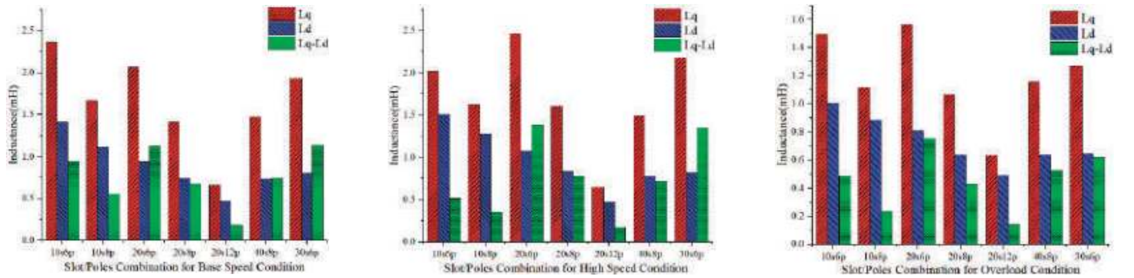
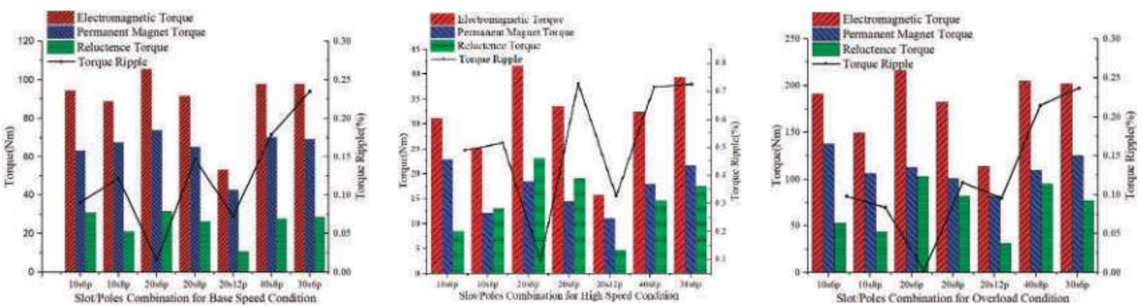
To ensure the stator outer diameter, each phase conductor number and PM mass is the same, and the torque and inductance performance of selected seven slot/pole combinations under three operating conditions are solved; the results are summarized in Figures 3 and 4.



(**a**) (**b**) (**c**)

Figure 3. Inductance characteristics in different slot/pole combinations: (**a**) base speed condition, (**b**) high-speed condition, and (**c**) overload speed condition.



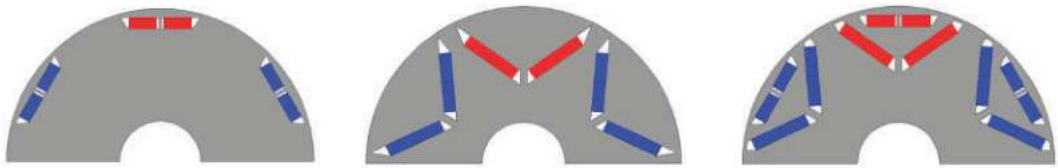
(**a**) (**b**) (**c**)

Figure 4. Torque characteristics in different slot/pole combinations: (**a**) base speed condition, (**b**) high-speed condition, and (**c**) overload speed condition.

Through the comparative analysis, it is found that when the pole pair number is the same, the difference in value between the L_d and L_q becomes larger as the slot number increases, and the torque performance under high-speed conditions is better at the same time; when the slot number is the same, the torque performance under high-speed condition becomes worse as the pole pair number increases. Compared with the integer slot winding, the fractional slot has a smaller torque ripple. It can be concluded that a motor that adopts a fractional slot can obtain better torque performance with a low torque ripple. Compared with the concentrated winding, the distributed winding has a larger difference in inductance, causing a larger reluctance torque. Although most motors choose FSCW or ISDW in basic design, they often benefit from the traditional three-phase motor but not the five-phase motor design. In order to make the five-phase motor with more choices of high winding coefficient slot/pole combination, FSDW can be seen as one suitable choice. Five-phase motor with an FSDW owns less end-winding length and torque ripple than the ISDW and owns a higher winding coefficient compared with a five-phase motor with FSCW. Considering both power density improvement and torque performance, 20s6p slot/pole combination FSDW is selected.

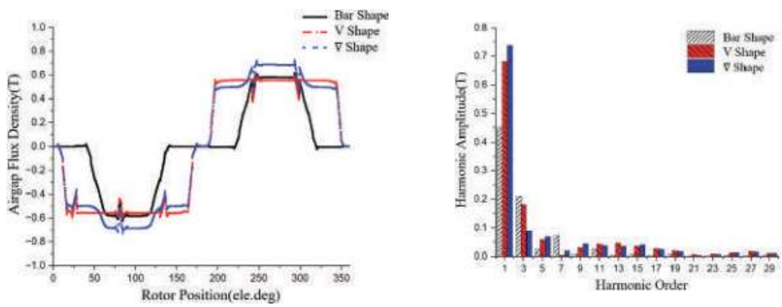
2.4. IPMSM Permanent Magnet Layer Number

The permanent magnet layer of the rotor affects the air gap-flux density and electromagnetic characteristics. It is significant to find how the PM layer number influences the motor electromagnetic characteristics and how to choose the PM layer number to obtain high power density. Figure 5 shows three rotor structures with bar shape, V shape, and ∇ shape, respectively, wherein the bar shape and V shape are single-layer rotor structures, and the ∇ shape is a double-layer rotor structure. Figure 6 shows the comparison of the no-load air gap flux density and its harmonic content under three different rotor structures.



(**a**) (**b**) (**c**)

Figure 5. IPMSM Rotor Structures: (**a**) Bar Shape. (**b**) V Shape. (**c**) ∇ Shape.



(**a**) (**b**)

Figure 6. IPMSM rotor structures air gap flux density performance: (**a**) air gap flux density waveform, (**b**) harmonic content.

Compared with the single-layer rotor structure, the double-layer rotor structure's air gap flux density peak value is larger, which, according to Equation (4), can improve motor power density. From the view of air gap flux density harmonic content, the fundamental content is higher than single-layer, while the other harmonics content is less than single-layer. If we define harmonic distortion rate as the ratio of all harmonics RMS, which excludes fundamental order, to the fundamental RMS, then the harmonic distortion rate of zero corresponds to the sinusoidal degree equal to 1. It can be clearly seen that the double-layer rotor structure has a better sinusoidal degree than the single-layer rotor structure.

By observing the double-layer ∇ shape rotor structure cloud map of the no-load magnetic field distribution in Figure 7, it is found that the nether layer PM pushes the upper layer magnetic field lines upward, resulting in lower magnetic flux leakage at the end of the upper layer PM. Increasing the air gap flux density peak value in the middle part of the double-layer rotor structure makes the no-load air gap flux density approach the sine waves, resulting in an improved sinusoidal degree of the air gap flux density. According to Equation (4), selecting the double-layer rotor structure contributes to motor high power density design.

Figure 7. No-load magnetic field distribution cloud map in ∇ shape rotor structure.

3. EM Analysis of Five-Phase CSDW High Power Density Motor

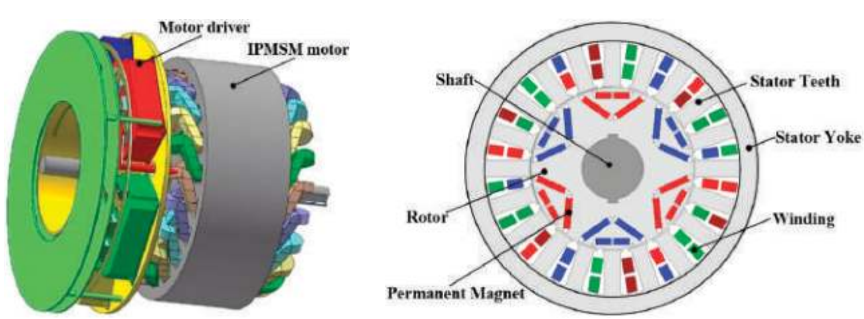
In the integrated motor drive system designed previously, called Ver1.0 type, it was found that an IMDS has a large volume and the motor inside has a low motor power density, which is still a distance from the DOE roadmap. In order to reflect high power density design method rationality and correctly, all indicators remain the same as in Ver1.0 type [18]. According to these indicators, the electromagnetic analysis and optimization of the Ver2.0 motor were carried out and compared with the Ver1.0 motor. The indicators for both Ver1.0 and Ver 2.0 types are shown in Table 3, which include motor indicators and part of motor driver indicators.

Table 3. Main indicators of HPDM.

Parameter	Value
Rate power	60 kW
Peak power	110 kW
Base speed	6000 rpm
Maximum speed	15,000 rpm
Peak Torque	>200 Nm
Torque Ripple	<5%
Current density	<12 A/mm^2
DC bus voltage	650 V
Peak current	<200 A
Stator outer diameter	$\leq$240 mm
Volume power density	>10 kW/L
Mass power density	>3 kW/kg

What needs illustration is that the peak current is affected by the selected switching device. Conventional IGBT devices cannot stand the extremely adverse working environment, but with the development of a wide bandgap semiconductor technology, SiC devices can adapt to high-power, high-temperature, and high-frequency applications. Considering the high power of IMD, the SiC switching device was selected, and the 1200 V and 300 A SiC half-bridge module of CREE series CAS300M12BM2 was selected as the switching device in the driver. The voltage and current indicators in Table 3 have taken the margins into account according to the chosen switching device.

The IMDS topology is shown in Figure 8a, which is the same structure as the Ver1.0 type. Through the analysis in Section 2, it is determined that the motor structure is a five-phase 20s6p double-layer rotor structure IPMSM, which can improve motor power density and ensure the torque and flux-weakening performance for an electric vehicle. The motor topology is shown in Figure 8b.



(**a**) (**b**)

Figure 8. Ver2.0 type topology: (**a**) IMDS overall structure, (**b**) Ver2.0 IPMSM motor structure.

3.1. Magnetic Bridge

The ∇-type rotor inner structure's connection to the outer relies only on two tangential magnetic bridges. The rotor stress is concentrated under high-speed conditions to cause rotor damage. In order to reduce the rotor stress at high speeds, radial magnetic bridges should be added between the permanent magnets in the same layer. There are three magnetic bridges that the motor can freely set, which are respectively recorded as magnetic bridges 1, 2, and 3, as shown in Figure 9. The effect of different magnetic bridge combinations on motor stress and electromagnetic performance will be discussed below.

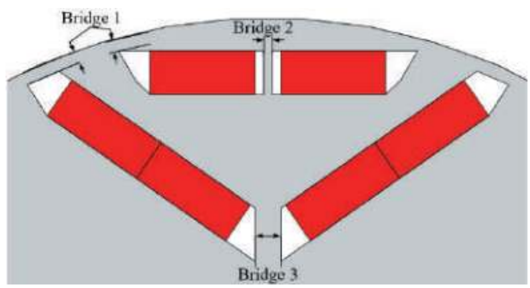


Figure 9. Rotor magnetic bridge schematic diagram.

Figure 10 shows the rotor stress results for different magnetic bridge combinations under Maximum speed conditions. It is found that when there is only bridge 1, or when adopting bridge 1 and bridge 2 combination, the rotor stress exceeds the yield strength of the 10JNEX900 silicon steel sheet (570 MPa) [22]. When adopting the bridge 1 and bridge 3 combination, the rotor's maximum stress is significantly reduced compared with the above two combinations. It can be concluded that the radial magnetic bridge 3 has an obvious improvement effect on the rotor intensity under maximum speed conditions.

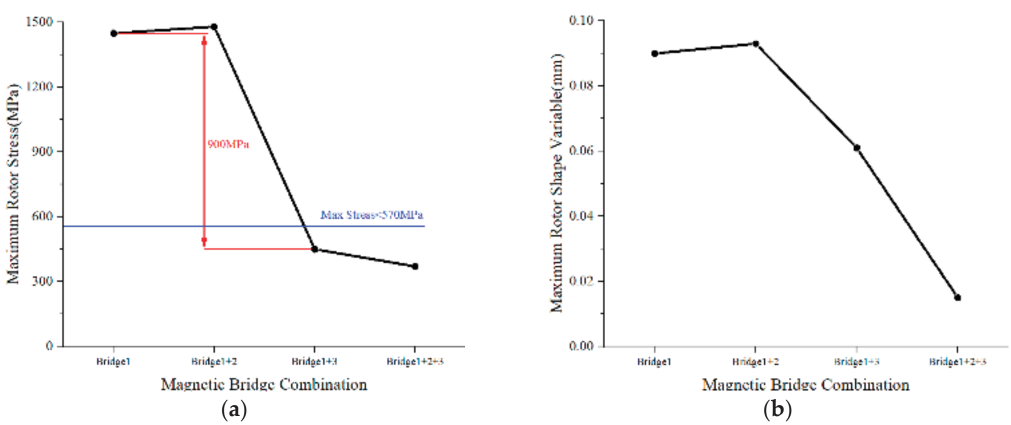
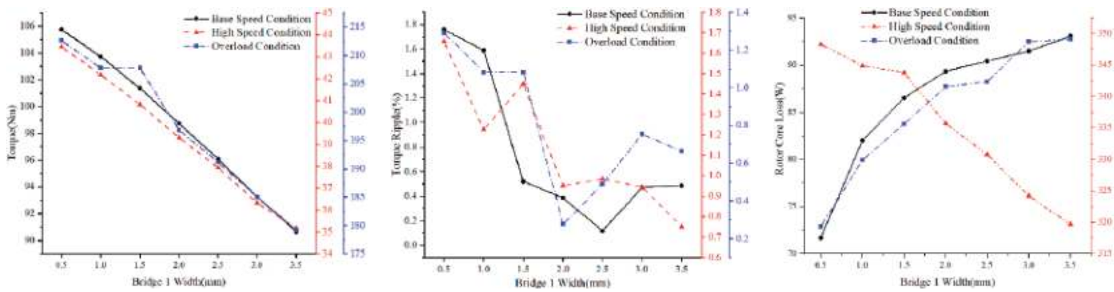


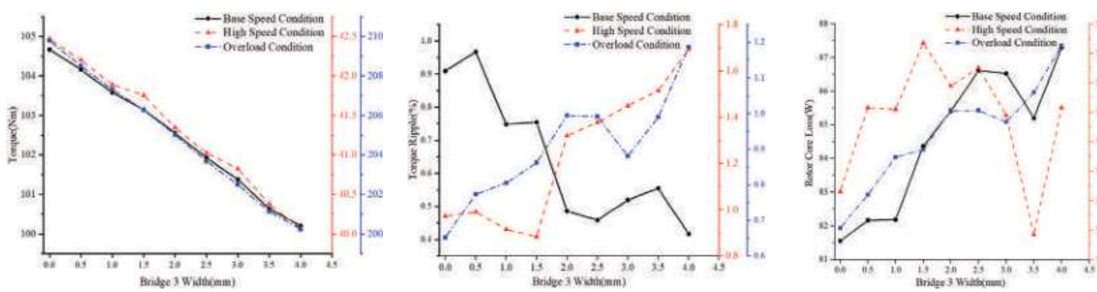
Figure 10. Rotor stress with different magnetic bridge combinations at maximum speed conditions: (**a**) rotor maximum von Mises stress, (**b**) rotor maximum deformation quantity.

Although widening a magnetic bridge helps to reduce rotor stress, an excessively wide magnetic bridge will lead to serious leakage flux of the permanent magnet ending, which will influence torque performance and motor power density. Figures 11 and 12 show the motor torque performance with different magnetic bridge 1 width and different magnetic bridge 3 width, respectively.



(**a**) (**b**) (**c**)

Figure 11. Motor electromagnetic characteristics with different bridge 1 width: (**a**) output torque, (**b**) torque ripple; (**c**) rotor core loss.



(**a**) (**b**) (**c**)

Figure 12. Motor electromagnetic characteristics with different bridge 3 width: (**a**) output torque, (**b**) torque ripple, (**c**) rotor core loss.

Magnetic bridge 1 is a tangential magnetic bridge, which mainly affects motor electromagnetic characteristics. A narrow magnetic bridge 1 will cause serious saturation at the magnetic bridge, leading to a rotor flux leakage increase. More permanent magnet flux leakage leads to permanent magnet utilization reduction, affecting motor torque performance and a larger rotor core loss under saturated conditions. Compared with the magnetic bridge 1, magnetic bridge 3 is a radial magnetic bridge. Although an excessively wide magnetic bridge can effectively relieve rotor stress, it also causes serious flux leakage onto the lower layer permanent magnet's terminal, leading to a motor output torque decrease. Torque ripple and rotor core loss under high-speed conditions and overload conditions will increase when the magnetic bridge is too wide. However, its relative, the tangential magnetic bridge has less influence on output torque.

In general, the radial magnetic bridge mainly affects the mechanical stress of the rotor, and the tangential magnetic bridge mainly affects the electromagnetic characteristics.

3.2. Stator Structure Optimization

The stator structure parameters are shown in Figure 13, which are mainly divided into the slot size and the stator yoke size. In terms of improving the power density, according to Equation (4), increasing slot pure copper area is beneficial to motor high power density. However, too large a slot area will lead to teeth narrowing and serious saturation. Therefore, the torque performance and motor loss should be comprehensively considered in optimizing stator structure.

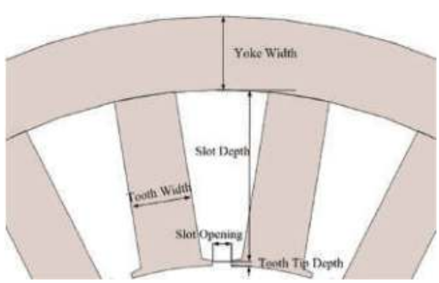


Figure 13. Stator structure schematic diagram.

Saturation can be defined as the following: stator core can be called an unsaturated state when the magnetic flux density is in the linear region of the B–H curve, and a saturated state is when the magnetic flux density is higher than the maximum magnetic density in the linear region. In the Ver2.0 motor, to alleviate excessive rotor loss at high speed and maintain motor torque performance, JFE super iron 10JNEX900, which has a high saturation magnetic flux density, low iron loss, and high permeability, is selected. The saturated magnetic flux intensity corresponding to 10JNEX900 is 1.8 T.

In order to find the optimal stator structure for the 60 kW electric vehicle high power density motor Ver2.0, stator tooth width was changed under the condition that the fixed stator outer diameter, each phase parallel branch current, and the slot copper space factor are the same. Under different stator tooth widths, the electromagnetic characteristic change curves at three operating conditions are shown in Figure 14.

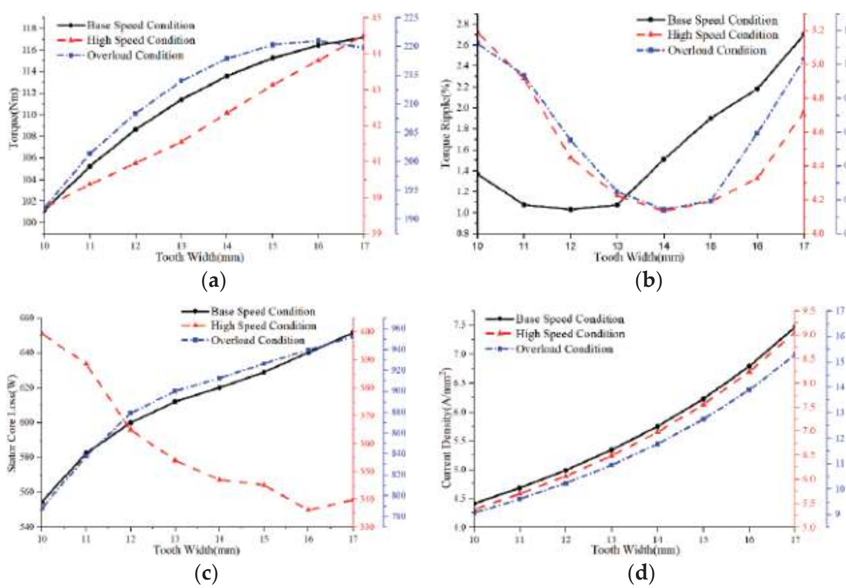
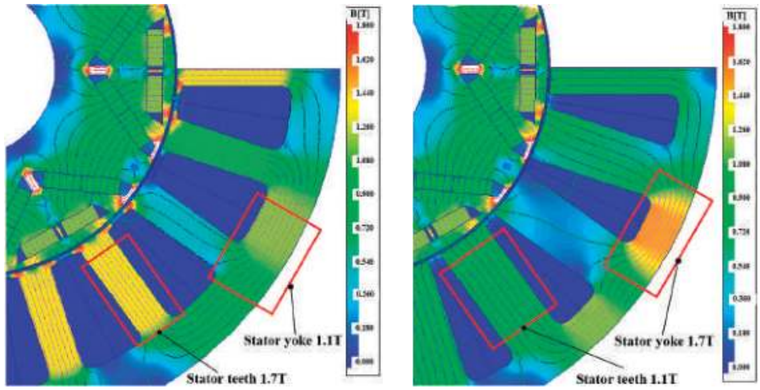


Figure 14. Electromagnetic characteristics vary with different tooth width: (**a**) output torque, (**b**) torque ripple, (**c**) stator core loss, (**d**) current density.

When the tooth width increases, the output torque becomes larger, and the torque ripple increases first and then decreases. Motor torque ripple at a high-speed condition is much larger than in the other two operating conditions. When the tooth width becomes wider, the tooth saturation can be effectively alleviated, and the output torque increases correspondingly. However, the magnetic circuit of the stator runs through the stator tooth and stator yoke. When the tooth width exceeds the yoke width, the yoke saturation degree deepens, and the torque ripple increases, so the torque ripple shows a trend of decreasing first and then increasing. Figure 15 shows the stator teeth and yoke saturation in stator teeth width of 11 mm and 20 mm, respectively.



(**a**) (**b**)

Figure 15. Stator flux density distribution in different stator tooth widths: (**a**) tooth width of 11 mm, and (**b**) tooth width of 20 mm.

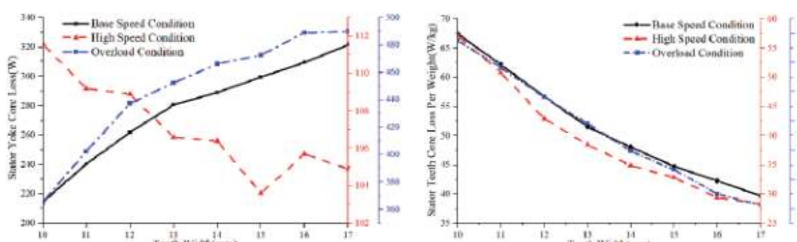
According to Figure 14c, it can be seen that, at the base speed condition and the overload condition, the stator core loss increases with the tooth width widening, while at a high-speed condition, the stator core loss decreases with the tooth width widening. To find the reason for the two types of stator loss, the stator tooth loss and stator yoke loss should be calculated separately according to the following expression:

$$p_{\text{yoke}} = p_{10/50} B_{\text{yoke}}^2 \left(\frac{f}{50}\right)^{1.3} G_{\text{yoke}} \tag{7}$$

$$p_{\text{tooth}} = p_{10/50} B_{\text{tooth}}^2 \left(\frac{f}{50}\right)^{1.3} G_{\text{tooth}} \tag{8}$$

In the formula, p_{yoke} and p_{tooth} are the core loss of the yoke and teeth, respectively, B_{yoke} and B_{tooth} are the average flux density of the yoke and tooth, respectively, in T, and G_{yoke} and G_{tooth} are the core weight of the yoke and the tooth, respectively, in kg.

The stator core loss at base speed condition generally shows an upward trend, in which the stator yoke core loss gradually increases with the tooth width, and the stator tooth core loss gradually decreases with the tooth width. When the tooth width becomes wider, the stator tooth saturation degree decreases, so the loss of the stator tooth decreases, as can be seen in Figure 16a.



(**a**) (**b**)

Figure 16. Stator core loss under different tooth widths: (**a**) stator yoke core loss, (**b**) stator teeth core loss per weight.

While stator tooth width increases, stator tooth allows more magnetic flux lines to travel, but it also requires that the stator yoke travels more magnetic flux lines. As described earlier, stator yoke width should be kept constant, so that magnetic flux lines may be borne under a too wide a stator tooth, as shown in Figure 15b, which will cause the saturation of the stator yoke and increase the stator yoke core loss. At the overload condition, the winding current density is about twice the density measured at the base speed condition. Even if the tooth width is widened, the loss of the yoke is still saturated, and the stator total core loss shows an upward trend. The motor in the high-speed condition is in the flux-weakening state, the motor magnetic field is further weakened, and both tooth flux density and yoke flux density drop significantly, so the total stator core loss shows a downward trend.

On the whole, considering the motor torque performance and motor loss, 13 mm tooth width is selected for the Ver2.0 motor.

3.3. Motor Electromagnetic Characteristic under Different Modulation Algorithms

Compared with a three-phase motor, a five-phase motor has higher reliability and a degree of control freedom. Reliability is mainly reflected in that when one phase or two phases have faults such as winding short circuit fault or winding open fault, the remaining phases can maintain the original operating condition. As for the degree of control, in a five-phase motor, it is reflected in the voltage vector space. In a three-phase motor, there is only one length between any two phases, while in the five-phase motor, there are two lengths. Therefore, expect a zero-coordinate system, the five-phase motor has the fundamental coordinate system and third harmonic coordinate system, which is called the fundamental α_1–β_1 coordinate system, and third harmonic α_3–β_3 coordinate system in this article.

Figure 17 shows the conventional five-phase half-bridge inverter topological structure. It can be seen that the five-phase inverter has 2^5 switching states, corresponding to 32 space voltage vectors. These 32 voltage vectors can be divided into 10 long vectors, 10 medium length vectors, 10 small vectors, and 2 zero vectors according to the vector length in the fundamental α_1–β_1 coordinate system. Moreover, these vectors in the fundamental α_1–β_1 coordinate system vector distribution and the third harmonic α_3–β_3 coordinate system vector distribution are shown in Figure 18, respectively.

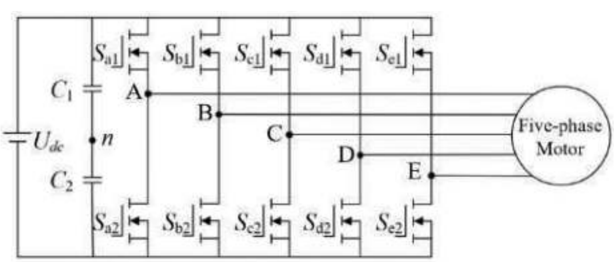


Figure 17. The topology of the five-phase half-bridge inverter.

Similar to the three-phase motor SVPWM modulation algorithm for using two adjacent voltage vectors to obtain the motor working voltage vector, the five-phase motor can also synthesize motor rotating vector using two adjacent long vectors in the fundamental coordinate system, which are usually called near two vectors SVPWM (NTV-SVPWM) [23]. However, when adopting the NTV-SVPWM modulation algorithm, the third harmonic will be generated in the third harmonic coordinate system, resulting in excessive voltage harmonics of the motor. At the same time, using NTV-SVPWM does not take advantage of the five-phase motor's high freedom control, causing bad motor electromagnetic characteristics such as high torque ripple, high line voltage, and high loss.

To solve the above problems, near four vectors SVPWM (NFV-SVPWM) modulation algorithm is proposed. NFV-SVPWM modulation algorithm uses two adjacent long vectors and two medium length vectors to synthesize the working voltage vector. NFV-SVPWM can adjust medium length vector length and long vector length by action time. In a fundamental coordinate system, the four vectors combined into a working voltage vector ensure the motor's normal operation as the NTV-SVPWM. However, in the third-harmonic coordinate system, the synthetic vectors multiplied by the vector length and action time can be arranged into a parallelogram so that the third harmonic can be eliminated in NFV-SVPWM.

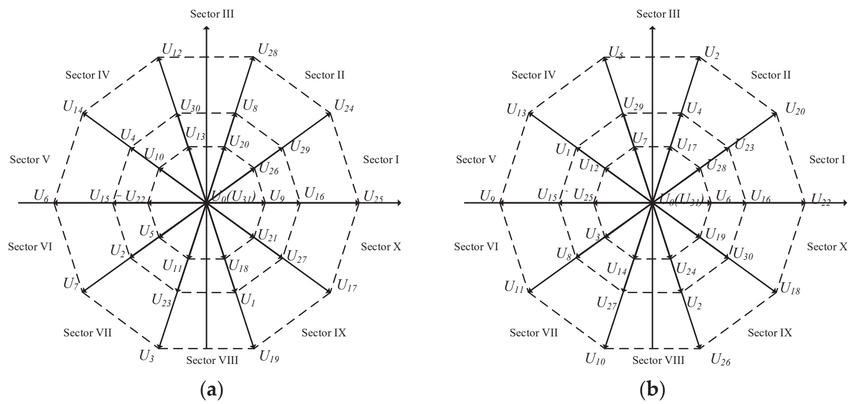


Figure 18. Space voltage vector distribution in the fundamental and third harmonic coordinate system. (**a**) Fundamental α_1–β_1 coordinate system vector distribution. (**b**) third harmonic α_3–β_3 coordinate system vector distribution.

Figure 19 shows the maximum torque per ampere (MTPA) control of the five-phase PMSM block diagram. The NTV-SVPWM and NFV-SVPWM modulation algorithms are respectively applied, and the corresponding current excitation results obtained are shown in Figure 20 for the power device operating at a 15 kHz switching frequency.

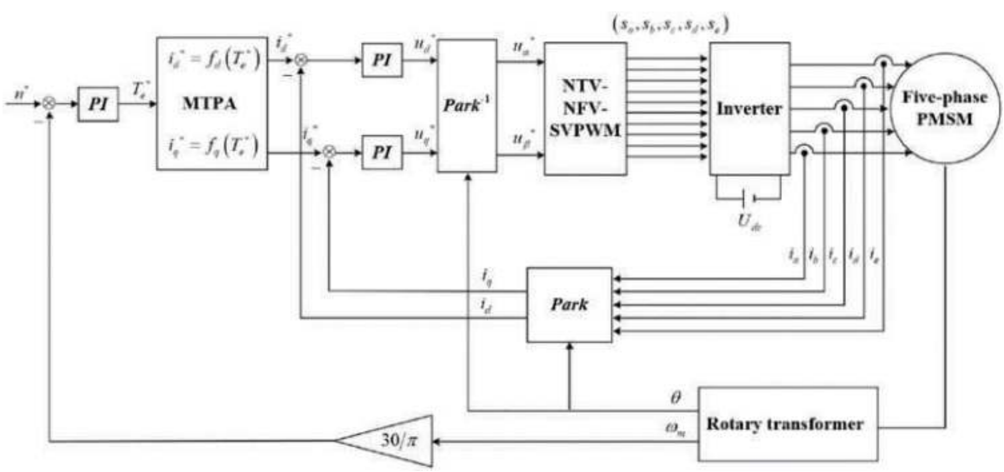
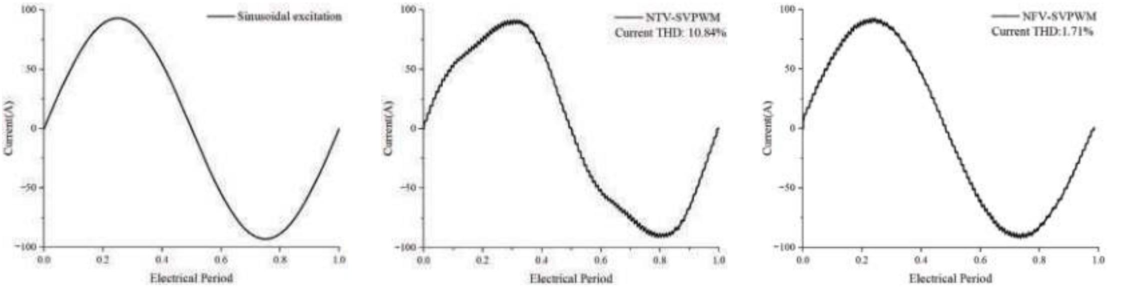


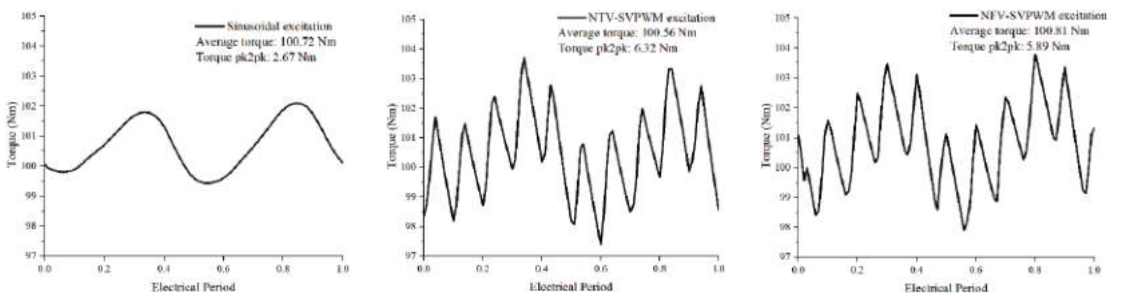
Figure 19. MPTA control of the five-phase PMSM block diagram.

Different modulation algorithms can influence the motor's electromagnetic characteristics. Figure 21 shows motor torque at the base speed condition using Figure 20 current excitation. In addition, motor loss in different modulation algorithms is summarized in Table 4.



(**a**) (**b**) (**c**)

Figure 20. Current excitation under different modulation algorithm: (**a**) ideal sinusoidal, (**b**) NTV-SVPWM, and (**c**) NFV-SVPWM.



(**a**) (**b**) (**c**)

Figure 21. Motor torque under different modulation algorithms (**a**) ideal sinusoidal, (**b**) NTV-SVPWM, and (**c**) NFV-SVPWM.

Table 4. Motor loss under different modulation algorithms.

Loss	Ideal Sinusoid	NTV-SVPWM	NFV-SVPWM
Stator core loss	492.36 W	1320.9 W	1230.4 W
Rotor core loss	29.24 W	66.8 W	62 W
Magnet loss	12.58 W	40.25 W	33.67 W
Copper loss	332.45 W	321.1 W	317.84 W

It can be seen from Figure 21 that, compared with the NTV-SVPWM modulation algorithm, the NFV-SVPWM modulation algorithm has the smallest torque ripple and an increase in core loss, and magnet loss caused by the NTV-SVPWM modulation algorithm's high current harmonics. Large torque ripple influences electric vehicle's comfort level, while high loss affects the IMDS temperature distribution, increasing the difficulty in IMDS thermal management. Based on electromagnetic characteristics, NFV-SVPWM is selected as the driver modulation algorithm.

When the inverter switching frequency is 15 kHz, the core loss in the stator is almost twice that of the motor under ideal sinusoidal excitation, which shows that the inverter can influence the motor's electromagnetic characteristic. Compared with ideal sinusoidal excitation, the current obtained by the modulation algorithm contains harmonics, which are generated by the comparison between modulating wave and carrier wave. When the modulation wave frequency is low, the current harmonic content is large; when the carrier frequency is high, the current harmonic content is small. The switching frequency affects the modulation frequency; a lower switching frequency leads to serious current distortion because of which the motor cannot perform with the electromagnetic characteristics of the ideal design. In order to reduce the driver-side influence on the motor, the switching frequency should be appropriately selected, and the motor's electromagnetic characteristics under different switching frequencies should be studied so as to achieve great motor electromagnetic characteristics. Table 5 shows the motor loss distribution of each part when the switching frequency changes from 3 kHz to 45 kHz.

Table 5. Motor loss under different switching frequencies using NFV-SVPWM.

Switching Frequency	Current THD	Stator Core Loss	Rotor Core Loss	Magnet Loss
Ideal sinusoid	0	492.4 W	29.2 W	12.6 W
3 kHz	8.06%	754.5 W	80.5 W	201.9 W
9 kHz	2.88%	675.5 W	68.5 W	75.9 W
15 kHz	1.71%	615.4 W	62 W	33.7 W
30 kHz	1.54%	568.2 W	46.8 W	23.5 W
45 kHz	1.37%	522.9 W	31.6 W	13.2 W

As can be seen from Table 5, with the increase in switching frequency, the distortion rate of the motor current waveform gradually decreases, and the motor loss component gradually approaches that of the sinusoidal excitation. It can be concluded that the increase in switching frequency is beneficial in reducing motor loss, especially the decrease in eddy current loss of permanent magnet and stator core loss.

However, when the switching frequency is too high, the power device switching loss will also increase. Although the motor driver adopts SiC devices, which can effectively reduce the power device switching loss, when the switching frequency is too high, power device switching loss will still be large. Therefore, the switching frequency of the power tube is selected as 30 kHz.

3.4. Ver2.0 Electromagnetic Characteristic

Through the selection of motor structure, slot/pole combination, IPMSM rotor structure, rotor magnetic bridge position, and optimization analysis of the stator and rotor, the final IPMSM topology structure is shown in Figure 22, where Figure 22a is Ver1.0 motor topology, and Figure 22b is Ver2.0 motor topology.

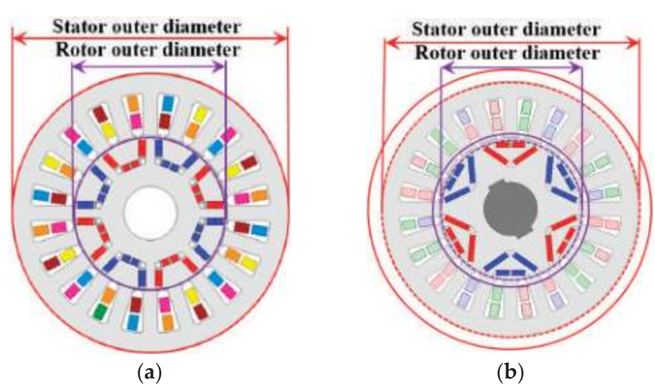


Figure 22. Motor in IMDS topology: (**a**) Ver1.0 motor topology, (**b**) Ver2.0 motor topology.

The Red and purple lines in Figure 22a,b represent the stator outer diameter and the rotor outer diameter, respectively, and the solid-line curve and dash-line curve represent Ver1.0 motor size and Ver2.0 motor size, respectively. It can be concluded that, through the design method of the high power density motor in this article, the stator and rotor diameters are significantly reduced in the Ver2.0 motor compared with the Ver1.0 motor.

