Iron Loss Analysis and Efficiency Calculation of Double Stator HTS Machine With Stationary Seal

WENZHONG MA1, YING HUANG1, DALONG GONG2, WEIGUO LI2, AND YUBIN WANG1

1College of New Energy, China University of Petroleum (East China), Qingdao 266580, China
2Shandong Energy Group Company Ltd., Jinan, Shandong 250014, China

Corresponding author: Wenzhong Ma (mawenzhong@126.com)

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ABSTRACT This paper mainly analyzes the iron loss of a double stator high-temperature superconducting (HTS) machine with stationary seal and calculates its efficiency. Firstly, according to the field-modulation theory, the air-gap flux density harmonic components of the excitation magnetic field and armature-reaction magnetic field are derived. Secondly, the air-gap flux density harmonic components corresponding to the flux density harmonic components of iron core are derived, and the main iron core flux density harmonic components are confirmed and verified by finite element analysis (FEA). When using the semi-FEA method to calculate the iron loss, according to the analysis results of main core flux density harmonic components, the appropriate step-interval of the FEA is selected. In addition, according to the characteristics of the iron core flux density, this paper proposes an alternative method of iron core flux density, which greatly shortens the calculation time of semi-FEA method of iron loss calculation. Then, iron loss model with variable coefficients is used to calculate the iron loss of the machine, and the calculation results are compared with those of the FEA iron loss calculation method. Finally, other losses and efficiency of the machine are calculated.

INDEX TERMS Field modulation, high-temperature superconducting machine, finite element analysis, core flux density, iron loss, machine efficiency.

I. INTRODUCTION

In recent years, high-temperature superconducting (HTS) brushless machines with stationary seal, whose HTS excitation windings have been located in the stator side, have attracted increasing attention in high-power direct-drive applications [1], [2], [3], [4], [5]. Due to the coupling device for cryogenic transfer and the torque tube omitted, the HTS coils of such machines are relatively static with respect to their refrigeration systems. In addition, such machines also do not need slip rings and carbon brushes, which greatly reduces the maintenance cost and improves the system reliability. Meanwhile such machines feature rich flux density harmonic components due to the modulation effect during the energy conversion, which makes iron loss analysis and efficiency calculation become complicated. But from the viewpoint of field-modulation theory, loss analysis and efficiency calculation are decoupled into contributions by each field harmonic component, which makes the analysis more organized. In this paper, a double stator high-temperature superconducting machine (DS-HTSM) with stationary seal [6] is taken as an example, and its iron loss and efficiency are calculated with the help of field-modulation theory.

Iron loss is one of the main parts of machine losses, which limits the improvement of machine efficiency. Therefore, it is required to analyze the source of iron loss and accurately calculate iron loss during machine design. The calculation methods of machine iron loss can be divided into analytical method, finite element analysis (FEA) method and semi-FEA method.

The analytical method of iron loss calculation is based on the analytical calculation of iron core flux density, and then the appropriate iron loss model is used to calculate the iron loss. In [7], the flux waveform of each part of the machine was predicted by measuring the flux waveform of the stator yoke, and it is assumed that the flux density at each point

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is equal to the average flux density of that part. Obviously, this experiment-based calculation of the core flux density is not applicable to the original design of machines. In [8], the equivalent magnetic circuit method was used to solve the iron core flux density of the machine. However, when the machine structure is relatively complicated, the calculation amount of this method will be greatly increased, and it is also difficult to establish an accurate equivalent magnetic circuit model. In addition to the above methods, there is also an analytical method based on the analytical calculation of the air-gap flux density. For example, [9] calculated the stator/rotor tooth flux densities by calculating the average value of the integral of the air-gap flux densities. Since the iron core flux densities obtained by the above methods are all average values, the iron loss calculation based on these methods will not be accurate enough.

The FEA method of iron loss calculation is based on the finite-element magnetic field analysis, and uses the iron loss analysis program to calculate the average iron loss or instantaneous iron loss [10], [11], [12], [13]. The semi-FEA method of iron loss calculation also requires finite-element magnetic field analysis to calculate the core flux density [14]. However, different from the FEA method, it can flexibly combine the loss experimental data of core material and select an appropriate iron loss model for iron loss calculation. In fact, whether it is the FEA method or semi-FEA method, it is necessary to set the appropriate step-interval for the finite-element magnetic field analysis. If the step-interval is too short, the computing time may be wasted, while if the step-interval is too large, some high-frequency harmonic components of core flux density may not be accurately calculated, causing errors in the iron loss calculation. In addition, for field-modulation machines, the core flux density periods of the stator and rotor are generally different [15], [16], [17], [18], which makes it difficult to set the duration of the finite-element magnetic field analysis. Through the above analysis, the key of the finite element magnetic field analysis is to select the appropriate duration and step-interval, which is based on the period of the core flux density and the frequency of the main core flux density harmonic components, respectively.

The duration of the FEA should be consistent with the flux density period. However, [19] proposed a computationally efficient FEA method when analyzing the permanent magnet rotor machine, which greatly shortened the FEA time of the permanent magnet flux density. To select the step-interval, it is only to determine the main components. Therefore, it is not necessary to accurately calculate the amplitudes of core flux density harmonic components. Literature [20] deduced the expression of the air-gap modulation field, and then obtained the frequencies of the core flux density harmonic components according to the rotating speeds and orders of the air-gap flux field harmonic components. By referring to the method in literature [20], the period (reciprocal of fundamental frequency) and main harmonic frequencies of core flux density of DS-HTSM can be calculated. However, unlike the single air-gap machine in the above literature, the double air-gap of DS-HTSM naturally divides itself into three parts. The modulation of each part core influences the flux density harmonic components of other part cores. Therefore, the flux density harmonic components of the core are more abundant.

