, it is not possible to use a priori information to reduce the heterogeneity of the input data. In such conditions, it is advisable to use statistical analysis methods. In particular, if we assume that abnormally high (in terms of modulus) values of gases growth rates are 'gross errors', it is quite logical to use statistical hypothesis testing methods to identify 'gross errors' from among similar measurements. The most common [28] statistical criterion used to identify 'gross errors' from among similar measurements is Student's t-test. However, it is used only if the random variable being analysed has a normal distribution. In this situation, the law of distribution of gases' growth rates is unknown and needs to be determined. By analogy with [29], Irwin's criterion was used to solve this problem. The identification of gross errors using Irwin's criterion was performed in the following sequence. For each of the transformers analysed, the values of the rise rates for each gas were sorted in ascending order. Next, the selective value of the Irwin criterion was determined:

$$\eta_{\text{est}} = \frac{\left(x_k - x_{k\,\text{prev}}\right)}{S}, \tag{2}$$

where x_k – suspicious value; $x_{k\,\text{prev}}$ – previous value in the variation series.

The value of the growth rate was considered a 'gross outlier' and was removed from the sample if the sample value of the Irwin criterion exceeded the critical value [29]. This procedure was repeated until the sample value of the Irwin criterion became less than the corresponding critical value. As an example of using Irwin's criterion to identify 'gross errors', Fig. 3 shows the dependence of the C_2H_2 growth rate on the duration of operation for a three-winding transformer with a rated power of 31.5 MVA and a rated voltage of 110/35/6 kV, which was exposed to short-circuit currents (see Fig. 2), after using the Irwin criterion.

Comparing the dependencies shown in Fig. 2 and Fig. 3, it is easy to see that the application of the Irwin criterion significantly reduced the spread between the values of gases growth rates.

The reliability and validity of conclusions obtained as a result of statistical analysis, in addition to the adequacy and validity of the choice of statistical criteria and the correctness of statistical procedures, are determined not least by the size of the sample analysed. In this regard, it is quite obvious that a sample compiled based on the results of calculating this parameter for a single transformer will clearly be insufficient for a reliable analysis of the law of distribution of gases growth rates.

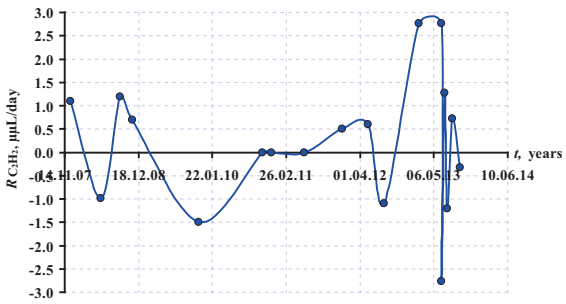


Fig. 3. Dependence of the C_2H_2 growth rate on the duration of operation for a three-winding transformer with a rated power of 31.5 MVA and a rated voltage of 110/35/6 kV, which was exposed to short-circuit currents, after using the Irwin criterion.

Therefore, the calculation results must be combined into a single data array. However, automatically combining the gases growth rates for different transformers, obtained after processing using the Irwin criterion, into a single array did not allow us to establish a distribution law. This is due to the heterogeneity of gases growth rates, caused primarily by differences in the operating temperature regimes of transformers (different load factors), different service lives, differences in design (cooling systems), and differences in the types of oil used. This conclusion is clearly illustrated in Fig. 4, which shows a histogram of the empirical distribution of C_2H_6 growth rate values for a combined data set consisting of 3264 values.

As in the previous case, given the limited a priori information about the operating modes of the transformers under analysis, statistical hypothesis testing methods were used to form arrays with uniform gases growth rate values. The main task of this stage of statistical processing of the initial data was to identify transformers for which the values of the mathematical expectations of gases growth rates and the values of gases growth rate dispersions would be statistically indistinguishable, and the distribution law of gases growth rates would be identical. To solve this problem, criteria (see Table I) were used, by analogy with [29], to test the following statistical hypotheses [28]:

1) statistical hypothesis about the identity of the distribution laws of two independent samples (Wilcoxon's nonparametric rank test, *W*, was used);

2) statistical hypothesis about the equality of mathematical expectations of two independent samples (*Z* criterion was used);

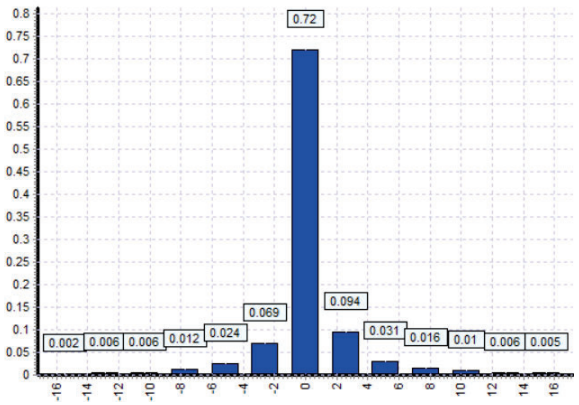


Fig. 4. Histogram of the empirical distribution of C2H6 growth rate values for a combined data set consisting of 3264 values.

TABLE I. STATISTICAL CRITERIA USED TO FORM HOMOGENEOUS ARRAYS OF GASES GROWTH RATES

Criterion	Statistical model	Main hypothesis	Criterion statistics	Comparison statistics
W [28]	$X_1 = (x_n^1 \dots x_{n1}^1) \sim f(x_1)$; $X_2 = (x_n^2 \dots x_{n2}^2) \sim f(x_2)$; $n = n_1 + n_2$	$f(x_1)=f(x_2)$	S_1 – sum of ranks of a set $\{x_i^1\}$ S_2 – sum of ranks of a set $\{x_i^2\}$	$S_{n1,n2,\,0,025}$ $S_{n1,n2,\,0,975}$
Z [28]	Sample 1: volume n_1, selective average m_1, selective dispersion δ_1^2 Sample 2: volume n_2, selective average m_2, selective dispersion δ_2^2	$m_1=m_2$	$Z = \dfrac{(m_1 - m_2)}{\sqrt{\dfrac{\delta_1^2}{n_1} + \dfrac{\delta_2^2}{n_2}}}$	Z_{crit}
F [28]	Sample 1: volume n, selective dispersion δ_1^2 Sample 2: volume m, selective dispersion δ_2^2	$\sigma_1^2=\sigma_2^2$	$F= \delta^2_{max}/ \delta^2_{min}$	$F_{(n-1),(m-1),\,0,95}$

3) statistical hypothesis about the equality of variances of two independent samples (Fisher-Snedecor *F*-test was used).

The formation of homogeneous arrays of gases growth rates was carried out in the following sequence. Each transformer was assigned a serial number. Next, general data arrays were formed for each of the gases analysed. After that, for each gas separately, the values of gases rise rates between different transformers were tested in pairs (first between the first and second, then between the first and third, etc.). The values of gases rise rates in the two transformers under analysis were considered uniform only if all three conditions were met simultaneously:

1. $S_1 > S_{n1,\,n2,\,0,025}$, $S_2 < S_{n1,\,n2,\,0,975}$ – test *W*;
2. $Z < Z_{crit,\,0,95}$ – test *Z*;
3. $F < F_{(n1\,-1),\,(n2\,-1),\;0,95}$ – test *F*.

In this case, the values of gases growth rates from two compared transformers were combined into a single data array.

The testing allowed us to form 3 to 10 data sets with uniform growth rate values for each gas. Unfortunately, the limited scope of this publication does not allow for a detailed analysis of all the data sets obtained. Therefore, Table II presents the statistical characteristics of homogeneous data sets with minimum (M_{min}) and maximum (M_{max}) values of growth rates for each of the nine gases analysed and the sum of gases in the hydrocarbon series (C_xH_y).

Analysing the data presented in Table II, it is easy to see that the values of the selective average gases growth rates in serviceable, normally operating transformers of non-hermetic construction differ significantly. This circumstance indicates the significant influence of the duration and operating modes of transformers, the characteristics of their construction, and the type of transformer oil used. It is clear that in order to ensure maximum reliability in detecting developing defects and minimising the risk of making erroneous decisions when using gases growth rates, their limit values must be determined taking these factors into account. Relatively low (close to zero) values of asymmetry coefficients suggest that the distribution law in question is symmetrical with respect to the mathematical expectation.

TABLE II. STATISTICAL CHARACTERISTICS OF HOMOGENEOUS ARRAYS OF GROWTH RATES OF GASES DISSOLVED IN OIL

Gas	Array	*N*	M_x	D_x	j_a	j_e
CH_4	M_{min}	169	0.107	0.412	0.332	4.316
	M_{max}	1192	0.431	41.270	0.173	5.603
C_2H_4	M_{min}	214	-0.001	0.228	0.272	4.357
	M_{max}	187	3.020	167.635	0.224	5.984
C_2H_6	M_{min}	141	0.017	0.069	0.568	4.778
	M_{max}	773	0.572	12.607	0.057	5.921
C_2H_2	M_{min}	416	0.028	0.281	0.101	3.627
	M_{max}	103	-0.162	7.849	0.009	5.452
C_xH_y	M_{min}	108	0.083	0.330	0.542	4.116
	M_{max}	566	1.009	69.986	0.137	5.929
H_2	M_{min}	148	0.112	0.840	-0.280	4.941
	M_{max}	179	0.227	43.089	-0.023	3.683
CO	M_1	245	0.346	23.197	-0.112	5.725
	M_2	441	12.268	3242.21	-0.152	3.658
CO_2	M_{min}	260	-0.932	380.163	-0.204	4.722
	M_{max}	1649	21.121	26805.1	0.253	6.527
N_2	M_{min}	51	-0.028	0.048	0.128	4.109
	M_{max}	277	0.546	6.638	0.526	7.738
O_2	M_{min}	87	0.008	0.079	0.213	5.717
	M_{max}	86	0.077	1.425	-0.025	4.621

[a.] In the table it is indicated: *N* – volume of selected values; M_x – selective average values, D_x – selective dispersion values; j_a – values of the selective asymmetry coefficient; j_e – selective excess coefficient values.

This proves that there is a certain equilibrium between the formation of new compounds and the diffusion of gases into the atmosphere. In turn, positive and non-zero values of the excess coefficients indicate that the distribution law in question has a more 'sharp' peak compared to the Gaussian distribution law.

The analysis of the distribution laws of gases growth rates in serviceable transformers of non-hermetic construction was performed using the 'ZR' programme developed at the Department of Electrical Power Transmission of the National Technical University 'Kharkiv Polytechnic Institute' [29]. This programme allows you to assess the conformity of an empirical sample to one of the following distribution laws: normal (Gaussian), exponential, extreme value distribution, beta distribution, gamma distribution, double exponential distribution (Laplace), logistic, lognormal, Rayleigh distribution, Weibull distribution, Cauchy distribution, etc. Pearson's criterion $\chi 2$ and the Kolmogorov-Smirnov criterion are used as criteria for agreement. The analysis showed that for all analysed homogeneous data sets, for each of the gases, the growth rate values can be described by a double exponential distribution (Laplace distribution). This law has a distribution density [28]:

$$p(x;\alpha;\beta) = \frac{\alpha}{2} \cdot e^{-\alpha \cdot |x-\beta|}, \quad (3)$$

where $-\infty<x<\infty$; $\alpha>0$ – scale parameter; $-\infty<\beta<\infty$ – shift parameter.

The values of the scale and shift parameters for the double exponential distribution law were calculated as follows [28]:

$$\alpha = \sqrt{\frac{D}{2}};\ \beta = M, \quad (4)$$

where *M* – selective average values; *D* – selective dispersion values.

Table III shows the values of the scale and shift parameters for the double exponential distribution law, as

well as the values of the χ^2 and Kolmogorov-Smirnov [28] goodness-of-fit criteria for homogeneous arrays with minimum and maximum values of gases growth rates. Fig. 5 shows a histogram of the empirical distribution and the density function of the double exponential distribution of the C_2H_2 growth rate for the M_{min} array.

As can be seen in Fig. 5, the density function of the double exponential distribution almost completely coincides with the histogram of the empirical distribution. This is also evidenced by the fact that the observed values of the Pearson and Kolmogorov-Smirnov criteria do not exceed the corresponding critical values of these criteria, which can be easily verified by analysing the results presented in Table III. Based on this, it can be concluded that the hypothesis that in serviceable non-hermetic transformers the values of gases growth rates correspond to a double exponential distribution is not rejected.

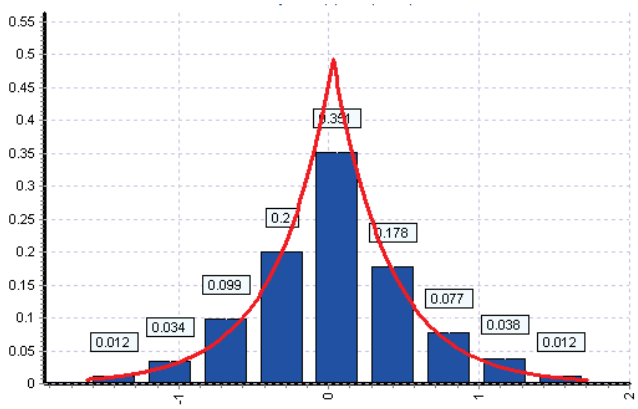


Fig. 5. Histogram of the empirical distribution and the density function of the double exponential distribution of the C_2H_2 growth rate for the M_{min} array.

TABLE III. VALUES OF THE SCALE AND SHIFT PARAMETERS OF THE DOUBLE EXPONENTIAL DISTRIBUTION LAW, AS WELL AS CALCULATED AND CRITICAL VALUES (AT α=0.05) OF THE PEARSON AND KOLMOGOROV-SMIRNOV CRITERIA FOR ARRAYS OF GROWTH RATES OF GASES DISSOLVED IN OIL IN SERVICEABLE TRANSFORMERS OF NON-HERMETIC CONSTRUCTION

Gas	Array	Parameters of the distribution law		The value of Pearson's criterion			The value of Kolmogorov-Smirnov criterion	
		α	β	f	χ^2_{est}	χ^2_{crit}	λ_{est}	λ_{crit}
CH_4	M_{min}	0.107	0.455	3	5.53	7.82	0.636	1.36
	M_{max}	0.431	4.544	6	11.32	12.6	0.533	1.36
C_2H_4	M_{min}	-0.001	0.339	4	8.52	9.49	0.734	1.36
	M_{max}	3.020	9.180	3	5.15	7.82	0.956	1.36
C_2H_6	M_{min}	0.017	0.187	3	6.01	7.82	0.963	1.36
	M_{max}	0.572	2.512	6	2.36	12.6	0.335	1.36
C_2H_2	M_{min}	0.028	0.376	6	9.28	12.6	0.750	1.36
	M_{max}	-0.162	1.991	2	3.49	5.99	0.723	1.36
C_xH_y	M_{min}	0.083	0.408	4	8.84	9.49	0.724	1.36
	M_{max}	1.009	5.921	5	2.86	11.1	0.405	1.36
H_2	M_{min}	0.112	0.650	3	6.47	7.82	0.942	1.36
	M_{max}	0.227	4.655	4	3.49	9.49	0.498	1.36
CO	M_{min}	0.346	3.413	3	0.10	7.82	0.119	1.36
	M_{max}	12.268	40.309	9	15.10	16.9	0.401	1.36
CO_2	M_{min}	-0.932	13.814	3	1.07	7.82	0.402	1.36
	M_{max}	21.121	115.804	6	7.42	12.6	0.483	1.36
N_2	M_{min}	-0.028	0.157	1	0.73	3.84	0.165	1.36
	M_{max}	0.546	1.825	2	4.53	5.99	0.524	1.36
O_2	M_{min}	0.008	0.199	2	2.69	5.99	0.553	1.36
	M_{max}	0.077	0.849	2	1.26	5.99	0.410	1.36

IV. ANALYSIS OF THE RESULTS OBTAINED

The analysis of transformer characteristics, based on DGA results that formed arrays with uniform gases growth rate values, revealed a number of interesting features. In particular, arrays with minimum gases growth rates were formed based on the results of DGA tests of transformers that were put into operation 10-15 years ago, loaded at no more than 20% of their rated capacity and filled with NITRO oil. Arrays with maximum gases growth rate values are formed based on the results of DGA transformers which have been in operation for more than 30 years, with a predominant composition of industrial consumers (coal mines, metallurgical and chemical enterprises), and have been operated for a long time with a load factor of more than 65% of the rated capacity. These transformers were originally filled with TkP oil, but during operation, oil of different grades was added. It should be noted that the results obtained fully coincide with the conclusions obtained in [30]. It is fundamentally important that that arrays with homogeneous gases rise rate values are formed based on the results of DGA tests of transformers with different rated voltages of 110 and 330 kV, different rated power and different types (two-winding, three-winding and split-winding).

The analysis showed that in transformers with identical operating times and modes, the rates of increase of different gases differ significantly, which is due, on the one hand, to differences in the energy levels required for gas formation and, on the other hand, to differences in the solubility coefficients of gases and their ability to diffuse into the atmosphere. Based on this, the limit values for the gases growth rates for each gas must differ.

The symmetry in the distribution of gases growth rates relative to mathematical expectations suggests that in serviceable, normally operating transformers of a non-hermetic construction, there is a certain equilibrium between the amount of gas formed and the amount of gas released into the atmosphere. At the same time, the probability of gases growth rates corresponding to the mathematical expectation far exceeds the probability of gases growth rates differing from the mathematical expectation, which determines the ‘sharpness’ of the distribution. Thus, the use of a double exponential distribution allows for adequate consideration of the characteristics of the gases formation mechanism in oils in serviceable non-hermetic transformers.

The obtained values of the scale and shift parameters for the distributions of gases growth rates in serviceable transformers of non-hermetic construction will allow determining the limit values of gases growth rates using statistical methods. This will make it possible to significantly reduce the risks that arise when making wrong decisions. However, for the practical implementation of this approach, it is necessary to analyse the laws of distribution of gases growth rates in transformers with various types of defects, which, given the multifactorial nature and complexity of the gas formation process, is a separate scientific and technical task.

V. CONCLUSIONS

It has been established that as a result of exposure to short-circuit currents, atmospheric and switching overvoltages, as well as prolonged overloads, the values of gases growth rates, in the absence of defects in the transformer, can significantly exceed the limit values of this

criterion, which can cause false rejection of transformers. It has also been established that the duration and operating modes of transformers, as well as the grade of transformer oils, have a significant impact on gases growth rates, which is fully consistent with the results obtained by other researchers. The statistical processing procedure is proposed that allows arrays with uniform gases growth rate values to be formed under conditions of limited a priori information. It has been established that in serviceable, normally operating transformers of non-hermetic construction, the values of gases growth rates can be described by a double exponential distribution, which adequately takes into account the peculiarities of the oil gas formation mechanism, in particular the presence of gases diffusion into the atmosphere. The scale and shift coefficients obtained from the analysis for transformers operating under different conditions will be used to adjust the limit values for each of the gases analysed, taking into account the established influence of operational factors.

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Detection Technology of the Start-Stop of a Car Using Accelerometer Data to Optimize Mobile Advertising

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***Abstract*—Advertising providers have long used mobile applications to deliver ads, which has proven effective. Drivers form a challenging audience for advertising because laws restrict phone use while driving, and reasonable rules govern drivers' phone usage. Nevertheless, advertisers still aim to reach this group with their ads. This article presents a method to detect vehicle start and stop events using only smartphone accelerometer data, without relying on GPS or other motion parameter measurements. This approach helps avoid legal risks associated with processing geolocation data. The algorithm follows a sequence of steps: it cleans accelerometer measurements from noise and gravity components; determines the direction of movement; projects acceleration onto this direction; calculates jerk; and classifies events using a neural network. A key aspect of the algorithm's operation is the recognition of the start and stop of the car if the mobile phone is mounted on the car's dashboard. The algorithm is implemented as a library integrated into a mobile app. The combination of a specific problem, a tailored solution, and large-scale real-world testing forms the scientific contribution of this research, compared to other works that address either broader tasks or different problems with alternative methods. The results demonstrate the algorithm's high accuracy and suitability for mobile advertising targeting drivers. Over 200,000 users worldwide use the commercial app monthly, incorporating this software solution.**

Keywords—information technology, safe advertising, accelerometer, car movement patterns, low pass filters, jerk, braking, class library

I. Introduction

In recent years, the number of electronic devices people hold and wear has significantly increased, including smartphones, smartwatches, fitness trackers, virtual reality glasses, and so on. These devices have built-in physical sensors, such as magnetic field, proximity, temperature, pressure, and acceleration sensors, used by software applications to identify specific aspects of a person's life. A whole trend related to human activity recognition has emerged it has found applications in fields such as healthcare [1], applications developed for smart homes, and so on. The human activity recognition technology is widely used in smartphones for real-time lifestyle tracking [2, 3]. The monitoring objectives are quite diverse, and sensor data is used in various combinations; often, GPS data is used alongside built-in sensors. For instance, almost all research on the movement of objects on roads [4] is conducted using GPS data, driving habits of drivers are studied [5], and sometimes inertial measurement units are also used [6].

Meanwhile, advertising providers have long utilized mobile applications for ad delivery. It is highly effective because we consistently carry our smartphones, allowing digital advertising to reach our awareness. As is known, advertisers categorize their audience, particularly by occupation. The audience of drivers is vast, and it is challenging for advertisers to reach them. On the one hand, many drivers use their mobile devices while driving, which presents an excellent opportunity for advertisers. On the other hand, many countries have restrictions on using such devices while driving, which is appropriate. There is a need among advertising companies to determine whether a car was moving or not without using GPS, as information about a person's location would impose additional legal responsibilities [7].

The advertising industry's growing need for secure methods of delivering content to drivers prompted this research. Campaigns aim to offer drivers ad impressions only when the vehicle is not participating in traffic and thus does not pose a road hazard. The main challenge has become the need to determine the mode of travel of mobile device users without using GPS modules, as location information requires additional legal obligations for protecting historical data. Only accelerometer data should be used, as this sensor is present even in older mobile device models, providing the application with competitive advantages. The use of GPS data is also prohibited, as this could lead to the disclosure of the driver's location, which is undesirable for the provider. The result of this work is information technology implemented in a commercial application.

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II. Analysis of Recent Research and task

Smartphones have long been used as practical tools for monitoring driver behavior. In their review [4], the authors concluded that accelerometers, gyroscopes, and GPS are the most commonly used sensors. In some cases, inertial measurement unit data is used in addition to GPS [5]. Popular data processing approaches include deep neural networks, event-based driver behavior recognition, and machine learning algorithms [6, 7].

In [7], driver behavior recognition was performed to activate accident detection. Four actions were recorded using the DriveNeXt neural model: driver entry/exit and vehicle boarding/alighting. The model achieved high accuracy but utilized a smartphone's accelerometer, gyroscope, and GPS.

It's worth noting that accelerometers are the most widely used sensors in wearable devices and are generally considered privacy-preserving, unlike sensors such as microphones and cameras. However, research [8] proves the opposite: accelerometers in mobile devices can reveal information about user activities, location, identification, device inputs, health status, body characteristics, age and gender, moods and emotions, and personality traits.

Using accelerometer data alone to determine motion elements is rare in modern research, as most studies use GPS data. For instance, work [9] presents a mobile application that collects accelerometer data to detect road defects. This data collection process involves two participants: a driver and an assistant who records the road situation. It is obvious that the data collection process described in [9] could be significantly streamlined by simultaneously recording driver voice feedback and accelerometer data during test drives. Furthermore, it's crucial to establish the correlation between the coordinate axes of the two objects—the vehicle and the accelerometer—*after* the phone has been placed in the desired position on the car's dashboard, rather than beforehand.

Work [10] focuses on investigating smartphone positioning variability. However, during the positioning process, its authors comprehensively utilize data from accelerometers, gyroscopes, linear accelerometers, and gravity sensors.

An accelerometer-based data acquisition system for public bus driver behavior recognition is detailed in [11]. While the processing algorithm effectively cleanses measurements from all three accelerometer axes using a sliding window method, it notably omits any conversion of these values into the vehicle's coordinate system. This limits the approach's applicability to scenarios where the accelerometer's coordinate system is perfectly aligned with the vehicle's direction of travel. It is known that a mismatch between the phone's coordinate system and the direction of motion is a typical problem in motion modeling [12]. Consequently, this method is unsuitable for general use cases where the accelerometer's coordinate system does not coincide with the direction of motion, such as when a phone is used inside a car.

To collect data in the work [13], a mobile phone was placed horizontally with the Z-axis pointing up and the Y-axis aligned with the car's direction of motion. Thus, they only considered the values of the two axes x and y, ignoring the Z-axis. This approach can lead to significant errors when driving uphill or downhill. The analysis of typical driving operations in [14] examined characteristic patterns in accelerometer data obtained from a mobile phone for each X, Y, and Z axes. However, this simplification can introduce grave errors if the phone isn't rigidly mounted vertically and oriented with the car's direction of travel. Therefore, to apply the proposed gesture recognition method, it's first necessary to determine the direction of motion vector within the phone's coordinate system and then translate the acceleration vector into the motion direction's coordinate system.

The reviewed articles demonstrate a high level of research, but they have limitations. These include implementation complexity, limited scalability, motion recognition delays of up to 5 minutes, and other issues. None of the methods reviewed can be used in a software tool that quickly detects when a vehicle stops or starts moving. Detective speed is crucial because the vehicle's stationary period might be brief. This is precisely when it makes sense to display advertisements on the phone, as the driver is more likely to view them during this window. The probability of viewing an ad significantly decreases while driving. Another important consideration is the risk of causing an accident due to untimely ad display, which could lead to financial losses for the advertiser instead of generating revenue.

Taking into account all the above-mentioned limitations and factors, it is necessary to develop an alternative method that takes into account all the relevant factors to create a technology for safe display of advertising in a car on a mobile device. The study aims to develop a technology for safely displaying advertising in a vehicle. To achieve this goal, it is necessary to solve the following tasks: analyze existing methods for determining the start-stop moment of a car and their application to the task at hand (which is presented above); develop an algorithm for processing accelerometer data for automatic detection of start-stop moments obtained from a mobile device mounted in a car (the only requirement for the location of the phone is its securely affixed to the vehicle's dashboard); conduct field experiments and obtain relevant data for further research.

III. Motion Detection Algorithm Using Mobile Phone Accelerometer Data

As is known, an accelerometer measures the acceleration applied to the device in which it is embedded. Like all similar sensors, a mobile phone accelerometer measures forces acting on its active mass, including Earth's gravity, and we can use equation (1) to obtain the acceleration value.

$$a = -g - \frac{1}{m}\sum F_s\,, \tag{1}$$

where a – device acceleration, g – gravitational acceleration, m – active mass, F_s – forces directly applied to the sensor.

Accelerometers adhere to a standard sensor coordinate system. Simplistically, when a device is stationary (e.g., placed on a table), the accelerometer reads $a \approx 9.81\ m/s^2$. During free fall (accelerating toward the ground), it registers $a \approx 0\ m/s^2$. Thus, gravity's contribution must be subtracted to measure the device's true acceleration.

A phone's position in a car is undetermined, and its coordinate system (Fig. 1a) isn't aligned with the vehicle's forward motion. The key objective is to resolve the car's acceleration vector f along the direction of travel (Fig. 1b) within the accelerometer's coordinate system.

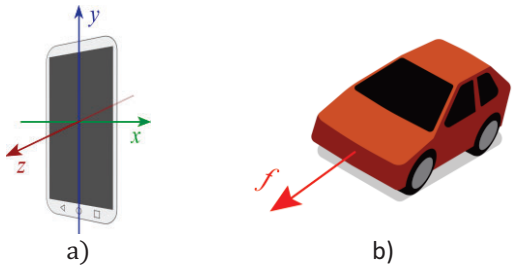


Fig. 1. a) Coordinate system of the mobile phone accelerometer. b) Vector of the forward direction of the car

The jerk vector, defined as the derivative of the acceleration vector, is key to analyzing an object's motion and dynamics. This vector indicates the rate of change in an object's acceleration and is a vital tool for diagnosing its motion and detecting events like the initiation or cessation of movement, braking, or acceleration. After determining the projection values of the acceleration vector onto the translational motion direction vector and the jerk values, a separate function can be used to recognize typical acceleration and braking patterns, as well as the start and end of motion. Such a function can be implemented by machine learning algorithms using neural networks. Thus, the generalized accelerometer data processing algorithm can be represented as a sequence of steps: select the motion direction vector, calculate the acceleration projection module onto the motion direction vector, calculate the jerk, and identify an event.

Briefly, regarding each step, the following can be said. Select the motion direction vector. To determine the motion vector in the phone's coordinate system, the gravity vector was first isolated using a low-pass filter. Then, the contribution of gravity was removed from the original data. The resulting vector represents the phone's own acceleration. The filtration coefficient was calculated as $t/(t + dT)$, where t is the time constant representing the latency the low-pass filter adds to sensor events, and dT is the interval between two consecutive events. For the first step, the time constant was chosen to be 2 seconds, and the average event delivery rate was configured to 100Hz (i.e., the interval between two events is 0.01 secs).

The low-pass filter is applied again at the next stage, but with a different filtering coefficient — the time constant was set to 0.25 seconds. Then, the same filter is applied using the parameters from the first phase to eliminate the effect of damped oscillations along the Z-axis (assuming this axis is considered the forward movement direction). However, we took the Z-axis values as absolute (without sign) this time. In this way, the vehicle's motion vector was obtained in the coordinate system of a mobile phone rigidly fixed to the dashboard with the screen facing the driver, and therefore, against the direction of movement. The phone's position on the dashboard can vary but must be fixed securely (placed in a phone holder).

Calculate the acceleration projection module onto the motion direction vector. The modulus of the projection of the phone's own acceleration vector onto the vehicle's motion vector was calculated by dividing the scalar product of these vectors by the modulus of the vehicle's motion vector.

Calculate the jerk. The jerk value was determined as the ratio of the change in acceleration to the change in time.

Identify the event. At the last stage, the algorithm determined the vehicle's current maneuver by comparing the calculated acceleration and jerk values over a specific time interval with patterns derived from typical car maneuvers. To recognize jerk and sudden braking events, it is implemented a perceptron-type neural network, trained it, and achieved 97% accuracy in correctly classifying these events.

IV. Conducting Experiments and Processing Data

To collect test data, an experiment was conducted in a real-world environment using an application developed for use in a car. The experiment was performed with a vehicle equipped with a 1.2-liter gasoline internal combustion engine and a continuously variable transmission. The car was driven by a driver with more than five years of driving experience. The circumstances described above affect driving style (i.e., acceleration and braking), which in turn influences the accelerometer readings. Six mobile phones with different configurations were used to collect data.

Initially, a control test was performed to determine the measurement accuracy of each smartphone's accelerometer. The essence of the control test was to record accelerometer data while the phone was stationary on a horizontal surface. According to the official Android API documentation, when a device is stationary on a table and not accelerating, the accelerometer should display a vector value equal to the acceleration due to gravity. It was found that for all selected test phones, the measurement error ranged from −2.41% to +2.20%, which is within the normal range according to the manufacturers' documentation.

During the experiment, five typical maneuvers and a series of compound maneuvers were performed using each phone. For each maneuver, the initial state—such as "standstill," "engine running," and "straight ahead"—was recorded. Among the operations considered, "engine start," "smooth start," "sudden start," "smooth braking," and "sudden braking" were highlighted. The car routes were selected from city streets with asphalt surfaces and a certain number of mandatory speed limiters (so-called "liar policemen"). The experiment was conducted during nighttime hours after 22:00 local time to avoid heavy traffic. These conditions allowed the desired maneuvers—and the experiment as a whole—to be carried out in a comfortable and safe environment. Each experiment was accompanied by voice comments from the driver, which were recorded synchronously with the accelerometer data. These voice comments were useful for reconstructing each experimental situation during data processing and when running the program in debug mode.

Based on the analyzed accelerometer data and transcribed voice comments, it was determined how the accelerometer readings changed during different maneuvers. After processing and reconstructing the experimental scenarios, the analysis focused on identifying patterns and characteristics in the accelerometer data. This included recognizing typical patterns corresponding to each type of maneuver, as well as detecting possible deviations or anomalies that could affect recognition accuracy. This stage was necessary to determine the key features used in training the perceptron model for recognizing the start and stop of movement.

The data obtained during the abrupt acceleration test can be seen in Figure 2, which shows changes in the acceleration vector values for each of the x, y, and z axes. During visual assessment, several notable sections of the graph were observed (around 0.25 s, 1.4 s, and 2.75 s), which stood out significantly from the overall background. The official

Android API documentation indicates that the actual sampling rate may vary within 55–220% of the nominal value. During data collection, the accelerometer was set to operate at a frequency of 100 Hz, or one reading every 10 ms.

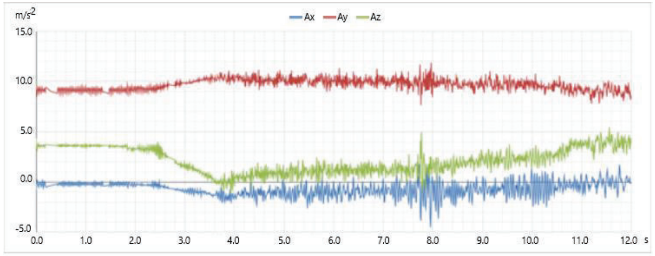


Fig. 2. Graph of accelerometer data obtained during hard acceleration of the car

Since the Earth's gravity always affects the measured acceleration, it must be separated by applying a low-pass filter. Figure 3 shows the three-axis free-fall acceleration obtained from the accelerometer output after filtering.

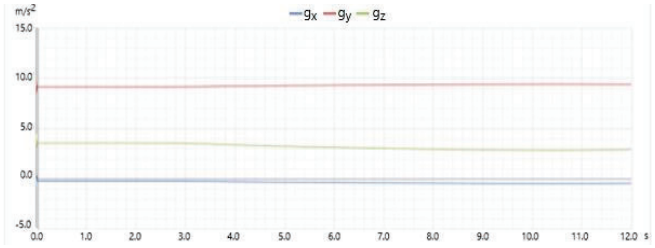


Fig. 3. Acceleration of free fall in 3 axes obtained by filtering from the accelerometer output

The algorithm's next step determines the device's true acceleration by removing the contribution of gravity, specifically by subtracting the two vectors coordinate by coordinate. The next step involves filtering to obtain acceleration values without noise. Figure 4 shows the result.

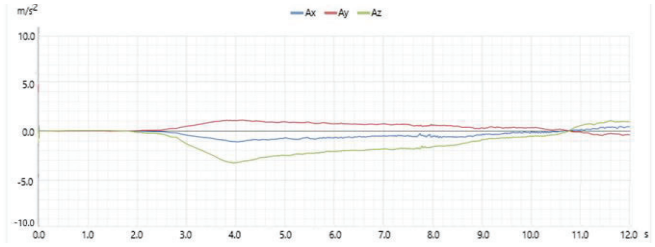


Fig. 4. Acceleration graph excluding the effects of Earth's gravity and noise

Figure 5 displays the magnitude of the projection of the acceleration vector onto the direction-of-motion vector (blue graph), calculated by dividing the scalar product by the magnitude of the direction-of-motion vector, as well as the magnitude of the jerk vector, which is the time derivative of acceleration (red graph). The graph clearly illustrates the pattern observed during the acceleration of a car: a sharp, short-term increase in jerk, followed by a smoother rise in acceleration, which then remains in the positive half-plane.

This same behavior was observed in data from other experiments. For example, Figure 6 shows a graph of the magnitude of the projection of the acceleration vector onto the direction vector and the magnitude of the jerk vector during sudden braking. This graph clearly illustrates the braking pattern: a sharp decrease in jerk and acceleration values. However, the graph also reveals that the acceleration values return to their previous levels, even though the vehicle has come to a complete stop after braking. Thus, it was found that accelerometer error can increase significantly in this situation.

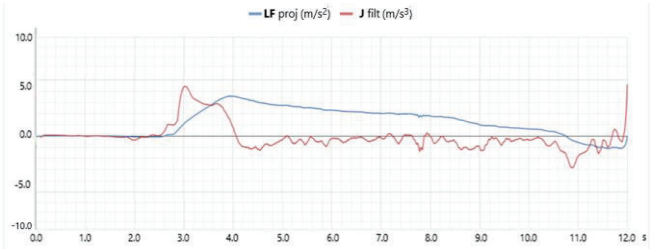


Fig. 5. Graph of the magnitude of the projection of the acceleration vector onto the direction-of-motion vector and the magnitude of the jerk vector.

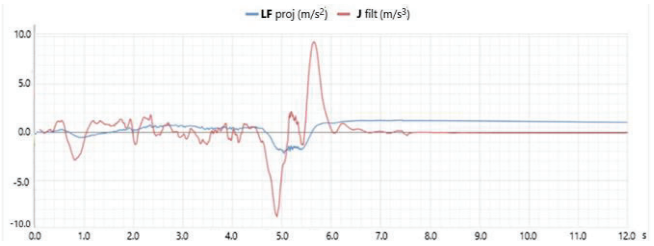


Fig. 6. Graph of the modulus of the projection of the acceleration vector onto the vector of the direction of movement and the modulus of the jerk vector during sharp braking

As a result of the experiment, we obtained 69 accelerometer data files in CSV format and 69 corresponding audio files in WAV format. Analyzing the data graphs, we identified a correlation between calculation accuracy and the driver's voice comments recorded during the experimental runs. We used the collected data to train a neural network.

V. Implementation of Research Results

The developed algorithm for recognizing when a car starts and stops moving was packaged as a JAR-format class library for integration into mobile applications. The first application to implement this library was Private Screenshots, which, according to an analytical report from Google Analytics for the period from March 1 to March 31, 2025, has over 220,000 active users. A unique feature of this application is that some of its functionality is integrated with foreground services, meaning the process remains active even if the user switches to another application. Additionally, since the application already supports accelerometer usage, implementing the library does not significantly alter the application's behavior. It should be noted that this library can be used in a microservice front-end architecture [15].

The implementation was carried out to collect data about the accelerometer models used on users' devices and to record the number of users for whom start and stop events were detected. These users can be segmented into a separate audience for whom specific advertisement display settings can be applied. The application's functionality remained unchanged, and event recording and data collection only took place for users who did not opt out of analytical data collection and processing in the application's settings. The library can be controlled via the Firebase Remote Config service, making it possible to stop or resume data collection without the need to publish a new version of the application.

Data collection from devices was conducted over the course of one week using the built-in sensors of smartphones belonging to current application users. During this period,

75677 users participated in the data collection. A total of 3,038 different phone models were identified (see Table 3, row "Total"), and 159 different accelerometer models were detected. For 45,808 (60.53%) users, representing 987 (32.49%) different phone models, the accelerometer model could not be determined. For these users, the default value for the accelerometer model was set to a predefined string.

The list of accelerometer models received was sorted in descending order by the number of phone models equipped with each accelerometer. In this list, the top 10 most common accelerometer models were found in 41.34% of all phone models and among 21.79% of users (see Table 1, row "Total").

TABLE I. THE 10 MOST COMMON ACCELEROMETERS BUILT INTO APP USER'S PHONE.

Position	Accelerometer model	Phone model		Users	
		number	%	number	%
1	LSM6DSO	357	11.75	4725	6.24
2	LSM6DSM	194	6.39	1364	1.80
3	ICM4X6XX	169	5.56	2663	3.52
4	BMI160	166	5.46	1270	1.68
5	LSM6DSL	139	4.58	2604	3.44
6	BMI26X	55	1.81	227	0.30
7	SC7A20	48	1.58	1387	1.83
8	LIS2HH12	45	1.48	1053	1.39
9	LSM6DS3C	42	1.38	753	1.00
10	LIS2DS	41	1.35	442	0,58
Total in the top 10		1256	41.34	16488	21.79
All in research		3038	100	75677	100

Thus, it was found that 5 out of 6 phones used during the experiment had accelerometers included in the list of the 10 most common among users of the application. Such accelerometer models are present in 41.34% of all phones in the sample. Therefore, the selection of test phones can be considered representative. Overall, the results indicate that the implemented algorithm is functional and can be applied within a certain narrowly specialized category of applications.

VI. CONCLUSION

This paper presents technology for displaying ads on portable devices in vehicles, based on an algorithm detecting vehicle start-stop events. The following tasks have been solved: analyzing existing methods and developing an algorithm to detect stop-and-go moments using accelerometer data from mobile devices mounted in vehicles.

The algorithm accurately identifies motion start and end regardless of phone orientation, unlike prior methods. Unlike other studies using multiple sensors or broader problems, this work relies solely on accelerometer data, avoiding legal risks linked to geolocation and enhancing compatibility with older phones.

The scientific contribution combines a targeted problem and solution using multi-stage filtering, vector math, neural network, and extensive real-world testing. The technology is implemented in a commercial app with over 200,000 monthly active users. It enables ads to display only when drivers are stopped, reducing distractions and accident risk. This ethical approach shifts mobile advertising from aggressive targeting to responsible, context-aware interaction.

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Mobile explosive object detection system for humanitarian demining needs

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***Abstract*—This work details an AI-driven framework designed for explosive object identification, leveraging unmanned aerial vehicles (UAVs) and a fusion of deep learning techniques. The core of the system combines the YOLOv8n architecture with transformer networks. To ensure relevant and effective training, we curated a bespoke dataset mirroring the explosive hazards present in the Ukrainian context. Performance metrics revealed a strong capability: the model scored 92.3% accuracy while maintaining 18 frames per second during real-time detection assessments, even under adverse conditions. Practical field deployments validated its applicability for rapid integration into existing emergency response protocols. Comparisons highlighted its advantages, surpassing numerous existing commercial solutions in terms of cost-effectiveness, operational efficiency, and ease of customization. Plans for further enhancements encompass the incorporation of ground-penetrating radar data and improvements to offline computational speed.**

Keywords—humanitarian demining, explosive object detection, YOLOv8, deep learning, UAV

I. Introduction

Modern Ukraine is facing one of the world's largest landmine threats. According to the State Emergency Service of Ukraine [1], as of the beginning of 2024, about 30% of the country's territory (more than 174,000 km²) is potentially contaminated with explosive objects of various types. This area is comparable to the territories of countries such as Tunisia or Uruguay and poses a serious threat to civilian life and the process of rebuilding infrastructure.

The problem is complicated by the variety of explosive objects, including both industrial anti-personnel and anti-tank mines and numerous improvised explosive devices of non-standard design [2]. Traditional demining methods, such as manual surveys with metal detectors or the use of specially trained dogs, are extremely labor-intensive and pose a great risk to the lives of deminers [3]. Over the past year, more than 1,200 cases of explosive detonations have been recorded, resulting in numerous casualties.

In view of this, the development of automated detection systems using unmanned aerial vehicles and artificial intelligence methods is becoming an extremely important task. Modern deep learning technologies, in particular computer vision algorithms, offer new opportunities to effectively address this problem [4]. However, the existing developments have not yet reached the required level of accuracy and reliability for their full practical application.

The aim of this work is to create a high-precision explosive object detection system that combines the advantages of unmanned technologies and advanced deep learning algorithms. The main objectives of the study are: analysis of existing methods, creation of a specialized data set, development of a hybrid neural network architecture, implementation of a software and hardware complex, and comprehensive testing [5, 6].

The scientific novelty of the work is the development of a unique hybrid architecture that combines the advantages of convolutional neural networks (YOLOv8) and transformers, specially optimized for the detection of explosive objects. The practical significance is that the system has already undergone successful field tests and is being prepared for implementation in the work of emergency services.

II. Literature Review

Modern scientific literature presents a wide range of approaches to the problem of detecting explosive devices. Traditional methods, such as metal detectors, are based on the detection of metal components of explosive devices. However, their effectiveness is significantly reduced when it comes to modern plastic mines or improvised explosives that contain minimal metal. Research by Heuschmid et al. has shown that the sensitivity of standard metal detectors to such objects reaches only 40-45% [5].

Ground penetrating radar, which analyzes the structure of the soil, shows better results, but, as Sato note in their work [7], they take a long time to survey even small areas and are extremely sensitive to soil type and moisture. In addition, both of these methods require the direct presence of the operator in the area of potential danger, which significantly limits their practical application in conditions of large-scale contamination of the territory.

Also interesting are the methods that use optical [8], hyperspectral [9] and multispectral [10, 11] imaging to search for explosive objects.

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Advances in artificial intelligence have opened up new perspectives in solving this problem. In particular, deep learning methods allow analyzing terrain images and automatically detecting potentially dangerous objects. Research conducted by the Ultralytics team has shown that architectures based on convolutional neural networks (CNNs), such as YOLO (You Only Look Once), demonstrate high processing speed and fairly good accuracy for standard cases [12, 13]. Their advantage lies in their ability to process images in real time, even on limited hardware resources, making them ideal for use on board unmanned aerial vehicles.

Transformational architectures, originally developed for natural language processing, have also found their way into computer vision [14]. The work of Dosovitskiy et al proves that they are particularly effective for analyzing complex scenes where global context is important [15]. However, their main disadvantage is high computational complexity, which limits their use in real time, especially on the onboard computers of drones.

Hybrid approaches that combine the strengths of CNNs and transformers are of particular interest. Research [16] shows that such architectures can achieve a good balance between accuracy and speed, although they require a more complex training process and a larger amount of training data.

An analysis of commercially available systems revealed that most of them have serious limitations. For example, the Mine Kafon Drone system developed in the Netherlands shows an accuracy of about 85% at a processing speed of 10 frames per second, but its cost exceeds $20,000, making it unaffordable for mass use. The HALO Trust system, which is used in the UK, has even worse accuracy (78%) at a price of about $15,000 [17].

Ukrainian developments in this area, such as the Phoenix system from NTU KPI, show encouraging results, but, as noted in the SES reports, have not yet reached the required level of accuracy for full practical use. The main problems are insufficient training data and limited opportunities for field testing.

An important aspect is the integration of different types of sensors. Most modern systems use only RGB cameras, which significantly limits their effectiveness in difficult conditions. At the state standard level, Ukraine has regulated using UAV and satellite imagery analysis [18] for surveying areas affected by explosive ordnance [19].

III. Image Analysis System

YOLOv8n is a model that doesn't waste time. Its neural networks operate at an extraordinary speed, processing images in a matter of moments. This makes it ideal for real-time applications where every second counts. Imagine a bomb squad using the YOLOv8n to survey an area for explosives. With this lightning-fast speed, he can respond to the threat immediately, saving lives.

This model is also extremely accurate. Its algorithms can clearly recognize objects in images without making mistakes. This makes it an indispensable tool in industries where mistakes can have dire consequences, such as medicine or mine clearance. For example, a doctor could use the YOLOv8n to analyze X-rays to detect a tumor. Thanks to this exceptional accuracy, he can make a timely diagnosis and start treatment, saving the patient's life. It is the same with sappers.

We chose this particular model because it has demonstrated high accuracy, speed and reliability, making it an excellent basis for developing a humanitarian demining service.

The development of a dataset with explosive objects for the YOLOv8n model is a key step in the process of creating an automated explosive detection system.

This dataset will be the basis for training the YOLOv8n model [20] to recognize and classify different types of explosive devices.

The first step in creating a dataset is data collection. It involves:

- Diversity: photographs should be taken from different viewing angles, under different lighting conditions and against different backgrounds, which will enable the model to generalize information and recognize explosive objects in a real-world environment;

- Quality: the use of high-resolution images with clear details and minimal noise significantly increases the likelihood of obtaining a well-trained model;

- Quantity: the larger the volume of collected images, the better the trained model will be.

High-quality images are crucial for training the YOLOv8n model, as they allow it to learn the visual characteristics of explosive objects with high accuracy. The variety, quality, and volume of images in the dataset play an important role in determining the model's performance.

As a result, in the future, a trained model for mine detection and classification will be developed, and the accuracy and efficiency of this model will be tested.

For the research, we used the open dataset "Explosive Object Detection Dataset" hosted on the Roboflow Universe platform [21]. This dataset consists of 440 images of explosive objects, including mines, with annotations in the YOLO format. The images show a variety of objects located in environments as close to reality as possible. This makes the dataset suitable for training and testing computer vision models for UXO detection.

The author's dataset was created using the above dataset. It was developed taking into account the specifics of the local theater of operations, the characteristics of explosive objects found on Ukrainian territory, and typical UAV shooting conditions [22]. The process of creating our own set involved increasing the volume of images, adjusting annotations, and adding variety to viewing angles and background images. This has a positive impact on the efficiency and stability of the models trained on the basis of this set [23].

In our case, the data source is photographs taken by drones that survey areas where there may be a threat. The RGB images are based on 640x640 pixels, which is the optimal compromise between accuracy and processing speed.

YOLOv8n requires images that meet the following minimum requirements:

- the image format must be JPEG or PNG;

- images must be square;

- the resolution should be 640x640 pixels.

Each image needs markup, i.e. information about the objects on it. We use the YOLO format, where for each object we define:

- class (0 – shell, 1 – patron, 2 – mine, 3 – grenade);

- coordinates of the center of the object (Fig. 1);

- overall dimensions of the object: X-axis object center, Y-axis object center, width object, height object (Fig. 2).



Fig. 1. Coordinate dimensions of an object (X-axis and Y-axis)

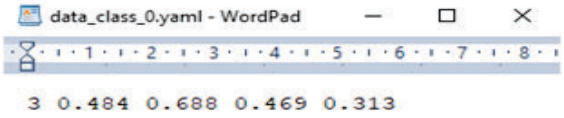


Fig. 2. Overall dimensions of the object (percentage from 0 to 1)

Since real photos can be taken from different angles, with mixed lighting and sharp changes in contrast, data augmentation is performed to improve the quality of the model:

- random rotation of images;

- change of brightness and contrast;

- adding noise;

- reflection of the reflection;

- adding "dirt" or partial overlap.

This makes it possible to train the model to work in conditions close to real life, when, for example, a mine is partially hidden in the grass.

The environment for training and testing the model was deployed on Google Colab (for initial training) and on a local server (for further integration and experimentation).

It was installed:

- Python 3.10;

- libraries: ultralytics, opencv-python, numpy, matplotlib;

- GPU acceleration systems: CUDA, cuDNN (on a local server);

- visualization of the results in Matplotlib.

This step allowed us to effectively test the model in conditions as close to the field as possible.

Using bash, execute the command and install the required library:

- Pip install ultralytics.

A specialized dataset containing more than 500 images of explosive objects (including grenades, anti-personnel mines, cluster munitions) was collected and annotated.