The electric vehicle motor designed in this article has three operating conditions which are called the base speed operating condition, high-speed condition, and an overload condition. Base speed is the motor's long time operation condition, which requires high efficiency and reliability. Overload is an instantaneous condition, which requires the motor to have high torque capacity, but high torque usually means large current density, large copper loss, and serious distorted iron loss, leading to motor electromagnetic characteristic reduction. High speed is also an instantaneous condition, which requires the motor to have a certain torque capacity at high speeds. In high-speed conditions, the current frequency is so high that the motor core loss increases heavily, affecting the motor's life and safety. Considering that the motor loses too much under high-speed and overload conditions, which brings high motor temperature problems, besides adopting segmented PM to reduce magnet eddy current loss, motor material needs to be selected emphatically. The materials selected for Ver2.0 are expressed in Table 6. The Ver2.0 motor selected 10JNEX900 as the stator and rotor core because of the material's high saturation magnetic flux density, low iron loss, and high permeability advantages, as stated above. In addition, the N38UH was selected for Ver2.0 as the magnet material on account of its high magnetic energy product and excellent demagnetization curve linearity at a high temperature of 150 °C.

Table 6. Ver2.0 motor's space dimensions.

Component	Material	Mass	Volume
Stator	10JNEX900	14.01 kg	1.87 L
Rotor	10JNEX900	6.17 kg	0.83 L
Winding	Copper	10.14 kg	1.14 L
Magnet	N38UH	1.49 kg	0.19 L
Shaft	20CrMnTi	2.88 kg	0.39 L
Overall	-	35.2 kg	4.42 L

The Ver2.0 motor's spatial dimensions of each part are shown in Table 6. The Ver2.0 motor's main size and structure after optimization can be seen in Table 7. The Ver2.0 driving performance is shown in Table 8.

Table 7. Ver2.0 motor's main size and structure.

Parameters	Ver1.0	Ver2.0
Stator outer diameter	260 mm	240 mm
Rotor inner diameter	145 mm	133 mm
Air gap length	1 mm	1 mm
Stator teeth width	14 mm	13 mm
Magnet segment	3	9
Axial length	90 mm	90 mm
Slot/pole combination	20/8	20/6
Rotor structure	U	∇

Table 8. Ver2.0 motor and Ver1.0 motor electromagnetic characteristics.

Parameters	Base Speed		High Speed		Overload	
	Ver1.0	Ver2.0	Ver1.0	Ver2.0	Ver1.0	Ver2.0
Speed	6000 rpm		15,000 rpm		5000 rpm	
Output torque	101 Nm	105 Nm	47 Nm	48 Nm	199 Nm	209 Nm
Torque ripple	3.9%	3.5%	19.9%	4.64%	12.54%	1.16%
Line voltage amplitude	626 V	635 V	641 V	596 V	639 V	569 V
Line voltage THD	22.9%	9.2%	75.4%	56.2%	46.5%	12.9%
Current density	5.3 A/mm^2		10.9 A/mm^2		6.5 A/mm^2	
Efficiency	95.6%	96.9%	90.6%	90.4%	94.6%	95.5%

Comparing the Ver1.0 motor and Ver2.0 motor designed by the above methods and processes, the motors have smaller torque ripple because of the larger output torque, which is mainly attributed to the double-layer rotor structure and suitable slot/pole combination to improve the PM utilization rate and reduce the MMF harmonic content. In addition, while improving the motor torque performance, line voltage THD is greatly reduced, and the line voltage pressure under high speed and overload conditions can be relieved.

The Ver1.0 motor's peak mass power density is 2.94 kW/kg, and the peak volume power density is 12.9 kW/L. The Ver2.0 motor, which is designed by the high power density method in this article, has a peak mass power density of 3.12 kW/kg and a peak volume power density of 15.19 kW/L. It is worth noting that when calculating the volume power density, the volume is selected as the cylinder volume composed of the total axial length, which includes winding end axial length, stator axial length, and the stator outer diameter. It can be seen that compared with the Ver1.0 motor, the Ver2.0 motor power density is higher.

According to the torque performance, the efficiency, DC bus voltage pressure, and the motor power density verify that the motor designed by the above method has better performance.

4. Analysis of Temperature Field of HPDM

HPDM requires a motor with a small volume and large output power, but too high output power will lead to large copper loss, core loss, and windage loss. However, HPDM has a small space/size for heat dissipation. If the motor temperature field is not analyzed, the temperature will become the limit of the HPDM. At higher temperatures, there will be hidden dangers of irreversible demagnetization of the permanent magnet and winding insulation inside the motor, resulting in the internal failure and reduced reliability of the motor.

4.1. Temperature Field Solution Model

Conventional cooling methods can be divided into active cooling and passive cooling. Passive cooling mainly relies on natural convection to dissipate heat and does not require additional cooling circuits. Active cooling mainly relies on forced convection of fluids and requires additional cooling circuits. The vehicle IMDS has high power, high speed, and a large loss of each component. Passive cooling cannot effectively dissipate heat from the vehicle's high-power machine. Therefore, this paper uses water-cooled active cooling with strong heat dissipation capacity to dissipate heat.

Using the traditional steady-state thermal circuit method to calculate the temperature rise cannot output the accurate 3D temperature field distribution of the motor. Therefore, this section calculates the temperature field of the motor based on the numerical finite element method. The temperature field solution model includes a water jacket, front and back end caps, stator and rotor cores, windings, magnets, shafts, and bearings. The temperature field solution model of the established 60 kW HPDM is shown in Figure 23.

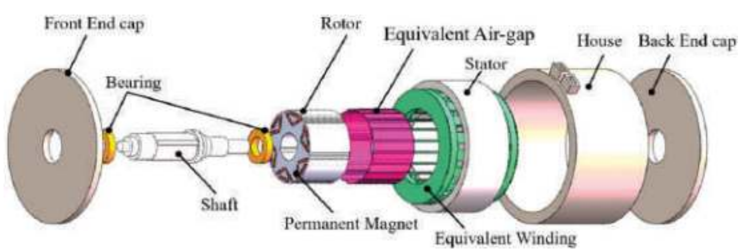


Figure 23. A 60 kW high power density machine temperature field solution model.

It is worth mentioning that, in order to reduce the model complexity, the rotary transformer installed at the front end of the bearing for position detection is not drawn in the temperature field solution model. Additionally, for the same reason of reducing the simulation difficulty and obtaining more accurate thermal simulation results, the temperature field solution model is equivalent to the winding and the stator–rotor air gap. The winding replaces all bare copper wires in the slot with a copper winding, which is placed in the center of the slot; the other parts in the slot, except the copper winding, are represented by equivalent insulation, and the heat transfer coefficient is treated equivalently according to the quality.

4.2. Water Jacket Structure

The water jacket provides radial heat dissipation for the motor and takes away most of the heat. However, the water jacket has various structures, and the heat dissipation capacity and pressure loss of different water jackets are different for HPDM because of the motor's large power and small heat dissipation area. Therefore, the most effective heat dissipation capacity should be taken as the key consideration factor in the selection of water jackets. This paper compares the three water jacket structures: spiral type water jacket structure I, dial type water jacket structure II, and axial water jacket structure III, as shown in Figure 24.

(**a**) (**b**) (**c**)

Figure 24. Different cooling water jacket structures (**a**) spiral, (**b**) dial, and (**c**) axial.

To ensure that different water jackets have the same axial length and 6 L/min fluid flow, a simplified motor temperature field solution model including water jackets, end caps, stator core, and equivalent winding is used. The heat source density of the stator core and the winding heat source density are set as the base speed condition corresponding to the heat sources 329,196 W/m^3 and 544,613 W/m$^{3.}$ At the initial temperature of the water flow of 65 °C, the flow rate, pressure loss, and temperature comparison of each component of the three structures are shown in Figure 25.

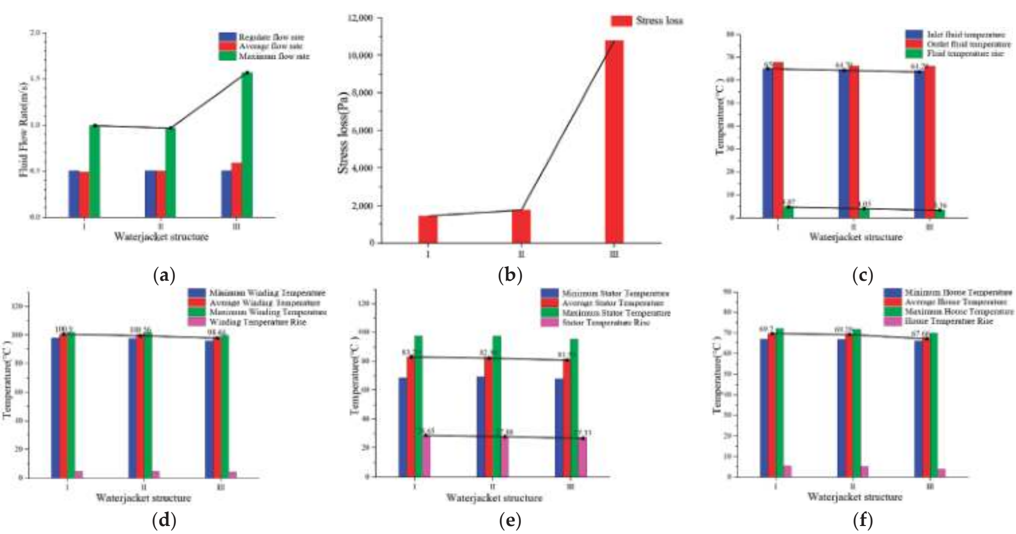


Figure 25. Comparison of three water jacket structures: (**a**) fluid flow rate, (**b**) stress loss, (**c**) fluid temperature, (**d**) winding temperature, (**e**) stator temperature, and (**f**) water jacket temperature.

By comparing the flow velocity, pressure loss, and temperature results of the three structures in Figure 25, it is found that the axial water jacket has the strongest heat dissipation capacity, the dial water jacket is the second, and the spiral water jacket structure is the worst. However, from the perspective of pressure loss, it can be seen that the spiral water jacket has the smallest pressure loss because of almost no turning heads, followed by the dial water jacket structure. The axial water jacket structure has the largest number of turning heads and the largest pressure loss. The dial channel structure is the best choice for comprehensive heat dissipation capacity and pressure loss. However, for HPDM, the base speed condition is not the most serious internal heating condition. Considering the extreme conditions of high speed and overload, the axial water jacket structure III had the best heat dissipation capacity. So, the axial water jacket structure should be selected for better thermal management inside the Ver2.0 motor.

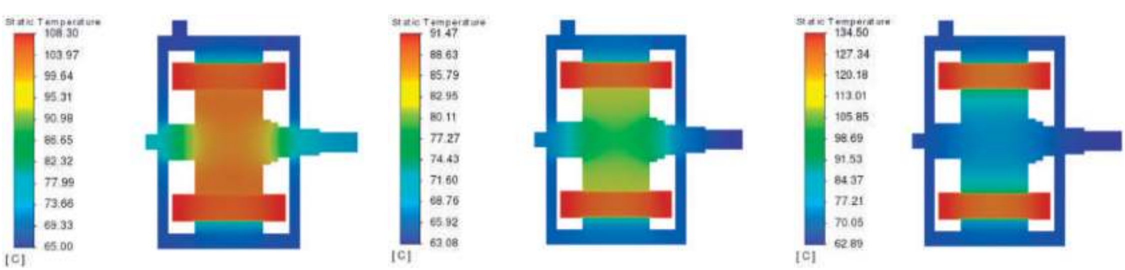
4.3. Temperature Solution of Main Operating Conditions

The main heating components of the HPDM designed in this paper are stator core, rotor core, winding, and the permanent magnet. Under the three main operating conditions, the heat generation rate of the heating element is obtained by dividing the loss of each heating element by the volume of the heating element. The heat generation rate of each element is shown in Table 9.

Table 9. Motor heat generation rate under main operating condition.

Component	Base Speed	High Speed	Overload
Stator yoke	268,248 W/m^3	110,862 W/m^3	451,831 W/m^3
Stator tooth	491,935 W/m^3	278,326 W/m^3	430,712 W/m^3
Winding	543,908 W/m^3	799,036 W/m^3	2,275,536 W/m^3
Rotor	165,534 W/m^3	450,413 W/m^3	145,721 W/m^3
Permanent magnet	15,772 W/m^3	46,481 W/m^3	37,487 W/m^3

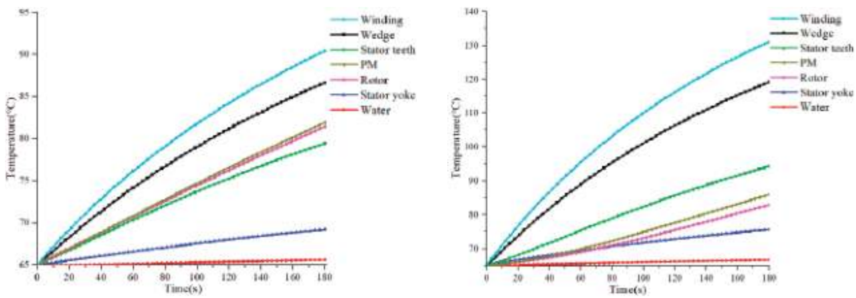
Using the established HPDM temperature field solution model, the temperature field is solved for the base speed condition, high-speed condition, and an overload condition by assigning relevant boundary conditions, material properties, and heat generation rate of heat-generating components. Among them, the base speed condition is a long-term working point, and its steady-state temperature is observed. The high-speed condition and the overload condition are short-time working conditions, and the temperature distribution of each part of the motor is observed under the operation for 180 s. The initial temperature of the water flow is given as 65 °C, the temperature section diagrams under three working conditions are shown in Figure 26, and the temperature change curves of each component under transient conditions are shown in Figure 27.



(**a**) (**b**) (**c**)

Figure 26. Temperature field solution results under three main operation conditions: (**a**) base speed conditions, (**b**) high-speed conditions, and (**c**) overload conditions.

From the temperature field solution results, it can be found that under the three main operating conditions of the base speed, high speed, and overload, the windings are the most serious parts of heat generation, and the highest temperature occurs at the end of the winding at the overload point, which is 135 °C. The temperature rises to 70 °C, and the maximum temperature rise do not exceed 155 °C and 105 °C for Class F insulation. The PMs in both the steady-state and the transient-state do not exceed the maximum allowable temperature of the material N38UH, which is 150 °C. The rotary transformer selected for Ver2.0 is TAMAGAWA TS2225N903E102; its allowable temperature, according to the manual, is −40 °C to 150 °C. In this article, the motor's cooling system solutions' rotary transformer can operate normally. It can be seen that the axial water jacket structure can effectively manage the heat of the designed 60 kW high power density motor.



(**a**) (**b**)

Figure 27. Temperature change curve under transient operating conditions: (**a**) high-speed conditions, (**b**) overload Conditions.

5. Conclusions

This paper researched the design method of the HPDM for electric vehicles IMDS. Through the method, a five-phase high power density motor with a rated power of 60 kW, a maximum speed of 15,000 r/min, and a peak power density of 15.19 kW/L has been designed. Compared with the Ver1.0 motor designed before, the Ver2.0 motor power density is higher. In addition, the temperature field is checked for the three main operating conditions of base speed, high speed, and overload. The following conclusions are drawn:

(1) The motor power density is related to the phase number, winding utilization, copper space factor, air gap flux density, and motor rotational speed. When designing a high power density motor, the above factors can be weighed;

(2) Increasing IPMSM rotor PM layer number can make air gap flux density step up, increase air gap flux density amplitude, and improve the sinusoidal degree at the same time, which is beneficial to improving the motor torque performance. For a five-phase motor, FSDW has less end winding length and torque ripple than an ISDW and has a higher winding coefficient compared with a five-phase motor with FSCW. FSDW is suitable for the five-phase motor with the high power density and electromagnetic characteristics in some cases;

(3) Driver modulation algorithm affects motor electromagnetic characteristics. Comparing the NTV-SVPWM modulation algorithm with the NFV-SVPWM modulation algorithm, NFV-SVPWM can eliminate harmonics in the third harmonic plane by taking advantage of the five-phase motor's two vector planes and obtain higher sinusoidal output excitation. Through the comparative analysis of the output current excitations with different switching frequencies, it is found that a high switching frequency is more conducive to improving motor electromagnetic characteristics and reducing motor loss. However, the switching frequency should be selected considering power device switching loss;

(4) For the three water jackets—the spiral, dial, and axial—the spiral structure has the smallest pressure loss and the worst heat dissipation capacity. The axial structure has the largest pressure loss and the strongest heat dissipation capacity.

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Abbreviations

The following abbreviations are used in this manuscript:

MDS	Motor drive system
IMDS	Integrated motor drive system
THD	Total harmonic distortion
HPDM	High power density motor
SPMSM	Surface-mounted permanent magnet synchronous machine
IPMSM	Interior permanent magnet synchronous machine
PM	Permanent magnet
RMS	Root mean square
MMF	Magnetic motive force
FSCW	Fractional slot concentrated winding
FSDW	Fractional slot distributed winding
ISDW	Integral slot distributed winding
MTPA	Maximum torque per ampere

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Article

Design and Analysis of a 30 kW, 30,000 r/min High-Speed Permanent Magnet Motor for Compressor Application

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Abstract: In this paper, the design and analysis of a 30 kW, 30,000 r/min high-speed permanent magnet motor (HSPMM) for compressor application are provided. In order to provide a reasonable electromagnetic design scheme, the electromagnetic performances of the HSPMM under different structures are analyzed and compared by the finite element method (FEM). The thermal performances and cooling system of the HSPMM are, respectively, analyzed and designed by computational fluid dynamics (CFD). Finally, the HSPMM's rotor strength is studied by both FEM and analytical methods, and the influencing factors of which are also researched in this paper.

Keywords: high-speed permanent magnet motor; FEM; electromagnetic design; CFD; thermal analysis; analytical method; rotor strength

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1. Introduction

High-speed permanent magnet motors (HSPMMs) have the advantages of compact structure, high power density, and high efficiency [1–3]. Compared with the traditional method of driving high-speed loads by using a constant speed motor and gearbox, it is more effective to reduce the volume of the driving system, increase power density and improve operation efficiency by using HSPMM directly, so that HSPMM has a promising future and a wide range of applications such as high-speed compressors, flywheel energy storage, and so on. However, the performances of HSPMM in multiple physical fields also need to be further investigated because of their high frequency, high speed, and high loss density [4]. Therefore, a 30 kW, 30,000 r/min HSPMM for compressor application is designed and analyzed in multi-physics fields in this paper.

Firstly, this paper provides analyses of the influencing factors of the HSPMM's electromagnetic performances such as the motor main dimensions, the number of poles, the number of stator slots, and the magnetization method of PM.

Due to the high power density and compact size, the heat dissipation capacity of HSPMM is limited, which may result in the demagnetization of PM. Therefore, the thermal analysis and cooling system design of the HSPMM are particularly necessary.

Considering the PM materials are fragile in tensile, it is necessary to install a sleeve outside the rotor for mechanical protection. This paper analyzes the rotor stress distribution of the HSPMM by both analytical and FEM methods. The influencing factors such as sleeve thickness and interference are also analyzed in this paper.

2. Electromagnetic Analysis

The electromagnetic performances of HSPMMs with different structures are presented and compared by FEM in this chapter [5,6]. As for material selection, NdFeB is selected as the PM material due to its high remanence and coercivity, and a silicon steel sheet is selected as the core material.

2.1. Main Dimensions of Motor

The stator inner diameter D and the motor active length L_{ef} which are called the main dimensions of the motor play an important role in motor design. The two main dimensions can be calculated by the following formulas:

$$D^2 L_{\mathrm{ef}} = \frac{6.1}{\alpha'_{\mathrm{P}} K_{\mathrm{nm}} K_{\mathrm{dp}} A B_{\delta}} \cdot \frac{P'}{n} \tag{1}$$

$$\lambda = \frac{L_{\mathrm{ef}}}{\tau} \tag{2}$$

$$\tau = \frac{\pi D}{2P} \tag{3}$$

where P' is the apparent power, n is the motor speed, α_{P}' is the equivalent pole arc coefficient, K_{nm} is the air gap flux factor, K_{dp} is the winding factor, A is the line load, B_{δ} is the air gap flux density, λ is the main dimension ratio, τ is the pole distance, and P is the number of pole pairs.

The estimated values of each parameter in the above formula are shown in Table 1. Therefore, it can be roughly obtained that the D and the L_{ef} are 45 mm and 105 mm, respectively, which still needs to be further adjusted according to the subsequent analysis results.

Table 1. Calculation parameters of main dimensions of motor.

Parameters	Value	Parameters	Value
P' (kW)	27	A (A/m)	50,000
n (r/min)	30,000	B_{δ} (T)	0.7
α_{P}'	0.7	λ	3
K_{nm}	1.11	P	2
K_{dp}	0.92		

Then, the air gap thickness and the outer diameter of the rotor can be set to 1.5 mm and 42 mm, which also need to be further optimized.

2.2. Number of Poles

The number of poles plays an important role in HSPMMs' electromagnetic performances. In order to limit the operating frequency of HSPMMs and suppress the losses caused by high-frequency current and high-frequency magnetic flux density, the number of poles of HSPMMs should better be set as two or four. The structural parameters and electromagnetic performances of two HSPMMs with two poles and four poles, respectively, are analyzed and compared in this paper. The two HSPMMs are designed with the same power (30 kW), same speed (30,000 r/min), and the same stator slots number (12).

The two HSPMMs' structural parameters are presented and compared in Table 2. It can be found the two poles motor has a lower frequency (500 Hz) than the four poles one (1000 Hz), which is beneficial to reduce stator iron loss and winding copper loss. However, the two poles motor has a larger size and lower power density than the four poles motor.

The two motors' back EMF and air gap magnetic flux density with FFT analysis results under no-load conditions are compared and shown in Figures 1 and 2, respectively.

Table 2. Structural parameters of 2 and 4 poles HSPMMs.

Parameters	2 Poles	4 Poles
Power (kW)	30	30
Speed (r/min)	30,000	30,000
Frequency (Hz)	500	1000
Stator outer diameter (mm)	118	103
Stator inner diameter (mm)	45	45
Slot number	12	12
Core length (mm)	130	105
Rotor outer diameter (mm)	42.2	42.2
Number of winding turns	7	8
Winding pitch	5	2

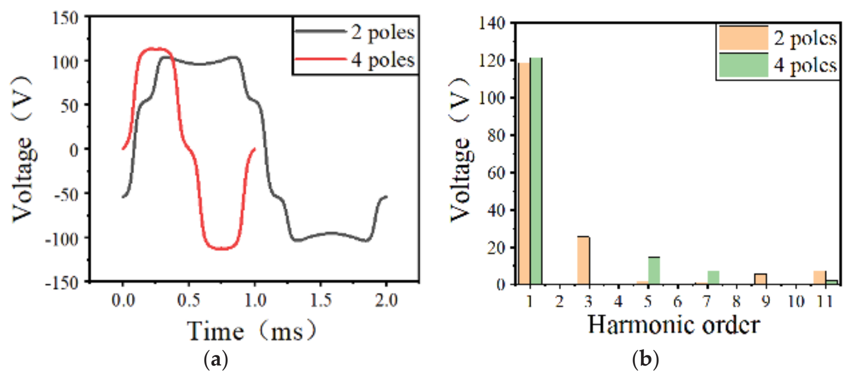


Figure 1. Waveforms and FFT analysis results of back EMF for 2 poles and 4 poles HSPMMs. (**a**) Waveforms; (**b**) FFT analysis results.

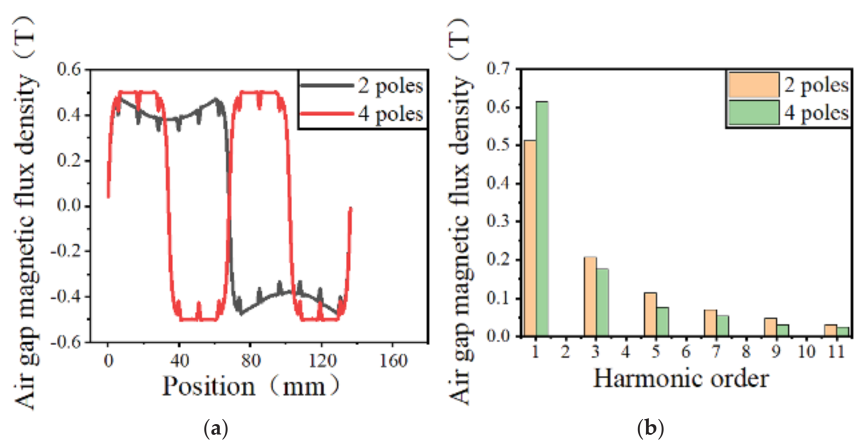


Figure 2. Waveforms and FFT analysis results of air gap magnetic flux density for 2 poles and 4 poles HSPMMs. (**a**) Waveforms; (**b**) FFT analysis results.

Figure 1 shows the two motors' back EMF waveforms with FFT analysis results. It can be found that the four poles motor has a larger fundamental component, the fifth harmonic component, and the seventh harmonic component, while other harmonic components are smaller than the two poles motor, especially the third harmonic component. In general, the back EMF of the four poles motor has lower harmonics, and the waveform of which is closer to the sine wave.

The two motors' air gap magnetic flux density waveforms with FFT analysis results are presented and compared in Figure 2. Each harmonic component of the air gap magnetic flux density of the four poles motor is lower than the two poles motor except for the fundamental component, indicating the four poles structure has better performance in reducing harmonics. Additionally, adopting the four poles structure for HSPMMs can also improve the utilization of PMs due to their higher fundamental component of air gap magnetic flux density.

The two motors' performances under full-load condition are also compared in this paper and shown in Figure 3 and Table 3. Due to the lower frequency, the two poles HSPMM has lower stator iron loss and winding copper loss than the four poles. Although the four poles HSPMM has lower eddy current loss and wind friction loss because of their smaller size and lower harmonics, the efficiency of the two poles HSPMM is still higher than the four poles HSPMM. Additionally, it can be found from Figure 3b and Table 3 that the two motors have the same average torque under full-load conditions, while the four poles motor has a larger torque ripple (13.84%) than the two poles (5.32%).

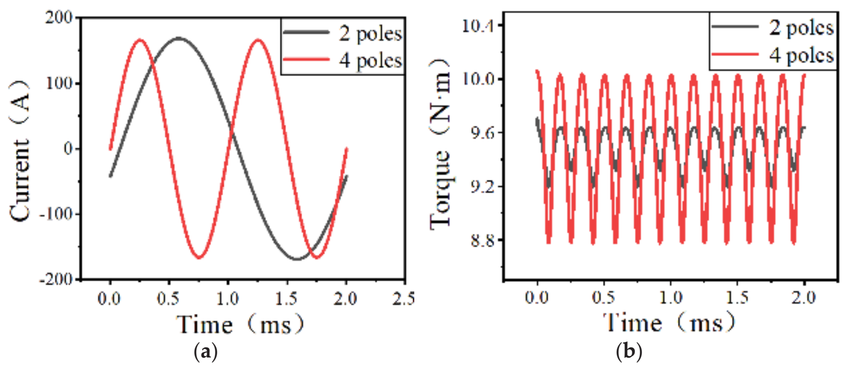


Figure 3. The comparisons of current and torque waveforms of the HSPMM under full-load conditions. (**a**) Current waveforms; (**b**) torque waveforms.

Table 3. Performances of 2 and 4 poles HSPMMs under full-load condition.

Parameters	2 Poles	4 Poles
RMS current (A)	117.10	117.28
Torque (N·m)	9.48	9.49
Torque ripple (%)	5.32	13.84
Stator iron loss (W)	321.43	558.68
Winding copper loss (W)	257.22	264.09
Rotor eddy current loss (W)	218.19	210.48
Wind friction loss (W)	72.53	58.58
Efficiency (%)	97.10	96.36

Finally, this paper chooses to adopt the four poles structure due to its lower harmonics, smaller size, higher power density, and lower rotor eddy current loss.

2.3. Number of Slots

As shown in Figure 4, there are three normally used stator structures of HSPMMs including slotless structure, less slot structure, and multi-slot structure. Considering the advantages of higher air-gap flux density, higher utilization of PMs, and lower harmonic amplitude, there is no doubt that the multi-slot structure has been the first choice in the design of HSPMM.

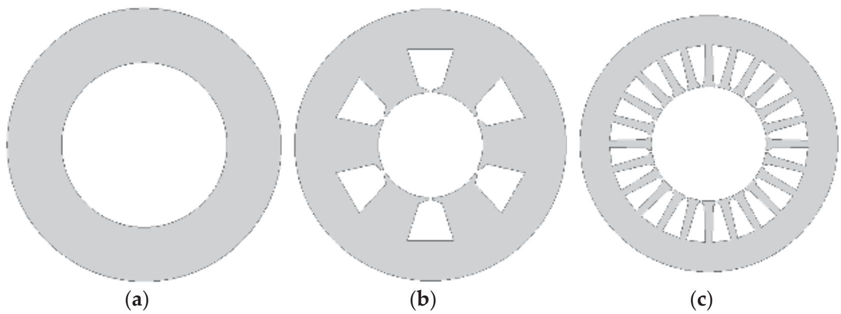


Figure 4. Stator structures of HSPMMs. (**a**) Slotless structure; (**b**) less slot structure; (**c**) multi slot structure.

In this paper, 3 HSPMMs with 12 slots, 18 slots, and 24 slots, respectively, are designed and compared to figure out the influence of the number of slots on HSPMMs. The three HSPMMs are designed with the same power, speed, frequency, number of poles, and size for a fair comparison.

Figure 5 shows the three HSPMMs' back EMF waveforms with FFT analysis results. It can be found that the 24 slots HSPMM has the largest fundamental component. Except for the third harmonic component, the other order harmonic components of the 24 slots HSPMM are lower than other motors, indicating that the multi-slot structure can help reduce harmonic components of back EMF.

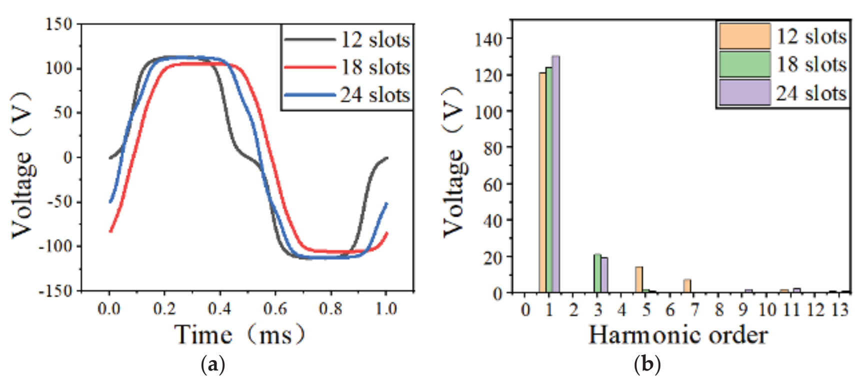


Figure 5. Waveforms and FFT analysis results of back EMF for 12 slots, 18 slots, and 24 slots HSPMMs. (**a**) Waveforms; (**b**) FFT analysis results.

The three HSPMMs' air gap magnetic flux density waveforms with FFT analysis under no-load conditions are presented in Figure 6. Due to the same rotor structure, the air gap magnetic flux density waveforms and FFT analysis results of the three motors are very similar to each other.

The performances of the three HSPMMs under full-load conditions are also compared and presented in Table 4. With the increase in the number of slots, the RMS current and torque ripple of HSPMMs is decreasing gradually. The three motors have nearly equal stator iron loss and wind friction loss since they have the same frequency and rotor size. The 18 slots motor has the lowest winding copper loss, though the RMS current of which is larger than the 24 slot one. The eddy current loss of 24 slots motor is much lower than other motors, which further proves that increasing the number of slots has good performance in the reduction of the harmonic components and the eddy current loss of HSPMMs.

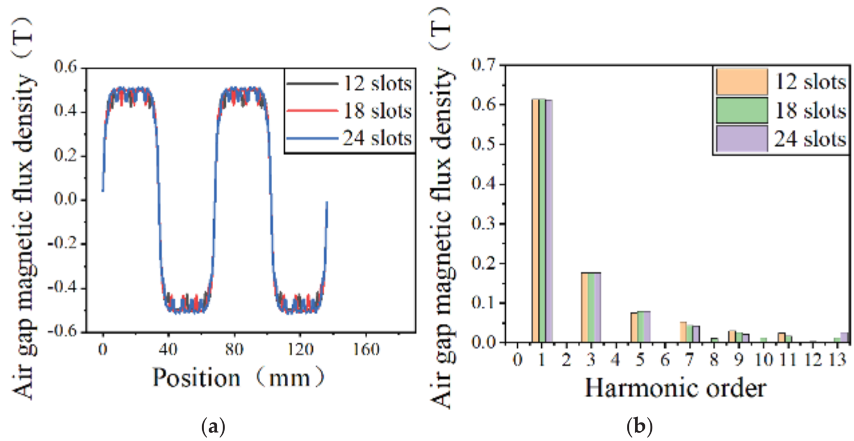


Figure 6. Waveforms and FFT analysis results of air gap magnetic flux density for 12 slots, 18 slots, and 24 slots HSPMMs. (**a**) Waveforms; (**b**) FFT analysis results.

Table 4. Performances of 12 slots, 18 slots and 24 slots HSPMMs under full-load condition.

Parameters	12 Slots	18 Slots	24 Slots
RMS current (A)	117.28	114.61	108.59
Torque (N·m)	9.49	9.54	9.54
Torque ripple (%)	13.84	5.05	2.95
Stator iron loss (W)	558.68	554.87	558.81
Winding copper loss (W)	264.09	225.73	244.81
Rotor eddy current loss (W)	210.48	47.15	21.22
Wind friction loss (W)	58.58	58.58	58.58
Efficiency (%)	96.36	97.05	97.06

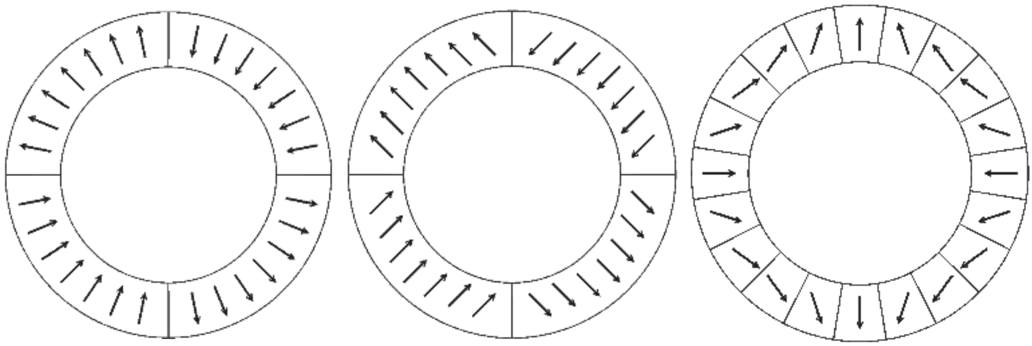
In summary, the 24 slots structure is a more reasonable choice for HSPMMs.

2.4. PM Magnetization Methods

The PM magnetization methods of HSPMMs mainly include radial magnetization, parallel magnetization, and Halbach magnetization, as shown in Figure 7. In order to figure out the influence of PM magnetization methods on the electromagnetic performances of HSPMMs, three HSPMMs with different PM magnetization methods are designed and compared in this paper. The three HSPMMs have the same power, speed, and frequency.

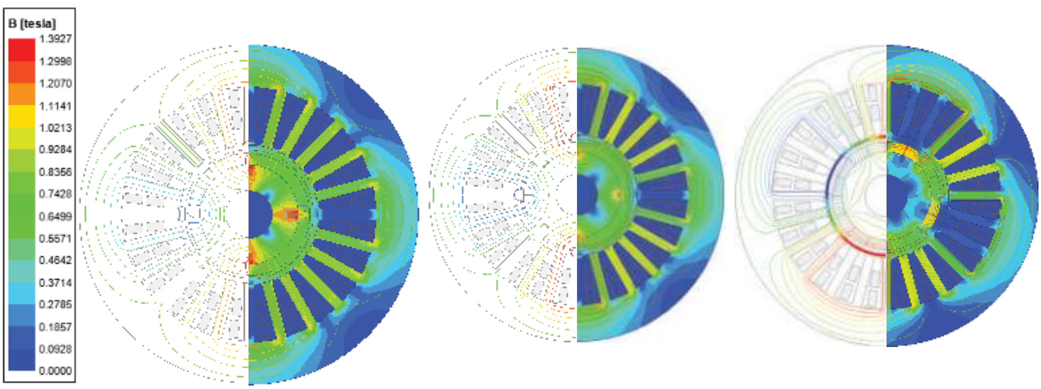
Figure 8 shows the magnetic flux density and line distributions of the three HSPMMs. According to Figure 8, It can be found that the color of the magnetic flux density cloud diagram in Figure 8c is generally lighter, indicating that the magnetic flux density of the motor with the Halbach magnetization method is lower than the other motors.

The three HSPMMs' back-EMF waveforms with FFT analysis results are presented in Figure 9. Then the RMS value of the total back EMF of the three motors calculated by FEM are 91.31 V (radial), 84.96 V (parallel), and 79.63 V (Halbach), respectively, indicating the HSPMM with the Halbach magnetization method has the smallest RMS value of total back EMF. Additionally, the back EMF waveform of the HSPMM with the Halbach magnetization method is closest to a sine wave due to its lower harmonic components.



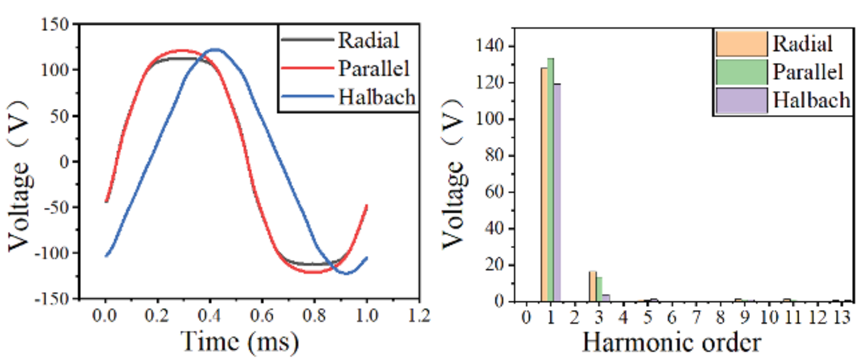
(**a**) (**b**) (**c**)

Figure 7. PM magnetization methods. (**a**) Radial magnetization; (**b**) parallel magnetization; (**c**) Halbach magnetization.



