Electrodynamics of Axial-Flow Rotary Blood Pumps

IGOR V. NESTERENKO1, ALEXANDER A. PUGOVKIN1,2, STEFFEN LEONHARDT3, (Senior Member, IEEE), MARIAN WALTER3, (Senior Member, IEEE), PATRICK BORCHERS3, ALEKSANDR MARKOV2, JAMSHID H. KARIMOV4, SERGEY V. SELISHCHEV1, (Senior Member, IEEE), AND DMITRY V. TELYSHEV1,2, (Member, IEEE)

1Institute of Biomedical Systems, National Research University of Electronic Technology, Zelenograd, 124498 Moscow, Russia
2Institute for Bionic Technologies and Engineering, I. M. Sechenov First Moscow State Medical University, 119991 Moscow, Russia
3Medical Information Technology, Helmholtz Institute of Biomedical Engineering, RWTH Aachen University, 52074 Aachen, Germany
4Department of Biomedical Engineering, Lerner Research Institute, Cleveland Clinic, Cleveland, OH 44195, USA

Corresponding author: Alexander A. Pugovkin (pugovkin@bms.zone)

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ABSTRACT In this work, electric machine of implantable rotary blood pumps (RBPs) of axial-flow type was theoretically investigated. Electromagnetic coupling of rotor and stator in axial-flow RBPs was described with the Maxwell’s equations of the classical electrodynamics given quasi-magnetostatic approximation of electromagnetic field in the simplified model consisting of permanent magnet and conductive loop rotating in free space with one spatial degree of freedom. Additionally, relative field error, created by the neglect of geometric deviations introduced through manufacturing tolerances, was estimated for a typical axial-flow RBP. Upper limit of error introduced by the simplifications was estimated less than 0.2 %, leading to the finite accuracy of the description and clearly determining the influence on the control signal. Based on the presented description and additional engineering considerations, the electric machine of axial-flow RBPs was defined as a three-phase non-salient pole synchronous machine with a permanent magnet rotor. Two key features were shown: a) unlike the conventional electric motors, signal of back electromotive force tends to be a sinusoidal waveform in any construction of axial-flow RBP with significant non-magnetic gap; b) the optimal waveform of control signal in this case is sinusoidal. Initial design and control parameters of the electric machine in axial-flow RBPs can be accurately determined with presented theoretical description. Based on the description, control system of an axial-flow RBP with the optimal waveform of control signal can be developed.

INDEX TERMS Axial-flow, back electromotive force, electrodynamics, permanent magnet rotor, rotary blood pump, sinusoidal waveform, synchronous motor, ventricular assist device.

I. INTRODUCTION

According to recent reports, about 64.34 million people worldwide are suffering the heart failure, and this number is expected to further increase in the near future [1], [2]. Due to donor hearts scarcity, mechanical circulatory support (MCS) is the frequently used treatment for patients suffering the end-stage heart failure. Among other systems of MCS, continuous-flow left ventricular assist devices (LVADs) are applied in clinical practice most often [3]–[7]. LVAD acts as a bypass device pumping additional amount of blood from the left ventricle to the aorta. It shows high durability and can easily be implanted in the patient’s thorax. Inflow and outflow cannulae are also required to ensure proper connection to the cardiovascular system.

Generally, continuous-flow LVADs are rotary blood pumps (RBPs) of axial- or radial-flow type. Basically, RBP is an electric motor with a path for blood flow generated by a rotating impeller driven through electromagnetic coupling with stator windings. Rotating impeller instead of a shaft is the essential distinction of RBP and conventional electric motor. In axial-flow RBPs, blood flow is created along
the pump central axis. Autonomous operation of an RBP is implemented by a controller supplied with rechargeable batteries as a source of electric energy. Percutaneous cable provides the electric connection between RBP and controller.

General principle of operation in axial-flow RBPs is conversion of supplied electric energy into mechanical energy of rotation and, eventually, into hydraulic energy of blood flow. Therefore, structure of rotary blood pumps separates into electric machine and hydraulic machine coupled through mechanical machine [8]–[16]. The electric machine consists of electromagnetically coupled stator and rotor. Bearings of various types represent the mechanical machine. And major elements of the hydraulic machine are impeller and diffuser, producing hydraulic work determined by the pressure head across the pump $H_h$ and the generated blood flow rate $Q_h$.

The sequential chain, as shown in Fig. 1, represents energy conversion and losses occurring during operation of axial-flow RBPs. Since the influence of interconnection between the structural units is negligible, each unit can be considered independently through the analysis of input and output parameters.

For conversion of electric energy into mechanical energy, an external magnetic field modulated by relative rotor position is necessary in the stator. Therefore, rotor should be a permanent magnet rigidly embedded into the impeller and driven by the electric current in stator windings, providing rotational movement with angular velocity $\omega$ (Fig. 2). This solution does not require connection of rotor circuit, as for electromagnetic rotor, greatly simplifying RBP design and increasing operational reliability. The torque $T_1$ created by the electric machine is not equal to the torque $T_2$ applied to the impeller, since it is partially compensated by the drag torque in the front and rear impeller bearings. The consistent nature of energy conversion also determines a relatively high level of losses due to their accumulation at each stage. In addition, efficient operation of the hydraulic machine with certain parameters requires a significant non-magnetic gap between the permanent magnet and the stator windings (in comparison to industrial electric motors).

A deep, fundamental description of the electromagnetic coupling between stator and rotor in terms of electrodynamics is important to determine the electric machine type and to optimize the signals of RBP control system. However, to authors’ knowledge, derivation of the electrodynamic characteristics of axial-flow RBPs was not yet presented in literature.

In this paper, we describe the electromagnetic coupling of stator and rotor in axial-flow RBPs based on the Maxwell’s equations given the commonly utilized simplifications. We estimate upper limit of error introduced by these simplifications, leading to the finite accuracy of presented description and determining influence on the control signal. Also, we theoretically justify the optimal control signal of axial-flow RBPs as sinusoidal waveform. Finally, we show and describe the main difference of RBPs from industrial electric motors and determine the actual type of RBP electric machine as a three-phase non-salient pole synchronous machine with a permanent magnet rotor.

II. ELECTROMAGNETIC APPROACH

The electromagnetic interaction in a medium with polarization and magnetization is described by the Maxwell’s equations represented in differential form as follows (Gauss’ law for electric fields is (1), Ampere’s law is (2), Faraday’s law is (3) and Gauss’ law for magnetic fields is (4)) [17]:

$$\nabla \cdot (\varepsilon_0 \vec{E} + \vec{P}) = \rho, \quad (1)$$
form of the Poynting’s theorem:

\[ \nabla \times \vec{H} = \vec{J} + \frac{\partial}{\partial t} (\varepsilon_0 \vec{E} + \vec{P}), \]  
\[ \nabla \times \vec{E} = -\frac{\partial}{\partial t} \mu_0 (\vec{H} + \vec{M}), \]  
\[ \nabla \cdot \mu_0 (\vec{H} + \vec{M}) = 0, \]  

where \( \vec{E} \) and \( \vec{H} \) represent the electric and magnetic field strengths, respectively; \( \vec{P} \) is the electric polarization field; \( \vec{M} \) is the magnetization field; \( \rho \) is the volumetric density of electric charge; \( \vec{J} \) is the electric current density; and \( \varepsilon_0, \mu_0 \) are the permittivity and permeability of free space, respectively.