For the grenade in Fig. 1 (640×640), the coordinates are:

x_center = 0.484,

y_center = 0.688,

width = 0.469,

height = 0.313.

A string is formed from this (Fig. 2).

The model was trained for about 45 minutes on a GPU (Google Colab), providing high quality recognition after 60 epochs:

- the previously trained model (yolov8n.pt) is used;

- augmentation is used: flip, mosaic, brightness shift;

- saving the best version – best.pt.

Two key criteria are used to determine the quality of model training: the loss function and precision. They allow us to assess how well the model has mastered the recognition of explosive objects in images [19].

The loss function determines the difference between predicted and actual values. In the context of YOLOv8, it includes several components: coordinate loss (bounding box regression), objectness loss (confidence loss), and classification loss. General view of the formula:

$$Total\ Loss = \lambda_{box} \cdot Loss_{box} + \lambda_{obj} \cdot Loss_{obj} + \lambda_{cls} \cdot Loss_{cls},$$

where: $Loss_{box}$ – error in determining the coordinates of the object; $Loss_{obj}$ – error in predicting the presence of the object; $Loss_{cls}$ – error in classifying the object; λ – weighting coefficients of each component.

Precision reflects the proportion of objects identified as positive (i.e. explosive) that are actually explosive. It is calculated by the formula:

$$Precision = \frac{TP}{TP+FP},$$

where: TP (True Positives) – the number of correctly detected explosive objects; FP (False Positives) – the number of objects that were mistakenly recognized as explosive.

The graph below visualizes these indicators, showing the changes in the loss function and accuracy of the model over 60 epochs of training.

Figure 3 shows the dynamics of changes in loss and accuracy during the training of the YOLOv8n model over 60 epochs.

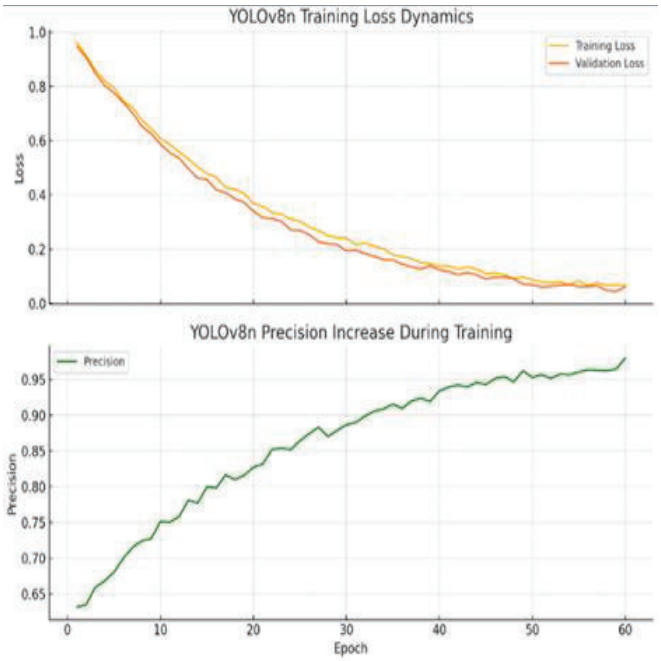


Fig. 3. Training dynamics of the YOLOv8n model

The upper part of the graph shows a decrease in the loss function on both training and validation samples. This indicates a gradual alignment of the model with the training data and effective error minimization.

The lower part of the graph shows the increase in precision, which steadily increases with the number of epochs. This dynamics confirms that the model is gradually learning to better recognize explosive objects [23].

After training was completed, the model was tested on fresh images. A photo of a grenade against a vegetation background was used as an example of an input image. Recognition was performed by overlaying a rectangular frame that highlighted the object.

After testing, we obtained the following results:

- grenade detected with a probability of 0.89;

- clear positioning on the image;

- processing time: 15 ms per 1 frame;

- successful integration into a prototype recognition system on an unmanned aerial vehicle.

We present the confusion matrix obtained after testing the YOLOv8n model (Fig. 4), when the training process has already been completed.

The predominant number of correct classifications for all four classes – projectiles, ammunition, mines and grenades – is observed along the main diagonal of the matrix. Minor deviations from this diagonal indicate erroneous predictions, in particular, cases where mines were identified as grenades. In general, the matrix analysis confirms the model's ability to successfully distinguish between different types of explosive objects in most cases.

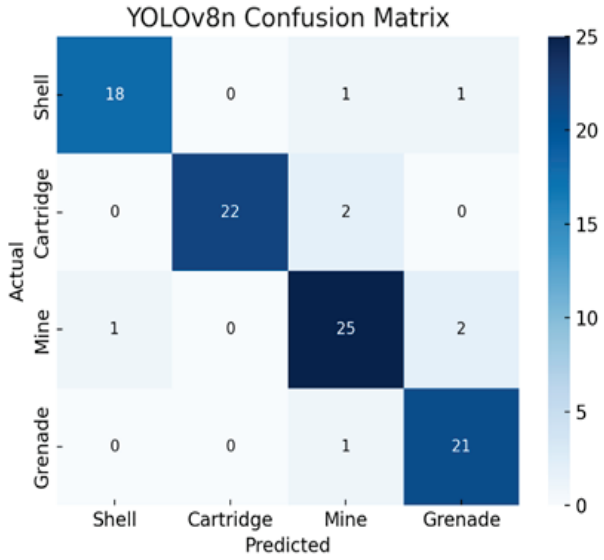


Fig. 4. Confusion matrix

We demonstrate the practical use of the trained model in conditions (Fig. 5) that are as close to realistic as possible.



Fig. 5. Relative detection probability

The image shows the detection of a grenade-like object partially hidden in the grass. The model successfully detected it with a rectangular detection box and identified the class with a probability of 0.89. The high probability and accurate positioning demonstrate the system's efficiency even in difficult visual conditions. This example emphasizes the practical applicability of the model for highly accurate and fast identification of hazardous objects in real field conditions.

IV. Analizis of Results

The data obtained from the study allow us to draw a number of key conclusions regarding the effectiveness of the proposed method of detecting explosive objects. The system has demonstrated a consistently high level of accuracy under various conditions of use, which confirms the correctness of

the chosen architecture and methods of information processing.

A comparative analysis with existing analogs reveals significant advantages of the developed solution. For example, the commercially available Mine Kafon Drone system shows an accuracy 7-9% worse than our development, at a much higher price of equipment [21]. It is worth emphasizing that the proposed system demonstrates better performance in detecting small-sized objects that pose a particular threat to the civilian population.

The efficiency of work in difficult conditions deserves special attention. The 5-8% decrease in accuracy in the rain or in the presence of dense vegetation is much lower than in most similar systems. This was achieved through careful augmentation of the training data and the use of auxiliary sensors (thermal imager and lidar).

A characteristic feature of our development is its adaptability to different types of explosive objects. The system performs equally well in detecting both standard-type industrial mines and improvised explosive devices, which is especially relevant in the context of modern conflicts. This was achieved by carefully selecting a training sample that included all major types of threats.

The energy efficiency of the system is another of its strengths. Power consumption of 25-30 watts allows the use of standard batteries of unmanned aerial vehicles without significantly reducing their operating time. This distinguishes our development from analogs that often require additional power sources.

However, there are certain limitations that need to be taken into account. The most noticeable is the decrease in accuracy in extreme weather conditions (heavy fog, rain). The system also cannot yet effectively detect objects deeply buried in the ground (more than 20 cm). These aspects require further improvement of the algorithms.

An important result is the practical applicability of the system for use in real-world conditions. Field tests have proven that the development is ready for implementation in the work of emergency services. The system can already be used for:

- initial survey of potentially hazardous areas;

- monitoring of already cleared areas to identify possible residual threats;

- creation of digital mine risk maps;

- training of personnel in remote detection of ERW.

Future improvements to the system focus on several important aspects. The most promising is the integration of subsurface sensing radars, which will significantly improve the identification of hidden objects. It is planned to expand the training data using images with new types of potential threats. An important area is the optimization of algorithms for work in conditions of limited access to communication, which involves carrying out part of the calculations directly on the unmanned aerial vehicle.

The social value of this development is hard to overestimate. Implementation of the system has the potential to significantly reduce the danger to sappers and speed up the process of clearing areas of explosive ordnance. This is especially important for agricultural regions, where mine danger poses serious obstacles to the resumption of normal economic activity.

The economic effect of using the system is to reduce the financial costs of demining. Preliminary calculations show that the use of the development can reduce costs by 30-40% compared to traditional methods by reducing the duration of work and the number of personnel involved.

Compared to its international counterparts, the Ukrainian development looks competitive both in terms of technical performance and cost. This opens up opportunities for its export to other countries facing the problem of mine threats.

V. Conclusion

The study confirmed the effectiveness of the proposed approach to detecting explosive objects using unmanned aerial vehicles and deep learning methods. The developed system showed an average accuracy of 92.3% and a processing speed of 18 frames per second, which is a significant improvement over traditional approaches.

The created hardware and software system integrated into the UAV combines data from various sensors and demonstrates stable operation in various conditions, including perfect visibility and difficult weather conditions. An important feature of the system is its ability to detect both industrial mines and improvised explosive devices with high efficiency.

Field tests have demonstrated the practical applicability of the development for real-world use. The system improves results by 25-30% compared to analogues, while reducing costs and energy consumption. This opens up opportunities for its large-scale implementation in the activities of rescue services.

Future research will focus on improving the algorithms to work in extreme conditions and integrating additional sensors. Special attention should be paid to adapting the system to the needs of different agencies and developing training programs for operators. The data obtained indicate great potential for further development in this area.

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MatViz3D Software Tool for Simulation of Polycrystalline and Hybrid Materials

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Abstract – **This study presents a significant extension of the MatViz3D software environment, initially designed to generate polycrystalline microstructures, to support a broader range of material systems. In response to the growing demand for versatile computational tools in materials science, the enhanced MatViz3D platform now enables the modeling of diverse microstructure types, including mesoporous architectures, composite materials, and thin films. These extensions are achieved by integrating new generation algorithms and configurable parameter sets seamlessly into the existing modular framework. The improved system allows for precise geometric control, supports high-resolution voxel-based representations, and facilitates export to external simulation platforms for mechanical analysis. Comprehensive testing confirms the generated structures' stability, scalability, and reproducibility. As a result, MatViz3D emerges as a flexible software tool suitable for computational research, digital material design, and data generation for machine learning workflows in materials engineering.**

Keywords – microstructure modeling, MatViz3D, simulation tools, polycrystalline, hybrid materials.

I. Introduction

Advancements in computational methods, machine learning, and Integrated Computational Materials Engineering (ICME) [1] have led to a growing demand for versatile software tools capable of generating, visualizing, and analyzing three-dimensional models of material microstructures. The need for flexible modeling approaches, interactive visualization, scalable computation, and precise control over microstructure parameters in digital materials science drives these developments.

Recent progress in image-based reconstruction techniques has significantly expanded the range of available methods. Algorithms such as SliceGAN [2] and GAN architectures for porous materials [3] enable efficient reconstruction of complex structures while reducing dependence on costly experimental techniques like FIB-SEM. Combined with deep learning, these approaches support property prediction and inverse design tasks [4]. Machine learning methods have also contributed to multiphysics simulations, weld joint analysis [5], and Bayesian optimization of microstructures [6], although challenges remain regarding interpretability, data quality, and interdisciplinary expertise [1], [7], [8].

Numerous specialized software solutions address particular aspects of microstructure analysis. PuMA focuses on porous materials and thermal properties [9], while μ2mech integrates phase-field modeling with finite elements [10]. The HPR-FE2 approach leverages model reduction techniques to lower computational costs in large-scale simulations [11], [12]. Optimization-based frameworks have demonstrated efficiency in reconstructing microstructures [13], and tools like MCRpy [14] and aflow++ [15] enhance flexibility in descriptor customization and property calculations. Statistically equivalent models of graphite-containing cast iron have been applied for yield surface analysis [16], and OpenCV has proven useful in mapping inclusion distributions in composite microstructures [17]. General-purpose platforms, including MOOSE [18], MetaFEM [19], and FEniCS Mechanics [20], provide scalable environments for advanced simulations but often require specialized expertise. On the visualization and segmentation front, Aphelion [21], MATBOX [22], and DCM [23] offer valuable functionalities with limited computational integration.

On the other hand, MTEX [24] and MIPAR [9] offer solutions for crystallographic texture and image analysis in a user-friendly manner, making them accessible to a broader audience. Recent developments include the Portilla–Simoncelli algorithm for fast microstructure reconstruction using descriptors [25], deep learning for strain field estimation [26], unsupervised ML for real-time analysis [27], and enhanced EBSD techniques supported by AI [28].

Despite these developments, the availability of universal tools that combine flexible modeling, accurate control of microstructure parameters, and interactive visualization remains limited. MatViz3D addresses this gap by providing

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an intuitive interface for generating polycrystalline, composite, porous, and mesoporous structures with 3D visualization support. Parallel cellular automata algorithms enable efficient modeling across a wide range of tasks. Research utilizing MatViz3D includes automated image analysis and grain segmentation by morphometric features [29] and polycrystalline structure modeling with comparisons to experimental data [30].

Understanding microstructure behavior under load remains a critical area of materials research. Studies highlight how heterogeneous microstructures influence mechanical responses. Aluminum alloys produced by laser selective sintering exhibit a stress plateau associated with submicron structures [31], while the phase composition of HSLA steels significantly affects mechanical behavior [32]. In entropystabilized aluminum alloys, local stress distributions are linked to particle damage [33], and multilayer ceramic coatings on titanium composites demonstrate improved crack resistance [34]. In addition, recent probabilistic modeling of containers with discrete extinguishing agents under impact loads confirms the importance of stress analysis for predicting failure probabilities based on material properties and external conditions [35]. These findings emphasize the role of detailed microstructural analysis in predicting mechanical performance and failure probability [36], [37].

With its accessibility, computational efficiency, and integration capabilities, MatViz3D offers a modern, adaptable platform aligned with the evolving needs of scientific and engineering communities in materials research.

II. Problem Statement

In contemporary research on numerical analysis and modeling of complex microstructures, there remains a pressing need for universal tools capable of generating geometric representations of material structures suitable for testing, parameterization, and machine learning. Such tools must not only support the flexible construction of three-dimensional models but also provide precise control over the internal structural parameters.

The MatViz3D software system was initially developed as a specialized environment for generating microstructures with controlled geometric properties. However, its initial functionality was primarily limited to simulating polycrystalline structures based on cellular automaton algorithms. This constrained its applicability to problems involving alternative morphologies or complex hybrid structures. In light of these limitations, the present study is devoted to extending the capabilities of MatViz3D by incorporating support for additional classes of microstructural models, including:

- mesoporous structures (such as aerogels and hydrogels),
- composite materials,
- thin-film architectures.

To accomplish this, the following objectives have been identified:

- the implementation of new algorithms for microstructure generation;
- the integration of extended parameter configuration options within the existing user interface;
- comprehensive testing of the enhanced system across different structural types, including model stability, scalability, and reproducibility evaluation.

The anticipated outcome of this work is the development of a versatile modeling framework that enables researchers to efficiently generate mathematically rigorous representations of diverse microstructures, facilitating their application in computational simulations and advanced material analysis.

III. The Basic Functionality of MatViz3D

A. Algorithms for the generation of polycrystalline microstructures

Polycrystalline materials are composed of numerous individual grains or crystals, each with a distinct crystallographic orientation. The structural arrangement and spatial distribution of these grains play a decisive role in determining the physical and mechanical properties of the material, including strength, hardness, and electrical

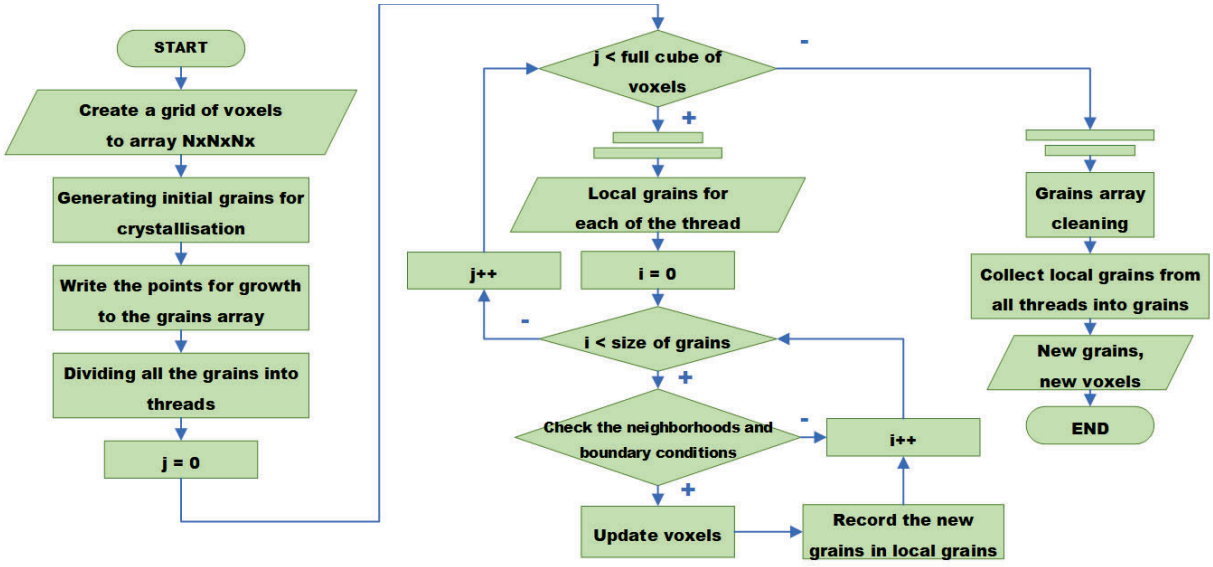


Fig. 1. Parallel grain growth algorithm: key computational steps

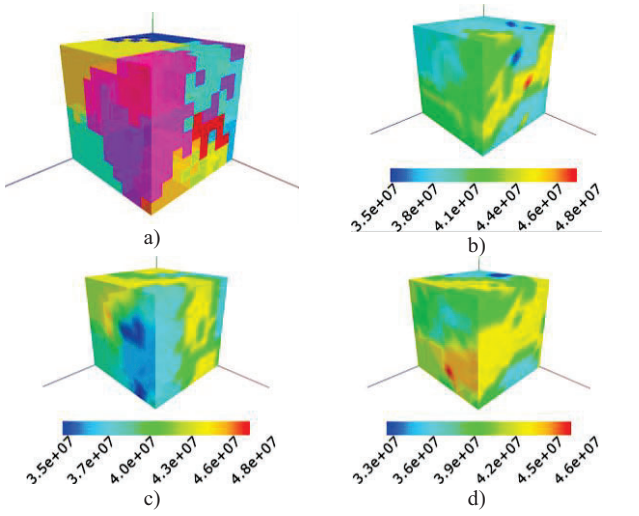


Fig. 2. a) Three-dimensional representation of the generated microstructure. b – d) Distributions of normal stress components: b) σ_x, c) σ_y, d) σ_z visualized within the same spatial domain (Pa)

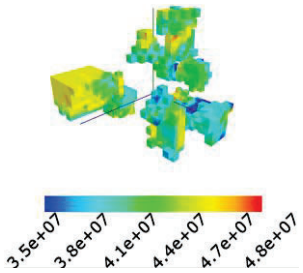


Fig. 3. Grain-resolved stress distribution in an elongated-grain microstructure (Pa)

conductivity. Consequently, the task of modeling polycrystalline microstructures involves reproducing a realistic three-dimensional grain topology that adheres to predefined geometrical and statistical parameters.

In earlier studies, particularly in [29], [30], foundational algorithms for microstructure generation were implemented using the cellular automaton (CA) approach. This method simulates grain growth in three-dimensional space by applying simple, local interaction rules, making it well-suited for capturing realistic morphological evolution. However, the original implementation relied on a sequential computational scheme, which inherently limited scalability and computational efficiency for large-scale structures.

Significant improvements have been introduced to the algorithmic framework as part of the present research. The primary enhancement involves integrating parallel computing techniques using the OpenMP library. This upgrade enables multithreaded processing of the three-dimensional voxel grid, drastically reducing generation time for large microstructures while maintaining structural fidelity and grain morphology.

A schematic representation of the improved parallel algorithm is provided in Fig. 1, outlining the principal computational stages: parameter initialization, grain seed placement, local neighborhood state updating, termination condition checking, and global structure updating. The introduction of parallelization not only enhances performance but also enables the modeling of more complex and statistically representative polycrystalline systems suitable for high-fidelity simulations.

B. Core principles of operation and stress analysis

The MatViz3D software environment was developed as a specialized tool for generating and analyzing geometric models of microstructures, offering a flexible configuration of grain parameters, spatial orientations, and morphological features. Its architecture is built on a modular design principle, enabling seamless integration of structure generation modules, visualization tools, and data interfaces for export/import to external simulation platforms. A comprehensive overview of the software implementation and underlying algorithms is provided in a previous study [29], which established the foundation for generating polycrystalline microstructures using cellular automaton and stochastic approaches.

As part of the present development, functionality for automated stress analysis has been integrated through coupling with the external ANSYS simulation package. Microstructures generated within the MatViz3D environment are exported in a format compatible with ANSYS Workbench, where predefined loading scenarios can be initiated directly through the user interface. Although the mechanical computations are executed outside the core application, the resulting stress tensor data are re-imported into MatViz3D for interactive visualization. This workflow enables a closed-loop modeling process – from microstructure synthesis to mechanical response evaluation – without requiring manual data processing during simulation setup or post-analysis.

An example of a polycrystalline structure with the superimposed stress components in the x, y, and z directions is illustrated in Fig. 2. The regions exhibiting elevated stress align with grain boundary configurations, confirming the validity of the mechanical-structural correlation. A different representation is provided in Fig. 3, where a microstructure with elongated grains is shown. In this case, the stress distribution is visualized in such a way that the internal state of each individual grain can be distinctly observed, demonstrating the system's ability to resolve anisotropic microstructures – particularly those typical of metals after plastic deformation or rolling.

Through the integration of these new features, MatViz3D evolves into a comprehensive modeling platform, not only facilitating the creation of microstructural models but also enabling preliminary engineering analysis of their mechanical properties. This extended functionality represents a critical advancement for supporting computational materials science and virtual experimentation tasks.

IV. Enhancement of MatViz3D functionality for modeling diverse microstructure types

A. Mesoporous structures

One of the key directions in the functional extension of the MatViz3D environment is the modeling of mesoporous materials, which are distinguished by the presence of a highly branched porous phase with pore sizes typically ranging from 2 to 50 nanometers. Such structures are critical for a variety of applications, including catalysis, filtration, molecular storage, and energy systems.

To simulate mesoporous architectures, MatViz3D incorporates an approach based on diffusion-limited aggregation (DLA), a well-established physical mechanism for modeling particle clustering processes.

This method mimics the formation of porous networks exhibiting high fractal complexity, analogous to the growth of clusters in colloidal suspensions, sol-gel systems, or vapor-phase condensates. The motion of individual particles or clusters is modeled as Brownian motion, with the diffusion coefficient of each particle determined according to the Stokes-Einstein relation:

$$D_i = \frac{k_B T}{6\pi\eta r_i} \quad (1)$$

where, r_i is the radius of the cluster, k_B is the Boltzmann constant, T is the temperature and η is the dynamic viscosity of the medium.

At each discrete time step, a particle undergoes a random displacement governed by:

$$x_i(t + \Delta t) = x_i(t) + \sqrt{2D_i\Delta t} \cdot R \quad (2)$$

where R is a randomly oriented unit vector, ensuring isotropic diffusion. Irreversible aggregation occurs when a particle comes within a threshold distance ε of an existing cluster:

$$\left\|x_i - x_j\right\| \leq \varepsilon \quad (3)$$

At this point, the particle adheres to the structure, creating a new growth site. The porous framework is thus formed through the repeated attachment of diffusing particles, giving rise to a hierarchical and spatially complex geometry.

A block diagram of the mesoporous structure generation algorithm in MatViz3D is shown in Fig. 5. The process begins with defining the spatial voxel domain and initializing a seed particle. Subsequent particles are released from random positions and undergo Brownian motion until they aggregate with the existing structure. Upon attachment, the particle becomes part of the growing cluster, and the cycle continues until a target pore fraction or cluster size is reached. The method offers flexible control over the resulting topology by adjusting initial nucleation parameters (e.g., concentration and number of seeds), dynamic variables (e.g., temperature and viscosity), and aggregation rules (e.g., interaction range and sticking conditions).

The generated mesoporous phase can be exported for further computational analysis. Fig. 4 illustrates a representative example of a three-dimensional mesoporous microstructure generated using this algorithm.

B. Composite materials

At the current development stage, MatViz3D supports the generation of a single composite type – unidirectional composites with cylindrical inclusions. Users can select constituent materials for the matrix and inclusions from the internal materials database, after which the system automatically generates a three-dimensional geometric model. This model fully integrates with the stress state assessment module, enabling seamless transfer to the ANSYS environment via a predefined loading scenario. Upon completion of the simulation, the computed results – particularly stress distributions in three spatial directions are imported back into MatViz3D for visualization and further analysis.

A schematic representation of the algorithm used to construct cylindrical inclusions within a voxelized domain is provided in Fig. 6. The generation process begins by initializing a three-dimensional voxel array of size N × N × N, initially filled entirely with matrix material. Based on either the specified number of inclusions or the overall volume fraction, the radius of each cylindrical inclusion is computed. The algorithm determines the center coordinates of the inclusion in the xy-plane and checks for each (x, y) position whether the following condition is satisfied:

$$(x - x_{center})^2 + (y - y_{center})^2 \leq r^2 \quad (4)$$

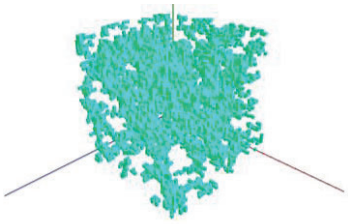


Fig. 4. Example of a 3D mesoporous microstructure generated in MatViz3D

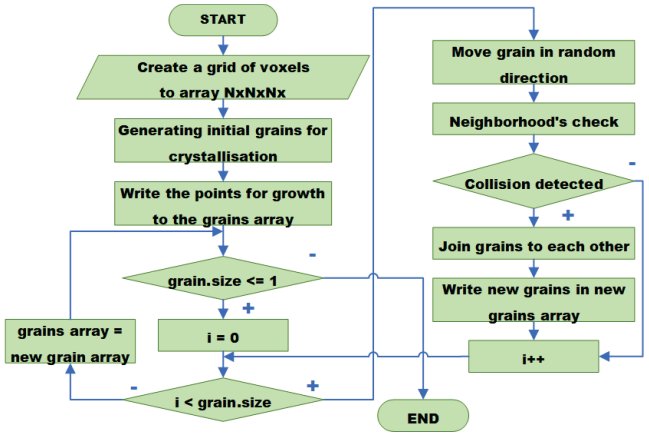


Fig. 5. Block diagram of the mesoporous structure generation algorithm

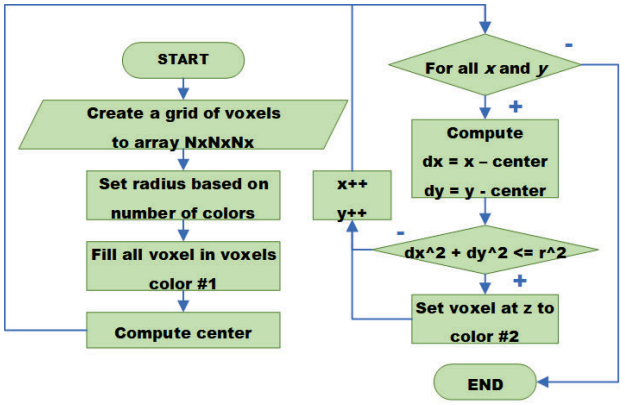


Fig. 6. Block diagram of the cylindrical inclusion generation algorithm

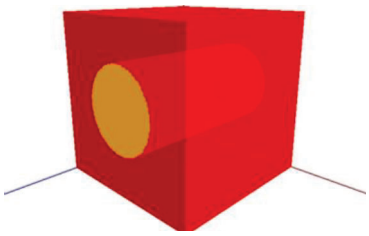


Fig. 7. Example of composite material modeling

If this condition holds, all voxels along the z-axis at that position are marked as inclusions, forming a continuous cylindrical shape. This method ensures that the inclusion geometry remains uniform across all cross-sections along the z-direction, thus accurately representing a classical unidirectional composite architecture. An example of modeling composite material in the environment is shown in Fig. 7.

The current implementation provides a robust foundation for further expansion. Future versions of MatViz3D will incorporate support for randomly distributed, spherical, ellipsoidal, and clustered inclusions, along with an extended library of composite morphological configurations to support more complex material systems.

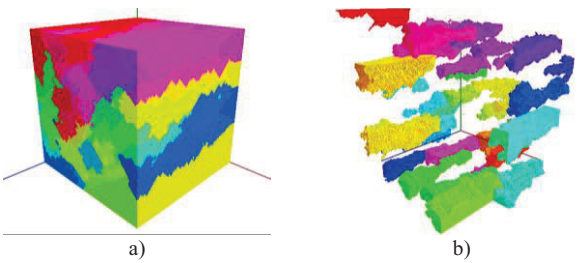
C. Thin-film architectures

In contrast to bulk polycrystalline materials – where grain nucleation occurs randomly throughout the volume – thin-film structures are characterized by a distinct initial condition: grain formation is initiated at a specific model surface. This setup reflects the physical nature of film deposition or crystallization processes on a substrate. It enables the morphology simulation with pronounced growth directionality, typically aligned with the surface normal.

In the implemented algorithm, the generation of a thin-film microstructure begins by initializing nuclei along a predefined boundary of the simulation domain (e.g., the plane $Z = 0$). These nuclei then grow into the volume, expanding their regions of influence based on local interaction rules, similar to those used in classical cellular automata. As a result, an anisotropic grain structure emerges, exhibiting elongation in the direction of growth, which is consistent with the physical characteristics of deposited films.

This approach not only facilitates simulation of the key stages of thin-film formation but also allows for variation in initial parameters – such as nucleus density, spatial distribution, and orientation. Such flexibility makes it possible to study the influence of technological factors on the resulting microstructure morphology. Moreover, the implementation ensures the structural continuity of the film, allows for precise control over boundary conditions, and supports the formation of physically realistic grain boundary networks.

An example of a simulated thin-film microstructure is provided in Fig. 8, illustrating the typical vertical elongation of grains from the nucleation surface toward the opposite side of the model domain – characteristic of structures formed through deposition or directional crystallization processes.



a) b)

Fig. 8. Thin-film microstructure a), and individual grain visualization b)

CONCLUSIONS

As a result of this study, the MatViz3D software environment was substantially enhanced to support the generation of microstructures beyond conventional polycrystalline models. The newly developed algorithmic modules enable the creation of geometrically consistent representations of composites, aerogels, and thin-film systems, characterized by complex morphologies and a high degree of parametric control.

The proposed extension of MatViz3D is closely connected to physical materials. Its algorithms reproduce real microstructure formation, such as diffusion-limited aggregation in porous media, fiber alignment in composites, and anisotropic grain growth in thin films. Unlike many existing tools, the updated version offers a unified environment for generating polycrystalline, porous, composite, and thin-film structures. It also supports their further engineering analysis. Benchmarking against experimental data shows a close match in geometric and morphological characteristics. The deviations remain within acceptable limits, confirming the reliability of MatViz3D as a tool for digital materials design.

The extension of the configuration framework and the seamless integration of additional settings into the software architecture ensured the system's coherence, scalability, and alignment with current modeling and simulation requirements. Furthermore, the implementation of stress data import and its subsequent three-dimensional visualization significantly broadened the functional scope of the platform. This integration allows for a continuous workflow within a unified environment, ranging from microstructure generation to analyzing their physical and mechanical behavior.

In summary, the updated MatViz3D system demonstrates strong potential as a versatile platform for the geometric modeling of material structures, particularly in research contexts that demand precise internal parameter control and flexible model selection tailored to the objectives of computational analysis.

ACKNOWLEDGMENT

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Improving Energy Efficiency and Environmental Cleanliness of a Mobile Power Plant by Modernizing the Fuel Supply Control System of a Diesel Generator

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Abstract — **The paper deals with the problems of building fuel supply control systems for mobile electric stations based on internal combustion engines, which are systems for producing electrical energy with high technical, economic, and environmental indicators, i.e. with improved energy efficiency. Optimization of fuel supply phase control and assessment of its uncertainty in the engine of the power plant is carried out taking into account changes in the speed and load mode of operation during the operation of the plant with gas turbine supercharging. When forming the fuel supply phase, the information on the piston entry moments not only at the top but also at the bottom dead center is used.**

Keywords — power plant, diesel generator, fuel consumption, exhaust emission, control, optimization, fuel phase, fuel injection, top dead center, bottom dead center.

I. Introduction

In [1], a mobile power plant based on a diesel generator was considered, in which a number of control actions, including the fuel injection advance angle θ (fuel supply phase), were regulated to ensure high technical, economic, and environmental indicators. Using the methodology of [2, 3] and developing it, it is possible to obtain a quasi-optimal dependence of the fuel supply phase on the mode parameters of the diesel generator. The dependence of this phase on angular velocity ω and load moment M_L is practically linear (where there is no exceeding of the maximum combustion pressure, the rate of its change by the crankshaft rotation angle and other parameters). It is of interest not only to analyze the methodological uncertainty of realization of the mentioned dependencies in real conditions of diesel generator operation, but also to develop measures for its reduction, as well as the synthesis of the block diagram for the technical realization of such a system, which is the subject of this paper.

To form the fuel supply phase, it is necessary to take information from the diesel generator about the moments of diesel piston entering the top dead center (TDC) and bottom dead center (BDC), load moment M_L and speed mode of the diesel generator. Note that usually for these purposes are used only the marks of TDC. The authors propose to use both TDC and BDC marks to form the fuel supply phase.

It is known that the fuel supply phase θ (fuel injection advance angle) significantly affects the technical, economic and environmental indicators of the engine, which is mentioned in [4–9]. This fact confirms the relevance of the work under consideration.

II. Basic Theoretical Provisions For Building Models Of Fuel Supply Phase Formation

Let's synthesize mathematical models of the processes of conversion of angular velocity of the shaft, load torque and other mode parameters in the fuel supply phase, considering the diesel generator serviceable and not taking into account the deviation of the rotation velocity ξ due to its small value compared to the average value of the angular speed of the crankshaft and equality to zero of the average value of the angular velocity deviation per revolution of the shaft. Let us consider the mathematical model of linear transformation proposed by the authors.

When the diesel generator is running, the fuel injection pulse duration is measured:

$$t_i = C_P \cdot M_L \tag{1}$$

where C_P – coefficient of proportionality between the duration of the fuel supply pulse and the diesel engine load moment M_L.

The fuel injection pulse duration t_i is converted by means of frequency pulses $f' >> \omega/2\pi$, where ω – angular velocity of diesel generator shaft, in digital code:

$$N_3 = f' \cdot t_i = f' \cdot C_P \cdot M_L \tag{2}$$

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At the moment the piston reaches top TDC, a impulse of duration t_{im} is generated. The time interval t_{ti} from the end of this impulse to the moment the piston arrives at BDC is converted by means of high-frequency impulses with frequency $f >> \omega/2\pi$ into some number N_1, which can be expressed as follows

$$N_1 = f \cdot t_{ti} = f \cdot \left(\frac{\pi}{\omega} - t_{im} \right). \quad (3)$$

From the moment the piston enters the BDC, the time interval t_{ti} is converted into the second number N_2 by means of pulses of the same frequency $f >> \omega/2\pi$, and

$$N_2 = f \cdot t_{ti} \,. \cdot \quad (4)$$

The numbers N_2 and N_3 are summed up and at the moment of equality $N_2 + N_3 = N_1$, that is, at observance of the relation

$$f' \cdot C_P \cdot M_L + f \cdot t_{ti} = f \cdot \left(\frac{\pi}{\omega} - t_{im} \right) \quad (5)$$

the fuel supply start signal is formed. The impulse repetition rate f is obtained by converting the repetition rate of TDC and BDC impulses into voltage $U = C_\omega \cdot \omega/\pi$, where C_ω – the coefficient of conversion of frequency into voltage, and voltage into frequency

$$f = C_{U1} \cdot U = C_{U1} \cdot C_\omega \cdot \frac{\omega}{\pi}, \quad (6)$$

where C_{U1} – the first voltage-to-frequency conversion coefficient.

The impulse frequency f' is also obtained by converting a voltage U of constant sign into an impulse signal

$$f' = C_{U2} \cdot U = C_{U2} \cdot C_\omega \cdot \frac{\omega}{\pi}, \quad (7)$$

where C_{U2} – the second voltage-to-frequency conversion coefficient.

Substituting (6) and (7) into expression (5), we obtain

$$t_{ti} = \frac{\pi}{\omega} - t_{im} - \frac{C_P \cdot M_L \cdot C_{U2}}{C_{U1}}.$$

Therefore, the advance time of fuel injection is equal to

$$t_{ai} = \frac{\pi}{\omega} - t_{ti} = t_{im} + \frac{C_P \cdot M_L \cdot C_{U2}}{C_{U1}},$$

and the fuel supply phase is defined as

$$\theta = \left(t_{im} + \frac{C_P \cdot M_L \cdot C_{U2}}{C_{U1}} \right) \cdot \omega. \quad (8)$$

The expression (8) is a mathematical model of the processes of linear transformation of angular velocity and load torque of the diesel engine into the fuel supply phase, since the fuel supply phase depends linearly on the mode parameters of the diesel generator (the influence of weather conditions, altitude, chemical composition of fuel, temperature conditions of the engine, generator, etc. on the operation of the system is not taken into account here). The diagram of fuel advance angle formation is presented in Fig. 1.

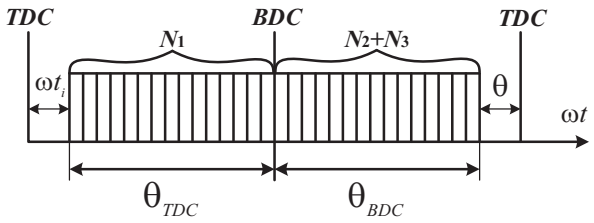


Fig. 1. Diagrams of fuel supply phase formation

Let us consider the mathematical models of the processes of conversion of angular velocity and load into the fuel supply phase in the transient speed modes of the unit under the previous assumptions, namely: acceleration (deceleration) of the engine crankshaft within one revolution constantly appears at the moment of the piston arrival at TDC and takes place within at least one revolution of the shaft. Taking into account expressio (3), the number N_1 can be written as follows

$$N_1 = f \cdot \left(\frac{\sqrt{\omega_0^2 + 2 \cdot \varepsilon \cdot \pi} - \omega_0}{\varepsilon} - t_{im} \right). \quad (9)$$

where ω_0 – is the angular velocity of the shaft at the moment when the diesel generator piston enters the TDC in the considered shaft revolution; ε – angular acceleration of the diesel generator shaft.

The second number N_2 is determined by (4). The ratio symbolizing the beginning of fuel injection, taking into account (9), can be represented as follows

$$f' \cdot C_P \cdot M_L + f \cdot t_{ti} = f \cdot \left(\frac{\sqrt{\omega_0^2 + 2 \cdot \varepsilon \cdot \pi} - \omega_0}{\varepsilon} - t_{im} \right). \quad (10)$$

Since the rotation of the diesel engine shaft is considered to be equally accelerated, the average angular velocity of the shaft as the piston moves from TDC to BDC is equal to

$$\omega_{m1} = \frac{\omega_0 + \sqrt{\omega_0^2 + 2 \cdot \varepsilon \cdot \pi}}{2}, \quad (11)$$

and respectively

$$f = \frac{C_{U1} \cdot C_\omega}{2\pi} \cdot \left(\omega_0 + \sqrt{\omega_0^2 + 2 \cdot \varepsilon \cdot \pi} \right). \quad (12)$$

By analogy with (11), the average angular velocity of the shaft as the piston moves from the BDC to the TDC can be represented as

$$\omega_{m2} = \frac{\sqrt{\omega_0^2 + 2 \cdot \varepsilon \cdot \pi} + \sqrt{\omega_0^2 + 4 \cdot \varepsilon \cdot \pi}}{2},$$

as a result of which

$$N_2 = \frac{C_{U1} \cdot C_\omega}{2\pi} \cdot \left(\sqrt{\omega_0^2 + 2 \cdot \varepsilon \cdot \pi} + \sqrt{\omega_0^2 + 4 \cdot \varepsilon \cdot \pi} \right). \quad (13)$$

Since fuel injection happens near the point at which shaft acceleration occurs, we can consider

$$f' = C_{U2} \cdot C_{\omega} \cdot \frac{\omega_0}{\pi}. \quad (14)$$

Taking into account (12), (13) and (14), expression (5) is transformed to the view

$$\frac{C_P \cdot M_L \cdot C_{U2} \cdot \omega_0}{\pi} + \frac{C_{U1} \cdot t_{ti}}{2\pi} \cdot \left(\sqrt{\omega_0^2 + 2 \cdot \varepsilon \cdot \pi} + \sqrt{\omega_0^2 + 4 \cdot \varepsilon \cdot \pi}\right) =$$

$$= \frac{C_{U1}}{2\pi} \cdot \left(\omega_0 + \sqrt{\omega_0^2 + 2 \cdot \varepsilon \cdot \pi}\right) \cdot \left(\frac{\sqrt{\omega_0^2 + 2 \cdot \varepsilon \cdot \pi} - \omega_0}{\varepsilon} - t_{im}\right). \quad (15)$$

Using the notation $\omega_{BDC} = \sqrt{\omega_0^2 + 2 \cdot \varepsilon \cdot \pi}$ and $\omega_{TDC} = \sqrt{\omega_0^2 + 4 \cdot \varepsilon \cdot \pi}$ we write from (13)

$$t_{ti} = \frac{\left(\omega_0 + \omega_{BDC}\right) \cdot \left(\frac{\omega_{BDC} - \omega_0}{\varepsilon} - t_{im}\right) - \frac{2C_P \cdot \omega_0 \cdot M_L \cdot C_{U2}}{C_{U1}}}{\omega_{BDC} + \omega_{TDC}}. \quad (16)$$

During the time t_{ti} the crankshaft of the diesel engine rotates by an angle $\theta_{BDC} = \omega_{BDC} \cdot t_{ti} + \frac{\varepsilon \cdot t_{ti}^2}{2}$ or taking into account (16)

$$\theta_{BDC} =$$

$$= \left[\omega_{BDC} + \frac{\varepsilon}{2}\left[\frac{\left(\omega_0 + \omega_{BDC}\right) \cdot \left(\frac{\omega_{BDC} - \omega_0}{\varepsilon} - t_{im}\right) - \frac{2C_P \cdot \omega_0 \cdot M_L \cdot C_{U2}}{C_{U1}}}{\omega_{BDC} + \omega_{TDC}}\right]\right] \times$$

$$\times \left[\frac{\left(\omega_0 + \omega_{BDC}\right) \cdot \left(\frac{\omega_{BDC} - \omega_0}{\varepsilon} - t_{im}\right) - \frac{2C_P \cdot \omega_0 \cdot M_L \cdot C_{U2}}{C_{U1}}}{\omega_{BDC} + \omega_{TDC}}\right]. \quad (17)$$

Then the fuel supply phase (fuel injection advance angle) taking into account (17) can be written as follows

$$\theta = \pi - \theta_{BDC} = \pi -$$

$$- \left[\left[\omega_{BDC} + \frac{\varepsilon}{2}\left[\frac{\left(\omega_0 + \omega_{BDC}\right) \cdot \left(\frac{\omega_{BDC} - \omega_0}{\varepsilon} - t_{im}\right) - \frac{2C_P \cdot \omega_0 \cdot M_L \cdot C_{U2}}{C_{U1}}}{\omega_{BDC} + \omega_{TDC}}\right]\right] \times\right.$$

$$\left. \times \left[\frac{\left(\omega_0 + \omega_{BDC}\right) \cdot \left(\frac{\omega_{BDC} - \omega_0}{\varepsilon} - t_{im}\right) - \frac{2C_P \cdot \omega_0 \cdot M_L \cdot C_{U2}}{C_{U1}}}{\omega_{BDC} + \omega_{TDC}}\right]\right]. \quad (18)$$

The obtained model reflects the dependence of the fuel supply phase on the angular velocity of the diesel engine shaft, angular acceleration and load moment in transient modes of diesel generator operation.

III. Analysis Of Methodological Uncertainty Of Fuel Supply Phase Formation

When analyzing the uncertainty, we will assume the same assumptions as before. Remember that in dynamics the uncertainty of the fuel supply phase formation is caused by the fact that the angular velocity of the shaft at the beginning and at the end of a given revolution is different if we assume that the diesel torque changes abruptly just before the impulse appears t_i. In this case, we can state the following: since the torque has already changed abruptly to some new constant value, we can consider that $M_L = \text{const}$, moreover, since fuel injection occurs near TDC, we should write by analogy with (8)

$$\theta = \omega_{TDC}\left(t_{im} + \frac{C_P \cdot M_L \cdot C_{U2}}{C_{U1}}\right), \quad (19)$$

Then, taking into account (18), the relative methodological uncertainty of fuel supply phase formation in transient modes of diesel generator operation can be represented by the following expression

$$\delta = \frac{\pi - S \cdot \left(\omega_{BDC} + \frac{\varepsilon}{2} \cdot S\right)}{\omega_{TDC}\left(t_{im} + \frac{C_P \cdot M_L \cdot C_{U2}}{C_{U1}}\right)} - 1, \quad (20)$$

where $S = \left(\omega_0 + \omega_{BDC}\right) \cdot \left(\frac{\omega_{BDC} - \omega_0}{\varepsilon} - t_{im}\right) - \frac{2 \cdot C_P \cdot \omega_0 \cdot M_L \cdot C_{U2}}{C_{U1}}$,

$\omega_{BDC} = \sqrt{\omega_0^2 + 2 \cdot \varepsilon \cdot \pi}$ and $\omega_{TDC} = \sqrt{\omega_0^2 + 4 \cdot \varepsilon \cdot \pi}$.

Consequently, the relative methodological uncertainty is a function of many parameters $\delta = f\left(\omega_0, \varepsilon, t_{im}, C_P, M_L, C_{U1}, C_{U2}\right)$. As an example for a particular mode, its plots are shown in Fig. 2. 3.

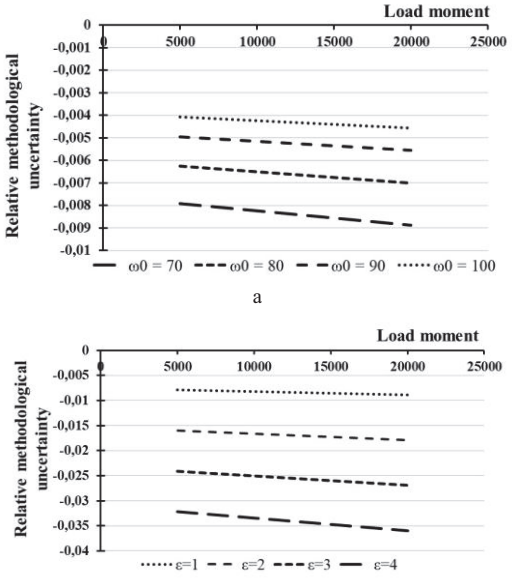


Fig. 2. Dependence of relative methodological uncertainty of linear transformation of mode parameters in the fuel supply phase as a function of the load moment of the standard system; a – at different initial angular velocities of the diesel generator crankshaft; b – at different angular accelerations of the diesel generator crankshaft.

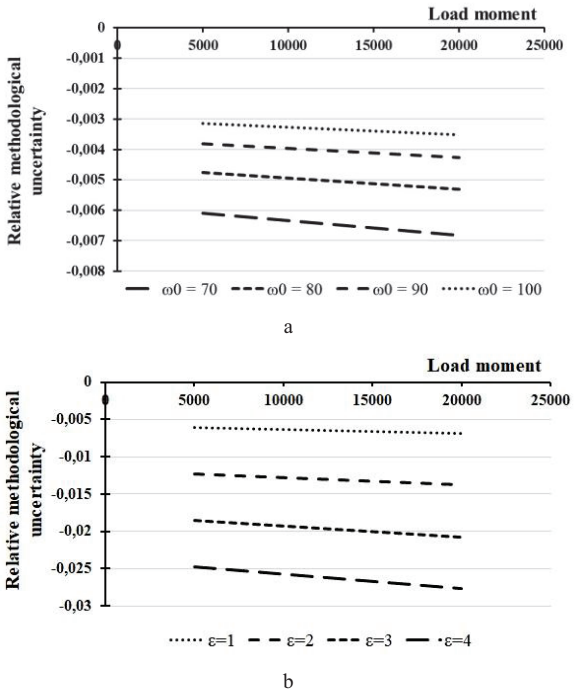


Fig. 3. Dependence of relative methodical uncertainty of linear transformation of mode parameters into fuel supply phase as a function of load moment of the modernized system; a – at different initial angular velocities of the diesel generator crankshaft; b – at different angular accelerations of the diesel generator crankshaft.

The curves are plotted for several fixed values of angular acceleration of the shaft at smooth variation of the initial angular velocity of the shaft and constancy of all other variables in the case of linear transformation of angular velocity and load torque in the fuel supply phase.

The graphs show that the modernized fuel supply system proposed in this work provides an average reduction in the uncertainty of fuel supply phase formation by a factor of 1.2 to 1.3 compared to the system [1].

IV. Technical Realization Of The System

Block diagram of the modernized fuel supply control system is shown in Fig. 4. and its blocks are also decoded there. Let us briefly describe the operation of the system taking into account that the corresponding signal conversion models were considered earlier.

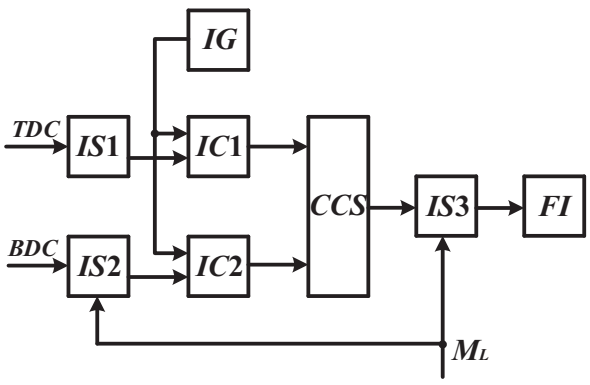


Fig. 4. Block diagram of the modernized fuel supply control system.