The above iron loss calculation methods are compared, as shown in Table 1. In this paper, based on the analytical calculation of the main harmonic components of the core flux density of the machine, a method for obtaining flux density data of iron core is proposed, which makes the semi-finite iron loss method both accurate and fast.

In addition to iron loss, DS-HTSM loss also includes copper loss, mechanical loss, and HTS loss. In a low-speed direct-drive machine, the skin effect of the armature winding is generally not considered, so Joule’s law can be used directly to calculate the copper loss of DS-HTSM. The HTS loss of HTS excitation machines is generally negligible. Therefore, the key of efficiency calculation of DS-HTSM is still iron loss calculation.

This paper is organized as follows. Section II introduces the machine topology and derives the expression for its air-gap flux density from the perspective of field modulation. Section III analyzes the frequencies of the core flux density harmonic components, selects the main core flux density harmonic and uses the FEA method to verify. Section IV proposes a method to shorten the time of FEA. On this basis, the semi-FEA method based on the iron loss model with variable coefficient is used to calculate the iron loss of the machine, and the obtained results are compared with the results of the FEA method. Then, Section V calculates other losses and overall efficiency of the machine. Finally, Section VI concludes this paper.

### Table 1. Calculation methods of iron loss.

| Calculation method                  | Initial design | Accuracy | Computation speed |
|------------------------------------|----------------|----------|------------------|
| Analytical method [8];[9]          | √              | Bad      | fast             |
| Analytical method (Based on the experiments) [7] | ×              | Bad      | fast             |
| FEA method [10];[13]               | √              | good     | Slow             |
| Semi-FEA method*                   | √              | good     | Relatively fast  |

* The semi-finite element method in this paper.

### II. CONFIGURATION MODEL OF AIR-GAP FLUX DENSITIES

#### A. CONFIGURATION

The configuration of DS-HTSM is shown in Fig. 1. It can be seen that it mainly includes an outer stator with the armature windings, an outer air-gap, a sandwiched rotor consisting of the alternating magnetic blocks and nonmagnetic blocks, an inner air-gap, and an inner stator with HTS excitation coils and cryogenic Dewars. The inner stator and the outer stator of the machine are both stationary, and they are collectively referred to as stator.
As mentioned above, DS-HTSM works based on field-modulation theory. In order to maximize torque capability by taking advantage of the magnetic gearing effect, the relationship between the rotor magnetic block number $P_r$, the excitation coils pole-pairs number $P_f$ and armature winding pole-pairs number $P_a$ should satisfy

$$P_r = P_f + P_a \quad (1)$$

**B. ANALYTICAL MODEL OF NO-LOAD AIR-GAP FLUX DENSITIES**

In this paper, the analytical calculation of air-gap flux densities is mainly to obtain the main core flux density harmonic components. So it is not necessary to calculate the amplitude of the air-gap flux density harmonic components accurately, and the magnetomotive force (MMF) waveforms and air-gap permeance waveforms are simplified to rectangular waves.

The polarity of two adjacent excitation coils in DS-HTSM is opposite. Fig. 2 shows the excitation MMF waveform, where $\theta_1$ is half of the radian of the inner stator tooth, $\theta$ is the mechanical angle along the machine’s circumference, and $F_m$ is the maximum value of the excitation MMF. By using Fourier decomposition, the excitation MMF $F_f(\theta)$ can be expressed as

$$F_f(\theta) = \sum_{j=1,3,5,...}^{+\infty} [F_j \cos(jP_f\theta)] \quad (2)$$

$$F_j = \frac{4F_m}{j\pi} \sin^2 \left(\frac{j\pi}{2}\right) \sin \left(j\pi\alpha_1\right) \quad (3)$$

where $F_j$ is the amplitude of the $j$-th harmonic component of the excitation MMF, and $\alpha_1$ is the pole arc coefficient of the inner stator. Since DS-HTSM is a double-polarity machine, the excitation MMF is expressed as an odd harmonic function. According to the properties of Fourier series, $j$ is a positive odd number.

The rotor is composed of alternating magnetic blocks and nonmagnetic blocks, which makes its air-gap permeance function alternate periodically along the circumference. Fig. 3 shows a cycle of the rotor permeance waveform, where $\theta_2$ is half of the radian of the rotor magnetic block, $\lambda_{rmax}$ and $\lambda_{rmin}$ are the maximum and minimum values of the rotor permeance function, respectively, and $\omega_r$ is the mechanical angular velocity of the rotor. By using Fourier decomposition, the air-gap permeance $\lambda(\theta, t)$ can be expressed as

$$\lambda_r(\theta, t) = \lambda_{r0} + \sum_{i=1}^{+\infty} \left[ \lambda_{ri} \cos \left( iP_r(\theta - \omega_r t + \theta_0) \right) \right] \quad (4)$$

$$\lambda_{ri} = \frac{2(\lambda_{rmax} - \lambda_{rmin}) \sin \left(i\pi\alpha_2\right)}{i\pi} \quad (5)$$

$$\lambda_{r0} = (\lambda_{rmax} + \lambda_{rmin}) \alpha_2 \quad (6)$$

where $\lambda_{r0}$ is the DC component of the rotor permeance function, $\lambda_{ri}$ is the amplitude of the $i$-th harmonic of the rotor permeance function, $\theta_0$ is the initial position angle of the rotor, and $\alpha_2$ is the rotor pole arc coefficient. Due to the influence of the saturation of the rotor magnetic blocks and the structure of the dovetail slots, $\alpha_1$ is not equal to 0.5. Therefore, based on (4)-(6), the rotor permeance function has even-order harmonic components.