(**a**) (**b**) (**c**)

Figure 8. Magnetic flux density and lines distributions of HSPMMs with different magnetization methods under no-load condition. (**a**) Radial magnetization; (**b**) parallel magnetization; (**c**) Halbach magnetization.



(**a**) (**b**)

Figure 9. Waveforms and FFT analysis results of back EMF for HSPMMs with different magnetization methods. (**a**) Waveforms; (**b**) FFT analysis results.

The air gap magnetic flux density waveforms with FFT analysis results for the three HSPMMs are presented in Figure 10. It is learned that each harmonic component of air gap magnetic flux of the HSPMM with the Halbach magnetization method is lower than other motors except for the fifth harmonic component.

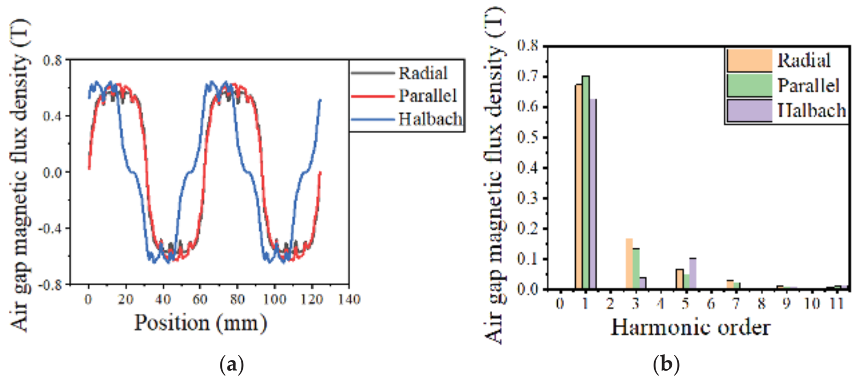


Figure 10. Waveforms and FFT analysis results of air gap magnetic flux density for HSPMMs with different magnetization methods. (**a**) Waveforms; (**b**) FFT analysis results.

The three HSPMMs' performances under full-load conditions are presented in Table 5. It can be learned that the HSPMM with PM parallel magnetized has the lowest current, the lowest torque ripple, the lowest rotor eddy current loss, and the highest efficiency.

Table 5. Performances of the three HSPMMs with different PM magnetization methods under full-load condition.

Parameters	Radial	Parallel	Halbach
RMS current (A)	110.89	106.15	119.02
Torque (N·m)	9.55	9.56	9.55
Torque ripple (%)	2.07	1.42	1.99
Stator iron loss (W)	511.09	478.63	437.87
Winding copper loss (W)	303.32	273.81	378.37
Rotor eddy current loss (W)	11.84	10.23	11.37
Wind friction loss (W)	33.16	33.16	33.16
Efficiency (%)	97.22	97.42	97.21

2.5. Final Electromagnetic Design Scheme

Based on the previous analysis, the electromagnetic design scheme of a 30 kW, 30,000 r/min HSPMM is finally completed in this paper. Figure 11 shows the current and torque waveforms of the HSPMM under full-load conditions, and the designed motor's structure and electromagnetic parameters are presented in detail in Table 6.

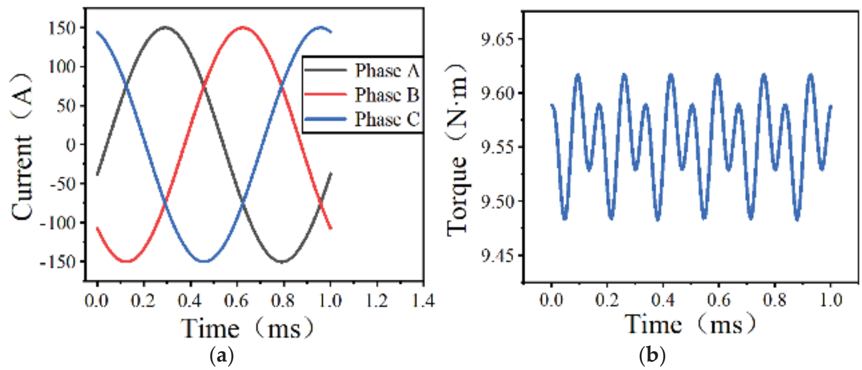


Figure 11. The current and torque waveforms of the HSPMM under full-load conditions. (**a**) Current waveform; (**b**) torque waveform.

Table 6. Structure and electromagnetic parameters of the 30 kW, 30,000 r/min HSPMM.

Parameters	Value	Parameters	Value
Power (kW)	30	Pole number	4
Speed (r/min)	30000	Slot number	24
Frequency (Hz)	1000	Stator outer diameter (mm)	100
RMS current (A)	106.15	Stator inner diameter (mm)	41
Torque (N·m)	9.56	Core length (mm)	103
Torque ripple (%)	1.42	Air gap thickness (mm)	1.5
Stator iron loss (W)	478.63	Rotor outer diameter (mm)	38
Winding copper loss (W)	273.81	Sleeve thickness (mm)	1
Rotor eddy current loss (W)	10.23	PM material	NdFeB
Wind friction loss (W)	33.16	PM thickness (mm)	3.4
Efficiency (%)	97.42	Magnetization direction	Parallel

3. Thermal Analysis

HSPMMs have limited heat dissipation capability due to their compact size and high loss density so a reasonable cooling system design is particularly necessary for HSPMMs. In this paper, CFD is selected to provide the thermal analysis for HSPMMs and study the cooling effects of different water-cooling methods [7,8].

3.1. Cooling Channel Structure

The structure of the three cooling channels is presented in Figure 12. Then, the comparative analysis of the thermal performances of HSPMMs with different cooling channel structures is performed in this paper. The three cooling channels are designed with the same inner surface area (13,229 mm^2), thickness (3 mm), and volume (40,774 mm^3) for comparison.

The calculation results of the heat transfer coefficient for the three HSPMMs are presented in Figure 13. It can be found the heat transfer coefficient of the spiral structure channel is higher than the other structures, indicating the spiral structure channel has better performance in heat dissipation. Therefore, as shown in Table 7, the maximum temperature of each component of the HSPMM with a spiral structure channel is the lowest of the three HSPMMs due to its higher heat transfer coefficient.

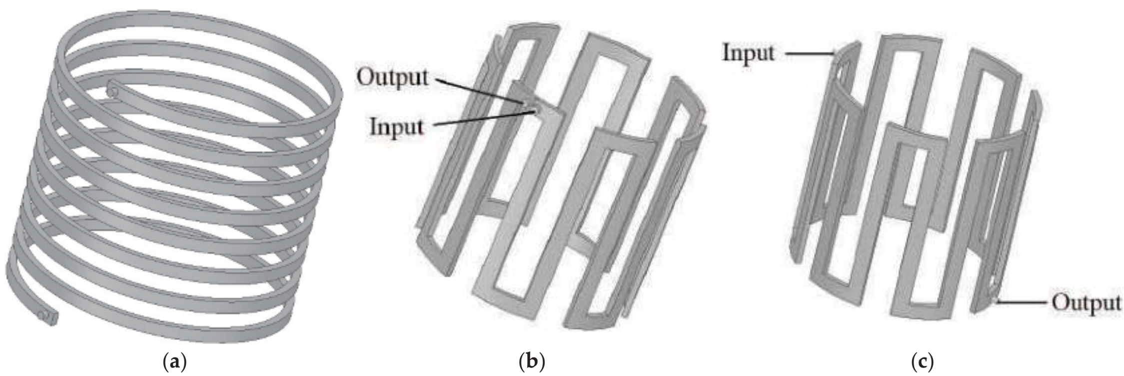


Figure 12. Cooling channel structures. (**a**) Spiral structure; (**b**) axial structure 1; (**c**) axial structure 2.

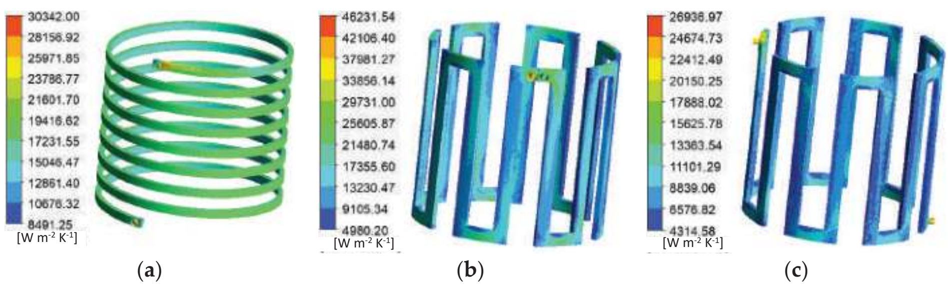


Figure 13. Heat transfer coefficient distribution of the three cooling channel structures. (**a**) Spiral structure; (**b**) axial structure 1; (**c**) axial structure 2.

Table 7. The maximum temperature of each part of motors with different cooling channel structures.

Parameters	Spiral	Axial 1	Axial 2
Cooling water (°C)	36.49	38.67	44.53
Stator core (°C)	66.08	66.63	68.80
Stator winding (°C)	69.96	70.84	73.99
Sleeve (°C)	124.89	125.01	131.91
PM (°C)	124.81	124.92	131.79

Figure 14 shows the pressure distribution of the three cooling channel structures. The axial channel 1 has the largest inlet pressure due to its larger channel angle and flow resistance. Correspondingly, the spiral channel has a smaller channel angle and flow resistance which contribute to its lower inlet pressure. In axial channel 2, the cooling water flows along two paths on both sides of the channel so that the actual flow distance is only half of the spiral channel and the axial channel 1. Therefore, the inlet pressure of the axial channel 2 is lower than other channels though it has a larger channel angle and flow resistance, too.

3.2. Number of Cooling Channel Turns

Considering the spiral cooling channel has a better cooling effect than other channel Structures, this paper provides further analysis of the number of spiral cooling channel turns.

The thermal analysis results of HSPMMs with a different number of cooling channel turns are shown in Figure 15. It can be found that with the rise of the number of cooling channel turns, the maximum temperature of cooling water and motor gradually reduced so that increasing the number of cooling channel turns is helpful to enhance the cooling effect.

However, when the number of channel turns is larger than eight, each motor component's maximum temperature hardly changed anymore, and even the cooling water maximum temperature rose slightly. Meanwhile, the eight-turn cooling channel has the largest heat transfer coefficient. Therefore, the number of cooling channel turns should be designed more reasonably with the considerations of both cooling effect and manufacturing difficulty.

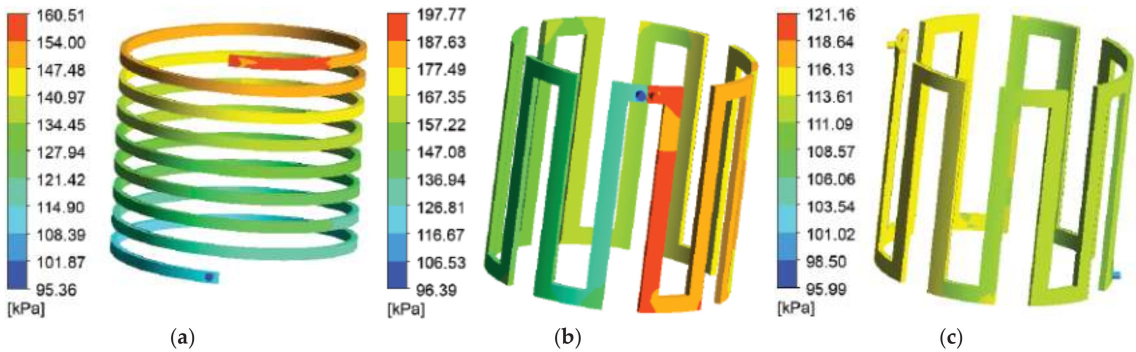


Figure 14. Pressure distribution of the three cooling channel structures. (**a**) Spiral structure; (**b**) axial structure 1; (**c**) axial structure 2.

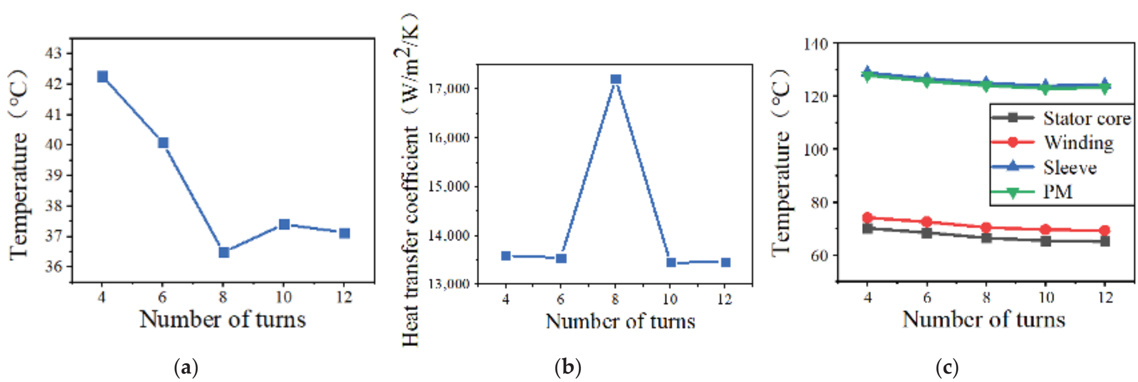


Figure 15. Thermal analysis results of HSPMMs with different number of cooling channel turns. (**a**) Cooling water maximum temperature; (**b**) Heat transfer coefficient; (**c**) Motor maximum temperature.

3.3. Cooling Water Flow Rate

The flow rate of cooling water also plays an important role in motor heat dissipation. Figure 16 shows the thermal analysis results of an HSPMM at different cooling water flow rates. It is learned that with the rise of the cooling water flow rate, the maximum temperature of the cooling water and the motor gradually decreased while the heat transfer coefficient of the cooling channel increased, indicating that increasing the cooling water flow rate can effectively enhance the cooling effect. However, when the flow rate is up to 3 m/s, the maximum temperature of the motor and the cooling water changes more slowly, indicating that continuing to increase the flow rate has few influences on the cooling effect, though the heat transfer coefficient still keeps increasing. Additionally, a high flow rate will also lead to an increase in inlet pressure and water pump power consumption.

Therefore, it is more reasonable to set the cooling water flow rate to 3 m/s.

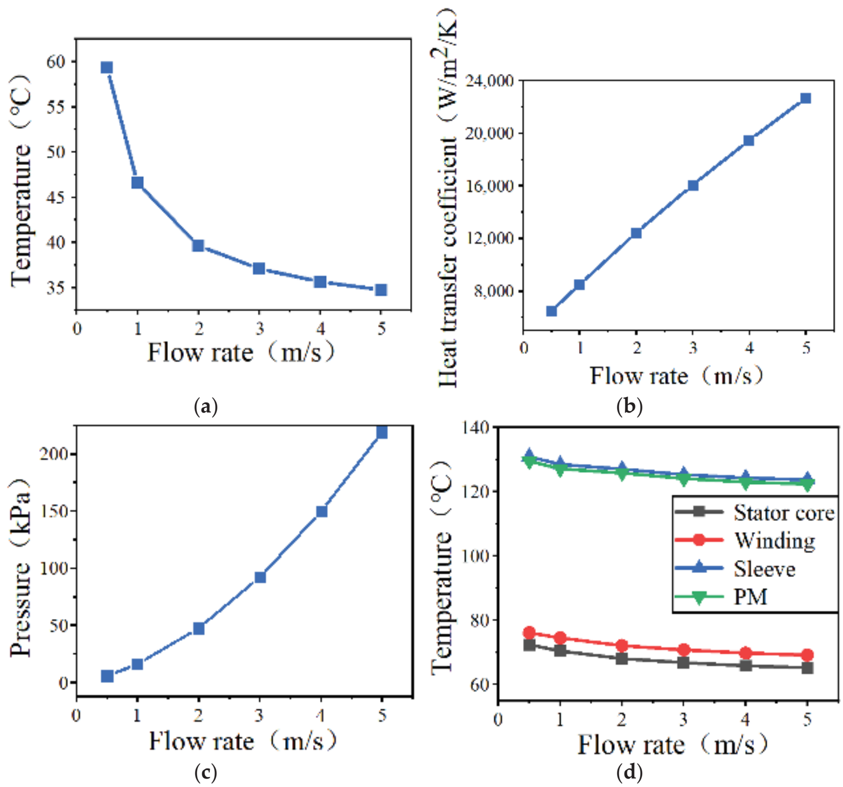


Figure 16. Thermal analysis results of HSPMMs at different cooling water flow rates. (**a**) Cooling water maximum temperature; (**b**) heat transfer coefficient; (**c**) inlet pressure; (**d**) motor maximum temperature.

3.4. Temperature Calculation Result

Finally, an eight turns spiral water channel is selected to protect the designed motor from overheating. The water flow rate is set as 3 m/s. The motor's temperature distribution calculated by CFD is shown in Figure 17. It can be found that the maximum temperature of the PM is 81.74 °C, which is much lower than the NdFeB's maximum permitted working temperature (180 °C), indicating the HSPMM designed in this paper has lower rotor loss density and the designed cooling method has better heat dissipation effect.

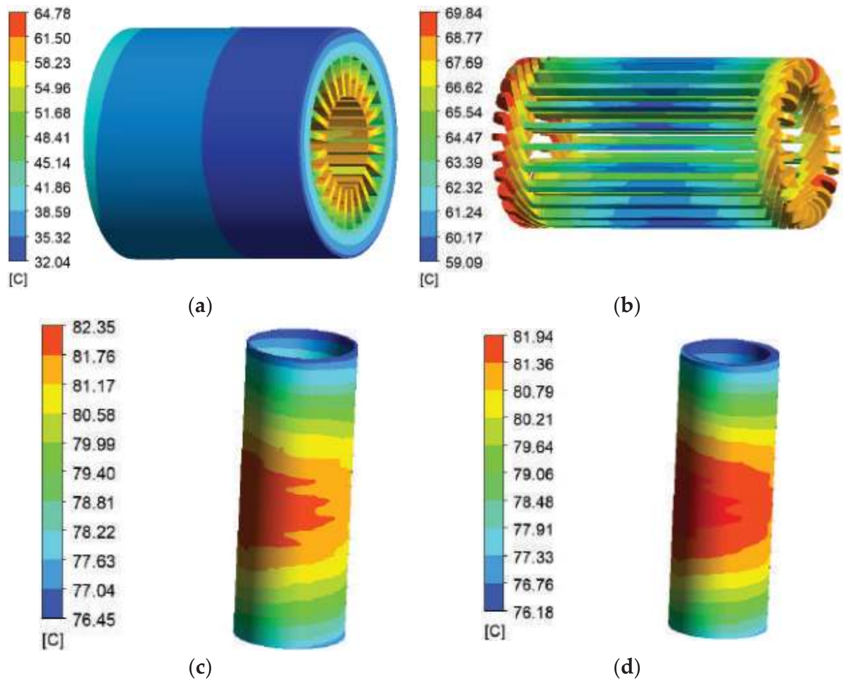


Figure 17. Temperature distribution of the designed motor. (**a**) Stator; (**b**) winding; (**c**) sleeve; (**d**) PM.

4. Stress Analysis

This paper provides the stress analysis for the sleeve–PM–core–shaft structure rotor of HSPMM by both the analytic method and FEM method [9]. The stainless sleeve with interference is adopted to protect the PM due to its better heat dissipation and strength. The stress analysis model of that structure rotor is shown in Figure 18.

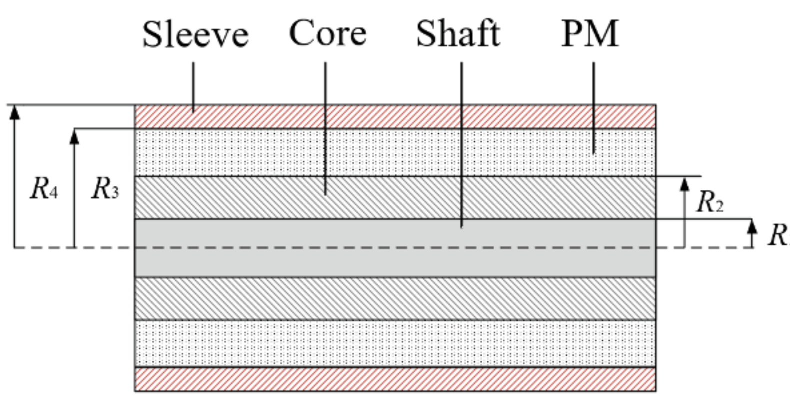


Figure 18. Stress analysis model of the sleeve–PM–core–shaft structure rotor.

4.1. Analytic Method

The sleeve–PM–core–shaft structure rotor can be equivalent to four thick-walled cylinder models, and the adjacent models are closely fitted. The pressure on the inner and outer surfaces of each cylinder model can be considered uniformly distributed. All the relevant properties of the cylinder model materials are isotropic. Then, a single-cylinder model is analyzed at first.

According to the material mechanics theory and geometric relationship, the strain-displacement relationship of the cylinder model can be described by [10]:

$$\begin{cases} \varepsilon_{\mathrm{r}} = \frac{du}{dr} \\ \varepsilon_{\theta} = \frac{du}{dr} \end{cases} \tag{4}$$

where the ε_{r} is the cylinder radial strain, ε_{θ} is the cylinder tangential strain, u is the cylinder displacement.

According to the Hooke's law, the stress–strain relationship can be described by:

$$\begin{cases} \varepsilon_{\mathrm{r}} = \frac{\sigma_{\mathrm{r}} - \mu\sigma_{\theta}}{E} + \alpha\Delta T \\ \varepsilon_{\theta} = \frac{\sigma_{\theta} - \mu\sigma_{\mathrm{r}}}{E} + \alpha\Delta T \end{cases} \tag{5}$$

where the σ_{r} is the cylinder radial stress, σ_{θ} is the cylinder tangential stress, μ is the cylinder Poisson's ratio, E is the cylinder young's modulus, α is the cylinder thermal expansion coefficient, ΔT is the cylinder temperature rise.

The force balance equation of the cylinder in the rotating condition can be described by:

$$\frac{d\sigma_{\mathrm{r}}}{dr} + \frac{\sigma_{\mathrm{r}} - \sigma_{\theta}}{r} + \rho\omega^2 r = 0 \tag{6}$$

where the ρ is the cylinder density, ω is the cylinder rotational angular velocity.

According to Equations (4)–(6), the u, σ_{r}, and σ_{θ} of a single-cylinder model can be obtained as follows:

$$u = Mr + \frac{N}{r} + \frac{(\mu^2 - 1)\rho\omega^2 r^3}{8E} \tag{7}$$

$$\begin{cases} \sigma_{\mathrm{r}} = \frac{ME}{1-\mu} - \frac{NE}{(1+\mu)r^2} - \frac{(3+\mu)\rho\omega^2 r^2}{8} - \frac{E\alpha\Delta T}{1-\mu} \\ \sigma_{\theta} = \frac{ME}{1-\mu} + \frac{NE}{(1+\mu)r^2} - \frac{(1+3\mu)\rho\omega^2 r^2}{8} - \frac{E\alpha\Delta T}{1-\mu} \end{cases} \tag{8}$$

where the M and N is the calculation coefficient.

According to Equations (7) and (8), the calculation equations for the displacement and stress of the sleeve–PM–core–shaft structure rotor can be obtained as follows:

$$u_{\mathrm{a}} = M_{\mathrm{a}} r + \frac{N_{\mathrm{a}}}{r} + \frac{({\mu_{\mathrm{a}}}^2 - 1)\rho_{\mathrm{a}}\omega^2 r^3}{8E_{\mathrm{a}}} \tag{9}$$

$$\begin{cases} \sigma_{\mathrm{ra}} = \frac{M_{\mathrm{a}}E_{\mathrm{a}}}{1-\mu_{\mathrm{a}}} - \frac{N_{\mathrm{a}}E_{\mathrm{a}}}{(1+\mu_{\mathrm{a}})r^2} - \frac{(3+\mu_{\mathrm{a}})\rho_{\mathrm{a}}\omega^2 r^2}{8} - \frac{E_{\mathrm{a}}\alpha_{\mathrm{a}}\Delta T_{\mathrm{a}}}{1-\mu_{\mathrm{a}}} \\ \sigma_{\theta\mathrm{a}} = \frac{M_{\mathrm{a}}E_{\mathrm{a}}}{1-\mu_{\mathrm{a}}} + \frac{N_{\mathrm{a}}E_{\mathrm{a}}}{(1+\mu_{\mathrm{a}})r^2} - \frac{(1+3\mu_{\mathrm{a}})\rho_{\mathrm{a}}\omega^2 r^2}{8} - \frac{E_{\mathrm{a}}\alpha_{\mathrm{a}}\Delta T_{\mathrm{a}}}{1-\mu_{\mathrm{a}}} \end{cases} \tag{10}$$

where the a = 1, 2, 3, and 4, corresponding to the equations of the shaft, core, PM, and sleeve, respectively. Considering that the displacement at the center of the shaft is 0, so that the N_1 is 0.

The stress and displacement boundary conditions for the sleeve–PM–core–shaft structure rotor are as follows:

$$\begin{cases} u_1(r = R_1) = u_2(r = R_1) \\ u_2(r = R_2) = u_3(r = R_2) \\ u_4(r = R_3) - u_3(r = R_3) = \delta \\ \sigma_{\mathrm{r}1}(r = R_1) = \sigma_{\mathrm{r}2}(r = R_1) \\ \sigma_{\mathrm{r}2}(r = R_2) = \sigma_{\mathrm{r}3}(r = R_2) \\ \sigma_{\mathrm{r}3}(r = R_3) = \sigma_{\mathrm{r}4}(r = R_3) \\ \sigma_{\mathrm{r}4}(r = R_4) = 0 \end{cases} \tag{11}$$

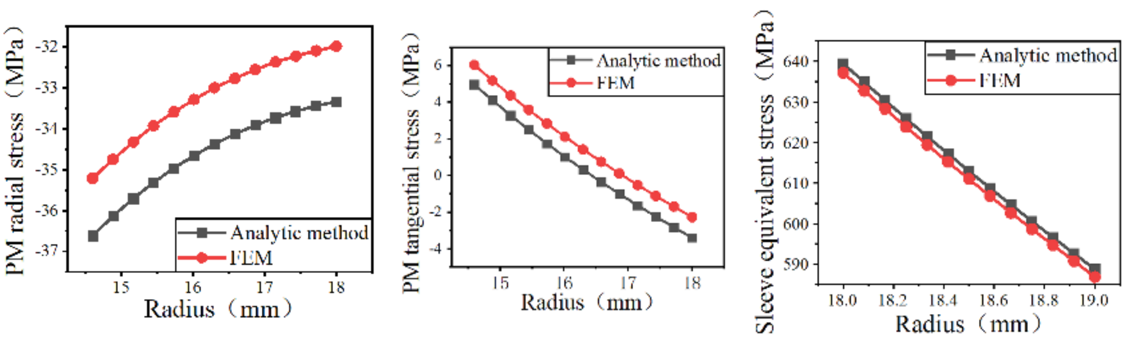
where the δ is the interference between sleeve and PM.

Finally, the calculation coefficients in Equations (9) and (10) and the complete stress analytical equations of the sleeve–PM–core–shaft structure rotor can be obtained by Equation (11).

4.2. FEM Verification

In order to verify the accuracy of the proposed analytic method, this paper adopts both the analytic method and FEM to calculate the stress distributions of a sleeve–PM–core–shaft structure rotor for comparison.

The stress distributions calculated by the two methods are shown in Figure 19. It can be found that there are certain acceptable errors in both the PM radial stress and tangential stress calculated by the two methods. Meanwhile, the analytical result of sleeve equivalent stress agrees well with the FEM one and the calculation error rate between the two methods is only 0.36%. Therefore, it can be considered that the analytic method provided by this paper can accurately calculate the stress distribution of the sleeve–PM–core–shaft structure rotor.



(**a**) (**b**) (**c**)

Figure 19. Stress distribution of the sleeve–PM–core–shaft structure rotor. (**a**) PM radial stress; (**b**) PM tangential stress; (**c**) sleeve equivalent stress.

4.3. Influencing Factors

Both the sleeve thickness and the interference have significant effects on the rotor stress distribution. The maximum stress of the rotor variations with sleeve thicknesses and interferences is presented in Figure 20. It can be found that increasing the interference is more conducive to reducing the maximum tensile stress of the PM in both radial and tangential directions, while the maximum equivalent stress of the sleeve will increase accordingly. Additionally, the larger thickness of the sleeve also has the advantage of reducing the tensile stress of the PM, but it has little effect on the reduction of sleeve equivalent stress.

4.4. Final Rotor Mechanical Protection Scheme

According to the above analysis, this paper finally adopts a stainless steel sleeve of 1 mm to mechanically protect the rotor of the HSPMM, and the interference between the sleeve and the PM is set to 0.05 mm. Then, the designed motor's rotor stress distribution is calculated by FEM and the result of which is shown in Figure 21. It is learned that the PM bears compressive stress in both radial and tangential directions, while the PM material has a strong tolerance to compressive stress. Meanwhile, the maximum equivalent stress of the sleeve is 537.19 MPa, which is lower than the permitted value of stainless steel (1100 MPa). Therefore, the final rotor mechanical protection scheme designed in this paper is reliable.

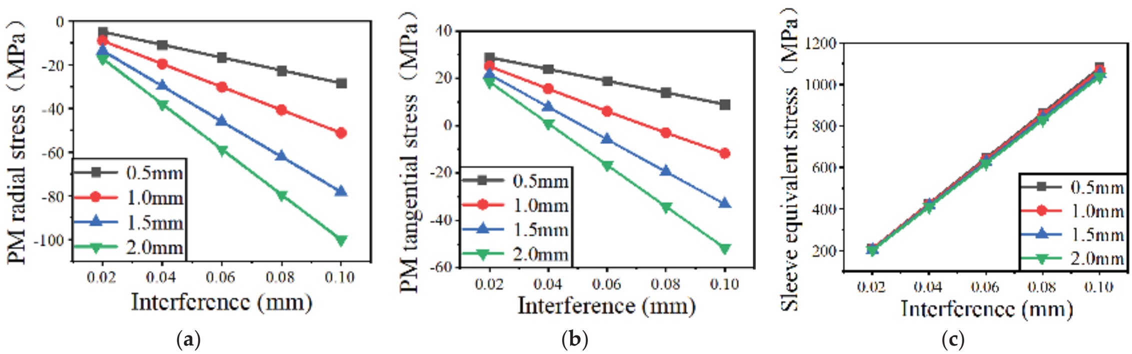


Figure 20. The maximum stress of the rotor variations with sleeve thickness and interference. (**a**) PM maximum radial stress; (**b**) PM maximum tangential stress; (**c**) sleeve maximum equivalent stress.

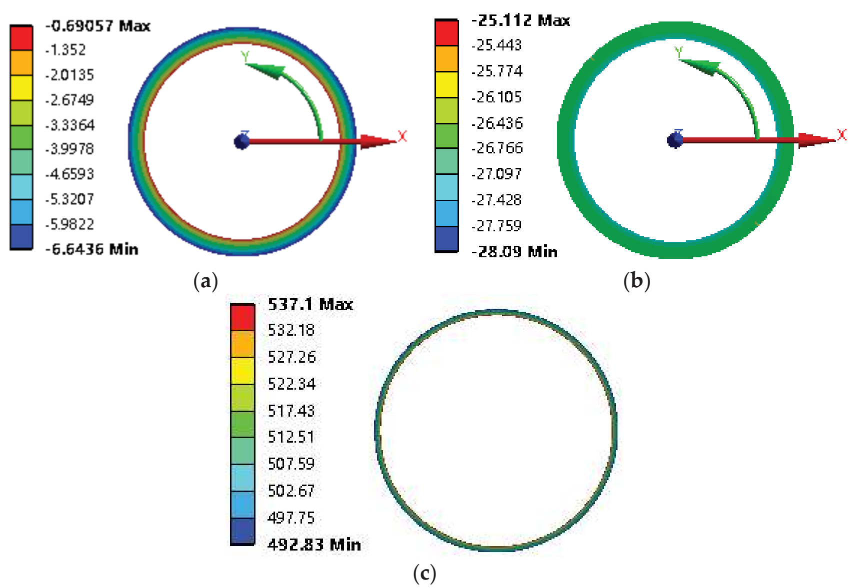


Figure 21. The designed motor's rotor stress distribution. (**a**) PM radial stress; (**b**) PM tangential stress; (**c**) sleeve equivalent stress.

5. Conclusions

This paper provides the design and analysis of a 30 kW, 30,000 r/min high-speed permanent magnet motor (HSPMM) for compressor application. Firstly, the effects of the number of poles, the number of slots, and the PM magnetization methods on the electromagnetic performance of the motor are studied. It is found the four poles, and twenty-four slots motor with PM parallel magnetized has better electromagnetic performance. Then, the CFD is adopted to provide thermal analysis for HSPMMs with different cooling channel structures and different channel turn numbers, and it is found the eight turns spiral water channel with a flow rate of 3 m/s is more beneficial for heat dissipation. The analytic method of the sleeve–PM–core–shaft structure rotor stress distribution is proposed in this paper. This paper provides a comparative analysis between the analytic method and the FEM of the rotor stress, proving the analytic method is reliable. Finally, the influencing factors of rotor stress distribution are analyzed in this paper. According to the analysis results, this paper adopts a stainless steel sleeve of 1 mm to mechanically protect the rotor of the HSPMM, and the interference between the sleeve and the PM is set to 0.05 mm. The rotor stress FEM calculation results of the designed motor show the fact that the rotor protection scheme adopted in this paper is reliable.

Author Contributions: Conceptualization, Z.Q. and Y.Z.; methodology, Z.Q. and Y.Z.; software, Y.Z., S.Y. and Z.X.; validation, Y.Z., S.Y. and Z.X.; formal analysis, Y.Z.; investigation, Z.Q. and Y.Z.; resources, Y.Z.; data curation, Z.Q.; writing—original draft preparation, Z.Q.; writing—review and editing, Z.Q. and Y.Z.; visualization, Z.Q. and Y.Z.; supervision, Y.Z.; project administration, Y.Z.; funding acquisition, Y.Z. All authors have read and agreed to the published version of the manuscript.

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Insulation Degradation Analysis Due to Thermo-Mechanical Stress in Deep-Sea Oil-Filled Motors †

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Abstract: With the wide application of motors in deep sea exploration, deep-sea motors require a higher power density and a longer lifetime. Motor lifetime mainly depends on the thermo-mechanical stress (TMS) load on its stator insulation. Unlike normal motors, deep-sea motors are usually filled with oil to compensate for the high pressure generated by seawater, which leads to high additional viscous drag loss. This, combined with the high pressure, will greatly change the TMS distribution and further influence motor insulation lifetime. Thus, the insulation degradation analysis of deep-sea oil-filled (DSOF) motors due to TMS has become important. This paper presents a TMS analytical model of DSOF motor insulation, considering the joint effects of high pressure and motor temperature. The CFD method is adopted to perform motor thermal analysis, considering temperature effects on viscous drag loss. The FEA method is adopted for thermo-mechanical analysis and to verify the analytical model accuracy. Rainflow counting and the Miner fatigue method are adopted to evaluate motor lifetime. Results show that compared with motors working in normal environments, TMS on DSOF motor insulation can be reduced by up to 59.5% due to high pressure and the insulation lifetime can be increased by up to 16.02%. Therefore, this research can provide references for high power density DSOF motor design.

Keywords: viscous loss; oil-filled motor; thermo-mechanical stress; CFD analysis; insulation degradation

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1. Introduction

Stator insulation systems (SISs) play a key role in affecting motor service life. The differences in thermal expansion coefficients among insulation components, including ground insulation, copper wires, coatings and epoxy fillings in stators, will induce thermo-mechanical stress (TMS) in the SIS. TMS is considered to be a critical factor causing insulation degradation [1]. Deep-sea motors usually adopt oil-filled methods to balance the seawater pressure, which brings two challenges to TMS assessment compared to motors running under normal conditions. One is that the oil filled in the air gap will lead to high additional temperature-dependent viscous drag loss and changing motor heat dissipation conditions, which will greatly influence the motor thermal distribution. The other is that the motor SIS is subjected to high pressure caused by seawater, which combined with the thermal distribution will change the TMS distribution greatly. In order to analyze the insulation degradation of a DSOF motor due to the TMS, these two challenges have to be considered.

In recent years, many researchers have addressed the thermal analysis issue of motors working on land, but few studies have focused on DSOF motors. The motor loss and heat dissipation conditions in DSOF motors are quite different from normal motors. The viscous loss induced by oil may account for more than 50% of the total loss [2]. Additionally, the thermal conductivities and heat transfer coefficients are tightly related to the fluid and temperature fields of DSOF motors. Therefore, numerical analyses such as the finite element method (FEM) and computational fluid dynamics (CFD) method should be adopted for the motor thermal analysis instead of analytical calculation. The advantage of the CFD method in calculating viscous drag loss compared with analytic methods has been demonstrated on a motor test rig [3]. The necessity of coupled field analysis by taking the viscous loss component into consideration in DSOF motor design has been illustrated in [4]. Reference [5] constructs a three-dimensional fluid-thermal coupled model of an oil-filled motor to obtain the heat transfer coefficients, fluid flow characteristics and temperature distribution. However, most of the published literature ignores the influence of temperature on the gap oil viscosity and the viscous loss. This paper focuses on this issue by using the viscosity user-defined function (UDF) to improve the accuracy of the DSOF motor temperature field calculation, based on which TMS analysis and lifetime assessment of the DSOF motor are performed.