The signal of the TDC mark is received by the formulator F1, which produces a rectangular impulse, which puts the impulse counter IC1 in the mode of recording information coming to one of its inputs from the output of the impulse generator IG. Some number is recorded in the counter IC1 up to the arrival of the piston diesel generator in the BDC. Impulse mark BDC comes to the input of the shaper IS2, which produces a rectangular impulse, the duration of which impulse is proportional to the load moment M_L. This impulse puts IC2 in the mode of recording information and its other input receives high-frequency impulses IG. Thus, the recording of numbers in IC2 occurs during the movement of the piston of the diesel generator from BDC to TDC. When the numbers in IC1 and IC2 coincide, the CCS circuit produces a signal that goes to the input of IS3. The latter produces an impulse to control the injector FI. The duration of this impulse corresponds to the load moment M_L, and its leading edge is determined by the fuel supply phase.

V. Experimental Studies Of The Designed System And Processing Of The Results

The authors conducted experimental studies of the modernized fuel supply system on a real diesel generator with a rated power of 2200 kilowatts and a speed of 1000 revolutions per minute. At the same time the following were measured: effective power *Ne*, relative specific emissions of nitrogen oxides qNO_X, relative specific emissions of soot *C*, specific effective fuel consumption *Ge* and controlled the speed mode of the unit. In addition, a number of operating process parameters were monitored and maintained within the permissible range of variation according to the standard diesel generator test protocol. The measurement results were processed by the least squares method [10,11].

Fig. 5 shows the dependence of specific and effective fuel consumption on the effective power of the diesel generator, which shows a significant increase in fuel efficiency, that is, energy efficiency of diesel generator with a modernized system compared to the standard one.

Figs. 6 and 7 show comparative dependences of nitrogen oxide qNO_X, and soot emissions *C* as a function of specific effective fuel consumption *Ge*, which shows that the fuel supply control system modernized by the authors provides a

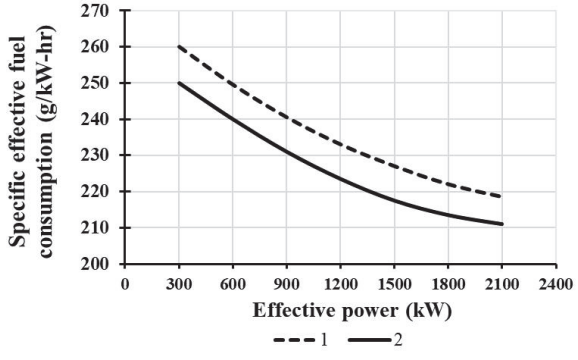


Fig. 5. Dependence of specific effective fuel consumption on the effective power of the diesel generator, 1 – standard fuel supply system; 2 – modernized fuel supply system.

reduction in harmful emissions and thus increases the environmental cleanliness of the diesel power plant.

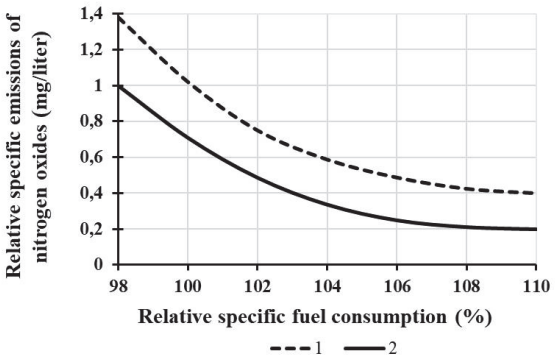


Fig. 6. Dependence of relative specific emissions of nitrogen oxides on relative specific fuel consumption at change of fuel supply phase, 1 – standard fuel supply system; 2 – modernized fuel supply system.

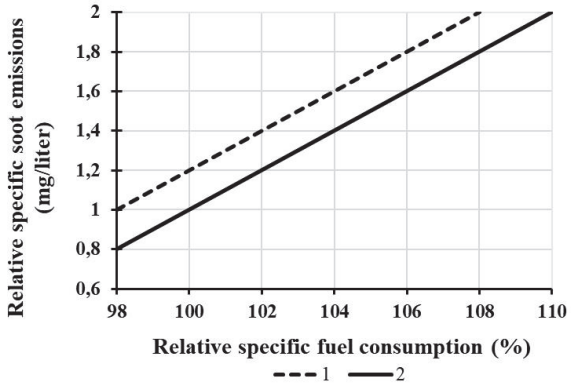


Fig. 7. Dependence of relative specific soot emissions on relative specific fuel consumption at change of fuel supply phase; 1 – standard fuel supply system; 2 – modernized fuel supply system.

VI. Conclusions

On the basis of the theory of optimal control, the theory of structured optimization, the theory of information-measuring systems, the theory of probability and mathematical statistics, as well as the theory of multivariate experiment, the following scientific and practical results are obtained:

1. A mathematical model of fuel supply phase formation has been synthesized, the methodological uncertainty of this phase has been investigated, and a block diagram of a quasi-optimal fuel supply control system for a power plant engine has been constructed, ensuring high technical, economic, and environmental indicators of the power plant, i.e., improved energy efficiency and environmental cleanliness, as well as fuel savings per kilowatt-hour of electricity produced in the system compared to serial installations.

2. The results of experimental studies of the power plant operation in real conditions have confirmed the correctness of theoretical provisions underlying its synthesis, and the attached as an example dependences of specific effective consumption of fuel on the effective power, dependences of nitrogen oxide emissions and soot emissions on fuel consumption testify to the increased energy efficiency and environmental cleanliness of the power plant with the developed fuel supply control system in comparison with the standard serial model.

3. Scientific novelty of the work consists in the fact that the authors propose to take into account the angular acceleration of the crankshaft of the diesel generator when forming the fuel supply phase and to use in the process of time-impulse and impulse-time transformation of information not only marks TDC of the diesel generator piston (and along with this in some cases to use angular marks of the crankshaft position), as it was in the standard serial fuel supply control systems, but also marks of the BDC. This made it possible to reduce the methodological uncertainty of fuel supply phase formation and, thus, to increase the energy efficiency and environmental cleanliness of the diesel power plant.

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Influence of Nonstationary Electromagnetic Processes on Electromagnetic Compatibility and Technical Condition of Special-Purpose Transformer Equipment

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Abstract – Studies have been conducted on the influence of nonstationary electromagnetic processes on additional losses, electromagnetic compatibility, and the current technical condition of power transformers under rapidly varying load conditions. It has been shown that nonstationary electromagnetic phenomena lead to increased additional losses and a reduction in equipment service life. Engineering methods have been proposed for assessing surface losses in bulk ferromagnetic materials by approximating the nonlinear magnetic permeability of steels under simultaneous magnetization by a periodic nonsinusoidal magnetic field and a constant component.

Keywords – electromagnetic compatibility (EMC), power transformer, rapidly varying loads, technical condition, additional losses, approximation, nonlinearity, magnetic permeability, structural steel, simultaneous magnetization, nonsinusoidality, magnetic field

Power transformers with oil cooling are the main type of electrical equipment in power supply systems. They provide the interconnection of individual sections of power supply systems with different voltage classes and industrial enterprises.

Power transformers are the most numerous type of power equipment in power supply systems. Thus, there are more than 12–15 transformers per unit of generated power. At the same time, in the power supply systems of energy-intensive industrial enterprises, the indicators of electric power quality significantly exceed the requirements of state standards and regulatory and technical documentation for general-purpose transformers. Therefore, for such power supply systems, it is advisable to develop special-purpose transformers or reconstruct general-purpose ones. This is explained by the fact that the operating modes of such transformers are sharply variable and probabilistic in nature.

In power transformers, such operating modes lead to the formation of nonstationary electromagnetic processes. The latter cause violations of electromagnetic compatibility and a reduction in transformer service life.

The most unfavorable operating modes of power transformers occur in DC systems and links, as well as in substations for converting alternating current of industrial frequency 50 Hz into 60 Hz.

Thus, for the optimal functioning of power supply systems as a whole, it is necessary to take into account not only the operating conditions of general-purpose electrical equipment but also those of rapidly varying loads.

In addition, it should be noted that, in practice, in the higher harmonic circuits of reactive power compensators of electric arc furnaces, ferroresonance processes often occur. They are formed as a result of parameter deviations of reactors and capacitor banks in the corresponding electrical circuits. These processes are excited by rapidly varying operating modes and significantly aggravate the operating conditions of power and furnace transformers.

During the design and operational conditions of energy-intensive electrical equipment in industrial power supply systems, special attention is paid to issues of electromagnetic compatibility and service life depending on operating conditions and load modes. These issues become particularly relevant for power supply systems of electric arc steelmaking furnaces, rolling mill drives, direct current lines and inserts, and others [1-21]. In such cases, achieving optimal operating conditions for power transformers, electro-technological installations, and other electrical equipment is practically impossible. This is explained by the presence of nonlinearities and the rapidly varying nature of loads, which disrupt the ability of technical systems to operate under normative parameters during exploitation. At the same time, the equipment itself becomes a partial source of interference and violation of electromagnetic compatibility with adjacent electrical equipment [1-15].

It should be noted that international technical organizations uphold rather strict requirements of the international standard IEC 61000-3-2:2014 regarding the limitation of harmonic current components in groups of electrical equipment, which are practically impossible to achieve using domestic technologies. This standard was developed by highly qualified specialists of the Subcommittee “Low-voltage Electromagnetic Phenomena” of Technical Committee 77 of the IEC “Electromagnetic Compatibility.” At the same time, it is noted that it is

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necessary to establish the maximum possible level of electromagnetic processes that may affect associated technical systems under specific operating conditions

Despite rather strict theoretical requirements for electromagnetic compatibility, in practice there are significant technical and economic losses [5-15, 10-25]. As an example, we present statistical data on losses in economically developed countries. In the United States (as of 2021) – $13.3 billion; the European Union – $15 to $20 billion; France – $25 billion; Canada – $5 billion. In the domestic power system, statistical data are unavailable due to the rather complex situation. In the United States, expenditures aimed at improving power quality indicators are estimated at $3.4 billion per year; in the European Union – no less than 5% of the estimated losses.

The present study is devoted to the influence of nonstationary electromagnetic processes on electromagnetic compatibility, specifically in terms of increasing overall additional losses in non-active structural elements, which act as a source of compatibility violations and reduce the operational reliability of power transformers.

During the design of power transformers, energy-intensive electromechanical systems, electrical machines and electric drives, as well as electrotechnological complexes, first of all, electromagnetic processes are studied in order to satisfy the parameters of technical requirements and technical specifications [1-17]. One of the defining issues is the calculation of local additional losses in structural details under different operating modes and conditions of operation, since their excess leads to a reduction in the service life of the equipment.

Inactive structural elements and power units are made of structural steels and, during equipment operation, are exposed to stray electromagnetic fields. Within electrical equipment, stray electromagnetic fields are distributed quite unevenly, varying in intensity and shape depending on operating modes and conditions. The stray electromagnetic field affecting inactive components induces additional losses, leading to local overheating, reduced dielectric properties of insulation, loss of electrodynamic stability, formation of undesirable cumulative effects, and emergency failures. Such studies are particularly relevant in the design and development of power transformers for supplying arc steelmaking furnaces, DC inserts, rolling mill drives, and other applications [26-35].

According to the requirements of state standards and regulatory technical documentation, each piece of electrical equipment corresponds to specific nominal parameters and possible levels of their exceedance. Exceeding these parameters and operating conditions leads to an increase in additional losses and the subsequent development of a range of undesirable consequences and emergency failures [4-17]. The intensity of the electromagnetic field, the geometric dimensions and design of inactive components, and their electromagnetic characteristics have a significant impact on additional losses. The electromagnetic field is excited by corresponding currents and has the same nature and shape [9]. These currents may have a rapidly varying nature, contain higher harmonics, and include a constant component [9]. As an example, Figure 1 presents oscillograms of phase currents of a high-power network transformer supplying arc steelmaking furnaces.

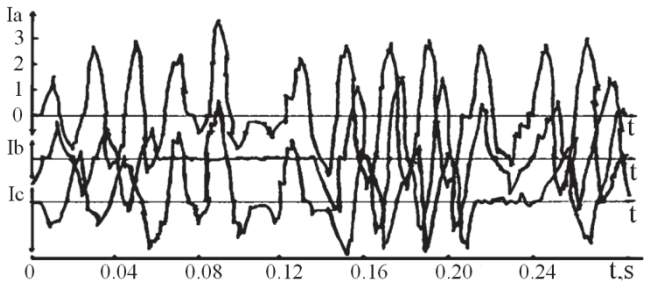


Fig. 1. Phase currents of a 160 MVA network transformer supplying an arc steelmaking furnace of type DSP-50 in the "melt" technological mode.

As seen in Fig. 1, at the initial stage of the oscillogram, a two-phase short technological short circuit occurs. The phase currents contain higher harmonics and a DC component. In the next stage, a three-phase technological short circuit occurs, and the currents contain higher harmonics. Subsequently, alternating technological short circuits are observed. Under such operating conditions, additional losses in the structural elements of this transformer increase compared to steady-state operation. For example, additional losses in the transformer tank (relative to the nominal mode) increase up to 23.0%, in the pressing plates of the magnetic system up to 60-73%, in the first package up to 23%, and in the lower consoles up to 52%. Multiple emergency inspections have shown that the primary causes are local overheating, loss of electrodynamic stability, insulation charring, winding deformations, and other issues. It is evident that rapidly varying operating modes of equipment:

- have a significant impact on electromagnetic compatibility, reliability, and service life of electrical equipment;

- require, during design and operation, consideration of the influence of the nature of the stray electromagnetic field on additional losses and overheating of structural components, as well as the cumulative effect of cause-and-effect factors leading to reduced technical reliability of electrical equipment;

- necessitate, in the calculation of additional losses, accounting for the influence of the stray electromagnetic field on the ferromagnetic properties and magnetic permeability of structural steels;

- demand research and development of engineering methods for considering the characteristic features of incident stray electromagnetic fields in structural steels used in electrical machine construction, which allows for significant refinement of existing methods for ensuring electromagnetic compatibility of energy-intensive electrical equipment and becomes particularly relevant.

The relevance of the research is defined by the practical need to develop methods for adjusting electromagnetic compatibility and assessing the current technical condition of equipment through the calculation of surface losses in massive components made of structural steels used in electrical machine engineering, under conditions of rapidly varying incident magnetic fields combined with a constant component and a nonlinear dependence of magnetic permeability on the field intensity.

The scientific novelty lies in deriving analytical dependencies that describe the influence of a time-varying stray magnetic field of electrical equipment - changing

according to a harmonic law and containing a constant component—on surface losses in massive structural steels with nonlinear ferromagnetic properties.

The formulation of the scientific and technical problem. It is known that inactive components of electrical power equipment exhibit a pronounced surface effect, characterized by the concentration of eddy currents in a thin surface layer at locations with the highest electromagnetic field intensity. This allows, with sufficient practical accuracy, the consideration of a ferromagnetic half-space instead of massive bodies with complex configurations. This assumption is valid when all geometric dimensions of the studied components significantly exceed the skin depth.

We will research surface losses depending on the nature of electromagnetic field penetration into the ferromagnetic half-space, i.e. $\chi \geq 0$, characterized by conductivity σ and the nonlinear dependence between B induction and the magnetic field $H - B = \mu(|H|) * H$. The field is set on the surface $\chi = 0$ as a linearly polarized tangential component of the magnetic field.

I. Research Method and Result

In our case, the periodic (resulting) magnetic field contains a constant component H_0 and a first harmonic $H_1 \cos(\varphi - \omega t)$ with a period $T = 2\pi / \omega$.

Thus, further research consists of solving a nonlinear differential equation regarding the propagation of the electromagnetic field in ferromagnetic structural steels within the framework of given boundary and initial conditions**.**

The specificity of the electromagnetic field is rapidly varying**,** which excites non-stationary electromagnetic processes in electrical equipment. One of the most important parameters is surface losses in ferromagnetic structural steels. Based on the above analysis, the total electromagnetic field can be represented in the following form:

$$H_{\xi}(0) = H_{\Sigma}(0,t) + H_o + H_n \ , \qquad (1)$$

$$H_{\Sigma}(0,t) = \sum_{j=1}^{n} H_o(0,t) \ , \qquad (2)$$

$$H_{\nu}(0,t) = \left\{ \frac{1}{2} \int_{o}^{T} \left[H_m \cos(\nu\omega t + \varphi_{\nu}) \right]^2 \right\}^{0,5} . \qquad (3)$$

The indices (∑), (0), (n), (ν) denote rapidly varying**,** non-sinusoidal**,** constant**,** pulsating**,** and higher harmonic components. In power transformers**,** non-stationary electromagnetic processes occur. Therefore, the study of surface losses in massive ferromagnetic structural steels presents a rather complex nonlinear problem**.**

To specify the non-stationary electromagnetic process in the surface layer of a massive ferromagnet**,** we will briefly conduct a physical-analytical study in a logical sequence of its manifestation and further development.

The generalized boundary parameter of the ferromagnetic half-space is represented as complex surface resistance**.** This is a highly effective physical-analytical approach for studying the surface effect**,** as the geometric dimensions of inactive equipment components exceed the penetration depth of the electromagnetic field into them. Thus, we will further investigate the ferromagnetic half-space**.**

The use of surface resistance allows the calculation of surface losses based on averaged external parameters of the electromagnetic field, defined or measured on the surface, without investigating the field inside.

For massive ferromagnetic structural components, surface resistance under rapidly varying electromagnetic fields is expressed through its tangential components, similarly to a field that varies over time according to a harmonic law [6-8]:

$$Z_{\xi}(H_{\Sigma}(0,t), H_o) = R_{\xi} + jX_{\xi} = j\left\{ \left[2\pi \sum_{\nu=1}^{n} f_{\nu} \cdot \left| \dot{\mu}_e(H_{\Sigma}(0,t), H_o) \right| \right] / \sigma \right\}, \qquad (4)$$

where R and X are the active and reactive components of the surface impedance; $\sum_{\nu=1}^{n} f_{\nu}$ – the spectrum of higher harmonics (including the fundamental harmonic) of the rapidly varying field; σ – the electrical conductivity of structural steel; $|\mu_e|$ – the modulus of the equivalent magnetic permeability of structural steel.

The equivalent magnetic permeability is a generalized magnetic characteristic of structural steel. It depends on the physical properties and chemical composition of the steel and exhibits a nonlinear dependence on the magnetic field intensity and its constant component. Under rapidly varying magnetic fields, it is determined based on the averaged parameters of the electromagnetic field, similarly to harmonic fields, and at equal effective values, it assumes equal values. Thus, surface resistance and equivalent magnetic permeability are key parameters of the rapidly varying surface effect, reflecting the physicochemical properties of ferromagnetic structural steel and exhibiting a nonlinear dependence on magnetic field intensity.

Surface losses in massive ferromagnetic structural steel are concentrated in a thin surface layer. Therefore, they are expressed through the active component of surface resistance in combination with the intensity of the rapidly varying magnetic field and its constant component:

$$\omega_{\xi}(H_{\Sigma}(0,t), H_o) = \left[z_{\xi.o} \cdot R'(H_{\xi}(0,t)) \cdot H_{\xi}(0,t)^2 \right]^{0,5}, \qquad (5)$$

$$Z_{\xi.o} = \sum_{j=1}^{n} z_{o.\nu} \ ; z_{o.\nu} = (2\pi f \cdot \mu_o / 2\sigma)^{0,5} \ , \qquad (6)$$

– active resistance of structural steel is a linear parameter;

$$R'(H_{\xi}(0,t)) = \left[2\dot{\mu}'_e(H_{\Sigma}(0,t), H_o) \right]^{0,5} \cos\varphi(H_{\Sigma}(0,t), H_o) \, , \qquad (7)$$

– relative surface resistance of structural steel (a parameter reflecting nonlinear dependence on the magnetic field);

$$\dot{\mu}'_e(H_{\Sigma}(0,t), H_o) = \left[2\dot{\mu}_e(H_{\Sigma}(0,t), H_o) \right] / \mu_o \, , \qquad (8)$$

– relative equivalent magnetic permeability of steel;

$$\varphi = \frac{\pi}{4} - \frac{1}{2}\arg \dot{\mu}_e(H_\Sigma(0,t), H_o) \ , \qquad (9)$$

where $\mu_o = 4\pi \times 10^{-7}\, H/m$ – vacuum permeability.

Thus, the calculation of surface losses in massive ferromagnetic structural steels is reduced to determining the generalized magnetic characteristic – equivalent magnetic permeability**.** It is determined experimentally or analytically based on the known magnetic permeability at the first harmonic [8]. The most convenient approach for practical use is the analytical dependence of equivalent magnetic permeability on the parameters of the electromagnetic field applied to the surface:

$$\dot{\mu}_e(H_\xi(o,t), H_o) = j\left\{\left[\sigma \dot{H}_\xi(o,t)\right] / \left[2\pi f H_\xi(o,t)\right]\right\} , \quad (10)$$

$$H_\xi(0,t) = \left\{\frac{1}{T}\int_o^T \left[\sum_{\nu=1}^{n} H_{m\nu}\cos(\nu\omega t + \varphi_\nu)\right]^2 dt\right\}^{0,5} , \quad (11)$$

$$E_\xi(0,t) = \left\{\frac{1}{T}\int_o^T \left[\sum_{\nu=1}^{c} E_{m\nu}\cos(\nu\omega t + \varphi_\nu)\right]^2 dt\right\}^{0,5} , \quad (12)$$

where H_{mv} and E_{mv} are the maximum instantaneous values of the magnetic and electric fields; ω is the fundamental angular frequency; φ is the initial phase; *n* and *c* represent the harmonic composition of the magnetic and electric fields.

It should be noted that the harmonic composition of the magnetizing current is close to the harmonic composition of the magnetic field. However, the harmonic composition of the electric field differs significantly from that of the magnetic field, as seen in (11-12). This becomes more evident when analyzing the relationship between the parameters of the electromagnetic field and the electromagnetic characteristics of structural steel:

– for surface resistance

$$Z_\xi(H_\Sigma(0,t), H_o) = \frac{E_\xi(0,t)}{H_\xi(o,t)} = \sum_{\nu=1}^{n}\left\{\left[\omega\nu\mu\left(H_\Sigma(0,t), H_o\right)\right] : \sigma\right\}^{0,5} \ ; \quad (13)$$

– for surface losses

$$\omega_\xi(H_\Sigma(0,t), H_o) = \sum_{\nu=1}^{n}\left\{\left[\omega\nu\mu\left(H_\Sigma(0,t), H_o\right)\right] : \sigma\right\}^{0,5} \sum_{\nu=1}^{n} H_\nu(0,t) \ ; \quad (14)$$

– for the distribution of electromagnetic induction flux in the surface layer

$$F_\xi(H_\Sigma(0,t), H_o) = \sum_{\nu=1}^{n}\left\{\left[\mu\left(H_\Sigma(0,t), H_o\right)\right] : (\omega\nu\sigma)\right\}^{0,5} \sum_{\nu=1}^{n} H_\nu(0,t) \ ; \quad (15)$$

– for the intensity of the electric fiel

$$E_\xi(H_\Sigma(0,t), H_o) = \sum_{\nu=1}^{n} \omega\nu F_\xi(H_\Sigma(0,t), H_o) \ . \quad (16)$$

Equations (13)–(16) allow for an analytical investigation of ammeter readings and voltage drop across a non-inductive shunt in the magnetization circuit of the sample.

Losses were measured using the wattmeter method. The intensity of the magnetic and electric fields, as well as the distribution of electromagnetic induction flux in the surface layer of structural steel, were determined based on the parameters of the samples, the current in the primary winding, and the voltage in the secondary winding, respectively.

The results of experimental studies on the electromagnetic parameters of ferromagnetic structural steel St.3 are presented in Figures 2–4 and in tables. The harmonic composition of the corresponding electromagnetic field parameters can be determined if at least one of them is known or specified.

Experimental studies and numerical calculations of the parameters of the nonstationary surface effect in ferromagnetic structural steels were carried out on toroidal samples made of structural steels St.2, St.3, St.3sp., and St.09G2s**,** similarly to the methodology described in [8].

The cross-sectional diameter of the samples significantly exceeded twice the penetration depth of the variable component of the electromagnetic field and was more than ten times the average diameter of the toroid**.** Under these geometric conditions, the surface effect in the samples manifested approximately as in a ferromagnetic half-space**.**

Magnetizing and measuring windings were wound onto the samples. The number of turns was selected to cover regions of moderate and strong magnetic fields, which occur in power transformers, as well as in shielding and supporting components of electrothermal devices. The magnetizing winding was connected to a voltage generation source via a non-sinusoidality and DC component generator [4].

The magnetizing current varied from sinusoidal to strongly non-sinusoidal. The harmonic composition of the current was measured using a spectrum and frequency characteristic analyzer. The non-sinusoidality of the magnetizing current:

$$\beta_i = \alpha_i / \left\{\left[1 - (\alpha_i/100)\right]^2\right\}^{0,5} \qquad (17)$$

was monitored using a distortion analyzer, which measured the ratio of the sum of higher harmonic components starting from the second to the sum of the same components, including the fundamental (first) harmonic

$$\alpha_i = \left(\sum_{\nu=2}^{n} I_\nu^2 / \sum_{\nu=1}^{n} I_\nu^2\right)^{0,5} . \qquad (18)$$

The DC component of the magnetizing current was maintained constant relative to the AC component and was monitored through ammeter readings and voltage drop across a non-inductive shunt in the magnetization circuit of the sample. Losses were measured using the wattmeter method.

The intensity of the magnetic and electric fields, as well as the distribution of electromagnetic induction flux in the surface layer of structural steel, were determined based on the parameters of the samples, the current in the primary winding, and the voltage in the secondary winding, respectively.

The results of experimental studies on the electromagnetic parameters of ferromagnetic structural steel St.3 are presented in Figures 2–4 and Tables 1–2.

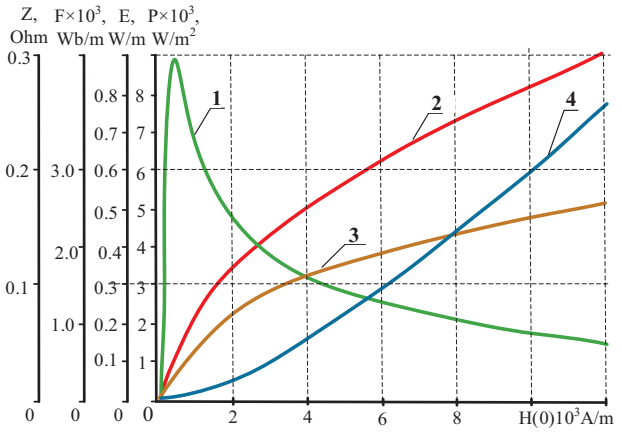


Fig. 2. Dependencies of surface resistance (1), electric field intensity (2), distribution of magnetic induction flux (3), and surface losses (4) of structural steel St.3sp as a function of alternating magnetic field at industrial frequency.

TABLE I. SURFACE LOSSES IN STRUCTURAL STEEL ST.3 DEPENDING ON THE DC COMPONENT OF THE SHARPLY VARYING MAGNETIC FIELD AT FIXED NON-SINUSOIDALITY COEFFICIENTS AND PERIODIC COMPONENT

Magnetization, %	28,3*			47,0*		
	1500	8000	12000	1500	8000	12000
0	427,2	4715,6	8277,5	513,4	5665,7	9945,3
10	389,4	4287,1	7525,2	467,6	5151,0	9041,1
30	296,5	3273,1	5745,2	356,3	3932,1	6902,4
50	184,3	2032,6	3567,8	222,1	2442,3	4286,7
70	154,7	1710,1	2996,5	185,9	2051,0	3600,2

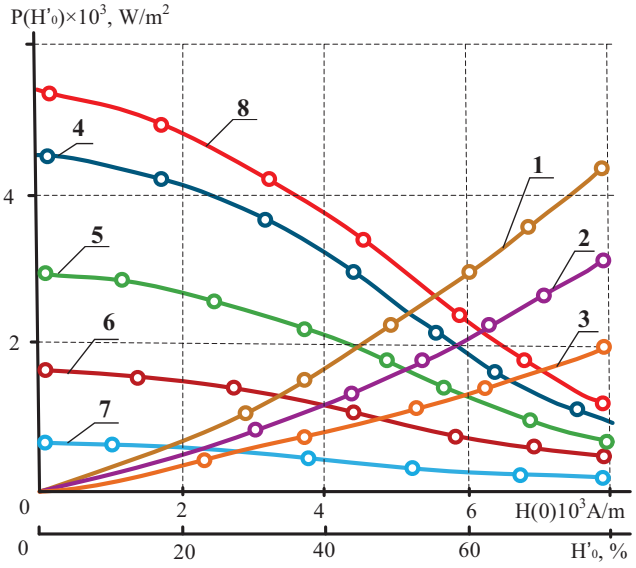


Fig. 3. Surface losses as a function of the constant biasing field at a fixed non-sinusoidal field 1 - $k\omega_p$= 1,18; 2 - $k\omega_p$= 1,0; 3 – H'_0 = 42 %; 4 – H(0) = 8000 А/м; 5 – H(0) = 6000 А/м; 6 – H(0) = 4000 А/м; 7 – H(0) = 2000 А/м; 8 – $H_2(0)$ = 8000 А/м.

TABLE II. SURFACE LOSSES IN STRUCTURAL STEEL ST.3 DEPENDING ON THE NON-SINUSOIDALITY COEFFICIENTS OF THE EXCITATION CURRENT AT FIXED MAGNETIC FIELD INTENSITIES

THD, %	1500		8000		12000	
	Measurement	Calculation	Measurement	Calculation	Measurement	Calculation
2,1	398,1	395,7	4398,7	4396,9	7721,4	7722,1
28,3	427,2	428,4	4715,6	4716,3	8277,5	6275,3
37,2	461,8	462,3	5102,7	5102,3	8957,1	8958,3
47,0	513,4	513,9	5665,7	5686,4	9945,3	9945,9
51,1	545,8	546,3	6031,2	6032,5	10586,5	10587,4
67,3	595,7	596,3	6584,6	6585,1	11558,4	11561,2

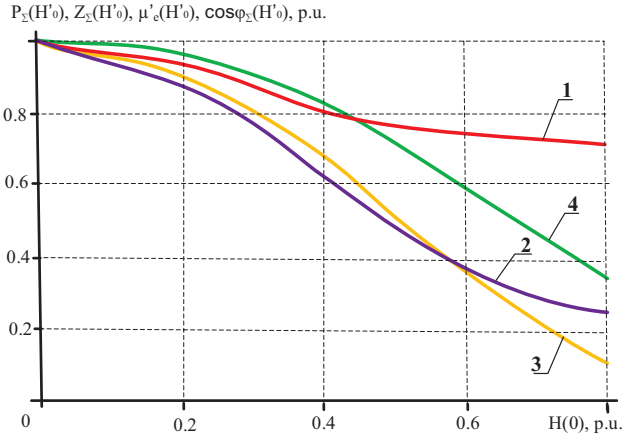


Fig. 4. Normalized surface effect parameters as a function of the biasing DC field at a fixed AC component for the region of moderate and strong fields; 1 – cos φ_Σ; 2 – P_Σ; 3 – $|\mu_E|$; 4 – Z_Σ

II. CONCLUSION

1. For the power supply systems of energy-intensive industrial enterprises with sharply variable operating modes, it is necessary to use special-purpose power transformers. For their design, it is essential to develop engineering methods for calculating nonstationary electromagnetic processes, as well as methods for reducing local and overall additional losses.

2. To specify the design of special-purpose transformers, it is advisable to develop new *"Technical Requirements"* and *"Technical Specifications"*. In these documents, in addition to the rated operating modes, it is necessary to include the permissible overloads, their multiplicity, duration, and characteristics.

3. Studies of nonstationary electromagnetic processes in special-purpose transformers have been conducted, enabling the assessment of the maximum possible impact on associated technical systems through changes in additional losses under rapidly varying loads and specific operational conditions, in accordance with the resolution of IEC Technical Committee 77 "Electromagnetic Compatibility".

4. The initial and boundary conditions have been formulated to represent the dynamics of nonstationary electromagnetic processes in massive ferromagnetic structural steels. This simplifies the task of studying nonstationary electromagnetic phenomena in engineering calculations of losses in non-active parts of electrical equipment structures. The calculation error for additional losses does not exceed 6% in weak magnetic field regions and 3.5% in magnetic field conditions characteristic of electrical equipment operation.

5. As a nonlinear magnetic characteristic of massive structural steels (with ferromagnetic properties), it is effective to use the equivalent magnetic permeability depending on the simultaneous influence of a sharply varying magnetic field and its DC component.

6. As a boundary parameter for structural steels (with a nonlinear dependence of the magnetic characteristic on the magnetic field), it is advisable to use complex surface resistance, which allows obtaining analytical dependencies between surface losses, electric and magnetic fields, and the distribution of electromagnetic induction flux without resorting to studies of field penetration into the half-space.

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Magnetic Pulse Complex for Attraction of Thin-Walled Metals with Resonant Charge of a Capacitive Energy Storage Device

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***Abstract*—The purpose of this work is to propose and substantiate the workability of a magnetic-pulse complex for attracting thin-walled metals, consisting of two successive circuits with a common capacitive energy storage device, the first of which is a resonant charger powered by an industrial voltage and frequency network, and the second unit is a discharge circuit with an inductor-tool for performing the assigned production task. When deriving the main analytical relationships for calculating the characteristics of the proposed resonant magnetic-pulse complex designed to attract sheet metals and operating as a resonant amplifier of the energy of an electric signal from an industrial voltage and frequency network, the mathematical apparatus of the theoretical foundations of electrical engineering and methods of integrating differential equations were used. The high efficiency of resonant formation of charge characteristics of a capacitive electric energy storage device from an industrial network of ~ 220 V at a frequency of ~ 50 Hz for ~ 0.1 s to a voltage of ~ 4595.85 V (gain is ~ 14.78 times) with a stored energy of ~ 2640 J (energy of the source ~ 8.47%, resonant energy ~ 91.53%) is shown. For a discharge circuit at a frequency of ~ 1500 Hz, it was obtained that the maximum energy in the inductor-tool is ~ 8.25 times higher than the energy consumed from an external power source. The discharge current in the inductor-tool is ~ 45270.4 A. With this current value there should be a sufficiently effective force impact on the workpiece.**

Keywords—electromagnetic forming, series active-reactive electric circuits, voltage resonance, amplifier of electric energy

I. Introduction, a Brief Review of the Literature

Achievements in the area of the creation of magnetic-pulse installations (MPI) as powerful energy sources ultimately determine the prospects for using pulsed electromagnetic fields in the implementation of modern environmentally friendly, resource-saving and high-performance technologies for metal pressure processing. Without going through numerous works devoted to this topic, we can highlight some of the main scientific publications [1, 2, 3, 4, 5]. Thus, the authors of the scientific publications [1, 2, 3] describe in sufficient detail examples of magnetic-pulse installations created at different times by world electrical manufacturers. The paper highlights the achievements in the development of metal forming tools by companies in Germany (Siemens), Great Britain (Wickmen Machine Tools), the USA (General Dynamics Corporation), and others [6, 7]. An extensive review in the papers [3, 8, 9], presented by a large group of co-authors from various German organizations, is devoted to describing the developments of technical aspects in the implementation of various stamping, assembly, and welding production operations [10, 11, 12, 13]. The proceedings of the international conferences [4, 5] highlight the key problems and achievements of modern metal-forming methods.

Finally, scientists under the supervision of Professor, Doctor of Engineering Sciences G.A. Shneerson describe in sufficient detail the problems and approaches to their solution in the field of obtaining super-strong pulsed electromagnetic fields not only for industrial purposes but also for large-scale physical research [14].

Some of the works, opening a new scientific direction, is devoted to the attraction of specified sections of thin-walled metals using low-frequency pulsed electromagnetic fields [15, 16]. The cited articles theoretically and experimentally illustrate the physics of the processes taking place, formulates the conditions for their implementation and examples of application in current modern technologies.

Successes in the study of natural phenomena make it possible to reach a qualitatively new level of technical capabilities in various areas of human activity. Focusing on electrical engineering and applied electrodynamics, it should be noted that the phenomenon of resonance and the introduction of resonant elements into the circuits of pulsed magnetic field generators should significantly increase the efficiency of technologies where they are used as sources of electrical power [17, 18, 19]. The first to point out this fact was Nikola Tesla. His developments, completed at the beginning of the last century, showed Mankind the way to the creation of powerful energy based on resonant electrical structures [17]. An example of the practical development of N. Tesla's ideas is a two-circuit resonant amplifier of electric power for magnetic-pulse processing of metals, presented in the work [19]. As is known, in electrical engineering, resonant effects appear when exciting exclusively harmonic oscillations in serial and parallel active-reactive circuits [20]. Theoretical

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and experimental attempts to amplify reactive electric power at voltage resonance are covered in works [21, 22]. Their conclusions allow us to move on to the consequences of their generalization, according to which the use of active-reactive electrical circuits can provide a solution to the pressing problems of modern electric power engineering. We are talking about the creation of an amplifier of the active electric power of currents and voltages. In the technical implementation, this device consists of two sequential active-reactive circuits with a common capacitive storage of electrical energy. In general, the operating principle of the proposed circuit consists of charging the capacitive storage to a given level of stored energy with its subsequent discharge to the active load of the amplifier. It should be noted that from a physical point of view, a magnetic-pulse installation made according to the above-described resonant circuit, ultimately represents a resonant energy generator for mechanical processing of metals according to a given production program.

An undoubted and significant advantage of using serial active-reactive circuits instead of bulky power equipment of traditional charging devices is the structural simplicity and the ability to increase or decrease voltage due to variation of the characteristics of the charging circuit (quality factor and process time). Moreover, the gain in the resonant circuit will be incomparably higher than in the circuit of a traditional charging device.

It should be emphasized that since the essential novelty of the subject of this article consists in the use of a resonance phenomenon for charging a capacitive energy storage device, the main attention will be paid to this practical application of the noted natural effect.

The purpose of this work is to propose and substantiate the actability of a magnetic-pulse complex for the attraction of thin-walled metals, consisting of two successive circuits with a common capacitive energy storage device, the first of which is a resonant charging device powered by an industrial voltage and frequency network, and the second block is a discharge circuit with an inductor-tool for performing the assigned production task.

II. Scheme, Principle of Operation. Statement of the Problem

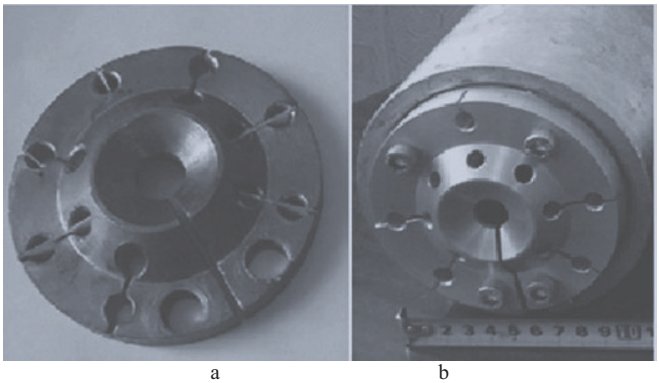
In announcing the subsequent consideration, it should be noted that all the characteristics of the equipment, as the initial data for the present numerical estimates, necessary to substantiate the legal capacity of the subject of consideration as a whole, are taken from the work [15].

A. Experimental Magnetic-Pulse Complex – MIUS-2

Charging circuit: source - three-phase network (~380 V), step-up voltage transformer to ~2500 V, rectifier, electric energy storage with a capacity of 1200 μF.

Discharge circuit: voltage on the capacitor – 2000 V, discharge current – 38000 A, frequency – 1900 Hz (at the output of the pulse current transformer with a gain of ~ 5, Fig. 1), circuit inductance – 5.84 μH, active resistance – 0.04 Ohm.

Result of magnetic-pulse attraction: a dent of ∅ 0.02 m in a steel sheet with a thickness of ~ 0.00075 m and a depth of ~ 0.0015 m was removed.



a b

Fig. 1. The tool in experiments on the magnetic-pulse attraction of sheet ferromagnets: a – single-turn inductor-tool, b – inductor with pulse current transformer assembly

B. Description of the Proposed Scheme

Fig. 2 shows the basic scheme of the resonant magnetic-pulse complex, which is, from the physical point of view, a pulse resonant amplifier of active electrical energy and is intended for the attraction of specified areas of the surface of sheet metals.

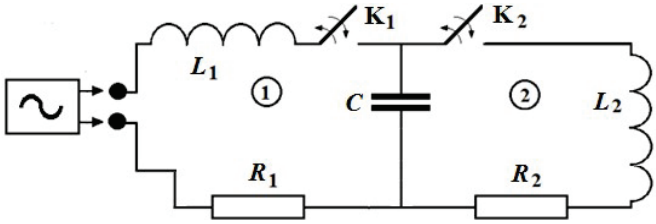


Fig. 2. Schematic diagram of a magnetic-pulse complex consisting of 2 series resonant circuits with a common capacitive energy storage device – C, 1 – charging circuit with inductance – L_1 and active resistance – R_1, 2 – discharge circuit with inductance – L_2, including the inductance of the inductor tool with current leads, and active resistance – R_2

Note. Due to equivalence, the object of this study (Magnetic Pulse Complex) in the further presentation of the material may be called either a resonant amplifier of active electrical energy or a resonant MPI (Magnetic Pulse Installation).

On the whole, the circuit of the proposed complex consists of two blocks with a common capacitive storage C. The first of the blocks (a series active-reactive circuit, designated by the number – 1) is a capacitance charging block, from a physical point of view, representing a reactive power amplification block. The second block (a series active-reactive circuit, designated by the number – 2) is a discharge block, physically representing a block for converting the energy stored in the capacitive storage into magnetic-pulse attraction energy generated by the inductor-tool.

In Fig. 2 L_1, R_1, L_2, R_2 – designations of the integral parameters of the circuit under consideration, respectively.

C. Operating Principle

Initial state: switch – K_1 is closed, switch – K_2 is open. The electric signal from the network excites the series charging contour – 1. The oscillating and increasing voltage on the capacitor is interrupted upon reaching a predetermined value by opening the switch K_1. The charge stops. Switch K_2 is being opened. The capacitive storage is discharged onto the winding of the inductor-tool. In its working area, attractive forces are excited due to the magnetic properties of the processed ferromagnetic sheet. Note that these attractive forces in the low-frequency mode of the acting fields significantly exceed the repulsive Lorentz forces. The selected area on the surface of the thin-walled metal will be attracted to the working surface of the inductor-tool [16].

Upon completion of the specified discharge time, switch – K2 is opened and K1 is closed. The circuit returns to its original state of readiness to repeat the process of charging the capacitive storage device of the electrical energy.

D. Statement of the Problem

- The objective of this consideration is to obtain the output characteristics of the proposed resonant scheme, ensuring the practical implementation of a given production operation.
- External power source – industrial voltage network ~ 220 V (amplitude value $E_m = 311$ V) with a frequency of $f_1 = 50$ Hz.
- The choice of the capacity of the electric energy storage device is arbitrary $C = 250 \cdot 10^{-6}$ F.
- Natural frequency of the charging circuit – $f_1 = 50$ Hz.
- Natural frequency of the discharge circuit – $f_2 = 1500$ Hz.
- Active resistances of the charging and discharge circuits – $R_{1,2} = 0.1$ Ohm.

Note. The choice of the discharge circuit frequency is determined by the requirement of its sufficiently low value: $f_2 = 1500$ Hz < 1900 Hz (discharge circuit). The active resistances of the element base should not disturb the oscillatory mode in active-reactive circuits: $R_{1,2} = 0.1$ Ohm < 0.042 Ohm (discharge circuit).

III. Calculation Ratios

All the calculation ratios given were obtained by well-known methods of electric circuit theory described in works [20, 23, 24]. To avoid cumbersome presentation, mathematical transformations are omitted.

According to the adopted formulation of the problem, we will separately consider and analyze the transient processes in each of the amplifier blocks.

A. Charging Circuit

The voltage $U_C(\varphi)$ on the capacitor:

$$U_C(\varphi) \approx - E_m \cdot Q_1 \cdot \left(1 - e^{-\frac{1}{2 \cdot Q_1} \cdot \varphi}\right) \cdot \cos(\varphi) \qquad (1)$$

where Q_1 is circuit quality factor – 1, $\varphi = \omega_1 \cdot t$ is a current phase, ω is the cyclic frequency, t is a time.

Charging current $J_1(\varphi)$ in the circuit:

$$J_1(\varphi) \approx \frac{E_m}{R_1} \cdot \left(1 - e^{-\frac{1}{2 \cdot Q_1} \cdot \varphi}\right) \cdot \sin(\varphi) \qquad (2)$$

The energy stored by the capacitor in relation to the energy of the power source:

$$\Delta W_C(\varphi) = \frac{Q_1}{2} \cdot \frac{\left(1 - e^{-\frac{1}{2Q_1}\varphi}\right)^2 \cdot \cos^2(\varphi)}{\int_0^{\varphi} \left(1 - e^{-\frac{1}{2Q_1} \cdot x}\right) \sin^2(x) \cdot dx} \qquad (3)$$

Equations (3) allows to make a quantitative assessment of the resonant charge of the capacitive storage device in comparison with the energy consumed from an external source. It is clear that the process of charging to the maximum voltage and stored energy is limited in time. At the same time, the energy consumption from an external source grows indefinitely over time. Thus, its saving and the consumption of resonant electrical energy for charging the capacitive storage device are interconnected. Their ratio is established by choosing the time t_1 of the charging process.

On the whole, formulas (1 – 3) completely describe the amplitude and phase-time dependencies of the operating characteristics of the circuit – 1.

B. Discharge circuit

The voltage $U_C(\varphi)$ on the capacitor in the discharge:

$$U_C(\varphi) = U_C(\varphi_1) \cdot e^{-\frac{1}{2 \cdot Q_2} \cdot \varphi} \cdot \sin(\varphi) \qquad (4)$$

where $U_C(\varphi_1) \approx - E_m \cdot Q_1 \cdot \left(1 - e^{-\frac{1}{2 \cdot Q_1} \cdot \varphi_1}\right) \cdot \cos(\varphi_1)$ is the voltage at the end of the charging process, $\varphi_1 = \omega_1 \cdot t$ is the voltage phase at the end of charging, ω_1 is the cyclic frequency in the circuit is 1, t is the charging time.

The discharge current $J_2(\varphi)$ in the circuit:

$$J_2(\varphi) = \frac{U_C(\varphi_1)}{R_2 \cdot Q_2} \cdot e^{-\frac{1}{2 \cdot Q_2} \cdot \varphi} \cdot \sin(\varphi) \qquad (5)$$

where $\varphi = \omega_2 \cdot t$ is a current phase, R_2 and Q_2 is the equivalent active resistance and quality factor of the discharge circuit, respectively.

Field energy in the inductance:

$$W_{L_2}(\varphi) = W_m \cdot e^{-\frac{1}{Q_2} \cdot \varphi} \cdot \sin^2(\varphi) \qquad (6)$$

where $W_m = \frac{L_1}{2} \cdot \left(\frac{E_m}{R_1} \cdot \left(1 - e^{-\frac{1}{2 \cdot Q_1} \cdot \varphi_1} \right) \cdot \cos(\varphi_1) \right)$ is the maximum.

The field energy in the inductance relative to the energy of the power source:

$$\Delta W_{L_2}(\varphi) = \Delta W_C(\varphi_1) \cdot e^{-\frac{1}{Q_2} \cdot \varphi} \cdot \sin^2(\varphi) \qquad (7)$$

Equations (4 – 7) completely describe the electromagnetic processes in the discharge circuit of the proposed circuit in Fig. 2.

IV. Numerical Estimates, Analysis

A. Charging cirquit – 1.

1) Circuit parameters.

a) Frequency $f_1 = 50$ Hz, $\omega_1 = 314.159$ s^{-1}.

b) Capacity $C = 250$ μF.

c) Inductance $L_1 = \frac{1}{\omega_1^2 \cdot C} = 0.0405$ H.

d) Equivalent active resistance $R_1 = 0.1$ Ohm.

e) Relative damping decrement, $\delta_1 = \frac{R_1}{2 \cdot L_1 \cdot \omega_1} = 0.0039$, which ensures the oscillatory nature of the charging process.

f) Quality factor $Q_1 = \frac{1}{\omega_1 \cdot R_1 \cdot C} = \frac{\omega_1 \cdot L_1}{R_1} = 127.324$.

2) Characteristics of the charging process.

a) Phase duration of charging $\varphi_1 = 10\pi$, charging time $t_1 = 0.1$ s.

b) Stored charge energy relative to source energy $\Delta W_C(\varphi_1) = 11.8$.

c) Stored charge energy $\Delta W_C(\varphi_1) = 2640$ J.

d) Charging voltage on the capacitor relative to the source voltage, $\frac{U_C(\varphi_1)}{E_m} = 14.78$.

Brief comments. The performed estimates are quite "transparent". Their results are easily interpreted from the electrophysical point of view. Thus, the starting point can be the known directly proportional relationship in the stationary mode between the voltage on the capacitor and the voltage of the source with a coefficient equal to the quality factor Q of the circuit. In the transient mode, the coupling coefficient begins to depend on time, but a significant increase in the voltage on the capacitor takes place, which is shown by the calculation results (Q = 14.78).

Simple approximations can be made by taking the ratio of the energies stored in a capacity and extracted from a source –

$$W_C(t) \Big/ W_i(t) = \frac{Q \cdot \int_0^t E(t) \cdot I(t) dt}{\int_0^t E(t) \cdot I(t) dt} = Q.$$

e) Charging voltage on the capacitor $U_C(\varphi_1) = 4595.85$ V.

f) Resonant part of the stored energy of the capacitance ~ 91,53 %.

g) Part of the stored energy of the container received from the source ~ 8,47 %.

Below are graphic illustrations of electromagnetic processes in the charging circuit – 1 of the magnetic-pulse complex circuit under consideration.

