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Methodic of calculation of fully HTS salient-pole electrical machine

***Abstract**. In this paper, a fully HTS salient-pole electrical machine with a ferromagnetic core is considered. Analytical expressions for main parameters of the machine are obtained based on the solution of the Puasson's equation for vector magnetic potential. Obtained solution takes into account dimensions of active zone, HTS tape's properties, especially relative permeability.*

Streszczenie.** W artykule analizowano maszynę elektryczną z rdzeniem ferromagnetycznym wykonanym z nadprzewodniak HTS. Równanie analityczne bazuje na rozwiązaniu róa)nania Poissona dla wektora potencjału magnetycznego. W analizie uwzględniono wymiary części aktywnej i właściwości taśmy HTS.* ***Matematyczny model maszyny elektrycznej wykonanej z taśmy HTS

Keywords: superconducting motor, HTS, high volumetric power electrical machines, electrical machines
Słowa kluczowe: in the case of foreign Authors in this line the Editor inserts Polish translation of keywords.

Introduction

High performance and specific power of HTS electrical machines make them the most promising in the way of develop of future electrical aircraft, high speed transport systems, wind turbines and etc [8, 9]. Design of HTS machine with HTS windings is the most interesting in case of high specific output parameters. In this paper fully HTS electrical machine with silent-pole rotor is described.

Analytical expressions for the distribution of magnetic fields in the active zone of a fully HTS electrical machine.

The design scheme of a multipolar salient-pole synchronous electrical machine with HTS windings on a magnetically soft rotor core is shown in Figure 1. In general, the stator winding can have different configurations and vary in the number of phases, slots, shortening, distribution, etc. To determine the analytical solution for the problem, the armature winding can be replaced by an equivalent current layer at the bore diameter. In particular, this approach was used in the studies of electrical machines with bulk HTS elements [1,2] and HTS windings [3, 4, 7].

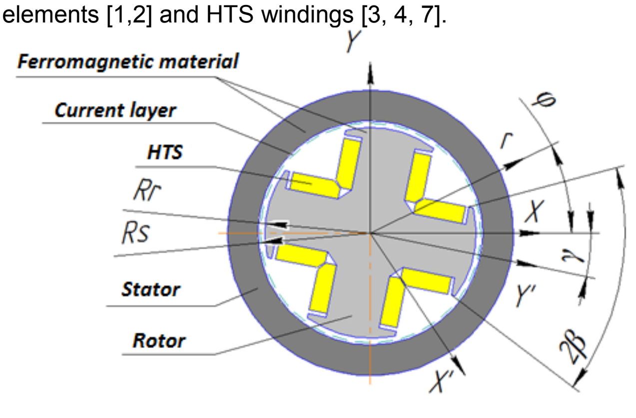


Fig. 1. Design scheme of an electrical machine with a ferromagnetic rotor and HTS windings

It is possible to use the superposition of the fields created by the currents in excitation and armature windings with the assumption of the constancy of relative magnetic permeability in ferromagnetic sections. The parameters of the machine under consideration are determined on the basis of approaches described earlier in [1, 4]. Therefore, in this paper obtained expressions take into account the relative magnetic permeability of HTS windings which can differ from [1], the magnetic properties of the salient-pole rotor can be taken into account only without saturation.

The distribution of magnetic fields is described on the basis of Maxwell's equations with the following conditions on the boundaries of areas with different magnetic permeabilities [5].

To set the problem of calculating two-dimensional magnetic fields, the following basic assumptions are subsequently adopted:

– Relative magnetic permeability of ferromagnetic segments of the rotor $\mu_{Fe}>>1$;

– Approximation of the active zone of the stator winding by an equivalent current layer on the radius R_S is used (Fig. 1);

– The machine is considered to be sufficiently long τ/L (τ - pole division, L - active length of the machine).

To date, there are a number of papers dedicated to the study of the magnetic field penetration in bulk HTS elements, and to the determination of their relative permeability. Similar studies devoted to HTS tapes are given in a limited amount and are mainly focused on the problem of AC losses in a tape. In this regard, the question of determining the magnetic permeability of coils made of HTS tape is currently open. Therefore, in this paper the relative magnetic permeability of areas with HTS windings in the rotor is assumed to be $\mu_s=1$. However, the developed technique allows us to take into account the variation of μ_s in the superconducting state.