TMS has been extensively studied in various industrial fields. The mismatch of materials' thermal expansion coefficients can induce TMS in electronic assemblies not only by temperature cycling during normal operation, but also due to the high temperatures experienced during fabrication, shipping and storage [1]. References [6,7] present a degradation model investigating the thermo-mechanical fatigue in insulated gate bipolar transistor modules and apply it to degradation estimation as well as to the accurate lifetime assessment of power modules. References [8,9] analyze the failure of a ceramic ball grid array package used for an FPGA chip caused by TMS and both analytical modeling and numerical simulations are performed to analyze the TMS, which is validated to be the main factor causing FPGA chip failure. In the paper [10], fatigue life evaluation for multiple epoxy laminate composites considering the stress versus cycle life has been investigated. Reference [11] focuses on the thermo-mechanical fatigue of silicon and molybdenum (SiMo) alloyed ductile iron by performing FEA and testing the stresses of the specimen. In IEEE Standard-56, TMS is regarded as one of the aging factors of motor insulation [12]. As for electric motors, researchers have derived the TMS analytical equation of a single wire in epoxy-impregnated traction motor stator windings and established 3D FEA models to demonstrate it [13]. Reference [14] analyzes cooling methods to reduce the mechanical stress, which will cause fatigue and the degradation of motor insulation, induced by the thermal loading on motor windings impregnated with epoxy, and builds an experimental set-up to validate the numerical analysis results. Some researchers have performed thermo-structural simulations on a segmented stator winding geometry numerically to quantify the TMS induced in the motor windings due to high temperature [5]. However, the work on insulation degradation analysis caused by TMS in electric motors is far from sufficient, especially for motors working in the deep sea. For DSOF motors, the SIS bears a high pressure varying with the operating depth, which can reach more than 100 MPa for the working depth of 10,000 m. The high pressure will change the TMS distribution and further affect the motor lifetime. This paper seeks to address the insulation TMS distribution issue considering a high-pressure environment based on the accurate thermal calculation results of DSOF motors.

This paper will take the brushless DC (BLDC) motor as the target due to its outstanding performance and wide application in deep-sea equipment. The main contributions of this paper are as follows:

1. Considering the deep-sea high-pressure effects and structural characteristics of the DSOF motor SIS, an analytical model of TMS for stator insulation in DSOF motors is proposed. The effects of slot dimension and copper fill factor on the TMS are investigated, and the analytical model shows that TMS in DSOF motor stator insulation is closely related to the external seawater pressure, the motor temperature rises and the motor structure dimensions.

2. To obtain the DSOF motor temperature rises for the further TMS analysis, a 3D CFD model of a DSOF motor is established considering the sensitivity of oil viscosity to temperature. Two UDFs are compiled for the temperature field calculation. In addition, the influence of viscosity on the heat transfer coefficient of the motor is analyzed, and the importance of the viscous loss and viscosity UDFs in the temperature field in respect of the accurate calculation of the DSOF motor is demonstrated.

3. Based on the calculation results of the motor temperature distribution, the analytical TMS model is used to obtain the stator insulation TMS variation trend with the seawater pressure. It is found that the seawater pressure can decrease the TMS to 59.5% at most. Then, based on the DSOF motor model, numerical analysis is carried out to obtain the coupled TMS through thermal-mechanical simulation, which further demonstrates the accuracy of the proposed analytical model.

4. The TMS spectra of dangerous points on the stator insulation are synthesized. By counting the rainflow of the TMS spectra, combining the S-N curve of stator insulation materials and using Miner fatigue theory, the lifetime of the DSOF motor is predicted. The research can afford a new perspective for high power density DSOF motor design. Figure 1 shows the flow chart of fatigue lifetime evaluation of DSOF motors.

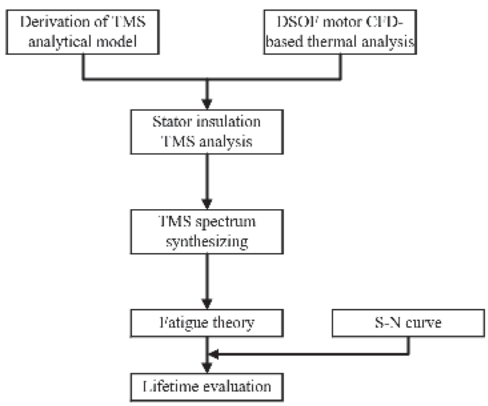


Figure 1. Flow chart of DSOF motor fatigue lifetime prediction.

This paper is organized as follows. Section 2 derives the analytical model of TMS in DSOF motor SISs by considering deep-sea high pressure, motor temperature rises and structure dimensions. Section 3 focuses on the temperature field calculation of DSOF motors based on fluid-thermal coupling analysis. In Section 4, the temperature results obtained by CFD analysis are applied to further calculate the TMS. Thermal-structure simulations are conducted to verify the accuracy of the proposed TMS analytical model. In addition, the motor insulation degradation and lifetime prediction are investigated based on the TMS analysis results. Conclusions are given in Section 5.

2. Analytical Model

Figure 1 shows the flow chart of fatigue lifetime evaluation of DSOF motors. The derivation of the TMS analytical model is performed first. For a DSOF motor, the SIS is adhered to stator laminations and compressed copper wires. The polymer coating and copper have similar thermal expansion coefficients, which are quite different from those of the stator core and epoxy impregnation. Thus, high TMS will be introduced into the stator insulation when the motor temperature rises. Moreover, the high density of the copper wires and irregular shape of the stator slots will make derivation of the TMS analytical model very complex. In order to facilitate the task while ensuring the model accuracy, the derivation is implemented by a combination of preliminary derivation and post-modification. That is, we disregard the effect of stator slots and windings in the preliminary derivation and investigate them in the post-modification.

2.1. Preliminary Derivation of TMS Analytical Model

The stator model ignoring the stator slots and windings is shown in Figure 2, where the outer layer is the stator core layer and the inner layer is the impregnation insulation layer. The preliminary aim is to obtain the TMS analytical model of the inner layer. In a deep-sea high-pressure environment, the inner and outer surfaces of the model are subjected to the same pressure due to the oil-filled method.

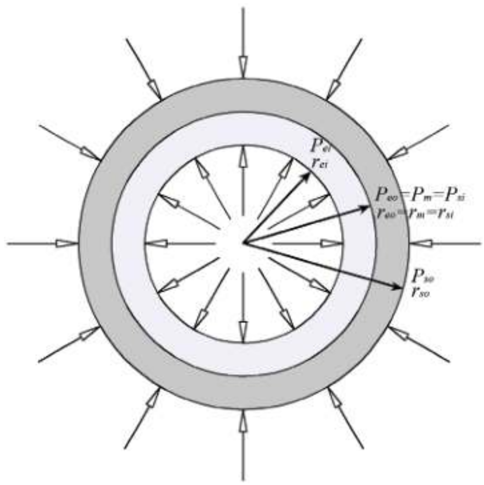


Figure 2. Cross-section of simplified DSOF motor stator model subjected to both internal and external pressures.

Since the geometry of motor stator, the operating temperature distribution and the pressure constraints are all symmetrical about the central axis, the derivation of the TMS analytical model can be performed based on a cylindrical coordinate system. According to the strain–stress theory, strain-displacement correlations and Lamé equations for spatially axisymmetric hollow cylinders [15–17], the stress components of the inner layer can be obtained provided the pressure in the common boundary is known. The two layers have equal radial displacements at the common boundary. Accordingly, the outer surface pressure of the insulation layer can be determined. As shown in Figure 2, P_{si} and P_{so} are the internal and external pressures of the stator core layer, respectively. r_{si} and r_{so} are the stator core layer inner and outer radii. As for the insulation layer, P_{ei} and P_{eo} are its internal and external pressures, respectively. r_{ei} and r_{eo} are its inner and outer radii, respectively. The pressure at the common surface is P_m, which is equal to P_{eo} and P_{si}. The radius of the common boundary is r_m, which is equal to r_{eo} and r_{si}. By means of Lamé equations, the stresses at the outer surface of the insulation layer and the stresses at the inner surface of the stator core layer are obtained, which can be further substituted into the strain–stress equation. Then, the radial displacements of the insulation layer and stator core layer at the common boundary, u_{eo} and u_{si}, can be obtained by means of strain-displacement correlation as follows:

$$u_{eo} = \frac{r_{eo}}{E_e}\left[\frac{2P_{ei}-P_{eo}\left(R_e^2+1\right)}{R_e^2-1} - \mu_e\left(\frac{P_{ei}-R_e^2P_{eo}}{R_e^2-1} - P_{eo}\right)\right] + r_{eo}\alpha_e T \quad (1)$$

$$u_{si} = \frac{r_{si}}{E_s}\left[\frac{P_{si}\left(R_s^2+1\right)-2R_s^2P_{so}}{R_s^2-1} - \mu_s\left(\frac{P_{si}-R_s^2P_{so}}{R_s^2-1} - P_{si}\right)\right] + r_{si}\alpha_s T \quad (2)$$

where

$$R_e = \frac{r_{eo}}{r_{ei}}$$

$$R_s = \frac{r_{so}}{r_{si}}$$

R_e denotes the ratio of the insulation layer outer radius to its inner radius. R_s denotes the ratio of the stator layer outer radius to its inner radius. E_e is Young's modulus, μ_e is Poisson's ratio and α_e is the coefficient of thermal expansion of the insulation layer. E_s, μ_s and α_s denote the same properties of the stator core layer. T denotes model temperature rise. Considering a seawater pressure P equal to P_{ei} and P_{so}, and with Equation (1) equal to Equation (2), the pressure P_m can be expressed as

$$P_m = \frac{V_{io}P - V_n T}{V_m} \tag{3}$$

where

$$V_{io} = \frac{2-\mu_e}{R_e^2 - 1} + \frac{R_s^2(2-\mu_s)}{R_s^2 - 1} \cdot \frac{E_e}{E_s}$$

$$V_m = \frac{(1-2\mu_e)R_e^2 + \mu_e + 1}{R_e^2 - 1} + \frac{(\mu_s + 1)R_s^2 - 2\mu_s + 1}{R_s^2 - 1} \cdot \frac{E_e}{E_s}$$

$$V_n = E_e(\alpha_s - \alpha_e)$$

V_{io}, V_m and V_n are the coefficients related to the motor dimensions and material properties. Substituting Equation (3) and insulation layer parameters into the Lamé equations then using the von Mises theory, the average TMS on the stator insulation layer, σ_a, can be expressed as

$$\sigma_a = \varsigma(r) \cdot \sqrt{\frac{[(V_m - V_{io})P + V_n T]^2}{V_m^2}} \tag{4}$$

where

$$\varsigma(r) = 2.25\frac{r_{eo}^2}{(r_{eo} - r_{ei})(R_e^2 - 1)} \int_{r_{ei}}^{r_{eo}} \frac{1}{r^2}\mathrm{d}r$$

$\varsigma(r)$ is a function of the stator insulation layer radius. The preliminary TMS analytical model notes that there is a minimum value of average TMS on the SIS as the deep-sea pressure increases. The TMS will first decrease to a minimum value, and then increase with the deep-sea pressure rising.

2.2. Post-Modification

Subsequently, the influence of the stator slot dimension and copper fill factor on the TMS of the motor SIS are explored using Ansys software to refine the analytical model.

2.2.1. The Effect of Slot Dimension

For investigating the influence of slot dimension on the TMS, three stator models A_1, A_2 and A_3 with stator teeth widths of 9, 13.4 and 16 mm are analyzed, respectively. The results are presented in Figure 3. The black line is the preliminary analytical model solution, while the others are from the simulation results. The minimum TMS points of the three model simulation results show no difference from the analytical solution, but the slopes are inconsistent and the difference will become larger when the stator slots become narrower. Consequently, the slot shape factor (SSF), s, denoting the ratio of the stator slot width to the stator yoke height is introduced to weigh the slot shape. Interestingly, the SSF only affects the changing rate of the TMS, not the minimum value. Therefore, the slope-modified coefficient Q_s is introduced to refine the preliminary derivation model.

To further quantify the influence of SSF, another six FEA models A_4 to A_9 with different SSFs are constructed and simulated. The correlation of the modified coefficient Q_s with s is shown in Figure 4. Remarkably, the slope-modified coefficient Q_s will increase to a stable value of 0.9123 as the slot shape factor grows.

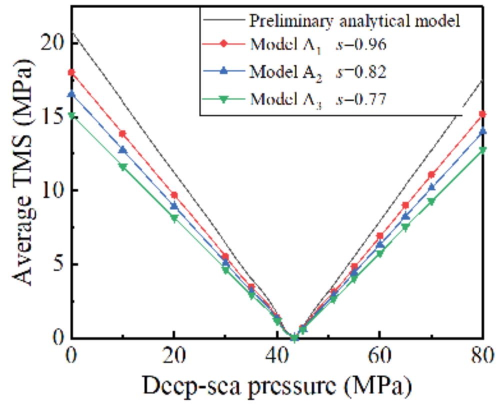


Figure 3. Average TMS of different slot shape factor variations with deep-sea pressure.

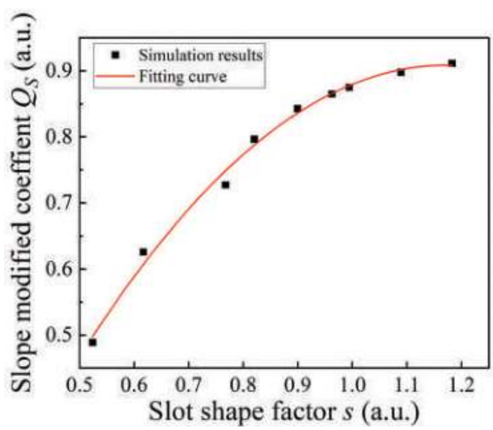


Figure 4. Slope-modified coefficient versus slot shape factor.

2.2.2. The Effect of Copper Fill Factor

For investigating the effects of the influence of the copper wire fill factor on the average TMS of the SIS, FEA simulations are performed on three stator FEA models with same SSF of 0.96 and different copper fill factors (CFFs), f, of 0, 0.3 and 0.4 are simulated. The results are presented in Figure 5. The black line still represents the analytical solution of the preliminary model, while others are from the simulation results. The three model simulation results of deep-sea pressure corresponding to the minimum TMS agree with the analytical solution, while the curve slope shows some inconsistencies. When the CFF becomes larger, the TMS changing rate will decrease, and the decrease rate will become smaller. Meanwhile, the lowest point moves upwards, and the slope closer to the minimum is gentler. Therefore, another slope-modified coefficient Q_f is introduced to improve the preliminary derivation model.

To further explore the influence of CFF, five FEA models with different CFFs are investigated. The variation in the slope-modified coefficient Q_f with f is shown in Figure 6. Notably, the slope-modified coefficient Q_f will decrease to a stable value of 0.7913 as the CFF grows.

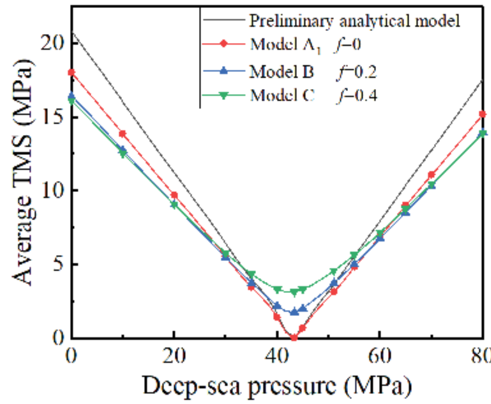


Figure 5. Average TMS of different copper fill factor variations with deep-sea pressure.

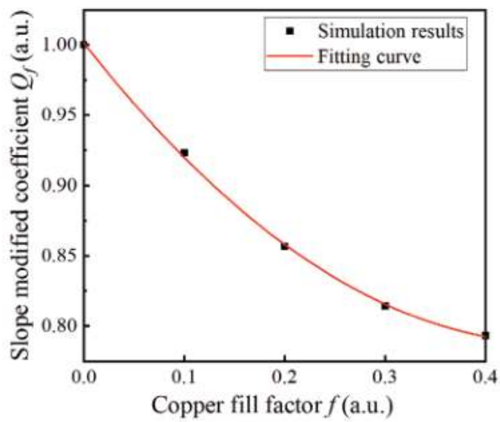


Figure 6. Slope-modified coefficient versus copper fill factor.

Eventually, the modified TMS analytical model can be expressed as

$$\sigma_a = Q_s Q_f \varsigma(r) \cdot \sqrt{\frac{[(V_m - V_{io})P + V_n T]^2}{V_m^2}} \tag{5}$$

In Equation (5), the average TMS for the DSOF motor SIS is not only related to the material properties and dimensions of the motor, but also to the temperature change and the ambient pressure. When the motor material, dimensions and temperature rise are fixed, the TMS on the insulation will first decline to a minimum value and then increase as the DSOF motor work depth increases. By means of the analytical model, the external pressure corresponding to the minimum TMS based on the DSOF motor temperature rise can be predicted.

3. Temperature Distribution of DSOF Motors Based on Fluid-Thermal Coupling Analysis

DSOF motor thermal analysis needs to focus on heat sources and heat dissipation. For heat sources, except copper loss, iron loss and eddy current loss, additional viscous loss generated by oil filled in the air gap has to be considered. For heat dissipation, direct contact with oil and seawater makes the heat removal of DSOF motors quite different from normal ones. Thus, the key to the thermal analysis of DSOF motors is the accurate calculation of viscous loss and heat transfer coefficients. Viscous loss is closely related to rotation speed and oil viscosity. For an oil-filled motor, the inner and outer radii of the air gap are r_1 and r_2, respectively, and according to the boundary conditions of the gap oil and fluid momentum equation, the mechanical loss consumed by the force of oil acting on the rotor surface can be determined. This mechanical loss, namely viscous loss, is defined as Equation (6) [18,19]:

$$P_{\text{oil}} = \frac{4\pi\mu\omega^2 L {r_2}^2 {r_1}^2}{\delta(r_2 + r_1)} \tag{6}$$

where μ denotes fluid dynamic viscosity ($\text{Pa} \cdot \text{s}$), ω denotes rotor angular velocity, L is rotor axial length and δ is the air gap width. It can be noted that viscous loss has a linear relation with dynamic viscosity for a motor with determined dimensions and rotational speed. Obviously, viscosity has a critical impact on viscous loss.

3.1. Viscous Drag Loss at Different Temperature

Viscosity has a close relation with fluid temperature. The distance between fluid molecules will expand and the fluid viscosity will decrease significantly as the temperature increases [20,21]. Hence, viscous loss calculation needs to take the temperature into account. The experience formula describing the viscosity variations with temperature, namely the Poisson formula, is shown in Equation (7) [22,23]:

$$\mu_t = v\rho = \mu_0 e^{-\lambda(t-t_0)} \tag{7}$$

where μ_t denotes the dynamic viscosity at temperature t, μ_0 denotes the dynamic viscosity at temperature t_0, and λ is the viscosity-temperature coefficient (VTC) of the liquid, reflecting the viscosity decreasing rate. Clearly, the viscosity has an e-exponential correlation with the temperature. Unlike no-oil motors, the empirical value of VTC 0.035 cannot be directly adopted for DSOF motors [24]. It is necessary to introduce CFD numerical calculation methods. A 24-slot and 8-pole DSOF motor (Sanao Electrical, Shanghai, China) with 2.62 KW rated power and 5000 rpm operating speed was adopted as the research subject. Its 3D CFD model was established as presented in Figure 7. Figure 8 shows the meshing details, comprising a total of 3,533,129 meshing cells.

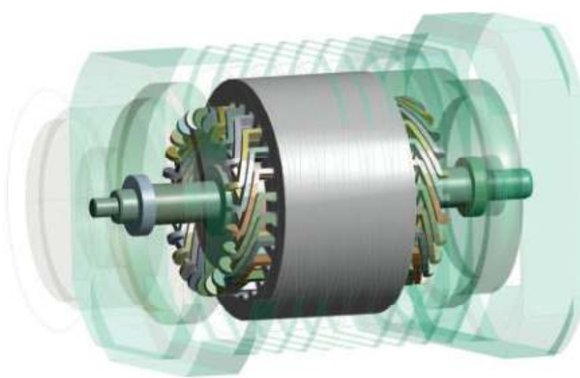


Figure 7. 3D CFD model of the DSOF motor.

According to the CFD fluid field model of the motor, oil viscosities at different temperatures are obtained and the corresponding viscous losses are determined. For subsequent motor temperature analysis, the temperature and corresponding viscous loss are fitted and compiled into a UDF, namely a viscous loss UDF. The relation between the viscous loss and temperatures is presented as Equation (8), where P_{oil} is viscous loss and T is the temperature of the gap oil.

$$P_{\text{oil}} = 3.576 \times 10^{-5} T^4 - 1.064 \times 10^{-2} T^3 + 1.154 \times 10 T^2 - 54.356 T + 1050.62 \tag{8}$$

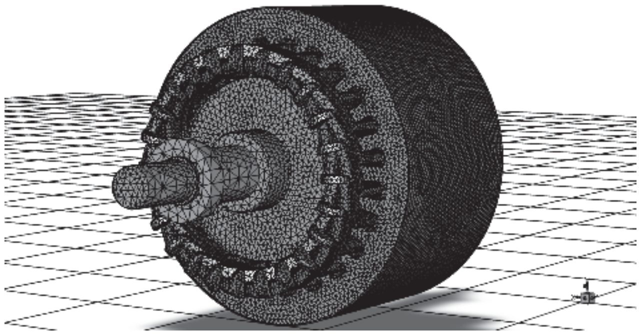


Figure 8. Meshing results of the DSOF motor.

3.2. The Need for a Viscous Loss UDF and Viscosity UDF

Viscosity changing with temperature not only affects the heat source (viscous loss) but also the heat dissipation of the DSOF motor. Therefore, the function of viscosity changing with temperature should be compiled into a viscosity UDF and adopted in the temperature field numerical calculation. By fitting with a polynomial, the relation between the oil viscosity and temperature is expressed as follows:

$$\begin{aligned} \mu = &-1.201 \times 10^{-10}T^5 + 5.387 \times 10^{-8}T^4 - 9.284 \times 10^{-6}T^3 + \\ &7.735 \times 10^{-4}T^2 - 3.162 \times 10^{-2}T + 0.533 \end{aligned} \tag{9}$$

where μ is the oil viscosity. Then, the viscosity UDF is compiled and added to the CFD analysis. After fully investigating the heat source and the dissipation condition, a thermal analysis of the DSOF motor is performed. CFD analysis both with and without the UDFs is carried out for investigating the influence of the UDFs. The temperature results of each motor component are compared in Table 1. The initial temperature is set to 293 K.

Table 1. Thermal analysis with and without viscous loss and viscosity UDFs.

Component	Temperature (K)					
	Minimum		Maximum		Average	
	No UDFs	With UDFs	No UDFs	With UDFs	No UDFs	With UDFs
Windings	331.7	325.1	332.7	326.1	332.2	325.6
Rotor	332.1	325.2	333.8	326.4	333.0	325.8
Magnets	333.9	326.5	335.2	327.6	334.7	327.2
Shaft	327.2	321.2	331.9	325.0	330.8	324.1
Stator	326.0	320.3	331.2	324.6	328.3	322.2

It is noted that motor component temperatures with viscous UDF and viscosity UDF are lower than those without the UDFs. Temperature rises of the rotor, permanent magnets and stator of the DSOF motor have decreased by 17.96%, 18.06% and 17.17%, respectively, indicating that viscosity indeed affects the heat transfer coefficient and heat source of the DSOF motor, and further demonstrating the necessity of viscous and viscosity UDFs in the accurate thermal calculation of DSOF motors.

3.3. Temperature Distribution of DSOF Motor

The temperature distribution of the whole DSOF motor is shown in Figure 9. The motor maximum temperature is 327 K and minimum temperature is 321 K. It is noted that the temperature rises of all motor components are almost the same.

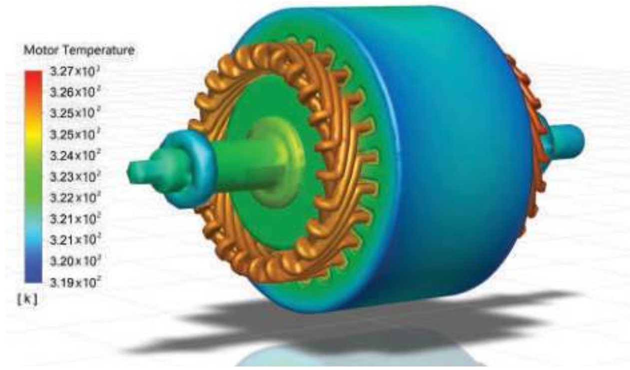


Figure 9. DSOF motor temperature distribution.

Figure 10 shows cross-section temperature contour details of the DSOF motor. Obviously, for the stator components, the maximum temperature difference is only 4 K. Therefore, the temperature distribution of the SIS can be regarded as uniform and spatially axisymmetric. The thermal analysis results will subsequently be used for the following TMS analysis of the DSOF motor.

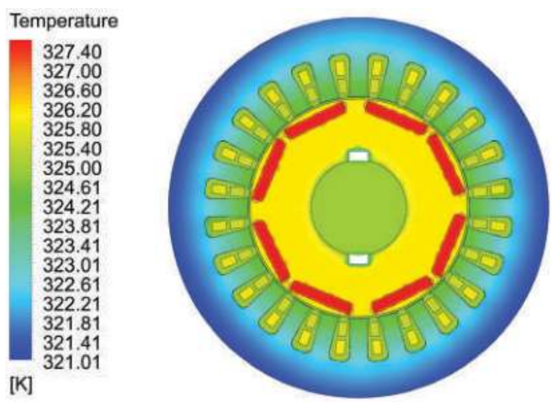


Figure 10. DSOF motor cross-section temperature field contour.

4. Thermo-Mechanical Simulation and Lifetime Analysis of the DSOF Motor

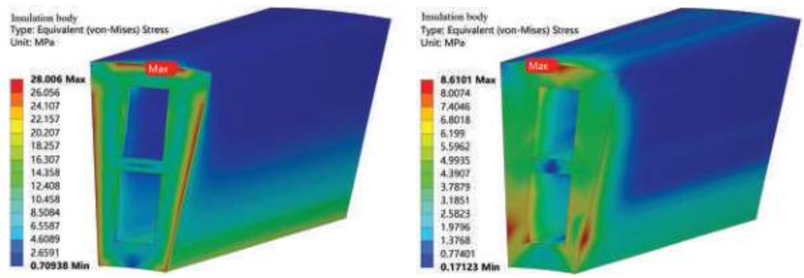
4.1. Thermo-Mechanical Simulation

Before analyzing the TMS of the DSOF motor by means of the analytical model proposed in Section 2, thermal-mechanical simulations are conducted to verify the model accuracy. The relevant material properties are listed in Table 2.

Table 2. Relevant material properties of the motor stator.

Properties	Electrical Sheet	Polymer Coating	Epoxy Impregnation
Young's modulus (MPa)	2×10^5	7.4×10^3	3.5×10^3
Poisson's ratio	0.29	0.42	0.34
Coefficient of thermal expansion (1/K)	1.2×10^{-5}	1.6×10^{-5}	5×10^{-5}
Thermal conductivity (W/m·K)	28	0.2	0.21
Heat capacity (J/kg·K)	440	1090	400
Density (kg/m^3)	7650	1530	1180

According to previous thermal processing results and the proposed TMS analytical model, the optimal working environment pressure for the DSOF motor is 20 MPa, where the TMS on the stator insulation system is the smallest. Therefore, thermal-mechanical simulations of the motor under normal pressure and 20 MPa are compared for further investigation. Figure 11 presents the simulation results.



(a) (b)

Figure 11. TMS in the DSOF motor stator insulation under seawater pressure. (**a**) Normal pressure. (**b**) Deep-sea pressure 20 MPa.

It is noted that the TMS at the stator insulation component is decreased when the seawater pressure changes from normal pressure to 20 MPa. For the stator insulation body, the most critical regions, where the maximum TMS is located, are on the top boundaries of the free ends. One is 8.61 MPa under the external pressure of 20 MPa, decreased by 69.25% compared to the other under normal pressure. Additionally, the average TMS in the SIS of the DSOF motor is further researched, and its variations with seawater pressure and temperature rise are illustrated in Figure 12. Obviously, the average TMS can be affected by both temperature rise and environment pressure, and the varying trends are consistent with the analytical model.

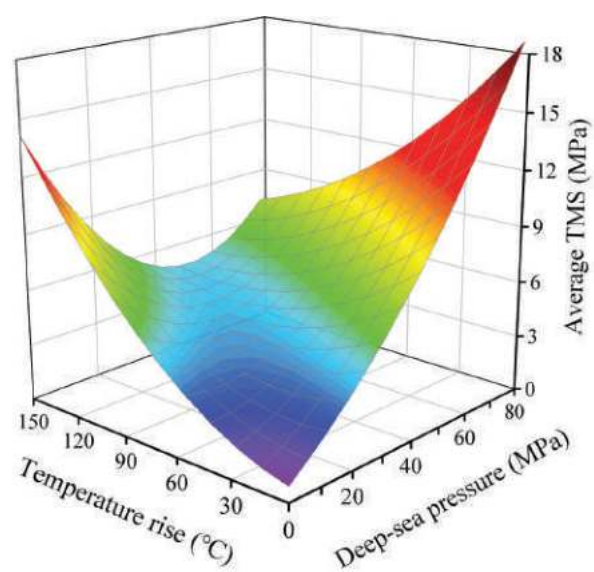


Figure 12. Average TMS in the SIS of the DSOF motor variation with seawater pressure and temperature rise.

Analytical solutions and simulation results of the external pressure variation corresponding to the minimum TMS with DSOF motor temperature rise are illustrated in Figure 13, where the analytical results agree well with the numerical ones and the maximum error rate is 4%, proving the accuracy of the TMS analytical model in this study.

For the targeted DSOF motor, the TMS in the motor SIS influenced by the deep-sea water pressure is then obtained as shown in Figure 14.

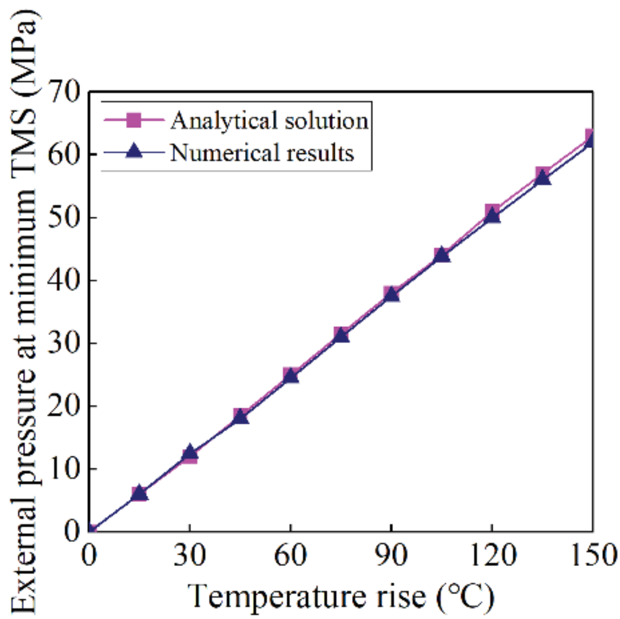


Figure 13. Comparison of analytical solution and numerical results of external pressure at the minimum TMS.

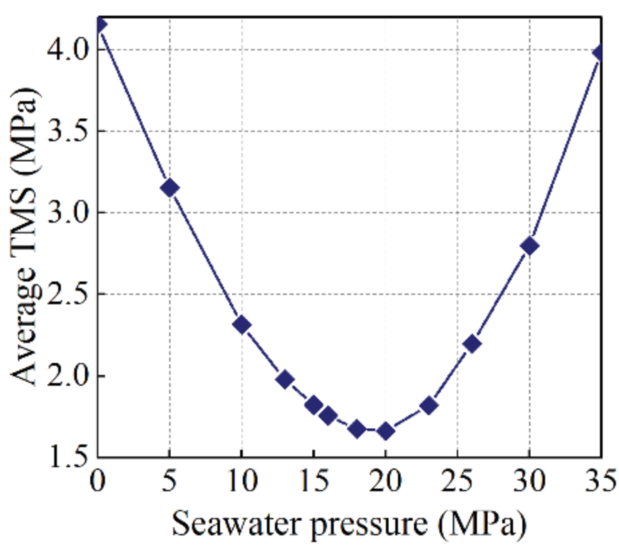


Figure 14. TMS in the SIS of the DSOF motor variation with seawater pressure based on the thermal results.

As working depth becomes larger, the average TMS in the DSOF motor insulation system will first decrease from 4.2 MPa to 1.7 MPa and then increase, and the stress can be reduced by up to 59.5%. These results indicate that the seawater pressure can ease stator insulation degradation and extend the DSOF motor service life. For further investigating the influence of the deep-sea working environment on the fatigue process of the DSOF motor, the insulation degradation analysis is then carried out.

4.2. Insulation Degradation Analysis

For replicating a typical DSOF motor operating cycle, an actual-use heating rate value of 0.02 °C/s and temperature change range of 5 °C relative to steady-state temperature are adopted [25]. Thus, for transient thermal loads of the targeted motor, the copper heating rate and thermal cycle are set to 0.02 °C/s and 350 s. Figure 15 shows the motor stator cycling temperature distribution results. Accordingly, the TMS spectra of the most dangerous point, where the maximum TMS is located, of the stator insulation are analyzed under normal pressure and 20 MPa, as shown in Figure 16. The stress spectra can be used for motor stator insulation fatigue lifetime prediction.

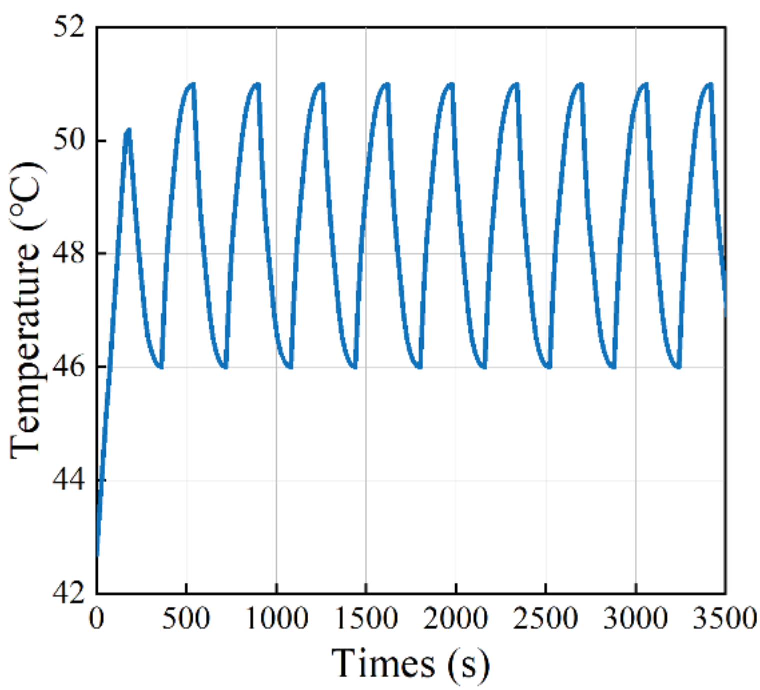


Figure 15. Motor stator temperature.

The mean amplitude and the number of TMS cycles are calculated by a rainflow cycle counter. The results are shown in Figure 17, where (a) represents the TMS results under normal pressure and (b) illustrates the TMS results under 20 MPa pressure. Obviously, there are 10 typical cycles in each case.

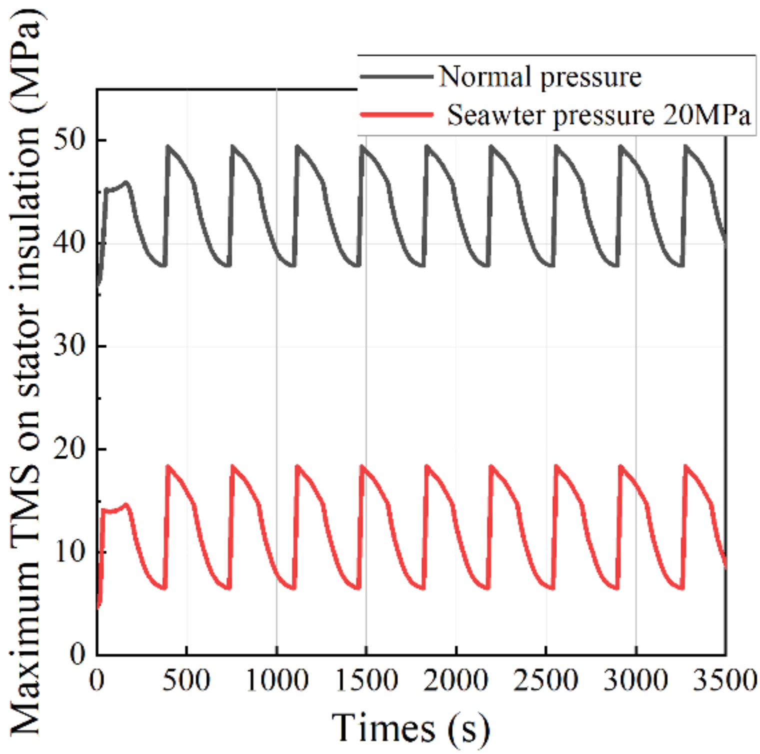


Figure 16. TMS spectra of stator insulation under normal pressure and 20 MPa.

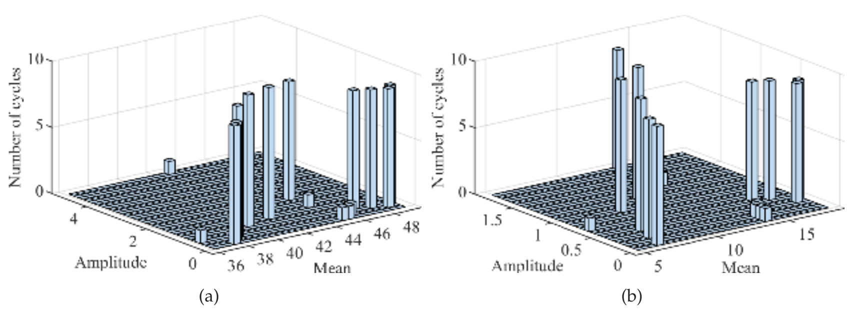


Figure 17. Results from rainflow cycle counter of the maximum TMS of stator insulation. (**a**) Normal pressure. (**b**) Deep-sea pressure 20 MPa.