Possibility of conversion of electromagnetic energy into mechanical energy is based on emergence of the Lorentz force described by (5) for the case of continuous charge and current distribution in the environmental volume, where \( \vec{f} \) represents the emerging force and \( \vec{B} \) is the magnetic induction:

\[ \vec{f} = \rho \vec{E} + \vec{J} \times \vec{B}. \]  

The magnetic induction is related generally to the magnetic field strength by the following equation:

\[ \vec{B} = \mu_0 (\vec{H} + \vec{M}). \]  

Further description of the energy conversion in axial-flow RBPs is based on the energy conservation law in differential form [18] of the Poynting’s theorem:

\[ \nabla \cdot \vec{S} + \frac{\partial w}{\partial t} + P_C = 0, \]  

where \( \vec{S} \) is the density of energy flux [19]; \( w \) is the spatial density of energy; and \( P_C \) is the spatial density of power converted from one form to another.

If (2) and (3) are multiplied in dot product by \( \vec{E} \) and \( \vec{H} \), respectively, and then added up, as described in [17], [20] and [21], the following is obtained:

\[ -\nabla \cdot (\vec{E} \times \vec{H}) = \frac{\partial}{\partial t} \left( \frac{1}{2} \varepsilon_0 \vec{E} \cdot \vec{E} \right) + \frac{\partial}{\partial t} \left( \frac{1}{2} \mu_0 \vec{H} \cdot \vec{H} \right) + \vec{E} \cdot \frac{\partial \vec{P}}{\partial t} + \vec{H} \cdot \frac{\partial \mu_0 \vec{M}}{\partial t} + \vec{E} \cdot \vec{J}. \]  

Terms of (8) correspond to (7) as follows:

\[ \vec{S} = \vec{E} \times \vec{H}, \]  
\[ \frac{\partial w}{\partial t} = \frac{\partial}{\partial t} \left( \frac{1}{2} \varepsilon_0 \vec{E} \cdot \vec{E} \right) + \frac{\partial}{\partial t} \left( \frac{1}{2} \mu_0 \vec{H} \cdot \vec{H} \right), \]  
\[ + \vec{E} \cdot \frac{\partial \vec{P}}{\partial t} + \vec{H} \cdot \frac{\partial \mu_0 \vec{M}}{\partial t} + \vec{E} \cdot \vec{J}. \]  

This helps to describe (8) in terms of energy conservation law, matching (7) and (8). The first and the second terms on the right side of (8) correspond to the rate of density change of the stored energy for electric and magnetic field components, respectively. The third and the fourth terms show the rate of density change of the energy stored in medium due to its polarization and magnetization. And the last term describes the electromagnetic energy conversion from one form to another at the considered point in space.

A. FEATURES AND SIMPLIFICATIONS IN AXIAL-FLOW ROTARY BLOOD PUMPS

Thus, (1)–(5) and (8) describe electromagnetic interactions in general form for an arbitrary environment, in cases the laws of classical electrodynamics are applicable. Further, these relations are applied for processes occurring in the axial-flow RBPs.

Based on the above qualitative description of the RBPs, a simplified model of electromagnetic coupling is proposed (Fig. 3) to construct analytical description. This model consists of a rigid loop made of an ohmic conductor with relatively high conductivity, powered by electric current and surrounded by a material with relatively low polarization and conductivity. The conductive loop is electromagnetically coupled with a permanent magnet and has one spatial degree of freedom, i.e. rotates about a fixed axis in a magnetic field of the magnet, with significant non-magnetic gap between the loop and the magnet.

This consideration allows to apply further simplifications. In the fully static state of the system with the loop powered with DC, energy of the magnetic field considerably prevails over the electric energy, since the voltage drop produced by high conductivity of the loop is insignificant and the capacitive coupling between the loop segments is negligible compared to the inductive coupling. This behavior of the electromagnetic system creates the prerequisites for considering it quasi-magnetostatic. Because dimensions of RBPs should not exceed \( l_{\text{max}} = 10 \text{ cm} \) and expected maximum fundamental frequency of the supply current is \( v_{\text{max}} = 10000 \text{ min}^{-1} \), we can estimate the relative error \( \delta \) caused by this...
approximation, as described in [20]:

\[ \delta = \mu_0 \varepsilon_0 L^2 \frac{v^2}{v_{\text{max}}^2} \approx 10^{-15}. \quad (12) \]

This estimated error is negligible, allowing to use the quasi-magnetostatic approximation and, therefore, to exclude the displacement currents in (2) from consideration.

Significant gap between the permanent magnet and the conductive loop influences the density of magnetic field energy in the volume of materials with weak magnetic properties, making the magnetic circuit linear. In the RBPs, gap between the magnet rotor and the stator windings is significant, so the magnetic field leakage is negligible, and the magnetic field strength \( \vec{H} \) depends only on the terminals connecting the loop to an external current source. Therefore, another simplification is that permanent magnet with residual magnetization is the only element of the system having pronounced magnetic properties.

The electric current flowing through the material of conductive loop is described by the following equation, also known as Ohm’s law:

\[ \vec{E} = \frac{\vec{J}}{\sigma}, \quad (13) \]

where \( \sigma \) is the conductivity of the loop material.

Given the introduced simplifications and (13), (1)–(4) and (8) take the following form:

\[
\begin{align*}
\nabla \cdot \varepsilon_0 \vec{E} &= \rho \\
\nabla \times \vec{H} &= \vec{J} \\
\n\nabla \times \vec{E} &= -\frac{1}{\mu_0} \frac{\partial \vec{H}}{\partial t} + \vec{M} \\
\n\nabla \cdot \mu_0 \vec{H} &= 0 \\
-\nabla \cdot (\vec{E} \times \vec{H}) &= \frac{1}{\mu_0} \frac{\partial}{\partial t} \left( \frac{1}{2} \mu_0 \vec{H} \cdot \vec{H} \right) + \vec{H} \cdot \frac{\partial \vec{M}}{\partial t} \\
&\quad + \frac{\vec{J} \cdot \vec{J}}{\sigma} + \vec{\omega^\prime} \cdot \vec{\omega^\prime}.
\end{align*}
\]

Both last terms of (15) correspond to the term \( P_C \) in (7) and to \( \vec{E} \cdot \vec{J} \) in (8). The first of them is the energy dissipation due to the heat produced by electric current flowing through the loop material. And scalar product of the torque density \( \vec{\omega} \) and the angular velocity field \( \vec{\omega}^\prime \) introduces power consumption of the work performed during the relative rotation of the system elements about a fixed \( z \) axis.