Thus, when calculating the magnetic field in the active zone of an electrical machine, it is possible to distinguish the following areas, differing in electrical and magnetic properties (the polar coordinate system is used):

– Area of the composite rotor of the electric machine $\rho < R_r$: $\mu_{Fe}>>1$ и $\mu_s=1$;

– Air gap area ($R_r \le \rho < R_S$): $\mu_\delta=1$;

– The stator area of the electric machine ($\rho \ge R_S$): $\mu_a>>1$.

Taking into account the assumptions made, the problem of the distribution of two-dimensional stationary magnetic fields reduces to solving the Poisson equation with respect to the vector magnetic potential A (***B***=*rot**A***), which in the case under consideration will have the form:

$$\Delta A = \mu_0 J \Delta\delta(\rho - R_s)$$

with the corresponding boundary conditions on the interfaces of media with different properties.

Here $\delta(\rho - R_s)$ is the delta function, *J* is the current density, and Δ is the thickness of the current layer.

Calculation of two-dimensional fields created by armature winding, in the active zone of the synchronous machine.

Using the solutions obtained earlier in [1, 4], and taking into account the properties of the areas into which the active zone is divided, the form of the vector magnetic potential function in the air gap is defined as follows:

(1) $$A_\delta = \left(\left(\left(\frac{\rho}{R_s} \right)^p + a_p \left(\rho^p + \frac{R_s^{2p}}{\rho^p} \right) - \left(\frac{R_s}{\rho} \right)^p \right) \sin(p\varphi) + c_p \left(\rho^p + \frac{R_s^{2p}}{\rho^p} \right) \cos(p\varphi) \right) \frac{\mu_0 I k_a}{2p}$$

where μ_0 is the magnetic permeability of vacuum; *I* is the amplitude value of the phase current; *p* is the number of pole pairs; a_p and c_p are unknown constants; k_a - armature winding factor; R_s is the radius of the stator boring; ρ and φ are the coordinates in the polar system.

The solution of equation (1) in the region of the rotor is also known and has the following form [6]:

(2) $$A_R = c_p \rho_p \sin(p\varphi) + c_{p1} \rho_p \cos(p\varphi)$$

The defining of the unknown constants in expressions (1) - (2) in the case of a salient-pole rotor is a complicated problem due to the presence of an area with different magnetic permeabilities. For this, an approach is proposed that consists in modulating the penetration of the magnetic field of the armature winding into the rotor by means of a current layer on its surface. Due to the small gaps at the points of conjugation of HTS windings and soft magnetic cores, the distribution of the surface current of the rotor can be found from the solution of auxiliary problems. In particular, the field distribution in the air gap of the machine with a "sinusoidal" current on the stator and homogeneous rotors with a given magnetic permeability ($\mu_{Fe}>>1$ and $\mu_s=1$) can be used. In this case, the total magnetic field in the air gap from the stator currents and the equivalent surface current layer of the rotor will be the same as for the rotor with a given μ_r.

The current layer of the rotor I_R is a cross-continuous function and can be represented in the following form (Fig. 3):

(3) $$IR = \begin{cases} I_{RFe} \sin(p\varphi), (-\pi + \gamma) \le p\varphi \le (-\pi + \gamma + \beta) \\ I_{RS} \sin(p\varphi), (-\pi + \gamma + \beta) \le p\varphi \le (\gamma - \beta) \\ I_{RFe} \sin(p\varphi), (\gamma - \beta) \le p\varphi \le (\gamma + \beta) \\ I_{RS} \sin(p\varphi), (\gamma + \beta) \le p\varphi \le (\pi + \gamma - \beta) \\ I_{RFe} \sin(p\varphi), (\pi + \gamma - \beta) \le p\varphi \le (\pi + \gamma) \end{cases}$$

where I_{RFe} is the amplitude of the current layer in the area of the ferromagnetic parts of the rotor; I_{RS} is the amplitude of the current layer in the HTS area of the rotor winding, *p* is the number of pole pairs, γ is the angle between the stator current vector and the rotor axis *d*, and β is the angle of the pole opening of the rotor along the ferromagnetic segment (Fig. 1).