In addition to the rotor, the outer stator teeth also modulate the no-load air-gap flux densities. Fig. 4 shows a cycle of the outer stator permeance waveform, where $\theta_3$ is half of the outer stator tooth angle, $\lambda_{omax}$ and $\lambda_{omin}$ are the maximum and minimum values of the permeance function of the outer
As the rotor, the outer stator is stationary relative to the excitation MMF source. By using Fourier decomposition, its air-gap permeance function $\lambda_{os}(\theta)$ can be expressed as:

$$\lambda_{os}(\theta) = \lambda_{os0} + \sum_{k=1}^{+\infty} [\lambda_{osk} \cos(k \alpha_3 \theta)]$$  (7)

$$\lambda_{osk} = \frac{2}{\pi} \sum_{k=1}^{+\infty} \frac{\lambda_{omax} - \lambda_{omin}}{k} \sin(k \pi \alpha_3)$$  (8)

$$\lambda_{os0} = 2 \left[ \lambda_{omax} (1 - \alpha_3) + \lambda_{omin} \alpha_3 \right]$$  (9)

where $\lambda_{os0}$ is the DC component of the outer stator permeance function, $\lambda_{osk}$ is the amplitude of the $k$-th harmonic of the outer stator permeance function, and $\alpha_3$ is the pole arc coefficient of the outer stator.

Fig. 5 shows the flux path of DS-HTSM excitation field. It can be seen that the excitation flux can be divided into main flux and leakage flux according to whether the excitation flux intersects the armature winding or not. If flux leakage is not considered, the flux density expressions of the inner air-gap and outer air-gap are exactly same, which can be expressed as

$$B_f(\theta, t) = F_f(\theta) \cdot \lambda_r(\theta, t) \cdot \lambda_{os}(\theta)$$

$$= \lambda_{r0} \lambda_{os0} \sum_{j=1,3,5,...}^{+\infty} \left[ F_j \cos(j \Phi_f) \right] + \frac{1}{2} \lambda_{r0} \sum_{j=1,3,5,...}^{+\infty} \sum_{i=1,k=1}^{+\infty} \left[ F_{j,i} \lambda_{osk} \cos\left( \left( j \Phi_f \pm k \alpha_3 \right) \theta \right) \right] + \frac{1}{2} \lambda_{os0} \sum_{j=1,3,5,...}^{+\infty} \sum_{i=1,k=1}^{+\infty} \left[ F_{j,i} \lambda_{osk} \cos\left( \left( j \Phi_f \pm i \alpha_3 \right) \theta \right) \right]$$

$$\times \left( \theta \mp \frac{ip_r \omega_0 t - ip_r \theta_0}{j \Phi_f \mp iN_r} \right)$$

In fact, the leakage flux of DS-HTSM cannot be ignored. Fig. 5 shows that the flux leakage can be divided into two types. The permeance of the first type is constant. The permeance of the second type is also constant in the circumferential direction and contains part of the rotor AC permeance in the diametrical direction. Therefore, these two types of leakage fluxes do not bring additional harmonic components of the air gap flux density, but cause the amplitudes of the inner air-gap flux density harmonic components to be different from those of the outer air-gap.

C. ANALYTICAL MODEL OF ARMATURE-REACTION AIR-GAP FLUX DENSITIES

The three-phase armature current with neglected harmonic components can be expressed as

$$i_A = \sqrt{2} I \sin(\omega_0 t + \varphi_0)$$

$$i_B = \sqrt{2} I \sin(\omega_0 t - 2\pi/3 + \varphi_0)$$

$$i_C = \sqrt{2} I \sin(\omega_0 t + 2\pi/3 + \varphi_0)$$

where $I$ is the effective value of the current, $\omega_0$ is the electrical angular velocity of the current, and $\varphi_0$ is the initial angle of the current.

Armature windings of DS-HTSM are arranged as double-layer concentrated windings. Fig. 6 shows MMF of three-phase armature windings, where $N$ is the number of turns of the armature windings. The resultant MMF of the three-phase armature windings $F_a(\theta, t)$ can be expressed as

$$F_a(\theta, t) = \sum_{v=1}^{+\infty} \left[ F_v \sin(\omega_0 t - v\Delta \Phi_0 + \varphi_0) \right]$$

$$F_v = \frac{2\sqrt{2} NI \sin(2\pi v \alpha_3/3)}{\pi \nu}$$

$$\delta = \begin{cases} 1, & v = 3n - 2 \\ -1, & v = 3n - 1 \end{cases}$$

where $F_v$ is the amplitude of the $v$-th harmonic of the armature-reaction MMF $\delta$ is 1 or -1, and $n$ is a positive integer.
Since the number of pole-pairs of the armature-reaction magnetic field is similar to the number of rotor magnetic blocks, most of the fluxes of the armature-reaction magnetic field start from the outer stator and return to the outer stator through the rotor magnetic blocks. In addition, the wide inner stator pole tips provide paths for the armature-reaction magnetic field. Therefore, the modulation effect of the inner stator pole tips on the armature-reaction magnetic field is ignored. The air-gap flux densities of armature-reaction field can be expressed as

\[ B_a(\theta, t) = F_a(\theta, t) \cdot \lambda_{tr}(\theta, t) \]

\[ = \lambda_{r0} \sum_{v=1}^{+\infty} \left[ F_v \cos \left( \frac{v \delta P_a}{2} \left( \theta - \frac{\omega_a t + \phi_0 - \pi/2}{v \delta P_a} \right) \right) \right] \]

\[ + \frac{1}{2} \sum_{v=1}^{+\infty} \sum_{i=1}^{+\infty} \left[ F_r \lambda_{rc} \cos \left( \frac{(v \delta P_a \pm i P_r) + (v \delta P_a \pm i P_r)}{v \delta P_a + i P_r} \right) \right] \]

\[ \times \left( \theta - \frac{(\omega_a \pm i P_r \omega_r) t \mp i P_r \theta_0 + \phi_0 - \pi/2}{v \delta P_a + i P_r} \right) \]