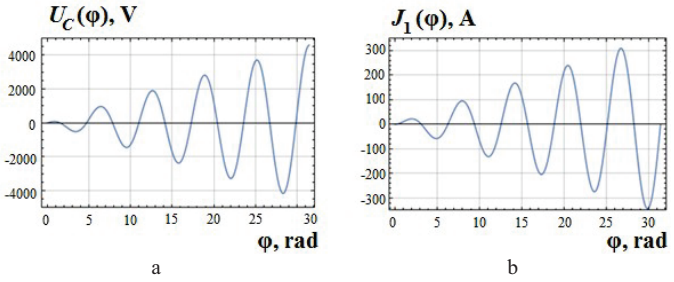


Fig. 3. Phase dependences of charging voltage – a and current – b

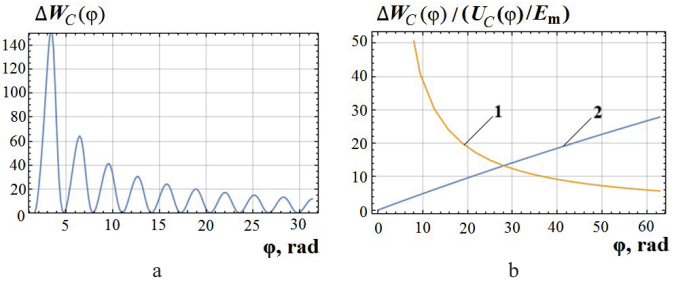


Fig. 4. Phase dependence of the energy stored by the capacitance, relative to the energy of the source – a, phase dependences of the maximum values of the relative energy of a capacitive storage device – b

Brief comments. Fig. 3. Both the voltage on the capacitance and the current in the circuit exponentially grow over time. The peculiarity of these processes is that at the moment of interruption of the charge at the maximum voltage, the current tends to zero. This creates favorable conditions for practical switching of the charging circuit.

In Fig. 4a the process of charging a capacitive storage device, unlike traditional circuits with rectifiers, is wave-like. During almost half of every period, the capacitance stores the corresponding maximum energy, and this maximum significantly exceeds the energy taken from the source (more than ~140 times at the beginning of the process). Fig. 4b, where $\Delta W_C(\varphi)$ – curve 1, and the relative charging voltage, $(U_C(\varphi)/E_m)$ – curve 2. The relative resonant energy of the capacitive storage device charge naturally drops to zero over time. Ultimately, only the energy of the source should and will be consumed. At the same time, the voltage on the capacitance will grow to its steady-state value. It can be assumed that the duration of the most effective charge (relative energy and the magnitude of the charging voltage in combination) is determined by the abscissa to the point of intersection of the curves in this figure.

3) Summary.

The considered charging circuit operates from an industrial network of ~ 220 V at a frequency of ~ 50 Hz. The charging time of the capacitive storage device is ~ 0.1 s to a voltage of ~ 4595.85 V (the gain is ~ 14.78 times) with a stored energy of ~ 2640 J (source energy ~ 8.47%, resonant energy ~ 91.53%).

B. Discharging cirquit – 2.

1) Circuit parameters.

a) Frequency $f_1 = 1500$ Hz, $\omega_2 = 9424.78$ s^{-1}.

b) Capacity $C = 250$ μF.

c) Inductance $L_2 = \dfrac{1}{\omega_2^2 \cdot C} = 45$ μH.

d) Equivalent active resistance $R_2 = 0.1$ Ohm.

e) Quality factor $Q_2 = \dfrac{1}{\omega_2 \cdot R_2 \cdot C} = \dfrac{\omega_2 \cdot L_2}{R_2} = 4.244.$

f) Relative damping decrement $\delta_2 = \dfrac{R_2}{2 \cdot L_2 \cdot \omega_2} = 0.1178$, which ensures the oscillatory nature of the charging process.

2) Characteristics of the discharging process.

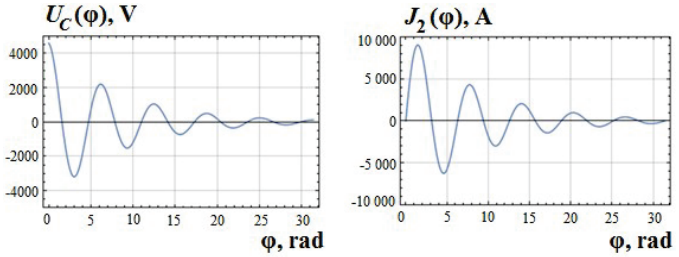
a) Phase duration of discharge, the choice is arbitrary, let it be the charging time in calculations $t_2 = 0.0012 \div 0.0067$ s.

b) Stored charge energy $\Delta W_C(\varphi_1) = 2640$ J.

c) Charging voltage on the capacitor $\Delta W_C(\varphi_1) = 4595.85$ V.

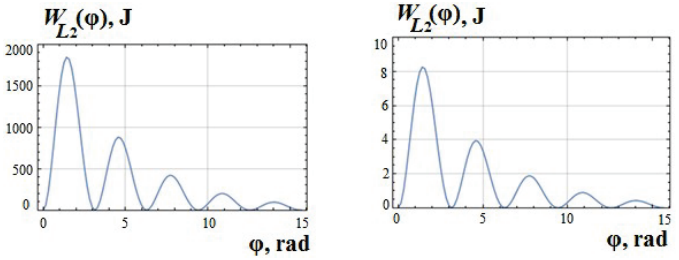
d) The maximum current in the circuit is ~ 9054 A.

e) Maximum current in a single-turn inductor-tool ~ 452704 A.



a b

Fig. 5. Phase dependences of the discharge voltage on the capacitor – a, and the current in the circuit – b



a b

Fig. 6. Phase dependences of energy in the inductance of the inductor-tool in absolute units of measurement – a, in relation to the energy of the external power source – b

Note. An increase in the current amplitude in a single-turn inductor-tool (Fig. 1) is achieved by including a step-up pulse transformer with a gain of ~ 5 in the discharge circuit – 2.

Above are graphs of the main characteristics of electromagnetic processes in the discharge circuit – 2 of the magnetic-pulse complex circuit under consideration.

Brief comments. Fig. 5. The dependences of voltage and current in the discharge circuit are exponentially decreasing. It should be noted that the current in the single-turn inductor-tool, increased to a value of more than ~ 45 kA, exceeds the current of ~ 38 kA (by ~ 7 kA) in the experiment, where a dent in a thin-walled steel sheet was successfully removed [15].

Fig. 6. The presented graphs relate to the energy of the inductor-tool. The energy of the magnetic field, which does the work of deforming the object of processing, has a wave-like character that decays over time. Due to resonance, its amplitude is more than 8 times greater than the energy consumed from an external source. In absolute terms, it is ~ 1850 J. Losses in comparison with the energy stored by the capacitive storage ~ 2640 J are obviously explained by dissipation in the active resistance of the circuit.

3) Summary.

The considered discharge circuit operates at a frequency of ~ 1500 Hz. The frequency can be varied by selecting the circuit inductance. The maximum energy in the inductor-tool is ~ 8.25 times greater than the energy consumed from an external power source. Resonance energy is ~ 88%, external source energy is ~ 12%. The discharge current in the inductor-tool, increased by turning on the step-up pulse transformer (gain factor ~ 5), is ~ 45270.4 A. With this current value, there

should be a sufficiently effective force impact on the object being processed.

CONCLUSIONS

The workability of a magnetic-pulse complex for attracting thin-walled metals, consisting of two successive circuits with a common capacitive energy storage device, the first of which is a resonant charging device powered by an industrial voltage and frequency network, and the second block is an active-reactive discharge circuit with an inductor-tool for performing the assigned production task, is proposed and substantiated.

High efficiency of resonant formation of charge characteristics of a capacitive electric energy storage device from an industrial network of ~ 220 V at a frequency of ~ 50 Hz is shown. Charging time is ~ 0.1 s to a voltage of ~ 4595.85 V (gain is ~ 14.78 times) with stored energy of ~ 2640 J (source energy ~ 8.47%, resonant energy ~ 91.53%).

For a discharge circuit at a frequency of ~ 1500 Hz (the frequency can be varied by the circuit inductance), it was found that the maximum energy in the inductor-tool is ~ 8.25 times greater than the energy consumed from an external power source. Resonance energy is ~ 88%, external source energy is ~ 12%. The discharge current in the inductor-tool, increased by turning on a step-up pulse transformer (gain factor ~ 5), is ~ 45270.4 A. With this current value, there should be a sufficiently effective force impact on the workpiece.

In comparison with the traditional MPI circuit the advantages of the proposed circuit of the magnetic-pulse complex are significant savings in consumed electricity due to resonance phenomena and a significantly lower cost, due to the absence of expensive traditional components of the charging device of the capacitive storage device of electrical energy.

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Analytical Upper Bound on the Intensity of Cherenkov Radiation at a Planar Interface Between Media

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***Abstract*—This work focuses on one specific aspect of phenomena associated with the electromagnetic radiation generated when a plane, density-modulated electron beam moves parallel to a vacuum-medium interface. In the considered case of a planar interface, such an electron beam gives rise to Cherenkov radiation. Within this configuration, by means of an exact analytical approach, we have derived an upper bound for the intensity of Cherenkov radiation. This theoretical finding is corroborated by the presented results of numerical experiments, which match the predicted intensity limit across a range of beam parameters and material properties. Obtaining a closed-form analytical solution to an electromagnetic problem of this complexity is quite rear and thus constitutes a significant result. Not only it deepens our understanding of diffraction radiation phenomena, but also provides a rigorous benchmark for testing and validating various numerical methods.**

Keywords—analytical models, closed-form solutions, electromagnetic radiation, electron beams, diffraction radiation, Cherenkov radiation

I. Introduction

When a plane, density-modulated electron beam flies parallel to a periodic structure, a set of homogeneous plane electromagnetic waves is generated. The number of propagating and evanescent waves, along with their respective wavelengths and directions of propagation, are determined by the electron beam's speed and modulation, as well as by the geometric and material properties of the structure. A similar beam traveling at a constant speed through a medium, where the speed of light is lower than that of the charged particles, generates two homogeneous plane electromagnetic waves diverging from the beam's axis. The wavelength and propagation direction of these waves depend on the beam's modulation period, the ratio of its speed to the speed of light in the medium, and on the material properties of the medium. These two phenomena, discovered in the mid-20th century, are known as Smith-Purcell effect and Cherenkov radiation, respectively [1], [2]. These effects could be viewed as particular manifestations of the more general phenomenon of diffraction radiation [3]–[6]. The huge potential of these phenomena was recognized almost immediately after discovery, sparking a great deal of interest from both academic and applied perspectives – a level of attention that remains strong to this day [3]–[9]. This sustained interest has yielded a wide variety of devices in which diffraction radiation and related effects play a key role. These include such applications as leaky wave and diffraction radiation antennas [10]–[12], diffraction radiation generators [13], [14], non-invasive beam parameters monitors [15], terahertz radiation sources [16], electromagnetic energy compressors [17], [18], and the development of metamaterials and metasurfaces [19].

Rapid progress in comprehension and harnessing diffraction radiation effects would be impossible without numerical modeling. One of the most trusted numerical methods employed for this purpose is the method of analytical regularization (MAR) [5], [20], [21]. MAR is an umbrella term for a family of methods based on the transformation of first-kind or strongly-singular operator equations into Fredholm second-kind operator equations via the analytical inversion of their singular components. This transformation ensures the convergence of discretizations and, consequently, the controllable accuracy of numerical solutions. The modern variant of MAR, based on the given-current approximation, is presented in the book [5] and the series of papers [22]–[25]. In that framework, the continuity conditions for field components are formulated as an ill-conditioned infinite system of linear algebraic equations; through analytical regularization, this system is converted into an equivalent Fredholm operator equation, whose numerical solution can be obtained with any predetermined accuracy via the truncation method. The analysis is carried out with the use of the so-called wave analog of effects under study. That is, a real, physical electron beam is not present in the models. Instead, its field is identified with that of an inhomogeneous electromagnetic wave, which, under certain conditions, gives rise to the same field as a plane, density-modulated electron beam. This is a legitimate substitution, as the field of plane waves propagating inside or outside a structure is generated by this electron beam's field. This generalization allows to study excitation not only by electron beams but also, for instance, by surface waves of open waveguides [4], [5], [17] [22].

As emphasized in [23], [24], the analysis of diffraction radiation on a new periodic structure should begin with its limiting case where the corrugation depth tends to zero – a planar structure. It has been shown that planar and periodic structures exhibit similar spectral characteristics, namely properties of their eigenwaves, which play a defining role in the nature of diffraction radiation. Eigenwaves of structures with planar boundaries transition smoothly into eigenwaves of corresponding periodic structures as the corrugation depth changes smoothly. Naturally, periodic structures exhibit a much richer resonant behavior than their planar counterparts. Nevertheless, a planar interface, being a simpler configuration to study, allows for the investigation and classification of

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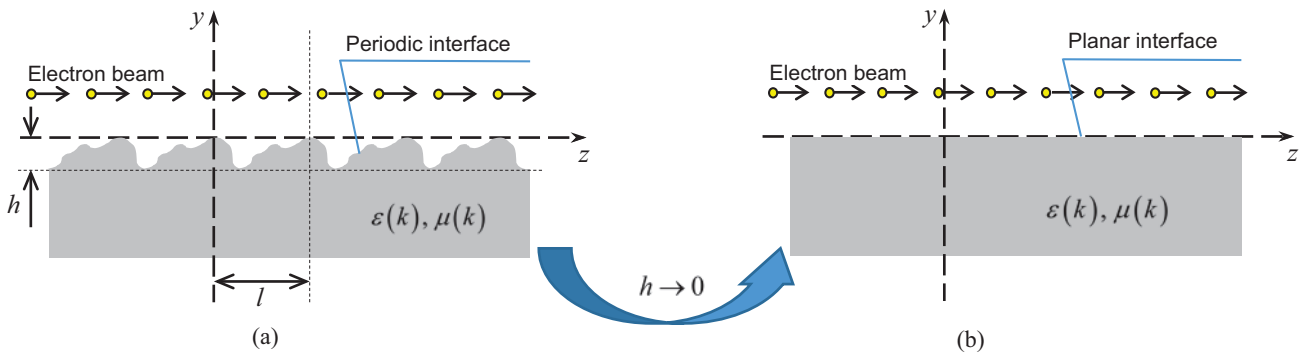


Fig. 1. Planar interface between vacuum and a medium as a limiting case of a periodically corrugated interface when the corrugation depth tends to zero. (a) Periodically corrugated interface ($h>0$); (b) Planar interface ($h=0$).

these eigenwaves [23], [24], and, most importantly, it enables the identification of parameter values that have the potential to support regimes of high-intense diffraction radiation on periodic structures. All this constitutes a good starting point for the subsequent analysis of periodic structures.

In this work, we present a significant result obtained while studying the interaction of an electron beam with a planar vacuum-medium interface: an analytically derived upper bound for the intensity of diffraction radiation from planar structures when excited by an inhomogeneous excitation field equivalent to that of a modulated electron beam. The obtained result is of both theoretical and applied importance. It deepens our understanding of diffraction radiation phenomena, allows for the determination of conditions under which the maximum possible level of diffraction radiation is achieved, which is particularly relevant for the design of high-efficient devices, and offers a rigorous benchmark for testing and validating various numerical solvers.

II. Problem Formulation

We begin with a formulation of the problem for the most general case of a periodic vacuum-medium interface, as it is our main object of interest. Then we consider a planar interface as its extreme limit corresponding to vanishing corrugation depth. From this simplified case, we analytically derive the upper bound for the radiation intensity.

A. General Case of a Periodically Corrugated Interface

Consider the structure shown in Fig. 1a. It is an interface between vacuum in the upper half-space and a medium in the lower half-space, whose permittivity and permeability are $\varepsilon(k)$ and $\mu(k)$. Here, k is a wavenumber, it plays the role of a frequency parameter. The structure is homogeneous along x direction ($\partial_x \equiv 0$) and periodic in z direction with period l, its profile is set by the periodic function $y = f(z)$, $-h \le f(z) \le 0$. Here, h is the interface corrugation depth.

Let a plane, density-modulated electron beam flies parallel to this structure, Fig. 1a. The beam is characterized by its modulation wavenumber k and relative (to the speed of light c) speed $0 < \beta < 1$. The interaction of the beam with the structure forms an H-polarized ($E_x = H_y = H_z = 0$) total electromagnetic field $\vec{E}, \vec{H}$, whose H_x component's behavior within $0 \le z \le l$ (one period of the structure) is descried by the following boundary value problem:

$$\begin{cases} \left[\partial_y^2 + \partial_z^2 + \varepsilon(y,z,k)\mu(y,z,k)k^2\right] H_x(y,z,k) = 0, -h < y < 0; \\ H_x(y,l,k) = \exp(2\pi i\zeta) H_x(y,0,k) \quad \text{and} \\ \quad \partial_z H_x(y,l,k) = \exp(2\pi i\zeta)\partial_z H_x(y,0,k), \quad -h \le y \le 0; \\ \vec{E}_{\text{tg}}, \vec{H}_{\text{tg}} \text{ are continuous when crossing the interface} \\ \quad \text{and at } y = 0, \ y = -h; \end{cases}$$

$$H_x(y,z,k) = \exp\left(-i\Gamma_0^+ y\right)\varphi_0(z) + \sum_{n=-\infty}^{\infty} R_n(k)\exp\left(i\Gamma_n^+ y\right)\varphi_n(z), \quad y \ge 0;$$

$$H_x(y,z,k) = \sum_{n=-\infty}^{\infty} T_n(k)\exp\left(-i\Gamma_n^-(y+h)\right)\varphi_n(z), \quad y \le -h. \tag{1}$$

Here, $\vec{E}_{\text{tg}}, \vec{H}_{\text{tg}}$ are tangential components of electric and magnetic field vectors, $\varphi_n(z) = l^{-1/2}\exp(i\Phi_n z)$ is a complete orthonormal system of transverse functions, $\Phi_n = (n+\zeta)2\pi/l$ are horizontal propagation constants, $\zeta \in \mathbb{R}$ is the phase shift along the period, $\Gamma_n^+ = \sqrt{k^2 - \Phi_n^2}$: $\operatorname{Re}\Gamma_n^+ \ge 0, \quad \operatorname{Im}\Gamma_n^+ \ge 0$ and $\Gamma_n^- = \sqrt{k^2\varepsilon(k)\mu(k) - \Phi_n^2}$: $\varepsilon^{-1}(k)\operatorname{Re}\Gamma_n^- \ge 0, \quad \operatorname{Im}\Gamma_n^- \ge 0$ are vertical propagation constants of outgoing harmonics in upper (index "+") and lower (index "–") half-spaces, respectively. The constitutive parameters ε and μ represent vacuum above the interface and a medium below it. If the beam's parameters k and β set $\Phi_0 = \zeta 2\pi/l = k/\beta$, then the excitation $H_x^i(y,z,k) = \exp\left(-i\Gamma_0^+ y\right)\varphi_0(z)$ is an inhomogeneous ($\operatorname{Re}\Gamma_0^+ = 0, \quad \operatorname{Im}\Gamma_0^+ > 0$) plane electromagnetic wave associated with the field of the electron beam (the wave analog concept detailed in [4], [5], [17], [22]). The time factor $\exp(-i\omega t)$, $\omega = kc$ and all dimensions (the SI system is used) are omitted. A detailed discussion of the problem (1) could be found in [5], [22].

Solving the problem (1) for the unknown $R_n(k)$ and $T_n(k)$ (utilizing MAR [5], [20]) enables computation of the intensity of diffraction radiation at upward and downward outgoing harmonics, i.e., the efficiency of energy extraction from the electron beam into the *n*-th propagating ($\operatorname{Im}\Gamma_n^{\pm}=0$) harmonic [5], [22]:

$$\begin{aligned} W_n^+(k) &= \left|R_n(k)\right|^2 \frac{\operatorname{Re}\Gamma_n^+}{\operatorname{Im}\Gamma_0^+} \\ W_n^-(k) &= \varepsilon^{-1}(k)\left|T_n(k)\right|^2 \frac{\operatorname{Re}\Gamma_n^-}{\operatorname{Im}\Gamma_0^+}. \end{aligned} \tag{2}$$

B. Planar Vacuum-Medium Interface

Now, let us consider the case of a planar interface between vacuum and a medium as a limit of a periodic interface when the corrugation depth $h \to 0$, Fig. 1b.

When the interface is a plane (the profile function $y = f(z) \equiv 0$), the problem (1) without excitation ($H_x^i \equiv 0$) reduces to an infinite set of independent homogeneous systems of linear algebraic equations [23]

$$\begin{cases} R_n = T_n \\ R_n\Gamma_n^+ = -T_n\Gamma_n^-\varepsilon^{-1}(k) \end{cases}; \quad n = 0, \pm 1, \pm 2, \ldots \tag{3}$$

with respect to unknown R_n and T_n. The system (3) is obtained from (1) by matching the tangential components of the field $\vec{E}, \vec{H}$ in the plane $y = 0$.

For each $k > 0$, non-trivial solutions of (3) determine an infinite set of identical eigenwaves of a planar structure [23]

$$\begin{aligned} &H_x(y,z,k) = \begin{cases} l^{-1/2} R_n \exp\left[i\left(\Gamma_n^+ y + \Phi_n z\right)\right], & y > 0 \\ l^{-1/2} T_n \exp\left[i\left(-\Gamma_n^- y + \Phi_n z\right)\right], & y < 0 \end{cases}; \\ &\Gamma_n^+ = \sqrt{k^2 - \Phi_n^2}, \quad \Gamma_n^- = \sqrt{k^2\varepsilon(k)\mu(k) - \Phi_n^2}; \\ &\Phi_n = \pm k\sqrt{\frac{\varepsilon(k)\left(\mu(k) - \varepsilon(k)\right)}{1 - \varepsilon^2(k)}}, \quad n = 0, \pm 1, \pm 2, \ldots \end{aligned} \tag{4}$$

There is no dependence on *n* in (4), the planar interface is not transforming. This means that an electron beam flying over a planar interface excites only principal (zeroth, $n = 0$) harmonic in the upper and lower half-spaces.

III. Analytical Results for Cherenkov Radiation

In the upper half-space, the only harmonic arising there is an evanescent one ($\operatorname{Re}\Gamma_0^+ = 0$ and $\operatorname{Im}\Gamma_0^+ > 0$ as $\Gamma_0^+ = \sqrt{k^2 - k^2/\beta^2}$ for any $k > 0$ and $0 < \beta < 1$), hence there is no radiation into the upper half-space.

In the lower half-space, diffraction radiation in the form of Cherenkov radiation is possible only if $\operatorname{Im}\Gamma_0^- = 0$. By the definition, $\Gamma_0^- = \sqrt{k^2\varepsilon(k)\mu(k) - \Phi_0^2}$ and $\Phi_0 = k/\beta$. Thus, the necessary condition for the existence of radiation in the lower half-space, $\operatorname{Im}\Gamma_0^- = 0$, is fulfilled when $\varepsilon(k)\mu(k) > \beta^{-2}$. It can be satisfied only when $\varepsilon > 0$ and $\mu > 0$ (conventional medium) or when $\varepsilon < 0$ and $\mu < 0$ (double-negative medium). The curve

$$\beta = \frac{1}{\sqrt{\varepsilon(k)\mu(k)}} \tag{5}$$

bounds the domain (within the range of *k* and β variation) where the intensity of Cherenkov radiation in the lower half-space, W_0^- (see (2)), is nonzero. Now, it is possible to obtain analytically an upper bound for the intensity of radiation into the lower half-space.

When $h = 0$, from the problem (1) and the definition (2) follows the explicit expression for the radiation intensity:

$$W_0^-(k,\beta) = \frac{4\sqrt{\varepsilon(k)\mu(k)\beta^2 - 1}\left|\varepsilon(k)\right|\sqrt{1-\beta^2}}{\varepsilon(k)\mu(k)\beta^2 - 1 + \varepsilon^2(k)\left(1-\beta^2\right)}. \tag{6}$$

Introducing

$$a = \sqrt{\varepsilon\mu\beta^2 - 1} \quad \text{and} \quad b = \left|\varepsilon\right|\sqrt{1-\beta^2}, \tag{7}$$

the relation (6) is rewritten as

$$W_0^-(k,\beta) = \frac{4ab}{a^2 + b^2}. \tag{8}$$

The denominator $a^2 + b^2$ could be written as $(a-b)^2 + 2ab$. As $(a-b)^2$ is nonnegative, then

$$\frac{2ab}{a^2 + b^2} = \frac{2ab}{(a-b)^2 + 2ab} \le 1. \tag{9}$$

From (8) and (9) follows that

$$W_0^-(k,\beta) \le 2. \tag{10}$$

It is easy to see that the maximum radiation intensity, $W_0^-(k,\beta) = 2$, is achieved when $a = b$. Going back to the originals from (7), this is equivalent to $\sqrt{\varepsilon(k)\mu(k)\beta^2 - 1} = \left|\varepsilon(k)\right|\sqrt{1-\beta^2}$. Raising both sides to the second power and solving for β yields

$$\beta = \sqrt{\frac{\varepsilon(k)^2 + 1}{\varepsilon(k)\mu(k) + \varepsilon(k)^2}}. \tag{11}$$

For k, β satisfying (11), Cherenkov radiation in the lower half-space, $W_0^-(k,\beta)$, reaches its maximum.

The obtained analytical results, namely, the border of the domain of existence of radiation (5), the upper bound on its

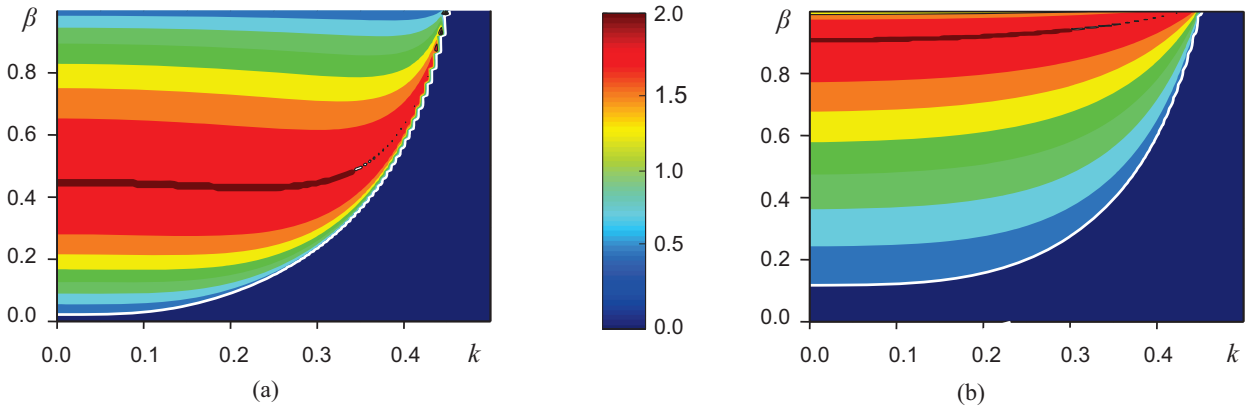


Fig. 2. Intensity of radiation W_0^- into the lower half-space filled with a plasma-like medium as a function of the beam parameters k and β. (a) $k_\varepsilon = 0.5$, $k_\mu = 1$; (b) $k_\varepsilon = 1$, $k_\mu = 0.5$. The curve (5) is marked white.

intensity (10), and the criterion for achieving maximum radiation (11), are of both theoretical and practical significance. These findings advance the fundamental understanding of diffraction radiation mechanisms and enable precise identification of the parameters that yield the highest possible radiation levels, which is particularly valuable for the design of high-efficiency devices, and also offer a stringent benchmark for the verification and validation of numerical solvers. This latter aspect is illustrated by the numerical results presented next.

IV. Numerical Validation

For the numerical analysis, we employ a versatile and reliable solver designed for the general case of a periodically corrugated interface. It implements one of the modern MAR variants [5], [22]–[25]. While the current study focuses on the case of a planar interface (i.e., zero corrugation depth), the use of the general solver ensures that the numerical results are obtained within a consistent and fully general framework.

First, we consider a plasma-like medium, which was extensively analyzed in [23], [24]. It fills the lower half-space ($y < 0$) and its constitutive parameters are set as

$$\varepsilon(k) = 1 - k_\varepsilon^2 / k^2 \quad \text{and} \quad \mu(k) = 1 - k_\mu^2 / k^2 . \qquad (12)$$

Here, $k_\varepsilon > 0$ and $k_\mu > 0$ are characteristic frequencies. Fig. 2 plots the radiation intensity $W_0^-(k, \beta)$ as a function of the excitation beam parameters k and β for two different sets of characteristic frequencies. It is clearly seen that the radiation intensity does not exceed the analytically derived limit (10), and the radiation exists and reaches its maximum precisely within the predicted ranges of k and β variations (see (5) and (11)).

Second, we consider a well-known metamaterial [26]. Once again it fills the lower half-space and its constitutive parameters are set as

$$\varepsilon(k) = 1 - k_\varepsilon^2 / k^2 \quad \text{and} \quad \mu(k) = 1 - \theta k^2 / \left(k^2 - k_\mu^2\right). \qquad (13)$$

Here, $k_\varepsilon > 0$ and $k_\mu > 0$ are characteristic frequencies and $0 < \theta < 1$. Fig. 3 plots the radiation intensity $W_0^-(k, \beta)$ as a function of k and β for different values of characteristic frequencies. For this simulation, $\theta = 0.56$. Again, the radiation intensity remains within the theoretical limit (10), and the radiation is generated and reaches its maximum within the ranges of k and β set by (5) and (11), respectively.

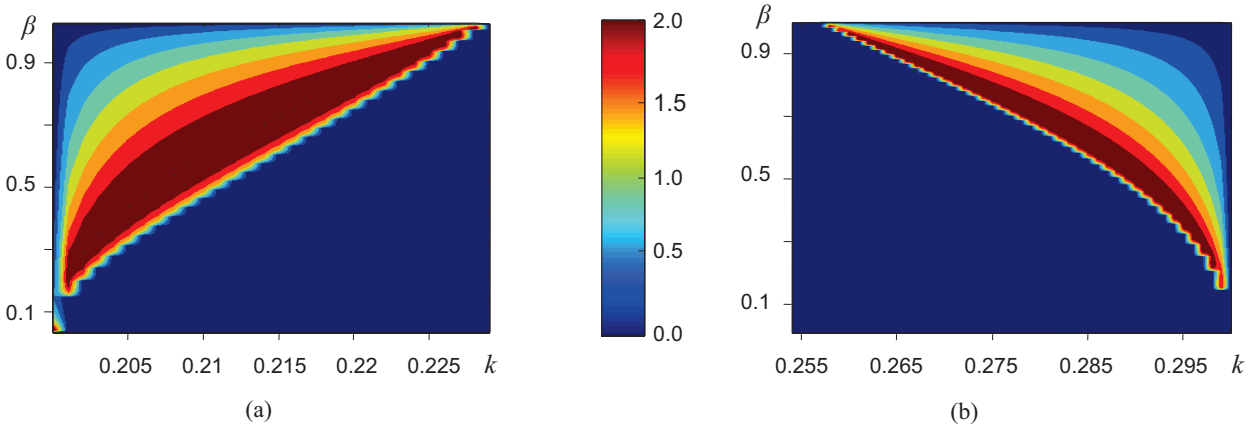


Fig. 3. Intensity of radiation W_0^- into the lower half-space filled with metamaterial as a function of the beam parameters k and β. (a) $k_\varepsilon = 0.3$, $k_\mu = 0.2$; (b) $k_\varepsilon = 0.2$, $k_\mu = 0.3$.

V. Conclusion

In this study, we have presented an analytically derived upper bound for the intensity of Cherenkov radiation generated by a plane, density-modulated electron beam interacting with a planar vacuum-medium interface. This result represents a rare instance of an exact solution to a nontrivial electromagnetic problem. It provides theoretical insight into the mechanisms underlying diffraction radiation and establishes conditions under which such radiation exists and reaches its maximum intensity. These findings are not only of fundamental interest, but also have practical relevance for the design and optimization of high-efficiency devices. Furthermore, the derived expressions could serve as an essential and rigorous benchmark for the validation and verification of numerical solvers, as confirmed by the accompanying simulation results.

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Intelligent Technologies of Remote Identification of Dynamic Objects

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***Abstract*— The remote identification of dynamic object parameters is a subfield of object detection that focuses on identifying object parameters from a distance using various sensor and imaging methods. The primary objective of object detection is to develop efficient and accurate algorithms for remote parameter identification, which can be used in hazardous environments or situations where direct contact with the object is not possible or undesirable. Object detection is a fundamental task in computer vision with diverse applications such as surveillance, autonomous vehicles, robotics, and smart manufacturing. The main goal of object detection is the precise identification of parameters of detected objects, including their position, size, shape, and orientation. In this research, we provide a comprehensive overview of various techniques and methods for remote object parameter identification to facilitate research and future development in this field. We begin by discussing the breakdown into stages. Then, we briefly survey different approaches to determine object parameters remotely, including feature-based methods, model-based methods, and machine learning-based methods. The machine learning-based approach involves the use of various machine learning algorithms, such as deep learning, to learn object parameters from a large labeled dataset of images. Finally, we describe some recent advancements and future research directions in this field, such as sensor technologies for object detection. Overall, this research highlights the importance of remote object parameter identification and provides valuable insights and resources for researchers and practitioners working in the field of computer vision and object detection.**

Keywords—Remote identification of dynamic objects, Object detection, Neural networks, Computer vision, Methods, Software, Deep learning, System architecture, Machine learning, Artificial intelligence

I. Introduction

Remote Parameter Identification of Dynamic Objects (RPIDO) is a subfield of Object Detection (OD) that focuses on identifying object parameters at a distance using various sensors and images. The primary goal of OD is to develop efficient and accurate algorithms for remote parameter identification, which can be used in hazardous environments or situations where direct contact with the object is impossible or undesirable. OD is a fundamental task in computer vision with diverse applications such as surveillance, autonomous vehicles, robotics, and smart manufacturing. The main objective of OD is to identify the parameters of detected objects, including position, size, shape, and orientation.

This article provides an analytical overview of various techniques and methods for remote parameter identification of objects with the aim of facilitating research and future development in this field. The analysis starts by categorizing the emergence of different technologies into stages. Over the past decades, it has been recognized that the development of OD includes two historical periods: the traditional period (pre-2014) and the deep learning-based period (post-2014). The main criteria for OD technologies are the time of their emergence and their performance [1]. A brief overview of approaches to remote object parameter estimation is provided, including feature-based methods, model-based methods, and machine learning-based methods. The machine learning-based approach involves various learning algorithms, such as deep learning, to estimate object parameters from large labeled image datasets. The transition from traditional OD methods to deep learning-based approaches has led to significant improvements in OD performance and has become dominant in the field. The recent advancements and future research directions in this field are described, including sensor technologies for OD. This research underscores the importance of remote object parameter identification and provides valuable references and resources for researchers and practitioners working in the field of computer vision and OD.

The goal of the article is to provide an overview of OD models for future research purposes and develop models on chosen methods. To achieve this goal, the following tasks were set: Analyze and supplement the roadmap of OD methods (2.A) Traditional OD methods, (2.B) Analyze two-stage, (2.C) Analyze one-stage OD methods, (2.D) Analyze methods that improve OD, (2.E) Analyze methods for estimating speed as a dynamic parameter, (3) Define conditions for the use of identification methods, (4) Describe the prospects of remote identification of dynamic objects, (5) Define case study: identifying parameters of a moving vehicle, (6) Analyze intelligent technologies of remote identification of dynamic objects, (7) Develop models of remote identification of dynamic objects.

II. Overview of Detection Methods

A. Traditional Methods of Object Detection

In the 1990s, early object detection algorithms were built based on handcrafted features. These methods include the Viola-Jones method, histogram of oriented gradients, and deformable part-based models.

In 2001, P. Viola and M. Jones achieved real-time detection of human faces [1–3]. Working on a Pentium III processor with a clock frequency of 700 MHz, the Viola-Jones (VJ). The VJ method used sliding windows to scan through all possible locations and scales on an image, checking if a window contains a human face.

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In 2005, N. Dalal and B. Triggs proposed the Histogram of Oriented Gradients (HOG) method [1, 4]. The HOG method is an improvement over the scale-invariant feature transform (SIFT) method [1, 5] and the shape contexts method [1].

The HOG method has become a crucial foundation for many object detection methods and has found wide applications in the field of computer vision for many years [1, 6-7].

The Deformable Parts Model (DPM) method, proposed by P. Felzenszwalb in 2008, is an extension of the HOG method [1, 8]. It follows the "divide and conquer" philosophy. The DPM method was further expanded by R. Girshick to work with real-time object detection and handle larger variations [1, 6, 7–9].

These methods were interesting and had the potential for development, modern object detection techniques based on deep learning are more effective due to their ability to automatically learn discriminative features from large datasets.

B. Two-stage Methods

In 2012, there was a resurgence of Convolutional Neural Networks (CNN) [1, 10]. As CNN can learn robust and high-level image features, the method of Region-based CNN (R-CNN) was proposed for object detection in 2014 [1, 11, 12].

In R-CNN, the first step involves generating a set of object proposals, which are candidate regions for object detection [1, 13].

Each proposal is then scaled to a fixed-size image and passed through a pre-trained CNN model to extract features [1, 10].Finally, linear Support Vector Machines (SVMs), known as SVM classifiers, are used to predict the presence of an object in each region and determine the object class. R-CNN significantly improves results on the sample of visual object classes [1, 14].

In 2014, spatial pyramid pooling networks (SPPNet) [1, 15] were proposed. Previous CNN models required input images of fixed-size, for example, AlexNet used images of size 224 by 224 [1, 10].

In 2015, R. Girshick proposed the Fast R-CNN [1, 16], which is an improvement over R-CNN and SPPNet [1, 11, 15].

In 2015, the concept of Faster R-CNN was introduced, which operates at nearly real-time speeds [1, 17, 18]. Faster R-CNN incorporates region proposal networks (RPN), enabling efficient proposal generation. In 2016, J. Dai proposed Region-based Fully Convolutional Network (R-FCN) was proposed [1, 20]. And in 2017 Light head R-CNN have been proposed [1, 21].

In 2017, Feature Pyramid Networks (FPN) were introduced [1, 22]. The use of FPN in the base system of Faster R-CNN achieved state-of-the-art object detection results using a single model. FPN has become a fundamental building block in many subsequent methods [1].

Two-stage methods evolved at an unprecedented pace, improved object detection by employing a top-down architecture and lateral connections to construct high-level semantics at all scales.

C. One-stage Methods

One-stage methods based on CNN can detect all objects in a single step. They are popular among mobile devices with real-time functionality and ease of deployment, but their performance noticeably decreases when detecting densely packed and small objects.

In 2015, the "You Only Look Once" (YOLO) method was proposed [23]. YOLO is characterized by a decrease in localization accuracy compared to two-stage methods, especially for certain small objects. Later YOLO modifications have been proposed [24-27].

In 2015, the Single Shot MultiBox Detector (SSD) method was also proposed [28]. In 2017, RetinaNet, a type of one-stage detector, was proposed [1, 29]. In 2018, H. Law et al proposed the CornerNet as an alternative approach to object detection [1, 30]. In 2019, X. Zhou et al. proposed the CenterNet method [1, 31], which also leverages key points (in CornerNet [1, 30], ExtremeNet [1, 32], etc), but eliminates the computationally expensive post-processing steps such as key point grouping (as in CornerNet) or non-maximum suppression (NMS).

In 2020, N. Carion et al. introduced the DEtection TRansformer (DETR) for object detection, where they viewed object detection as a set prediction task and proposed a detection network based on transformers [1, 33]. Later, X. Zhu et al. proposed Deformable DETR [34] to address the challenges of long convergence time in DETR and limited performance in detecting small objects, which achieved state-of-the-art results [1].

D. Improved Hybrid Methods

In 2017, Hu et al. proposed Relation Networks (RN) for OD: an adapted attention module and an object relation module that considers the interactions among different objects in an image, including their distinctive features and geometric information [34, 35].

In 2018, Dai et al. proposed Deformable Convolutional Networks (DCN). Deformable Regions of Interests Pooling (DRoIP) can also adapt to the spatial locations of object parts with different shapes. DCNv1 achieved significant improvement in accuracy, while DCNv2 employed more deformable convolutional layers than DCNv1 [35, 36].

In 2019 Neural Architecture Search for Feature Pyramid Networks (NAS-FPN) the authors from Google Brain conducted neural architecture search to discover a new architecture called NAS-FPN for feature pyramid generation. Adding additional pyramid networks, increasing the dimensionality of features, and using high-capacity architectures significantly enhance detection accuracy [35].

DCMv2 used more deformable layers and feature mimicking to improve the deformable effect and accuracy in object detection [35, 36].

NAS-FPN is an approach that combines neural architecture search with the Feature Pyramid Network (FPN) framework. NAS-FPN can generate customized FPN architectures that adapt to different datasets and object scales, leading to improved detection performance [35].

For improving the speed recognition of detection can be enhanced by learning the three-dimensional shape of an object [38].

E. Motion Estimation-Based Identification

In 1911, the proportional-integral-derivative (PID) controller was developed and used for object motion or velocity estimation. This controller consists of three components: proportional, integral, and derivative. The PID controller effectively estimates the motion or velocity of the object and aims to minimize the error between the desired value and the actual state of the object [39].

In 1981, the block-matching (BM) method was developed and used for motion estimation in video sequences. In 1981, the Lucas-Kanade optical flow method was developed and used pixel intensities to determine the optical flow and is based on approximating the local intensity difference equation using the least squares method. The Horn-Schunck optical flow method (1981) also uses pixel intensities and smoothness of the optical flow. In 2015, the pixel recursion methods involve analyzing individual pixels in an image. In 2003, the Farneback optical flow method was developed and used polynomial approximation to determine the optical flow. In 2015, the dense optical flow method was developed and uses a dense grid of points to estimate the optical flow across the image and provides motion information in different regions of the image [39, 40].

III. Conditions for the Use of Identification Methods

The choice of method depends on specific application constraints. Below are key criteria to consider on Table 1.

TABLE I. Conditions for the Use of Identification Methods

Criterion	Description	Methods Best Suited	Criterion
Accuracy	Precision in object localization and classification	Two-stage methods (e.g., Faster R-CNN, FPN)	Accuracy
Speed	Real-time performance requirements	One-stage methods (e.g., YOLO, SSD)	Speed
Hardware Constraints	CPU/GPU limitations in embedded systems	Traditional methods, YOLO-lite	Hardware Constraints
Object Size	Detecting small vs. large objects	FPN, CenterNet, Deformable DETR	Object Size
Scene Dynamics	Motion, occlusion, environmenta l variability	Optical Flow, PID, Farneback, DCNv2	Scene Dynamics
Training Data Availability	Limited labeled data vs. large datasets	Transfer learning with CNNs, handcrafted methods for few-shot	Training Data Availability

This section helps practitioners choose the most appropriate method based on practical considerations.

IV. The Prospects of Remote Identification of Dynamic Objects

RPIDO can be applied to environmental monitoring, medical diagnostics, robotics, components of video surveillance systems and security systems, detecting suspicious activities, tracking motion, and real-time identification of individuals or objects, the gaming and sports domains. It allows for measuring speed, trajectory, and other motion characteristics of objects in real-time [40-42].

V. Case Study: Identifying Parameters of a Moving Vehicle

To demonstrate the application of the approaches discussed, consider the following scenario:

A. Problem Statement

Task: Identify the type, speed, and trajectory of a moving vehicle in a monitored area.

Constraints:

Hardware: Edge device with embedded GPU (e.g., Jetson Nano).

Accuracy requirement: 90% for detection, ±5 km/h for speed.

Time: Real-time processing (< 100 ms/frame).

Environment: Urban setting, partial occlusions, changing light.

B. Method Selection

Model presented for selected method on Fig. 1.

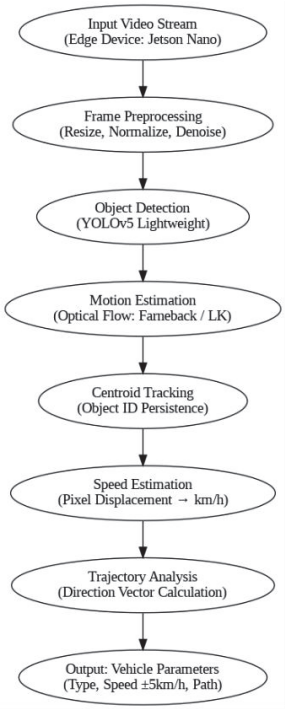


Fig. 1. ML-based Ensemble for Remote Identification.

Given these conditions. YOLOv5 with optical flow refinement (Farneback or Lucas-Kanade) offers a good balance between speed and moderate accuracy. For more robust tracking in occluded scenes, Deformable DETR or DCNv2 with FPN could be used, assuming better hardware availability. Speed estimation is handled by dense optical flow combined with centroid tracking. Implementation Outline. Frame capture and preprocessing. Object detection using YOLOv5. Motion estimation with Farneback optical flow. Data fusion to extract speed and direction vectors. This case illustrates how practical constraints guide the selection of methods for real-world deployment. This model begins with a live video stream captured from an edge device such as the NVIDIA Jetson Nano. The input frames are preprocessed to adjust resolution, normalize pixel values, and reduce noise for better detection accuracy. Next, object detection is performed

using a lightweight version of YOLOv5, optimized for real-time inference on constrained hardware. For motion estimation, the system applies dense optical flow algorithms like Farneback or Lucas-Kanade to track movement between frames. To maintain identity and trajectory of objects across frames, centroid tracking is used. Speed estimation is computed by translating pixel displacement into real-world velocity (e.g., km/h), assuming known camera calibration. Finally, the trajectory of the vehicle is analyzed by calculating direction vectors over time. The combined output includes the type of vehicle, its speed with an accuracy tolerance of ±5 km/h, and its predicted path—all processed in under 100 ms per frame.

VI. Intelligents Technologies of Remote Identification of Dynamic Objects

Recent studies have introduced advanced intelligent technologies for the remote identification of dynamic objects. These approaches emphasize high-accuracy classification, robustness under uncertainty, and ensemble methods. A major contribution is the integration of classification performance assessment techniques such as ROC analysis and multidimensional confusion matrices [43].

Chelak et al. [43] present a comprehensive framework that utilizes these tools to better evaluate and improve the performance of classifiers used in dynamic identification tasks. Their methodology provides a more nuanced understanding of classification boundaries and misclassification patterns in multidimensional spaces, which is vital for complex systems. Additionally, the use of fuzzy decision trees has been proposed for adaptive classification under uncertainty. This method allows the modeling of imprecise and vague knowledge often present in dynamic object monitoring scenarios [44].

Another set of approaches revolves around ensemble learning. In particular, bagging classifiers have been shown to be effective in identifying states of computer systems, which are analogous in complexity and variability to dynamic objects [45].

These ensemble-based models provide enhanced stability and generalization capabilities, which are crucial in environments with fluctuating input parameters or incomplete data. Hornostal and Gavrylenko [46] further explored the use of heterogeneous ensembles in system state identification problems, offering increased resilience and diversity in classification models applicable to dynamic object tracking tasks.

For object detection in large-scale spatial contexts, Khudov et al. [47] developed a method based on the firefly algorithm, which demonstrates effectiveness on satellite imagery and can inspire improvements in aerial or remote vehicle tracking systems.

Additionally, Gorokhovatskyi and Yakovleva [48] presented the use of medoids for packing ORB image descriptors, supporting fast and memory-efficient object representation—an important factor in embedded systems and constrained devices.

Filatov et al. [49] proposed an image classifier architecture tailored for fast retrieval in large datasets, which can aid in reference-based identification or archival analysis of dynamic objects. These intelligent technologies can be considered a part of the future research and development agenda for systems that require not only high accuracy, but also adaptability, explainability, and robustness in the remote identification of dynamic objects for Machine Learning (ML).

VII. Models for Intelligents Technologies

A. ML-based Ensemble for Remote Identification

On Fig. 2 presented model ML-based Ensemble for Remote Identification. The first model, ML-based Ensemble for Remote Identification, is a comprehensive machine learning architecture designed to extract and classify information from sensor data to identify dynamic objects remotely.

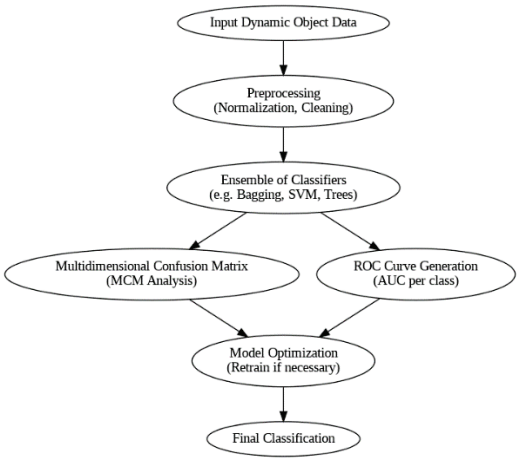


Fig. 2. ML-based Ensemble for Remote Identification.

It begins with input from sensors such as cameras, radar, or lidar. These raw signals undergo preprocessing steps, including frame extraction and normalization, to prepare the data for analysis. Feature extraction is then performed using machine learning techniques, particularly convolutional neural networks and other deep learning models, which are effective at detecting and encoding spatial and temporal patterns. These features are passed into an ensemble of classifiers that utilize bagging and boosting to improve generalization and accuracy. The model's outputs are evaluated using tools such as multidimensional confusion matrices and ROC curve analysis. Based on this evaluation, the system can refine its detection strategy by tuning internal parameters. Ultimately, the model supports the remote identification of an object's position, shape, speed, and other characteristics.

B. Fuzzy Decision Trees and Bagging

On Fig. 3 presented model ML-based Ensemble for Remote Identification. The second model focuses on a hybrid approach that integrates fuzzy logic with ensemble learning, titled Fuzzy Decision Trees and Bagging. It also starts with sensor input but applies fuzzy logic to manage uncertainty and imprecision in the dynamic environment. The system defines fuzzy rules through membership functions that express degrees of truth rather than binary classifications. These fuzzy rules guide the construction of decision trees trained on random subsamples of the data. Instead of relying on a single model, this approach aggregates the outputs of multiple fuzzy trees using bagging, which reduces variance and increases stability.

The final stage of the model enables remote parameter identification, estimating properties such as the state, speed, and orientation of the observed object.