A piecewise-continuous function I_R can be represented as a Fourier series [6]:

(4) $$I_R = \sum_{k=1,3,5..}^{\infty} a_k \sin(kp\varphi) + b_k \cos(kp\varphi)$$

where *k* is the number of the harmonic.

Since it is sufficient to use only the first harmonic of the magnetic field to determine the main parameters of an electric machine [1, 4, 6], then the distribution function of the current layer (3) will have the following form:

(5) $$I_R = J\Delta = a_k \sin(p\varphi) + b_k \cos(p\varphi)$$

where $a_k = \frac{1}{\pi} \int_{-\pi}^{\pi} I_R(p\varphi) \sin(p\varphi) d\varphi$,

$b_k = \frac{1}{\pi} \int_{-\pi}^{\pi} I_R(p\varphi) \cos(p\varphi) d\varphi$ are Fourier coefficients.

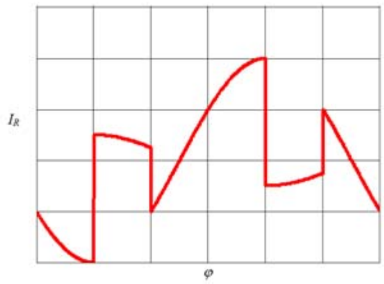


Fig. 2. Rotor current on HTS and magnetic segments

According to (3), the expressions for the Fourier coefficients a_k and b_k will have the form:

(6) $$a_k = \frac{I_{RS} - I_{RFe}}{\pi} \left(\sin(2p\beta) \cos(2p\gamma) - 2p\beta \right) + I_{RS} p$$

(7) $$b_k = \frac{I_{RS} - I_{RFe}}{\pi} \sin(2p\beta) \sin(2p\gamma)$$

The values of the amplitudes of the currents I_{RS} and I_{RFe} are found from the solution of two auxiliary problems.

The problem of the distribution of the vector magnetic potential in the air gap of the machine with a homogeneous rotor with a magnetic permeability $\mu_{Fe}>>1$.

The problem of the distribution of the vector potential in the air gap of a machine reduces to particular solutions of equations (1) and (2).

The general form of the current-layer distribution on the stator surface can be represented as:

(8) $$I_s = i_m \sin(p\varphi)$$

Whereas this distribution has only a sine component, the equation (1), which describes the magnetic field from the stator current layer without the influence of the composite rotor, will also have only a sine component and can be written as:

(9) $$A_\delta = \frac{\mu 0 Ika}{2p} \left(\left(a_p + \frac{1}{R_s^p} \right) \rho^p + \left(a_p - \frac{1}{R_s^p} \right) R_s^{2p} \rho^{-p} \right)$$

The general form of the distribution of the current layer on the surface of the rotor is:

(10) $$I_R = J\Delta \sin(p\varphi)$$

Equation (2), which describes the magnetic field from the current layer of the rotor without the influence of the stator windings, will also have only a sine component and can be written as:

(11) $$A_r = c_p \rho^p \sin(p\varphi)$$

Thus, the problem of the distribution of the vector magnetic potential in the air gap reduces to solving the following system [4, 6]:

(12) $$\begin{cases} A_\delta = \frac{\mu 0 Ika}{2p} \left(\left(a_p + \frac{1}{R_s^p} \right) \rho^p + \left(a_p - \frac{1}{R_s^p} \right) R_s^{2p} \rho^{-p} \right), R_r \le \rho < R_s \\ A_r = c_p \rho^p \sin(p\varphi), \rho < R_r \end{cases}$$

where $I = \frac{i_m m W_a}{\pi}$ is a stator current, i_m is an amplitude value of the stator current, m is the number of phases, W_a is the number of turns of the armature phase, μ_0 is the magnetic permeability of the vacuum; I is the amplitude value of the phase current; R_s is the stator radius; p is the number of pole pairs; a_p, and c_p are unknown constants.