(15)

### III. ANALYSIS OF CORE FLUX DENSITY HARMONICS

#### A. CORE FLUX DENSITY HARMONICS PRODUCTION PRINCIPLE

The space harmonic components of air-gap flux densities with different rotational speeds enter the core, resulting in a periodic and non-sinusoidal core flux density with respect to time. Each time the space harmonic of air-gap flux densities passes through a pole pitch, the corresponding temporal harmonics of stator core flux densities alternates once, and the alternating frequency \( f_{sc} \) can be expressed as

\[ f_{sc} = \frac{P_{air} |\omega_{air}|}{2\pi} \]  

(16)

where \( P_{air} \) and \( \omega_{air} \) are the order (represented by the number of pole-pairs) and the mechanical angular velocity of a certain air-gap flux density space harmonic, respectively.

Since the rotor is rotating at the angular speed \( \omega_r \), the alternating frequency \( f_{rc} \) of the rotor core flux density harmonic can be expressed as

\[ f_{rc} = \frac{P_{air} |\omega_{air} - \omega_r|}{2\pi} \]  

(17)

In the following analysis, the air-gap flux density harmonic components refer to the air-gap flux density space harmonic components, and the core flux density harmonic components refer to the core flux density temporal harmonic components.

#### B. ANALYSIS OF NO-LOAD CORE FLUX DENSITY HARMONICS

The no-load air-gap flux densities contain rich harmonic components. The orders and rotating speeds of harmonic components are calculated using Eq. (10), and then the frequencies of stator and rotor core flux density harmonic components are calculated according to Eq. (16) and Eq. (17), respectively.

#### III. ANALYSIS OF CORE FLUX DENSITY HARMONICS

**Table 2. No-load core flux density harmonic frequencies.**

| Order   | Rotation speed | Frequency (stator core) | Frequency (rotor core) |
|---------|----------------|-------------------------|------------------------|
| \( jP_f \)| 0              | 0                       | \( jP_f \)             |
| \( iP_{r+1} \times kn_o \) | 0               | \( iP_{r+1} \times kn_o \) | \( jP_{r+1} \times kn_o \) |
| \( iP_{r+1} \times kn_o \) | \( iP_{r+1} \times kn_o \) | \( iP_{r+1} \times kn_o \) | \( iP_{r+1} \times kn_o \) |

Table 2 shows the results of the above derivation, where frequency (stator core) and frequency (rotor core) are the frequencies of the stator and rotor core flux density harmonic components, respectively, and \( f_r \) is the mechanical frequency and is equal to \( \omega_r / 2\pi \).

Based on Table 2, the no-load air-gap flux density harmonic orders \( P_{no} \) can be expressed as

\[ P_{no} = |jP_f \pm aiP_r \pm \beta kn_o| \]  

(18)

where \( \alpha \) can take 0 or 1, which means that the harmonic components are modulated by the DC or AC permeance components of rotor, and \( \beta \) can take 0 or 1, which means that the harmonic components are modulated by the DC or AC permeance components of outer stator. But it should be noted that the two ‘±’ in Eq. (18) are independent, that is to say, the equation has four forms.

The corresponding harmonic frequencies of stator and rotor core flux densities are expressed as Eq. (19) and Eq. (20), respectively.

\[ f_{sc} = |\alpha iP_r| f_r \]  

(19)

\[ f_{rc} = |jP_f \pm \beta kn_o| f_r \]  

(20)

It can be considered that the excitation MMF is a uniform MMF modulated by the inner stator teeth. According to Eq. (19) and Eq. (20), it can be found that the no-load core flux density harmonic frequencies depend only on the AC components of permeance functions of the relatively moving modulator.

According to the law of electromagnetic induction, the armature current frequency should be equal to the fundamental frequency of outer core flux density, which can be expressed as

\[ f = P_{fr}f_r = P_r n_r / 60 \]  

(21)

where \( n_r \) is the rotor speed. The rated speed is 300 t/min, so the rated armature current frequency of DS-HTSM is 90 Hz.

#### C. ANALYSIS OF ARMATURE-REACTION CORE FLUX DENSITY HARMONICS

Similar to the previous part, according to Eq. (15), Eq. (16) and Eq. (17), the harmonic frequencies of armature-reaction core flux densities are calculated and shown in Table 3.

From the stator core flux density frequencies in Table 3, it can be seen that the outer stator core flux density harmonic
TABLE 3. Armature-reaction core flux density harmonic frequencies.

| Order | Rotation speed | Frequency (stator core) | Frequency (rotor core) |
|-------|----------------|-------------------------|------------------------|
| ωP/ν | ω/ωP | f | f−νωP/ν |
| νωP/ν±iνP | (ωP±iω)/(νωP±iνP) | iνP/ν | iνωP/ν |

FIGURE 7. The finite element model of DS-HTSM.