**B. DEVELOPMENT OF ANALYTICAL DESCRIPTION**

Further, (15) is integrated over a limited volume \( V \), since the current in the loop is limited by the finite volume of the loop itself, and a finite volume boundary, where the magnetic field strength \( \vec{H} \) becomes negligible, can always be found. Thus, difference between integral over the limited volume \( V \) and integral over infinite volume is always less than any predetermined positive number. Therefore, the integral form of the energy conservation law is:

\[ - \int_V \nabla \cdot (\vec{E} \times \vec{H}) \, dV = \int_V \left( \frac{1}{\mu_0} \frac{\partial}{\partial t} \left( \frac{1}{2} \mu_0 \vec{H} \cdot \vec{H} \right) + \vec{H} \cdot \frac{\partial \vec{M}}{\partial t} \right) \, dV \]

\[ + \int_V \frac{\vec{J} \cdot \vec{J}}{\sigma} \, dV + \int_V \vec{\omega} \cdot \vec{\omega}^\prime \, dV. \quad (16) \]

Then each integral in (16) is considered individually.

1) **ELECTROMAGNETIC ENERGY FLUX**

According to the divergence theorem [18],

\[ \int_V \nabla \cdot (\vec{E} \times \vec{H}) \, dV = \oint_{S_V} (\vec{E} \times \vec{H}) \cdot d\vec{s}, \quad (17) \]

where \( S_V \) is the surface bounding the volume \( V \).

The surface integral in (17) determines the rate of electromagnetic energy transfer from volume \( V \) through the surface \( S_V \) into the external space. Due to quasi-static approximation, there is no electromagnetic radiation. This allows to state that the energy flux through the surface \( S_V \) is nonzero only in the area \( S'_V \), where surface \( S_V \) intersects the terminals connecting the conductive loop to an external current source. According to [21], the integral of \( (\vec{E} \times \vec{H}) \) over the surface \( S'_V \) is opposite to the current in the loop \( I \) multiplied by the difference of potentials \( U_s \) between terminals connecting the loop to the current source:

\[ \oint_{S'_V} (\vec{E} \times \vec{H}) \cdot d\vec{s} = -U_s I. \quad (18) \]

2) **RATE OF CHANGE OF THE MAGNETIC FIELD ENERGY**

The first term in the right side of (16) is the rate of change of the magnetic field energy that given (6) can be represented as follows:

\[ \frac{dW}{dt} = \int_V \left( \frac{1}{\mu_0} \frac{\partial}{\partial t} \left( \frac{1}{2} \mu_0 \vec{H} \cdot \vec{H} \right) + \vec{H} \cdot \frac{\partial \vec{M}}{\partial t} \right) \, dV = \int_V \vec{H} \cdot \frac{\partial \vec{B}}{\partial t} \, dV, \quad (19) \]

whence the following equation for the energy increment \( \Delta W \) over \( (t - t_0) \) is obtained:

\[ \Delta W = W(t) - W(t_0) = \int_V \int_{t_0}^t \vec{H} \cdot \frac{\partial \vec{B}}{\partial t} \, dt \, dV \]

\[ = \int_V \int_{t_0}^{(t)} \vec{H} \cdot d\vec{B} \, dV. \quad (20) \]

Hereinafter, \( \int_{t_0}^{(t)} \) means integration of the corresponding integrands in terms of time, since all of them are time-dependent quantities. Also, most of explicit designations of time dependence are downed for conciseness.

Since in the considered electromagnetic system the magnetic field is generated by the current flowing in the conductive loop and by the permanent magnet rotor, resulting field \( \vec{H} \) is a superposition of the loop field \( \vec{H}_L \) and the permanent magnet field \( \vec{H}_{PM} \):

\[ \vec{H} = \vec{H}_L + \vec{H}_{PM}. \quad (21) \]
Therefore, change of magnetic field energy in the system splits into three parts: change of the loop field energy \( \Delta W_L \), change of the permanent magnet field energy \( \Delta W_{PM} \), and change of the coupling energy \( \Delta W_{L,PM} \):

\[
\Delta W = \Delta W_L + \Delta W_{PM} + \Delta W_{L,PM}. \tag{22}
\]

Terms in (22) are represented separately given the substitution of (21) and (6) into (20) as follows:

\[
\Delta W_L = \int_V \frac{1}{2} \mu_0 \left( H_L^2(t) - H_L^2(t_0) \right) dV,
\]

\[
\Delta W_{PM} = \int_V \frac{1}{2} \mu_0 \left( H_{PM}^2(t) - H_{PM}^2(t_0) \right) dV + \int_{V_{PM}} \int_{(t_0)}^{(t)} \mu_0 \vec{H}_{PM} \cdot d\vec{M} dV,
\]

\[
\Delta W_{L,PM} = \int_V \mu_0 \left( \vec{H}_L(t) \cdot \vec{H}_{PM}(t) - \vec{H}_L(t_0) \cdot \vec{H}_{PM}(t_0) \right) dV + \int_{V_{PM}} \int_{(t_0)}^{(t)} \mu_0 \vec{H}_L \cdot d\vec{M} dV,
\]

where \( V_{PM} \) is the volume of the permanent magnet as the only material with magnetic properties.

\( \Delta W_L \) is the energy of magnetic field produced by the loop current in free space. The energy of the permanent magnet \( \Delta W_{PM} \) includes the energy of magnetic field produced by the magnet in free space and the energy stored in its magnetized material. Both components of \( \Delta W_{PM} \) have arisen before initial time instant \( t_0 \) and remain constant during operation of the system with permanent magnet driven in a fully reversible recoil region. The first term of the coupling energy \( \Delta W_{L,PM} \) is the interaction between two sources of magnetic field in free space, and the second term is the impact of the loop field on the permanent magnet material, generally causing a displacement of the operating point on its demagnetization curve. This last term is due to the fact that magnet material is not absolutely magnetically hard.

Hence, with constant energy of permanent magnet and zero change of coupling energy in the entire volume \( V \) given one spatial degree of freedom for the conductive loop, total energy change of the system reduces to energy change produced by the loop. Therefore, we obtain

\[
\Delta W = \int_V \int_{(t_0)}^{(t)} \vec{H}_L \cdot d\vec{B} dV. \tag{26}
\]

Differential of magnetic induction \( d\vec{B} \) can be expressed through differential of magnetic vector potential \( d\vec{A} \), as in [20], with the equation:

\[
d\vec{B} = \nabla \times d\vec{A}. \tag{27}
\]

Considering substitution of (27) into (26) and property of divergence \( \nabla \cdot (\vec{H}_L \times d\vec{A}) \), we obtain the following:

\[
\Delta W = \int_V \int_{(t_0)}^{(t)} \vec{H}_L \cdot (\nabla \times d\vec{A}) dV
\]

\[
= \int_V \int_{(t_0)}^{(t)} (\nabla \cdot (\vec{H}_L \times d\vec{A}) + (\nabla \times \vec{H}_L) \cdot d\vec{A}) dV
\]

\[
= \int_V \int_{(t_0)}^{(t)} (\nabla \times \vec{H}_L) \cdot d\vec{A} dV, \tag{28}
\]

The term \( \nabla \cdot (\vec{H}_L \times d\vec{A}) \) is zero. Further, application of the Ampere’s law for magnetostatics, determined by (14), gives:

\[
\Delta W = \int_V \int_{(t_0)}^{(t)} J \cdot d\vec{A} dV. \tag{29}
\]

Since there is no current outside the conductive loop, it is possible to pass from the integral over the volume \( V \) to the integral along the loop \( C \) with current \( I \), and then, using Stokes’ theorem [18], move to the integral over the surface \( S_C \) bounded by \( C \):

\[
\Delta W = \oint_C \int_{(t_0)}^{(t)} id\vec{A} - d\vec{l} = \int_{S_C} \int_{(t_0)}^{(t)} I(\nabla \times d\vec{A}) \cdot d\vec{s}. \tag{30}
\]

Next, substitution of (27) into (30) provides:

\[
\Delta W = \int_{S_C} \int_{(t_0)}^{(t)} id\vec{B} \cdot d\vec{s} = \int_{(t_0)}^{(t)} d\Psi, \tag{31}
\]

where

\[
d\Psi = \int_{S_C} d\vec{B} \cdot d\vec{s} \tag{32}
\]

is the increment of the flux linkage of the loop. Thus, it is shown that the increment of the magnetic field energy in the considered system is equal to the integral of the current in the loop \( I \) over the flux linkage \( \Psi \).