The constants a_p and c_p are found from the following boundary conditions on the surface of the rotor:

$$(13)\quad \begin{cases} [B_n] = 0, \frac{1}{\rho}\frac{\partial A_\delta}{\partial \varphi}\Big|_{R_r} = \frac{1}{\rho}\frac{\partial A_r}{\partial \varphi}\Big|_{R_r} \\ [H_\tau] = 0, -\frac{1}{\mu_0}\frac{\partial A_\delta}{\partial \rho}\Big|_{R_r} = -\frac{1}{\mu_0}\frac{\partial A_r}{\partial \rho}\Big|_{R_r} \end{cases}.$$

Substituting (12) into (13), we obtain a system of algebraic equations for finding the unknown constants a_p and c_p. The result of the solution is:

$$(14)\quad a_p = \frac{\mu_{Fe} + l}{R_s^p(\mu_{Fe} l + 1)},\ c_p = \frac{4\mu_{Fe}\mu_0 I k_a R_s^p}{2p\left(R_s^{2p} + R_r^{2p}\right)(\mu_{Fe} l + 1)}$$

where $l = \left(R_s^{2p} - R_r^{2p}\right)/\left(R_s^{2p} + R_r^{2p}\right)$, μ_{Fe} is the relative magnetic permeability of ferromagnetic segments of the rotor (μ_{Fe}>>1).

The problem of the distribution of the vector magnetic potential in an air gap with a current layer on the surface of a rotor with I=JΔsin(pφ).

Just as in the case of the previous auxiliary problem, the general form of the distribution of the current layers on the surfaces of the stator and the rotor has only sine components.

The general form of the current-layer distribution on the stator surface can be represented as:

$$I_s = i_m \sin(p\varphi)$$

The general form of the distribution of the current layer on the surface of the rotor, as was indicated above, has the form:

$$I_R = J\Delta \sin(p\varphi).$$

Particular solutions of Eqs. (1) and (2) will also have only sine components. The problem of the distribution of the vector magnetic potential in the air gap reduces to solving the following:

$$(15)\quad \begin{cases} A_{\delta 1} = \frac{\mu_0 Ika}{2p}\left(\left(a_{p1} + \frac{1}{R_s^p}\right)\rho^p + \right. \\ \left.\left(a_{p1} - \frac{1}{R_s^p}\right)R_s^{2p}\rho^{-p}\right), R_r \le \rho < R_s \\ A_{r1} = c_{p1}\rho^p \sin(p\varphi), \rho < R_r \end{cases}$$

The unknown constants a_{p1} and c_{p1} are found from the boundary conditions on the rotor surface:

(16)

$$\begin{cases} [B_n] = 0, \frac{1}{\rho}\frac{\partial A_{\delta 1}}{\partial \varphi}\Big|_{R_r} = \frac{1}{\rho}\frac{\partial A_{r1}}{\partial \varphi}\Big|_{R_r} \\ [H_\tau] = J\Delta\sin(p\varphi), -\frac{1}{\mu_0}\frac{\partial A_{\delta 1}}{\partial \rho}\Big|_{R_r} + \frac{1}{\mu_0}\frac{\partial A_{r1}}{\partial \rho}\Big|_{R_r} = J\Delta\sin(p\varphi). \end{cases}$$

Substituting (15) into (16), we obtain a system of algebraic equations for finding the coefficients a_{p1} and c_{p1}.

Solving this system, we obtain an explicit form of the unknown constants a_{p1} and c_{p1}:

$$(17)\quad \begin{cases} a_{p1} = \frac{1}{R_s^p} - \frac{J\Delta R_r^{p+1}}{R_s^{2p}\mu_0 I k_a} \\ c_{p1} = \frac{2\mu_0 I k_a R_s^{-p} - J\Delta R_r^{p+1}\left(R_s^{-2p} + J\Delta R_r^{-2p}\right)}{2p} \end{cases}$$

Since the functions A_δ and $A_{\delta 1}$ are the same, then, equating the coefficients a_p and a_{p1}, one can obtain the dependences for the amplitude values of the currents on the surface of the ferromagnetic and HTS segments of the rotor in the following form:

$$(18)\quad J(\mu)\Delta(\mu) = 2\frac{R_s^p \mu_0 I k_a (1-\mu)}{(\bar{R}+1)R_r^{p+1}(\mu l + 1)}$$

In regard to (18), the coefficients I_{RS} and I_{RFe} will have the following form:

$$(19)\quad \begin{cases} I_{RS} = 2\frac{R_s^p \mu_0 I k_a (1-\mu_s)}{(\bar{R}+1)R_r^{p+1}(\mu_s l + 1)} \\ I_{RFe} = 2\frac{R_s^p \mu_0 I k_a (1-\mu_{Fe})}{(\bar{R}+1)R_r^{p+1}(\mu_{Fe} l + 1)} \end{cases},$$

where $\bar{R} = \frac{R_r^{2p}}{R_s^{2p}}$.