TABLE 4. Key parameters of DS-HTSM.

| Parameter | Value |
|-----------|-------|
| Rated power (kW) | 10 |
| Rated speed (r/min) | 300 |
| Number of outer stator teeth | 42 |
| Number of inner stator teeth | 8 |
| Number of pole pairs of armature windings | 14 |
| Number of pole pairs of exciting windings | 4 |
| Number of rotor magnetic blocks | 18 |
| Outer stator outside diameter (mm) | 720 |
| Outer stator inside diameter (mm) | 612 |
| Inner stator outside diameter (mm) | 552 |
| Inner stator inside diameter (mm) | 326 |
| Outer air-gap length (mm) | 1 |
| Inner air-gap length (mm) | 1 |
| Stack length (mm) | 100 |

components are generated by the modulation of the AC components of the rotor permeance function, and the rotor core flux density harmonic components are only related to the space harmonic of the armature-reaction MMF. Therefore, same as no-load, the armature-reaction core flux density frequencies depend only on the AC components of permeance function of the relatively moving modulator.

D. ANALYSIS AND FEA VERIFICATION OF MAIN CORE FLUX DENSITY HARMONICS

The finite-element model of DS-HTSM is shown in Fig. 7, and the key parameters are shown in Table 4. In order to facilitate the analysis of the characteristics of core flux densities, four points are marked in Fig. 7. In the figure, point A is located at the inner stator pole tip, point B is located at the outer stator teeth, point C is located at the rotor near the inner air-gap, point D is located at the rotor near the outer air-gap.

When j, i and k are taken as 1, the corresponding harmonic orders of the no-load air-gap flux densities are 4$^{th}$, 14$^{th}$, 20$^{th}$, 22$^{nd}$, 28$^{th}$, 38$^{th}$, 46$^{th}$, 56$^{th}$ and 64$^{th}$. Fig. 8 shows the waveforms and harmonic components of the air-gap flux densities when DS-HTSM excitation current is 40A. It can be found from Fig. 8(b) that the amplitudes of the above-mentioned harmonic components (j = i = k = 1) are relatively high, so they are confirmed as the main air-gap flux density harmonic components.

According to Table 2, the fundamental frequencies of the core flux densities in the stator and rotor are determined as 90 Hz and 10 Hz, respectively. Therefore, the core flux density periods of the stator and the rotor are 1/90s and 0.1s, respectively. That is to say, the stator core flux density period is equal to the armature current period, while the rotor core flux density period is 9 times of the armature current period.

According to Table 2, the frequencies of the core flux density harmonic components corresponding to the main air-gap flux density harmonic components are solved, and the obtained results are shown in Table 5. It is proved by the finite element magnetic field analysis below that these core flux density harmonic components in Table 5 are the main components.

Fig. 9 shows the no-load flux density waveforms and harmonic components of the point A and point B, where T is the armature current period (1/90s) of DS-HTSM. It can be seen from Fig. 9(a) that the periods of the flux densities in
the inner and outer stator core are 1/90s, which is consistent with the above analysis. It can be seen from Fig. 9(b) that among the harmonic components, the DC component and the 90 Hz fundamental component are relatively high, which is consistent with Table 5. This shows that the amplitudes of these stator core flux density harmonic components are also relatively high, so they are confirmed to be the main harmonic components.

Fig. 10 shows the no-load flux density waveforms and harmonic components of the point C and point D. It can be seen from Fig. 10(a) that the periods of the flux densities in the rotor core are 0.1s, which is consistent with the above analysis. It can be seen from Fig. 9(b) that the 20 Hz harmonic component is relatively high. In addition, the amplitudes of the 190 Hz and 230 Hz harmonic components at point D are also relatively high. The 190 Hz and 230 Hz harmonic components are generated by the modulation of the outer stator teeth. Therefore, their amplitudes at the point D near the outer stator teeth are higher than those at the point C. Overall, the harmonic components of 20 Hz, 190 and 230 Hz are also the main core flux density harmonic components in the rotor, which is consistent with Table 5.

In order to study the separate influence of DS-HTSM armature-reaction magnetic field, the excitation current is set to 0 A, and the armature coils are connected to the three-phase sinusoidal current source. The initial phase angle of the three-phase current source is the same as that of the no-load back EMF. Fig. 11 shows the armature-reaction air-gap flux densities. With the harmonic orders ($n = 1$) marked in Fig. 11(b), it can be seen that these harmonic components account for a large proportion of the total harmonic components, and these harmonic components are called main armature-reaction air-gap flux density harmonic components. It should be noted that the 32nd order harmonic is derived from the reverse-rotating armature-reaction MMF harmonic, but the 32nd order harmonic component does not exist in the no-load main air-gap flux density harmonic components.

Table 6 shows the core flux density harmonic components caused by the main armature-reaction air-gap flux density components.
TABLE 6. Armature-reaction air-gap flux density main harmonic orders and their corresponding core flux density harmonic frequencies.

| Order | Frequency (stator core) | Frequency (rotor core) |
|-------|-------------------------|------------------------|
| 14th  | 90 Hz                   | 20 Hz                  |
| 32nd  | 180 Hz                  | 20 Hz                  |
| 4th   | 0 Hz                    | 20 Hz                  |
| 28th  | 90 Hz                   | 230 Hz                 |
| 46th  | 0 Hz                    | 230 Hz                 |

It can be seen that the above-mentioned special 32nd order harmonic component generate an additional stator core flux density harmonic component with the frequency of 180 Hz.

Fig. 12 shows the armature-reaction flux density waveforms and harmonic components of the point B of the outer stator. Under the armature-reaction magnetic field, the flux density of point A at the pole tip is very small and its harmonic components are similar to those of point B, so the flux density waveform of point B is not shown here. It can be seen from Fig. 12(a) that the core flux density period of the outer stator under the armature-reaction magnetic field is also 1/90s. Fig. 12(b) shows that the frequencies of main core flux density harmonic components of outer stator are 90 Hz and 180 Hz and that a DC component exists in the flux density, which is consistent with the analysis in Table 6.