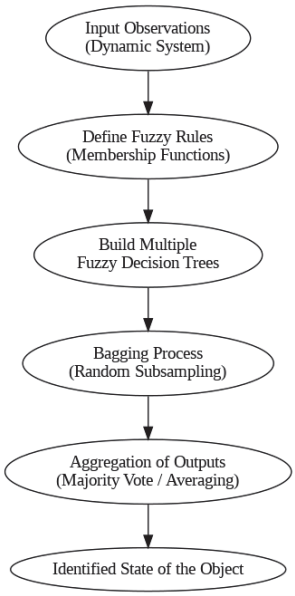


Fig. 3. Fuzzy Decision Trees and Bagging.

VIII. Conclusion

Over the past decades, object detection has undergone two significant historical periods: the traditional object detection period (before 2014) and the deep learning-based object detection period (after 2014). The traditional period was characterized using handcrafted features and shallow learning algorithms for object detection. In the deep learning-based detection period, there has been a rise in the deep neural networks that autonomously learn features and perform end-to-end object detection.

During the traditional period, several important methods emerged, including the VJ method in 2001, DPM in 2005, and the Faster R-CNN method in 2014, which was the last method before the deep learning-based object detection period. The Viola-Jones method employed Haar features and adaptive boosting algorithm for real-time face detection, while DPM relied on a pictorial structure model and used a mixture of parts for object detection. The Faster R-CNN was the first method to use multi-region pooling and achieved state-of-the-art results on various benchmark datasets for object detection. The Faster R-CNN introduced RPN and achieved state-of-the-art performance on various object detection benchmarks.

In the deep learning-based object detection period, several important methods based on deep neural networks have been notable, including YOLO in 2016, SSD in 2016, and RetinaNet in 2017. SSD was one of the first deep learning-based methods that used a single network for both object localization and classification. YOLO used a single neural network to predict bounding boxes and class probabilities directly from full images, while RetinaNet introduced a novel focal loss function to address the class imbalance issue in object detection.

Remote identification of dynamic objects can benefit from a combination of techniques such as the PID controller, block-matching method, and pixel recursion algorithms. These approaches enable the estimation of object motion or velocity, motion detection in video sequences, and analysis of individual pixels for changes in position or value. Additionally, optical flow estimation methods like Lucas-Kanade, Horn-Schunck, Farneback, and dense optical flow offer valuable insights into the motion patterns and flow of objects in images or video frames, further enhancing the remote identification of dynamic objects.

In conclusion, the progress in object detection has gone through two significant periods, each characterized by different technologies and methodologies. In the traditional period, reliability relied on handcrafted features and shallow learning algorithms, whereas in the deep learning-based detection period, deep neural networks that autonomously learn features and perform end-to-end object detection were used. Both periods have witnessed seminal methods, with their priority and timing of appearance serving as key indicators highlighting the underlying technologies. Remote parameter identification of dynamic objects pertains to the process of identifying and tracking the motion and behavior of objects in real-time. This process relies on object detection, which involves determining the presence and location of objects in an image or video frame.

The model based on chosen methods demonstrates an effective fusion of modern object detection and motion estimation techniques tailored to real-time remote identification on resource-constrained devices. By leveraging the efficiency of YOLOv5 and the precision of optical flow methods like Farneback and Lucas-Kanade, the system achieves a practical compromise between speed and accuracy. Centroid tracking further enhances object persistence across frames, while speed and trajectory estimation provide actionable parameters for dynamic object monitoring. The modularity of the approach allows for scalability: under enhanced hardware conditions, more advanced detectors like Deformable DETR or DCNv2 can be integrated without altering the core workflow. Ultimately, this case study highlights how theoretical advances in object detection and motion analysis translate into robust, real-world implementations, particularly for applications involving remote identification of moving objects in dynamic and partially occluded environments.

While the ML-based ensemble model emphasizes deep feature extraction and precision classification, the fuzzy tree model prioritizes interpretability and adaptability in uncertain and variable conditions. Both aim to enhance the robustness and accuracy of remote identification in complex environments.

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Electrodynamic characteristics of ceramic materials research for the creation of protective avionics elements

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***Abstract*—The paper presents the results of experimental studies of the electrodynamic characteristics of ceramic materials obtained on the basis of the target phases of the RO – Al_2O_3 – SiO_2 (RO – SrO, BaO, ZnO). The main attention is paid to the analysis of the transmission, reflection and absorption coefficients for slawsonite, celsian, celsian-slawsonite and slawsonite-willemite ceramics in the frequency range of 26.5…38 GHz. It is established that the studied ceramic materials demonstrate the stability of electrodynamic parameters in a given frequency range and provide the required level of radio transparency. The possibilities of using the studied materials in the designs of aviation and rocket technology in the manufacture of individual structural and protective elements are assessed. The feasibility of using heterophase ceramics of slawsonite-willemite composition for the manufacture of products with optimal dielectric properties, increased heat resistance and operating temperature up to 1100 °C is established. The results obtained can be used in the development of new types of radio-transparent protective structures for avionics.**

Keywords — radiotransparent materials, celsian, slawsonite, willemite, two-stage ceramic technology, electrodynamic characteristics, microwave range, complex profile products, waveguides

I. Introduction

The development of modern aviation and space systems places strict demands on materials used to create avionics protection elements, in particular radar fairings and protective screens. One of the key tasks is to ensure high radio transparency of materials in a wide frequency range while maintaining their mechanical strength, thermal resistance and chemical inertness. Materials and devices based on them must provide resistance to harsh environmental conditions, which significantly narrows the range of materials available for this purpose. The properties of microwave ceramics are determined by a number of factors, including the composition, purity of the initial components, technological processing conditions, features of the crystal structure, as well as the final density or porosity of the material. That is why researchers are particularly interested in ceramic materials that combine these properties and are able to work in extreme conditions [1-3].

For radiolucent ceramics, the dielectric constant (ε) and the dielectric loss tangent (tgδ) are regulated parameters that must meet established standards and technical requirements. These characteristics determine the degree of transparency of the material for radio waves and its ability to effectively transmit electromagnetic waves with minimal losses. To protect radio-controlled equipment elements, radio-transparent materials are used, which are structural dielectrics with a single-layer or multilayer structure. They are capable of transmitting electromagnetic waves in the microwave range with minimal losses and distortions, while the level of radio wave reflection should not exceed approximately 1%. In addition, additional parameters can be taken into account to determine the suitability of radiolucent ceramics, including wave impedance, reflection, transmission and attenuation coefficients of electromagnetic waves [3].

In the modern production of radiolucent materials, various composite materials are used, in particular fiberglass, fiberglass, aluminophosphorus ceramic plastics. Fused quartz, sitals based on slawsonite, spodumene or celsian, as well as ceramic materials with BeO, ZrO_2, Al_2O_3, BN, SiAlON and Si_3N_4, which are part of the main matrix, are also used. Radiotransparent ceramics, in particular based on aluminum oxides (Al_2O_3), silicon nitrides (Si_3N_4), magnesium oxides

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(MgO), as well as their composites, are characterized by low dielectric losses, stability of electrodynamic parameters at high temperatures and resistance to aggressive environments. This makes such materials promising for use in the manufacture of antenna fairings, aircraft waveguides, and other protective elements, where it is important to minimize the impact of the material on the passage of electromagnetic radiation [4-7].

Glass-crystalline materials, mainly formed by the phases of slawsonite and celsian, have high functional and operational characteristics, which makes them suitable for the manufacture of protective components of electronic equipment. Such elements provide reliable protection of ground-controlled radio equipment, affect the aerodynamic properties of aviation objects and the accuracy of their control, and are also characterized by a long service life. At the same time, the technological process of producing these materials is extremely energy-intensive, which significantly increases the cost of final products.

High-alumina ceramics have significant advantages over other materials due to their increased refractoriness, which is a key factor when used in the design of antenna fairings for high-speed missile systems. The phase transition temperature of high-alumina ceramics is about 2050 °C, while for sitals this indicator is in the range of 1200…1350 °C, and for ceramics based on quartz glass - does not exceed 1300 °C [8]. At the same time, under the action of a one-sided short-term thermal load, the performance of the product is determined by a set of factors, among which the phase transition temperature is only one of the parameters. In addition to ensuring radio transparency, such materials must have high mechanical strength to counteract aerodynamic loads, as well as withstand significant temperature fluctuations that accompany flight at high speeds [5-8]. The difficulty lies in finding the optimal ratio between the electrodynamic characteristics (dielectric permittivity, dielectric loss tangent) and the physical and mechanical properties of the material. Also, most existing ceramic materials based on quartz, alumina, nitrides and similar compounds are accompanied by a number of limitations due to the need for multi-stage high-temperature processing during their manufacture [9].

Modern development of technologies and innovations requires the creation of new materials with clearly defined properties. Oxides of elements, which were previously used mainly in laboratory research, are now gaining important practical importance. For effective work in this area, it is necessary to deeply understand the physicochemical features of the technology of refractory non-metallic and silicate materials. One of the key aspects is the study of equilibrium phase diagrams of the state of oxide systems and the professional use of the information they contain. The development of new functional ceramic materials, which includes the synthesis of materials with a given phase composition at reduced temperatures using energy-efficient technologies, should be based on knowledge of the subsolidus structure of basic oxide systems. This knowledge allows controlling the processes of phase formation and identifying stable phase combinations [2].

Promising systems for the creation of ceramic radiotransparent materials are systems based on RO – Al_2O_3 – SiO_2 (where RO – SrO, BaO, ZnO), in which controlled synthesis of such target crystalline phases as willemite, slawsonite and celsian is possible [10-12]. These phases are characterized by low dielectric losses (ε – 5…8, tgδ ⩽ (1…50)·10^{-4}), high melting point (over 1650…1750 °C), low coefficient of linear thermal expansion (α ⩽ 3.2·10^{-6} K^{-1}) and significant mechanical strength (E = 110…115 GPa).

In [13], the properties of materials synthesized on the basis of the BaO – ZnO – SiO_2 – B_2O_3 and BaO – B_2O_3 – Al_2O_3 – SiO_2 systems were investigated. To prepare the slurry, the starting components were weighed and ground together with deionized water for 12 hours at a mass ratio of water : solid phase = 4:1. The resulting slurry was dried at 80 °C, after which preliminary sintering was carried out at 700 °C with an isothermal holding time of 3 hours. Test samples were formed from sintered powders with a particle size of 0.5…1.0 μm under a specific pressure of 25 MPa. The dielectric characteristics were determined at a frequency of 1 MHz using a Precision LCR Meter (Agilent 4284A, USA). Hexacelsian ceramics demonstrated the following properties: TCLE – (8…9)·10^{-6} K^{-1}, dielectric constant – 7.0…7.4, density – 2.9…3.1 g/cm³, bending strength – 70…90 MPa.

In their studies, American scientists [14] found that the dielectric characteristics of the phases of celsian, slavsonite and their glass composition in a ratio of 3:1 remain stable in a wide temperature range from 400 to 1400 °C. In addition, even at maximum operating temperatures, the deviation of the dielectric constant is no more than 7...11%.

Materials structurally formed mainly from these phases are characterized by high indicators of functional and operational characteristics, which ensures their suitability for use as protective elements in radio systems of the aviation industry and ground control complexes.

Therefore, scientific and technological progress in the field of radio-controlled missile technology is largely determined by the creation of radio-transparent materials (RTM), capable of providing reliable protection and trouble-free operation of radio-electronic equipment in harsh operating conditions.

The purpose of the work

The purpose of this work is to study the electrodynamic characteristics of samples of radio-transparent ceramics of various compositions in order to assess their suitability for use as protective elements of avionics.

II. RESEARCH METHODS

To study the structure of the samples, the method of scanning electron microscopy was used on a PHENOM Pro scanning electron microscope (Thermo Fisher Scientific Inc., USA). During the research, an acceleration voltage of 15 kV and small beam currents were used in the mode of secondary electron flow. To compare the sizes and habits of the structural components of the crystalline phases and assess the homogeneity of the samples at the microlevel, photographs were taken at the same magnifications, which allows for detailed analysis.

To determine the qualitative phase composition of the experimental samples, the X-ray diffraction (XRD) method was used using DRON-3M and SHIMADZU XRD-6000 diffractometers with CuK_α-radiation and a nickel filter under standard conditions of use. Phase identification was carried out using a standard ASTM X-ray diffraction chart.

The electrodynamic properties of the materials were investigated in the millimeter range of microwave frequencies within 26.5…38 GHz. The complex scattering parameters of the composite samples were determined using a Keysight PNA N5227A vector analyzer of microwave circuits. Measurements were performed for samples in the form of a rectangular parallelepiped, which were located in a

hollow rectangular waveguide WR 28. The geometric dimensions of the opposite faces of the samples corresponded to the cross-section of the waveguide (7.112×3.556 mm²), taking into account permissible deviations. The thickness of the samples was 3…4 mm, which was due to the technological features of manufacturing and the need to ensure the proper level of interaction of the material with electromagnetic radiation.

III. PRESENTING MAIN MATERIAL

A. Obtaining samples using two-stage ceramic technology

The process of manufacturing ceramic products consists of several stages, during which the material undergoes various physicochemical transformations. As a result, a gradual transformation of the dispersed system occurs, in which the solid phase is represented by the mineral components of the ceramic mass. The technology of manufacturing fine ceramic products includes the following main stages: preparation of raw materials, grinding of the charge, mixing, forming briquettes for synthesis, drying, firing and, if necessary, additional processing. Depending on the type of products, these stages can be omitted, changed in sequence or combined with each other.

The experimental samples were manufactured using optimized technological parameters of the two-stage ceramic process for forming complex-profile radiolucent products, proposed by the authors [15]. At the first stage, the synthesis of target crystalline phases was carried out, and at the second stage, the firing of finished products. For the research, compositions based on single-phase ceramics were selected: stoichiometric slawsonite (S-CM), stoichiometric celsian (C-CM), mixed celsian-slawsonite (CS-CM) in the phase ratio $BaAl_2Si_2O_8$: $SrAl_2Si_2O_8$ = 1:3, as well as heterophase slawsonite-willemite (SW-CM) ceramics in the phase ratio $SrAl_2Si_2O_8$: Zn_2SiO_4 = 1:0.8.

To prepare the charge based on single-phase compositions, crushed quartz, technical alumina of the G-00 brand, as well as strontium carbonate and/or barium carbonate were used in proportions that ensure the stoichiometric composition of the synthesized phases. The formation of briquettes for the subsequent synthesis of the target compounds was carried out by the method of semi-dry pressing on a manual hydraulic press under a specific pressure of 20 MPa. To moisten the charge before pressing, a 10% aqueous solution of carboxymethylcellulose was used, the amount of which was selected to ensure the humidity of the press powder at the level of 7–8%. The synthesis of the target compounds of celsian, slawsonite or their solid solution of barium-strontium anorthite was carried out at a temperature of 1250 °C with a 2-hour isothermal holding.

To obtain samples of heterophase slawsonite-willemite ceramics, zinc whites, metallurgical alumina of the G-00 brand, strontium carbonate (chemically pure) and quartz were used as raw materials (pure quartz sand with a SiO_2 content of not less than 99.5% can be used). The formation of briquettes for the synthesis of target phases was carried out using a method similar to that given above. To intensify the phase formation process, it is advisable to introduce potassium feldspar into the composition of the charge in an amount of 4 wt. % (more than 100% on a dry basis), which contributes to the simultaneous synthesis of target phases of slawsonite and willemite at a reduced temperature of 1150°C with isothermal holding for 2 hours.

The synthesized materials were ground to a particle size of no more than 3 mm, after which they were loaded into a planetary mill for wet grinding using distilled water and a complex additive to obtain a slip with a humidity of about 30%. Grinding was carried out until the residue on sieve № 0063 was no more than 1%. Complex-profile products were formed by slip casting in plaster molds. After the casting reached the required strength, it was removed from the mold and dried to a final humidity not exceeding 1%.

The final products were fired in a muffle furnace at a temperature of 1350 °C with a holding time of 4 hours for single-phase ceramic compositions and at a temperature of 1200°C with a holding time of 2 hours for heterophase slawsonite-willemite ceramics.

B. Structural and phase features of the obtained ceramics

The structural and phase features of the developed ceramics were investigated using X-ray diffraction and SEM methods.

The results of qualitative X-ray diffraction, under the selected technological parameters, showed the presence of only the target phases: for slawsonite - $SrAl_2Si_2O_8$, for celsian - α-$BaAl_2Si_2O_8$, for celsian-slawsonite ceramics - a solid solution of $Ba_{0.25}Sr_{0.75}Al_2Si_2O_8$, which indicates the completeness of the reactions of formation of the target phase.

In SEM images of samples of the S-CM series, fired at a temperature of 1350 °C with a holding time of 4 h, the structure of the samples is densely sintered and more homogeneous. The size of individual pores does not exceed 1-2 μm. Slawsonite crystals acquire clear shapes.

In micrographs of samples of the C-CM series, obtained under similar technological parameters, with a magnification of ×5000, it is possible to observe individual grains of celsian with a size of 2 to 5 μm, which are in contact with each other, creating a dense, uniform fine-crystalline structure.

According to scanning electron microscopy, the microstructure of CS-CM ceramics is uniform and quite dense. At a temperature of 1350 °C, active formation of solid solution crystals occurs, the composition of which is identified by X-ray diffraction as $Ba_{0.25}Sr_{0.75}Al_2Si_2O_8$.

Data on the micromorphology of single-phase and heterophase ceramics obtained under optimal technological parameters indicate the implementation of directed solid-phase synthesis of target compounds (celsian, slawsonite, willemite and their combinations) provided that their stoichiometric composition is preserved. The exception is the single-phase ceramics of the CS-CM series, the phase composition of which is represented by a solid solution of celsian and slawsonite.

C. Determination of electrodynamic characteristics of ceramic materials

The parameters of the interaction of electromagnetic radiation with ceramic samples were determined using a vector analyzer of microwave circuits Keysight PNA N5227A in the frequency range of 26.5…38 GHz. For the convenience of comparative analysis of the electrodynamic characteristics of ceramic materials, the obtained values of the amplitude coefficients, given in dB, were converted into the corresponding coefficients, presented in fractions of a unit. Such conversion was carried out on the basis of the balance of the ERP (Electromagnetic Radiation Power) ($P_{ref} + P_{trans} + P_{abs} = P_{in}$) and the balance of the amplitude coefficients ($(K_{ref})^2 + (K_{trans})^2 + (K_{abs})^2 = 1$, where K_{ref} is Reflection Coefficient; K_{trans}

is Transmission Coefficient; K_{abs} is Absorption Coefficient. The results of the conversion are presented in Table 1.

TABLE I. ELECTRODYNAMIC CHARACTERISTICS OF THE CONSIDERED CERAMIC MATERIALS

Ceramics cipher	Frequency, GHz	Coefficient					
		$R_{trans(a)}$, dB	$R_{ref(a)}$, dB	$R_{abs(a)}$, dB	R_{trans}, times	R_{ref}, times	R_{abs}, times
S-CM	26.5	–4.7	–2.4	–10.7	0.34	0.58	0.09
	38	–2.5	–6.2	–7.1	0.56	0.24	0.20
C-CM	26.5	–4.8	–2.6	–9.2	0.33	0.55	0.12
	38	–2.9	–6.4	–5.9	0.51	0.23	0.26
CS-CM	26.5	–5.6	–1.8	–11.9	0.28	0.66	0.06
	38	–1.9	–5.9	–10.1	0.65	0.26	0.10
SW-CM	26.5	–5.8	–1.8	–11.2	0.26	0.66	0.08
	38	–2.3	–6.3	–7.5	0.59	0.23	0.18

From the obtained data, we see that with increasing frequency there is a general tendency to increase the transmission coefficient (K_{trans}) and decrease the reflection coefficient (K_{ref}), which indicates an improvement in the bandwidth of materials in the millimeter range. The most significant increase in the transmission coefficient was noted for CS-CM and SW-CM samples: at a frequency of 38 GHz, $R_{transmitted}$ was 0.65 and 0.59, respectively, with a simultaneous decrease in $R_{reflect}$ to 0.26 and 0.23.

The obtained results demonstrate the prospects of using the studied ceramic compositions for the manufacture of radio-transparent structures with controllable electrodynamic properties depending on the operating frequency range. In particular, CS-CM type materials are advisable to use in conditions where a combination of significant absorption and moderate transmission at lower frequencies is important, while for operation at higher frequencies the SW-CM composition is most effective due to its better transmission characteristics and reduced reflection.

CONCLUSIONS

Experimental studies of the electrodynamic properties of the developed ceramic materials of the S-CM, C-CM, CS-CM and SW-CM compositions in the frequency range of 26.5…38 GHz were conducted. It was established that all the developed materials meet the basic requirements for radio-transparent ceramics and are characterized by a transmission coefficient $K_{trans\,(a)}$, which is from –5.8 to –1.9 dB (equivalent to K_{trans} = 0.3…0.7). The studied characteristics demonstrate stability in the frequency range of 26.5–38 GHz, which is an important condition for the effective functioning of radio-transparent ceramic materials as protective elements of avionics.

It was established that according to the values of the above characteristics, the studied materials can be classified as radio-transparent ceramics and are suitable for use in the aerospace industry in the manufacture of individual structural elements. It is worth noting that the operating temperature of products based on slawsonite-willemite ceramics is limited to 1100 °C, however, their use in a number of areas of aircraft and rocketry, in particular as waveguides, seems promising given the set of technical characteristics and economic feasibility.

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Effect Decreasing Overheating in Hybrid 2D- 3D Heterostructure GaN- h-BN Oscillator Diode

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Abstract — Problems of obtaining high frequency generation by GaN-based diode is considered in the paper. The GaN planar structure with length of 1 μm is investigated to estimate its ability to generation under impact ionization and self-heating. Monte Carlo technique and 2D modeling are used to obtain DC and AC characteristics of the diode. The oscillation efficiency and current density characteristics versus bias are given. Influence of temperature effect and impact ionization on diode characteristics have been demonstrated in the course of the research. Estimating maximal temperature, we propose to decrease the temperature by applying heat-escaping top-surface heat spreaders in the form of single layer (SL) of hexagonal boron nitride (h-BN) on the top of the diode. Temperature reduction over 150 K in diode with SL h- BN is shown.

Keywords — temperature, mole fraction, impact ionization, scattering, self-heating, single layer, GaN, electric field strength, thermal conductivity, thermal boundary resistance.

I. Introduction

To obtain high frequencies in solid-state active elements including the terahertz range, we searched possible ways to increase their frequency responsibility. This problem is solved both by naturally reducing the size of the elements, and, therefore, reducing the flight time of charge carriers, as well as using the effective mechanisms to increase their function.

For example, in transferring electron diodes (TED), the main problem is to reduce the time of electron transition from satellite valley of conduction zone to the lowest central valley at conduction band (reverse transitions). This time usually exceeds the time of the direct transition. Practically, the single way to solve this problem is to search for suitable materials with small transition times.

Binary III –nitride compounds, such as a GaN, InN and AlN, and their ternaries along with the quaternary, are considered one of the most possible materials to develop the high-speed electronics including terahertz frequency generation devices. Reverse transition time in GaN is known to be of order 10^{-13} s. That is an order of magnitude less than in GaAs to be traditionally used for Gunn diodes. Moreover, it has a high saturation velocity for electrons (more than 2×10^7 cm/c), and owing to its wide energy bandgap, is also an excellent candidate for device operation at high temperatures.

Theoretical estimations of frequency properties of GaN and other III nitrides, oscillation efficiency demonstrate the ability to obtain generation up to 1 THz and powers up to 10 mW. For example, superiority of GaN as a Gunn diode over those based on materials like GaAs and InP was demonstrated in [1, 2]. Energy characteristics both of GaN diode and their ternaries, including graded band gap diode, were analyzed in [3].

However, until recently, fabrication of GaN-Gunn diodes is still difficult. Typical problems are the high series resistance leading to high DC losses, self-heating effects, and small thermal conductivity of substrates [1], [4]. Generation of fundamental frequency in the range of 0.3 - 0.4 THz by GaN-based diode was experimentally obtained in [5]. It is reported on the fabrication of GaN-based THz Gunn sources representing an active mesa-structure on GaN-substrate. According to the authors, it was possible due to stable side-contacts, good passivation of the active mesa, and good heat sink.

The aim of the study is 2-D simulations of GaN-based planar diode on sapphire substrate to analyze generation of electromagnetic oscillations, considering strong electric fields effects, including heating and impact ionization. Thus, diode structure and substrate were used to obtain maximal heating effect. Hexagonal boron nitride (h-BN) as a factor avoiding formation of localized overheating in hybrid 2D-3D heterostructure diode is considered.

II. The Model

A cross section of GaN-based planar structure considered in the study is shown in Fig.1. In fact, it is n^+–n–n^+ diode on a sapphire substrate, with a width of 0.64 μm and a length of

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1.28 μm. The donor concentration in the *n*-type region corresponded to Kremer criteria for Gunn based on GaN, and was specified in the range from 6×10^{16} cm^{-3} to 2×10^{17} cm^{-3} In the contact n^+-regions is of 0.16 μm×0.32 μm, with donor concentration of 1×10^{19} cm^{-3}. Thus, an active diode length is about 1 μm.

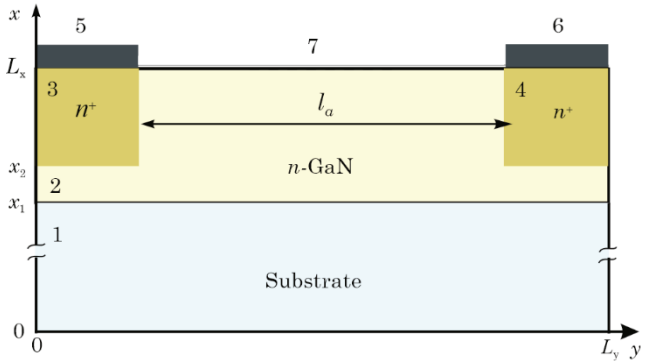


Fig.1. Diode structure: low conductivity substrate (1), *n*-type diode channel (2), high doped contact region (3,4), metal contacts (5,6), h-BN- layer (7)

To accurately evaluate the non-equilibrium electron transport in the diode under self-heating, Ensemble Monte-Carlo technique is implemented, self-consistently coupled with the solution of the Poisson equation and synchronized with the solution of the heat conduction equation. The time step of 5×10^{-16} s was chosen as a compromise to match the high carrier scattering intensity in GaN and the analysis of the diode heating time. Zone parameters and band structure model of GaN, and actual scattering mechanisms for electron and holes are taken into account according to [6, 7].

Realistic approach is a finite-thickness substrate with a fixed temperature at its bottom boundary at $x=0$, accepted as 300 K in the study. The sapphire substrate thickness is taken 1.28 μm, and its length matched the diode length. The adiabatic thermal boundary conditions are applied to diode sides at $y=0$ and $y=L_y$. Evaluating temperature distribution is obtained by solving the heat conduction equation:

$$c\frac{\partial T(r,t)}{\partial t}=-\nabla\left[k(T)\nabla T(r,t)\right]+F(r,t), \qquad (1)$$

where dependence $k(T)$ and the specific heat capacity c were taken according to [8], as well as thermal boundary conductance between interfaces (metal-GaN and GaN-sapphire). Here, ohmic contact is suggested to be gold with a titanium adhesion layer and the TBC at the metal–GaN interface is $2\cdot10^8$ W/(K·m^2) [9]. The TBC at the interface between the sapphire substrate and GaN is 10^7 W/(K·m^2) [10].

The function $F(r,t)$ in equation (1) represents distribution of internal heat sources [11]:

$$F=F_0\sum\left(\hbar\omega_{ems}-\hbar\omega_{abs}+\Delta E_{II}\right), \qquad (2)$$

where $\hbar\omega_{ems}$ and $\hbar\omega_{abs}$ are the energies of phonon emission and absorption, respectively, ΔE_{II} is the energy absorbed by the lattice due to impact ionization, and F_0 is a constant parameter of numerical model and device doping. Increased lattice temperature due to self-heating is incorporated into the temperature dependence of all considered scattering mechanisms and energy gap. The temperature dependence of the energy gap is obtained by the empirical Varshni relation [12,13]. To obtain a solution to equation (1), as well as Poison equitation for electrostatic potential, the FMG method is implemented.

III. Results and Discussion

A. DC and frequency characteristic of the diode

The evaluation of the frequency and energy characteristics of the GaN-based diode indicates a high potential of the GaN-based Gunn diode. [5,14,15]. Estimation made in the study was carried out under assumption of diode operation in single-circuit resonator with corresponding voltage $U(t)=U_0+U_1\sin 2\pi ft$, applied to diode. Here f is resonator frequency; U_1 –amplitude of the first voltage harmonic determined by resonator. Oscillation efficiency is determined as alternative power P to DC power ratio $\eta=P/P_0$. The maximal η is found by optimized choice of U_0 and U_1

The frequency dependences of the optimal efficiency, obtained without taking into account self-heating and impact ionization, are shown in Fig. 2 for different doping concentrations in the active lengths of the diodes, which is done in most publications [14,15].

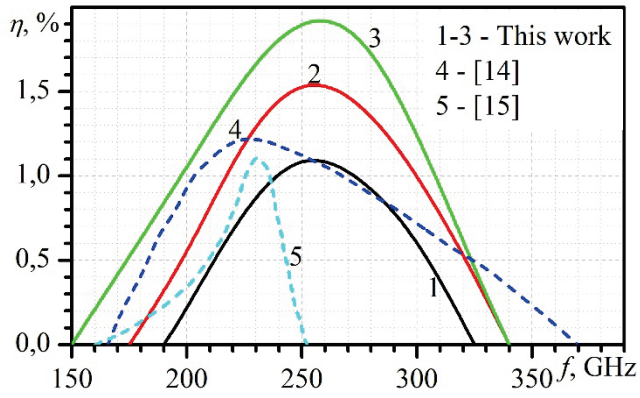


Fig.2. Dependence of maximal efficiency generation on frequency. For diode's with difference doping: 1 - N_d=6×10^{16} cm^{-3}; 2 – 10^{17} cm^{-3}; 3 – 2×10^{17} cm^{-3}.

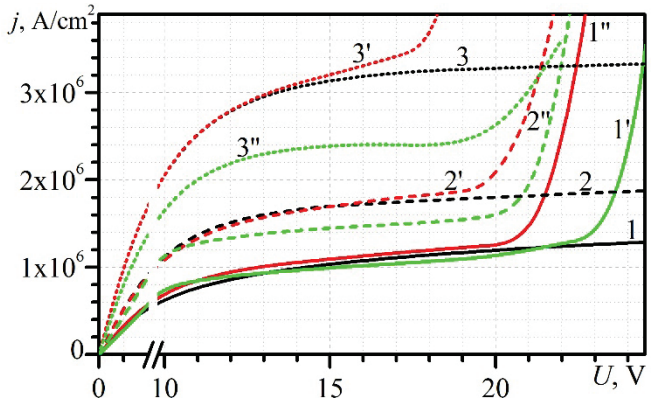


Fig.3. j-V characteristics of diodes: 1, 1',1'' – N_d=6×10^{16} cm^{-3}, 2, 2', 2'' – N_d=10^{17} cm^{-3}, 3, 3', 3'' - N_d=2×10^{17} cm^{-3}: 1-3 – w/o II and w/o self-heating; 1'-3' – with II and w/o self-heating; 1''-3'' – with II, with self-heating.

The maximal oscillation efficiency is about 250 GHz for all doping levels. It corresponds to U_1 =25 V and U_1 =10 V.

EMC simulation results obtained under heating and impact ionization are shown in Fig. 3.

The dependences demonstrate the impact of heating and II on diode operation. Firstly, we can see a narrowing of bias region available to obtain high frequency generation, associated with II in strong electric field at anode typical for n^+–n–n^+ diode. On the other hand, influence of II degree results in self-heating. However, heating is a long-term process having a little effect on formation of dynamic negative resistance in diode, shown in Fig. 4.

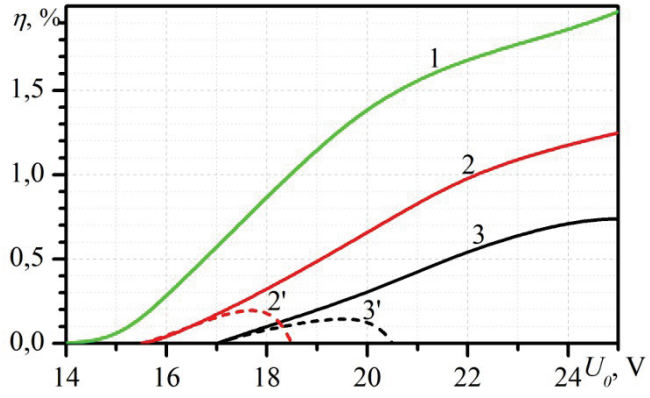


Fig. 4. Dependence of optimized efficiency generation at f=250 GHz for different bias voltage with different doping concentration: 1 – N_d=2×10^{17} cm^{-3}; 2, 2' – N_d=10^{17} cm^{-3}; 3, 3' - N_d=6×10^{16} cm^{-3}; 1-3 – w/o II and self-heating; 2', 3' – with II and self-heating.

Limitations for the generation are connected rather with II than with heating. For this reason, generation at 2×10^{17} cm^{-3} was not found. Negative aspect of heating is significant temperature rising in diode cannel. 2D temperature plot was shown in Fig. 5 for bias 22 V.

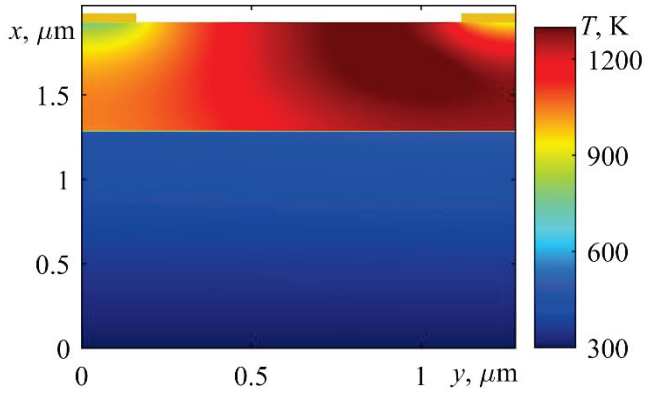


Fig. 5. Temperature distribution in the GaN diode at U=22 V; t_{sim}=100 ns.

As seen from Fig. 5 the maximal temperature exceeds 1200 K to be quoted significantly. The high electric field at anode of GaN diode and the resulting non-uniform distribution of dissipated power led to the formation of hotspots in the device channel. Despite good thermal properties of GaN, this results in decrease of mobility and current carriers through the diode, as seen from Fig.3. Yet, the main problem is the ability to remove heat from internal hot spots rear contacts constraining possible operation regimes of diode and can lead to contact degradation.

B. Characteristics of the diode with h-BN

Local thermal control of device can be substantially improved via formation of additional heat-escaping top-surface heat spreaders. For example, few layer Gr (FLG) films present are shown to be effective heat spreaders with respect to ordinary metals films [16]. Graphene, having a high thermal conductivity, is also considered for this purpose [17]. An advantage is relatively small TBR between interface of those material and various substrates. However, both Cr and graphene are high conductivity materials.

Atomic thin h-BN unlike zero band-gap graphene has wide band-gap, and can been considered the top ranked candidate for the dielectric material for next-generation electronic devices [18].

An in-plane thermal conductivity of one layer (1-L) h-BN were obtained experimentally in [19, 20] at room temperature. Thus, thermal conductivity of 585 W/(m·K) was obtained in [21], Raman-deduced average thermal conductivity of 751 W/m·K was obtained in [19]. The maximal value of thermal conductivity higher than 800 W/(m·K) is reported in [20]. Thus, heat can spread quickly laterally through a large area of one layer (1-L) h-BN, and localized overheating of a device can be avoided [21].

This approach is expected to be effective in heterostructures composed of distinct 2D materials. However, the use of h-BN with 3D-material combination is possible in the so-called hybrid heterostructures with mixed sizes. Such 2D/3D heterostructures are known to propose and practically realize [22], including combination 3D GaN with h-BN.

Simulation planar of GaN-based diode taking into account the thermal transport between diode, 1L h-BN layer, and substrate was carried out in [23]. Proposed numerical thermal model based on multigrid algorithm is applied in the study.

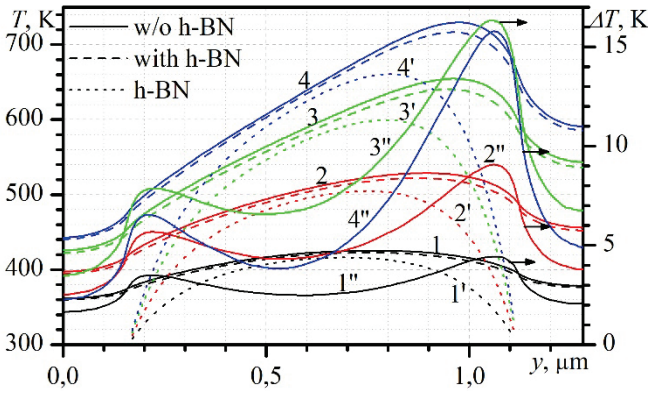


Fig. 6 – Temperature distribution on the top layer of diodes with N_d=10^{17} cm^{-3} at difference bias (1-4 – GaN channel with and without SL h-BN; 1'-4' - SL h-BN; 1''-4'' – temperature difference): 1 – U_0=10V, 2 – U_0=15V; 3 – U_0=20V; 4 – U_0=22 V.

The 1L h-BN layer is placed on the top of the diode, between the metallic contacts with contact between them. TBC at the GaN/1L h-BN interface is 3.5·10^7 W/(K·m^2) [24]. All details concerning modeling of the h-BN monolayer, such as thermal properties and thermal modelling, are discussed in [23].

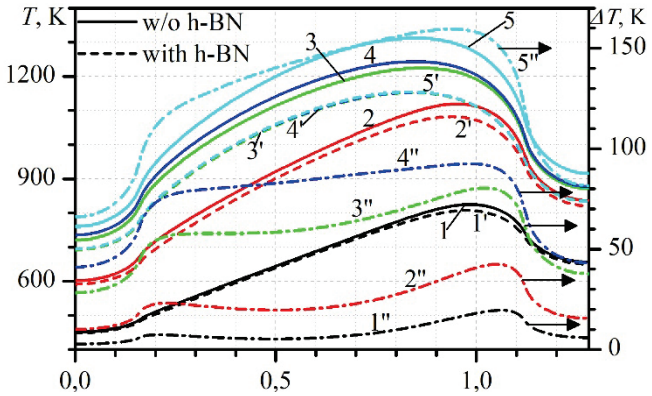


Fig. 7. Temperature distribution on top channel of GaN diode at U=22 V for different time: (1-5 - GaN channel without SL h-BN;1'-5' - GaN channel with SL h-BN; 1''-5'' – difference between temperature with and without SL h-BN) 1,1',1'' - t_{sim}=12.5 ns; 2,2',2'' - t_{sim}=25 ns; 3,3',3'' - t_{sim}=50 ns; 4,4',4'' - t_{sim}=75 ns; 5,5',5'' - t_{sim}=100 ns.

Temperature distribution on top GaN surface, h-BN monolayer formed per simulation time (t_{sim}) of 12 ns are shown in Fig. 6 at difference diode bias. Dependences correspond both to diodes with h-BN and diodes without monolayer. Here, temperature difference between surface temperature for both considered cases is presented. As seen from Fig. 6, introduction of SL h-BN leads to temperature decrease. Those changes depend on crystal local temperature, and are quite weak. However, modelling conducted on a larger time scale shows a more significant possible temperature increase, Fig. 7. This temperature can be 150 degrees lower in diodes with h-BN then without it. However, it can be important for diode contact, especially the anode one. For example, the ohmic Ti/Au contact to GaN can be stable during short-term heat treatment at 900ºC [25]. Most contacts degraded during heating at lower temperatures. Temperature at anode is 683 ºC in diode with SL h-BN, and 803 ºC without one.

IV. Concusion

The study has shown that the problem of obtaining high frequency generation in GaN diode is quite complicated. A number of factors influence this. The factor of electric field is the most significant. Temperature rise due to strong electric field appears because of carrier scattering on phonon as well as due to impact ionization. Temperature is also a reason for decreasing negative differential resistance. Low thermal conductivity of sapphire makes the implementation of heat sinks difficult. This leads to high DC losses. The studies have revealed that possible range of voltage bias is much narrower. The investigations indicated the effect of decreasing temperature due to application of SL h-BN on the top of the diode. The proposed way is most effective when localization of electric field within the device leads to large hot spots near the device surface.

The considered structure is not optimal for obtaining generation. Optimization of diode parameters is possible by replacing sapphire substrates with SiC substrates with a higher thermal conductivity 100–350W m^{-1} K^{-1}. The second way is formation of high resistance cathode contact to protect formation of anode strong electric field domain. The preferred solution is to use Schottky contact as cathode similar to [25]. The use of a planar diode allows for solving the problem of diode cooling, including by removing heat from the upper surface of the heat spreaders.

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Wind Tower with Gravity Energy Storage

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***Abstract*—A wind tower design with a solid-state gravitational energy storage device has been developed for stationary installation as a separate power facility. The features and advantages of the structural and technical solutions of the proposed wind tower, the specifics of its operation and practical application possibilities are considered. A complete breakdown of the wind tower's layout is provided, which will ensure its high technical characteristics and parameters during operation. The analysis of the operating modes of the proposed wind tower was carried out in the form of a mathematical description of its electrical parameters. Factors influencing the design features of the wind tower on the quantitative indicators of electricity generation and storage processes were identified. Reasonable conclusions were made regarding the energy efficiency, reliability, compactness, durability and environmental impact of the proposed wind tower design, on the basis of which further prospects for its implementation in the electric power industry were determined.**

Keywords—power plant, renewable energy source, wind tower, gravity energy storage, vertical axis wind turbine, operating mode

I. INTRODUCTION

The basis for energy independence of consumers is the introduction of power supply systems based on alternative energy sources. Today, renewable energy has a great potential in the world, but only a small part of it is used. However, the energy crisis that is gradually gripping countries is a major impetus for the development of hybrid and autonomous power supply systems that are targeted at different categories of consumers. At the same time, it should be noted that renewable energy is environmentally friendly, i.e. organic with the environment.

From a technical point of view and with the available renewable energy development technologies, a complete transition to autonomous power supply systems is not yet possible, but the world is moving in this direction and new generations of inventions with more advanced technical characteristics appear every day.

It should also be noted that, according to experts, energy efficiency, system reliability and high power quality are important characteristics when developing the latest designs of autonomous power plants. At the same time, the developed structures should be safe in operation, have high maintainability, i.e. be equipped with unified equipment. An additional important characteristic is the ease of deployment of the power plant on site [1].

The gradual transition from traditional to alternative energy is already included in the development programmes of the electric power industry in many countries of the world and it takes into account the active development of autonomous power supply systems [1]-[3]. Programmes for developing the electricity sector with renewable energy focus on policies, financial instruments, and regulatory frameworks to support the growth and integration of renewable sources into the energy mix. These initiatives aim to incentivize investment, simplify permitting processes, and facilitate market integration of renewable energy projects [4].

That is why so much attention is currently paid to the development of power plants and their implementation in both full autonomy and hybrid modes as sources of distributed generation, which is reflected in scientific research [5]-[7] and practical developments [8]-[10].

An important task now is to pursue a policy of rapid transition to alternative energy powered by renewable energy sources. Also, the issue of full autonomy of power supply systems is relevant for Ukraine, which has suffered huge destruction of the generating capacities of the power system and its condition is critical in terms of functioning in a normal mode in compliance with the indicators of electricity quality and reliability of power supply to all consumers in the industrial and domestic sectors [11]-[13].

A significant advantage of power supply systems based on alternative energy sources is a number of issues in the energy sector, such as:

1. full decentralisation of power supply systems and the possibility of organising microgrids with intelligent control;

2. complete energy independence with reliable and high-quality energy supply to electricity consumers;

3. building energy supply systems in areas where connection to the centralised energy system is not possible;

4. the possibility of simultaneous generation of electricity and heat to fully cover the needs of consumers;

5. improvement of the environmental situation.

Thus, the design or improvement of autonomous power plants based on the latest alternative energy and intelligent

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control technologies is an urgent task that requires further development and research.

In this regard, today there are many scientific papers devoted to the development of the design of autonomous power plants [14]-[16] and the study of their operating modes under different operating conditions [17]-[19], analysis and improvement of the energy efficiency of both individual components and power plants as a whole [20]-[22].

It should be noted that the operation of autonomous electric power plants will not be effective for consumers without the use of energy storage devices designed to operate in the modes of duplication of renewable energy sources or increase the system capacity in cases of insufficient generation of electricity due to the climatic conditions of the installation area.

The most developed and researched energy storage technology today is batteries, which have a very extensive classification according to various parameters. However, mechanical energy storage devices also deserve attention, and a large number of scientific papers [23]-[25] have been devoted to their development and research.

The purpose of this article is to develop the design of an electric power plant in the form of a wind tower with a gravitational energy storage system that can operate both autonomously and in parallel with a traditional power grid, and to study and select an algorithm for operation characterised by increased energy efficiency compared to similar plants.

II. Design of a Wind Tower with Gravitational Energy Storage

Today, there is already a large number of design solutions for autonomous power plants using the latest developments based on environmentally friendly alternative energy technologies. Most of these designs are equipped with energy storage and an intelligent control unit for generation, distribution and consumption. In fact, each design is a masterpiece of scientific research. The desire to improve existing designs is the main driver of the development of power supply systems in terms of both structural and functional parameters.

The main factors for the introduction of autonomous power plants for consumers are to ensure reliable power supply and, at the same time, affordability and the lowest possible operating costs.

Almost all existing designs of autonomous power plants contain a standard set of components:

- one or more renewable energy sources that have design differences in terms of the type of conversion of primary energy into electricity;

- energy storage system with the same or different types of storage devices according to the principle of operation;

- additional auxiliary devices that perform the functions of converting and distributing electrical energy (motors, generators, inverters);

- control system for the power plant as a whole and its individual components;

- other accessories and equipment intended for the operation of the power plant and related to its design features.

The authors have developed a wind tower with a gravitational energy storage system [26], which is used as an integral part of an autonomous power plant to supply consumers.

Wind energy has a fairly high energy potential in the world and is already quite developed, and therefore wind turbines are currently being implemented very actively both for industrial consumers and households.

Depending on the design, wind turbines are most suitable for wind speeds of 3-9 m/s. In Ukraine, the average wind speed is 4.5 m/s, and many areas are classified as zones with higher and more constant wind speeds.

All low-speed wind turbines are designed with a vertical axis of rotation. These wind turbines include: straight wing blade rotor (orthogonal), Savonius rotor, helicoidal rotor, Darrieus rotor, sail wing unit, multi-blade unit with guide, plate rotor, cup rotor.

Vertical wind turbines have a number of advantages compared to horizontal wind turbines, namely

- the rotor of this type is technically absolutely safe, virtually silent, and does not generate vibrations and can be installed anywhere according to sanitary standards;

- such wind turbines start at a fairly low wind speed (from 1.5 m/s) and start operating at rated capacity at wind speeds of 3-4 m/s;

- the rotation speed of such wind turbines is absolutely uniform and does not require stabilisation, there are no swirling phenomena;

- has a fairly high resistance to wind loads and can operate at any height.

Taking into account these advantages and the conditions of further operation, the authors decided to use the vertical location of the wind turbine axis. The design of the developed wind tower is shown in Fig. 1 [26].

The design of a wind turbine with a vertical axis of rotation developed by the authors looks like a tower with a gravity storage device in the middle. Accordingly, the taller the structure, the more powerful it will be, firstly, due to the increase in wind speed, which increases with height, and secondly, due to the amount of stored energy, which depends on the length of the load descent path, which is the main component of the storage device.

The developed wind tower structure shown in Fig. 1 consists of a vertical wind turbine 1, a frame housing in the form of a column 2, on which the wind turbine is mounted on angular contact bearings 3, a magnetic disc 4 mechanically connected to the turbine, coils 5 of the turbine generator, centrifugal regulator 6, which is necessary to limit the turbine speed at very high wind speeds, electromechanical coupling 7,

which disconnects the turbine from the storage facility if necessary, magnetic disc 8 and generator coils 9 of the gravity storage facility, centrifugal regulator 10 with a braking system necessary to regulate the discharge rate of the storage device and stop it if necessary, angular gearbox 11, drum 12 for winding the cable 13 for lifting and lowering the massive load 14 of the storage device, blocks of polyspace 15 on which the load is suspended, roller guides 16 and shock absorbing springs 17 necessary to dampen shocks when the load is fully lowered [26].

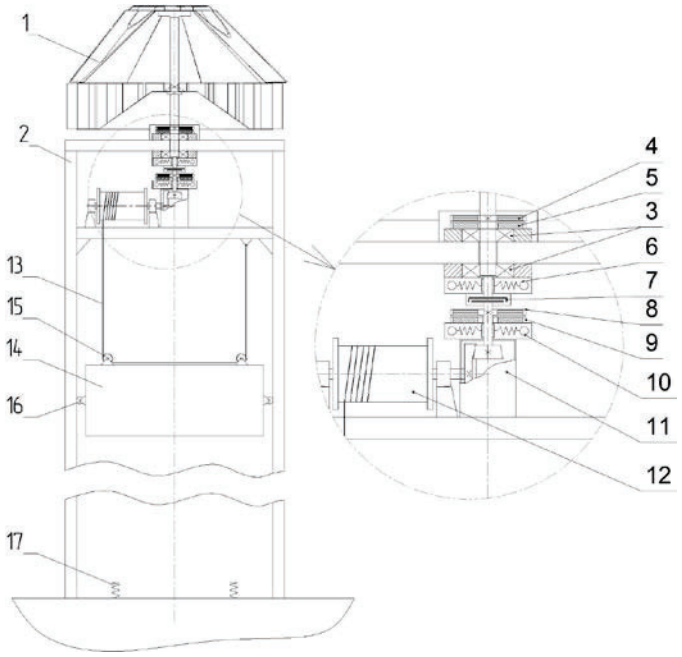


Fig. 1. Design of a wind tower with gravitational energy storage [26].

To increase the stability of the structure, it is additionally equipped with cable ties 18 and screw anchors 19 as shown in Fig. 2.