By applying the Goodman equation to the rainflow counting results of the TMS spectra, the equivalent zero-mean stress amplitudes are obtained [26]. Then, combining the S-N curve of the insulation material and Miner fatigue theory, the cumulative damage of the dangerous point in the stator insulation is obtained as 6.08×10^{-5} under normal pressure and 5.24×10^{-5} under 20 MPa. It is assumed that when fatigue damage reaches 1, fatigue failure occurs. Thus, the number of cycles to the insulation fatigue failure is 16,436 under normal pressure and 19,069 under 20 MPa. As presented in Figure 16, the cycle time is 0.97 h. By multiplying the number of cycles to failure, the stator insulation lifetimes are 15,943 h and 18,497 h under normal pressure and 20 MPa, respectively. It is noted that compared with motors working in normal conditions, the lifetime of a motor running in a deep-sea environment will increase by 16.02% at most, which indicates that deep-sea high-pressure can slow down the degradation of stator insulation under TMS.

5. Conclusions

This paper has made an effort to analyze the insulation degradation for deep-sea oil-filled (DSOF) motors due to TMS. This was done by first proposing a TMS analytical model of DSOF motor stator insulation, considering the two key factors; namely, deep-sea pressure and motor temperature rise. Then, CFD analysis was performed by investigating the thermal effects on oil viscosity and viscous drag loss to obtain the motor temperature rise. Based on the temperature rise, thermo-mechanical simulations on a DSOF motor were performed to quantify the TMS and verify the proposed TMS analytical model. According to the TMS analysis results, the DSOF motor lifetime was finally evaluated and compared for two different working pressure conditions. The thermo-mechanical simulation results were compared with the analytical solutions obtaining a 4% deviation rate, which demonstrates the accuracy of the proposed TMS analytical model. Results show that the TMS in the DSOF motor insulation system will first decrease then increase as the DSOF motor working depth becomes larger. Compared to the normal pressure working conditions, the TMS on the motor stator insulation can be reduced by 59.5% at most under seawater pressure, and DSOF motor insulation lifetime can be increased by 16.02% at most, which indicates that the seawater pressure can ease stator insulation degradation and extend the DSOF motor service life. This research can provide references for high power density DSOF motor design. In the future, the relevant accelerated aging experiments for a DSOF motor will be carried out to further verify the proposed model. The high power density DSOF motor design will be conducted based on this research.

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Article

Comprehensive Comparison of a High-Speed Permanent Magnet Synchronous Motor Considering Rotor Length–Diameter Ratio

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Abstract: For high-speed permanent magnet machines (HSPMMs), many design schemes of rotor length–diameter ratios can satisfy the constraints of multiple physical fields during the motor design period. The rotor length–diameter ratio greatly impacts the comprehensive performances of multiple physical fields. However, these analyses are missing in the existing literature. Therefore, this paper focuses on the influence of the rotor length–diameter ratio on comprehensive performances. Firstly, finite element models (FEM) of multiple physical fields are built by ANSYS Workbench platform and Motor-CAD software. Then, the comprehensive performances of multiple physical fields are comparatively analyzed. Finally, the designed HSPMM is implemented, based on one prototype of 60 kW, 30,000 rpm to verify the results of comparative analysis. Based on the comparative analysis above, the influent laws of rotor length–diameter ratios on comprehensive performances of multiple physical fields are discussed and summarized, which can be used as a reference for the rotor structural design of HSPMMs.

Keywords: high-speed permanent magnet machines; comprehensive characteristics; comparative; rotor length–diameter ratio

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1. Introduction

In recent years, with the development of electronic devices, high-speed permanent magnet machines (HSPMMs) access industry and life more and more widely, because of their excellent electromagnetic performance and simple control characteristics [1–3]. However, when the HSPMMs rotate at high speed, various problems occur, which hinder the development of HSPMMs, and solving these problems has become a research focus and hot spot for researchers [4–6]. In the design of HSPMMs, the key structural parameters and performances must meet the constraints of multiple physical fields, which include electromagnetic, mechanical, and thermal physical fields [7,8]. However, many rotor shapes can satisfy the constraints of multiple physical fields, from short and thick to long and thin with rotor shape, which leads to the uncertain selection of the rotor shape in the design period. Therefore, the influence of the rotor length–diameter ratio on the comprehensive performances is worthy of discussion and research. In this paper, the influence of rotor length–diameter ratio on comprehensive performances is studied and discussed, in order to obtain better comprehensive performances for HSPMMs.

In the existing literatures, the comprehensive performances of HSPMMs are comparatively investigated through key performance parameters. For electromagnetic and loss properties, some key electromagnetic parameters are comparatively analyzed, such as line back-EMF at no load, radial air-gap flux density distribution, stator iron-core loss, permanent magnet loss, air-friction loss, etc. [9,10]. For rotor stress characteristics, the rotor stresses are calculated, including radial and tangential stresses of permanent magnets, and von-Mises stress distributions of sleeve [11]. For rotor dynamics, the rotor mode and critical speed are simulated [12]. For the thermal behavior, the cooling system and temperature distribution are investigated according to the designed parameters and calculated loss characteristics [13]. It can be seen that the design of HSPMMs must meet the needs of comprehensive performances of multiple physical fields. For HSPMMs, many rotor shapes can satisfy all the needs of comprehensive performances, but different rotor length–diameter ratios have great influence on the comprehensive performances of multiple physical fields. A shorter and thicker rotor shape can effectively reduce the motor temperature rises and improve the rotor dynamics' reliability, but the permanent magnet stress and the sleeve stress are increased due to the poor centrifugal force of the larger rotor diameter [14,15]. A longer and slender rotor shape can effectively improve the rotor stress, while the rotor dynamics and thermal characteristics may not be satisfied due to the longer rotor core length or the smaller rotor diameter [16]. It is clear from the above description that the rotor length–diameter ratio can affect motor performances, including electromagnetic, mechanical, and thermal characteristics. Hence, the influence of rotor length–diameter ratio on comprehensive performances cannot be ignored. In the existing literatures, the influence of rotor length–diameter ratio on motor performance mainly focuses on the analysis of single physical fields and the influence of the rotor length–diameter ratio on the comprehensive performance of multiple physical fields is not clear. To obtain an accurate relationship between the rotor length–diameter ratio and comprehensive performances, comparative analysis of multiple physical fields (such as electromagnetic, mechanical, and thermal) is necessary. The rotor length–diameter ratio should be investigated based on a comparative analysis of comprehensive performances, which include electromagnetic and loss characteristics, mechanical strength, rotor dynamics, and thermal performance.

Based on the above analysis, the rotor length–diameter ratio has a great influence on the comprehensive performances of multiple physical fields. However, the effect has rarely been focused on in the existing literature, which leads to the uncertain selection of the rotor length–diameter ratio in the design period of high-speed permanent magnet motors. Therefore, in this paper, the influence of the rotor length–diameter ratio on comprehensive performances is discussed and comparatively investigated. Firstly, finite element models of multiple physical fields are built. Then the influent laws of rotor length–diameter ratio on comprehensive performances are summarized. Finally, a prototype is manufactured and the comprehensive performances are experimentally measured. The conclusions obtained can be used as a reference for the rotor structural design of HSPMMs. The initial electromagnetic design scheme is introduced in Section 2. In Section 3, the process of comparative analysis is determined and three cases with representative rotor length–diameter ratios are selected. In Section 4, the comprehensive performances of these three cases are comparatively analyzed in detail. In Section 5, the designed HSPMM are manufactured and experiments are conducted. Finally, the conclusions summarized are drawn in Section 6, and some key simulation results and the influence of the rotor length–diameter ratio on comprehensive performances are summarized, which can be used as a reference for rotor structural design of HSPMMs.

2. HSPMM Structure and Main Performance

In the design of a high-speed permanent magnet motor, electromagnetic performance, losses, temperature distribution, rotor stress, and rotor dynamics should be considered comprehensively. It is necessary to analyze the comprehensive characteristics of the motor, and then the designed motor scheme can meet all needs. In the design process, it is important to determine the mechanical structure of the motor firstly, because the different mechanical structures will directly affect other performance of the designed motor [17].

In this paper, a high-speed permanent magnet machine for 60 kW at 30,000 rpm is designed and the main parameters of the motor can be obtained, as shown in Table 1. A surface-mounted permanent magnet rotor is designed. To ensure the reliability of rotor permanent magnets (PMs) rotating at high speed, a carbon fiber sleeve of 5 mm is used to improve the rotor stress, and the tensile strength of carbon fiber materials is 1400 MPa. The stator core is composed of 0.35 mm cold-rolled silicon of low-loss steel.

Table 1. Main parameters of the designed motor for HSPMM.

Parameters	Values
Output power (kW)	60
Rated rotating speed (rpm)	30,000
Rated load voltage (V)	380
Rated power factor	0.98
Stator slots	24
Carbon fiber sleeve thickness (mm)	5

3. Initial Comprehensive Characteristics Comparative Analysis for HSPMM

3.1. Comprehensive Comparative Analysis Process for HSPMM

To analyze the influence of main structural parameters on the comprehensive characteristics, a process is presented, as shown in Figure 1.

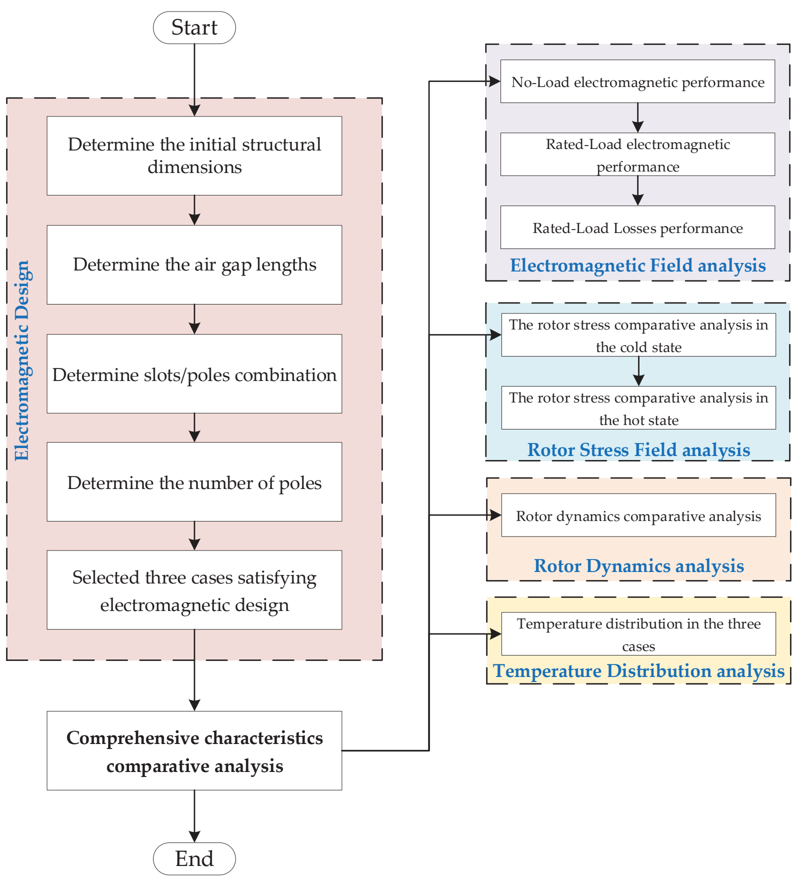


Figure 1. Comprehensive characteristics comparative analysis and design process.

Step 1: According to the basic performance parameters and rotor diameter constraints, the initial structural dimensions are determined. Step 2: The electromagnetic characteristics of different air-gap lengths are compared and analyzed, and the air-gap length is determined. Step 3: The influence of slots/poles combination on the electromagnetic characteristics is analyzed. Step 4: The number of poles is determined through the influence of the variable number of poles on electromagnetic performance. Step 5: Three cases satisfying electromagnetic design are selected, mainly considering the difference in rotor shapes. Step 6: The electromagnetic and loss characteristics are comparatively analyzed under three cases. Step 7: The rotor stresses are comparatively analyzed under three cases. Step 8: The rotor dynamics are comparatively analyzed under three cases. Step 9: The temperature distribution is comparatively analyzed under three cases.

3.2. Multiple Physic Fields Constraints

The rotor dimensions of HSPMMs are limited by the electromagnetic, mechanical, and temperature characteristics. For the electromagnetic performance, the rotor dimensions must satisfy the electromagnetic load. For the mechanism, the rotor length–diameter ratio must be within a specific range, so that the rotor will not be damaged by centrifugal force and resonance. For temperature rises, the heat dissipation design of the motor should take into account the selected cooling system and working mode, so that the temperature distribution of the motor is within the constraints. Three aspects restrict each other and determine the rotor dimensions of HSPMMs. In this paper, the following basic constraints are determined according to the actual requirements.

1. Geometric constraints: interference fit $\delta_s = 0.15$ mm, the thickness of the sleeve $h_{sleeve} = 5$ mm, the thickness of the magnets 7 mm $\leq h_{PM} \leq$ 8.5 mm, and the stator outer diameter $D_{is} = 157$ mm.
2. Electromagnetic constraints: the amplitude of the line to line Back-EMF at no-load is between 500 V and 540 V, and the air-gap magnetic flux density is between 0.4 T and 0.6 T. The output power in the rated load is $P_{out} \geq 60$ kW, the thermal load required $AJ \leq 200$ A^2/mm^3(determined by experience).
3. Strength constraints: the tensile strength of permanent magnet should be less than 80 MPa, and the tensile strength of carbon fiber sleeve should be less than 1400 MPa. Safety margins should be considered in the engineering, so the tensile strength limitation of the permanent magnet at rated speed is 64 MPa, and the tensile strength limitation of the sleeve at rated speed is 1100 MPa.
4. Critical speed constraints: the rated speed of the rotor is required to be less than 0.7 times the first-order critical speed.
5. Thermal constraints: the limited maximum working temperature of the HSPMM is 130 °C, and stator housing water cooling system is adopted.

3.3. Influence Analysis of Main Parameters on Electromagnetic Performance

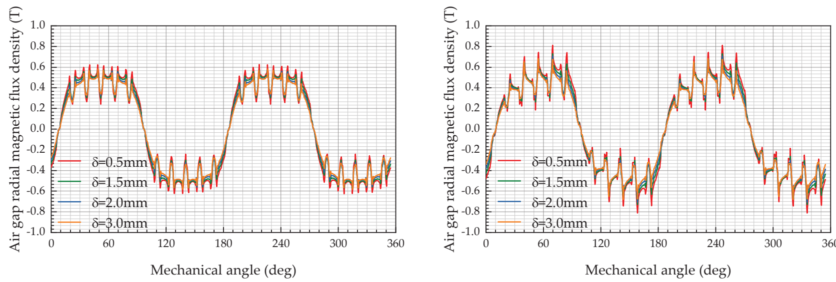
3.3.1. Air-Gap Length

For different air-gap lengths, the electromagnetic performance of the motor is analyzed, such as the air-gap lengths of 0.5, 1.5, 2.0, and 3.0 mm. In this analysis, rated output power is maintained, and the thickness of the permanent magnet is increased with the increase of air-gap length. The permanent magnet thickness and electromagnetic characteristics are shown in Table 2.

Table 2. Electromagnetic characteristics under different air-gap lengths.

Case	1	2	3	4
Air-gap length (mm)	0.5	1.5	2.0	3.0
Magnet thickness (mm)	6.5	8	8.8	13
No-load line back-EMF (V)	500.93	497.70	490.46	488.13
RMS current (A)	94.35	95.25	95.13	94.21
Cogging torque (%)	2.70	2.20	1.93	1.67
Efficiency (%)	97.09	97.05	97.01	96.94

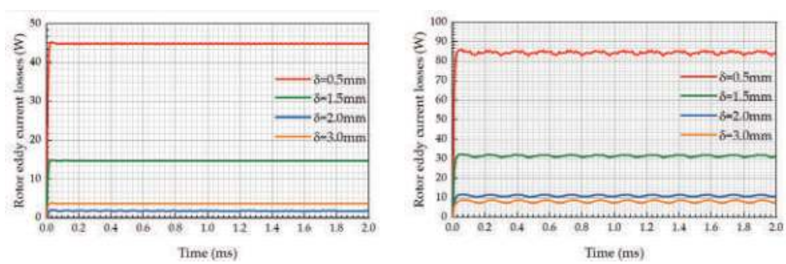
The radial air-gap flux density distribution with the mechanical angle under no load and rated load is shown in Figure 2. The amplitude of the radial air-gap flux density is mainly affected by the air-gap length, and the impact of the air-gap length on harmonic order and the distribution of the radial air-gap flux density of the motor is little.



(**a**) (**b**)

Figure 2. Radial air−gap flux density under no load and rated load: (**a**) no-load; (**b**) rated-load.

The rotor eddy-current losses under no load and rated load with different air-gap lengths are shown in Figure 3. It can be shown that the air-gap length is smaller and the rotor eddy-current loss is larger because a smaller air-gap length will cause more current harmonics to enter the rotor, resulting in more rotor eddy-current losses. As the air-gap length increases, more current harmonics will be shielded, so the rotor eddy-current losses gradually decrease. When using a smaller air-gap length, the dosage of magnets is smaller, while the rotor eddy-current loss is larger. When using a larger air-gap length, the dosage of magnets is larger, and the rotor eddy-current loss is smaller.



(**a**) (**b**)

Figure 3. Rotor eddy-current losses on the magnets under no load and rated load: (**a**) no-load; (**b**) rated load.

Therefore, the air-gap length is a compromise between rotor eddy-current losses and electromagnetic performance. Although the rotor eddy-current loss is smallest under the 3.0 mm air-gap length, electromagnetic performance cannot be guaranteed, especially for the no-load back-EMF. The rotor eddy-current losses of the 0.5 mm and 1.5 mm air-gap lengths are larger, which ultimately lead to a higher motor temperature rise. According to the level of influence of the air-gap length on rotor eddy-current loss in Figure 4, rotor eddy-current loss decreases gradually with the increase of air-gap length. Electromagnetic performance and the rotor eddy-current loss are comprehensively considered and the air-gap length of 2 mm is selected as a compromise solution.

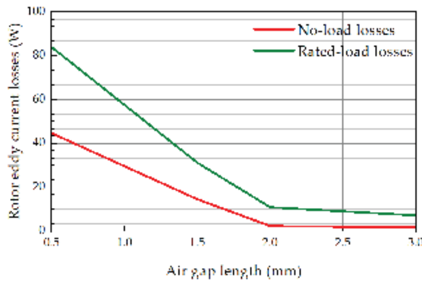


Figure 4. Effect of air-gap length on rotor eddy-current losses under no load and rated load.

3.3.2. Slots/Poles Combination

The slots/poles combination will affect the winding factor and the cogging torque period. Therefore, the impact of slots/poles combination on electromagnetic characteristics is not ignored.

The winding factor, cogging torque period, and load power frequency with different pole numbers are discussed in Table 3. Because of the same minimum common multiple of the poles and slots, the cogging torque period is the same. However, the stator winding factors of two-pole and four-pole motors are 0.949, and 0.933, respectively, indicating the high stator windings utilization ratios of two-pole and four-pole motors. When the power frequency is high, it will greatly increase the stator iron-core loss and rotor eddy-current loss. Therefore, in the HSPMMs, a large power-supply frequency should not be used. In conclusion, this paper adopts 24 slots and two poles or four poles in the designed motor.

Table 3. Electromagnetic performance comparative analysis under different slots/poles combination.

Slots/Poles	Slot per Pole per Phase	Pole Pitch/Pitch	Winding Factor	Cogging Torque Period	Load Frequency
24/2	4	12/11	0.949	24	500 Hz
24/4	2	6/5	0.933	24	1000 Hz
24/6	4/3	4/3	0.885	24	1500 Hz
24/8	1	3/1	0.5	24	2000 Hz

3.3.3. Number of Poles

The number of poles plays a major role in HSPMMs performance. Electromagnetic characteristics of two-pole and four-pole motors under no load and rated load are analyzed. In Figure 5, for the two-pole motor, the magnetic flux density of stator yoke is greater than 1.8 T, which makes the stator yoke saturated under the same stator outer diameter of 157 mm. Therefore, the stator outer diameter of the two-pole motor should be expanded.

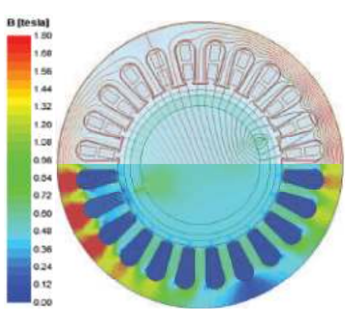


Figure 5. Magnetic flux density and lines distributions of two-pole motor with the stator outer diameter 157 mm.

The stator outer diameter of the two-pole motor must be expanded to the stator yoke non-saturation. The main structural parameters of the two-pole motor and the four-pole motor are compared, as shown in Table 4. The stator outer diameter of the two-pole motor is extended by 15% over the four-pole motor, ensuring basic electromagnetic characteristics.

Table 4. Structural parameters of two-pole and four-pole motors for HSPMM.

Parameters	Two-Pole	Four-Pole
Power (kW)	60	60
Speed (rpm)	30,000	30,000
Frequency (Hz)	500	1000
Slot number	24	24
Stator outer diameter (mm)	181	157
Stator inner diameter (mm)	89	89
Rotor outer diameter (mm)	86	86
Effective core length	110	110

Electromagnetic characteristics of two-pole and four-pole motors are shown in Table 5. Although the four-pole motor has almost double stator iron-core loss density compared to the two-pole motor, the stator iron-core loss of the four-pole motor has almost 3/2 times that of the two-pole motor, as they have different iron volumes.

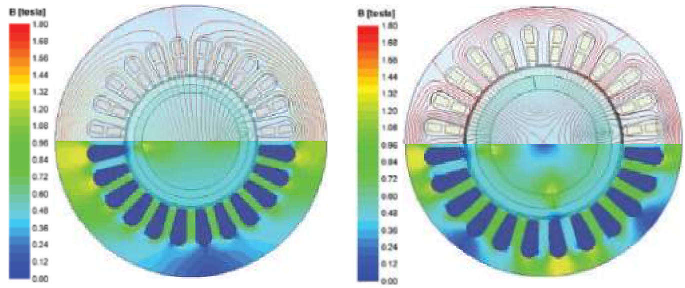
Table 5. Electromagnetic characteristics comparison of two-pole motor and four-pole motor for HSPMM under rated load.

Parameters	Two-Pole	Four-Pole
RMS current (A)	95.38	95.25
Line back-EMF (V)	497.83	497.70
Torque (Nm)	19.10	19.11
Stator-teeth flux density (T)	1.04	1.04
Stator-yoke flux density (T)	1.31	1.31
Thermal load (A^2/mm^3)	168.8	162.07
Power density (kW/kg)	2.66	4.10
Materials of stator	DW310-35	DW310-35
Iron loss density (W/kg)	20.28	41.85
Stator-core loss (W)	457.581	611.792

The two-pole and four-pole motor are compared and analyzed from the stator yoke magnetic flux density distribution, rotor eddy current, winding end length, radial air-gap flux density distribution, and line-to-line back-EMF.

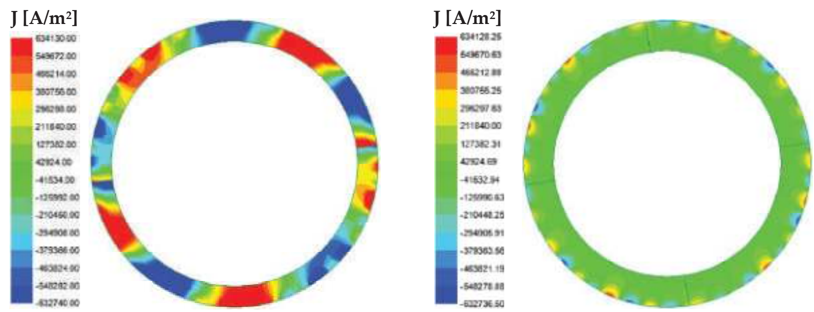
In Figure 6, the maximum values of flux densities are basically the same for the two-pole and the four-pole motors, which are about 1.38 T in the stator yoke. However, the stator outer diameter is 157 mm for the four-pole motor and 181 mm for the two-pole motor, thus the utilization of the stator-yoke material is higher for the four-pole motor. The four-pole motor allows the smaller dimensions and an improved material utilization ratio.

Rotor eddy-current loss in permanent magnets is part of the cause of motor temperature rise. Rotor eddy-current density distribution on permanent magnets of the two-pole and the four-pole motors is shown in Figure 7. Obviously, the rotor eddy current of the two-pole motor is much higher than that of the four-pole motor. According to the 2D-FEA calculation, the rotor eddy-current loss of the four-pole motor is 31.52 W, which is lower than the 235.56 W of the two-pole rotor. The four-pole motor is formed by an arrangement of four magnets, and the two-pole motor consists of two magnets.



(**a**) (**b**)

Figure 6. Magnetic flux density and lines distributions of two-pole and four−pole motors at no-load: (**a**) two-pole motor; (**b**) four-pole motor.



(**a**) (**b**)

Figure 7. The rotor magnet eddy-current density distribution of two-pole and four-pole motors under rated load: (**a**) two-pole motor; (**b**) four-pole motor.

The winding end length of the two-pole motor is longer than that of the four-pole motor, because the stator winding coils of the two-pole motor have a larger pitch than that of the four-pole motor. The stator weights and coils' half-turn lengths of two-pole and four-pole motors are compared in Table 6. According to the comparison in Table 6, the coils' half-turn length of the four-pole motor is much shorter than that of the two-pole motor, and the weight and material consumption are much smaller than that of the two-pole motor. The radial air-gap flux density distribution with the mechanical angle and Fourier transforms results for the two-pole and the four-pole motors are shown in Figures 8 and 9.

Table 6. Structure parameter value comparison of stator winding coils for HSPMM.

Parameters	Two-Pole Motor	Four-Pole Motor
Coils half-turn length (mm)	299.492	197.258
Total winding weight (kg)	22.5634	14.6193
Armature core steel consumption (kg)	24.1621	17.4648

The fundamental amplitude of radial air-gap flux density of thefour4-pole motor is better, and the odd harmonic amplitude is lower than that of the two-pole motor. The amplitudes of the 11th and 13th harmonics of the four-pole motor are larger than that of the two-pole motor, because the two-pole motor adopts a shorter pitch winding, which weakens the 11th and 13th harmonics.

At rated load, the radial air-gap flux density distribution of two-pole and four-pole motor is almost the same at no load. However, at rated load, the fundamental amplitude of radial air-gap flux density of the four-pole motor is much greater than that of the two-pole motor, which can provide a higher air-gap flux density to ensure the motor generates sufficient electromagnetic torque.

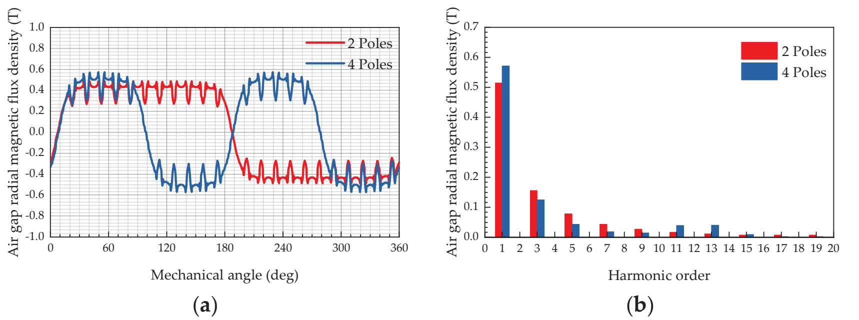


Figure 8. Radial air-gap flux density of two-pole and four−pole motors under no load: (**a**) radial air-gap flux density with mechanical angle; (**b**) Fourier transforms.

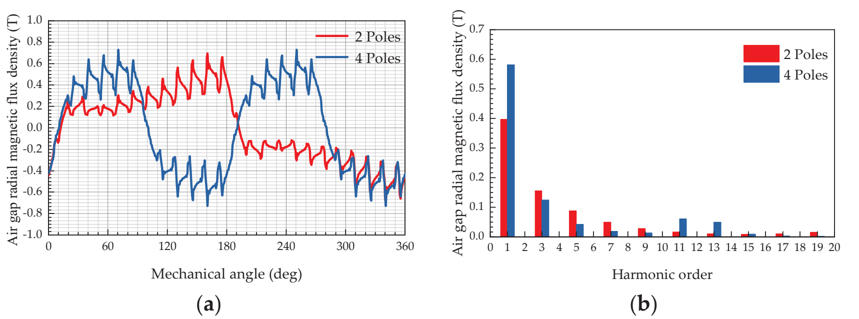


Figure 9. Radial air-gap flux density of two-pole and four-pole motor under rated load: (**a**) radial air-gap flux density with mechanical angle; (**b**) Fourier transforms.

The line-to-line back-EMF and the Fourier transforms results of two-pole and four-pole motors at no-load are shown in Figure 10. It can be found that the line-to-line back-EMF waveform of the two-pole motor is the flat-top wave, because the fifth and seventh harmonics amplitudes of the line-to-line back-EMF of the two-pole motor is higher, and the fundamental amplitude of the two-pole motor is slightly greater than that of the four-pole motor.

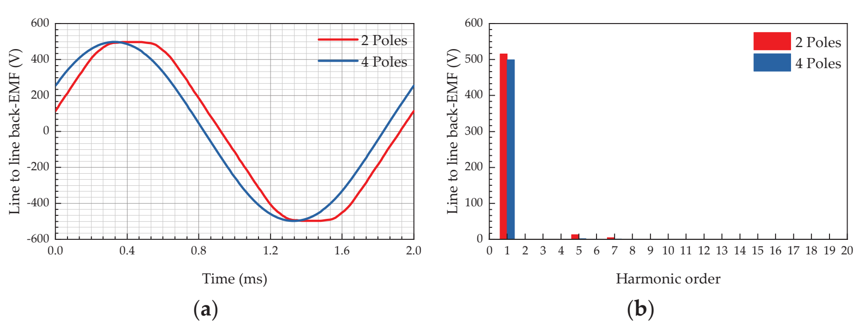


Figure 10. Line to line back−EMF of two-pole and fours-pole motor at no load: (**a**) line to line back-EMF waveform; (**b**) Fourier transform.

Based on the above analysis, for electromagnetic performance between two-pole and four-pole motors, the following results can be obtained.

- For a two-pole motor, due to the stator core's magnetic saturation, the motor dimension, coils length, and the total weight are larger than those for a four-pole motor. It is reasonable to consider a four-pole motor for the strict requirements on motor size applications.
- For the stator-core loss, a four-pole motor is larger than that of a two-pole motor, and the main reason is the lower load frequency for the four-pole motor.
- For the rotor eddy-current density, rotor eddy-current loss of the four-pole motor is 31.52 W, which is much lower than the 235.56 W of the two-pole rotor.
- For the radial air-gap flux density, the fundamental amplitude of a two-pole motor is lower than that of a four-pole motor, and the high-order harmonic amplitudes of a two-pole motor are greater than that of a four-pole motor.
- For the line-to-line back-EMF, the waveform of a four-pole motor is more sinusoidal. The fifth and seventh harmonic amplitudes of the line-to-line back-EMF of a two-pole motor are much larger than those of a four-pole motor.

3.3.4. Permanent Magnetic Materials

At present, permanent magnetic materials of HSPMMs are SmCo permanent magnet and NdFeB permanent magnet. Permanent magnetic materials usually have low mechanical strength and high thermal sensitivity. The permanent magnetic material SmCo has better temperature stability, while NdFeB has the low bending strength, low tensile strength, and compressive strength. Compared to SmCo permanent magnet, the permanent magnetic material NdFeB has poorer temperature stability, but the better magnetic property, better mechanical properties, and the lower price. The performances of the two permanent magnetic materials are shown in Table 7. To ensure the mechanical strength of the rotor, permanent magnet NdFeB is used.

Table 7. Performance of permanent magnet materials for HSPMM.

Parameters	NdFeB	SmCo
Residual flux density	$\leq$1.47 T	0.85 T $\sim$1.15 T
Coercive force	$\leq$1200 kA/m	$\leq$800 kA/m
(BH)max	$\leq$398 kJ/m^3	$\leq$258.6 kJ/m^3
Density (kg/m^3)	7400	8300
Thermal conductivity $(\mathrm{W}/(\mathrm{m}\cdot\mathrm{K}))$	8.9	11
Poisson ratio	0.27	0.24
Working temperature	$\leq$150 °C	$\leq$300 °C
Coefficient of the remanent	$-(0.095\sim0.15)\%/\mathrm{K}$	$-(0.03\sim0.09)\%/\mathrm{K}$
Tensile strength (MPa)	80	35

3.4. Comprehensive Comparative Analysis Schemes for Rotor Shape

For the requirements of electromagnetic performance, some designers design the rotor short and thick, and some designers design the rotor long and thin. The length–diameter ratio of the rotor is defined as $\lambda = L_{ef}/D_r$.

There are few studies in the literature on the influence of the length–diameter ratio on the multiple physical characteristics of the motor. In this paper, under the constraints of the multiple characteristics, three design cases are selected for comprehensive comparative analysis. Case 1 has the rotor shape of short and thick, and the length–diameter ratio is smallest. Case 3 has the rotor shape of long and thin, and the length–diameter ratio is biggest. Case 2 is an intermediate state between the rotor shapes of Case 1 and Case 3. The structural parameters of the rotor shape for the three cases are shown in Table 8.

Table 8. Structural parameters of the rotor shape for HSPMM under three cases.

Parameters	Case 1	Case 2	Case 3
Effective core length (mm)	55	110	200
Air-gap length (mm)	2	2	2
Stator outer diameter (mm)	203.6	165.4	131.1
Stator inner diameter (mm)	114	92.6	73.4
Rotor outer diameter (mm)	110	88.6	69.4
Length–diameter ratio	0.5	1.24	2.88
Sleeve thickness (mm)	5	5	5
Slots/poles combination	24/4	24/4	24/4

4. Comprehensive Comparison for Rotor Shape

4.1. Electromagnetic and Losses Characteristics Analysis

The two-dimensional finite element model (2D-FEM) is established for the HSPMM with three rotor shapes. The magnetic flux density distribution and flux line distribution are calculated for the rated load condition, as shown in Figure 11. In Case 1, the magnetic flux density is the highest, which is about 1.72 T. The magnetic flux density of the motor in Case 2 is the lowest, which is about 1.38 T. For Case 3, the magnetic flux density is about 1.506 T. If the magnetic flux density is too high, the iron core would be saturated, and the iron-core loss and temperature rise increased.

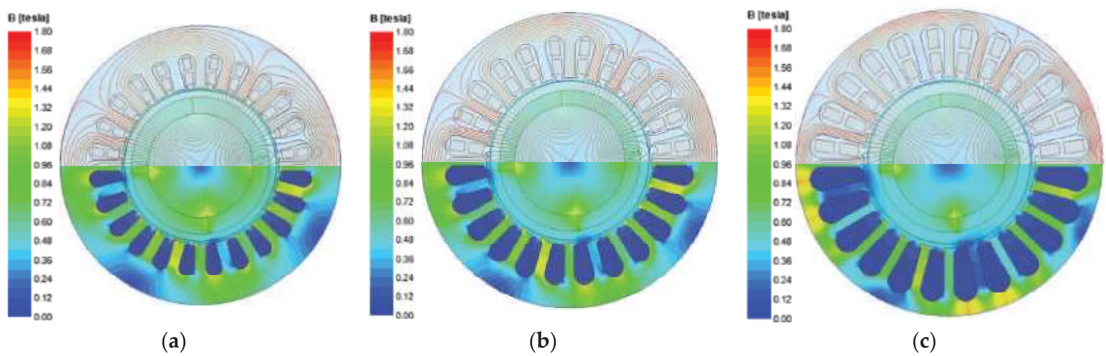


Figure 11. Magnetic flux density and lines distributions under different cases at rated load: (**a**) Case 1; (**b**) Case 2; (**c**) Case 3.

The line-to-line back-EMF and its Fourier transform results in three cases under no-load conditions are shown in Figure 12. In Case 1, the fundamental amplitude is slightly lower, and the 5th and 11th harmonics are slightly higher. Case 2 and Case 3 are the same.

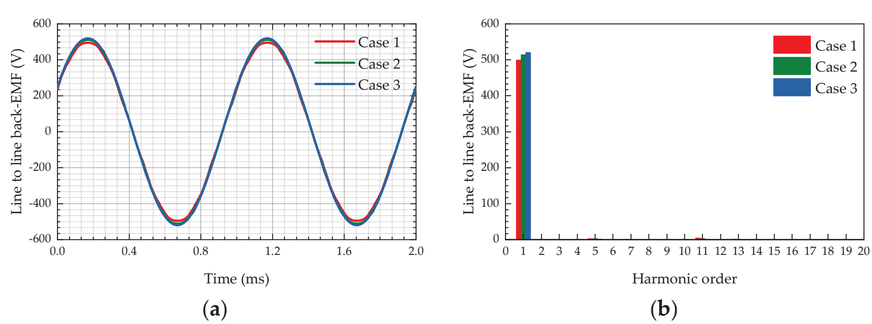


Figure 12. The line to line back-EMF and its Fourier transforms under the three cases at no load: (**a**) line to line back-EMF waveform; (**b**) Fourier transform results.

The radial air-gap flux density with mechanical angle waveform and Fourier transforms results are compared and analyzed for the three cases under the no load and rated load, as shown in Figure 13. The fundamental amplitude of Case 1 is the biggest, while the corresponding high-order harmonics amplitude is also highest. The fundamental amplitude of Case 2 is lower than Case 1, while the amplitude of the high-order harmonics is also smaller than Case 1. In Case 3, the fundamental amplitude is about 0.46 T, which is smallest and is not conducive to the electromagnetic performance of the motor.