According to [21], the Ampere’s law for magnetostatics (14), the vector potential definition (27), and the Coulomb gauge \( \nabla \cdot \vec{A} = 0 \) provide the equation \( \Delta \vec{A} = -\mu_0 \vec{J} \) that is the Poisson’s equation with the following solution:

\[
\vec{A} = \frac{\mu_0}{4\pi} \int_V \frac{\vec{J}(\vec{r})}{|\vec{r}_0 - \vec{r}_1|} dV, \tag{33}
\]

where \( \vec{r}_0 \) is the radius vector to the observation point; and \( \vec{r}_1 \) is the radius vector to the element \( dV \). Substituting the (33) into (29), we obtain:

\[
\Delta W = \int_V \int_{(t_0)}^{(t)} \vec{J}(\vec{r}_0) \cdot d \left( \frac{\mu_0}{4\pi} \int_V \frac{\vec{J}(\vec{r}_1)}{|\vec{r}_0 - \vec{r}_1|} dV \right) dV. \tag{34}
\]

As it was shown previously, we can go to the integrals over two contours instead of both integrals over the volume \( V \), considering zero electric current at \( t_0 \) and locating the observation point at the conductor axis of the loop:

\[
\Delta W = \oint_{C_0} \int_{(t_0)}^{(t)} id \left( \frac{\mu_0}{4\pi} \oint_{C_1} \frac{d\vec{r}_1}{|\vec{r}_0 - \vec{r}_1|} \right) d\vec{r}_0
\]

\[
= \frac{\mu_0}{4\pi} \oint_{C_0} \oint_{C_1} \int_{(t_0)}^{(t)} d\vec{r}_1 d\vec{r}_0 |\vec{r}_0 - \vec{r}_1|
\]

\[
= \frac{\mu_0}{4\pi} \int_{(t_0)}^{(t)} d\vec{r}_0 \int_{(t_0)}^{(t)} \int_{C_0} \oint_{C_1} \frac{d\vec{r}_1}{|\vec{r}_0 - \vec{r}_1|}. \tag{35}
\]

Contours \( C_0 \) and \( C_1 \) should not coincide, since this will lead to an uncertainty in the result. One of the contours runs along the conductor axis and another one, along the conductor...
surface. Indirectly, this fact confirms that it is impossible to apply the approximation using zero thickness conductors in some cases of electrodynamics.

The double integral in the result of (35) depends only on geometric parameters of the conductive loop and corresponds to the loop inductance $L$. The change in the magnetic field energy of the loop is expressed through the inductance $L$ as follows:

$$\Delta W = \frac{LI^2}{2}. \quad (36)$$

Combining (31) and (36), we obtain:

$$\int_{(i)} \mathcal{J} \cdot \mathcal{J} \, dV = \frac{LI^2}{2}. \quad (37)$$

As a result, we obtain the following equation for flux linkage $\Psi$:

$$\Psi = LI. \quad (38)$$

3) ENERGY DISSIPATION

The following equation holds with the loop made of an ohmic conductor and at zero current outside the loop:

$$\int_V \tau \cdot \phi \, dV = T_z \omega, \quad (40)$$

where $T_z$ is the torque produced by relative rotation of the loop and the permanent magnet about fixed $z$ axis; and $\omega$ is the angular velocity of this rotation.

Further, we consider the loop with the current $I$ (Fig. 4) rotated at angle $d\phi$ relative to the $z$ axis from position $C_0$ to position $C_d\psi$, in the magnetic field $\mathbf{B}$ and apply the approach from [24].

The force exerting the segment of the loop $d\tilde{l}$ in the magnetic field $\mathbf{B}$ is determined by the magnetic component of the Lorentz force in (5) as follows:

$$dF = I \left( d\tilde{l} \times \mathbf{B} \right). \quad (41)$$

The torque produced by the given force relative to the $z$ axis is equal to:

$$dT_z = \tau_z \cdot (\tilde{r} \times d\tilde{l}) = I\tilde{e}_z \cdot (\tilde{r} \times (d\tilde{l} \times \mathbf{B})). \quad (42)$$

Hereinafter, $\tilde{e}_z, \tilde{e}_\rho, \tilde{e}_\phi$ are unit vectors directed along the corresponding axes of the cylindrical coordinate system $(\rho, \phi, z)$. We transform (42) to the cylindrical coordinates with the substitution $\tilde{r} = \rho \tilde{e}_\rho + z\tilde{e}_z$:

$$dT_z = I \left( \tilde{e}_z \times \tilde{r} \right) \cdot (d\tilde{l} \times \mathbf{B}) = I \left( \tilde{e}_z \times (\rho \tilde{e}_\rho + z\tilde{e}_z) \right) \cdot (d\tilde{l} \times \mathbf{B}) = I\rho \left( \tilde{e}_z \times \tilde{e}_\rho \right) \cdot (d\tilde{l} \times \mathbf{B}). \quad (43)$$

Further, given $\tilde{e}_z \times \tilde{e}_\rho = \tilde{e}_\phi$, we obtain:

$$dT_z = I \rho \tilde{e}_\phi \cdot (d\tilde{l} \times \mathbf{B}) = I \rho \tilde{B} \cdot (\tilde{e}_\phi \times d\tilde{l}). \quad (44)$$

Next, we introduce a surface element $d\tilde{s}$ equal to $d\tilde{r} \times d\tilde{l}$ and to $\rho d\phi (\tilde{e}_\phi \times d\tilde{l})$. Therefore, it is possible to make the substitution $\rho (\tilde{e}_\phi \times d\tilde{l}) = d\tilde{s}/d\phi$ into (44) with the following result:

$$dT_z = I \frac{\tilde{B} \cdot d\tilde{s}}{d\phi}. \quad (45)$$

Integrating (45) over the surface $\Delta S$ bounded by $C_0$ and $C_d\psi$, we obtain the equation for the loop torque $T_z$:

$$T_z = \int_{\Delta S} \tilde{B} \cdot d\tilde{s}. \quad (46)$$

Taking (32) into account, one can note that the integral in the resulting part of (46) is the increment of the flux linkage $d\Psi$:

$$T_z = I \frac{d\Psi}{d\phi}. \quad (47)$$

allowing to write (40) as follows:

$$\int_V \tau_z \cdot \phi \, dV = T_z \omega = I \omega \frac{d\Psi}{d\phi}. \quad (48)$$

Analysis of the terms in (16) allows to rewrite it with lumped parameters from (18), (36), (38), (39) and (48):

$$U_s \cdot I = I \frac{\partial \Psi}{\partial t} + I^2 R + \omega I \frac{\partial \Psi}{\partial \phi}. \quad (49)$$

Thence, the equation of voltage drop in the considered loop is obtained:

$$U_s = \frac{\partial \Psi}{\partial t} + IR + \omega \frac{\partial \Psi}{\partial \phi}. \quad (50)$$
The first term describes the voltage drop due to the self-induction phenomenon. The second term corresponds to the voltage drop at the ohmic resistance of the loop, and the third term is the back electromotive force (EMF) produced by the voltage drop at the ohmic resistance of the loop. And the third induction phenomenon. The second term corresponds to the back EMF by the time derivative \( \frac{d}{dt} \psi \).