Substituting (19) into (6) and (7), we obtained expressions for the coefficients a_k and b_k:

$$(20)\quad a_k = \frac{\bar{R}\mu_0 I k_a}{R_s^p R_r^{1-p}}\left(\frac{m_1 - n_1}{\pi}\left(\sin(2p\beta)\cos(2p\gamma) - 2p\beta\right) + \right. \\ \left. + (1 - n1)p\right),$$

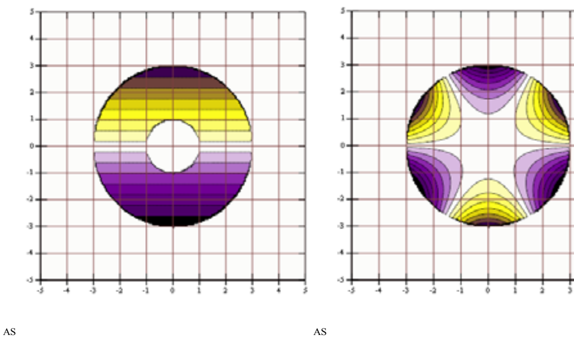
$$(21)\quad b_k = -\frac{\bar{R}\mu_0 I k_a}{R_s^p R_r^{1-p}}\frac{m_1 - n_1}{\pi}\sin(2p\beta)\sin(2p\gamma),$$

where $m_1 = (\mu_{Fe} + l)/(\mu_{Fe} l + 1)$, $n_1 = (\mu_s + l)/(\mu_s l + 1)$.

The general solution for the problem of the distribution of the vector magnetic potential in the air gap with allowance for (22) and (23) will have the form:

(22)

$$A_\delta = \left(\left(\left(\frac{\rho}{R_s}\right)^p + a_{k1}\left(\rho^p + \frac{R_s^{2p}}{\rho^p}\right) - \left(\frac{R_s}{\rho}\right)^p\right)\sin(p\varphi) + \right. \\ \left. + c_{k1}\left(\rho^p + \frac{R_s^{2p}}{\rho^p}\right)\cos(p\varphi)\right)\frac{\mu_0 I k_a}{2p}$$



a) p=1 b) p=3

Fig. 3. Distribution of magnetic fields

Using the boundary conditions (16) and expression (22), we obtain the values of the coefficients a_k and b_{k1}:

(23) $$a_{k1} = -\frac{m_1 - n_1}{\pi R_s^p}\sin(2p\beta)\cos(2p\gamma) + 2\frac{m_1 - n_1}{\pi R_s^p}p\beta + \frac{n_1}{R_s^p},$$

(24) $$b_{k1} = \frac{m_1 - n_1}{\pi R_s^p}\sin(2p\beta)\sin(2p\gamma).$$

Based on the obtained analytical expressions, the distribution of magnetic fields in the active zone of a synchronous machine with HTS windings on the rotor and stator was build. Figure 4 represents examples of such distributions with the number of pole pairs p = 1 and 3.

Inductive resistance of the armature winding.

In order to obtain the characteristics of a synchronous machine and determine its output parameters with known initial data, it is necessary to determine the main inductive resistances of the armature winding and the EMF [54].

The considered electric machine is a salient-pole since the magnetic resistances along the d and q axes do not coincide. In this regard, it is necessary to determine the inductive resistances along the d and q axes separately.

To determine the main inductive resistances x_d and x_q, we find the energy of the magnetic field in the active zone of the machine. The energy of the magnetic field in the active zone of the machine is defined as [6]:

(25) $$W = \frac{1}{2}\int JA_\delta dV = \frac{1}{2}J_0 L_s R_s \int_0^{2\pi} A_\delta \sin(p\varphi)d\varphi,$$

where $J_0 = m i_m w_a k_a / \pi R_s$, L_s is the active length of the machine, m is the number of phases, w_a is the number of turns in the armature phase, k_a is the winding coefficient, R_s is the stator radius, and A_δ is defined from (22) with allowance for (23) and (24).