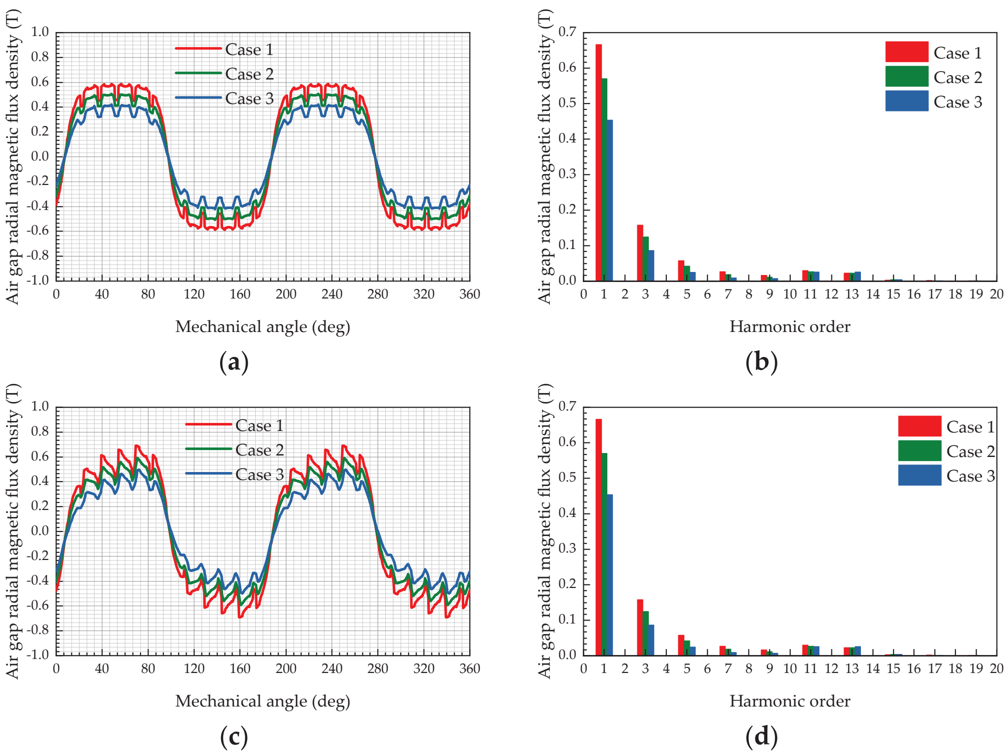


Figure 13. Radial air-gap flux density and Fourier transforms under three cases at no load and rated load: (**a**) radial air-gap flux density at no load; (**b**) Fourier transforms results at no load; (**c**) radial air-gap flux density at rated load; (**d**) Fourier transforms results at rated load.

4.2. Losses Comparative Analysis

4.2.1. Rotor Eddy-Current Loss Comparative Analysis

The rotor eddy-current density in the three rotor shapes at rated load is shown in Figure 14. It can be found that the rotor eddy-current density on the permanent magnets in Case 3 is the smallest, and the rotor eddy-current density on the permanent magnets in Case 1 is the largest. Rotor eddy-current losses of the permanent magnets are calculated, as shown in Figure 15. The rotor eddy-current loss in Case 1 is the highest, which is about 121.4 W, indicating that the rotor temperature rises in Case 1 would be serious. Case 3 has smallest rotor eddy-current loss, which is about 5.7 W. Obviously, as the rotor length–diameter ratio grows, the rotor eddy-current loss decreases.

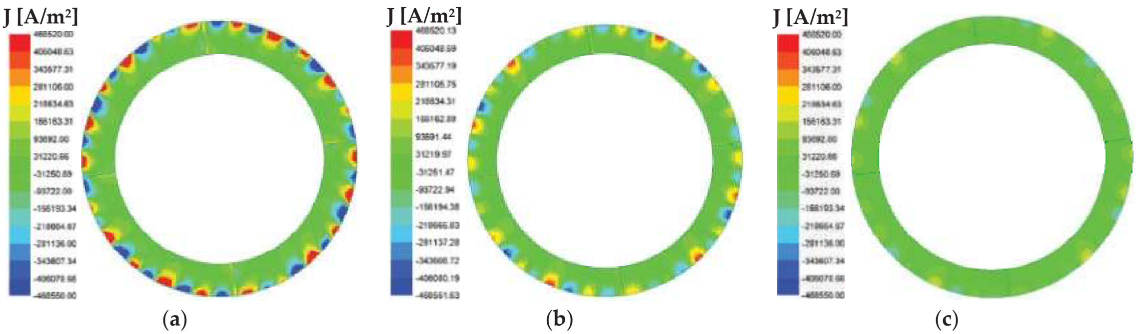


Figure 14. Rotor eddy-current density in three cases at rated load: (**a**) Case 1; (**b**) Case 2; (**c**) Case 3.

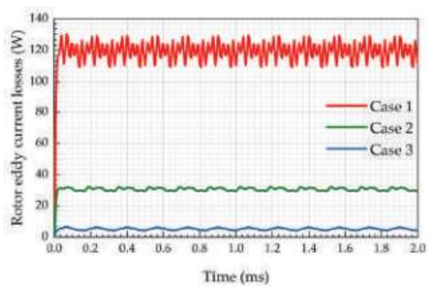


Figure 15. Eddy-current losses in the three cases at rated load.

4.2.2. Stator Iron-Core Loss Comparative Analysis

The stator iron-core losses calculation results under the three cases are shown in Figure 16. According to Berttoti's discrete model of iron-core losses, the iron-core losses are related to the amplitude and frequency of magnetic flux density in the stator core. From the magnetic flux density distribution in Figure 11, the magnetic flux density amplitude of the stator yoke in Case 3 is the highest, so the iron-core loss is the largest.

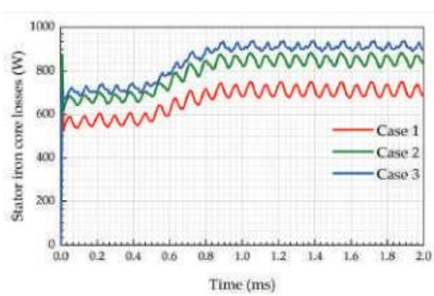


Figure 16. Stator iron-core losses in the three cases at rated load.

4.2.3. Air-Friction Losses Comparative Analysis

The rotor surface linear velocity of HSPMMs is higher, air-friction loss is larger than that of the low-speed motor. Air-friction loss of HSPMMs has a large influence on the design of the rotor shape structure and motor temperature rise. Air-friction loss calculation values of the three cases are shown in Figure 17. It can be found that the bigger the rotor diameter, the larger the air-friction loss at rated speed.

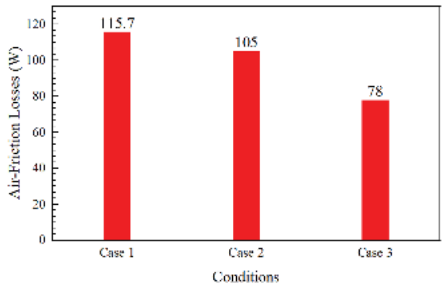


Figure 17. Air-friction losses in the three cases at rated speed.

4.3. Rotor Stress Characteristics Analysis

4.3.1. Rotor Stress Comparative Analysis at a Cold State

In this paper, 3D-FEM are established for the stress of permanent magnets and sleeve under three cases. The radial and tangential stress of permanent magnets and the Von-Mises stress of the sleeve at a cold state are obtained. Figure 18 shows the Von-Mises stress distribution diagram of the sleeve. It can be found that the Von-Mises stress of the sleeve decreases with the decrease in the rotor diameter at a cold state, and the results are consistent with the theoretical analysis.

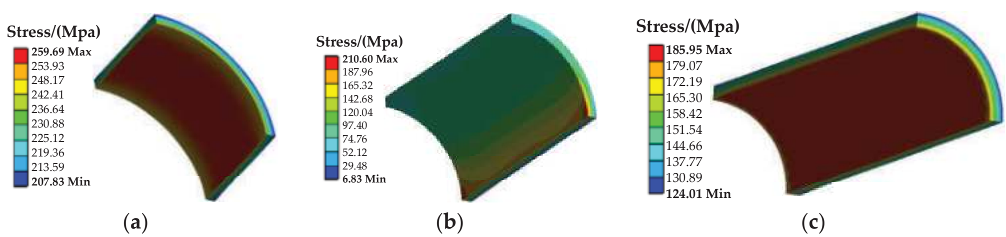


Figure 18. Sleeve Von-Mises stress distribution under the three cases at rated speed at a cold state: (**a**) Case 1; (**b**) Case 2; (**c**) Case 3.

Figure 19 indicates the tangential stress distribution of permanent magnets. It can be found that the tangential stress of permanent magnets in Case 1 is the highest at 40.246 MPa in a cold state. The tangential stress of permanent magnets in Case 3 is the lowest at 8.74 MPa in a cold state. The tangential stress of permanent magnets is mainly related to the interference fit. When the interference fit increases, the assembly of permanent magnets becomes closer, and its tangential stress decreases correspondingly. However, excessive interference fit would increase the radial pressure of permanent magnets and increase the technology assembly and manufacturing difficulty.

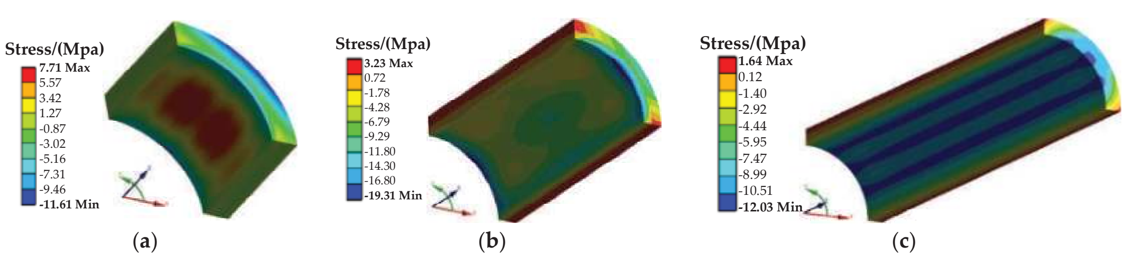


Figure 19. Permanent magnets tangential stress distribution under the three cases at rated speed at a cold state: (**a**) Case 1; (**b**) Case 2; (**c**) Case 3.

The radial stress diagram of the permanent magnets under the three cases is shown in Figure 20. The radial stress of permanent magnets indicates the centrifugal force of the permanent magnets when the rotor is rotating at the rated speed. It can be found that the stress at the periphery of the permanent magnets is minimal, due to the radial pressure exerted by the sleeve on the permanent magnets and the centrifugal force of the rotor canceling each other out.

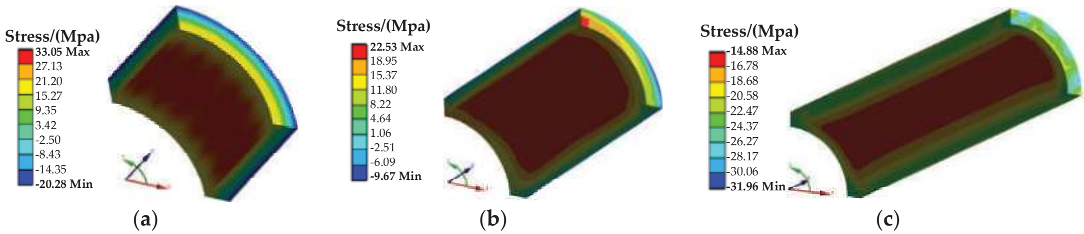


Figure 20. Permanent magnets radial stress distribution under the three cases at rated speed at a cold state: (**a**) Case 1; (**b**) Case 2; (**c**) Case 3.

4.3.2. Rotor Stress Comparative Analysis at a Hot State

For the three cases, the Von-Mises stresses of sleeve, tangential stress, and radial stress of permanent magnets are comparatively analyzed at a hot state. The Von-Mises stress distribution of the sleeve is shown in Figure 21, the tangential stress distribution of permanent magnets is shown in Figure 22, and the radial stress distribution of permanent magnets is shown in Figure 23. At the hot state, it can be found that the radial and tangential stress of magnets increase with the temperature rise, while the carbon fiber sleeve Von-Mises stress did not change, indicating that the thermal stability of the permanent magnets is poorer, and the performance of the permanent magnets is strongly influenced by temperature rise. According to the stress calculation results at a hot state, the tangential stress of permanent magnets under Case 1 is 87.5 MPa, which has exceeded the thermal constraints. Therefore, the rotor shape of Case 1 cannot meet the constraints of rotor stress.

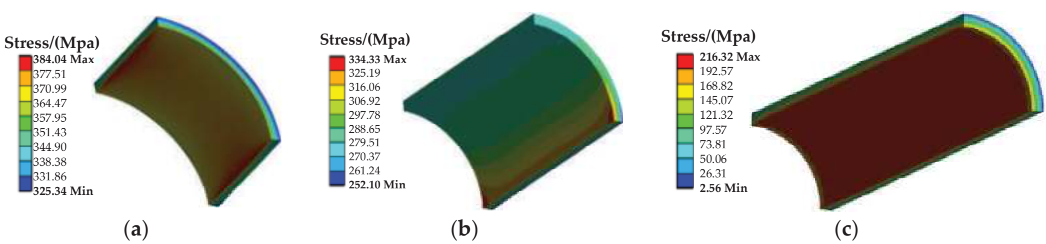


Figure 21. Sleeve Von−Mises stress distribution under the three cases at rated speed in a hot state: (**a**) Case 1; (**b**) Case 2; (**c**) Case 3.

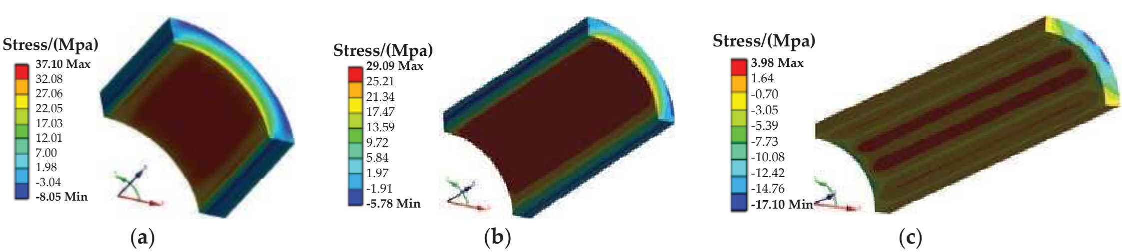


Figure 22. Permanent magnets' tangential stress distribution under the three cases at rated speed at a hot state: (**a**) Case 1; (**b**) Case 2; (**c**) Case 3.

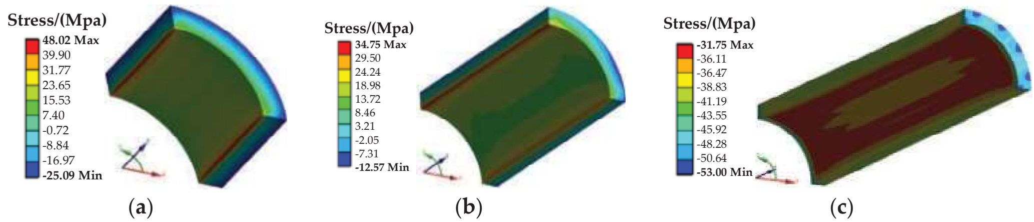


Figure 23. Permanent magnets' radial stress distribution under the three cases at rated speed at a hot state: (**a**) Case 1; (**b**) Case 2; (**c**) Case 3.

From the above rotor stresses analysis under the cold and hot state, it can be found that as the rotor length-diameter ratio increases, the rotor outer diameter gradually decreases, thus the rotor stress gradually decreases, indicating that the mechanical reliability of the rotor gradually increases.

4.4. Rotor Dynamics Characteristics Analysis

When the high rotor speed of HSPMMs reaches its critical speed, the rotor would resonate and damage the rotor. To avoid resonance, the rated speed of the rotor should be less than 0.7 times the first-order critical speed.

The first-order critical speed of the rotor is related to the diameter and length of the rotor. For case 1, $D_r/L_{ef}^2 = 2.5 \times 10^{-2}$, for case 2, $D_r/L_{ef}^2 = 5.12$, and for case 3, $D_r/L_{ef}^2 = 1.2 \times 10^{-3}$.

The supporting stiffness of the rotor is set as 3.5×10^4 N/mm. Modes of oscillation are obtained by the 3D-FEM calculation. Figure 24 shows the first order and second order modes of the rotor for Case 1, Case 2, and Case 3. It is found that the first critical speed of the rotor is 43,800 rpm for Case 1, 168,220 rpm for Case 2, and 44,663 rpm for Case 3, which indicates that the rotor dynamics meet the stability requirements for the three cases, and smaller and larger length–diameter ratios may cause the rotor to resonate more easily.

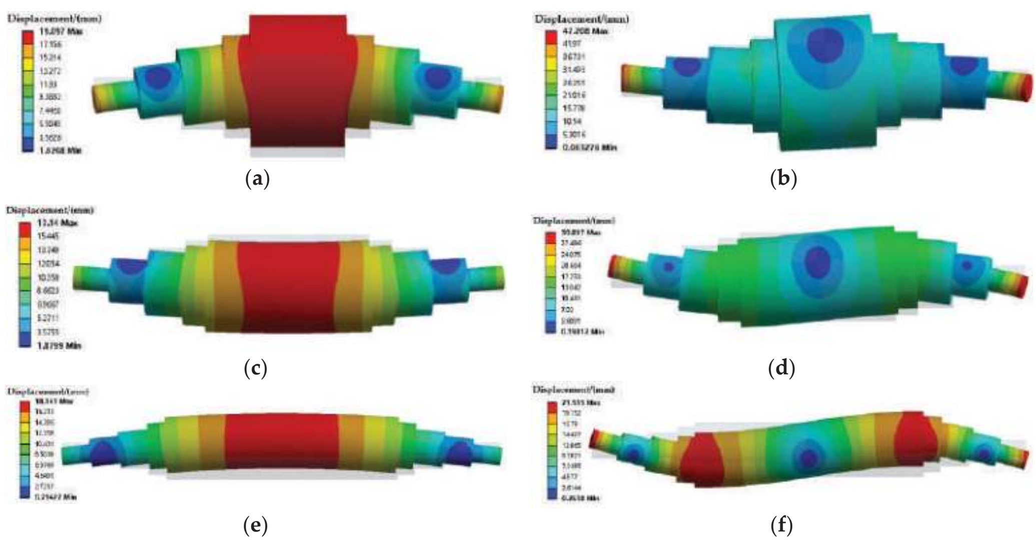


Figure 24. Rotor dynamic characteristics under the three cases: (**a**) first-order mode for Case 1; (**b**) second-order mode for Case 1; (**c**) first-order mode for Case 2; (**d**) second-order mode for Case 2; (**e**) first-order mode for Case 3; (**f**) second-order mode for Case 3.

4.5. Thermal Characteristics Analysis for Three Cases

In this paper, the housing water cooling system is used for heat dissipation design. The cooling system must ensure the safe operation and lifetime of the designed motor. The maximum tolerable temperature of the permanent magnets is limited to less than 130 °C to increase the reliability. Furthermore, the maximum winding temperature is set to 130 °C to increase the service life of the HSPMM.

To meet the temperature constraints, a housing water cooling system is designed for the motor. The water duct cools the motor through the spiral waterway on the stator housing. The distribution of spiral ducts is shown in Figure 25. In this design, the cooling water temperature and the ambient temperature are both 30 °C, and the water flow rate is set to 2 m^3/h.

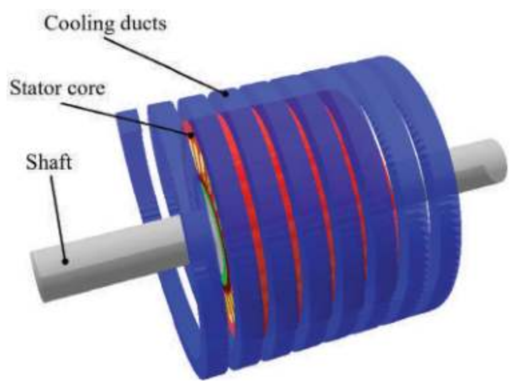


Figure 25. Spiral water ducts for designed motor stator housing cooling system.

The temperature distribution and comparison of the stator windings, stator cores, and permanent magnets are shown in Figures 26 and 27. For the three cases, the maximum temperatures of the winding and rotor are lower than the limiting temperature values. However, the temperature of permanent magnets of Case 1 is slightly lower than that in the other two cases. When the motor temperature requirements are extremely strict and the cooling system is also limited, the smaller rotor length–diameter ratio design can reduce the permanent magnet temperature slightly. In Figure 27, it can be found that as the rotor length–diameter ratio increases, the temperature of stator reduces gradually, and the temperature of rotor increases.

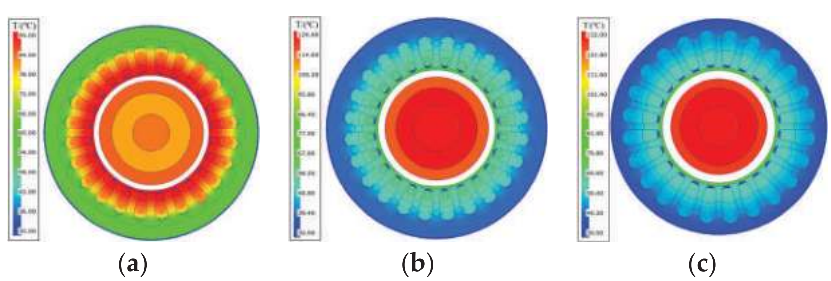


Figure 26. The temperature distribution under the three cases at rated speed: (**a**) Case 1; (**b**) Case 2; (**c**) Case 3.

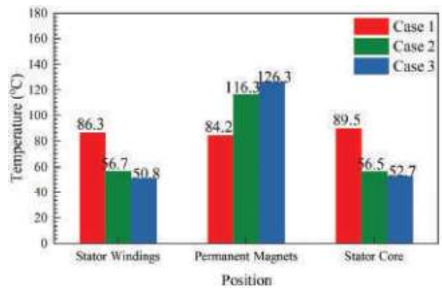


Figure 27. The temperatures on the different positions under the three cases at rated speed.

5. Prototype and Experimental Tests

Based on the above analysis, Case 2 can satisfy the comprehensive performance constraints best, so Case 2 is selected as the final design parameters of the 60 kW at 30,000 rpm HSPMM. Then, a prototype is manufactured, as shown in Figure 28. Electromagnetic and thermal tests are also measured, including the power, back-EMF, and winding temperature, as shown in Table 9.

Figure 28. Prototype.

Table 9. Structure parameters of the rotor shape for HSPMM under the three cases.

Parameters	Measurement	Calculation
Power (kW)	60	60
Back-EMF (V)	520	523
Stator winding temperature (°C)	60.9	56.7

The line-to-line back-EMF of no-load operation is measured with the prototype driven by another motor. The measured peak-to-peak value of line-to-line back-EMF at 30,000 rpm is 520 V, which is nearly the same as the designed voltage. The stator windings temperature at rated load is measured by the Pt100 resistance temperature detectors installed in the stator slots. It can be found that the temperature of the measured windings is 60.9 °C, which is close to the FEM calculation results of 56.7 °C. Obviously, for the designed HSPMM, the measured value of electromagnetic performance and temperature characteristics are close to the calculated results, to satisfy the electromagnetic and thermal constraints.

In addition, the prototype is shown to work at the rated speed for a long period. There is no damage to the rotor, which indicates that the design of the HSPMM is also reasonable and that the designed motor satisfied all the comprehensive performance constraints in the actual industry.

Overall, the experimental prototype tests show that the designed motor meets all physical field constraints, including electromagnetic, mechanical, and thermal characteristics, and as this paper proposed, the influent law of rotor length–diameter ratio on comprehensive characteristics is beneficial for the HSPMMs.

6. Conclusions

For the high-speed permanent magnet machines design, many rotor shape schemes can satisfy the multiple physical fields' comprehensive performances. A variety of rotor length–diameter ratios are available, from stubby to slender. However, different rotor length–diameter ratios impact on multiple physical fields' performances. The effect of rotor length–diameter ratio on multiple physical fields' comprehensive performances have rarely been focused on in the existing literature, which leads to an uncertain selection of the rotor length–diameter ratio in the design period of high-speed permanent magnet motors.

In this paper, finite element models of electromagnetic properties, rotor stresses, rotor dynamics, and temperature are built by the corresponding finite element analysis software, and the multiple physical fields performances are analyzed numerically and comprehensively comparatively analyzed through the finite element models.

The following key simulation results can be drawn from a comprehensive comparative analysis.

- When keeping the same stator outer diameters for the two-pole motor and four-pole motor, the stator yoke of the two-pole motor is magnetically saturated. The stator outer diameter of the two-pole motor is enlarged by 15% to avoid the magnetic flux density of stator-core saturation. The stator coils half-turn length of the four-pole motor is 197.258 mm, which is much shorter than the 299.492 mm of the two-pole motor. Rotor eddy-current loss of the four-pole motor is 31.52 W, which is much lower than the 235.56 W of the two-pole rotor. Compared to the two-pole motor, the four-pole motor has the higher power density.
- With the smallest rotor length–diameter ratio, Case 1 has the largest rotor eddy-current loss (121.4 W), lowest core loss (720.8 W), and largest air-friction loss (115.7 W). With the largest rotor length–diameter ratio, Case 3 has the smallest rotor eddy-current loss (5.7 W), highest core loss (923.3 W), and smallest air-friction loss (78.0 W).
- For the three design cases, Case 1 has the smallest rotor length–diameter ratio and the largest rotor outer diameter, and thus the sleeve stress and permanent magnet tangential and radial stresses are both the largest regardless of operating in cold and hot states, which are 259 MPa, 7.71 MPa, and 33.051 MPa for the cold state, and 384 MPa, 37.1 MPa, and 48.02 MPa for the hot state. Case 3 has the largest rotor length–diameter ratio and the smallest rotor outer diameter, and thus has the smallest sleeve stress and permanent magnet tangential and radial stresses in both the cold state and the hot state, which are 185.95 MPa, 1.64 MPa, and 14.881 MPa for the cold state, and 216.32 MPa, 3.98 MPa, and 31.745 MPa for the hot state.
- The influence of the rotor length–diameter ratio on rotor dynamics performance is compared and investigated. The first critical speed of the rotor is 43,800 rpm for Case 1, 68,220 rpm for Case 2, and 44,663 rpm for Case 3.
- The temperatures of the stator windings for the three cases are 86.3 °C, 56.7 °C, and 50.8 °C, respectively. The temperatures of permanent magnets for the three cases are 84.2 °C, 116.3 °C, and 126.3 °C, respectively.

In this paper, the analysis results show different rotor length–diameter ratios greatly influence multiple physical fields' comprehensive performances, and these influent laws are crucial to the design of high-speed permanent magnet motors. Thus, in the design period of high-speed permanent magnet machines, the influence of rotor length–diameter ratio on multiple physical fields' comprehensive performances should be considered and investigated, in order to obtain better multiple physical fields' comprehensive performances.

In order to verify the theoretical analysis, based on the comprehensive characteristics comparative analysis results, one 60 kW at 30,000 rpm HSPMM is designed. Multiple physical fields' comprehensive comparative analysis results are tested through prototype experiments. The RMS value of the no-load line back-EMF during rated operation is

520 V, which is very close to the calculated result 523 V and the stator winding stabilized temperature is 60.9 °C, which agrees with the analysis result 56.7 °C. The analytical method is confirmed by the experimental results of the prototype.

In the existing literature, for high-speed permanent magnet motor design, the influence of rotor length–diameter ratios on multiple comprehensive physical performances is not usually considered, so the rotor shape can be designed in various forms, such as a short and thick rotor or long and thin rotor. However, different comprehensive physical properties are produced by different rotor shapes, which are very important for the design of high-speed motors. In this paper, the influence of rotor length–diameter ratio on the comprehensive performance of multiple physical fields is considered and investigated, which includes electromagnetic characteristics, losses properties, rotor stresses distributions, rotor dynamics, and thermal behavior. Through comprehensive comparative analysis of different rotor length–diameter ratios, the following conclusions can be obtained, which have not been proposed in the previous literatures.

- For the electromagnetic characteristics, as the rotor length–diameter ratio goes up, the fundamental amplitude of the no-load line back-EMF increases gradually and the fundamental and high-order harmonical amplitudes of the no-load and rated-load radial air-gap flux density decreases.
- For the loss properties, as the rotor length–diameter ratio grows, the rotor eddy-current loss decreases, the stator-core loss increases, and the air-friction loss decreases.
- For the rotor stresses' distributions, as the rotor length–diameter ratio increases, the rotor outer diameter gradually decreases, thus the rotor stress gradually decreases, indicating that the mechanical reliability of the rotor gradually increases.
- For the rotor dynamics, smaller and larger length–diameter ratios may cause the rotor to resonate more easily.
- For the thermal behavior, as the rotor length–diameter ratio increases, the temperature of stator reduces gradually, and the temperature of rotor increases.

Based on the analysis above, it can be concluded that the influence of the rotor length–diameter ratio on multiple comprehensive physical performances cannot be ignored and different rotor length–diameter ratios greatly impact on the multiple comprehensive physical performances of HSPMMs. The conclusions obtained in this paper can be used as a reference for rotor structural design of HSPMMs.

In this paper, the destructive experiments usually performed to test rotor stresses could not be directly tested on to the rotor at high-speed rotation, due to the limitation of the test equipment. Thus, the reliability of the rotor is indirectly verified through the long-term operation of the HSPMMs.

In the future, to provide more valuable conclusions, the rotor length–diameter ratio should be further optimized by adopting an optimization analysis method, based on a coupled multi-physics performance, which could achieve a better multi-physics performance of the HSPMMs.

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Nomenclature

δ_s	Interference fit.
h_{sleeve}	The thickness of the sleeve.
h_{PM}	The thickness of the permanent magnet.
D_{is}	Stator outer diameter.
P_{out}	Output power at the rated load.
AJ	Thermal load.
δ	Air-gap length.
B	Magnetic flux density.
J	Current density.
λ	Rotor length–diameter ratio.
L_{ef}	Effective core length.
D_r	Rotor outer diameter.

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A Novel Model-Free Predictive Control for T-Type Three-Level Grid-Tied Inverters

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Abstract: The model-free predictive control (MFPC) scheme is an effective scheme to enhance the parameter robustness of model predictive control. However, the MFPC scheme can be affected by the current gradient updating frequency. This paper proposes an improved MFPC scheme for a T-type three-level inverter. First, a novel current gradient updating method is designed to estimate all current gradients per control period, which uses the current gradient relationship between different voltage vectors and eliminates the effect of current gradients updating stagnation. Moreover, a sector judgment method based on the current gradient is proposed. Redundant small vectors are accurately judged and the computational burden is greatly reduced. Finally, simulation and experimental comparisons on a T-type three-level inverter verify the effectiveness of the proposed MFPC scheme.

Keywords: model-free predictive control; T-type three-level inverter; current gradient; stagnation effect; sector judgment

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1. Introduction

In recent years, grid-tied inverters, including two-level grid-tied inverters and multi-level grid-tied inverters, have been widely applied in various new energy generation systems, such as wind power generation systems [1] and photovoltaic power generation systems [2]. As the crucial interface between renewable energy generation systems and the grid, the grid-tied inverters are needed to meet the various requirements with better steady-state performance, faster dynamic performance, higher reliability, and so on [3,4]. Although two-level inverters are commonly used because of their flexible control and reliable operation, they also have some problems such as high harmonics. Compared to two-level inverters, multi-level inverters have many advantages. Among them, the T-type three-level inverter has been studied by many scholars because of its advantages of low output harmonic, low switching loss, and high efficiency [5,6].

With the increasing application of T-type three-level inverters, various control schemes have also been developed. In particular, model predictive control (MPC) schemes have been studied in T-type three-level inverters, with the advantages of high-speed response, comprehensive control schemes, and multi-objective control ability [7–9]. However, the control performance of MPC schemes mainly depends upon the accuracy of model parameters [10]. Many methods have been designed to improve the parameter robustness of MPC. System parameters are identified online to reduce parameter errors [11–13]. The disturbances generated by inaccurate parameters are calculated and compensated in the prediction [14,15]. In these studies, it is essential to tune parameters in different observers, which increases the calculation burden in traditional processors. Moreover, an MPC based on sum of squares optimization has been developed in [16], which improves output performance by formulating the largest possible region of attraction for the considered equilibrium point and guaranteeing the stability of the Lyapunov function. Recently, model-free predictive control (MFPC) has been reported to overcome this challenge by providing robust and accurate state predictions without a complete model of the system.

Various MFPC schemes have been reported based on the ultra-local model [17,18], the auto-regressive with exogenous input (ARX) model [19], and the look-up table (LUT) [20–31]. The model-free predictive control (MFPC) scheme based on LUT was firstly introduced in [20] for permanent magnet synchronous motor (PMSM) drives, which is a computationally light yet very effective scheme. This scheme uses the current gradient of each voltage vector to predict currents. However, the current gradients of remaining voltage vectors keep the old values, which is considered as the current gradient updating stagnation effect. In the most serious case, the long-term stagnation effect can affect the stability of the control system.

To improve the updating frequency of current gradients, an MFPC scheme based on minimum updating frequency is proposed to update each gradient within the specified time [21]. The implementation process is that, if a voltage vector is not used in the past predefined frequency, it will be forcibly used in the next period and its current gradient will be updated. However, this updating method [21] affects the control performance since the non-optimal voltage vector is used frequently. In [22,23], the remaining current gradients are estimated according to the updated current gradients of the used voltage vector in the past three times. This method is effective only when the voltage vectors of the past three consecutive periods are different from each other; otherwise, the update can stagnate. In [24], the current gradients of all unused voltage vectors are estimated by using the current gradient relationship between different voltage vectors. Nevertheless, the updating can be affected when the applied vectors are the same for two consecutive periods. In [25–29], multi-vector MFPC is studied, in which two or three current gradients of the two or three applied voltage vectors are updated by using multiple adaptive sampling points. Although the updated number of current gradients increases with the increase in the number of applied voltage vectors, the stagnation effect still exists. The MFPC scheme is used in the three-level inverter-fed interior PMSM system [30], and all the current gradients are divided into seven categories. Each type of current gradient is updated according to the amplitude relationship of the corresponding voltage vector. However, among the 27 current gradients corresponding to the 27 voltage vectors of the three-level inverter system, up to six current gradients can be updated. In [31], the extended adjacent state scheme is used to reduce the possible number of remaining gradients, then the measured current gradient is used to update the remaining current gradients.

The stagnation effect becomes more obvious as the number of voltage vectors increases when the MFPC scheme is used in a T-type three-level grid-tied inverter system. To enhance the parameter robustness of predictive control, eliminate the current gradient updating stagnation effect, and improve the efficiency of voltage vector optimization, this paper proposes an improved MFPC scheme. The main contributions are shown below.

(1) A simplified look-up table (LUT) is designed. The redundant vector formed by two or three voltage vectors is considered as one corresponding current gradient. Hence, the number of LUTs decreased from 27 to 19.
(2) A novel current gradient updating method is proposed to eliminate the stagnation effect caused by the conventional updating method. The current gradient relationship between different voltage vectors is derived, so all the current gradients can be estimated in each period.
(3) A sector judgment method based on the current gradient is proposed. The proposed judgment method avoids using mathematical models to calculate the reference voltage. Hence, the number of candidate voltage vectors is reduced from 27 to 3, and the calculation speed is greatly improved.

The contents of this paper are organized as follows. In Section 2, the topology, vectors, and conventional MPC scheme are reviewed. In Section 3, the proposed scheme is introduced in detail. In Section 4, the proposed MFPC scheme is validated. In Section 5, the conclusion is given.

2. Conventional MPC Scheme

2.1. Topology and Voltage Vectors

The topology of the T-type three-level grid-tied inverter studied in this paper is shown in Figure 1, where u_{dc} is the dc-link voltage, and C_1 and C_2 are the upper and lower dc-link capacitors, respectively. O is the neutral point of dc-link capacitors and the zero-potential reference point. i_{C1} and i_{C2} are the current of C_1 and C_2, respectively. i_g is the output current. It can be seen from Figure 1 that each T-type three-level inverter bridge arm contains four switching tubes, which are S_1~S_4. There are three switching modes, including P, O, and N modes, based on the different switching combinations. All bridge arms share the neutral point of dc-link capacitors.

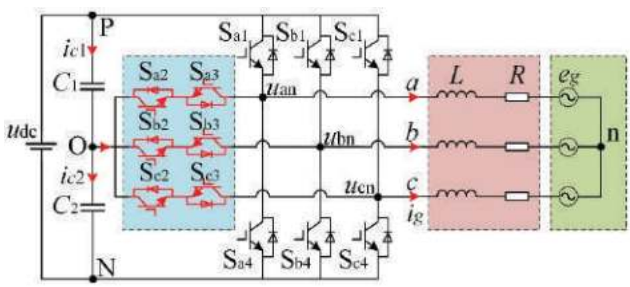


Figure 1. Topology of the T-type three-level grid-tied inverter.

There are 27 voltage vectors that can be generated by a three-phase three-level inverter, as shown in Figure 2. The specific positions in Figure 2 corresponding to multiple space vectors are called redundant vectors, whose output characteristics are equivalent. According to the amplitude of the voltage vector, the 27 switching states can be divided into a zero vector (PPP, OOO, NNN), small vector (POO, PPO, OPO, OPP, OOP, ONN, OON, NON, NOO, NNO, ONO), medium vector (PON, OPN, NPO, NOP, ONP, PNO), and large vector (PNN, PPN, NPN, NPP, NNP, PNP).

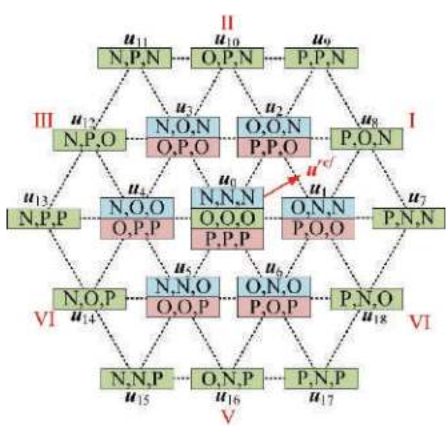


Figure 2. Basic voltage vectors of the three-level three-phase inverter.

2.2. Conventional Predictive Model

The principle of the conventional MPC scheme has been studied in many papers. Firstly, the mathematical model based on the LR-filtered three-level inverter connecting to grids in the α-β coordinate system can be expressed as

$$L\frac{di_g}{dt} = u_x - Ri_g - e_g \tag{1}$$

where $u_x = [u_{x\alpha}, u_{x\beta}]^T$, $i_g = [i_{g\alpha}, i_{g\beta}]^T$, and $e_g = [e_{g\alpha}, e_{g\beta}]^T$ represent the output voltage vectors of the inverter, output current vectors, and grid voltage vectors, respectively. L represents the filter inductance and R represents the stray resistance.