Fig. 13 shows the comparison between the core flux density harmonic components under no-load and on-load conditions. It can be seen from Fig. 13(a) that the flux density period in rotor core under the armature-reaction magnetic field are 20 Hz and 230 Hz, which are consistent with the results obtained in Table 6.

Fig. 14 shows the comparison between the core flux density harmonic components under no-load and on-load conditions. It can be seen from Fig. 14(a) that the harmonic component of radial flux density with a frequency of 180 Hz harmonic components ($n = 1$). It can be seen that the above-mentioned special 32nd order harmonic component generate an additional stator core flux density harmonic component with the frequency of 180 Hz.

Fig. 12 shows the armature-reaction flux density waveforms and harmonic components of the point B of the outer stator. Under the armature-reaction magnetic field, the flux density of point A at the pole tip is very small and its harmonic components are similar to those of point B, so the flux density waveform of point B is not shown here. It can be seen from Fig. 12(a) that the core flux density period of the outer stator under the armature-reaction magnetic field is also 1/90s. Fig. 12(b) shows that the frequencies of main core flux density harmonic components of outer stator are 90 Hz and 180 Hz and that a DC component exists in the flux density, which is consistent with the analysis in Table 6.

Fig. 13 shows the armature-reaction flux density waveforms and harmonic components of the point D of the rotor. It can be seen from Fig. 13(a) that the flux density period in rotor core under the armature-reaction magnetic field is also 0.1s. It is inferred from Fig. 13(b) that the frequencies of the main flux density harmonic components in the rotor core under the armature-reaction magnetic field are 20 Hz and 230 Hz, which are consistent with the results obtained in Table 6.
is almost zero under no-load condition and has a certain amplitude at on-load condition, so this harmonic component is unique to the armature-reaction magnetic field. In addition, the harmonic amplitudes of other harmonic components under no-load and on-load conditions are not much different. Through the above analysis, it can be seen that the amplitudes of the harmonic components of the flux densities the core is different when the DS-HTSM is loaded and no-load, but the periods of the core magnetic densities are same, and the highest values of the harmonic frequency are both 230 Hz. Therefore, the same finite element model can be used under no-load and on-load conditions to obtain the core flux density data required for iron loss calculation.

IV. IRON LOSS CALCULATION
A. FLUX DENSITY DATA OBTAINED BY FEA METHOD
In order to calculate the iron loss, it is necessary to obtain the flux density data of the machine core. When using FEA method to obtain flux density data, the number of FEA steps per second $N_{\text{step}}$ (reciprocal of step-interval) can be calculated using the following equation:

$$ N_{\text{step}} = k_c f_{\text{max}} $$  \hspace{1cm} (22)

where $f_{\text{max}}$ is the highest frequency of the flux density harmonic components that need to be accurately accounted for in the iron loss calculation and $k_c$ is a coefficient. In general, the higher the value of $k_c$, the more accurate the flux density data will be, but the computation time will be correspondingly longer. The maximum frequency of the main core flux density harmonic component is 230 Hz, so $f_{\text{max}}$ is taken as 230 Hz. In order to take into account the influence of the 230 Hz harmonic component on the iron loss, the coefficient $k_c$ should be at least two, which can be explained by Shannon’s sampling theorem. In this paper, $k_c$ is taken as ten to obtain more accurate iron loss values. According to Eq. (22), the number of FEA steps per second is 2,300, and the FEA step-interval is 0.43 ms. By the iron loss harmonic analysis method, it is found that the iron loss generated by the harmonic components below 230 Hz accounts for 83.5% and 89.6% of the total iron loss under no-load and on-load conditions, respectively. It can be seen that the iron loss in the main harmonic frequency range (0-230 Hz) accounts for the main part, which shows that the step-interval of the finite element analysis is reasonable.

In the previous section, it is concluded that the no-load or on-load flux density period in DS-HTSM stator core is 1/90s ($T$), while the period of the rotor core is 0.1s ($9T$). Therefore, the conventional FEA time is set to 0.1s to obtain complete flux density data. On top of that, this paper proposes a method to reduce the FEA time to one-ninth of the conventional calculation as follows.

The time taken for a rotor magnetic block to rotate from the current position to the position of the next one is $T$, and the period of the armature current is $T$. Therefore, the flux density at a point inside the machine at time $t_0$ should be the same as at time $t_0 + T$. Fig. 15 shows the rotor core flux density contour at different times when the machine is loaded. The flux density contour of 1# magnetic block in the Fig. 15 (e)-(h) is exactly same as that of the 2# magnetic block in the Fig. 15 (a)-(d). It is not difficult to infer that the flux density distribution of the 1# magnetic block in the time interval ($2T - 9T$) can be replaced by that of the 2#-9# magnetic blocks in the time interval (0-$T$).

Fig. 16 shows flux density waveforms at the center of the 1# magnetic block using the method I and the method II, where method I is the conventional method and method II is the proposed alternative method. It can be seen from Fig. 16 that the flux density waveforms obtained by the two methods are basically the same. Therefore, only the time $T$ is needed to obtain the flux density data of the 1# magnetic block within $9T$. In addition, the flux densities of different magnetic blocks are only different in the initial phase angles,

![FIGURE 15. Flux density contour of rotor core at different times. (a) $t = 0$, (b) $t = T/4$, (c) $t = 2T/4$, (d) $t = 3T/4$, (e) $t = T$, (f) $t = 5T/4$, (g) $t = 6T/4$, (h) $t = 7T/4$.](image-url)
which is expressed as

\[ P_{Fe} = k_1 \cdot B_m^2 \cdot f_c + k_2 \cdot B_m^2 \cdot f_c^2 \]  (23)

where \( B_m \) and \( f_c \) are the amplitude and frequency of core flux density, respectively, and \( k_1 \) and \( k_2 \) are the hysteresis loss coefficient and the eddy current loss coefficient, respectively.