The system works as follows. In the presence of wind, the turbine 1 starts to rotate, driving a generator consisting of a magnetic disc 3 and coils 4. The generated energy is supplied to the consumer. In the event of excess energy, the electromechanical clutch 6 connects the turbine to the gearbox 10, which has a drum 11 on its low-speed shaft, and the gravity storage device starts charging, i.e. lifting the load 13. At this time, part of the wind energy can be supplied to the consumer in the form of electricity from the turbine generator. If a large amount of electricity is needed for the consumer, the clutch 6 is disengaged and the load starts to descend, rotating the generator consisting of a magnetic disc 7 and coils 8 through the drum 11 and gearbox 10. The energy is supplied to the consumer from two generators at once, satisfying the consumer's need to increase the load. If the consumer no longer needs a large amount of energy, the process of charging the drive starts again. If there is no wind or insufficient wind, the clutch 6 disengages and the brake built into the centrifugal regulator 9 is activated, and the entire system goes into standby mode. Similar actions are also taken when the drive is fully charged. The system is controlled by an electronic system based on a microcontroller, which is powered by a built-in low-capacity rechargeable battery [26].

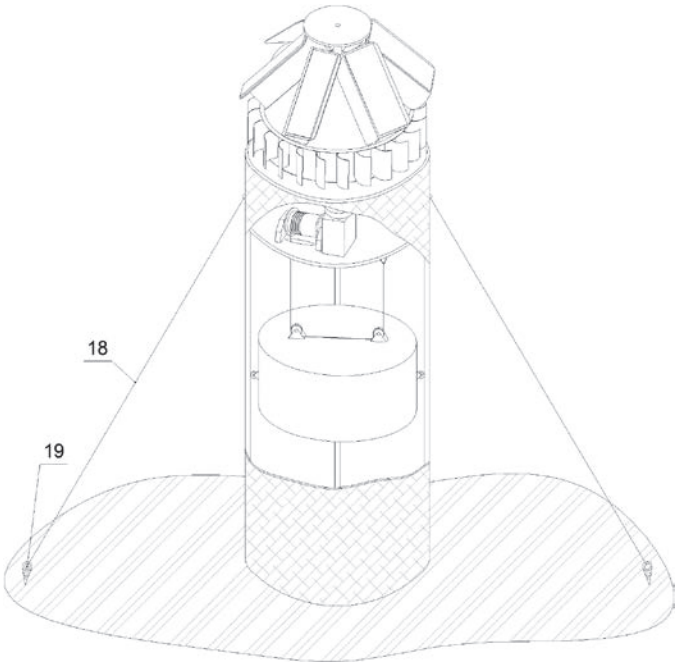


Fig. 2. General view of the installed wind tower [26].

The advantages of the developed wind tower design are:

1. Increased resistance to wind loads due to the presence of a frame-type column with a heavy gravity storage inside, which makes it possible to implement a wind tower with the required height depending on the wind energy potential of the area;

2. the ability to quickly increase capacity to cover even unpredictable peak loads due to the presence of a gravity storage device in the wind turbine design;

3. increased reliability of the wind tower due to the absence of self-discharge in the gravity storage;

4. long service life of the wind tower with virtually no reduction in technical characteristics, as well as a high level of operational maintainability;

5. complete environmental friendliness of the processes of electricity generation and storage.

The presented wind tower design is a fully autonomous energy facility that can also be a source of distributed generation in microgrid systems.

Depending on the size of the manufactured wind tower structure, its power output changes in direct proportion, and thus it can be used both in power supply systems for industrial facilities and in power supply systems for small households.

III. Analysis of Wind Tower Operating Modes

As mentioned in the previous section, the developed wind tower design can operate in both stand-alone and hybrid modes.

When the wind tower operates in stand-alone mode, the total power of the installation must exceed the maximum power of the consumer:

$$P_{mvv} \geq P_{mc}, \quad (1)$$

where P_{mvv} – maximum capacity of the wind tower; P_{mc} – maximum consumer power.

The maximum capacity of a wind tower consists of the power generated by the wind turbine (P_v) and the power stored in the gravity energy storage (P_{ga}):

$$P_{mvv} = P_v + P_{ga}. \quad (2)$$

The amount of electricity that can be generated from a wind tower is determined by the formula:

$$W_{\max v} = W_v + W_{ga} = 0,5\xi\rho A V^3 t \pm mgh\eta_c, \quad (3)$$

where W_v – the amount of electricity generated by the wind turbine; W_{ga} – potential energy of the lifted load; ξ – wind energy utilisation rate; ρ – wind density; A – surface area swept by the rotor; v – wind speed; m – weight of the drive load; g – acceleration of free fall; h – load lifting height; η_c – efficiency of the load lowering and lifting system.

The efficiency of the load lowering and lifting system is determined by the expression:

$$\eta_c = \eta_{cr} \cdot \eta_{cz}, \quad (4)$$

where η_{cr} – efficiency of the gravity storage device during discharge, i.e. during the lowering of the load; η_{cz} – the efficiency of a gravity drive when charging, i.e. when lifting a load.

It should be noted that the '+' sign in formula (3) corresponds to the maximum mode of operation of the wind turbine, in which the gravity storage facility supplies electricity to the consumer's power supply system by lowering the load.

The sign '-' in formula (3) corresponds to the minimum mode of operation of the wind turbine, in which the gravity storage device takes the electrical energy generated by the wind turbine to lift the load, i.e. is charged.

Thus, according to formula (3), it can be concluded that the amount of electricity will be greater the higher the wind tower is, because with increasing height, the wind speed increases, which leads to an increase in wind turbine speed and, accordingly, an increase in electricity generation. In addition, an increase in the height of the wind tower increases the amount of energy that can be obtained from the gravity storage.

When a wind farm operates in a hybrid mode with the grid, the charge cycles of the gravity storage system can be provided by both the wind farm and the grid, and the discharge cycles are driven by the grid's need to cover load peaks.

IV. Investigation of the Efficiency of the Developed Wind Tower

In order to quantify the efficiency of the developed wind tower, we modelled its operating modes when providing power to consumers in autonomous and hybrid modes. This task was solved using the program-calculating complex DigSILENT PowerFactory [27].

The electrical grid scheme and simulation results for one of the calculated operating modes with a wind tower are presented in Fig. 3. The power generation parameters of the developed wind tower were determined in accordance with the given calculation equations and the current climatic conditions of the installation's operation. The authors considered the case of locating the proposed wind tower in the Kharkiv region. The distribution electrical grid to which the proposed wind tower is connected is made up of 0.4 kV cable power lines. The load was represented by a complex composition of domestic and industrial consumers with a total installed capacity of 5.5 kW. Radial and trunk non-redundant schemes are used to supply consumers, as the vast majority of consumers in the electrical grid fall into the third category in terms of power supply reliability.

The quantitative results of modelling the investigated operating modes of the electrical grid under various conditions of power output by a wind turbine with a gravitational energy storage device are presented in Table I.

In order to assess the degree of improvement in the efficiency of the electrical grid under consideration, assuming the installation of the wind tower developed by the authors, a mode without power generation by the wind tower was investigated. This mode is basic for the electrical grid, since under such conditions all consumer needs are met by transporting the necessary amount of electrical energy from the energy system to the investigated distribution electrical grid. The obtained modelling results showed that the use of the developed wind tower, which is connected as close as possible to the place of consumption of the generated electrical energy, increases the efficiency of the electrical grid. This is due to a reduction in power losses during the transmission of electrical energy. For the electrical network under consideration, this indicator is 0.066 kW (35.48%) for autonomous mode and 0.052 kW (27.96%) for hybrid operating mode.

In addition, the analysis of the modelling results has shown that the developed wind tower fully ensures power supply to consumers in all considered operating modes. Moreover, it is important to reduce energy losses in the power grid when generating power from a wind turbine, which allows the installation proposed by the authors to be used as an integral part of an autonomous energy system with improved efficiency.

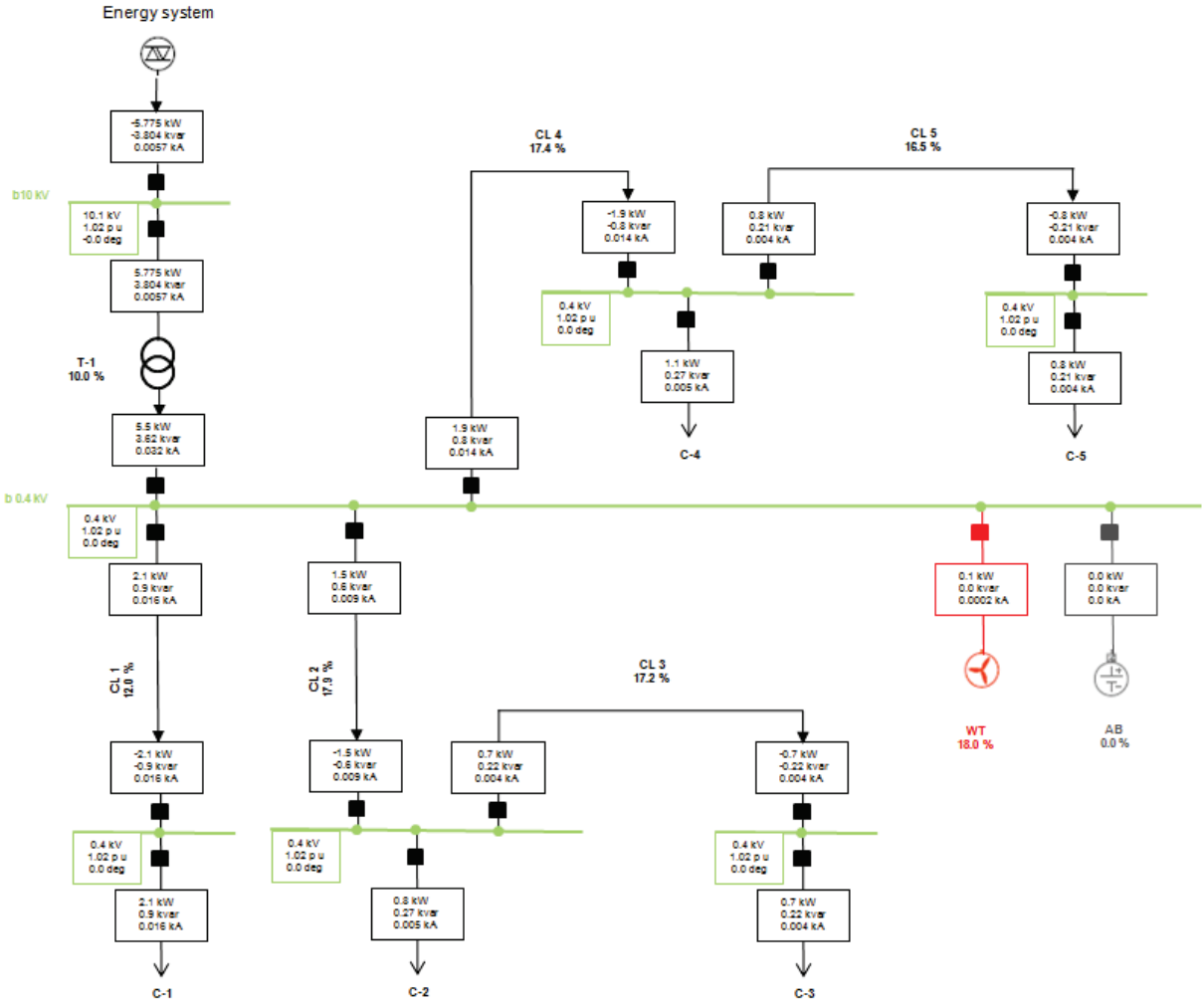


Fig. 3. Calculation of hybrid mode operation of a distribution electrical grid with a wind tower.

TABLE I. RESULTS OF CALCULATING THE MODE OF POWER OUTPUT BY A WIND TOWER WITH A GRAVITATIONAL ENERGY STORAGE

Mode of operation	Consumer power, kW	Wind tower generation capacity, kW		Power consumed from the energy system, kW	Power losses in the electrical grid, kW
		P_v	P_{ga}		
without generation	3	0	0	3	0,186
autonomous	3	3	0	0	0,12
	3	2	1	0	0,125
	1,5	2	-0,5	0	0,14
	0,5	1	-0,5	0	0,147
hybrid	3	2	0	1	0,16
	3	1	0	2	0,17
	1,5	0,5	0,5	0,5	0,153
	0,5	0,2	0,1	0,2	0,142
	0,5	0,3	0	0,2	0,136
	0,5	0,4	0,1	0,2	0,134

Thus, he implementation of the developed wind tower with a gravitational storage system is not only promising but also necessary, as Ukraine's energy sector currently requires a complete overhaul and the application of the latest technologies to build modern power supply systems that can operate in both autonomous and hybrid operating modes.

CONCLUSIONS

The paper proposes a design of a wind tower with a gravitational solid-state energy storage device, which is a modern engineering solution based on alternative energy sources and new environmental technologies for electricity accumulation. The developed design has a number of features and advantages that allow us to draw conclusions about its energy efficiency, reliability, compactness, durability and minimal environmental impact. The implementation of the developed wind tower design is reasonable and relevant today due to the possibility of its use both in autonomous power supply systems and as a source of distributed generation in the latest intelligent microgrid systems.

Modelling of the operating modes of the distribution electrical network with the connection of the developed wind tower has proven that the efficiency of electrical grids increases due to the reduction of losses from the transmission of generated electrical energy to consumers. Thus, the introduction of the proposed wind towers of improved design as sources of distributed generation or autonomous power supply will contribute to improving the efficiency of the power system as a whole.

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Hybrid Radio-absorbing Coatings Using Glass-crystalline Microspheres of Technogenic Origin

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***Abstract* — In the current conditions of rapid radar detection and guidance means development, the importance of measures to reduce military facilities' effective scattering surface is growing. One of the key ways in this area is creating and improving radio-absorbing coatings (RAC), which reduce the electromagnetic radiation reflection in the radio wave range. In modern conditions of rapid radar detection and guidance development, the importance of measures to reduce the effective scattering surface of military objects is increasing. One of the key directions in this field is the creation and improvement of radio-absorbing coatings (RACs), which reduce the reflection of electromagnetic radiation in the radio wave range. The paper discusses the research results on hybrid (polymer matrix) coatings obtained using technogenic microspheres as a filler. The effectiveness of using low-temperature short-term heat treatment of microspheres for targeted crystallization of magnetically active phases (maghemite, hercynite, hematite) is shown.**

Keywords — hybrid coatings, functional fillers, microspheres, radio absorption mechanism, broadband absorption, phase formation, structure

I. Introduction

The main parameter that determines the military equipment visibility when the enemy uses radar surveillance equipment is the effective scattering surface (ESS). This parameter value largely depends on the object's surface shape and the camouflage means presence capable of reducing its visibility. The main parameter that determines the military equipment's visibility when the enemy uses radar surveillance equipment is the effective scattering surface (ESS), the value of which largely depends on the object's surface shape and the presence of camouflage means capable of reducing its visibility. To reduce the visibility of military equipment, a special concealment technology from radar systems (stealth technology) is required [1]. It includes a set of measures, including developing special geometric shapes and radio-absorbing coatings, which lead to a signal power reduction reflected in the radar direction (Fig. 1).

Radio-absorbing coatings (RAC) significantly reduce the probability of military equipment detection by enemy radar reconnaissance means, providing an advantage in combat conditions and increasing the technical capabilities of the army. Given the trends towards automation, miniaturization, and increased sensitivity of radar systems, the issue of effective shielding and radio wave absorption becomes particularly relevant [2].



a)



b)

a) Northrop Grumman B-2 Spirit stealth bomber, Photo Credits: Northrop Grumman/U.S. Air Force [3],
b) Sea Shadow stealth ship, Photo Credits [4]
Fig. 1. Examples of using Stealth technology to conceal military equipment

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RAC is an essential element of stealth technologies that reduce the military equipment's radar visibility, particularly aircraft, drones, missiles, and armored vehicles, for enemy radars. Such coatings reduce the effective scattering area (ESR) by absorbing electromagnetic waves in the radar range. The vast majority of research aimed at creating radar concealment devices is focused on the following areas [3]:

- improving absorption efficiency in a wide frequency range (from 1 GHz to 40 GHz, including millimeter waves for more modern radars);

- reducing the weight and thickness of the RAC, which improves equipment aerodynamic characteristics and maneuverability;

- increasing resistance to external factors: environmental temperature and humidity, mechanical damage, chemical degradation.

The materials science development and innovations in this area open up new opportunities for creating hybrid composite materials with specified electromagnetic properties that can be adapted to a wide range of frequencies and operating conditions. Therefore, creating effective, lightweight, technological, and durable coatings capable of effectively absorbing electromagnetic radiation in the radio frequency range is a pressing task [4]. Solving this problem will contribute to developing electronic camouflage, ensuring the viability of military equipment, and increasing the state's defense capability.

In our opinion, no less important is the possibility of simultaneously implementing other coatings functions, such as the ability to self-heal, heat absorption, and adaptive masking in the visible spectrum range, as well as meeting the conditions of their production, environmental friendliness, and cost-effectiveness.

In this work, we will not touch on the issues of creating radio-absorbing nanomaterials [5], metamaterials [6] and RAC, capable of adapting to the enemy radars frequencies thanks to "smart" materials [7], as well as the use of quantum materials with ultra-high absorption efficiency [8]. Their effectiveness is, of course, significant, but the specificity is too high. We focused on reducing the negative impact on the aircraft's aerodynamics by reducing the weight of the radio-absorbing coating, increasing its temperature stability (which is important for hypersonic vehicles), and enabling broadband absorption (from L- to Ka-band).

Analysis of the latest developments in this materials science field has shown that most of the RAC belong to the class of composites based on a polymer or ceramic matrix with various filler combinations (dielectric, magnetic, electrically conductive). Graphite, carbon, metal, or metalized polymer fibers, iron oxides, ferrites, carbonyl iron, etc., have been used as fillers for RAC [9-11].

In this sense, a promising direction for reducing radar signature is to use silicate microspheres as fillers in a polymer matrix of hybrid RAC. As a rule, microspheres have a diameter of 1 to 150 μm and a low density (especially hollow ones) [12]. It allows the creation of lightweight coatings that are essential for aviation equipment. Silicate microspheres' properties depend on their composition, which is why they are divided into two types: glass and ceramic microspheres. Both glass and ceramic microspheres can be solid or hollow. Hollow microspheres have high thermal insulation properties, essential for protection against infrared sensors used to detect and visualize thermal radiation of objects.

Glass microspheres, made of borosilicate glass (for example, 3M Glass Bubbles), are characterized by a low density (0.1–0.6 g/cm^3) [12]. Ceramic microspheres, made of aluminum and silicon oxides, are also resistant to extreme temperatures (over 1000 °C) and temperature fluctuations. This will allow the obtaining of composite materials, which are ideal for operation in high-speed flight conditions of hypersonic aircraft. In addition, such microspheres' high chemical and abrasive resistance will contribute to a longer service life of radio-transparent coatings of structural external elements [13].

The authors [10] report on the prospects of using ceramic microspheres to obtain radio-transparent materials capable of transmitting radio waves without significant absorption or reflection. Instead, ferrite or carbon microspheres are used as fillers to manufacture radio-absorbing composite materials and coatings, which absorb electromagnetic waves, excluding their reflection [9]. It can be assumed that microspheres can effectively scatter or absorb radio waves due to ash microspheres' dielectric properties and the possibility of filling with magnetic or conductive particles. Their use in combination with magnetic fillers will allow for the expansion of the radio waves' absorption range.

It should be noted that the producing technology for artificial silicate microspheres is complex and energy-intensive, which determines their high cost as a commercial product. Therefore, the study of the use of technogenic origin microspheres as an alternative is of considerable interest.

Ash microspheres are formed during the pulverized coal solid fuel combustion at thermal power plants and are part of the ash-slag mixture, waste that accumulates in ash dumps. Therefore, the phase composition of ash microspheres depends on the coal type burned at the thermal power plants. Thus, the ash microspheres of the Kryvyi Rih thermal power plant, which were used in the work, were obtained due to the anthracite combustion of the Donetsk coal basin.

Information on the composition and properties of Ukrainian thermal power plants' microspheres is provided in works [14–15]. Finding ways to adjust their composition and properties will allow expanding the scope of use of this valuable technogenic product. Research on using ash microspheres in producing radio-absorbing composites is essential to creating cheaper means of effectively masking ground, aviation, and naval military equipment from radar surveillance.

The work aims to investigate the effect of microspheres extracted from ash from thermal power plants on the radio-absorbing ability of the hybrid coatings.

The research plan provided for:

- Determination of the hollow ash microspheres composition and properties,

- Possibility investigation of preserving glass-crystalline microspheres by magnetically-active phases targeted crystallization in the ash microspheres shell,

- Comparative analysis of hybrid coatings that contain glassy, glassy-crystalline microspheres as a filler and a combination of glassy microspheres and ferrite.

II. Materials and Research Methods

The ash microspheres of the Kryvyi Rih TPP were studied as a research object. Microspheres' chemical and granulometric compositions and structural and phase features were determined.

The phase composition of the ash microspheres was determined by the X-ray phase analysis method using a DRON-3M diffractometer with CuKα radiation and a nickel filter for standard operating conditions. The morphological features were studied using a PHENOM Pro scanning electron microscope in the secondary electron flow mode (input voltage 15 kV, probe current 20 mA).

To determine the dielectric characteristics of polymer matrix composites based on epoxy resin, samples with a diameter of 30 mm and a thickness of 3 mm were manufactured. The dielectric permittivity and dielectric loss tangent were determined on an RLC E7-8 parameter-measuring device. The property indicators value was defined as the arithmetic mean of 3÷5 parallel studies, the difference of which did not exceed 5%.

To determine the magnetic permeability, samples were made in the form of tablets with a thickness of 3 mm according to the cross-section of the waveguide. The measurements were carried out by the waveguide transmission/reflection method using an Anritsu VNA vector network analyzer in the frequency range of 0.01–20 GHz. According to the measurement results, the complex values of the dielectric and magnetic permeability of the samples were determined, with which the reflection coefficient RL was calculated:

$$RL = 2 - \log\left|\frac{Z_{in} - Z_0}{Z_{in} + Z_0}\right|$$

де Z_{in} – is the impedance of the absorber layer, $Z_0 \approx 377\ \Omega$ is the impedance of free space.

III. Discussion of the Results

A. Determination of the hollow ash microspheres composition and properties

As in most other cases, the main components of microspheres extracted from the ash dumps of the Kryvyi Rih TPP are silicon, iron, and aluminum oxides. According to the average sample chemical analysis results, they include the following components, wt.%: 45.0 SiO_2, 21.17 Al_2O_3, 26.32 Fe_2O_3, 2.52 MgO, 2.54 CaO, 2.0 K_2O+Na_2O, 0.25 TiO_2, 0.20 MnO. Using the obtained data, the microspheres' properties were determined by calculation with the methods [16]: softening temperature 1460 °C, proper density $\rho = 2458$ kg/m^3, thermal conductivity $\lambda = 0.0657$ W/m·K, linear coefficient of thermal expansion $\alpha = 3.93 \cdot 10^{-6}$ 1/K, bulk weight 400 kg/m^3, mechanical strength 7.5–10.0 МПа.

The ash microspheres have predominant sizes of 100–160 μm and are practically X-ray amorphous. X-ray phase analysis established the glassy nature of the shell with signs of vitrification on the inner surface (Fig. 2).

Considering the ash microspheres' chemical composition, an assumption was made regarding the possibility of changing their phase composition by forming iron-containing compounds during heat treatment. It was subsequently found that ash microspheres heated for 30 min at a temperature of 850 °C cause crystallization of the initially X-ray amorphous material.

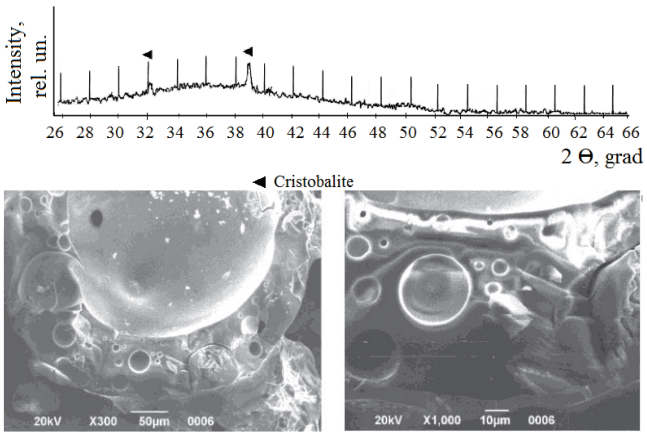


Fig. 2. Structural and phase features of ash microspheres

B. Hollow glass-crystal microspheres: obtaining and study of phase composition

According to the results of X-ray diffraction (fig. 3) in the samples after heat treatment, were identified mainly the phases of Hercynite $FeAl_2O_4$ (d = 0.1899; 0.2470; 0.2037; 0.1171; 0.1577 nm), and Maghemit γ-Fe_3O_4 (d = 0.2949; 0.2519; 0.1607; 0.1480 nm); Hematite α-Fe_2O_3 (d = 0.3703; 0.2686; 0.2519; 0.2217; 0.1841; 0.16593; 0.1484; 0.1449 nm). The sample also contained phases: Mullite $Al_6Si_2O_{13}$ (d = 0.543; 0.3437; 0.3398; 0.2899; 0.2217; 0.1606 nm), Spinel $MgAl_2O_4$ (d = 0.2855; 0.2843; 0.2436; 0.2017; 0.1550; 0.1424 nm) and α-Quartz SiO_2 (d = 0.3348; 0.2458; 0.2662 nm).

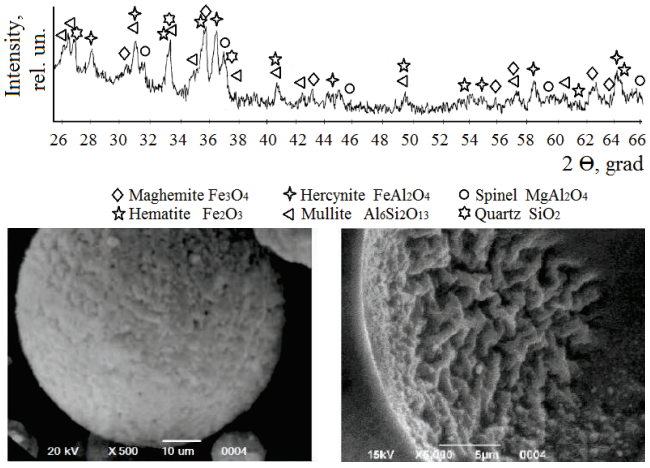


Fig. 3. Structural and phase features of ash microspheres after heat treatment (850 °C)

Table 1 presents the characteristics of the identified phases according to Landolt–Börnstein Numerical Data and Functional Relationships in Science and Technology. As can be seen from the presented data, maghemite is ferromagnetic, while hematite and hercynite have weak magnetic properties. However, maghemite and hercynite remain the most effective for both types of absorption (ε" and μ"), even at frequencies below 1 MHz [17]. Hematite can be advantageous due to its very high ε_r. In contrast, mullite, spinel, and quartz maintain consistently low ε_r and μ_r values. These phases will be useful for polymer matrix structural stabilization or the formation of gradient layers (for example, to reduce the ε_r/μ_r reflection coefficient at the "air – coating" interface).

TABLE 1. PROPERTIES OF CRYSTALLINE PHASES IDENTIFIED IN ASH SPHERES AFTER THEIR HEAT TREATMENT AT 950 °C

Crystalline phase	Melting temperature, °C	Mohs Hardness	Dielectric Permeability	Magnetic Permeability
			при f=1 кГц	
Hercynite* $FeO{\cdot}Al_2O_3$	1440	8.0	25–45	20–60
Maghemite γ–Fe_3O_4	–**	6.0	40–70	50–100
Hematite α–Fe_2O_3	1565	6.5	100–200	1.1–1.5
Mullite* $Al_6Si_2O_{13}$	1810	7,0	6.5-8.0	□ 1.0
Spinel* $MgO{\cdot}Al_2O_3$	2135	8,0	8-10	□ 1.0
Quartz α–SiO_2	1713-1728	7,0	3.8-4.5	□ 1.0
*prone to isomorphic substitution; ** above 400 °C it transforms into hematite				

It should be noted that the magnetically active phases (hercynite, maghemite, and hematite) have different mechanisms of radio wave absorption. Thus, hercynite and maghemite absorb radio waves' electromagnetic energy due to magnetic losses and magnetohysteresis, and hematite – due to the mechanism of high-frequency polarization. These phases also differ in the absorption frequency range, which is 2–12 GHz for hercynite, 0.5– 8.0 GHz for maghemite, and 8– 18 GHz for hematite [18].

Therefore, heat treatment of ash microspheres according to the specified parameters gives them primary advantages in increased magnetic permeability and, as a result, consequential losses on resonances and relaxation. Thereby, adjusting the ash microspheres' phase composition by ferromagnetic phases forming provides a change in their dielectric and magnetic characteristics and an increase in absorption capacity.

Ash microspheres containing maghemite, hercynite, and hematite phases have much better absorption capacity, especially in the low- and mid-frequency ranges, where glassy microspheres practically do not work. This approach makes cheap technogenic microspheres a valuable component for improving radio-absorbing coatings capable of reducing the reflection effective area of electromagnetic radiation from radars and reducing the elevation of military-technical facilities.

C. Comparative analysis of the hybrid coatings' properties obtained using glassy and glass-crystalline ash microspheres

When assessing the radio-absorbing ability of the developed hybrid coatings, it should be noted that reducing the effective area of electromagnetic wave energy dissipation is implemented due to dielectric losses, magnetic losses, and structural resonance. Given this, the ash microspheres' role as a filler in polymer matrix radio-absorbing coatings can be characterized as follows.

Microspheres have a low dielectric constant, which allows them to dissipate radio wave energy. Microspheres containing magnetically active phases have a low dielectric constant, enabling them to dissipate radio wave energy. Hollow microspheres work as microresonators that scatter electromagnetic waves, reducing the ESS.

For comparative analysis, the dielectric and magnetic loss indices of microspheres before and after their heat treatment for 30 min at a temperature of 850 °C were experimentally determined. It was found that for the case of X-ray amorphous microspheres $\varepsilon r = 3.1$ and $\mu r = 1.0$, while for glassy crystalline microspheres containing magnetically active phases, $\varepsilon r = 11.81$ and $\mu r = 5.8$ for (see Table 1).

TABLE 2. REFLECTION COEFFICIENTS OF COATINGS

Code	Filler	Reflection Loss (dB) at Frequency (GHz)								
		0.5	1.0	2.0	4.0	6.0	8.0	10.0	14.0	18.0
GR	amorphous AM* - 70wt.%	-0.5	-1.5	-2.8	-4.0	-5.5	**-7.0**	-4.8	-3.5	-1.0
GCR	glass-crystalline AM* - 50wt.%	-9.0	-14.0	-25.0	**-30.0**	-28.0	-25.2	-20.5	-15.0	-12.0
GRF	amorphous AM* - 50wt.% $Ni_3ZnFe_2O_4$-15wt.%	-9.5	-15.0	-28.0	**-31.0**	-30.0	-27.5	-22.0	-16.8	-13.5
AM* – ash microspheres										

Further studies were conducted on RAC samples of three compositions obtained based on epoxy resin as a polymer matrix and fillers, which differed in qualitative and quantitative characteristics (Table 2). As an effectiveness indicator of the experimental coatings with a thickness of 4 mm, the reflection coefficient RL (dB) was used. RL values were determined as the logarithmic measure of signal loss during reflection for the frequency range 0.5–18 GHz.

The results of the studies showed that the GR coating contains 70 wt. % amorphous microspheres are inefficient because the lowest absorption coefficient characterizes them. It is characterized by weak absorption in the X- and Ku-bands (8–18 GHz) and is very weak in the L- and S-bands of frequencies (1–4 GHz). The GCR coating, which contains 50 wt. % magnetically active phases, has a high absorption coefficient and can absorb up to 99 % of the electromagnetic waves' energy of (RL < –20 dB) in the 2–6 GHz frequency range. In general, the absorption range of this coating is ~1.2–12 GHz, including a sufficient level of absorption in the X- and Ku-frequencies bands, but its highest efficiency is observed in the L- and S-bands. The GRF coating, which contains 15 wt.% % nickel-zinc ferrite, demonstrates slightly better indicator values.

Thus, the coating with magnetically active microspheres provides much better absorption in the experimental frequency range, especially in its low-frequency zone, which is critical for long-range detecting radar stations. Instead, glassy microspheres may be appropriate only as an auxiliary filler to reduce the coating's density and adjust the ε_r/μ_r reflection coefficient at the “air – coating” interface.

Considering the obtained data, a modified hybrid coating formulation was proposed, which differs in the following composition, wt. %: polymer (epoxy) matrix – 55 %, glass-crystalline microspheres containing magnetically active phases – 30 %, amorphous microspheres – 10 %, graphite – 5 %. Technological additives: (dispersant, plasticizer) – 0.5–1% over 100%; the amount of hardening catalyst depends on the epoxy resin type. According to mathematical modeling results, the modified coating with a thickness of 4 mm will have the following characteristics: density 1–1.4 g/cm³, dielectric constant (ε_r ~8–12), magnetic permeability (μ_r ~4–6), absorption range of electromagnetic waves 1–12 GHz, reflection coefficient (RL up to –30 dB at frequencies 2–6 GHz), heat resistance 250–300 °C.

IV. Conclusions

The possibility of RAC modification with the ferromagnetic phases directed synthesis, metal coating, or filling with magnetic particles allows microspheres to adapt to different frequency ranges. The developed radio-absorbing composite coatings obtained using glass-crystal microspheres can ensure effective masking of ground, aviation, and naval military equipment from radar surveillance.

The comparative analysis results of the hybrid coatings' absorption efficiency based on an epoxy matrix and various fillers (glass-crystalline, amorphous hollow microspheres, and their combinations with finely dispersed Zn-Ni ferrite) confirmed the hypothesis put forward regarding the effectiveness of the method of phase composition directional correction of ash microspheres. The ash microspheres' heat treatment causes the formation of ferromagnetic compounds, the magnetic losses of which contribute to the effective radio wave absorption.

Further research will be aimed at creating structured multilayer coatings, each layer of which will be optimized for the specific frequency range absorption: 1–2 GHz (L-band), 2–4 GHz (S-band), 4–8 GHz (C-band), 8–12 GHz (X-band), 12–18 GHz (Ku-band), 18–26.5 GHz (K-band), 26.5–40 GHz (Ka-band) and 40–50 GHz (Q-band).

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Planar Active Element with Graded Band Gap InGaN Mesa Structure

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Abstract— Active GaN-based element for wide frequency application was proposed. Diode structure represents a conductive channel with cathode mesa structure based on InGaN graded layer, and two ohmic contacts both to cathode and anode. The conditions for obtaining the necessary electric field distribution in diode cathode are considered in the paper. Ability to form the uniform field distribution in diode channel was shown in the research. Diode characteristics are analyzed under impact ionization and self- heating while carrying out 2D modeling by Monte Carlo technique. DC characteristic of diode was obtained. Power spectral density diode with difference in doping profile of mesa structure and doping concentration in the channel were studied. The ability to develop a wide band generator of noise signal at frequencies close to invert time of carrier transit through channel was shown.

Keywords — GaN, electric field strength, temperature, impact ionization, self-heating, power spectral density.

I. Introduction

Active diode elements based on transferred electron effect have wide functionality. These devices can be sources of oscillation, amplifiers providing difference functionalities as an oscillation modulation, ultrasound generation, etc. [1]. However, there are several limitations caused by high frequency and functional application of those devices. Thus, transferred electron devices (TED) exhibit a significant reduction in alternating current power at a frequency correlated with the inverse intervalley transfer electron times, and frequency operation range is narrow that limits some wide frequency application.

Significant changes in device properties can be achieved utilizing the non-lattice-matched materials, the so- called graded semiconductors [2]. The application of graded-band semiconductors in transferred electron devices is well known for modification of electron transport in Gunn diode [3,4]. One of the prime properties of the graded-gap diodes is good frequency propertied that provide frequency tuning, including bias [4] to be contrary to ordinary diode with weak frequency-bias dependence. Frequency extension has been also demonstrated in a planar diode structure with active lateral borders (LBD) [5,6]. These structures consist of a conductive channel with two ohmic contacts with semiconductor element formed on top face as mesa structure. Noteworthy, the wider frequency tuning range was obtained for LBD in the form of graded gap layer [7].

The idea of using transfer electron effect in the diode channel and electronic processes in LBD to modify diode characteristic can be applied to develop wide frequency active element. Today, it is primarily electronic warfare (EW), and a problem caused by the increasing use of lethal autonomous weapons (LAW) and defenses to prevent the enemy from using the electromagnetic signal and impairing its ability to use it.

Thus, the aim of the study is to consider the development ability of active element based on III – nitride compounds, using a graded band layer. Application of InGaN alloys allows us to fabricate the high-power devices with features such as the high electric field, resistance to high temperature, high thermal conductivity.

The emphasis is on formation of condition for stable impact ionization in diode, and diode scalability to obtain required frequencies and signal parameters.

II. The Structure and Model of the Diode

Device structure is determined from condition when high frequency oscillation like Gunn oscillations is impossible. In this case, the incoherent or noise current can be obtained in diode. This regime corresponds to transferred electron amplifiers where the domain propagation is suppressed. To obtain the condition, a field distribution must be controlled by the external field.

There are several ways to achieve stability of TED. Supercritical doped devices may be stabilized by a special doping profile. For example, it can be a cathode doping "notch" or Schottky barrier near the cathode. The same effect can be achieved by special geometry of diode cathode contact or a combination of special device geometry with the mentioned above ways.

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A proposed GaN-based diode structure is shown in Fig.1:

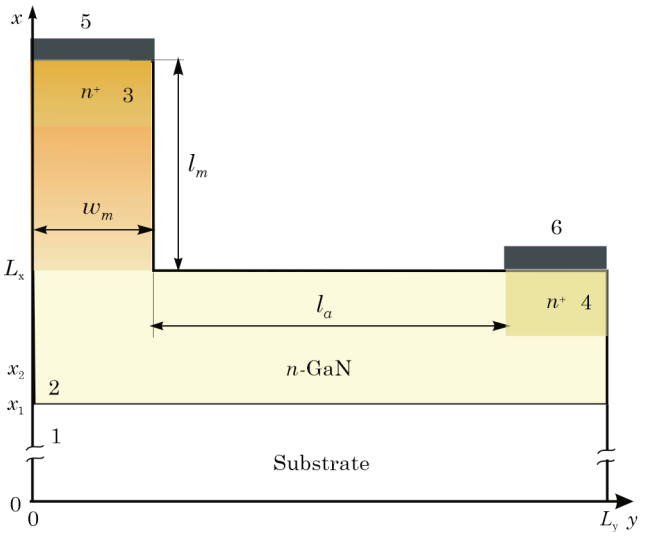


Fig.1. Diode structure: low conductivity substrate (1), *n*-type diode channel (2), high doped contact region (3,4), metal contacts (5,6)

The diode consists of GaN-based planar part (2) with thinness of L_x = 0,32 μm, It placed on a sapphire substrate (1) with thinness of 1.28 μm (point x_1). The mesa with a length of l_m and width of w_m is placed on the top of the channel formed diode cathode. Donor concentration in heavily doped contact regions (3) and (4) was $5\cdot10^{24}$ m^{-3}. All contacts in the structure are considered ohmic. The width of metal contact (4) is corresponding to w_m. It is varied from 0.16 μm to 0.32 μm. The width of metal contact (6) taken 0.16 μm. Thus, the active region length is considered to about 1 μm, the full structure length including mesa is of 1.3 μm.

Donor profile and dependence InGaN composition on coordinate along x – axis shown in Fig. 2. The dependence corresponded to diode cross section involving both cannel and mesa. The coordinate dependence of the alloy composition z in $Ga_zIn_{1-z}N$ is represented as follows:

$$z(x)=\begin{cases}1, & x\le x_2\\ z(x_2)-C(x_2,x_3)\exp\left(-\dfrac{(x-x_2)^2}{2\sigma^2}\right), & x\in(x_2,x_3)\\ 0, & x\ge x_3\end{cases}, \quad (1)$$

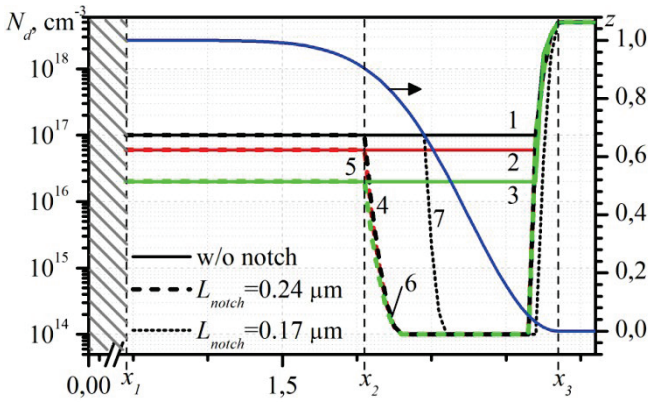


Fig.2. The doped profiles and composition vs *x*-coordinate.

Two type diodes were taken in consideration. The doping profiles 1-3 corresponded to diodes without "notch" (type 1). The doping profiles 4-7 represent diode having low resistance region ("notch") with doping concentration of 10^{20} m^{-3} (type 2).

To obtain close to real time current dependence through the diode depicted the stochastic process in diode the non-equilibrium charge carrier transport in the diode was considered synchronous Ensemble Monte Carlo (EMC) technique. Time step was 5×10^{-16} s. Three-valley model of conduction band represented by lower Γ- and upper Γ_1- and M-L-valleys was used. Initial stage of impact ionization occurred due to strong electric field is considered. For this reason, the ionization is suggested to course by one carrier type (electrons), and energy and momentum conservation laws were applied accurately. Details of simulation of carriers scattering are given in [7]. Material parameters are chosen in according to [7-10].

Anisotropy of material properties was taken into account in coordinate dependence on dielectric permittivity, impact ionization coefficients, scattering, and thermal parameters of materials and interface between them.

Due to the importance of thermal diode properties at strong electric fields, the self-heating effect in diode was taken into account similar to the model [11,12]. Ensemble Monte-Carlo technique was implemented self-consistently with Poisson and heat conduction equation solved numerically by FMG method [13].

The temperature of bottom sapphire surface is taken 300 K as well as temperature on the contact top. Temperature dependence on thermal parameters of InGaN is taken from [14]. Ohmic contact is suggested to be gold with a titanium adhesion layer. Thermal boundary conductance (TBC) at the metal–InN interface and at the metal–GaN interface are given as the same of $2\cdot10^8$ W/(K·m^2) [15]. TBC of interface between GaN and sapphire substrate was taken 10^7 W/(K·m^2) [16]. Self-heating resulted from phonon and energy of impact ionization gain.

III. Results and Discussion

A. DC characteristic of the diode

The energy band diagram and kinetic energy of modeled particles versus coordinate show on Fig. 3 for diode of type 2 with doping concentration of $6\cdot10^{22}$ m^{-3}.Quasi-electric field is known to be compensated by the electrons displacement for a *n*-type semiconductor. A quasi-electric field acts on holes significantly stronger than on electron. Holes appearing due to impact ionization move out into cathode rapidly. However, because of high electron concentration in n^+-region, the electric fields in mesa are almost unchanged.

The electric field distribution obtained for diode of type 1 and type 2 are shown in Fig. 4 for difference DC bias applied to diodes. In all DC voltage the magnitude of electric field domain is still the same and depends weakly on bias. In case of diode of type 2, the maximal magnitude of electric field domain occurs almost twice as large as in diode of type 1due to partial ionization in diode channel.

Thus, holes can be accumulated in the channel, contributing to conductivity growth and current increase, correspondently.

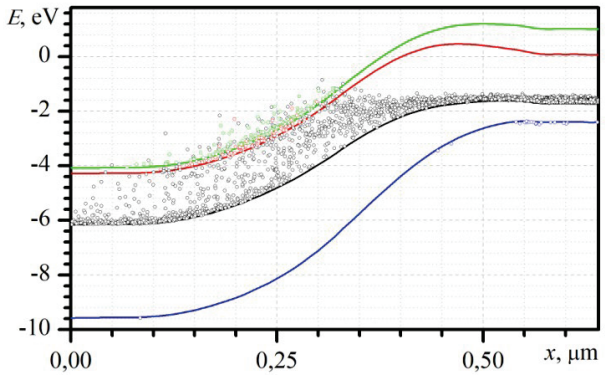
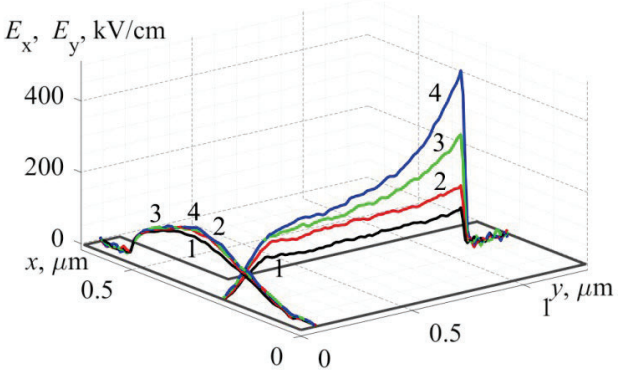
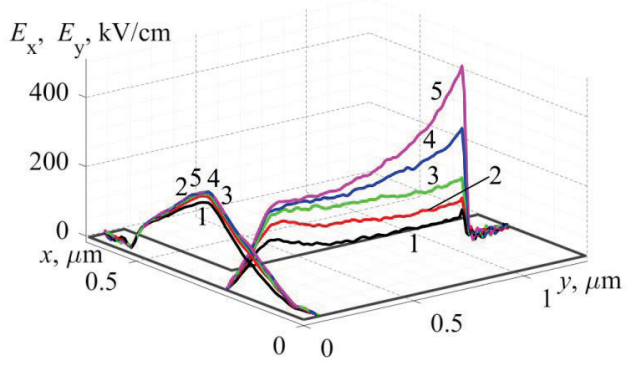


Fig.3. Zone diagram of diode cathode at U = 25V with N_{Dn} = 6×10^{16} cm^{-3} and N_{Dnotch} = 10^{14} cm^{-3}.

Electric field distribution in wide bias range is still fairly uniform or has a little deviation from the uniform one. That condition is optimal for creation of transferred electron amplifiers. In this case, larger increase in the electric field appears near the cathode and a nearly uniform field distribution across the rest of the diode. Exception is a short region near the beginning of the channel, where the electric field is low.



a)



b)

Fig.4. Distribution of the E_y in cross section of diode at x = 0.24 μm and the E_x in section y = 0.08 μm for diode w/o (a) and with notch (b) at doping concentration in channel 6×10^{17} cm^{-3}: 1 – U_0=10 V; 2 – U_0=15 V; 3 – U_0=20 V; 4 – U_0=25 V; 5 - U_0=30 V.

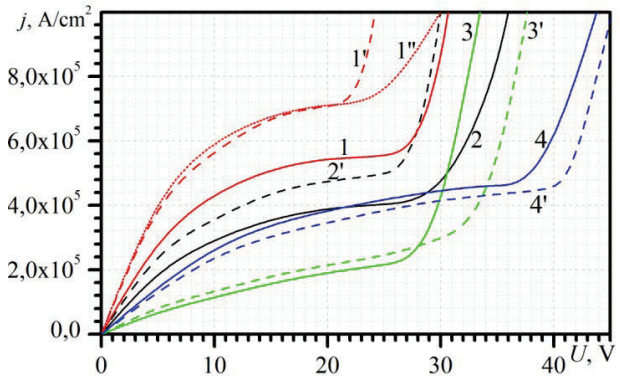


Fig.5. Dependences of current density on applied bias voltage of diodes with and w/o notch (1-3 – w/o "notch"; 1'-4',1'' – with "notch") : 1,1',1'' – N_{Dn} = 10^{17} cm^{-3}; 2,4,2',4' – N_{Dn} = 6×10^{17} cm^{-3}; 3,3' - 2×10^{17} cm^{-3}; 1'-3' - N_{Dnotch} = 10^{14} cm^{-3}, L_{notch} = 0.24 μm; 1'' - N_{Dnotch} = 10^{14} cm^{-3}, L_{notch} = 0.17 μm; 1-3,1'-3',1''4' – L_y = 1.28 μm; 4,4' – L_y = 2.56 μm

Dependences of current density on applied bias voltage for difference type diode with doping level of 2·10^{22} m^{-3}, 6·10^{22} m^{-3}, and 10^{23} m^{-3} are shown in Fig. 5. Diode characteristics obtained at 300 K are studied with account and without account possible self-heating effects.

In Fig 5, all solid lines correspond to diode of type 2. For these diodes, the biggest impact ionization threshold appears if concentration is less than 10^{23} m^{-3}. When doping concentration is large, the impact ionization leads to significant rise in electron concentration in channel, causing rise in current. In diode of type 1, the additional current rise is associated with increasing holes concentration in channel. When width of low resistance region is small, the diode properties become closer to diode of type 1. This case corresponded to doping profile 7 in Fig, 2.

Self-heating effect in diode determined by heating of diode region near interface between channel and mesa, is shown in Fig. 6. Heat sources are mainly impact ionization.

The temperature rise in diode of type 1 is higher, compared to diode of type 2 at the same bias. That is conditioned by diode heating in the stronger electric fields near anode. At bias voltage of 25 V, this difference is about 120 K.

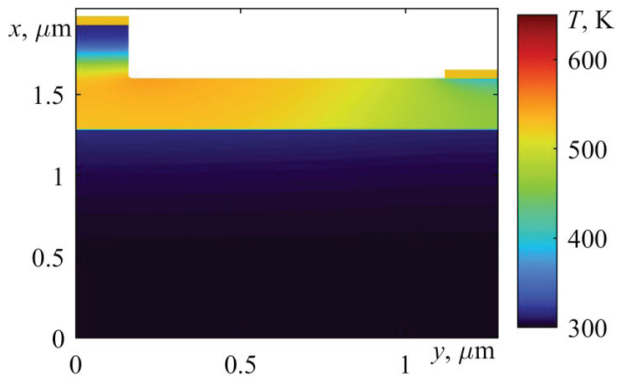


Fig. 5. Temperature distribution in the GaN diode of type 2 with doping of 10^{23} m^{-3} at U=25 V; t_{sim}=12.8 ns.,

B. PSD characteristics of the diodes

To estimate the possible usage of diodes for noise generators, the power spectral density of diode current was

estimated and temporary time samples of diode were analyzed. The numerical algorithm corresponded to [17]. Power spectral density (PSD) $S(\omega)$ was evaluated by the autocorrelation function (ACF) of current sequences, Fig. 6.