The discrete model of the inverter can be expressed as

$$i_g(k+1) = i_g(k) + \frac{T_s}{L}\left[u_x(k) - Ri_g(k) - e_g(k)\right] \tag{2}$$

where T_s is the control period.

The voltage vector selected in the last control period can be used to calculate the prediction current $i_g(k+1)$ by (2). Then, all the voltage vectors shown in Figure 2 can be further used to calculate prediction currents $i_{gx}(k+2)$ by (3).

$$i_{gx}(k+2) = \left(1 - \frac{RT_s}{L}\right)i_g(k+1) + \frac{T_s}{L}\left[u_x(k+1) - e_g(k+1)\right] \tag{3}$$

Finally, to evaluate the control performance of each voltage vector, a cost function shown in (4) is defined. The voltage vector that minimizes the cost function is used as the optimal vector and applied to the next control period.

$$G_x = \left(i_{ref}(k+2) - i_{gx}(k+2)\right)^2 \tag{4}$$

where $i_{ref}(k+2)$ are the reference currents.

In a three-level inverter control system, in addition to the output current, the neutral point voltage needs to be controlled. To avoid the design of weighting factor values when multiple control targets exist, MPC schemes without weighting factors have been proposed. In the MPC scheme without weighting factors, the P-type small vectors (POO, PPO, OPO, OPP, OOP) are considered as candidate vectors when the capacitor voltage $u_{c1} \geq u_{c2}$, and the N-type small vectors (ONN, OON, NON, NOO, NNO, ONO) are considered as candidate vectors when the capacitor voltage $u_{c1} < u_{c2}$. The control block diagram of MPC is shown in Figure 3.

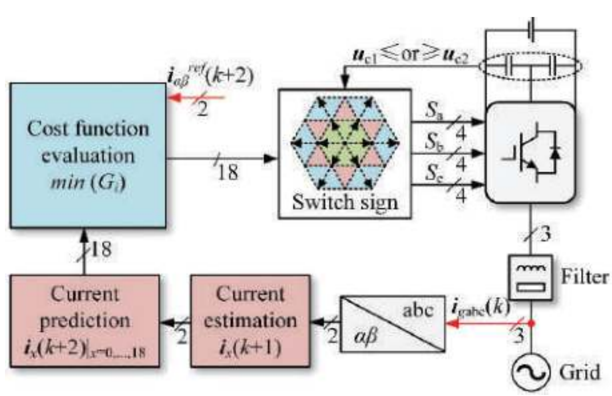


Figure 3. Control block diagram of the conventional MPC scheme.

However, the prediction currents are affected when model parameters are mismatched with the controller parameters. The prediction currents with parameter errors can be expressed as

$$i_{gerr}(k+1) = i_g(k) + \frac{T_s}{L+\Delta L}\left[u_x(k) - (R+\Delta R)i_g(k) - e_g(k)\right] \tag{5}$$

where ΔL represents the inductance error between its actual value and its controller value, and ΔR represents the resistance error between its actual value and its controller value. Hence, it is important to enhance the parameter robustness of predictive control.

3. The Proposed MFPC Scheme

To eliminate the dependence on model parameters, an improved MFPC scheme based on a novel current gradient updating method is proposed in this section, including the basic principle of MFPC, the current gradient updating stagnation effect analysis, the proposed current gradient updating method, the sector judgment method, and the implementation steps.

3.1. Basic Principle of MFPC

In the MFPC scheme, the mathematical model (1) can be rewritten as [17]

$$\begin{aligned}\Delta i_x(k-1) &= \tfrac{T_s}{L}\left(u_x(k-1) - e_g(k-1) - Ri_g(k-1)\right)\\ &= i_g(k) - i_g(k-1)\end{aligned} \tag{6}$$

where $\Delta i_x(k-1) = [\Delta i_{x\alpha}(k-1), \Delta i_{x\beta}(k-1)]^T$ are the current gradients of the applied voltage vector $u_x(k-1)$. Although the switching states of redundant vectors are different, their coordinate components are the same. The multiple current gradients corresponding to redundant vectors are considered to be consistent. For example, voltage vector u_1 corresponds to two switching states of ONN and POO, so the current gradient corresponding to the two states is uniformly defined as Δi_1. Hence, nineteen current gradients Δi_x are stored in the look-up table (LUT) for current predictions, as shown in Table 1.

Table 1. The LUT of the T-type three-level inverter.

Voltage Vector	u_0	u_1	u_2	u_3	u_4	u_5	u_6	u_7	u_8	u_9	u_{10}	u_{11}	u_{12}	u_{13}	u_{14}	u_{15}	u_{16}	u_{17}	u_{18}
Current Gradient	Δi_0	Δi_1	Δi_2	Δi_3	Δi_4	Δi_5	Δi_6	Δi_7	Δi_8	Δi_9	Δi_{10}	Δi_{11}	Δi_{12}	Δi_{13}	Δi_{14}	Δi_{15}	Δi_{16}	Δi_{17}	Δi_{18}

Then, the prediction currents at the (k + 1)th instant and the (k + 2)th instant can be calculated based on (7) and (8), respectively.

$$i_{gx}(k+1) = i_g(k) + \Delta i_x(k) \tag{7}$$

$$i_{gx}(k+2) = i_{gx}(k+1) + \Delta i_x(k+1) \tag{8}$$

Finally, the optimal voltage vector u_{op} selected by cost function (4) is applied in the next control period.

3.2. Current Gradient Updating Stagnation Effect Analysis

It is obvious that the current prediction of MFPC is based on the current gradients which directly affect the accuracy of current prediction and the result of optimal voltage vector selection. The current gradients of applied voltage vectors are calculated based on (6) in the conventional MFPC scheme [20], however, the updating of the remaining current gradients stagnates. When the MFPC scheme is used in a three-level inverter system, the stagnation effect can be more obvious with the increasing number of voltage vectors, which not only affects the current performance but also affects the control of the neutral point voltage.

To reduce the updating stagnation effect, various current gradient updating methods have been reported in [21–24,30]; however, the updating stagnation is still existing. Hence, it is necessary to design an updating method to totally eliminate the stagnation effect.

3.3. The Proposed Current Gradient Updating Method

In the proposed updating method, sampling points are set before switching states in each control period to avoid the current spikes. Based on (6), the relationship between the current gradient of remaining voltage vector Δi_y and the current gradient of applied voltage vector Δi_x at $(k-1)$th instant and $(k-2)$th instant can be expressed as (9) and (10), respectively.

$$\Delta i_y(k-1) - \Delta i_x(k-1) = \frac{T_s}{L}\left(u_y(k-1) - u_x(k-1)\right) \tag{9}$$

$$\Delta i_y(k-2) - \Delta i_x(k-2) = \frac{T_s}{L}\left(u_y(k-2) - u_x(k-2)\right) \tag{10}$$

The filter inductance L can be considered as a constant when T_s is short enough. Then, T_s/L of (9) and (10) can be eliminated by the division between (9) and (10), and the relationship between the current gradient in the two control periods can be written as

$$\Delta i_y(k-1) = \frac{u_y(k-1) - u_x(k-1)}{u_y(k-2) - u_x(k-2)}\left(\Delta i_y(k-2) - \Delta i_x(k-2)\right) + \Delta i_x(k-1) \tag{11}$$

However, (11) cannot be used to update the current gradient when $u_y(k-2) - u_x(k-2) = 0$. In an α-axis coordinate system, seven situations may cause the stagnation effect. They are $u_{8\alpha} = u_{18\alpha}$, $u_{1\alpha} = u_{9\alpha} = u_{17\alpha}$, $u_{2\alpha} = u_{6\alpha}$, $u_{0\alpha} = u_{10\alpha} = u_{16\alpha}$, $u_{3\alpha} = u_{5\alpha}$, $u_{4\alpha} = u_{11\alpha} = u_{15\alpha}$, and $u_{12\alpha} = u_{14\alpha}$, respectively. In a β-axis coordinate system, five situations may cause the stagnation effect. They are $u_{9\beta} = u_{10\beta} = u_{11\beta}$, $u_{2\beta} = u_{3\beta} = u_{8\beta} = u_{12\beta}$, $u_{0\beta} = u_{1\beta} = u_{4\beta} = u_{7\beta} = u_{13\beta}$, $u_{5\beta} = u_{6\beta} = u_{14\beta} = u_{18\beta}$, and $u_{15\beta} = u_{16\beta} = u_{17\beta}$, respectively. To eliminate the stagnation effect, (9) under the α-axis and β-axis is derived as (12) and (13), respectively. For example, when the applied voltage vector is u_2, $\Delta i_{2\alpha\beta}$ can be obtained based on (6). Then, in addition to $\Delta i_{6\alpha}$ (calculated based on (12)), other α-axis current gradients can be calculated based on (11). Moreover, in addition to $\Delta i_{3\beta}$, $\Delta i_{8\beta}$, and $\Delta i_{12\beta}$ (calculated based on (13)), other β-axis current gradients can be calculated based on (11).

$$\Delta i_{y\alpha}(k-1) = \Delta i_{x\alpha}(k-1) \tag{12}$$

$$\Delta i_{y\beta}(k-1) = \Delta i_{x\beta}(k-1) \tag{13}$$

As a result, the stagnation effect is eliminated and the nineteen current gradients can be updated in each control period, as shown in Figure 4.

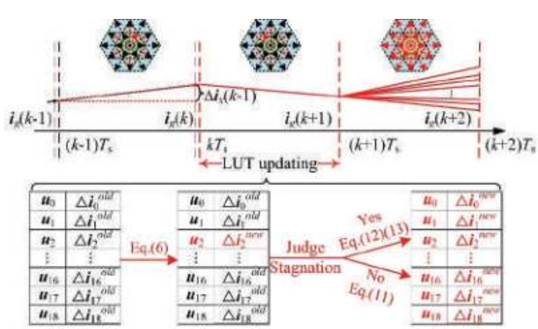


Figure 4. Proposed current gradient updating method.

3.4. Proposed Sector Judgment Method

Conventionally, in order to find the optimal voltage vector through the cost function (4), 27 voltage vectors need to be traversed. When the redundancy state of the zero vector is ignored and the small vectors are selected as P-type small vectors or N-type small vectors according to the dc-link capacitor voltage difference, there are still 19 voltage vectors to be traversed. Since the calculation of reference voltage depends on the mathematical model, the sector judgment method based on reference voltage may cause sector judgment deviation when the model parameters are mismatched. Therefore, a sector judgment method based on the current gradient is proposed in this section.

Figure 5 shows the 19 voltage vectors and their corresponding 19 current gradients. The three-level inverter space vector coordinate system is divided into 6 large sectors and 24 small sectors. The position of the large sector where the reference current is located is judged according to the prediction currents of the medium vectors. For example, if the reference current is in large sector I, the cost of u_8 calculated by (4) is the smallest compared with the cost of u_{10}, u_{12}, u_{14}, u_{16}, and u_{18}.

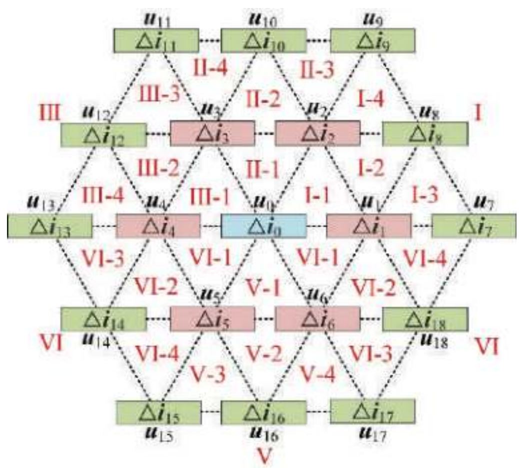


Figure 5. Big and small sectors of the three-level three-phase inverter.

After judging the large sector, it is necessary to judge the small sector. First, the prediction currents can be calculated by (8). Then, the weighted error square ε^2 of the four small sectors can be calculated by (14) and selecting the smallest ε^2 corresponding sector as the target sector. Finally, there are only three voltage vectors as candidate vectors.

$$\varepsilon^2 = \sum_{x=1}^{3} \left(i_{ref}(k+2) - i_{gx}(k+2) \right)^2 \tag{14}$$

3.5. Implementation Steps

Figure 6 shows the control diagram of the proposed MFPC scheme. The proposed MFPC scheme is realized to eliminate parameter dependence, the novel current gradient updating method is realized to eliminate the stagnation effect, and the sector judgment method is also realized to further reduce the computational burden over the conventional MFPC scheme. The detailed implementation steps are as follows.

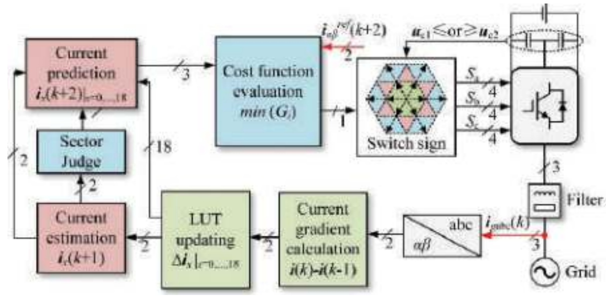


Figure 6. Control block diagram of the proposed MFPC scheme.

Step I: Sample the dc-link capacitor voltage $u_{c1}(k)$ and $u_{c2}(k)$, select P-type basic voltage vectors or N-type basic voltage vectors based on $u_{c1}(k) - u_{c2}(k) \leq$ or >0.
Step II: Sample the output current $i_g(k)$ and calculate current gradient $\Delta i_x(k-1)$ by (6).
Step III: Update the remaining current gradients without stagnation effect by (11), and update the remaining current gradients with stagnation effect by (12) and (13).
Step IV: Calculate prediction current $i_{gx}(k+1)$ and $i_{gx}(k+2)$ by (7) and (8), respectively.
Step V: Judge large sectors by (4), and judge small sectors by (14).
Step VI: Evaluate the cost of the three voltage vectors by (4) and select the optimal vector.

4. Simulation and Experimental Evaluation

In order to verify the effectiveness of the proposed MFPC scheme, simulations in MATLAB/Simulink and experiments under the conventional MPC scheme, conventional MFPC schemes, and the proposed MFPC scheme are carried out. Figure 7 depicts the T-type three-level three-phase grid-connected inverter experimental setup. The main control chip of the inverter is DSP28335. The sampling frequency is set as 20 kHz. The inverter parameters are listed in Table 2.

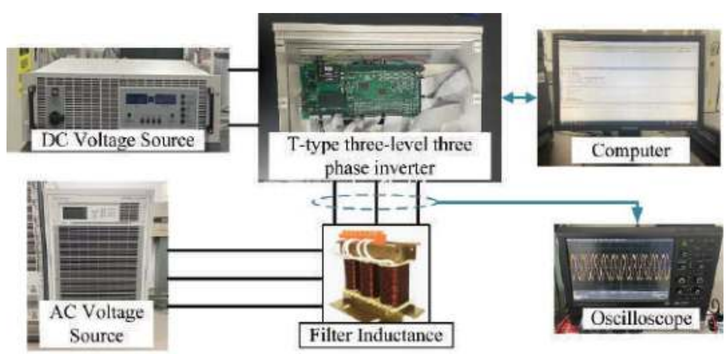


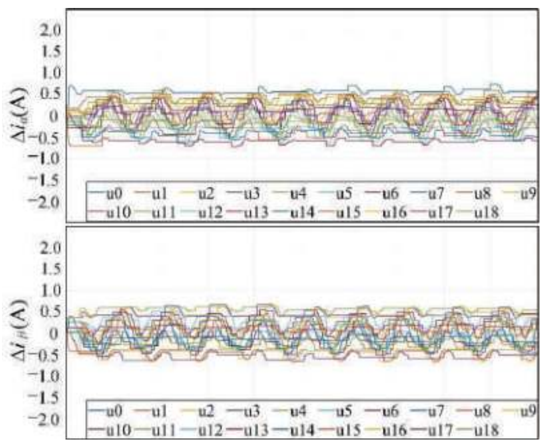
Figure 7. T-type three-level three-phase grid-connected inverter experimental platform.

Table 2. System and control parameters.

Parameters	Symbol	Values
DC-link voltage	u_{dc}	300 V
Peak of grid phase voltage	e	150 V
Grid angular frequency	ω_g	314.16 rad/s
Control time	T_s	50 µs
Parasitic resistance	R	0.05 Ω
Filter inductance	L_0	10 mH

4.1. Impact of Current Gradient Updating Stagnation

The proposed current gradient updating method is compared with the updating method in [20] and the updating method in [30]. Figure 8a illustrates the α-axis and β-axis current gradients when the updating method in [20] is used. In this current gradient updating method, there is a notable stagnation effect in both the current gradients Δi_α and Δi_β because the current gradients of the applied voltage vector can be updated for the LUT; however, the remaining current gradients keep the old values. Compared to the two-level inverter system [20], the stagnation effect becomes more obvious with the increasing number of voltage vectors when the updating method [20] is used in the three-level inverter system. As shown in Figure 8b, although the updating frequency is improved, the stagnation effect still exists in the α-axis and β-axis current gradients when the updating method in [30] is used. When the updating method is converted to the proposed updating method, it can be seen that there is no stagnation effect existing in the current gradient waveforms, as shown in Figure 8b, which shows the effectiveness of the proposed updating method.



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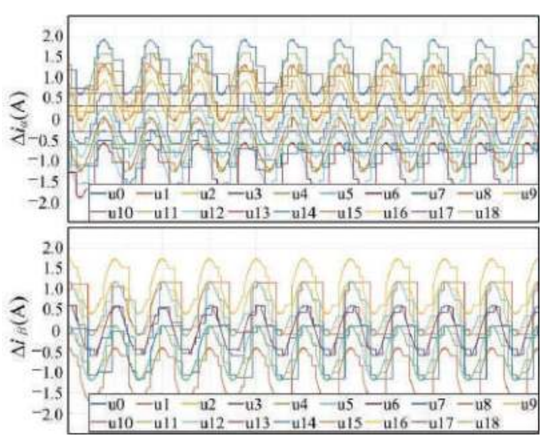

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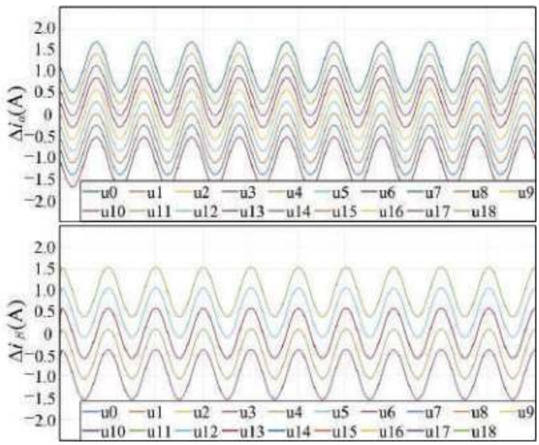
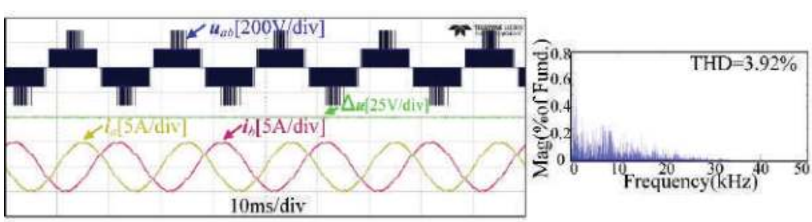


(**c**)

Figure 8. Current gradient comparisons under different updating methods (i_{ref} = 5A). (**a**) Updating method in [20]. (**b**) Updating method in [30]. (**c**) The proposed updating method.

4.2. Steady-State Experimental Evaluation

Figure 9 shows the steady-state performance comparisons of the conventional MPC scheme, conventional MFPC scheme in [20], conventional MFPC scheme in [30], and the proposed MFPC scheme when model parameters match controller parameters. Based on the steady-state experimental waveforms, it can be seen that both the conventional MPC scheme, the conventional MFPC scheme in [30], and the proposed MFPC scheme can achieve tracking the reference currents and balancing the neutral point voltages. However, the neutral point voltage of the conventional MFPC scheme in [20] is affected by the current gradient stagnation effect. The harmonic spectra of the current are drawn by MATLAB/Simulink with the experimental data obtained from the oscilloscope, as shown in Figure 9.



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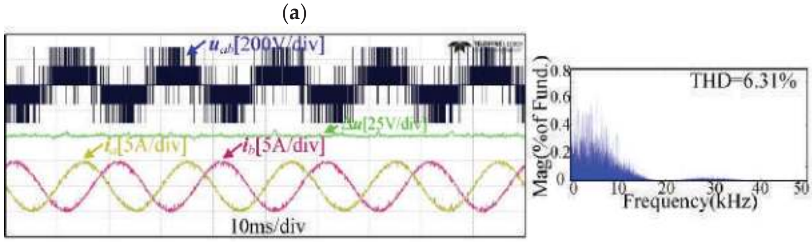

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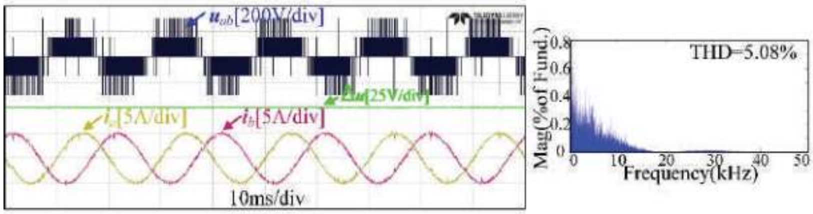
Figure 9. *Cont.*



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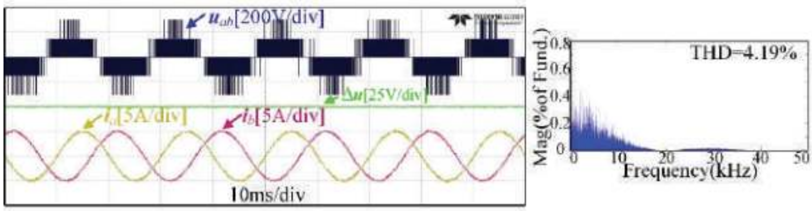


(**d**)

Figure 9. Experimental waveform comparisons (i_{ref} = 5A). (**a**) Waveforms under the conventional MPC scheme. (**b**) Waveforms under the conventional MFPC scheme in [20]. (**c**) Waveforms under the conventional MFPC scheme in [30]. (**d**) Waveforms under the proposed MFPC scheme.

For the conventional MPC scheme, the total harmonic distortion (THD) of phase-a output current is 3.92%. For the conventional MFPC scheme in [20], the THD of phase-a output current changes from 3.92% to 6.31% because the current performance is affected by the stagnation effect and the sampling noise effect. For the conventional MFPC scheme in [30], the THD of phase-a output current changes from 6.31% to 5.08% because the stagnation effect is improved, but it cannot be eliminated. Moreover, the current spikes caused by the stagnation effect can be observed simultaneously in Figure 9b,c. Compared with the conventional MFPC scheme in [30], the THD of phase-a output current reduces from 5.08% to 4.19% because the stagnation effect is eliminated, as shown in Figure 9d. However, the current performance of the proposed scheme is worse than that of the conventional MPC scheme because the current performance of the MFPC scheme is affected by the sampling noise.

To further verify the effectiveness of the proposed MFPC scheme, other experimental comparisons under different reference currents are carried out. The THDs of phase-a current under four control schemes are shown in Figure 10. From Figure 10, the THDs of the four control schemes are reduced with the increased reference currents. The proposed MFPC still has a similar current THD to the MPC and has a lower current THD than the MFPC in [20] and the MFPC in [30].

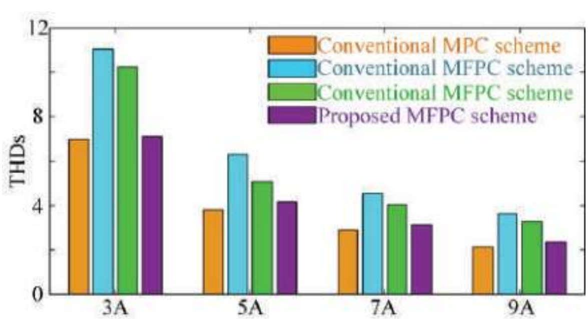
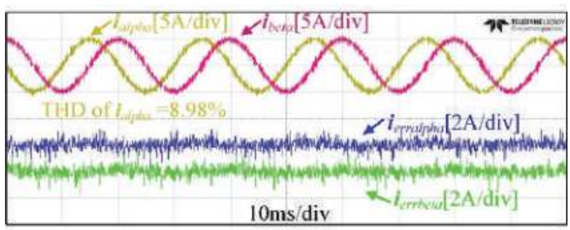


Figure 10. The comparison of phase-a current THDs under different reference currents.

4.3. Experimental Evaluation under Mismatched Model Parameters

To verify the effectiveness of the proposed MFPC scheme in enhancing parameter robustness, the control performance of the proposed scheme and the conventional MPC scheme are compared in this section when the filter inductance as the model parameter is taken as different actual values from the controller values L_0. It can be seen that compared with the conventional MPC scheme, the proposed scheme reduces the THD from 8.98% to 6.25% when the actual filter inductance is set as 0.05 H (i.e., $0.5L_0$) as shown in Figure 11. Additionally, the proposed scheme reduces the THD from 2.78% to 1.87% when the actual filter inductance is set as 0.2 H (i.e., $2L_0$), as shown in Figure 12. Under the above two experimental conditions, the current prediction errors of the proposed scheme are lower than that of the conventional MPC scheme. The current waveforms, the THDs, and the current prediction errors verify that the proposed scheme has stronger parameter robustness than the conventional MPC scheme.



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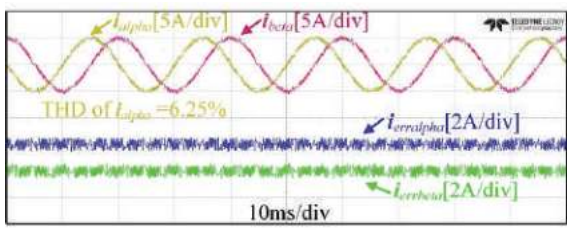
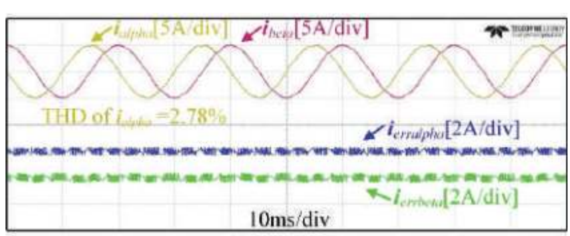


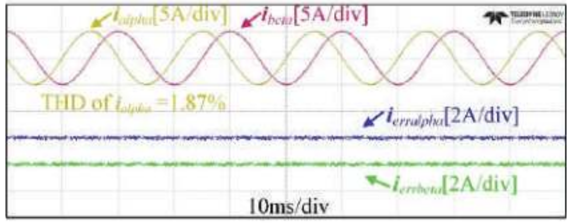
(**b**)

Figure 11. Experimental waveforms comparisons when L/L_0 = 0.5 (i_{ref} = 5A). (**a**) Waveforms under the conventional MPC scheme. (**b**) Waveforms under the proposed MFPC scheme.



(**a**)

Figure 12. *Cont.*



(**b**)

Figure 12. Experimental waveforms comparisons when L/L_0 = 2 (i_{ref} = 5A). (**a**) Waveforms under the conventional MPC scheme. (**b**) Waveforms under the proposed MFPC scheme.

4.4. Performance Comparison of All Schemes

In order to further demonstrate the advantages of the proposed MFPC, various performances of all schemes are compared and summarized in Table 3, including THDs, computational burden, parameter robustness, and stagnation effect. The proposed MFPC has a similar THD to MPC when model parameters are accurate and have better robustness against mismatched parameters than MPC. Though 18 current gradients in the proposed MFPC should be updated per control period, the speed of optimization is effectively reduced by the proposed sector judgment method. Hence, the proposed MFPC has a similar computational burden to MPC. For the MFPC in [20,30], although they cannot be affected by the mismatched parameters, their THDs are higher than those of the proposed MFPC because of the stagnation effect. For the MFPC in [20], it has the smallest computational burden because there is no current gradient updating exists, which leads to the largest THD. For MFPC in [30], it has the largest computational burden because there is all current gradient is updated and candidate voltage vectors are not reduced.

Table 3. Performance comparison of different control schemes.

Schemes	THDs	Computational Burden	Parameter Robustness	Stagnation Elimination
Conventional MPC	3.92%	35.9 μs	No	-
Conventional MFPC in [20]	6.31%	31.8 μs	Yes	No
Conventional MFPC in [30]	5.08%	41.2 μs	Yes	No
Proposed MFPC	4.19%	36.2 μs	Yes	Yes

5. Conclusions

This paper proposes an improved MFPC scheme based on the novel current gradient updating method and the effective sector judgment method for a T-type three-level grid-tied inverter. The proposed scheme has the following salient features: First, compared to conventional MPC schemes without a weighting factor, it totally eliminates the effect of mismatched model parameters and enhances the parameter robustness, especially when the actual parameters are less than the control parameters. Then, compared to conventional MPC schemes using a weighting factor, it avoids designing the weight factor and still realizes the good control of current and neutral point voltage. Moreover, compared to conventional MFPC schemes, it eliminates the current gradient stagnation effect by the designed updating method. The THDs of the proposed scheme decrease by up to 2.12% and the current spikes caused by the long updating stagnation are avoided. Finally, the discrimination accuracy of redundant small vectors is improved by using the proposed sector judgment method based on the current gradient, and the speed of optimization is increased by 31.5%.

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Conflicts of Interest: The authors declare that there is no conflict of interest.

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Article

A General Equivalent Modeling Method for DFIG Wind Farms Based on Data-Driven Modeling

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Abstract: To enhance the stable performance of wind farm (WF) equivalent models in uncertain operating scenarios, a model-data-driven equivalent modeling method for doubly-fed induction generator (DFIG)-based WFs is proposed. Firstly, the aggregation-based WF equivalent models and the equivalent methods for aggregated parameters are analyzed and compared. Two mechanism models are selected from the perspective of practicality and complementarity of simulation accuracy. Secondly, the simulation parameters are set through two sampling methods to construct a training database. Next, the whole fault process is divided into five phases, and the weight coefficient optimization model is established according to the data-driven idea to achieve the adaptive configuration of the weight. Finally, the electromechanical transient simulations of the power systems with a DFIG-based WF is carried out by using the MATLAB/Simulink platform. Compared with the detailed WF model, the simulation time of the WF equivalent proposed in this paper can be significantly reduced by about 87%, and simulation results show that the proposed method can effectively improve the adaptability of the WF equivalent model in different wind scenarios and voltage dips.

Keywords: DFIG-based wind farm; general equivalent modeling; mechanism model; data-driven; weight coefficient

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1. Introduction

Increasing the penetration of large-scale wind farms (WFs), wind power has become an important power source in power systems [1,2]. Wind power has randomness and volatility, which brings deep changes in the operation mechanism [3–5]. To support rapid development and reduce the operation risks, simulation technology is increasingly indispensable to reflect the behavior of the actual power systems and lead scientific construction of new-type electric power systems. Large-scale WFs may consist of hundreds or even thousands of wind turbines (WTs), which could significantly enlarge the size of A model and then cause the "curse of dimensionality" [6]. Therefore, the equivalent model, on the basis of reasonable reduction from the detailed model, is essential to be developed.

Currently, the aggregation-based method, which was originally used in synchronous generators, is widely applied to model a large-scale WF in the literature [7,8]. The aggregated model can be divided into a single-machine equivalent model (SEM) and a multi-machine equivalent model (MEM) [9]. The single-machine equivalent method, which aggregates the whole WF into one equivalent WT, requires a small amount of calculation, but it is hard to represent dynamic behaviors of a whole WF due to the distribution of collector cables and wind speed differences across the WF. The multi-machine equivalent method, which separates WTs of a WF into several clusters and aggregates each cluster into one equivalent WT, generally represents WF characteristics better and has wider applications in practice [10]. The MEM includes two steps: (1) identify the group of WTs with similar dynamic characteristics, and (2) obtain the aggregated parameters of the WTs, transformers, and collector cables [11].

During the past few decades, to obtain a better performance on WT clustering, wind speed [12,13], rotor speed [14], the power characteristic curve [15,16], crowbar action [17,18], chopper action [19] and other quantities have been fully investigated to be selected as the clustering indicators. Wind speed is regarded as the primary clustering indicator. Crowbar or chopper action is considered more suitable for grouping WTs when simulating low voltage ride through characteristics. However, the selection of these clustering indicators is generally based on specific time spots, and limitations universally exist in these studies. For example, WT characteristics dynamically change along with the time that passes, and the time spot-based clustering results are suitable in limited situations. In [20], WTs connected to the same feeder were aggregated into one equivalent WT, but this approach, might introduce large equivalent errors when wind speeds of WTs within a feeder have large deviations. Recently, a WF equivalent modeling method based on feature influence factors and improved back propagation (BP) neuron networks algorithm was proposed, which improves the efficiency and accuracy when grouping large amounts of WTs [21]. The clustering result is still calculated at a specific steady state, and keeps changing due to the unpredictability of environmental scenarios. Therefore, the MEM based on the operation point is stochastic, and both structures and parameters of the equivalent model are random. Conclusions under different scenarios will also be different.

As for the uncertainty problem of the MEM, several studies propose a probabilistic clustering concept for aggregate modeling of WFs. In [22], a weighted graph representing relationship of the power of WTs was used to build the Markov chain in order to estimate the probability that WTs belong to the same cluster. In [23], a probabilistic equivalent model was constructed considering the probability distribution characteristics of wind speed and wind direction in an actual DFIG-based WF. On this basis, the probability distribution characteristics of the fault type is taken [24]. Moreover, the paper merges clustering results with insignificant differences through the significance test of the Fisher discriminant, and improves the generality of the probabilistic WF equivalent model. In [25], historical meteorological data were utilized to investigate the probability distribution of key equivalent parameters, such as capacity, wind speed and electrical impedance to the point of common coupling. In [26], an equivalent model for mixed WFs based on BP neural networks was constructed. However, it is difficult for this data-driven modeling approach to achieve the desired simulation accuracy.

Accordingly, this paper proposes an innovative general equivalent modeling method for DFIG-based WFs. The main contributions are listed as follows:

(1) The method of combining model-driven and data-driven is introduced into the research on general equivalent modeling of DFIG-based WFs. For uncertain scenarios, the established model has a wide range of adaptability.
(2) Meaningful insights into how to select mechanism models are provided. Considering different calculation methods for the equivalent parameters of the collector cables, two mechanism models with complementary characteristics are selected.

The remainder of this paper is organized as follows. In Section 2, the ideas and methods based on a data-driven model are introduced. The two sampling methods and training database construction are presented in detail in Section 3. In Section 4, the proposed model is verified with different voltage dips and changing wind scenarios. The discussion and conclusions are discussed in Sections 5 and 6.

2. Ideas and Methods

The following issues need to be taken into account for researching the general equivalent modeling of DFIG-based WFs. One is how to cope with the simulation demand of uncertain operation scenarios and enhance the adaptability of the general model, and the other is how to improve the convenience of using the model while ensuring the accuracy of the model and enhancing the engineering utility value.

To address the two issues, the framework of the model-data-driven general equivalent modeling method for DFIG-based WFs is shown in Figure 1, which includes selecting the mechanism model, setting the weight coefficients, and constructing the training database.

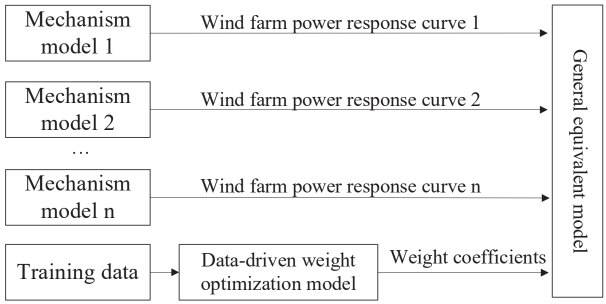


Figure 1. Framework of general equivalent modeling method.

2.1. Mechanism Model Selection

Suitable mechanism models can effectively improve the accuracy of the general equivalent model. The mechanism model of DFIG-based WFs is developed with an aggregation-based method, of which the SEM is the simplest, and the MEM by clustering WTs with similar wind speeds or electrical distances is more common. Therefore, the three equivalent models are tentatively selected as the mechanism models in the paper.

For the three mechanism models, the MATLAB/Simulink platform was used to conduct simulation studies and model accuracy analysis. The simulation results show that the dynamic performances of the SEM and the MEM by clustering WTs with similar electrical distances were almost the same, and the equivalent accuracy of both was lower than that of the MEM by clustering WTs with similar wind speeds. Considering that the selected mechanism model needs to have high accuracy, the MEM by clustering WTs with similar wind speeds was finally chosen as the mechanism model.

In addition, the mechanism model needs to obtain the aggregated parameters of the WTs, transformers, and collector cables. At present, the aggregated parameters of the WTs have been widely agreed upon, i.e., the calculation of aggregated wind speed is based on equal total wind energy, and the calculation of aggregated parameters of WTs is based on the capacity weighting method. In addition, the calculation of aggregated parameters of transformers is also based on the capacity weighting method. However, there are two different calculation methods for the aggregated parameters of the collector cables. One is the equal loss power method [27], and the other is the equal voltage dip method [28]. The corresponding methods are detailed in Equations (1) and (2), respectively:

$$Z_{\mathrm{eq}} = \frac{\sum\limits_{i=1}^{m} (P_{Z\_i}^{2} Z_{1\_i})}{\left(\sum\limits_{i=1}^{m} P_i\right)^2} \tag{1}$$

$$Z_{\mathrm{eq}} = \frac{\sum\limits_{i=1}^{m} (\sum\limits_{j=1}^{i} (Z_{1\_j} P_{Z\_j}) P_i)}{\left(\sum\limits_{i=1}^{m} P_i\right)^2} \tag{2}$$

where Z_{1_i} is the cable impedance of the ith WT branch, P_{Z_j} is the total active power flowing through the impedance Z_{1_j}, P_i is the active power of the ith WT, and m is the number of WTs in the clustering groups.