Further, we determine how the flux linkage
\[
\psi = LI + \psi_{PM},
\]
where \( \psi_{PM} \) is the flux linkage with the magnetic field of a cylindrical permanent magnet with transverse magnetization.

III. DETERMINATION OF ELECTRIC MACHINE TYPE AND OPTIMAL CONTROL SIGNAL WAVEFORM

A. BACK ELECTROMOTIVE FORCE WAVEFORM

Further, we determine how the flux linkage \( \psi_{PM} \) changes during relative rotation of the loop and the permanent magnet with angular velocity \( \omega \). We also determine the waveform of the back EMF by the time derivative \( d\psi_{PM}/dt \). For this, we find analytical expression for spatial distribution of the permanent magnet field with transverse magnetization, shown in Fig. 5. We use scalar potential of the magnetic field, defined through the concept of magnetic charges, as proposed in [25]:

\[
\Phi_{PM}(\vec{r}) = \int_V \frac{\varrho(\vec{r}')}{4\pi|\vec{r} - \vec{r}'|} d^3\vec{r}',
\]
where \( \varrho(\vec{r}') \) is the volumetric density of magnetic charges; \( \vec{r} \) is the radius vector of the observation point; \( \vec{r}' \) is the radius vector to an elementary magnetic charge. Then, we turn to the cylindrical coordinate system \((\rho, \varphi, z)\) and to the surface density of magnetic charges \( \zeta \), determined through the projection of the magnetization vector \( \vec{M} \) as follows:

\[
\zeta(\varphi) = M \cos \varphi.
\]

This transition allows to integrate over the magnet surface:

\[
\Phi_{PM}(\vec{r}) = \frac{M}{4\pi} \int_0^{2\pi} d\varphi' \left( \frac{a \cos \varphi'}{\sqrt{\rho^2 + a^2 - 2\rho a \cos(\varphi - \varphi') + (z - z')^2}} \right) dz'.
\]

where \( a \) and \( L \) are the dimensions of the magnet. The transition to the magnetic field is carried out through the definition of the scalar potential:

\[
\vec{H} = -\nabla \Phi_{PM}.
\]

Equations (55) and (56) show that dependence of spatial distribution of the magnetic field strength on the coordinate \( \varphi \) is sinusoidal for a finite cylindrical magnet. Therefore, given the previous calculations, the expression for flux linkage \( \psi_{PM} \), included in (52), is obtained as follows:

\[
\psi_{PM} = \psi_a \sin \varphi,
\]
where \( \psi_a \) is the flux linkage amplitude determined by the magnetization of the permanent magnet and its geometric parameters.

From (51), (52) and (57), one can obtain the following differential equation describing the processes in the considered system:

\[
U_s = IR + \frac{d}{dt}LI + \omega \psi_a \cos \omega t.
\]

B. SUPPRESSION OF GEOMETRIC DEVIATIONS

Equation (57) is obtained for a permanent magnet of ideal cylindrical shape. However, geometric deviations occur during the RBP assembly as a result of manufacturing tolerance in various parts, such as deviations in permanent magnet shape and/or misalignment between axis of rotation and axis of symmetry of the magnet. These deviations lead to a distortion in the change of flux linkage \( \psi_{PM} \), introducing high-frequency harmonics that destabilize the control and decrease overall efficiency of the system. Hereinafter, we consider that any geometric deviation introducing field error can be reduced to axis misalignment leading to the same error.

Further, we evaluate the magnetic field error produced with this effect by providing the example of the asymmetric installation of the permanent magnet, illustrated in Fig. 6. For this, we again use (53) and the following expansion in a series [26]:

\[
\frac{1}{|\vec{r} - \vec{r}'|} = \sum_{i=0}^{\infty} \frac{\rho^i}{\rho^{i+1}} P_i \left( \cos \left( \frac{\vec{r} \cdot \vec{r}'}{\rho^2} \right) \right).
\]
where \( P_i(\vec{r}, \vec{r}') \) are the Legendre polynomials forming an orthogonal system. As a result, we obtain:

\[
\Phi_{PM}(\vec{r}) = \Phi_{PM}^0(\vec{r}) + \Phi_{PM}^1(\vec{r}) + \frac{1}{4\pi} \sum_{i=2}^{\infty} \int_V a(\vec{r}') \frac{r_i}{r_i+1} \cdot P_i(\cos(\vec{r}\vec{r}')) d^3\vec{r}'.
\]

(60)

The value of \( \Phi_{PM}^0(\vec{r}) \) is zero, since there are no magnetic charges. Given \( P_1(\cos(\vec{r}\vec{r}')) = \cos(\vec{r}\vec{r}') \), \( \Phi_{PM}^1(\vec{r}) \) can be expressed in terms of the dipole magnetic moment \( \vec{m} \):

\[
\Phi_{PM}^1(\vec{r}) = \frac{1}{4\pi r^2} \int_V \vec{r}' \cdot \vec{r} \cos(\vec{r}\vec{r}') d^3\vec{r}' = \frac{1}{4\pi r^2} \int_V \vec{r}' \cdot \vec{r} d^3\vec{r}' = \frac{\vec{r} \cdot \vec{m}}{4\pi r^3}.
\]

(61)

The dipole scalar potential \( \Phi_{PM}^1(\vec{r}) \) corresponds to the scalar potential of an ideally cylindrical shape magnet defined in (55). The dipole magnetic moment \( \vec{m} \) can be expressed for a permanent magnet through the material magnetization \( \vec{M} \):

\[
\vec{m} = \int_V \vec{M} dV.
\]

(62)

For \( r > a \gg r_{err} \), values of the Legendre polynomials \( P_i(\cos(\vec{r}\vec{r}')) \) tend to unity. With \( i \) increasing, the expression \( r_i/r_i+1 \) in (60) that determines the scalar potential of higher-order magnetic moments decreases stronger than the term \( \Phi_{PM}^1(\vec{r}) \). Thus, prevalence of the sinusoidal field component of the rotating permanent magnet increases with distance from the surface of the permanent magnet. Given this and the monotonic change in the scalar potential with increasing distance, we conclude that influence of deviations decreases with increasing non-magnetic gap.