B. COEFFICIENT FITTING OF IRON LOSS MODEL

In general engineering calculations, a simplified constant-coefficient iron loss model is used to calculate iron loss, which is expressed as

\[ P_{Fe} = k_1 \cdot B_m^2 \cdot f_c + k_2 \cdot B_m^2 \cdot f_c^2 \]  (23)

where \( B_m \) and \( f_c \) are the amplitude and frequency of core flux density, respectively, and \( k_1 \) and \( k_2 \) are the hysteresis loss coefficient and the eddy current loss coefficient, respectively.

Due to the rich harmonic components of the core flux densities in DS-HTSM, the constant-coefficients iron loss model will cause a large error. In order to fit the coefficients, the independent hysteresis loss coefficients and eddy current loss coefficients should be separated from the measured loss data at different flux density amplitudes and frequencies. Literature [21] divided both sides of Eq. (23) by \( B_m^2 f_c \) at the same time, and regarded the obtained \( P_{Fe}/B_m^2 f_c \) as a linear function of frequency \( f_c \). The intercept of the linear function is the hysteresis loss coefficient, and the slope of the linear function is the eddy current loss coefficient. In fact, when the flux density amplitude or frequency is slightly higher, the frequency and the influence on the loss at high frequency cannot be ignored. According to the variation rules of the frequency and the influence on the loss at high frequency cannot be ignored. Therefore, this paper attempts to fit the iron loss data using a cubic polynomial function. When \( f_c \) is zero, the iron loss \( P_{Fe} \) is also zero, so the constant term of the cubic polynomial function should be zero. This fitting function can be expressed as

\[ P_{Fe} = a(B_m)^3 + b(B_m)^2 + c(B_m)f_c \]  (24)

where \( a(B_m) \), \( b(B_m) \) and \( c(B_m) \) are related functions of the flux density amplitude \( B_m \) only.

Fig. 17 shows the measured values and fitted curves of iron loss of non-oriented silicon steel 50WW470. In Fig. 17, the fit goodness of each curve is greater than 0.9999, so the relationship between iron loss and frequency can be well expressed by using the Eq. (24). The parameters obtained by fitting the iron loss curve are shown in Table 7. Five sets of experimental data are used when fitting each curve, but these fitting curves must pass through the origin, so it is equivalent to using six sets of experimental data for cubic fitting, and the obtained results are obviously credible.

It can be seen from Table 7 that the absolute value of \( a \) is very small. However, \( a \) is the coefficient of the third power of the frequency and the influence on the loss at high frequency cannot be ignored. According to the variation rules of \( a, b \) and \( c \) in Table 7, polynomials of different degrees are used for fitting. The fitting polynomials are shown in Eq. (25), where \( a_1 - a_4, b_1 - b_2 \) and \( c_1 - c_4 \) are constants, and the results after fitting are recorded in Table 8.

\[
\begin{align*}
  a(B_m) &= (a_1 B_m^3 + a_2 B_m^2 + a_3 B_m + a_4) B_m^2 \\
  b(B_m) &= (b_1 B_m + b_2) B_m^2 \\
  c(B_m) &= (c_1 B_m^3 + c_2 B_m^2 + c_3 B_m + c_4) B_m^2
\end{align*}
\]  (25)
When the core flux density is alternating, according to the law of electromagnetic induction, an eddy-current-like elec-

tromotive force will be induced in the iron core, so a current will be generated, which is called eddy current. Since the essence of eddy current loss is Joule heating, the harmonic analysis method in circuit analysis can be used. However, there is also a rotating magnetization in DS-HTSM core, which will generate additional eddy current losses. Aiming at the characteristic of the flux density of DS-HTSM core, an orthogonal decomposition model is used in this paper to approximate the loss caused by the elliptical rotational magnetization.

The flux density of each point in the machine core can be decomposed into two components, the tangential component and the radial component. Using the principle of harmonic analysis, the radial and tangential components can be decomposed into a series of flux density harmonic components, respectively. The radial and tangential flux density harmonic components of the same order at one point can form a flux density vector, and the trajectory of the flux density vector in the Cartesian coordinate system is an ellipse whose center point is at the origin.

The schematic diagrams of the two magnetization models are shown in Fig. 20. In the general harmonic analysis model, the elliptical rotational magnetization is treated as radial and tangential alternating magnetization, and $B_{rm}$ and $B_{tm}$ are the radial and tangential flux density amplitudes, respectively. In the orthogonal decomposition model, the elliptical rotational magnetization is treated as two alternating magnetizations with flux density amplitudes $B_{am}$ and $B_{bm}$, where $B_{am}$ and $B_{bm}$ are the semi-major axis length and semi-short axis length of the elliptical trajectory, respectively. The transformation from the general harmonic analysis model to the orthogonal decomposition model is equivalent to a rotation change of the rectangular coordinate system. In Fig. 20, the $x$-axis and $y$-axis represent the radial and tangential directions of the machine, respectively, and the $x'$-axis and the $y'$-axis represent the long-axis direction and the short-axis direction of the ellipse, respectively. The $xy$ coordinate system rotates $\theta_k$ counterclockwise to get $x'y'$ coordinate system. It can be seen from Fig. 20 that when the flux density amplitude in the long axis direction of the ellipse is the largest, the corresponding amplitude in the short axis direction will be zero, and vice versa. Therefore, the alternating magnetization

C. SEMI-FEA METHOD AND FEA METHOD OF IRON LOSS CALCULATION

When the core flux density is alternating, according to the law of electromagnetic induction, an eddy-current-like elec-
in these two directions has little influence on each other, and the alternating magnetization in these two directions can be used to replace the elliptical rotation magnetization. The orthogonal decomposition model of eddy current loss is expressed as

\[
P_e = \int_1^\infty \sum_{k=1}^\infty k^2 \left\{ k_e(B_{amk}, k f_1) B_{amk}^2 + k_e(B_{bmk}, k f_1) B_{bmk}^2 \right\}
\]

(29)

where \( k \) is the harmonic order of the core flux density, \( f_1 \) is the fundamental frequency of the flux densities, and \( B_{amk} \) and \( B_{bmk} \) are the semi-major axis length and semi-minor axis length of the elliptical trajectory of \( k \)-order flux density harmonic, respectively.