After integration, we obtain the expression for the energy of the magnetic field:

(26) $$W = \frac{\mu_0 k_a^2 i_m^2 m^2 w_a^2 L_s R_s^p}{2p\pi}a_{k1},$$

Energy of one phase is $W_1 = W/m$. Using the following relations $W_1 = LI_c^2/2, x = \omega L, i_m = I_c\sqrt{2}$ we obtain expressions for the main inductive resistances x_d and x_q along the *d* and *q* axes:

(27) $$X_d(\gamma = 0) = \frac{2\mu_0 m\omega L_s w_a^2 k_a^2}{p\pi}\left(\frac{m_1 - n_1}{\pi}\left(-\sin(2p\beta) + 2p\beta\right) + n_1\right)$$

(28) $$Xq(\gamma = \pi/2) = \frac{2\mu_0 m\omega L_s w_a^2 k_a^2}{p\pi}\left(\frac{m_1 - n_1}{\pi}\left(-\sin(2p\beta)\cos(p\pi) + 2p\beta\right) + n_1\right),$$

where m is the number of phases, w_a is the number of turns of the phase of the armature winding, ω is the angular frequency, μ_0 is the magnetic permeability of the vacuum; L_s is the active length of the machine, k_a is the winding factor; p is the number of pairs of poles, $m_1 = (\mu_{Fe} + l)/(\mu_{Fe}l + 1)$, $n_1 = (\mu_s + l)/(\mu_s l + 1)$.

Determination of the excitation flux.

The effective value of the EMF of idling E_0 can be found using the main magnetic flux of field winding at zero stator currents with the help of the following relation [4, 6]:

(29) $$E_0 = \pi\sqrt{2}k_a w_a f\Phi_0,$$

where, f is the electrical frequency, and Φ_0 is the main magnetic flux.

Finding of the main magnetic flux under the above assumptions can be accomplished with the use of Ampere's law [5]:

(30) $$\oint Hdl = \sum_k I_k$$

where H is the magnetic-field intensity, dl is the element of the length of the closed contour L, $\sum I_k$ is the algebraic sum of the currents associated with the contour l.

The magnetic circuit of the machine can be divided into sections on which the intensity H is constant, on the basis of expression (32), it is possible to obtain Ohm's law for a magnetic circuit [5]:

(31) $$F = \sum \Phi R_\mu,$$

where Φ is the magnetic flux, R_μ is the magnetic resistance, $F=IW$ is the magnetomotive force (I is the current in the winding, and W is the number of turns).

The design scheme of the machine shown in Figure 1, can be used to develop constructing scheme of an active zone. Taking into account the mechanical properties of modern HTS tapes, the design of the machine with the number of slots per pole and phase q <1 can be chosen. This allows the use of racetrack coils in the stator and rotor of the machine. A construction scheme of such a machine is shown in Fig. 4.

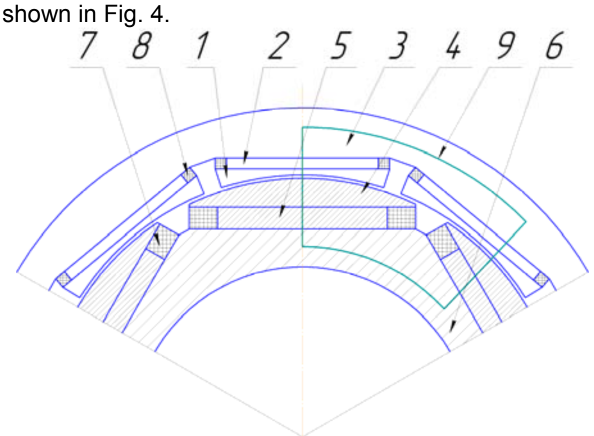


Fig. 4. Constructive diagram of a fully HTS electrical machine
1 - the clamp of the stator tooth; 2 - stator tooth; 3 - stator yoke; 4 - rotor pole clamp; 5 – pole core; 6 - rotor yoke; 7 - HTS coil; 8 - HTS coil; 9 - main magnetic flux line

It can be seen from Fig. 4 that the main magnetic flux produced by the field winding is closed along the following sections of the magnetic circuit: the core of the rotor pole, the pole clamp, the air gap, the clamp of the stator tooth, the tooth, the stator yoke, the tooth, the clamp of the stator tooth, the pole clamp, the rotor pole, the rotor yoke. In this case, all the structural elements are made of ferromagnetic material. Thus, the relative magnetic permeability μ of all parts of the magnetic circuit, other than the air gap, is nonzero and nonlinear.