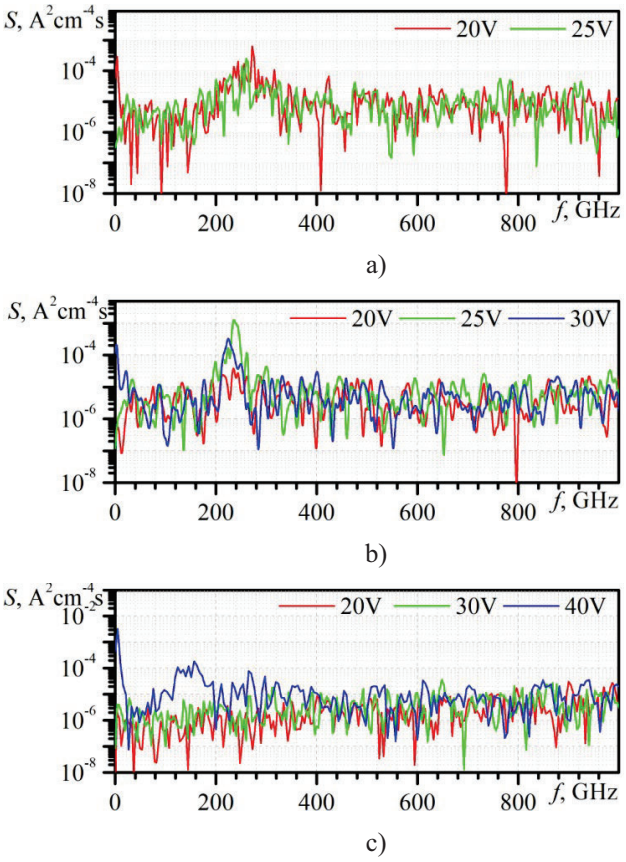


Fig. 6. PSD for diodes w/o notch (a, c) and with notch (b) at doping concentration in channel 6×10^{16} cm^{-3} with L_y = 1.28 μm (a, b) and L_y = 2.56 μm (c).

The rise of PSD, occurred at bias voltage, corresponded to beginning of impact ionization in diode at frequencies of 160-360 GHz. The frequency range of PSD is wide. As can be seen from Fig.6 a), the wider frequency range is in diode type 1. In diodes of type 2, the frequency band is narrower if PSD rises. It is noteworthy that Gunn generation is not observed in all represented cases. The study of diode with channel of 2.56 μm is shown in Fig.6 c), PSD rises in lower frequency and in wider frequency range (80- 400 GHz).

IV. Concusion

Active GaN-based element with cathode graded gap in InGaN mesa structure has been studied by Monte Carlo technique. In the considered structure, the electric field stabilization can be achieved. A wide band noise signal can be obtained due to electric field distribution.

A power spectral density diode with a differential mesastructure doping profile and channel doping concentration was investigated. It has been shown that by increasing the bias voltage in the diode, an increase in PSD of more than an order of magnitude can be achieved. Maximal frequencies rise closes to invert time of carrier transit through the channel Thus, the immediate result is the possibility of adjusting the maximum frequency by changing the diode channel, and therefore to expand the possible range of application of the proposed structure. The following problems are the optimization of the diode parameters, the study of diode cooling as well as the use of the structure as an amplifier.

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3D Modeling Method for Endoprosthesis of Damaged Finger Using CT Scan Data

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***Abstract*—This paper presents a comprehensive workflow for the 3D modeling and printing of a damaged human finger from computed tomography (CT) scan data, tailored for the analysis of finger joint endoprosthesis. The process begins with image acquisition and DICOM-to-STL conversion in RadiAnt DICOM Viewer, followed by refinement and segmentation in Autodesk Meshmixer, and slicing for 3D printing in OrcaSlicer. Both anatomical and cross-sectional models are produced to enable implant fitting, biomechanical testing, and preoperative surgical planning. The novelty of this work lies in the integration of a triadic software chain and the dual-model approach, specifically adapted to the unique requirements of small finger joints, which distinguishes it from conventional workflows developed for larger bones. The methodology improves segmentation accuracy, facilitates implant design and selection, and supports more effective surgical planning. The proposed approach demonstrates significant potential for reducing operative time, enhancing patient-specific implant fitting, and contributing to better clinical outcomes in orthopedic surgery. This workflow may serve as a foundation for clinical trials and future integration into computer-assisted surgical systems**

***Keywords*—3D printing, biomedical engineering, endoprosthesis, finger joint, computed tomography (CT), anatomical modeling, surgical planning, biomechanics**

I. Introduction

The application of 3D printing, also known as additive manufacturing, in the medical field has witnessed significant growth in recent years. This technology offers the capability to create patient-specific anatomical models, surgical guides, and even customized implants [1, 2]. The ability to translate medical imaging data, such as CT scans, into tangible 3D structures has opened new avenues for improving diagnostic accuracy, surgical planning, and patient outcomes [1, 2]. Moreover, recent research emphasizes that patient lifestyle patterns also strongly influence the acceptance and development pathways of 3D-printed prostheses, particularly in hand applications [3].

Finger joint endoprosthesis is a complex surgical procedure that requires precise planning and implant fitting to restore function and alleviate pain [4, 5]. The use of 3D-printed models can provide surgeons with a valuable tool to visualize the patient's anatomy, simulate surgical procedures, and optimize implant design before the actual surgery. Given the high burden of musculoskeletal injuries and their contribution to long-term disability [6], the development of accurate, reproducible, and patient-specific models is of particular clinical relevance. Recent user-centered design approaches to pediatric prostheses have further demonstrated the importance of personalization in improving patient satisfaction and clinical outcomes [7].

The novelty of this study lies in the development of a comprehensive workflow for finger joint endoprosthesis, combining a triadic software chain (RadiAnt DICOM Viewer → Autodesk Meshmixer → OrcaSlicer) with the dual use of anatomical and cross-sectional models. Unlike previously reported methods focused on large bones or single-stage modeling, the proposed approach is specifically adapted to the challenges of small finger joints, enabling more accurate segmentation, implant fitting, and biomechanical evaluation. Furthermore, our methodology aligns with emerging research on hybrid degradable artificial joints incorporating bioactive nanofibers [8] and complements recent work on dynamic orthotic solutions for finger rehabilitation [9], thus underscoring its potential impact across surgical planning, implant development, and post-operative care.

II. Methodology

The methodology employed in this study comprises a series of sequential steps, each crucial to achieving the final objective of producing accurate 3D printed finger models.

A. Abbreviations and Acronyms

The initial step involves the acquisition of CT scans of the damaged finger. CT scans utilize X-rays and computer processing to create detailed cross-sectional images of the body [10]. The data is stored in the DICOM format, which is the standard for medical imaging.

RadiAnt DICOM Viewer software (version 2025.1) was used to open and process the DICOM files. RadiAnt DICOM Viewer is a free, open-source DICOM viewer that allows for the visualization, manipulation, and analysis of medical images. In this study, RadiAnt DICOM Viewer was specifically used for visualizing CT scan slices (Fig. 1) in various planes (axial, coronal, sagittal), adjusting the window level and width to optimize the visualization of bone structures (Fig. 2), isolating the finger region from surrounding tissues, and exporting the selected finger region as an STL (Stereolithography) file.

The STL format is a widely used file format for 3D printing. It represents the surface geometry of a 3D object as a collection of triangles. The conversion of DICOM data to STL format is a critically important step, as it translates the medical image data into a format that can be interpreted by 3D modeling and slicing software.

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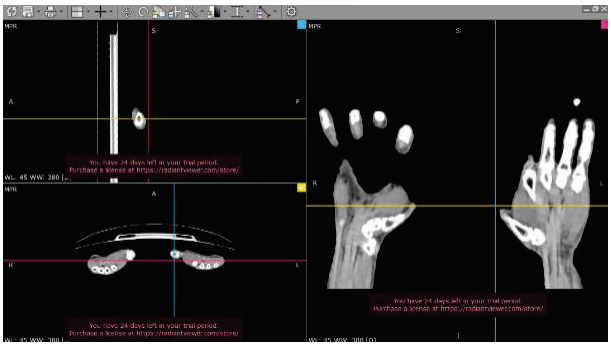


Fig. 1. CT scan

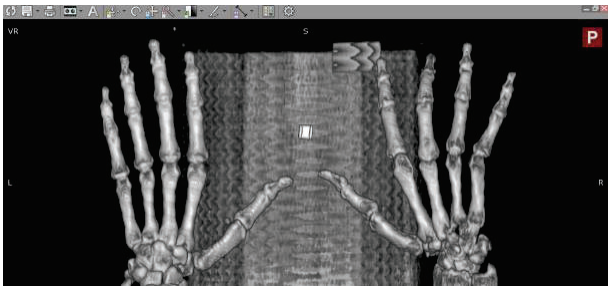


Fig. 2. Bone structure visualizations

B. Model Processing with Autodesk Meshmixer

The exported STL file from RadiAnt DICOM Viewer was then imported into Autodesk Meshmixer (Fig. 3), a free 3D modeling software developed by Autodesk. Autodesk Meshmixer (version 3.5.0) provides a user-friendly interface and a variety of tools for editing, refining, and optimizing 3D models [11].



Fig. 3. Exported stl file

In this study, Autodesk Meshmixer was used for noise reduction and smoothing, where tools were applied to smooth the surface and remove any unwanted irregularities that can sometimes arise from CT scan data noise and artifacts. Accurate segmentation of the finger into its individual bone segments (phalanges) (Fig. 4) is crucial for endoprosthesis analysis; therefore, Autodesk Meshmixer's selection and separation tools were used to isolate each phalanx, and the joints between the phalanges were then articulated to simulate the finger's range of motion. To reduce material consumption and printing time, the finger model was hollowed out using Autodesk Meshmixer's hollowing tools, with wall thickness carefully controlled to maintain structural integrity. Furthermore, Autodesk Meshmixer was utilized to analyze the model and identify areas that would require support structures during 3D printing, generating and optimizing these structures to ensure successful printing while minimizing material usage.



Fig. 4. Phalangeal bones

C. Figures and Tables

The processed STL model from Autodesk Meshmixer was then imported into OrcaSlicer. OrcaSlicer (version 2.3.0) is powerful and user-friendly slicing software that prepares the 3D model for printing (Fig. 5). It converts the 3D model into a series of thin layers and generates the G-code, which contains the instructions for the 3D printer.

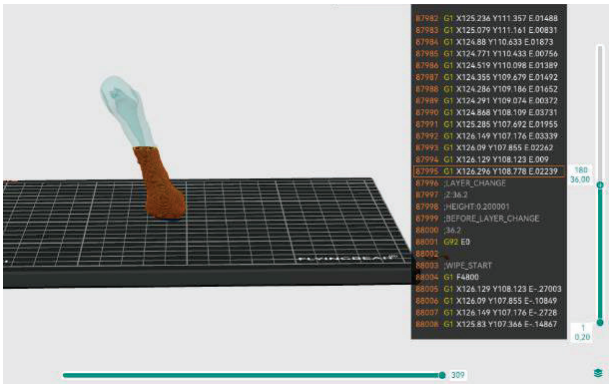


Fig. 5. 3D model slicing

OrcaSlicer allows for the adjustment of various printing parameters, including layer height (which affects print quality and resolution), infill density (the amount of material used to fill the interior of the model, affecting its strength and weight), printing speed (which affects printing time and quality), support structures (the type and density needed for overhanging parts), and nozzle temperature and bed temperature (which affect adhesion and quality of the printed material).

In this study, OrcaSlicer was used to generate two types of 3D prints. The first was a full anatomical model, a complete 3D model of the finger providing an accurate representation of its external anatomy, useful for visualizing the overall shape and size of the finger, as well as for surgical planning. The second was a cross-sectional model, a model of the finger virtually cut along a specific plane, revealing its internal structures, which is particularly valuable for analyzing the interaction between the endoprosthesis and the bone, and for assessing the fit and placement of the implant.

The 3D models were printed (Fig. 6) using a Flyingbear Ghost 5 3D printer with PLA filament. Printing parameters were optimized to achieve a balance between print quality, speed, and material consumption.

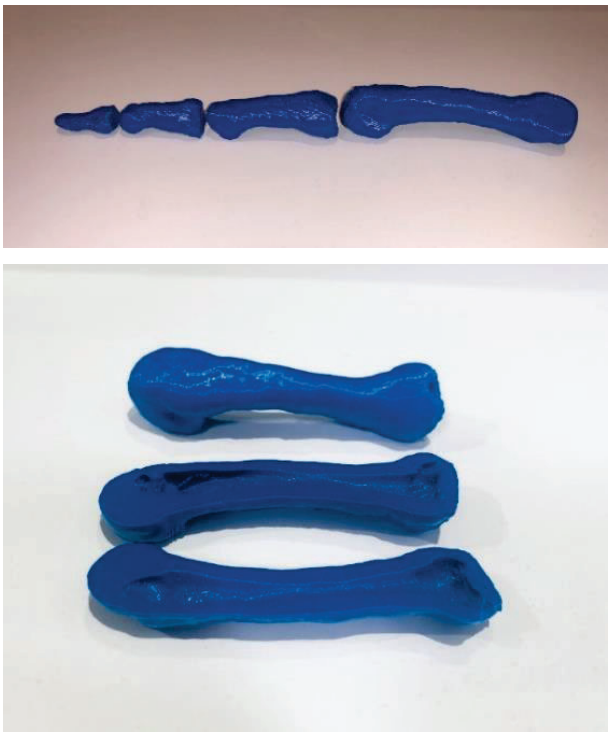


Fig. 6. Printed 3D models

D. Endoprosthesis Study

The 3D printed models were then used to facilitate a comprehensive study of finger joint endoprosthesis. This included implant fitting and placement, where the models were used for physical testing of the fit and placement of different endoprosthesis designs, allowing surgeons to visualize the implant's interaction with the bone and identify any potential issues before actual surgery. Biomechanical analysis was also performed, where the 3D printed models, in conjunction with biomechanical testing equipment, can be used to simulate finger movements and analyze the stresses and strains on the implant and surrounding bone, helping to optimize implant design and predict its long-term performance.

Furthermore, 3D printed models served as a valuable tool for surgical planning and simulation, enabling surgeons to practice surgical procedures, determine the optimal approach, and anticipate potential challenges, particularly with the use of telemedicine services [12]. This approach of computer planning based on mathematical modeling and virtual visualization of proposed solutions is highly relevant at the current stage [13, 14].

Finally, the models were used for patient education and communication, helping to improve patient understanding of their condition and the planned surgical procedure, thereby reducing anxiety.

III. RESULTS

The methodology described in this paper successfully enabled the creation of accurate and detailed 3D printed models of a damaged finger from CT scan data. The use of RadiAnt DICOM Viewer, Autodesk Meshmixer, and OrcaSlicer software provided an optimized and efficient workflow for processing medical images and preparing models for 3D printing.

The 3D printed models demonstrated a high degree of anatomical accuracy, reflecting the intricate details of the finger's bone structure and joint articulation. Cross-sectional models proved particularly valuable for visualizing internal anatomy and the interaction between the endoprosthesis and the bone, providing opportunities for stress–strain evaluation consistent with biomechanical approaches reported in previous studies [15].

The use of these models in endoprosthesis studies yielded several significant benefits, including improved implant fitting (through physical testing of implant fit on the 3D printed model, allowing selection of the most appropriate implant size and design, minimizing complications), enhanced surgical planning (providing surgeons with a realistic representation of patient anatomy, enabling more effective planning and anticipation of challenges), reduced surgical time (due to pre-surgical planning, minimizing patient exposure to anesthesia and infection risk), and improved patient outcomes (by enhancing surgical precision, reducing complications, and facilitating faster recovery).

IV. DISCUSSION

This study demonstrates the applicability of 3D modeling and printing for preoperative planning in finger joint endoprosthesis. Although additive manufacturing is widely used in orthopedic practice, most reported methods are tailored for large bones or generalized anatomical modeling, which limits their precision when applied to small joints.

The scientific novelty of this work lies in a comprehensive methodology that integrates a triadic workflow (RadiAnt DICOM Viewer → Autodesk Meshmixer → OrcaSlicer) into a streamlined sequence from CT data acquisition to optimized 3D printing. Unlike existing approaches focusing on single stages, the proposed method ensures consistency across segmentation, refinement, and slicing. Moreover, the combination of

anatomical and cross-sectional models provides additional advantages for implant analysis, addressing the limitations of conventional replicas that do not adequately represent implant–bone interaction.

A further distinction is the adaptation of the workflow to the unique requirements of small finger joints, which differ significantly from large bones in terms of scale, biomechanics, and surgical complexity. The methodology supports precise phalanx segmentation, controlled hollowing to reduce material use, and model preparation suitable for both biomechanical evaluation and surgical rehearsal. The resulting physical models enabled implant fitting trials, improved visualization for surgical planning, and enhanced patient communication. Similar to recent user-centered design studies on pediatric prostheses [7], this highlights the importance of patient-oriented approaches in improving clinical outcomes.

Recent advances in degradable hybrid 3D-printed artificial joints incorporating nanofibers loaded with pharmaceuticals and growth factors [8] highlight the broader potential of combining additive manufacturing with regenerative medicine. These developments, together with emerging research on dynamic orthotic solutions for rehabilitation [9], suggest that the integration of preoperative planning, implant design, and post-operative care into a single workflow is a promising future direction. In addition, biomechanical studies on graft fixation and stress distribution in musculoskeletal systems [15, 16] reinforce the relevance of stress–strain testing on 3D-printed finger models for predicting implant stability and long-term outcomes.

Nevertheless, the approach is not without limitations. Model accuracy depends on CT image quality, processing requires specialized expertise, and current printing materials do not fully replicate bone properties. Cost and regulatory considerations also remain barriers to broader clinical integration. Future research should focus on automating segmentation, testing advanced printing materials, and validating the method through clinical studies. Despite these challenges, the presented methodology represents a significant step toward patient-specific solutions for small joint surgery and provides a foundation for further integration into computer-assisted and telemedicine systems.

V. Conclusion

This study presents a comprehensive workflow for 3D modeling and printing of a damaged finger from CT scan data, adapted to the challenges of finger joint endoprosthesis. The integration of RadiAnt DICOM Viewer, Autodesk Meshmixer, and OrcaSlicer with both anatomical and cross-sectional models enable accurate segmentation, implant fitting, and biomechanical evaluation. The workflow supports patient-specific surgical planning, implant testing, and stress–strain analysis, contributing to improved outcomes in musculoskeletal injury cases. Beyond surgery, the approach enhances patient communication and aligns with user-centered prosthetic design. Future work should focus on automation, advanced biomaterials, and clinical validation, extending the methodology toward regenerative implants and rehabilitation solutions.

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Giant magnetoresistance in $Bi_{88.08}Mn_{11.92}$

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***Abstract*— Magnetoresistance (MR) measurements in $Bi_{88.08}Mn_{11.92}$ showed that the *MR* remained positive over the whole range of magnetic fields. At 90 kOe, the maximum magnetoresistance reached about 3169 % when the magnetic field was perpendicular to the current (*H* ⊥ *I*) and about 379 % for the parallel orientation (*H* // *I*). For comparison, previous results on $Bi_{95.69}Mn_{3.69}Fe_{0.62}$ - characterized by a lower Mn content - showed much higher *MR* values under identical conditions: almost 3877 % for *H* ⊥ *I* and 742 % for *H* // *I*. It is proposed that the main reason for the difference in magnetoresistive properties between $Bi_{88.08}Mn_{11.92}$ and $Bi_{95.69}Mn_{3.69}Fe_{0.62}$ is the effect of the intrinsic magnetism of the α-BiMn inclusions, which significantly affects the transport behavior of electrons and holes within the bismuth-based matrix.**

Keywords - Bismuth–manganese solid solutions*, *valleytronics*, *magnetoresistance*, *Dirac fermions, α-BiMn phase, bismuth, quantum limit

I. Introduction

Bismuth is a material with unique electronic properties. Notable properties include a small Fermi energy (EF) of only a few hundredths of an electron volt, a large electron free path length of up to several millimeters, small effective masses of charge carriers of $10^{-2} \div 10^{-1}$ of the free electron mass, and a reduced density of charge carriers of 10^{-5} electrons per atom. Bismuth also has a large dielectric constant of about 10^{-5} and a large diamagnetic permeability of approximately 100, as well as a g-factor as high as 200. Bismuth was the first material in which large positive magnetoresistance, cyclotron resonance in a metal, the Shubnikov–de Haas and de Haas–van Alphen effects, oscillating magnetostriction, undamped microwave waves and a number of dimensional effects were experimentally detected [1], [2]. It has recently been suggested that some of the electrons in bismuth could be considered massive Dirac fermions [3]–[7].

Recently, a new branch of electronics has emerged, called valleytronics, which is supposed to control such degrees of freedom as valleys, local maxima/minima in the valence or conduction band of multi-valley semiconductors and semimetals [8]–[10]. It is even proposed to create qubits for quantum computers based on valleys [11].The list of materials for valleytronics includes bismuth [12], [13]. Partly because of this, a comprehensive study of bismuth and materials based on it is promising these days.

Compounds based on bismuth are no less fascinating than pure bismuth itself. Among such materials, Bismuth–Manganese solid solutions occupy a special place, exhibiting ferromagnetism above room temperature is approximately 360 °C and high coercivity at room temperature [14]. With increasing temperature, the coercivity rises to values exceeding those of conventional rare-earth permanent magnets [15], [16]. Until recently, the electrical transport properties of these materials have received very little attention

Several years ago, we studied the electrical resistivity of the solid solution $Bi_{95.69}Mn_{3.69}Fe_{0.62}$ in magnetic fields. It was found that, under a magnetic field, the temperature dependence of the electrical resistivity exhibits nonmonotonic behavior, and the magnetoresistance is positive over the entire temperature range, reaching a maximum value of several thousand percent in a magnetic field of 140 kilooersted [17]–[20].

The aim of this work was to demonstrate how the magnetoresistive behavior evolves with increasing manganese content by performing a detailed analysis of the field-dependent magnetoresistance in the compound $Bi_{88.08}Mn_{11.92}$ under magnetic fields up to 90 kOe at various temperatures.

II. Samples and Experimental Method

The samples were prepared using the Bridgman method. This crystal growth technique can be used to produce either single crystals or textured polycrystalline samples. In this case, textured polycrystals were obtained. For electrical resistivity measurements, the samples were cut into parallelepipeds with approximate dimensions of 7 x 2 x 2 mm, along the base of the original cylindrical ingot (perpendicular to the c-axis, nominally).

The quantitative elemental composition was determined using a Philips 515 scanning electron microscope with an EDAX PV9800 spectrometer. The results of this study revealed the following average quantitative elemental composition by volume of the material: 88.08 at% Bi and 11.92 at% Mn. Deviations from this composition did not exceed ±2% at different points on the sample surface (approximately eight measurements were taken). No other elements except bismuth and manganese were detected.

According to the phase diagram (Figure 1) [21], at these manganese concentrations, the sample was a bismuth matrix containing the magnetic α-BiMn phase.

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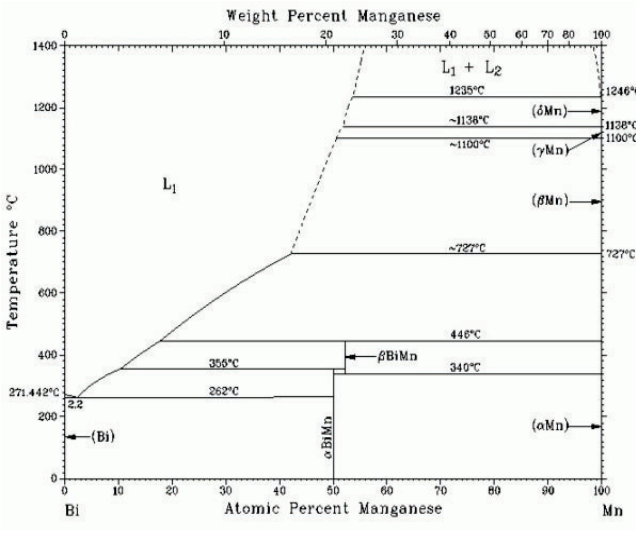


Fig. 1. Bi-Mn phase diagram [21]

Resistive and magnetoresistive measurements were performed using the standard four-probe method on a fully automated Quantum Design PPMS-9T (Figure 2). Current and voltage contacts were made using silver paste. The measurements were carried out with an alternating current (I = 30 mA, f = 17 Hz) applied along the longest dimension of the sample, both in zero magnetic field and in transverse and perpendicular magnetic fields up to 90 kilooersted using an insert with a rotary mechanism (Figure 3 and 4), in the temperature range from 2 to 300 K. In these measurements, a permanent magnetic field was generated using a superconducting solenoid.



Fig. 2. Commercial automated unit PPMS-9T by Quantum Design. 1 - cryogenic part (nitrogen, helium cryostats and related equipment); 2 - stand with measuring equipment; 3 - PPMS Model 6000 controller; 4 - computer; 5 - place for inserting measuring inserts.

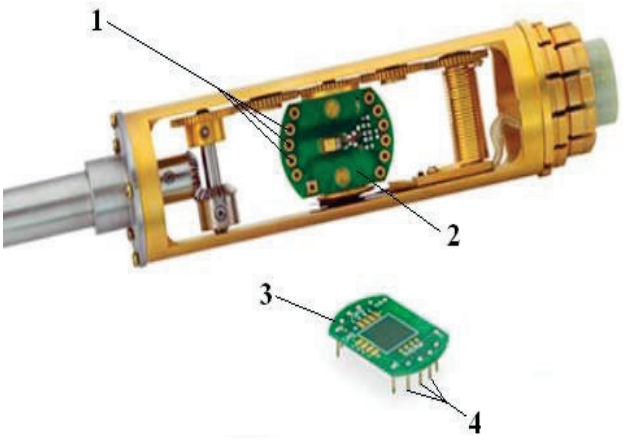


Fig. 3. Lower part of the insert with a rotary device for measuring electrical resistance. 1 - contact sockets, 2 - rotary platform, 3 - measuring holder, 4 - electrical connectors.

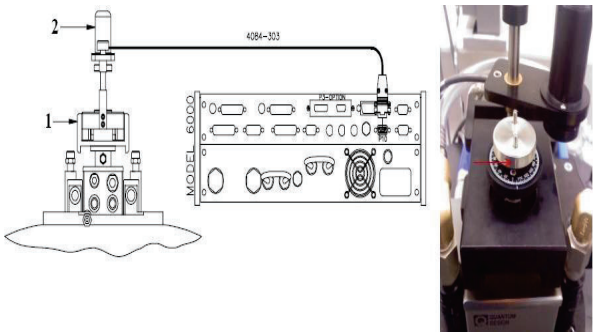


Fig. 4. Upper part of the insert with the rotary device for measuring electrical resistance. 1 - device for rotating the turntable, 2 - motor for the device for rotating the turntable.

III. Results and Discussion

The Figure 5 presents the relative magnetoresistance $MR = \Delta\rho / \rho_0 = ((\rho(H) - \rho(0)) / \rho(0)) \cdot 100\ \%$ for the transverse configuration ($\boldsymbol{H} \perp \boldsymbol{I}$). It is clear that the magnetoresistance is positive across the entire range of magnetic fields and temperatures. The maximum relative magnetoresistance at 90 kOe occurs at 120 K and reaches approximately 3169 %. The magnetoresistance curves exhibit distinctly different behaviors in different temperature intervals: below 80 K, $\Delta\rho / \rho_0$ rises sharply up to about 20 kOe, after which the growth rate slows so that at high fields the curves approach saturation. Above 80 K, no saturation is observed within our field range.

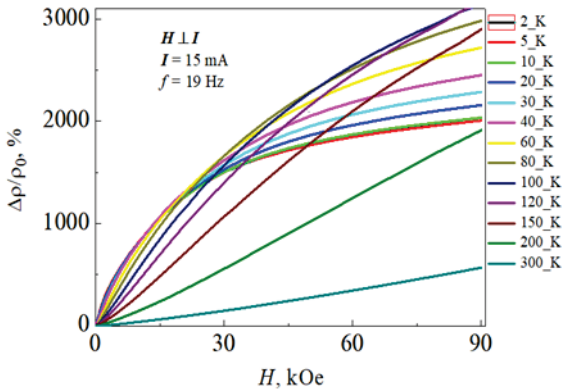


Fig. 5. Magnetic field dependences of the relative magnetoresistance of $Bi_{88.08}Mn_{11.92}$ for $\boldsymbol{H} \perp \boldsymbol{I}$ at different temperatures.

The Figure 6 shows the relative magnetoresistance $\Delta\rho/\rho_0$ for the longitudinal configuration ($\boldsymbol{H}$ // $\boldsymbol{I}$). Again, $\Delta\rho/\rho_0$ remains positive throughout all fields and temperatures. The maximum magnetoresistance in this case is nearly ten times lower than in the transverse case: at 90 kOe and 120 K it is approximately 379 %. Below 80 K, the field-dependencies of $\Delta\rho/\rho_0$ become non-monotonic, displaying a pronounced maximum in low magnetic fields that shifts toward higher fields as temperature increases. Earlier literature [22]–[24] also observed a significant maximum in the magnetic field dependence of the longitudinal magnetoresistance of bismuth at approximately the same magnetic field values as for $Bi_{88.08}Mn_{11.92}$. This maximum was associated with the quantum limit being reached, at which point only one Landau level remains below the Fermi level in the electron zone. In a semimetal, this results in charge carriers 'overflowing' between the hole and electron zones. Consequently, the concentration of electrons and holes increases, resulting in a decrease in electrical resistance [23]. Above 80 K, the magnetoresistance field-dependencies of $Bi_{88.08}Mn_{11.92}$ become monotonic.

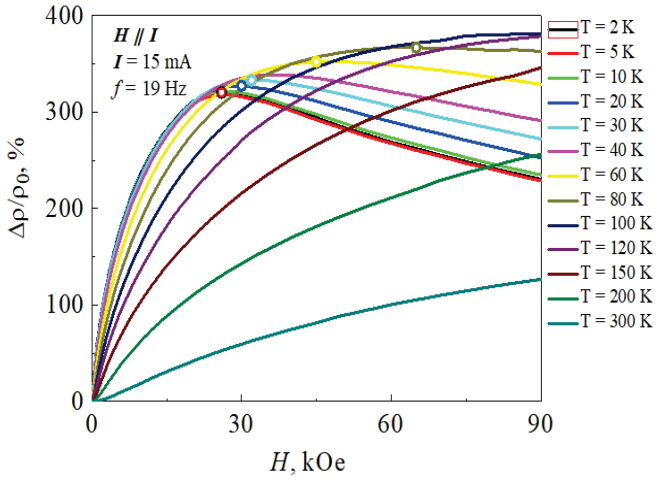


Fig. 6. Magnetic field dependences of the relative magnetoresistance of $Bi_{88.08}Mn_{11.92}$ for $\boldsymbol{H}$ // $\boldsymbol{I}$ at different temperatures.

Table 1 compares the magnetoresistance values of compounds with different manganese content. As the manganese concentration increases, the relative magnetoresistance values decrease.

Table 1. Maximum magnetoresistance values $MR = \Delta\rho/\rho_0 = ((\rho(H) - \rho(0))/\rho(0))\cdot 100$ % for $Bi_{95.69}Mn_{3.69}Fe_{0.62}$ and $Bi_{88.08}Mn_{11.92}$

Compounds	MR_{max}, $\boldsymbol{H}$ // $\boldsymbol{I}$, at 90 кOe	MR_{max}, $\boldsymbol{H} \perp \boldsymbol{I}$, at 90 кOe
$Bi_{95.69}Mn_{3.69}Fe_{0.62}$	3877 %	742 %
$Bi_{88.08}Mn_{11.92}$	3169 %	379 %

According to the literature [5], the Fermi energy of bismuth is only a few hundredths of an electron volt. External perturbations (e.g. temperature, magnetic field or strain) can affect the degree of overlap between the electron and hole pockets, consequently affecting the kinetic properties (e.g. electrical conductivity). For instance, when the electron and hole Fermi surfaces coincide, a metal-to-insulator transition may occur, accompanied by potential intermediate states.

Bismuth has almost an equal number of electrons and holes. As reported in the literature, the Fermi surface of bismuth consists of one hole pocket at the *T*-point and three electron pockets at the *L*-points within the first Brillouin zone of the rhombohedral unit cell (see Fig. 7). These three equivalent electron pockets are considered to be triply degenerate electron valleys (*e*1, *e*2, *e*3) in a zero magnetic field (see Fig. 7).

However, due to the strongly anisotropic shape of the electron pockets (see Fig. 7), this valley degeneracy can be lifted by an applied magnetic field [12, 13]. At the *L*-points, bismuth has narrow band gaps ($E_g \approx 15.3$ meV) between the conduction and valence bands (see Fig. 8). Applying a magnetic field along the binary and bisectrix axes initially reduces the gap at the *L*-points, eventually leading to a level crossing at the critical field [25, 26].

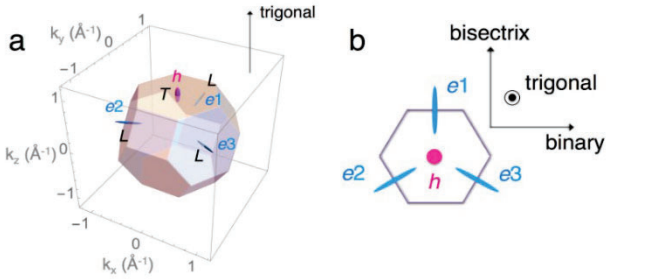


Fig. 7. (a) Three-dimensional schematic pictures of the first Brillouin zone and the Fermi surfaces of bismuth. Three valleys of electron pockets, labelled *e*1, *e*2 and *e*3 (blue), are located at *L*-points, and one valley of hole pockets, labelled *h* (purple), is located at the *T*-point. (b) Diagram of Fermi surfaces projected onto a binary bisector plane [5].

As the field increases further, the hole and electron subzones will move in opposite directions, eventually leading to a transition from a semimetal to a semiconductor. The differences in electrical resistance behaviour observed in the solid solutions studied in this work, compared to pure bismuth, may be due to the additional influence of magnetism in the α-BiMn phase on the material's conductive properties.

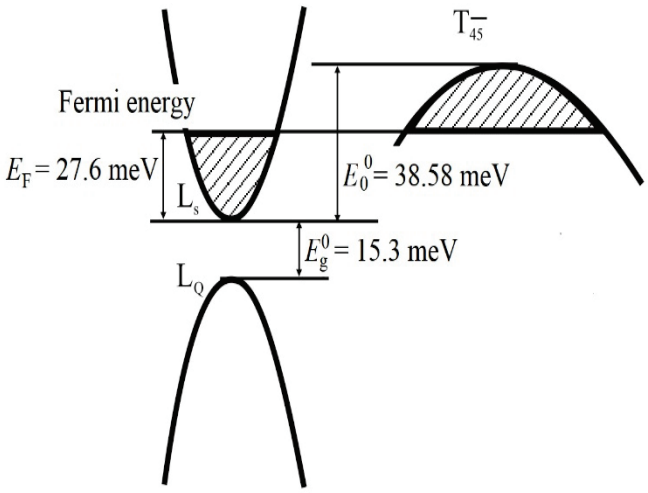


Fig. 8. The energy position of the extrema of the valence band and conduction band at points L and T of the Brillouin zone of a bismuth crystal in the vicinity of the Fermi energy at T = 4.2 K.

Such influence can cause changes in the degree of overlap of the electron and hole regions of the Fermi surface, and, as a result, a change in the conductive properties relative to pure bismuth. Changing the angle of inclination of the magnetic field relative to the crystallographic directions can lead to an additional change in the degree of overlap of the electron and hole regions, changes in the magnetic structure of the αBiMn phase, and deviations in electrical resistance behavior from that observed in the absence of a magnetic field, as well as the emergence of strong anisotropy in the temperature dependencies of electrical and magnetic resistance, which was revealed in our studies.

IV. Conclusion

1) For the first time, the magnetic-field dependencies of magnetoresistance were investigated for the textured polycrystal $Bi_{88.08}Mn_{11.92}$ in two configurations: magnetic field perpendicular to the transport current and magnetic field parallel to it.
2) The maximum relative magnetoresistance values for both configurations are observed at a field strength of 90 kOe and a temperature of 120 K. These values are 3169 % for $\boldsymbol{H} \perp \boldsymbol{I}$ and 379 % for $\boldsymbol{H} \parallel \boldsymbol{I}$. This is significantly lower than the values observed in solid solution with lower manganese content ($Bi_{95.69}Mn_{3.69}Fe_{0.62}$): 3877 % and 742 %, respectively.
3) It has been found that the magnetoresistance of $Bi_{88.08}Mn_{11.92}$ in the $\boldsymbol{H} \perp \boldsymbol{I}$ configuration differs significantly from that in the $\boldsymbol{H} \parallel \boldsymbol{I}$ configuration. In the first case, the dependencies are monotonic. In the second case, maxima are observed in the magnetoresistance-magnetofield dependencies below 80 K in weak magnetic fields. At the same time, above 80 K, the dependencies are monotonic.

Acknowledgments

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Implementation of a Battery Energy Storage System Model in MATLAB/Simulink

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***Abstract*— Modeling of energy storage systems is a very important condition for the development of energy storage systems. This article shows the development and modeling of a battery energy storage system (BESS) in MATLAB Simulink, which is aimed at settling peak loads in the grid. This model allows us to visualize the key parameters of the system, including the state of charge (SOC), the load profile before and after the implementation of the BESS, and the efficiency of peak load balancing. This, in turn, makes it possible to conduct a comprehensive analysis of the system and assess the feasibility of its implementation. The presented model can be used as a tool for further research, optimization of the storage device size, and feasibility study of BESS implementation in distributed energy systems.**

Keywords*— *Battery Energy Storage System, BESS, energy storage modeling, smart grid, simulation model, energy efficiency

I. Introduction

The Battery Energy Storage Systems (BESS) are an important component of modern power grids. They act like a temporary power buffer, allowing accumulating excess electricity and using it during peak load periods, which stabilizes the grid, regulates the frequency, smooth out daily fluctuations in electricity consumption and increases the efficiency of renewable energy sourcesx [1].

Contrary to traditional methods of energy storage based on thermal or mechanical processes, batteries operate based on electrochemical reactions that take place between the active components of the storage devices in the middle of the cells. Therefore, electrochemical storage systems are highly efficient, responsive to load changes, and allow for repeated charging and discharging without large power losses [2].

Since electricity consumption is not constant, and most generating capacities, such as thermal power plants or nuclear power plants, are not highly maneuverable, there is a need for fast and flexible load leveling. In this context, the implementation of energy storage systems will not only reduce the load on power plants during peak consumption hours, but also ensure more efficient and stable operation of the entire power system [3]. In addition, BESS can perform the function of backup power supply to the grid, for example, providing power to critical infrastructure in case of malfunctions or short-term maintenance of generating units at power plants.

Effective design and implementation of energy storage systems requires accurate modeling tools that will simulate the dynamics of the Battery energy storage system (BESS) in various operating modes. Among the modern modeling methods is MATLAB/Simulink software from The MathWorks Company, which provides fast design and simulation with customizable settings and well-designed visualization of output data [4]. Therefore, modeling BESS using Simulink will allow us to study the impact of battery parameters on the efficiency of load balancing, assess the state of the grid in real time, and conduct a preliminary feasibility analysis of the introduction of energy storage systems [5].

Figure 1 shows the general structure of how BESS works with the grid and consumers. The BESS acts as a buffer between the generating plant and the consumers, discharging during peak consumption hours and recharging during low-load hours, such as at night.

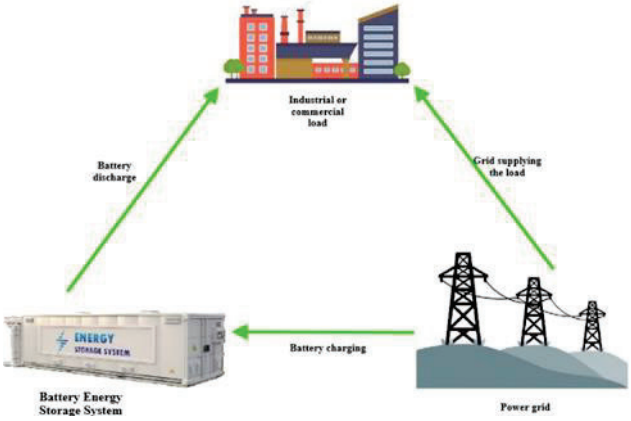


Fig.1 Operation of the battery storage system in the power grid

II. Overview of Existing Research

Energy storage systems (BESS) based on electrochemical storage devices are called batteries that use electrical energy to carry out chemical reactions, transform it into chemical energy (charge) and return electrical energy (discharge) [6]. Currently, the most popular types of storage devices are:

- lithium-ion batteries (Li-ion) are currently the most popular choice for modern energy storage systems due to their high energy density, efficiency, and service life;

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- lithium-iron phosphate (LFP) - known for their safety, long service life and thermal stability, they are usually used in stationary BESS;

- nickel Manganese Cobalt (NMC) battery offers higher energy density than LFP, but it is more expensive and has a shorter lifespan than LFP and requires careful temperature control to protect against overheating;

- sodium batteries (Na-ion) are a promising alternative to lithium-ion batteries (Li-ion), providing lower manufacturing costs, environmental performance due to the use of sodium instead of scarce metals such as sodium and cobalt, but they cannot yet provide the same or higher density as lithium-ion batteries (Li-ion).

Currently, modern developments are focused on autonomous and hybrid energy systems, ways to improve their optimization, stability and efficiency. In general, stand-alone photovoltaic systems with energy storage (BESS) show quite high efficiency in regions with limited access to centralized power supply grids. An example of modeling such systems in Matlab/Simulink is well described by Puneet Raj and Kirti Pal, which confirmed the model's ability to fully meet energy needs using only a solar power plant and energy storage [7].

The importance of energy storage systems in the future is well analyzed in the paper by Harshal Patel and Khalid Simjee [8]. They highlight the key role of BESS systems in keeping power grids stable and balancing energy generation and consumption, especially with the current growth of renewable energy.

Thus, existing studies highlight the significant potential of energy storage systems and the need for further scientific development and research. To optimize, improve the efficiency and stability of the energy system, which is a critically important component in the context of the ever-increasing integration of renewable energy sources, as demonstrated in the ALPHAESS article [9].

In summary, the analysis of recent publications shows that energy storage systems are a strategic element of modern energy systems. BESS provide reliability, flexibility, and economic benefits in the current transition to renewable energy sources. Therefore, further research should be directed toward improving methods for modeling and designing energy storage systems, reducing their cost, and fully integrating them into distribution power grids.

III. Evaluation of BESS Performance with Traditional Power Generation Sources

Amidst the ever-increasing demands for reliability and flexibility in power systems, the integration of energy storage systems into power grids with traditional generation sources is becoming particularly relevant. Combining these sources with BESS can significantly improve the reliability and economic performance of power grids [10].

A significant advantage of energy storage systems is their ability to respond instantly to load changes and smooth out peak electricity consumption by charging batteries during low load periods and releasing energy during peak load periods on the grid. In other words, they compensate for insufficient generation capacity from traditional sources without the need to start up additional capacity at power plants, reducing fuel consumption and extending the operating life of equipment at power plants, as shown in Figure 2. Therefore, BESS integration stabilizes voltage in power grids, especially under conditions of uneven load, with the energy storage system acting as a buffer between the generation source and the consumer (Figure 3). In the case of integration with non-constant generation sources such as solar power plants or wind power plants [10, 11].

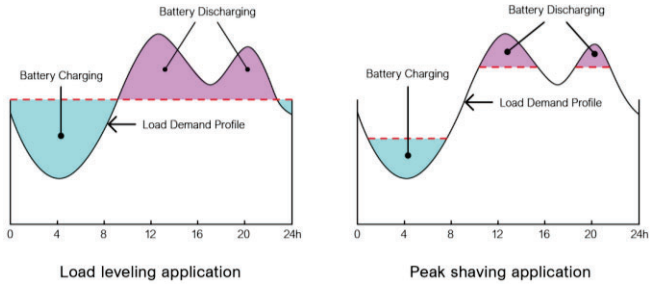


Fig. 2 BESS operation in load leveling mode and during peak loads [8]

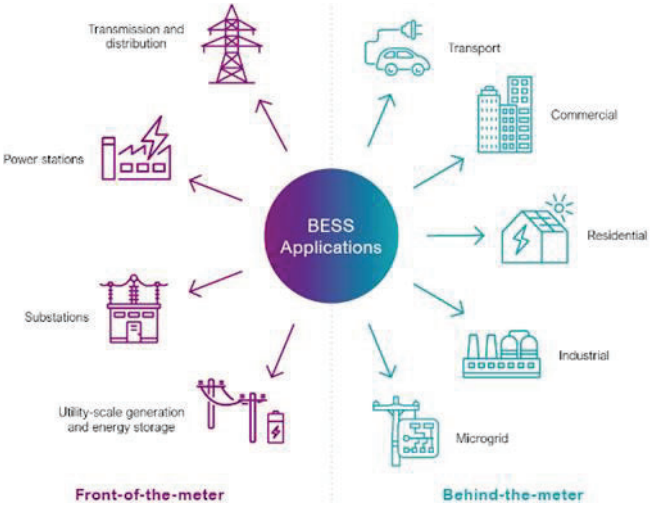


Fig. 3 BESS operation with generating stations and end consumers [8]

This allows compensating for the instability of generation by adapting power consumption profiles, enabling BESS to act as a balancing element that helps avoid over generation or under generation of electricity. Therefore, in this work, we are engaged in the research and implementation of the BESS model in the MATLAB/Simulink environment.

IV. Selection of Energy Storage System Capacity

Before starting the modeling, let's consider the parameters for determining the required battery capacity for reliable operation. The main objective is to ensure that energy consumption is covered in accordance with the calculated energy storage system and to meet the internal needs of the system [10, 12].

The key parameters influencing the choice of storage capacity are:

- load coverage, i.e., the storage capacity must compensate for power shortages or surpluses in the power system;

- BESS operating requirements (inverters, controllers, and other equipment).

Figure 4 shows the general structural diagram of BESS, which includes a storage battery, voltage inverter, controller, and transformer.

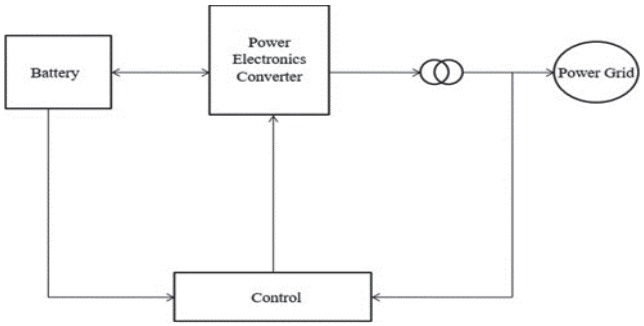


Fig. 4 General structural diagram of BESS [11]

Next, we will consider the method for calculating the required capacity of a BESS. The energy storage subsystem, the inverter, and the maximum permissible current of the storage batteries, which must be calculated, determine the power of a BESS.

To determine the minimum permissible capacity of the storage battery, it is first need to find the nominal voltage of the direct current circuit $U_{dc\ nom}$, V

$$U_{DC\ nom} = U_{bat\ nom} \cdot N, \quad (1)$$

where $U_{bat\ nom}$ – nominal voltage of a single battery, V; N – number of batteries connected in series.

N is set so that the voltage in the DC circuit is kept within the allowed limits at different battery charge levels.

$$\begin{Bmatrix} U_{bat\ min} \cdot N > U_{min\ allow} \\ U_{bat\ max} \cdot N < U_{max\ allow} \end{Bmatrix}, \quad (2)$$

where $U_{bat\ min}$ – minimum voltage of a battery, V; $U_{bat\ max}$ – maximum voltage of a battery, V; $U_{min\ allow}$, $U_{max\ allow}$ – allowed DC bus voltage range, V.

The capacity of a single series-connected storage unit ($C_{\text{min capacity}}$, Ah) is defined, as the minimum permissible value required ensuring that the storage subsystem delivers the specified power (P_{bat}, kW). This parameter reflects the necessary energy reserve that allows the battery to provide the specified power during operation, taking into account the voltage level and discharge duration.

$$C_{min\ capacity} = \frac{100 \cdot P_{bat}}{C_{rate} \cdot U_{DC\ nom}}, \quad (3)$$

where C_{rate} – charge/discharge rate in per unit/hour.

Next, calculate the nominal capacity of one serial battery cell ($C_{nom\ bat\ capacity}$, Ah) using the formula:

$$C_{nom\ bat\ capacity} = C_{bat\ nom} \cdot n \cdot m, \quad (4)$$

where $C_{bat\ nom}$ – nominal battery capacity, Ah; n – number of battery cells connected in parallel in one module; m – number of batteries connected in series and connected in parallel with each other;

For determining the energy capacity of the energy storage system selected based on the primary condition, the following formula is used:

$$E_{BESS} = C_{nom\ bat\ capacity} \cdot N \cdot \frac{U_{bat\ nom}}{1000} = C_{nom\ bat\ capacity} \cdot \frac{U_{DC\ nom}}{1000}, \quad (5)$$

where E_{bess} – energy capacity of the BESS according in kWh; $C_{\text{nom bat capacity}}$ – nominal capacity of one serial battery cell; N – number of batteries connected in series; $U_{bat\ nom}$ – nominal voltage of a single battery, V; $U_{dc\ nom}$ – nominal voltage of the direct current circuit, V.

This formula shows that the total energy stored in the BESS depends on the actual capacity in ampere-hours and the total voltage of the series-connected battery chain.