This suppression of deviation influence is inherent for the axial-flow RBPs, in contrast to industrial electric motors with a required minimum air gap in the magnetic circuit that causes relatively high sensitivity to deviations. Also, analysis of (60) shows the impossibility to generate arbitrary waveforms of the flux linkage in axial-flow RBPs (e.g., a waveform close to triangular shape that is necessary for producing the broadband back electromotive force of trapezoidal shape).

The exact solution of (55) and expressions for the magnetic field components of a cylindrical permanent magnet with transverse magnetization are given in [25]. These expressions are relatively complex, and we use some simplification to estimate the error produced by the asymmetric installation of the permanent magnet. Note that this magnetic field decays as a power of reciprocal distance from the magnet axis within the range of 2 (infinite cylindrical permanent magnet [27]) and 3 (point magnetic dipole [27]):

\[
\tilde{H}_c(r > a) \propto \frac{1}{r^2} \quad \text{and} \quad \tilde{H}_d(r > a) \propto \frac{1}{r^3},
\]

(63)

where \( \tilde{H}_c \) is the magnetic field strength of an infinite thin cylindrical magnet; and \( \tilde{H}_d \) is the magnetic field strength of a point magnetic dipole.

Hence, for a finite cylindrical magnet,

\[
\delta_{H}(r, r_{err}) = \left( r - r_{err} \right)^{g} - 1,
\]

(64)

where \( \delta_{H}(r, r_{err}) \) is the relative error of magnetic field value in case of axis misalignment; \( r_{err} \) is the magnet misalignment error; and \( g \) takes the value between 2 and 3. These expressions correspond to the geometric deviations in the description leads to the same estimate the error produced by the asymmetric installation of the permanent magnet. Note that this magnetic field decays as a power of reciprocal distance from the magnet axis within the range of 2 (infinite cylindrical permanent magnet [27]) and 3 (point magnetic dipole [27]):

\[
\tilde{H}_c(r > a) \propto \frac{1}{r^2} \quad \text{and} \quad \tilde{H}_d(r > a) \propto \frac{1}{r^3},
\]

(63)

Therefore, the relative field error decreases as a power of reciprocal \( r_{err} \) between square and cube. In the considered system, conductive loop is located at \( r = a + G \), where \( a \) is the radius of permanent magnet; and \( G \) is the non-magnetic gap. In industrial electric motors, size of the gap is close to the manufacturing tolerance of the magnet radius. And in axial-flow RBPs, the gap is of the magnet radius or larger. Therefore, analytical comparison of maximum relative field error in industrial electric motors and axial-flow RBPs gives the following inequality:

\[
\delta_{RBP} = \left( 1 - \frac{r_{err}}{a + G} \right)^{g} - 1 < \left( 1 + \frac{r_{err}}{a} \right)^{g} - 1 = \delta_{EM},
\]

where \( \delta_{RBP} \) is the relative field error in axial-flow RBPs; and \( \delta_{EM} \) is the relative field error of industrial electric motors.

For dimensions and tolerances in one of typical axial-flow RBPs (\( a = 4.75 \text{ mm}; G = 3.75 \text{ mm}; \text{and } r_{err} = 0.003 \text{ mm} \)) [11], distortion of the permanent magnet field in the winding volume is up to 0.11 %. Therefore, neglect of geometric deviations in the description leads to the same value of the relative field error.

Thus, previous theoretical justification and small value of the error indicate that increased non-magnetic gap suppresses geometric deviations in the construction of axial-flow RBPs.
and leads to attenuation of high-order harmonics, and therefore, provides sinusoidal waveform of back EMF. Another consequence of increased non-magnetic gap and cylindrical shape of the permanent magnet rotor is non-salient pole nature of the electric machine having parameters of the magnetic circuit and inductance of the stator windings independent of the rotation angle.

C. OPTIMAL CONTROL CURRENT WAVEFORM

Conversion of the supplied electric energy into mechanical energy transferred to the mechanical machine is described by (47). To find an optimal waveform of control current, we define the optimality criteria as the maximum efficiency of electric machine. Equation (49) shows that linear losses are proportional to the squared current. Thus, optimal waveform of control current should correspond to the minimum RMS current value at the maximum produced torque. To solve this problem, we use the optimization methods from [28]–[30]. A continuous periodic current \( I \) in the winding and a continuous periodic change in the flux linked with winding \( d\Psi_{PM}/d\varphi \) expand in trigonometric Fourier series as follows:

\[
I = \sum_{n=1}^{N} \left( p_{q} \cos(n\varphi) + p_{d} \sin(n\varphi) \right),
\]

\[
d\Psi_{PM}/d\varphi = \psi_{a} \sum_{m=1}^{M} \left( k_{q}^{m} \cos(m\varphi) + k_{d}^{m} \sin(m\varphi) \right),
\]

where \( p_{q}, p_{d}, k_{q}, k_{d} \) are the Fourier series coefficients given as \( q- \) and \( d- \) axis components of the rotor multiple reference frame [29], [30]; \( M \) and \( N \) are finite positive integers.

Substitution of (65) and (66) into (47) gives:

\[
T = \psi_{a} \sum_{n=1}^{N} \sum_{m=1}^{M} \left[ k_{q}^{m} p_{q} \cos((m-n)\varphi) + k_{d}^{m} p_{d} \cos((m+n)\varphi) \right] + \sum_{m=1}^{M} \left[ k_{q}^{m} p_{q} \sin((m+n)\varphi) - k_{d}^{m} p_{d} \sin((m-n)\varphi) \right] + \sum_{m=1}^{M} \left[ k_{q}^{m} p_{q} \sin((m+n)\varphi) + k_{d}^{m} p_{d} \cos((m-n)\varphi) \right] + \sum_{m=1}^{M} \left[ k_{q}^{m} p_{q} \cos((m-n)\varphi) - k_{d}^{m} p_{d} \cos((m+n)\varphi) \right]
\]

(67)

Herein, we consider only the average torque \( T_{av} \) and correspondingly use only the terms of series satisfying the conditions \( m-n=0 \). Also, we should increase \( M \) or \( N \) to achieve \( H = \max \{ M, N \} \), obtaining:

\[
T_{av} = \frac{\psi_{a}}{2} \sum_{m=1}^{H} \left( k_{q}^{m} p_{q} + k_{d}^{m} p_{d} \right). \tag{68}
\]

Next, we introduce the following notations:

\[
\bar{k} = \left[ k_{q}^{1} k_{q}^{2} \ldots k_{q}^{H} k_{d}^{1} k_{d}^{2} \ldots k_{d}^{H} \right], \tag{69}
\]

\[
\bar{i} = \left[ i_{q1} i_{q2} \ldots i_{qH} i_{d1} i_{d2} \ldots i_{dH} \right]^T, \tag{70}
\]

therefore, obtaining:

\[
T_{av} = \frac{\psi_{a}}{2} \bar{k} \cdot \bar{i}. \tag{71}
\]

Rearranged (71) is:

\[
\bar{k} \cdot \bar{i} = \frac{2T^*_{av}}{\psi_{a}}, \tag{72}
\]

where \( T^*_{av} \) is the desired average torque.