Determining the parameters of nonlinear systems is a complex task, the solution of which requires an iterative process. In addition, the presence of nonlinearity makes it impossible to apply the principle of superposition of magnetic fields. However, most electrical machines operate near the bend of magnetization curve [5]. This makes it possible to take a certain value of μ for the ferromagnetic sections, relying on the experience of designing traditional and HTS electrical machines.

Considering the magnetic flux loop, given in Figure 4, and the dimensions of the machine, on the basis of (31), one can obtain an equation for determining the magnetomotive force of the machine:

$$F = \Phi_\delta \cdot \left(R_{\mu i}/2 + 2\cdot R_{\mu s} + 2\cdot R_{\mu p} + 2\cdot R_{\mu\delta} + \right. \\ \left. + +3\cdot R_{\mu pz}/2 + 3\cdot R_{\mu z}/2 + R_{\mu a}/2\right) \quad (32)$$

where $R_{\mu i}$ is the magnetic resistance of the rotor yoke; $R_{\mu s}$ is the magnetic resistance of the rotor pole core; $R_{\mu p}$ is the magnetic resistance of the pole clamp; $R_{\mu\delta}$ is the magnetic resistance of the air gap; $R_{\mu pz}$ - magnetic resistance of the clamp of the stator tooth; $R_{\mu z}$ is the magnetic resistance of the stator; $R_{\mu a}$ is the magnetic resistance of the stator yoke; Φ_δ is the main magnetic flux of the machine.

The magnetic flux in the tooth and its clamp in one case is equal to the magnetic flux in the air gap, and in the other is equal to its half. This is due to the fact that the armature winding is made with the number of slots per pole and phase q = 0.5.

In general, the magnetic resistance is written as [5]:

$$R_\mu = l/(\mu \cdot \mu_0 \cdot S), \quad (33)$$

where l is the length of the magnetic field line; S is the area through which the magnetic flux is looped, μ_0 is the magnetic constant; μ is the relative magnetic permeability.

Considering the geometrical dimensions of the sections of the machine's magnetic circuit, the expressions for the magnetic resistances are:

- rotor pole core

$$R_{\mu s} = H_s/(\mu_s \mu_0 L_s B_s), \quad (34)$$

- rotor pole clamp

$$R_{\mu p} = H_p/(\mu_p \mu_0 L_s \tau_p), \quad (35)$$

- air gap:

$$R_{\mu\delta} = \delta/(\mu_0 L_s \tau_\delta), \quad (36)$$

- stator tooth:

$$R_{\mu z} = H_z/(\mu_z \mu_0 L_s B_z), \quad (37)$$

- the clamp of the stator tooth:

$$R_{\mu p\_st} = H_{p\_st}/(\mu_{p\_st} \mu_0 L_s \tau_{p\_st}), \quad (38)$$

- stator yoke:

$$R_{\mu a\_st} = \tau_a/(\mu_{a\_st} \mu_0 L_s H_{a\_st}), \quad (39)$$

- rotor yoke:

$$R_{\mu a\_r} = \tau_{a\_r}/(\mu_{a\_r} \mu_0 L_s H_{a\_r}), \quad (40)$$

where H_s, H_p, H_z, H_{p_st}, H_{a_st}, H_{a_r} are the heights of the core of the rotor pole, the pole clamp, the stator tooth, the clamp of the stator pole, the stator and rotor yokes respectively; δ is the air gap size; μ_s, μ_p, μ_z, μ_{p_st}, μ_{a_st}, μ_{a_r} is the relative magnetic permeabilities of the core of the rotor pole, rotor pole clamp, stator tooth, stator pole clamp, stator yoke and rotor yoke respectively; L_s is the active length; μ_0 is the magnetic permeability of vacuum; τ_p, τ_{p_st}, τ_a, τ_{a_r} are pole division on the diameter of the rotor pole clamp, stator pole clamp, the average diameter of the stator yoke and rotor yoke, respectively.