Also, in order to fully implement the functions of energy storage systems and ensure that the system can operate efficiently, it is necessary to fulfill the condition of BESS operation at each moment of time t, within a predetermined time interval [t_1; t_2], described with the following formula:

$$E_{bess\ min} \le E_{bess}(t_1) - \int_{t_1}^{t_i} P_{bat}(t)dt \le E_{bess\ max}, \quad (6)$$

where $E_{bess\ min}$ – lower limit of change in accumulated charge BESS, kWh; $E_{bess}(t_1)$ – starting value of accumulated charge BESS, kWh; $P_{bat}(t)$ – storage system capacity, kW; $E_{bess\ max}$ – upper limit of change in accumulated charge BESS in operating mode, kWh.

Therefore, when choosing the capacity of an energy storage system, it is necessary to take into account both technical characteristics and consumption patterns. Correct calculations ensure the efficient and reliable operation of the energy storage system in real conditions.

V. Design and Simulation of the BESS

The design and modeling of the energy storage system was carried out using MATLAB/Simulink software. This model is based on a mix of math functions and simple logic blocks that simulate how the energy storage system works [7, 11]. The model has three main blocks: the grid, the load, and the energy storage device. To show the load graph, we used the standard Simulink/Sources/From Workspace block.

In the model configuration parameter, should specify the solution mode with a fixed step of 1. This is necessary for hourly display of load graphs.

The load profile was chosen as the basis: office building, recorded at hourly intervals, shown in Figure 5 and Table 1, where the first row shows the hours and the second row shows the load (kWh). These parameters were, then added to MATLAB and linked to the From Workspace block.

TABLE I. The optimal angle of inclination of the PV module and the angle of inclination of the sun for each month

T	**0**	**1**	**2**	**3**	**4**	**5**	**6**	**7**
***P*, kWh**	100	120	110	125	125	130	140	191
T	8	9	10	11	12	13	14	15
***P*, kWh**	250	464	518	544	588	565	562	607
T	16	17	18	19	20	21	22	23
***P*, kWh**	576	494	409	401	150	120	110	101

The load graph shown in Figure 5 consumes a total of 7,500 kWh per day and has a peak power of 607 kW. For this purpose, an energy storage system model was designed in MATLAB/Simulink. The simulated BESS is designed to smooth out peak electricity consumption by storing excess energy during times of low load on the grid and releasing it during times of increased demand.

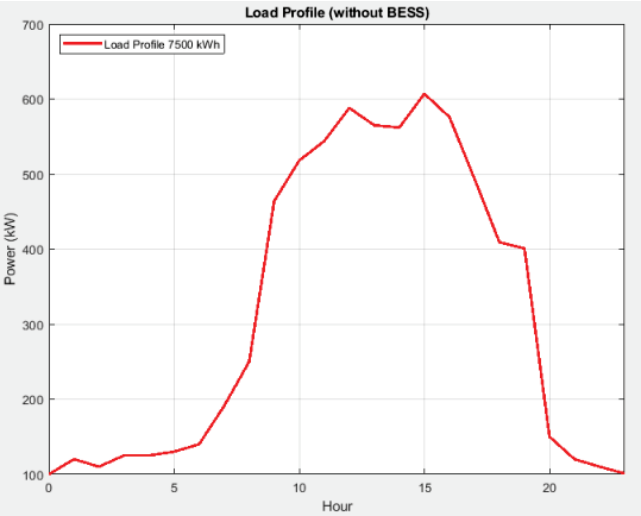


Fig. 5 Daily load graph for an office building

Figure 6 shows the implemented model. This model takes into account the dynamics of battery discharge and charge, considering the load profile, which in turn allows the effectiveness of the model to been assessed [12]. The main parameters used are battery capacity, efficiency, maximum charge and discharge power, minimum acceptable battery charge level, and parameters responsible for discharge and charge time.

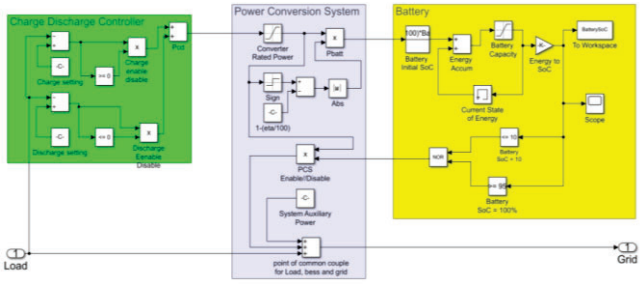


Fig. 6 General view of the BESS model

The BESS model consists of three control blocks, namely a discharge and charging controller, an energy converter, and a storage battery [13]. Together, they make sure the energy storage system works right, when the grid power is limited and the load changes.

Controller of discharge and charge controls how BESS discharges and charges the battery [14, 15]. The controller is configured in such a way that when the power consumption is less than the set value, the system switches to the battery-charging mode.

Switching the charge on and off, taking into account the systems own needs, is defined in (7). Where the charging power is subtracted from the consumption power and if the result is positive, the charge controller starts charging the battery, and if the result is negative, the controller stops charging the battery [16].

$$C_{en} = \begin{cases} 0, if (P_{ch\ set} - P_{load}) < 0 \\ 1, if (P_{ch\ set} - P_{load}) \geq 0 \end{cases}, \quad (7)$$

where C_{en} – charging enable signal; P_{set} – charge power setting, kW; P_{load} – current load power consumption, kW.

In addition, the discharge controller determines when the consumption exceeds the set discharge power, it automatically switches to the battery discharge mode to smooth the load, which in turn reduces the use of energy from the grid. The logic of disabling and enabling the discharge process is presented in formula 8, in which the value of the discharge power is subtracted from the current energy consumption [17, 18]. If the result of this subtraction is negative, the controller receives a discharge signal and gives the necessary power from the battery, but if the resulting value is positive, the controller stops discharging the battery.

$$D_{en} = \begin{cases} 0, if P_{dis\ set} - P_{load} < 0 \\ 1, if P_{dis\ set} - P_{load} \geq 0 \end{cases}, \quad (8)$$

where D_{en} – discharge enable signal; $P_{dis\ set}$ – configured discharge power setting, kW; P_{load} – current load power consumption, kW.

The output power from the charge-discharge controller P_{cd} is defined in formula 9, where the signals C_{en} and D_{en} are complementary, which ensures that the battery will never be charged and discharged at the same time [19, 20]. The addition block combines the output charge and discharge powers into one control signal, which is transmitted to the input of the energy converter.

$$P_{cd} = [(P_{ch\ set} - P_{load})\ C_{en}] + [(P_{dis\ set} - P_{load})\ D_{en}], \quad (9)$$

This approach allows us to efficiently manage the battery and ensure that the BESS operates effectively in accordance with the specified parameters.

The energy converter module is located between the charge-discharge control, battery, load, and network. This module directs the controller output to charge the battery or discharge the accumulator to power the load. Power from the charge-discharge controller in kW is supplied to the input of the converter module [21, 22]. The power passes through the "Saturation" block, which corresponds to the nominal value of the converter, i.e., the maximum value that the BESS model can process.

One of the most important parameters of the energy converter model is its energy conversion efficiency, which takes into account losses during battery charging and discharging. The effect of efficiency consumed from the battery or consumed from it is described by:

$$P_{batt} = \left| sign\left(1 - \frac{\theta}{100}\right) \right| \cdot P_{cd}, \quad (10)$$

where P_{batt} – actual power delivered to or drawn from the battery; P_{cd} – output power from the charge-discharge; Θ – power conversion efficiency (%), typically in the range of 80-95; *sign* – +1 during charging, –1 during discharging.

The battery module accepts an integer value in kW, which is received every hour from the energy converter module. A positive value indicates charging, while a negative value indicates discharging [23, 24]. The Initial Condition block allows you to set the initial charge level of the BESS before commissioning.

Our model provides memory cells that store the current battery status, expressed as a percentage. This battery status is

recorded in the controller's memory for further analysis and feedback to the energy conversion module [25, 26].

The battery capacity in the model is limited by setting upper and lower charge thresholds. A minimum charge level of 10% has been set to prevent deep discharge, which can negatively affect the battery life [27, 28]. Feedback to the converter module is generated based on the battery charge. That is, when the battery charge level is less than 10% or exceeds 100%, the converter module is automatically turned off by transmitting a zero logic signal PC Sen, which is implemented in formula 11.

$$PCS_{en} = \begin{cases} 1, & if\ 10 < SoC < 100 \\ 0, & otherwise \end{cases}, \qquad (11)$$

where PCS_{en} – logical signal to enable the power conversion system; SoC – battery state of charge, %.

This condition ensures that the controller will not continue to charge the battery after reaching 100% charge, and will not allow deep discharge below 10%. This approach is reasonable, ensures battery life, and allows the energy storage system to operate stably.

VI. BESS MODEL SETTINGS AND RESEARCH RESULTS

The settings menu was created to facilitate working with the BESS model configuration. It allows you to quickly make changes to the main parameters before starting the system simulation or during the calibration of simulation scripts.

The main parameters of the energy storage system include:

- nominal power of the BESS in kW. That is, the maximum charging or discharging power that the converter can provide. In our case, it is set at 300 kW;

- converter efficiency, which takes into account power losses during battery charging and discharging. We have assumed a value of 90% as this is typical for industrial converters;

- the total battery capacity used in our model is 2000 kAh;

The power consumption of the BESS itself is taken into account at 5%, which covers the system's own needs for cooling, protection, monitoring, distribution devices, etc.

The settings sub-item also includes the settings for the charge-discharge controller parameters, which are responsible for switching the BESS to charge or discharge mode [29, 30]. In our case, the controller is controlled by two parameters. The first is responsible for charging the battery when the load is below 380 kW, at which point the model uses excess power to charge the battery. The second is responsible for discharging the battery when the load exceeds 430 kW, in which case the model maintains the load using the battery. There is also a 50 kW difference between the modes, which avoids frequent switching between BESS operating modes and increases system reliability.

One of the most common scenarios for using BESS is to limit peak loads on the grid [31, 32]. This method is implemented by pre-charging the battery during periods of low load on the power grid, followed by its use during peak load hours.

To test the model presented in Figure 6, we used the load graph shown in Table 1 and Figure 5. The graph shows that the increased load begins after 9 a.m. and continues until 7 p.m. Total electricity consumption per day is 7500 kWh.

Figure 8 shows the results of BESS performance modeling. The initial load profile has a daily maximum of 607 kW, which was reduced to 430 kW after the integration of the storage system. The graph shows the dynamics of the SoC battery charge level, which changes throughout the day.

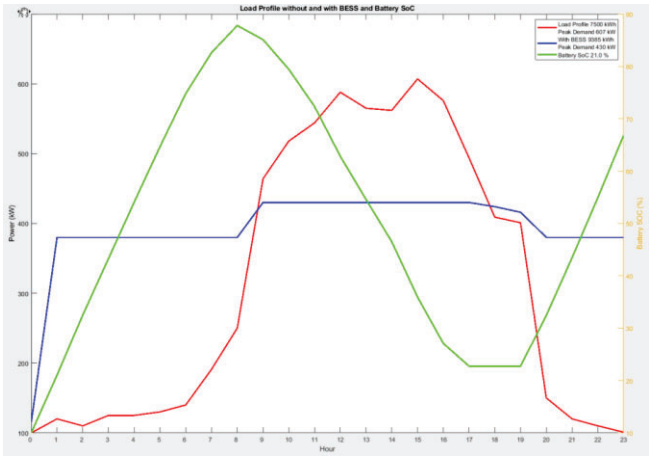


Fig. 8 Working with BESS with and without peak load smoothing, and battery charge level dynamics

During the night and morning hours, the load on the grid remains below the charging threshold, so the charge-discharge controller activates the battery charging process [33, 34]. As a result, the battery accumulates 87% of its charge, after which, during periods of increased load, the BESS switches to discharge mode. During this period, the load from the grid is maintained at 430 kW, while the battery charge level gradually decreases but does not reach the minimum threshold, stopping at 21% before the end of the peak load.

As shown in Figure 8, after 7 p.m., the load decreases and the battery switches to charging mode, which lasts until the end of the day. Therefore, it can be noted that, unlike the programmed discharge threshold of 10%, we managed to maintain a higher charge level. This means that the BESS will reach capacity faster and be ready for a new cycle.

Thanks to the integration of the BESS model, we were able to reduce consumption from 607 kW to 430 kW and stabilize it at this level throughout the entire period of increased load. This result confirms the effectiveness of implementing an energy storage system to smooth out peak loads on the power grid.

CONCLUSION

This article presented an energy storage model using MATLAB/SIMULINK software to study its effectiveness in reducing peak loads on the grid. This model includes basic BESS components such as a charge-discharge controller, battery, and converter.

Modeling using a typical load graph showed that BESS can reduce the load on the power grid from 607 kW to 430 kW, which is a reduction of more than 29% while ensuring stable operation of the system in cyclic mode. That is, charging during low load hours and discharging during high load hours. At the same time, the battery charge level did not fall below the set threshold of 10%, which indicates the effectiveness of the implemented model.

The results obtained confirm the feasibility of using BESS to smooth out peak loads in the grid and improve energy efficiency. The proposed model can be used for further research on energy consumption optimization or for integration with renewable energy sources.

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Influence of Substrate Temperature on Structural and Electrophysical Properties of Thin Mo Films Obtained by DC Magnetron Sputtering

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***Abstract*—This study explores thin molybdenum (Mo) films deposited by DC magnetron sputtering on polycrystalline Al_2O_3 substrates for microelectronics. Mo films exhibit high conductivity, thermal stability, and corrosion resistance, making them suitable for semiconductor contacts, such as in CdTe-based devices. Two sample sets were prepared: at room temperature and with substrate heating (130°C). XRD confirmed a cubic Mo phase, with heated substrates showing improved crystallinity and reduced microstrains. SEM revealed smooth surfaces, while EDS detected minor oxygen at grain boundaries without oxide formation. Heated substrates yielded films with resistivity of 0.4×10^{-4} Ω·cm lower than room-temperature films (2.5×10^{-4} Ω·cm)—approaching bulk Mo values (5.2×10^{-6} Ω·cm). The enhanced performance is attributed to better adatom mobility and adhesion. These results demonstrate the potential of DC magnetron-sputtered Mo films as low-resistance, high-quality contact layers for microelectronics and photovoltaics.**

Keywords—molybdenum thin films, magnetron sputtering, polycrystalline alumina, electrical resistivity, microelectronics.

I. Introduction

Thin molybdenum coatings obtained by DC magnetron sputtering represent a promising solution for modern electronics industry, particularly in Ukraine [1-3]. Due to its physicochemical properties, molybdenum demonstrates high temperature stability, excellent electrical conductivity and corrosion resistance, making it an ideal material for microelectronic devices.

The use of DC magnetron sputtering as a low-temperature method ensures uniform and high-quality deposition of thin films on various substrate types, particularly on polycrystalline alumina Al_2O_3. This technology is economically advantageous as it reduces energy consumption and overall production costs through lower resource usage and reduced need for high processing temperatures.

For Ukraine's electronics industry, this method holds special significance as it enables production of high-quality coatings with minimal costs, which is crucial for developing competitive manufacturing processes. Moreover, molybdenum films deposited this way demonstrate excellent adhesion to substrates, important for their further operation under high-temperature conditions [4, 5, 22].

Among potential applications of molybdenum is its use as a metal contact layer for semiconductors like cadmium telluride, which is actively used in solar energy and optoelectronics. Due to its high conductivity, molybdenum ensures stable contact with semiconductors, positively affecting the efficiency of electronic devices.

DC magnetron sputtering also enables precise control over thickness and stoichiometry of created thin films, which is critical for microelectronics requiring high precision in thin film structure creation. This allows formation of reliable metallization layers that maintain their properties under functional loads.

The high reliability and stability of molybdenum coatings facilitate their long-term use in complex electronic systems, which is relevant for devices operating under challenging conditions. Thus, using molybdenum as a metallization material opens new opportunities for development of Ukraine's electronics industry, improving product quality and its competitiveness in international markets.

II. Experimental

Thin Mo films were deposited by DC magnetron sputtering on commercially available polycor BK-100-1 substrates measuring 14×25 mm (polycrystalline Al_2O_3). The substrates were chemically cleaned in a diluted solution of H_2O:H_2O_2:H_2SO_4 (2:1:1 ratio), then dried in a stream of isopropyl alcohol vapor.

Deposition was performed in a VUP-5m magnetron sputtering system using a Mo target and argon as the sputtering gas. The substrate holder was equipped with 2 holders, each capable of holding 3 substrates. The base pressure in the vacuum chamber was 4×10^{-3} Pa. The sputtering power varied from 30 to 125 W, with working pressure ranging from 0.6 to 1 Pa. The total deposition time was 8 minutes. The samples differed in the substrate temperature during the sputtering process - the samples of the first series were obtained at room temperature, and the

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samples of the second series were obtained after heating the substrate material to a temperature of 130 ^{0}C. Immediately after deposition, the ~0.5 μm thick samples were sent for X-ray structural analysis and electrophysical property investigations.

III. Results and Discussion

The fabricated samples were examined using a DRON-4M diffractometer. Table 1 and Figure 1 present processed experimental data for samples obtained with substrate heating to 130°C [6, 8].

X-ray phase analysis determined the phase composition of these metallic films as the stable cubic Mo phase (JCPDS No. 42-1120). The X-ray diffraction patterns show peaks from (110) and (211) crystallographic planes at 2θ angles of 40.32° and 73.19° respectively. Other visible peaks on the diffraction patterns represent reflections from polycrystalline Al_2O_3. The varying intensities of these peaks are explained by differences in substrate material production batches and the material's high level of texturing.

The peak observed at 2θ angles of 38.5-42.5° is an integrated reflection combining signals from both polycrystalline Al_2O_3 and the molybdenum film. In diffraction patterns of Mo films deposited at room temperature, a halo from fine-crystalline phase was detected in this angular range (38.5-42.5°), which cannot be analyzed for structural parameters. For the 130°C Mo films, additional analytical processing was performed to separate these peaks and determine their parameters.

The results of X-ray diffraction studies were analytically processed and the values of the coherent scattering regions and microstresses were obtained for each family of crystallographic planes.

An increase in intensity and decrease in integral half-width of peaks is observed on X-ray diffraction patterns of thin Mo films both without preliminary heating and with substrate material heated to 130°C.

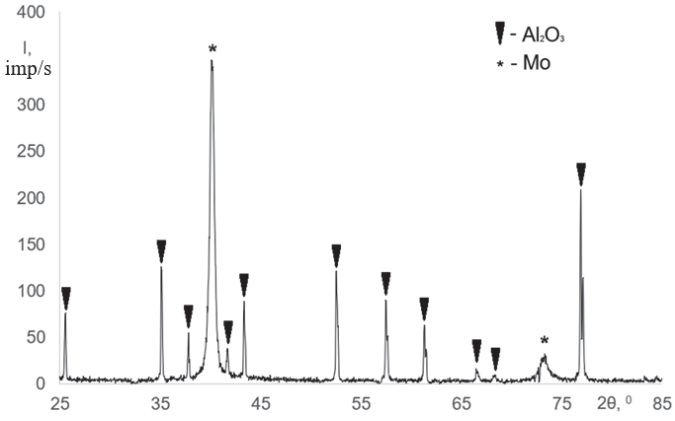


Fig. 1. Typical X-ray diffractogram of a thin Mo film on a polycor substrate

These changes can be explained by better adsorption of nucleation states to the polycor surface under heated substrate conditions, compared to magnetron sputtering on unheated substrates at room temperature, when atoms of the sputtered material lack sufficient free energy for surface migration in search of energetically favorable incorporation into the crystal structure. The calculated crystalline sizes measure 28 nm and 14 nm for the (110) and (211) crystallographic plane families, respectively.

The microstrain values indicate significant tensile stresses ($4.65\text{-}5.79\times10^{-3}$ a. u.) in the metallic film, caused by the high texturing of the substrate material. However, when using elevated substrate temperatures during DC magnetron sputtering, the combined effect of polycor's thermal expansion coefficient and the high free energy of sputtered material enables formation of cubic-phase Mo metal layers with strong adhesion.

The polycor BK-100-1 substrates (designated ЩЕ7.817.000) employed in this study possess critical parameters influencing Mo film formation [13]. The high chemical purity (≥99.7% Al_2O_3 content) effectively prevents parasitic chemical reactions during sputtering. The polished surface micro-roughness (Rz = 0.1 μm, Ra = 1.6 μm) [14] contributes to uniform Mo adatom adhesion, consistent with the defect-free morphology observed in SEM images (Fig. 2) [9, 10].

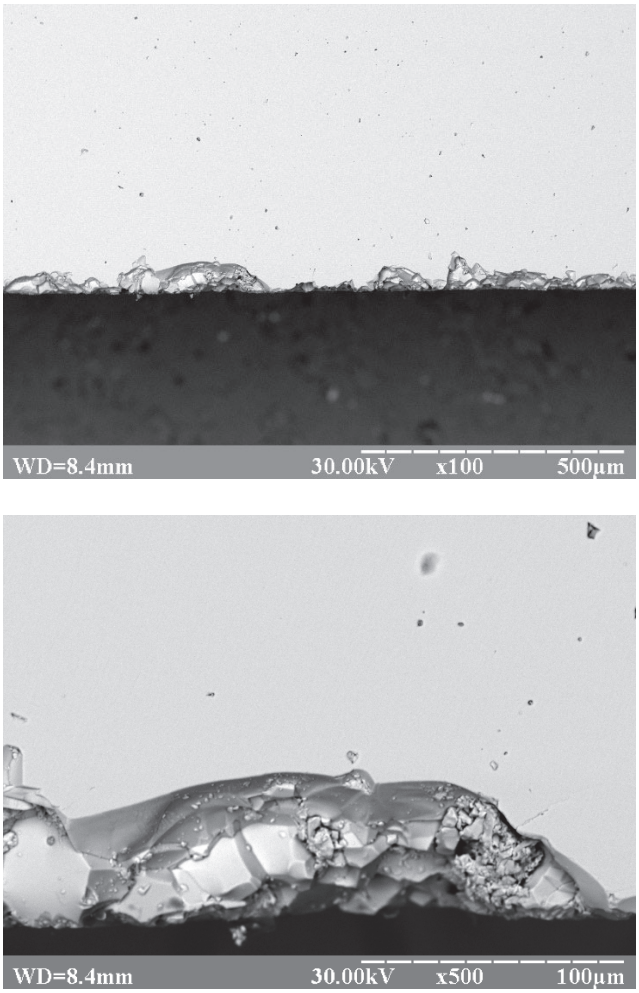


Fig. 2. Results of studying the fracture of a thin Mo film on a polycor substrate by SEM

The thermal expansion coefficient ($\alpha = 80\times10^{-7}$ K^{-1} within 20-900°C range) governs the thermomechanical behavior of the film-substrate system. At the deposition temperature of 130°C, the mismatch between α_{Mo} (54×10^{-7} K^{-1}) and $\alpha_{Al_2O_3}$ generates compressive stresses on the order of 10^{-3}, as confirmed by XRD data in Table 1 ($\Delta d/d = 4.65\text{-}5.79\times10^{-3}$). The substrate's mechanical strength (bending

strength of 313.8 MPa) prevents deformation under thermal loads during processing.

The combination of zero water absorption and high volume resistivity (ρ_V ≥2×10^{14} Ω·cm at 150°C) ensures measurement accuracy by eliminating parallel current paths through the substrate during film resistivity characterization. The ceramic's high thermal conductivity (≥24 W/(m·K) in comparable grades promotes uniform temperature distribution during heating, explaining the reduced crystallographic orientation gradients observed in films (Fig. 1). The low dielectric loss (tanδ ≤0.0001 at 8-10 GHz) indicates potential for high-frequency device applications.

The results obtained through scanning electron microscopy (SEM) of the thin Mo film surface are shown in Figure 2.

This figure displays both the film surface and a portion of the cross-section fracture. The image reveals the quality and grain size within the film thickness near the substrate interface, showing their gradual transition to a continuous metallic layer at the film surface. The film surface appears smooth, mirror-like, without topographic variations.

The density of visually detectable non-through defects on the film surface was 1 per 5000 μm². [18, 20]

TABLE I. RESULTS OF ANALYTICAL PROCESSING OF THE X-RAY DIFFRACTOGRAM OF THE MO SAMPLE AT A SUBSTRATE TEMPERATURE OF 130°C

hkl	2θ, ^{0}	d, nm	I, imp/s	FWHM, ^{0}	D, nm	Δd/d·10^{-3}, a. u.
(110)	40,32	0,224	25,11	0,59	28	4,65
(211)	73,19	0,129	23,22	1,34	14	5,79

Results of additional investigation of the Mo film atomic composition by energy-dispersive X-ray spectroscopy are presented in Figure 3. The analysis shows that during the formation of the metallic film by DC magnetron sputtering, an interaction occurs between the material at the grain boundaries of the film and adsorbed oxygen. The oxygen in the vacuum chamber may result from trace amounts of water present in the argon (working gas) atmosphere, which dissociates and is absorbed by the Mo film [11, 12].

The formation of oxides was not detected by X-ray structural analysis methods, therefore it should be assumed that only minor amounts of oxygen and possible oxide formation are present on the film surface and/or within the grain boundaries of the film.

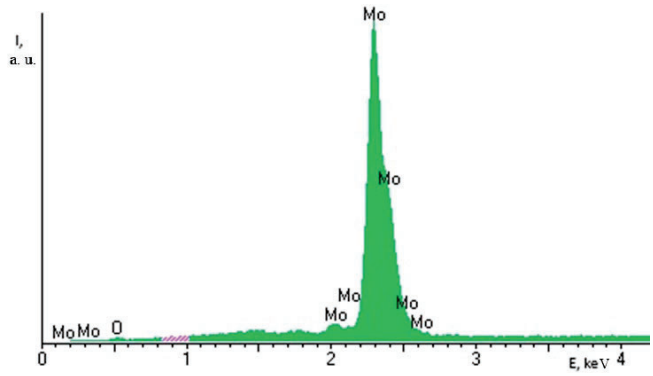


Fig. 3. Results of investigating the surface of a thin Mo film on a polycor substrate by energy-dispersive X-ray spectroscopy

To control the resistivity of metallic thin Mo films for their application in microelectronic devices, resistivity measurements were conducted using the four-point probe method. The studies revealed that molybdenum films deposited at room temperature exhibited a resistivity of 2.5×10^{-4} Ω·cm, while those obtained at 130°C showed 0.4×10^{-4} Ω·cm.

These results indicate that the enhanced structural perfection of metallic films achieved through substrate heating above 100°C helps reduce contact resistance by one order of magnitude compared to bulk Mo (5.2×10^{-6} Ω·cm) [23-24]. This finding confirms that heating the polycor substrate above 100°C during DC magnetron sputtering enables the formation of non-amorphous, continuous molybdenum film layers suitable for applications in microelectronic devices.

DC magnetron sputtering demonstrates distinct advantages over chemical vapor deposition (CVD) for molybdenum contact layers in CdTe photovoltaics [15-17, 25-27, 35]. Whereas conventional CVD processes (e.g., using $Mo(CO)_6$ precursors) achieve higher deposition rates (~200 nm/min) [28], they incur significant drawbacks: precursor costs exceed $800/kg [29], and require complex gas-handling systems to manage pyrophoric compounds (CAS 13939-06-5). In contrast, our DC sputtering process achieves industrially viable deposition rates (65 nm/min) at 0.4× the operational cost, primarily due to:

- Elimination of toxic precursors;
- Reduced energy consumption (no precursor decomposition required);
- Simplified vacuum infrastructure (base pressure 4×10^{-3} Pa suffices).

Critically, sputtered films deposited at 130°C exhibit superior electrical properties ($\rho = 0.4\times10^{-4}$ Ω·cm) compared to typical CVD-Mo contacts ($\rho = 1.2–5.0\times10^{-4}$ Ω·cm) [30]. This correlates with enhanced CdTe device efficiency: Han et al. [31] reported 16.2% efficiency for sputtered-Mo contacts versus 14.7% for CVD counterparts, attributable to reduced contact resistance (Rc < 10^{-3} Ω·cm²). The preferential (110) orientation in our XRD data (Fig. 1) further explains this advantage, as highly textured Mo layers minimize carrier scattering at grain boundaries [32].

Thermal stability remains a concern for both methods. While CVD films show delamination above 450°C due to carbon contamination [33], our sputtered films maintain adhesion at 130–150°C operational temperatures (validated by SEM cross-sections, Fig. 2). For high-temperature applications (>300°C), however, CVD may retain an advantage in conformal coverage on nanostructured substrates [34-36].

IV. Conclusion

In this work, the physical and technological parameters of DC magnetron sputtering for obtaining thin metallic molybdenum films on polycor substrates are described.

Among the obtained results, it should be noted that heating the substrate material above 100°C reduces the specific resistance of formed Mo metal layers, increases the level of crystalline perfection and enhances adhesion to polycor [21, 22]. The insignificant amount of molybdenum oxides on the surface of formed layers can be removed by chemical-mechanical polishing methods or ion-plasma surface etching during preparation for subsequent technological operations and is not critical for the structure's resistance value.

With a deposition rate of ~65 nm/min, cubic-phase Mo film condensates were obtained, exhibiting specific resistance of 4×10^{-5} Ω·cm. Such films can be used as contact layers in microelectronic industry, for example as contacts for photovoltaic converters based on CdTe.

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Algorithms for measuring lower limb angles for monitoring human gait using computer vision

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***Abstract*—This study explores low-cost, marker less options for monitoring knee flexion during gait. This study examines a deep-learning approach that uses BlazePose to derive hip-to-ankle key-points directly from RGB video, a classical Canny plus Hough pipeline that, after a one-time user initialization, tracks thigh-and-shank lines at more than 60 FPS on a CPU, and propose a hybrid strategy in which the neural network supplies periodic re-initializations while the classical tracker maintains high-rate tracking. Preliminary tests indicate each pathway can keep angular error within 5–8 degrees, suggesting all three merit further validation and possible extension to stereo or monocular 3-D reconstruction and remote-monitoring workflows. These approaches are particularly relevant for rehabilitation monitoring of patients with prostheses and orthoses, where a non-invasive and low-cost method can significantly improve long-term assessment. The possibility to run both pipelines on ordinary laptops and even smartphones highlight their potential for telemedicine and home-based gait tracking.**

Keywords - gait analysis, Human Pose Estimation (HPE), BlazePose deep-learning model, Canny–Hough classical tracking, Part Affinity Fields (PAF) grouping, Monocular 3-D reconstruction with Statistical Parametric Human Body Model (SMPL)

I. Introduction

The analysis of human gait is critical for the detection and monitoring of many neurological and orthopedic pathologies. Studies have shown that changes in gait parameters can precede clinical manifestations of diseases for years, serving as biomarkers of disorders in the corresponding structures of the nervous system [1]. Traditionally, gait kinematics is measured using hardware systems, such as optical systems with markers or wearable sensors. However, such solutions are expensive, require complex setup, and can be burdensome for patients (marker placement, calibration, etc.). As an alternative, more attention is being paid to marker less methods based on computer vision, which allow automatic estimation of a person's pose and movements from ordinary video. This is especially relevant for people with limited mobility, where non-invasive gait monitoring in the home or clinic will significantly improve rehabilitation assessment. In clinical biomechanics, special attention is paid to patients with prostheses or orthoses, since their gait differs significantly from healthy individuals. Most open-source pose estimation models are trained on healthy people and may fail when applied to atypical silhouettes or pathological movements. Therefore, developing adaptable algorithms that remain accurate for patients with mobility limitations is a pressing research need, and low-cost markerless solutions can expand access to rehabilitation monitoring.

Two approaches are considered in the current literature to solve the problems of estimating the peculiarities of human lower limb movement: traditional and deep learning methods. In the traditional approach, Human Pose Estimation (HPE) tasks are solved by classical computer vision algorithms. They use cascade feature-based detectors (Haar, HOG) to detect people or limbs, as well as graphical body models such as pictorial structures that optimize the location of body segments by minimizing energy. Such approaches have limited accuracy and robustness: they are sensitive to changes in lighting, background, clothing, and angles. For example, the Histogram of Oriented Gradients (HOG) method detects pedestrians well against a contrasting background, but in complex scenes with body parts interacting, its effectiveness decreases. In general, classical algorithms provide high CPU speed and ease of implementation but are significantly inferior to modern neural network models in terms of the accuracy of human joint positioning.

Deep 2D methods have been actively developed over the past decade and now dominate HPE. Deep neural network architectures can be of two types: top-down and bottom-up. In the top-down approach, the whole person is first detected (for example, via Faster R-CNN or YOLO), and then the neural network determines the coordinates of the skeletal joints within the found boundig box. This approach provides high accuracy for a single object but is dependent on the quality of the detection: if the detector misses a person or part of a limb, the pose estimation stage will be wrong. Bottom-up approaches, on the contrary, first find all the key points (joints) in the image and then group them by people using associative features such as Part Affinity Fields. They are better suited for crowded scenes and do not scale directly with the number of people but may be less accurate for a single person due to the global nature of the estimation. In general, current research offers numerous neural network models for 2D pose estimation: OpenPose is the first open real-time system for multi-human pose estimation [2]; AlphaPose is an accurate top-down method for regional body estimation [3]; R-CNN Keypoint Mask adds keypoint heads to the object detector [4]; HRNet is a high-precision architecture while maintaining high resolution of the feature map [5]; MediaPipe Pose is a lightweight model from Google optimized for mobile devices [6], etc. Thanks to large datasets (COCO, MPII, Human3.6M) and GPUs, these models provide impressive accuracy and robustness to various shooting conditions. In particular, BlazePose (Google MediaPipe) is capable of real-time detection of 33 key points on the human body on a mobile phone with a frequency of >30 frames, which opens the way to massive use in portable applications [7]. Recent studies also emphasize the importance of adapting neural pose models to pathological gait [8] and integrating gait monitoring into telemedicine platforms [9]. These works underline the

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growing demand for lightweight, accessible tools that can operate outside of specialized laboratories.

Assessing postures in three-dimensional space is the next step required for a full biomechanical gait analysis. There are multi-camera systems (stereo) and depth cameras such as Microsoft Kinect that directly provide 3D joint coordinates [10]. However, such solutions are only affordable in the laboratory (expensive equipment, sensitivity to conditions). Alternatively, monocular 3D reconstruction is possible - when 2D key points are first extracted from a regular video, and then their depth is restored using pre-trained neural network models or statistical body models (e.g., SMPL). Examples of such approaches: VideoPose3D - takes a sequence of 2D coordinates and produces joint trajectories in 3D; HMR, SPIN - regresses the parameters of a 3D human body model from a single RGB image. Monocular methods are cheaper and easier to deploy, but less accurate than stereo or depth systems, especially with complex movements and occlusions. In the context of tracking the gait of people with prostheses, current work recommends focusing on 3D analysis if high-precision parameters are required or combining 2D+3D approaches. At the same time, 2D data (joint flexion angles), which are quick and easy to obtain, are often sufficient for basic rehabilitation monitoring.

Most open source pose models are trained on datasets of healthy people. This poses a challenge for their application in the biomechanics of pathological gait: for example, a neural network may incorrectly recognize the postures of a limb with a prosthesis or orthosis. The researchers note that for the algorithms to work correctly, either model adaptation (fine-tuning on special data) or additional engineering solutions are required, such as imposing biomechanical constraints, filtering the results, etc. For example, standard models are often confused in the case of an atypical leg silhouette with a complex orthotic device. Therefore, when building a gait pattern tracking system for people with limited mobility, special attention should be paid to its adaptability. It is recommended to use open frameworks (OpenPose, MMPose, etc.) with the possibility of retraining on patients' own videos, as well as to provide for the transition to 3D reconstruction of steps if high accuracy is required. In addition, the system should operate in real time on available hardware (e.g., laptop or smartphone), which requires optimization and the ability to select a simpler pose assessment mode. That is why it is advisable to develop hybrid approaches that combine the advantages of deep models (full automation, stability) and classical algorithms (CPU speed, user control).

In this paper, we propose two complementary implementations of a gait monitoring system. The first solution uses a state-of-the-art pose estimation model (BlazePose) and fully automatically calculates the knee joint flexion angle at each video frame. The second solution deliberately abandons neural networks, relying on classical computer vision algorithms (Canny, Hough) to track leg segments with minimal user initialization. Below, we describe both approaches and their hybrid combination, present experimental verification results, and discuss future prospects.

II. Description Of The Software Solution

The first approach implements automatic video analysis using Google's MediaPipe BlazePose model. This deep neural network model determines in real time the coordinates of 33 key points on the human body in each frame, including the hip and shin joints needed to estimate the knee angle. The system's algorithm can be represented as a sequence of blocks (Fig. 1): (1) a frame is read from the video stream; (2) the BlazePose neural network finds human poses - the coordinates of the hip, knee, and ankle joints (for both legs); (3) based on these coordinates, the angle between the hip and the lower leg is calculated. In particular, the angle at the knee is defined as the angle between the hip vector (from the hip joint to the knee) and the shin vector (from the knee to the ankle) in the frame plane. The scalar product of the vectors is used for the calculation. At the output of each frame, the system produces an angle value (in degrees); thus, a time series (graph) of angle changes during the step is formed. In the implementation developed by the author, this series is automatically exported to a CSV file for further analysis. It is important to emphasize that the depth model works completely automatically, without the need for any markup from the user: it is enough to point the camera at the patient, and the algorithm independently "tracks" his or her feet. For clarity, visualization can be superimposed on top of the video - a skeleton of connected joints and a display of the current angle value (for example, in text or graph). Fig. 2 shows an example of a frame with a BlazePose skeleton overlaid and a measured knee angle. The performance of this approach depends on computing resources: on a modern GPU, it runs in real time (30+ FPS), while on a CPU, performance may be reduced. Nevertheless, the advantage is versatility: the neural network correctly recognizes joints even with foreign objects in the frame, different clothes, and typical changes in lighting, as proven by previous studies [11]. To improve reliability, the implementation provides filtering by the confidence factor that the model assigns to each point found: only joints with >50% confidence is considered, otherwise the angle is not calculated for a given frame (to avoid noise spikes).

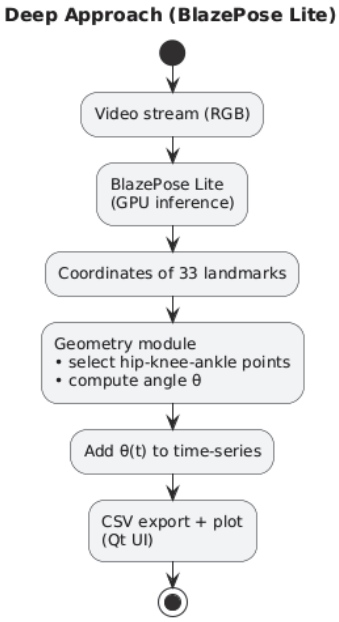


Fig. 1. Deep approach (BlazePose Lite)

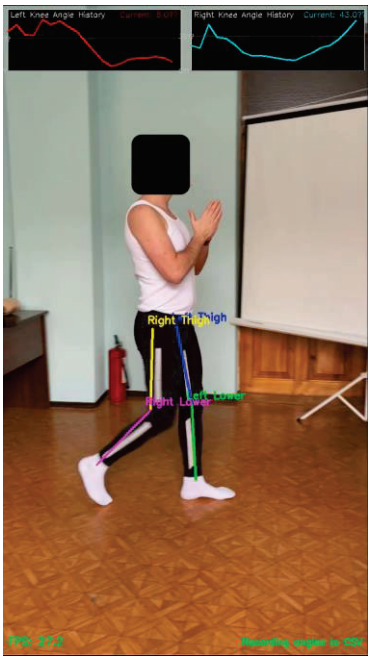


Fig. 2. Example of a frame with a BlazePose skeleton overlaid

Fig. 3 plots the bend of the left knee from frame to frame. The repeating highs and lows mark each step, showing the person walked at a steady pace.

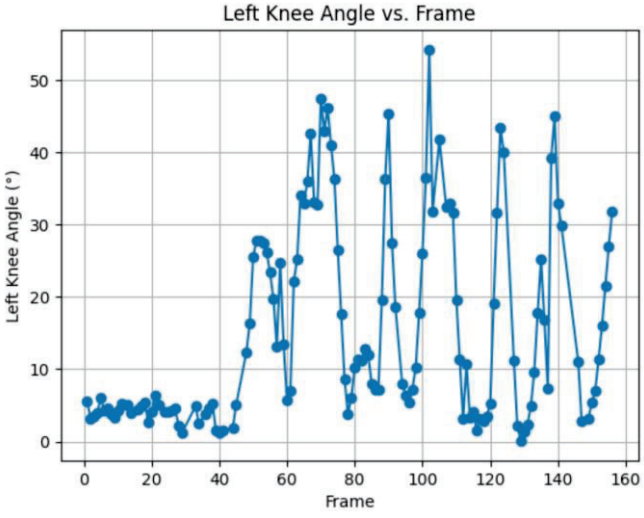


Fig. 3. Left-knee flexion angle versus frame number

Fig. 4 does the same for the right knee. Its curve looks similar, but the peaks are a bit lower, meaning this leg didn't flex quite as far as the left.

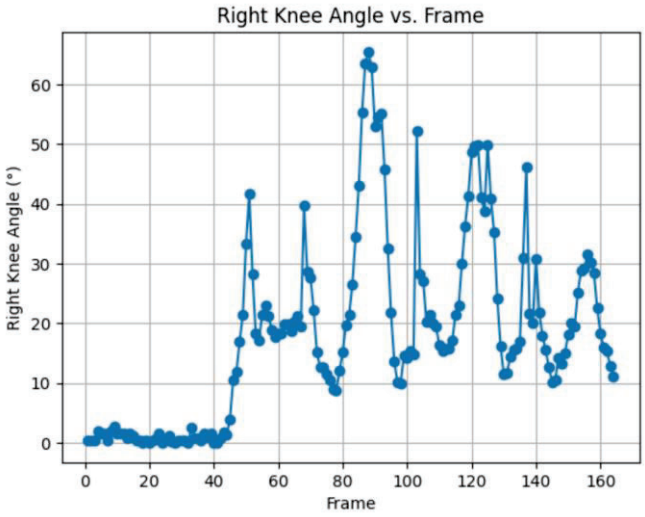


Fig. 4. Right-knee flexion angle versus frame number

Fig. 5 puts the left-knee angle on one axis and the right-knee angle on the other. The data trace a loop, a simple picture of how the two legs move out of phase yet stay in a regular rhythm—evidence of a stable, balanced gait.

The oscillations of knee flexion angles in Fig. 3 and Fig. 4 clearly correspond to the swing and stance phases of gait, confirming the ability of the algorithm to capture rhythmic walking patterns. In Fig. 5, the loop trajectory highlights the out-of-phase coordination between legs, which is a standard biomarker of balanced, symmetric gait.

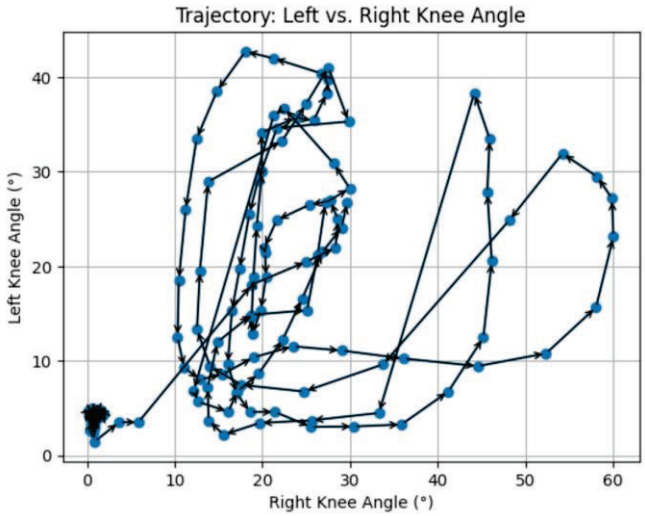


Fig. 5. Left-knee angle plotted against right-knee angle (cycle loop)

This paper also implements a second approach aimed at situations where a powerful GPU is not available, and the video stream needs to be processed at a high frame rate. The method is entirely based on classical image processing and has been tested on video of a robotic arm: this has safely allowed us to polish the geometric procedures before transferring them to video of human lower limbs. The algorithm (Fig. 6) starts with a one-time initialization: the user draws two tentative lines along the robotic arm links (analogous to the future "thigh" and "leg") in the first frame. Then each frame is processed automatically: the contour map is generated by the Canny detector, long elongated segments are selected from the resulting boundaries, and the adaptive Hough transform recovers straight lines corresponding to the axes of the selected links. The lines are sequentially compared with the

previous frame by position and slope, which ensures stable tracking, after which the angle between them is calculated and the frame values are saved to CSV; the angle graph and superimposed lines are displayed in the PyQt interface in real time (Fig. 7).

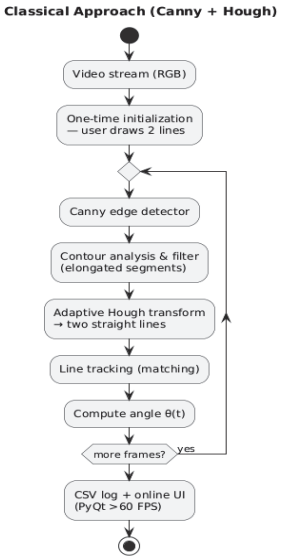


Fig. 6. Classical approach (Canny + Hough)



Fig. 7. Example of a frame with Canny + Hough approach

Fig. 8 shows the angle of the robotic arm relative to the horizontal line during a motion. The waveform is almost sinusoidal. This regular swing confirms that the arm base stays fixed at 90°, while the base joint makes periodic movements. The quasi-sinusoidal waveform in Fig. 8 illustrates the periodicity of motion. Such regularity is expected in normal human walking as well, which supports the transferability of the classical pipeline from robotic-arm validation to human lower-limb monitoring.

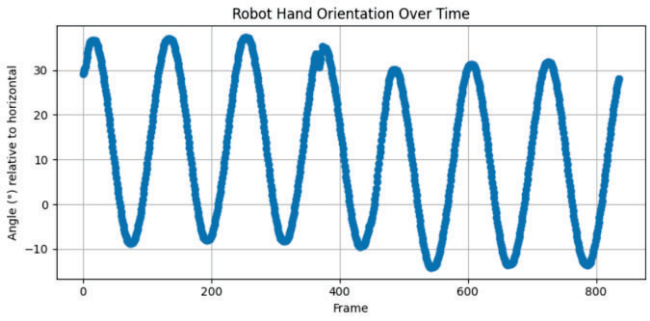


Fig. 8. Robot-hand orientation vs. time (frame index)

Due to the simplicity of operations, one frame is processed in a few milliseconds, so even on an average CPU, more than 60 FPS is achieved. This makes the approach suitable for low-cost single-board computers or field measurements without a hardware accelerator. However, the method has a number of limitations. Manual initialization requires accuracy: an error in the starting location of the lines can lead to loss of tracking. The algorithm is sensitive to Canny and Gap parameters, and they need to be adapted to specific shooting conditions (contrast, background, lighting). If there are foreign objects or overlapping lines, tracking can also fail, because the method does not use a high-level model of the human body.

Both implementations demonstrated stable operation with angular deviations within 5–8°, which is sufficient for basic clinical gait assessments and rehabilitation monitoring. The deep learning approach minimizes user intervention and is robust to environmental factors, while the classical Canny–Hough pipeline ensures high speed and low hardware requirements. Taken together, these results confirm the feasibility of low-cost computer vision tools for tracking gait in patients with limited mobility.

III. Conclusions

This paper presents two implementations of a computer vision system for automated gait pattern tracking focused on measuring lower limb flexion angles in people with limited mobility. The first approach is based on the BlazePose deep neural network and provides fully automatic angle detection from 2D human videos. The second approach uses classical algorithms (Canny + Hough) for fast tracking of leg segments and can work on less productive hardware in real time. Both methods showed accurate measurement results on the test videos: the error between them is a few degrees, which is acceptable for practical purposes of biomechanical gait analysis. The depth approach has the advantage of minimizing user involvement and better robustness to complex conditions (lighting, background, crowded frame), while the classical approach is characterized by high performance and ease of setup.

After analyzing the advantages and disadvantages of the two approaches, a hybrid idea is to combine them into a single system to compensate for the weaknesses of each. On the one hand, BlazePose provides full automation and robustness to complex cases, but it can be slow on some devices and sometimes inaccurately detects joints in people with prostheses. On the other hand, the OpenCV tracker is extremely fast and works even on the simplest hardware, but it needs initialization and may lose the target. The hybrid mode involves the following: the neural network module is used for rough initial pose estimation in the initial frames (for example, it automatically detects the position of the hip and lower leg and initializes the lines), after which the classical module refines and tracks the position of these lines between the joints in subsequent frames. In fact, BlazePose can “mark” the patient's legs once, and then a faster algorithm takes over and tracks the angles. If necessary (for example, if the classic tracking goes astray or the person leaves the frame and returns), the neural network detector can be periodically turned on for reinitialization. Thus, the user has the freedom to choose whether they can work in BlazePose offline mode, or in OpenCV economy mode, or combine them. The combination of “deep” and “traditional” logic is expected to increase the system's reliability: the neural network will insure the algorithm against gross errors, and the classic tracker will smooth out noise and speed up work on large videos. The proposed idea opens the way to adaptive gait monitoring

systems that can be flexibly configured to the patient's resources and specifics. Soon, we plan to implement such a hybrid algorithm and test its effectiveness on patient video data. Fig. 9 schematically demonstrates the concept of integrating the two approaches.

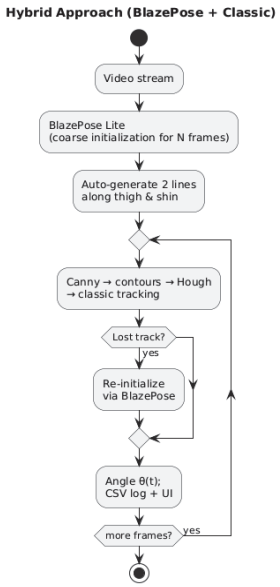


Fig. 9. Hybrid Approach (BlazePose + Classic)

In the future, it is planned to expand the system's functionality: to implement stereo imaging to reconstruct full 3D joint angles, to integrate a camera distance calibration module to estimate absolute spatial parameters of the step, and to establish remote data transmission (telemetry) to use the system in telemedicine rehabilitation protocols. This will turn high-precision gait analysis into a truly massive, affordable, and mobile technology suitable for widespread implementation in clinical practice and home monitoring of patients.

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