Due to the rich core flux density harmonic components of DS-HTSM, a flux density cycle may contain multiple hysteresis loops. Fig. 21 shows a schematic diagram of hysteresis loops, which includes a major hysteresis loop and two minor hysteresis loops. The flux density amplitude of the large hysteresis loop is \( B_1 \), and the flux density amplitude of the upper minor hysteresis loop is \( B_2 \). The hysteresis loss is mainly determined by the area of the hysteresis loop. Yamazaki’s model was proposed as \[14\], \[22\]

\[
P_h = f_c \left( \sum_{j=1}^{N_r} k_h(B_{rj}) B_{rj}^2 + \sum_{i=1}^{N_t} k_h(B_{tj}) B_{tj}^2 \right)
\]

(30)

where \( N_r \) and \( N_t \) are the numbers of radial and tangential hysteresis loops, respectively, and \( B_{rj} \) and \( B_{tj} \) are the amplitudes of radial and tangential hysteresis loops, respectively.

Based on the core flux density data obtained by FEA, the hysteresis loss is calculated using the MATLAB program. The key of this program is to obtain the flux density amplitude of each hysteresis loop. The program flow is shown in Fig. 22.

The semi-FEA method based on Eq. (29) and Eq. (30) are used to calculate the no-load and on-load core losses of DS-HTSM. Table 9 shows a comparison between the calculated iron loss of the semi-FEA method and the FEA method. It should be noted that FFT and hysteresis loops models are respectively used to calculate eddy current losses and hysteresis losses when using FEA method. It can be seen from Table 9 that the results obtained by the two methods are almost the same. However, when using semi-FEA method, an alternative method is proposed to shorten the calculation time of iron core flux density by 1/9, thus greatly saving the calculation time of iron loss.
So the DS-HTSM efficiency can be expressed as

$$\eta(\%) = \frac{\omega_r T_e - P_{Fe} - P_{cu}}{\omega_r T_e + P_b} \times 100 \quad (33)$$

where $T_e$ is the electromagnetic torque calculated by FEA.

When DS-HTSM operates with the rated armature current, the output power and efficiency corresponding to different excitation currents are shown in Fig. 24. When the excitation current of the machine exceeds 40A, the output power rises slowly. This is caused by the saturation of the iron core, so it is appropriate to choose the rated excitation current of 40A. It can be seen from Fig. 24 that when the excitation current is the rated excitation current, the efficiency of the machine can reach 94%.

VI. CONCLUSION

This paper mainly analyzes the iron loss and efficiency of DS-HTSM. In this paper, semi-FEA method is used to calculate the iron loss of the machine, and a method is proposed to shorten the calculation time of iron loss to about 1/9 of that of the traditional semi-FEA method. In addition, the main harmonic frequencies of the core flux densities are deduced, and the FEA step-interval is set based on the highest frequency (230 Hz) in the main core flux density harmonic. Then, the semi-FEA method is used to calculate the iron loss of the machine, and the obtained results are consistent with the calculation results of the FEA method. Finally, through the efficiency calculation, it is found that the output efficiency of the machine can reach 94% with the rated load.

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**Ying Huang** was born in Shandong, China, in 1997. He received the B.S. degree in electrical engineering and automation from the College of Information and Electrical Engineering, Shandong Jianzhu University, Jinan, China, in 2020. He is currently pursuing the M.Sc. degree in electrical engineering with the China University of Petroleum (East China).

His research interest includes loss and efficiency analysis of high temperature superconducting generators.

**Dalong Gong** was born in China, in 1997. He received the B.S. degree in electrical engineering from the China University of Petroleum, Shandong Jianzhu University, Jinan, China, in 2020. He is currently pursuing the M.Sc. degree in electrical engineering with the China University of Petroleum (East China).

His research interest includes flexible multistate switch.

**Weiguo Li** was born in Shandong, China, in 1984. He received the M.S. degree from Zhejiang Sci-Tech University, Zhejiang, China, in 2007. Since 2007, he has been with Shandong Energy Group Company Ltd., where he is currently an Intermediate Engineer with the New Energy Department. His current research interests include fuel cell distributed energy management and development of fuel cell control systems.

**Yubin Wang** was born in Shandong, China, in 1984. He received the M.Sc. degree in control theory and control engineering from the College of Electrical Engineering and Automation, Shandong University of Science and Technology, Jinan, China, in 2003, and the Ph.D. degree in electrical engineering from the Department of Electrical Engineering, Southeast University, Nanjing, China, in 2011.

Since 2003, he has been with the China University of Petroleum (East China), where he is currently a Professor with the College of New Energy. From September 2013 to December 2013, he was a Research Assistant with the Department of Electrical and Electronic Engineering, The University of Hong Kong. He is the author or coauthor of over 30 technical papers. His teaching and research interests include electromagnetic design and analysis of high temperature superconducting generators with stationary seal and permanent magnet machines.