Substituting (34) - (40) into (32), we obtained the expression for the main magnetic flux:

$$\Phi_0 = (2I_f B_k \mu_0 L_s)\cdot\left(4h_l b_l (H_1/\mu_s + H_2/\mu_p + H_\delta) + \right. \\ \left. + 3(H_3/\mu_z + H_4/\mu_{p\_st}) + H_5/\mu_{a\_st} + H_6/\mu_{a\_r}\right)^{-1} \quad (41)$$

where $H_1=H_s/B_s$, $H_2=H_p/\tau_p$, $H_3=H_z/B_z$, $H_4=H_{p\_st}/\tau_{p\_st}$, $H_5=\tau_a/H_{a\_st}$, $H_6=\tau_{a\_r}/H_{a\_r}$, $H_\delta=\delta/\tau_\delta$ are constructive coefficients, I_f is the excitation current of HTS winding, B_k is the width of the field winding coil, L_s is the active length of the machine.

Expression (41) allows to determine the main magnetic flux taking into account dimensions of the active zone of the machine, critical current of the HTS tape and the relative magnetic permeability of the magnetic circuit of the machine parts.

Rated calculation of the HTS synchronous machine.

Due to the steel magnetic core, the machine cannot provide a high specific power (kW/kg), but it allows to increase the volumetric power (MW/liter). In order to determine the specific parameters, using the obtained expressions for the inductive resistances of the armature winding and the magnetic flux of field winding, an estimated calculation of a 1 MW synchronous HTS motor has been performed. The initial data is given in the Table 1.

Table 1. Initial parameters of the machine

Parameter	Symbol	Value
Power, kW	P_2	1000
Phase voltage, V	U	1000
Number of phases	m	3
Rotation speed, min^{-1}	n	2500
Critical current of the HTS tape, A	I_s	60

At the first stage, an analytical calculation is performed, a sketch of the active zone of the machine is obtained, which is used to construct a three-dimensional model and further three-dimensional finite element simulation in Ansys Maxwell. The results of calculating of the EMF and inductive resistances according to analytical expressions and modeling are given in the Table 2.

Table 2. Comparision of results of analytical calculation and numerical modeling

Parameter	Analytical calculation	Modeling
EMF of idling, V	870	885
Inductive resistance along the d axis, Ohm	1.18	1.29
Inductive resistance along the q axis, Ohm	1.02	1.1

It can be seen from Table 2 that difference between numerical modeling and analytical calculation is less than 10%. It is connected with steel saturation (in 3D model real steel was used) and flux linkages. In general, it means that developed methodic could be used for determining the basic parameters of the machine in the first stage. The results of calculating the machine with 1 MW output power are given in the Table 3.

Table 3.REsults of calculating of 1 MW fully HTS machine

Parameter	Symbol	Value
Radius of stator boring, m	R_a	293
Active length, m	L_s	330
Number of pole pairs	p	9
Height of field winding, mm	H_f	16
Height of armature winding, mm	H_a	16
EMF of idling, V	E_0	938
Inductive resistance of field winding along the d axis, Ohm	X_d	1.29
Inductive resistance of armature winding along the d axis, Ohm	X_q	1.1
Specific power, kW/kg	P_2`	2.8
Volumetric power, kW/l	P_2``	22

Conclusions.

The paper considers a new technique for calculating the basic parameters of a fully HTS electrical machine with electromagnetic excitation and a salient-pole rotor. The feature of the obtained expressions is the possibility to consider the transport current of HTS coils, as well as their magnetic permeability, which was not accomplished in previous works. In addition, the magnetic permeability of sections of the magnetic circuit can also be taken into account. The developed technique allows to estimate the main dimensions of the active zone of the machine at the first stage. Analytical expressions for the inductive resistances and the main magnetic flux of a fully HTS machine have a simple analytical form, which allows for optimizing calculations. Finite element modeling showed high accuracy of the developed technique.

Thus, the proposed approaches and the results obtained can serve for estimating and optimizing the calculations of a fully HTS electric machine with increased volumetric power.

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