Equation (72) contains \( 4H \) unknowns yielding infinitude of solutions. From this infinitude, it is necessary to find a solution with a minimum current value \( I = I_{min} \), and the maximum efficiency of electric machine. This task is performed by the method of minimum norm solution [31]:

\[
i_{min} = \left( k k^T \right)^{-1} k^T \frac{2T^*_{av}}{\psi_{a}} \tag{73}
\]

or

\[
i_{min} = k' k^T, \tag{74}
\]

where \( k' = 2T^*_{av} / (\psi_{a} k k^T) \).

In our case, the quantity \( \Psi_{PM} \) is explicitly determined by (57), and then, solution to (74) is trivial:

\[
I_{min} = i_{q1} \cos(\omega t). \tag{75}
\]
Thus, it is shown that with flux linkage varying sinusoidally in the electric machine of axial-flow RBPs, the waveform of the control current coinciding with the waveform of back EMF is optimal. This analytical conclusion is confirmed in practice by [28], [30], [32]–[34].

### D. ELECTRIC MACHINE TYPE
Eventually, considered electromagnetic system consists of the permanent magnet and the conductive loop with electric current, which, if necessary, can be replaced by a multi-turn winding. However, this configuration does not provide the ability to start the machine at an arbitrary position of the rotor at rest. At least two (preferably orthogonal) windings (Fig. 7 (a)) with the possibility of independent current regulation are necessary to start the machine. Therefore, the stator should have at least three terminals in case of combining one terminal for two windings, or four terminals in case of independent connection of windings. It is desirable to minimize the number of stator electrical leads, since larger number complicates the driveline design and increases the number of output channels, reducing the reliability of the control system. Also, it is desirable to have the complete symmetry of stator windings. This allows to unify the output channels of the control system and to optimize the load on the power source. The requirements for minimum number of leads and symmetry of windings are met by a three-phase structure, classical for alternating current circuits (Fig. 7 (b)). Based on the foregoing, it is proposed to construct the stator of axial-flow RBP of a three-phase configuration. Similar expressions for stator voltage, mechanical torque and optimal current waveform in three-phase systems are given in detail in [28], [29], [34].

Finally, we examined the key features of axial-flow RBPs, allowing to define the device as a three-phase non-salient pole synchronous machine with a permanent magnet rotor and a sinusoidal waveform of back EMF. The device is described by (58), obtained from classical electrodynamics with a number of simplifications. These simplifications lead to errors presented in Table 1, yielding overall relative error less than 0.2 %.

### IV. CONCLUSION
In summary, thorough description of electromagnetic coupling of stator and rotor in axial-flow RBPs was performed. The actual equation of stator voltage was derived from fundamental Maxwell’s equations describing this coupling through quasi-magnetostatic approximation of electromagnetic field in the simplified model consisting of permanent magnet and conductive loop rotating in free space with one spatial degree of freedom. Additionally, relative field error, created by the neglect of geometric deviations introduced through manufacturing tolerances, was estimated for a typical axial-flow RBP. The upper limit of error introduced by the simplifications was estimated less than 0.2 %, allowing to determine the finite accuracy of the proposed description and its influence on the RBP control. Sinusoidal waveform was shown to be the only waveform for back electromotive force in axial-flow RBPs, in contrast to industrial electric motors. Also, it was theoretically justified as the optimal control signal waveform. Eventually, electric machine of axial-flow RBPs was formulated as a three-phase non-salient pole synchronous machine with a permanent magnet rotor, based on the provided description and additional engineering considerations.

The presented description can be used for reasonable construction of axial-flow RBP and development of control system with increased power efficiency and optimized weight, size, and reliability, crucial in MCS.

Future research is addressed to numerical and experimental validation of the proposed theoretical description and to subsequent creation of an RBP control system with the implemented optimal waveform of control signal.

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IGOR V. NESTERENKO received the Dipl.-Ing. degree in computer science from the National Research University of Electronic Technology, Zelenograd, Moscow, Russia, in 1998. He is currently a Lead Engineer with the Institute of Biomedical Systems, National Research University of Electronic Technology. He has more than ten years industry experience in medical devices development and certification. He has solved engineering problems in the projects of the first Russian automated defibrillator and the Russian ventricular assist device. His research interests include design of electrical defibrillators, mechanical circulatory support systems, and general problems of functional safety in medical devices.

ALEXANDER A. PUGOVKIN received the Dipl.-Ing. degree in biomedical engineering from the National Research University of Electronic Technology, Zelenograd, Moscow, Russia, in 2014. He is currently an Engineer with the Institute of Biomedical Systems, National Research University of Electronic Technology, and a Research Assistant with the Institute for Bionic Technologies and Engineering, I.M. Sechenov First Moscow State Medical University, Moscow. His research interests include simulation of cardiovascular systems and design of mechanical circulatory support systems.

STEFFEN LEONHARDT (Senior Member, IEEE) received the M.S. degree from SUNY at Buffalo, NY, USA, the M.D. degree from Goethe University Frankfurt, Germany, and the Dipl.-Ing. and Ph.D. degrees in control engineering from TU Darmstadt, Germany.

After working as the Research and Development Manager at Dräger Medical AG&Co KGaA, Lübeck, Germany, for almost five years, he was appointed as a Full Professor in biomedical engineering with RWTH Aachen University, in 2003, where he has been the Philips Endowed Chair of Medical Information Technology, from 2013 to 2017. He has been a fellow of the North Rhine-Westphalian Academy of Sciences, Humanities and the Arts, Düsseldorf, Germany, since 2014. From 2015 to 2018, he served as the President for the International Society for Bioelectromagnetism, from 2015 to 2016, he was appointed as an IEEE EMBS Distinguished Lecturer. In 2018, he received the Doctor Honoris Causa from Czech Technical University in Prague, Czech Republic. Furthermore, in that year he has been appointed as a Distinguished Professor at IIT Madras, Chennai, India. His research interests include closed-loop control and automation in medicine, bioimpedance, unobtrusive sensing, and cyber medical systems. For his publication details (https://scholar.google.de/citations?hl=de&user=4JIzYgAAAAJ)

Dr. Leonhardt is an Associate Editor of the Journal of Clinical Monitoring and Computing; the IEEE TRANSACTIONS ON BIOMEDICAL CIRCUITS AND SYSTEMS, the IEEE JOURNAL OF BIOMEDICAL AND HEALTH INFORMATICS, and the journal Biomedical Engineering/Biomedizinische Technik.
ALEKSIANDR MARKOV received the B.Sc. degree in physics from St.-Petersburg State Electrotechnical University, Russia, in 2012, the M.Sc. degree from the Peter Grünberg Institute, Forschungszentrum Jülich (FZJ), Germany, in 2014, and the Ph.D. degree from the Institute of Complex Systems (ICS-8), Forschungszentrum Jülich (FZJ), in 2017. He completed his physics thesis work at Peter Grünberg Institute (PGI-8), Forschungszentrum Jülich (FZJ), working on the fabrication of multilayer thin-film structures for microwave application. From 2014 to 2017, he worked with the Laboratory of Functional Oxides and Molecules, Forschungszentrum Jülich on Tailoring and characterization of bioelectronics interfaces. He completed doctoral thesis with the Institute of Complex Systems (ICS-8), Forschungszentrum Jülich (FZJ), in 2017, on self-assembled molecules and bioelectronics interfaces. In 2018, he joined the Research Group as a Postdoctoral Researcher with the Laboratory of Organic Electronics, LOE, Linköping University’s Campus Norrköping, Sweden. He focused there on the application of organic semiconductor nanocrystals in bioelectronics and electrochemical applications. In 2019, he has moved to the Laboratory of Wearable Bio-compatible Devices and Bionic Prostheses, Institute for Bionic Technologies and Engineering, I.M. Sechenov First Moscow State Medical University, Moscow, Russia.

MARIAN WALTER (Senior Member, IEEE) was born in Saarbrücken, Germany, in 1966. He received the Dipl.-Ing. and Dr.-Ing. degrees in electrical engineering, with a specialization in control engineering from Technische Universität Darmstadt, Germany, in 1995 and 2002, respectively, and the Habilitation degree from RWTH Aachen University, Germany, in 2021. He has worked in the medical engineering industry for three years and was appointed as a Senior Scientist at Medical Information Technology, RWTH Aachen University, in 2004. His research interests include non-contact monitoring techniques, signal processing, and control of life support systems.

PATRICK BORCHERS was born in Goslar, Germany, in 1993. He received the B.Sc. degree in electrical engineering, in 2016, and the M.Sc. degree in biomedical technology from Technische Universität Berlin, Germany, in 2019. He is currently pursuing the Ph.D. degree in electrical engineering with the Chair of Medical Information Technology, RWTH Aachen University, Germany. His research interests include signal processing for and control of biomedical systems, especially for ventricular assist devices.

ALEKSIANDR MARKOV received the B.Sc. degree in physics from St.-Petersburg State Electrotechnical University, Russia, in 2012, the M.Sc. degree from the Peter Grünberg Institute, Forschungszentrum Jülich (FZJ), Germany, in 2014, and the Ph.D. degree from the Institute of Complex Systems (ICS-8), Forschungszentrum Jülich (FZJ), in 2017. He completed his physics thesis work at Peter Grünberg Institute (PGI-8), Forschungszentrum Jülich (FZJ), working on the fabrication of multilayer thin-film structures for microwave application. From 2014 to 2017, he worked with the Laboratory of Functional Oxides and Molecules, Forschungszentrum Jülich on Tailoring and characterization of bioelectronics interfaces. He completed doctoral thesis with the Institute of Complex Systems (ICS-8), Forschungszentrum Jülich (FZJ), in 2017, on self-assembled molecules and bioelectronics interfaces. In 2018, he joined the Research Group as a Postdoctoral Researcher with the Laboratory of Organic Electronics, LOE, Linköping University’s Campus Norrköping, Sweden. He focused there on the application of organic semiconductor nanocrystals in bioelectronics and electrochemical applications. In 2019, he has moved to the Laboratory of Wearable Bio-compatible Devices and Bionic Prostheses, Institute for Bionic Technologies and Engineering, I.M. Sechenov First Moscow State Medical University, Moscow, Russia.

JAMSHID H. KARIMOV received the M.D. and Ph.D. degrees. He is currently an Expert in device therapies for heart failure, mechanical circulatory support, total artificial hearts, cardiac replacement, and minimally invasive cardiac surgical techniques. He is also affiliated in molecular medicine with the Cleveland Clinic, Department of Biomedical Engineering, Cleveland Clinic Lerner College of Medicine, Lerner Research Institute, Case Western Reserve University, OH, USA. His research interests include designing and development of the advanced cardiac therapies as a potential support or substitute for cardiac failure. Dr. Karimov is a fellow of the American Heart Association, and a fellow of the American College of Cardiology.

ALEKSIANDR MARKOV received the B.Sc. degree in physics from St.-Petersburg State Electrotechnical University, Russia, in 2012, the M.Sc. degree from the Peter Grünberg Institute, Forschungszentrum Jülich (FZJ), Germany, in 2014, and the Ph.D. degree from the Institute of Complex Systems (ICS-8), Forschungszentrum Jülich (FZJ), in 2017. He completed his physics thesis work at Peter Grünberg Institute (PGI-8), Forschungszentrum Jülich (FZJ), working on the fabrication of multilayer thin-film structures for microwave application. From 2014 to 2017, he worked with the Laboratory of Functional Oxides and Molecules, Forschungszentrum Jülich on Tailoring and characterization of bioelectronics interfaces. He completed doctoral thesis with the Institute of Complex Systems (ICS-8), Forschungszentrum Jülich (FZJ), in 2017, on self-assembled molecules and bioelectronics interfaces. In 2018, he joined the Research Group as a Postdoctoral Researcher with the Laboratory of Organic Electronics, LOE, Linköping University’s Campus Norrköping, Sweden. He focused there on the application of organic semiconductor nanocrystals in bioelectronics and electrochemical applications. In 2019, he has moved to the Laboratory of Wearable Bio-compatible Devices and Bionic Prostheses, Institute for Bionic Technologies and Engineering, I.M. Sechenov First Moscow State Medical University, Moscow, Russia.

PATRICK BORCHERS was born in Goslar, Germany, in 1993. He received the B.Sc. degree in electrical engineering, in 2016, and the M.Sc. degree in biomedical technology from Technische Universität Berlin, Germany, in 2019. He is currently pursuing the Ph.D. degree in electrical engineering with the Chair of Medical Information Technology, RWTH Aachen University, Germany. His research interests include signal processing for and control of biomedical systems, especially for ventricular assist devices.

SERGEY V. SELISHCHEV (Senior Member, IEEE) received the Dipl.-Ing. degree from the National Research University of Electronic Engineering, Moscow, Russia, in 1976, the Ph.D. degree in physics and mathematics from the Institute of Metallurgy and Material Sciences of Russian Academy of Sciences, in 1983, and the Dr.-Sc. degree in physics and mathematics from the Institute of Thermophysics, Russian Academy of Sciences, in 1988. In 1993, he has initialed the teaching of biotechnical and medical devices and systems with the National Research University of Electronic Technology. In 1999, he has been founded the Department of Biomedical Systems, Institute of Biomedical Systems, since 2018, at this institution and chaired it since foundation. He was the Project Leader in the development of the first Russian automated defibrillator and the first Russian ventricular assist device, namely, Sputnik VAD. He is currently the Director of the Institute of Biomedical Systems, National Research University of Electronic Technology, Zelenograd, Moscow. He has authored and coauthored more than 200 scientific and technical papers. He is also the Editor-in-Chief of the Russian journal Meditsynskaya Tekhnika, translated and published in English as the Biomedical Engineering.

DMITRY V. TELYSHEV (Member, IEEE) received the B.S., M.S., and Ph.D. degrees in electronic engineering from the National Research University of Electronic Technology, Zelenograd, Moscow, Russia, in 2006, 2008, and 2011, respectively. He is currently the Director of the Institute for Bionic Technologies and Engineering, I.M. Sechenov First Moscow State Medical University, Moscow, and an Associate Professor with the Institute of Biomedical Systems, National Research University of Electronic Technology. He has authored and coauthored more than 70 technical papers. His research interests include the design of mechanical circulatory support systems, recording and processing of ECG, and general problems in design of implantable medical devices.

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