The aggregated parameters of the collector cables calculated by the two methods had a significant difference, especially for the WTs at the end of one feeder. Therefore, we finally selected the mechanism model by clustering WTs with similar wind speeds, together with the aggregated parameters of the collector cables by the equal power loss method, as mechanism model 1 (MM1). The mechanism model clustering WTs with similar wind speeds, together with the aggregated parameters of the collector cables by the equal voltage dip method, was selected as mechanism model 2 (MM2).

2.2. The Data-Driven Adaptive Weight Optimization Model

To set the weight coefficients, the simplest way is to ignore the differences among models and assign the same weight. However, the accuracy of this general WF equivalent model will not significantly improve compared to the traditional model. Therefore, we adopted a time-varying weight coefficient, which means different models had different weight coefficients, and the same model also had different weight coefficients in differen phases.

To find the best weight coefficients at different periods, the particle swarm optimization (PSO) algorithm was adopted [29]. The fitness function of the PSO algorithm was set to the root mean square error of the general WF equivalent model in different phases, that is:

$$\begin{aligned} & optfunction(\omega_1, \omega_2) = \sqrt{\sum_{i=1}^{n} \left(\sum_{k=1}^{m} [\omega_1 y_1(k) + \omega_2 y_2(k) - y_0(k)]^2 \right)} \\ & s.t. \quad \omega_1, \omega_2 \in [0,\ 1] \end{aligned} \tag{3}$$

where ω_1,ω_2 are the weight coefficients of MM1 and MM2 for a specific phase, respectively. $y_1(k)$, $y_2(k)$ and $y_0(k)$ are the output values of MM1, MM2, and detailed WF model at the kth sampling point within a phase, respectively, m is the number of sampling points, and n is the sample size of the training data.

3. Data Source

3.1. Data Settings

The wind speed and external fault information are required for WF equivalent modeling. Considering the completeness of data samples, this paper adopts two ways to generate data samples. One is simple random sampling within the selected range, and the other is sampling based on the probability density function. For the first sampling method, six scenarios were considered, i.e., low wind speed range (4.5~8.5 m/s), medium wind speed range (8.5~12.5 m/s), high wind speed range (12.5~22 m/s), medium-low wind speed range (4.5~12.5 m/s), and medium-high wind speed range (8.5~22 m/s) are considered. Next, the input wind speed of each WT was randomly generated with equal probability by ignoring the wake effect. The voltage dip at the point of connection (POC) was drawn with equal probability in the range of 0.1~0.9 p.u. For the second sampling method, considering the statistical characteristics of wind speeds, the wind speed and wind direction of the WF were generated based on the probability density function. Next, the input wind speed of each WT was derived based on the wake effect. In addition, the ground resistance values were also extracted based on the probability density function.

In this paper, for the first sampling method, 50 sets of sample data were generated for each scenario. For the second method, 200 sets of sample data were drawn according to the probability.

3.2. Training Database

Taking the large capacity of data samples into account, we adopted the MATLAB 2020b platform to carry out numerical simulations using m-language programming and Simulink to call each other. The simulation results were automatically stored and called for the adaptive weight optimization model based on the PSO algorithm. The specific flow is shown in Figure 2.

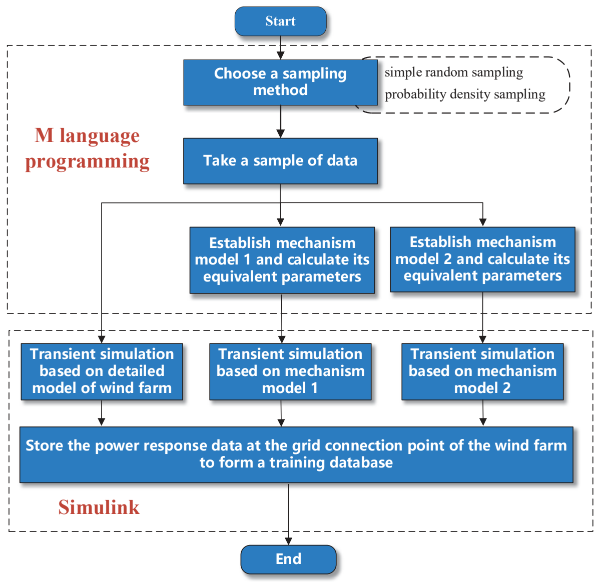


Figure 2. Training database construction flowchart.

4. Example Analysis

4.1. Example of Calculation

A detailed WF consisting of 28 × 1.5 MW DFIGs (DFIG_1—DFIG_28) was set up in the MATLAB/Simulink platform, as shown in Figure 3. The model parameters are given in Table A1.

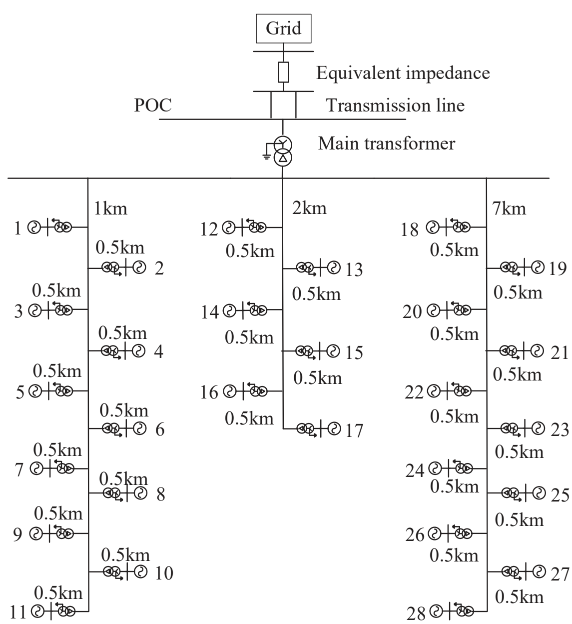


Figure 3. DFIG-based WF topology.

Compared with the simulation results of the detailed WF model (DM), error metrics are defined as:

$$E_{\mathrm{r}} = \frac{1}{m}\sum_{k=1}^{m}\left|\frac{y_{\mathrm{f}k} - y_k}{y_k}\right| \times 100\% \tag{4}$$

$$E_{\mathrm{a}} = \frac{1}{m}\sum_{k=1}^{m}\left|\mathrm{y}_{\mathrm{f}k} - y_k\right| \tag{5}$$

where y_k and y_{fk} are the output values of the detailed WF model and the equivalent model at the kth sampling point, respectively, and m is the number of sampling points.

4.2. Optimization of Adaptive Weight Coefficients

As previously commented in Section 2.2, the PSO algorithm was used to find the optimal weight coefficients of the mechanism models. The numbers of population size and iterations were set to 500 and 400, respectively. Moreover, inertia weight φ is an essential parameter of the PSO algorithm, and we adopted a dynamic adjustment inertia weight strategy, where the inertia weight is dynamically adjusted according to a linear decreasing approach:

$$\varphi(j) = \varphi_{\max} - (\varphi_{\max} - \varphi_{\min})(j/j_{\max}) \tag{6}$$

where $\varphi_{\max} = 0.9$, $\varphi_{\min} = 0.1$, $j_{\max}$ is the maximum number of iterations, and j is the number of current iterations.

According to the time window division method of IEC 61400-27-1 [30], the whole fault process is divided into five phases, namely, pre-fault $(t_0 - 1\mathrm{s},\ t_0)$, early fault $(t_0,\ t_0\ +\ 0.14\mathrm{s})$, the quasi-steady-state phase during the fault $(t_0 + 0.14s,\ t_1)$, early fault recovery $(t_1,\ t_1\ +\ 0.5\mathrm{s})$, and the quasi-steady-state phase during the fault recovery $(t_1 + 0.5\mathrm{s},\ t_1 + 5\mathrm{s})$ (t_0 and t_1 denote the initial time of fault and the time of fault clearance, respectively).

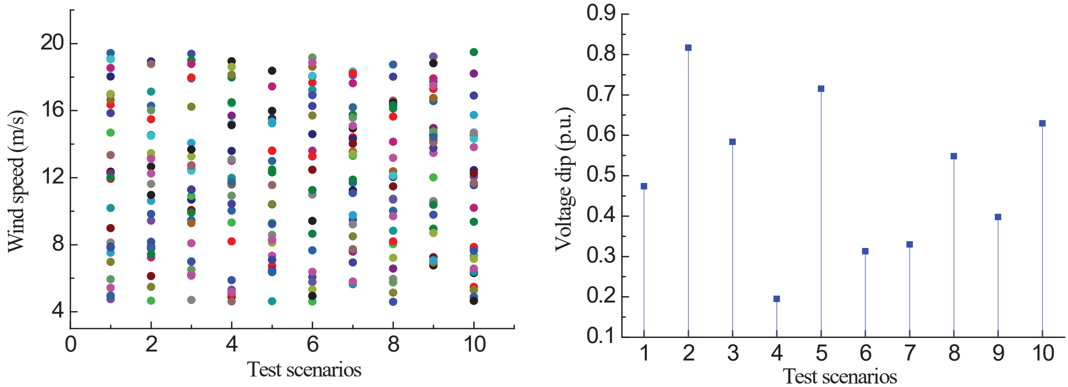
After the iterative optimization of the PSO algorithm, the weight coefficients of the above five time windows corresponding to MM1 and MM2 are shown in Table 1.

Table 1. Adaptive weight coefficients.

Models Category	**Active Power Weight Coefficients**				
	$(t_0 - 1s, t_0)$	$(t_0, t_0 + 0.14s)$	$(t_0+0.14s, t_1)$	$(t_1, t_1 + 0.5s)$	$(t_1 + 0.5s, t_1 + 5s)$
MM1	0.9713	0.3982	0.7163	0.0065	0.7270
MM2	0.0296	0.6048	0.2855	0.9982	0.2728
Models Category	**Reactive Power Weight Coefficients**				
	$(t_0 - 1s, t_0)$	$(t_0, t_0 + 0.14s)$	$(t_0+0.14s, t_1)$	$(t_1, t_1 + 0.5s)$	$(t_1 + 0.5s, t_1 + 5s)$
MM1	0.0003	0.0871	0.4679	0.0005	0.0043
MM2	0.8898	0.9973	0.6002	0.9555	0.9896

4.3. The Adaptability of the General Equivalent Model

To verify the adaptability of the proposed model (PM), ten groups of input wind speed and fault voltage dip were randomly generated, which are shown in Figure 4. In each scenario, the wind speeds were assigned to the WTs in WF, and the voltage dip was realized by setting a three-phase symmetrical short-circuit fault at the midpoint of one transmission line.



(**a**) (**b**)

Figure 4. Ten test scenarios. (**a**) Wind speed scenarios. (**b**) Voltage dip scenarios.

Under the operation conditions of Figure 4, the dynamic responses at the POC of the DM, MM1, MM2 and PM were sampled. The active and reactive power errors of different equivalent WF models were calculated according to Equations (4) and (5), respectively, as presented in Tables 2 and 3. According to the values given in Tables 2 and 3, the differences in accuracy of the three equivalent models are illustrated in Figures 5 and 6, respectively.

Table 2. Active power response error of three equivalent models.

Case	$(t_0 - 1s, t_0)$			$(t_0, t_0 + 0.14s)$			$(t_0 + 0.14s, t_1)$			$(t_1, t_1 + 0.5s)$			$(t_1 + 0.5s, t_1 + 5s)$		
	MM1	**MM2**	**PM**	**MM1**	**MM2**	**PM**	**MM1**	**MM2**	**PM**	**MM1**	**MM2**	**PM**	**MM1**	**MM2**	**PM**
1	0.034%	0.033%	0.022%	1.970%	1.602%	1.721%	1.536%	1.359%	1.412%	0.841%	0.750%	0.787%	0.698%	0.649%	0.665%
2	0.383%	0.384%	0.294%	0.701%	0.387%	0.324%	1.102%	0.822%	0.843%	0.446%	0.523%	0.295%	0.068%	0.093%	0.071%
3	0.161%	0.161%	0.080%	1.317%	1.008%	1.068%	3.373%	1.808%	1.734%	1.218%	0.651%	0.674%	1.028%	0.954%	0.988%
4	0.171%	0.170%	0.082%	4.407%	4.667%	4.282%	1.728%	2.062%	1.748%	4.474%	4.375%	4.202%	0.491%	0.427%	0.479%
5	0.362%	0.368%	0.273%	0.428%	0.477%	0.552%	0.401%	0.141%	0.169%	0.608%	0.717%	0.320%	0.142%	0.155%	0.132%
6	0.360%	0.361%	0.451%	1.921%	1.930%	1.799%	1.856%	1.761%	2.012%	1.231%	1.162%	1.597%	1.802%	1.822%	1.787%
7	0.051%	0.048%	0.045%	3.925%	2.108%	2.431%	0.435%	3.014%	0.883%	1.721%	1.256%	1.022%	1.019%	1.013%	0.998%
8	0.166%	0.179%	0.156%	2.057%	2.606%	2.557%	0.650%	4.800%	1.892%	1.512%	1.279%	0.915%	0.473%	0.436%	0.444%
9	0.406%	0.411%	0.316%	2.357%	1.779%	1.730%	0.103%	0.609%	0.358%	1.245%	0.862%	0.632%	0.095%	0.104%	0.102%
10	0.255%	0.255%	0.166%	0.642%	1.106%	1.062%	0.960%	1.218%	0.302%	1.732%	1.714%	1.280%	0.491%	0.572%	0.533%

Table 3. Reactive power response error of three equivalent models.

Case	$(t_0 - 1s, t_0)$			$(t_0, t_0 + 0.14s)$			$(t_0 + 0.14s, t_1)$			$(t_1, t_1 + 0.5s)$			$(t_1 + 0.5s, t_1 + 5s)$		
	MM1	**MM2**	**PM**	**MM1**	**MM2**	**PM**	**MM1**	**MM2**	**PM**	**MM1**	**MM2**	**PM**	**MM1**	**MM2**	**PM**
1	0.0046	0.0060	0.0002	0.0203	0.0210	0.0129	0.0179	0.0138	0.0064	0.0224	0.0178	0.0194	0.0141	0.0113	0.0127
2	0.0076	0.0087	0.0007	0.0192	0.0148	0.0080	0.0467	0.0112	0.0082	0.0649	0.0290	0.0323	0.0101	0.0053	0.0053
3	0.0057	0.0062	0.0005	0.0407	0.0185	0.0068	0.0212	0.0218	0.0045	0.0990	0.0534	0.0570	0.0176	0.0083	0.0086
4	0.0090	0.0091	0.0003	0.0069	0.0074	0.0081	0.0133	0.0141	0.0161	0.0468	0.0454	0.0397	0.0063	0.0065	0.0060
5	0.0079	0.0078	0.0001	0.0155	0.0139	0.0082	0.0048	0.0032	0.0220	0.0360	0.0319	0.0351	0.0064	0.0054	0.0054
6	0.0076	0.0080	0.0004	0.0084	0.0078	0.0074	0.0215	0.0227	0.0197	0.0396	0.0363	0.0365	0.0048	0.0050	0.0047
7	0.0052	0.0059	0.0001	0.0149	0.0082	0.0069	0.0090	0.0220	0.0102	0.0724	0.0347	0.0365	0.0083	0.0048	0.0049
8	0.0056	0.0057	0.0006	0.0384	0.0178	0.0061	0.0305	0.0307	0.0161	0.0888	0.0459	0.0496	0.0148	0.0070	0.0072
9	0.0089	0.0093	0.0010	0.0141	0.0107	0.0058	0.0245	0.0229	0.0200	0.0334	0.0263	0.0270	0.0050	0.0058	0.0052
10	0.0043	0.0053	0.0022	0.0277	0.0166	0.0059	0.0117	0.0115	0.0057	0.0639	0.0392	0.0424	0.0115	0.0065	0.0068

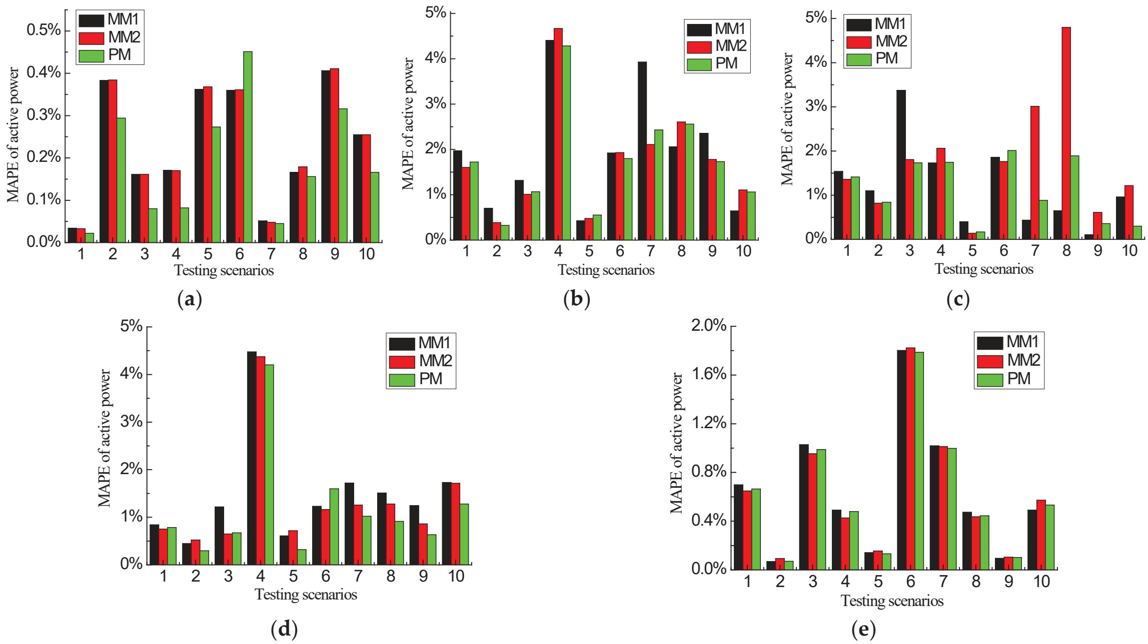


Figure 5. Active power response error of three equivalent models. (**a**) Pre-fault phase. (**b**) Early fault phase. (**c**) Quasi-steady-state phase during the fault. (**d**) Early fault recovery phase. (**e**) Quasi-steady-state phase during the fault recovery.

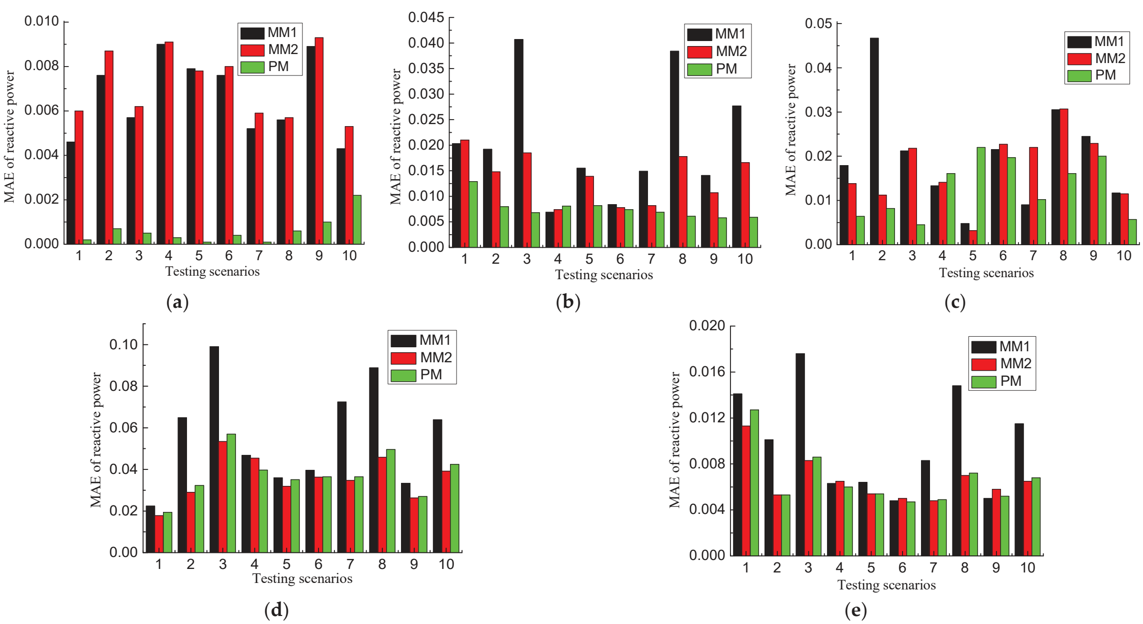


Figure 6. Reactive power response error of three equivalent models. (**a**) Pre-fault phase. (**b**) Early fault phase. (**c**) Quasi-steady-state phase during the fault. (**d**) Early fault recovery phase. (**e**) Quasi-steady-state phase during the fault recovery.

As shown in Table 2 and Figure 5, for the pre-fault phase $(t_0 - 1s, t_0)$, the PM constructed in this paper exhibited higher active power simulation accuracy except in Case 6. In this phase, the active power output of the WF is stable. Moreover, the performance difference of the two mechanism models is relatively fixed for different scenarios. It helps to improve the accuracy of the PM through the optimal configuration of weight coefficients. For the other four phases, the improvement in accuracy decreased slightly, and the accuracy of the PM was between the MM1 and MM2 in some scenarios. The reason lies in that the active power output of WF starts to fluctuate in those phases, and the performance difference of the two mechanism models also begins to change in different scenarios. However, taking the quasi-steady-state phase during the fault $(t_0 + 0.14s,\ t_1)$into account, where the phenomenon that the accuracy of the PM is between the MM1 and MM2 appears more often, the active power errors of each model are presented in Figure 7. It can be seen that the active power performance of the PM is more stable than that of the MM1 and MM2, indicating its better adaptability.

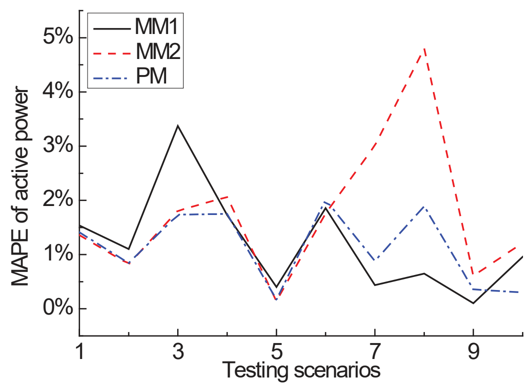


Figure 7. Active power errors of different models in quasi-steady-state process during fault duration.

As shown in Table 3 and Figure 6, for the pre-fault phase $(t_0 - 1s, t_0)$, the PM established in this paper also shows higher reactive power simulation accuracy. For the other four phases, the reactive power performance of the PM is between MM1 and MM2 in some scenarios. However, taking the early fault recovery phase into account, where the above phenomenon appears more often, the reactive power errors of each model are illustrated in Figure 8. It can be seen that the reactive power performance of the PM is also more stable.

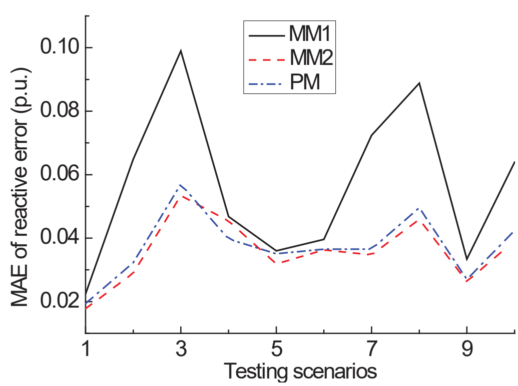


Figure 8. Reactive power errors of different models in quasi-steady-state process during fault duration.

To further demonstrate the effectiveness of the PM, the dynamic responses of the DM, PM, MM1 and MM2 in Cases 6 and 3 are presented in Figures 9 and 10, respectively. As mentioned above, for Case 6, the accuracy of the PM was lower than both the MM1 and MM2. From Figure 9, we can observe that the PM remains accurate and has high correspondence with the electromechanical transient responses of the DM. For the other nine cases, for example, Case 3, it can be seen that the responses of the PM were much closer to that of the DM than the MM1 and MM2, as shown in Figure 10.

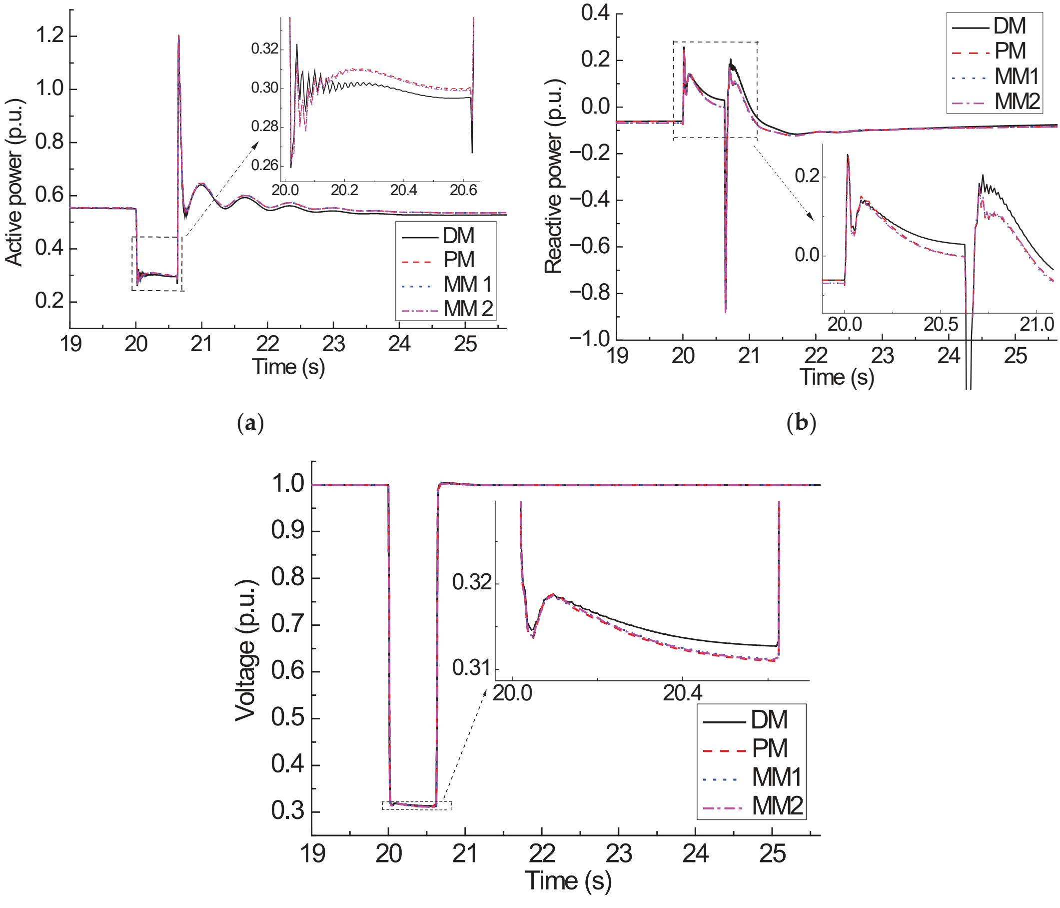


Figure 9. The dynamic responses of WF at POC in Case 6. (**a**) Active power. (**b**) Reactive power. (**c**) Voltage.

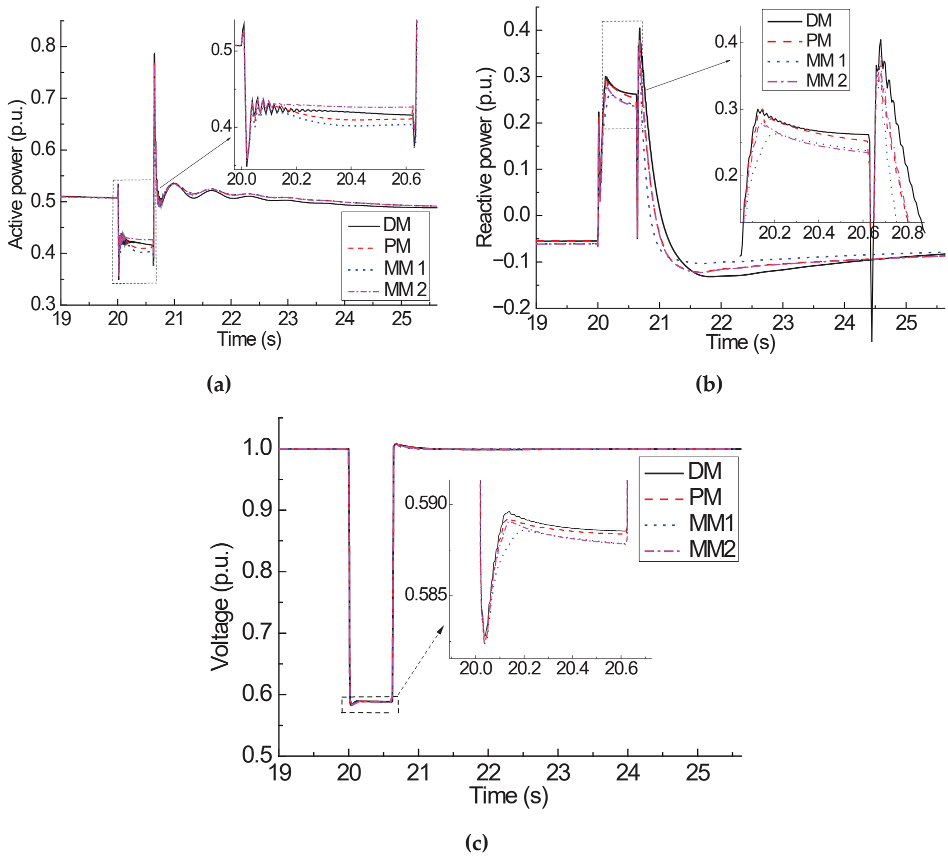


Figure 10. The dynamic responses of WF at POC in Case 3. (**a**) Active power. (**b**) Reactive power. (**c**) Voltage.

5. Discussion

Until now, the general equivalent modeling method for DFIG-based WFs has not been considered from a model-data driven point of view; only a few researchers have analyzed the possibility of identifying the crowbar action based on a data-driven model [31]. This means that the accuracy of WF-equivalent models has been improved, only considering coherent cluster divisions without regard to the adaptability of the equivalent model for different scenarios.

One of the novelties presented in this paper is to enhance the stable performance of WF-equivalent models in uncertain operating scenarios. As the scenario changes, the accuracy of the traditional WF equivalent models will also change, and the performance is often uncertain due to complex influencing factors. As shown in Figures 5 and 6, sometimes the performance of the MM1 is better, and sometimes the performance of the MM2 is better. It is not clear in which scenario the effectiveness of the equivalent model will fail. Therefore, in the view of enhancing the stability of the model performance, we used data-driven mining of complementary characteristics between mechanism models. As shown in Figures 7 and 8, the applicability to uncertain scenarios was significantly improved.

In addition, this paper provides meaningful insights into how to select mechanism models. We analyze the influence of the two calculation methods for the aggregated parameters of the collector cable on the accuracy of the equivalent model, and select two equivalent models with complementary characteristics. This helps to improve the accuracy of the general equivalent model. From Tables 2 and 3, we can observe that except in rare phases, the accuracy of the general equivalent model is smaller than the maximum error between the two mechanism models. Moreover, the average error of the general equivalent model is also smaller than those of the two mechanism models.

The simulations were carried out on a workstation with the following specifications: Intel Xeon(R) Platinum 8375C, 32 CPU @ 2.9 GHz, 128 GB of RAM. The average simulation times of the DM, PM, MM1 and MM2 were 2146 s, 283 s, 135 s and 135 s, respectively. By the PM, the simulation time of the detailed WF can be significantly reduced by about 87 %. Compared to the MM1 and MM2, about two times the simulation time is increased.

6. Conclusions

From the viewpoint of improving the applicability and engineering utility value of the DFIG-based WF equivalent model, this paper proposes a general model-data-driven equivalent modeling method. It verifies the effectiveness of the proposed model for uncertain operation scenarios through simulation.

The two mechanism models are selected based on clustering indicators and calculation methods for the aggregated parameters of the collector cable. The combination of the two models can not only take advantage of the high accuracy of each model, but also take advantage of their complementary nature to reduce simulation errors.

In this paper, the fault process is divided into five stages, and the data-driven adaptive weight optimization model is constructed to optimize the corresponding weight coefficients. It further improves the adaptability of the general WF equivalent model in uncertain scenarios.

However, the main drawback of this general equivalent modeling method is that the accuracy of the PM is sometimes between the MM1 and MM2. In future works, a possible extension could be to apply the time series analysis theory to analyze the fluctuation law of the dynamic responses, and we will further improve the accuracy of the general equivalent model by optimizing the time window division.

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Conflicts of Interest: The authors declare no conflict of interest.

Appendix A

Table A1. Simulation parameters.

Double-fed Asynchronous Wind Turbines	**Wind Turbines**			
	Blade radius /m	31	Shaft system stiffness factor /(pu/rad)	1.11
	Inertia time constant /s	4.32	Rated wind speed /(m/s)	12.5
	Cut-in wind speed /(m/s)	4.5	Cut-out wind speed /(m/s)	22
	Double-fed asynchronous generators			
	Rated power /MW	1.5	Rated frequency /Hz	50
	Rated voltage /kV	0.575	Stator impedance /pu	0.016 + j0.16
	Rotor impedance /pu	0.023 + j0.18	Stator and rotor mutual impedance /pu	j2.9
	Power converters			
	Rated capacity of rotor-side converter /MVA	0.525	Rated capacity of grid-side converter /MVA	0.75
	DC Bus Rated Voltage/kV	1.15	DC side bus capacitance /F	0.01
	Crowbar circuit input threshold /pu	2	Crowbar circuit cut out threshold /pu	0.35
	Crowbar resistance /pu	0.1		
Machine sidetransformer	Rated capacity /MVA	1.75	Rated frequency /Hz	50
	Rated Ratio /kV	25/0.575	Impedance /pu	0.06
Main Transformer	Rated capacity /MVA	150	Rated frequency /Hz	50
	Rated Ratio (kV)	125/25	Impedance /pu	0.135
Cable line	Unit resistance /(Ω/km)	0.1153	Unit inductance (/Ω/km)	j0.3297

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Article

Compressed Air Energy Storage System with Burner and Ejector

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Abstract: The timescale of the energy-release process of an energy storage system has put forward higher requirements with the increasing proportion of new energy power generation in the power grid. In this paper, a new type of compressed-air energy storage system with an ejector and combustor is proposed in order to realize short-timescale and long-timescale energy-release processes under the non-supplementary combustion condition and ejector supplementary combustion condition, respectively. A simulation model of the new system is established in APROS software. The results of this study show that the new system can realize continuous power output when energy storage and energy release operate simultaneously, and especially when the ejector coefficient is 0.8 and burner thermal power is 10 MW, the power-generation time is 12.45 h and the total generated power is 140,052 kW·h, which are 15.6 and 17.5 times greater those of the short-timescale condition, respectively. In summary, the compressed-air energy storage system with an ejector and combustor that is proposed in this paper can flexibly meet the demands of multiple timescales' power generation.

Keywords: compressed-air energy storage; multiple timescales; ejector; burner

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1. Introduction

In recent years, the clean and low-carbon process of energy utilization is accelerating, and the scale of new energy power generation, such as wind power and photovoltaic power, is growing larger and larger in China [1]. According to the data of the China National Energy Bureau, the installed capacity of wind power and solar power was about 330 million kW and 320 million kW, respectively, by the end of February 2022. However, new energy power generation, such as wind power and photovoltaic power, has the characteristics of strong intermittence, volatility, and randomness [2], which may have adverse effects on the power quality, safety, and stability of the power systems [3].

Large-scale electric energy storage technology is one of the effective ways to solve the above problems [4]. It has been validated that equipping a large amount of energy storage systems can effectively stabilize the gap and volatility of new energy power generation [5]. The energy storage systems will release energy to supplement the power generation when new energy power generation is insufficient, which can ensure the power balance, safe, and stable operation of power grid. Additionally, they will store energy to ensure that the new energy is consumed when the new energy generation is surplus. At present, the main types of large-scale clean power energy storage are pumped storage, compressed-air energy storage (CAES) [6], electrochemical energy storage [7], etc., and their typical rated power and maximum output time [8–10] are shown in Table 1. It can be seen that the longest continuous response time (that is, output power time) is 1–26 h of compressed-air energy storage.

Table 1. Typical rated power and maximum output time of various energy storage systems [8–10].

Types	Typical Rated Power	Continuous Response Time
Electrochemical energy storage	1 kW~50 MW	1 min~4 h
Pumped storage	100~2000 MW	2~8 h
Compressed-air energy storage	500 kW~300 MW	1–26 h
Flywheel energy storage	5 kW~5 MW	15 s~15 min
Superconducting magnetic energy storage	0.01~1 MW	ms~15 min
Supercapacitor	0.01~1 MW	1 s~15 min

The CAES system has the advantages of a large capacity, low pollution, moment of inertia, long storage cycle, etc. [11], and scholars and researchers have been widely concerned with it because of its broad development prospects. [12]. Since Stal Laval proposed to use underground caves to realize compressed-air energy storage in 1949 [13], domestic and foreign scholars have carried out a lot of research and practice [14] and established a number of CAES commercial power stations and demonstration projects [15]. At present, there are many types of CAES systems, which can be divided into supplementary combustion and non-combustion CAES systems from the perspective of auxiliary combustion [16]. The supplementary combustion CAES system has the advantages of strong reliability, stability, and good flexibility, but the disadvantage is obviously, e.g., consumption of fossil energy and emission of greenhouse gas. Two large-scale CAES power plants have been put into commercial operation in the world, namely the Huntorf power plant in Germany and the McIntosh power plant in the United States [17], both of which are supplementary combustion CAES systems. Compared with the supplementary combustion CAES system, the non-combustion CAES system has no supplementary combustion process, and the heat required in the energy-release stage mainly comes from the compression heat generated by the air-compression process in the energy-storage stage, so it has advantages of being environmentally friendly and pollution free [18,19]. At present, all the CAES demonstration projects in China adopt non-combustion CAES. The advanced adiabatic compressed-air energy storage (AA-CAES) power station of 50 MW in Feicheng, Shandong Province, was put into commercial operation in September 2021, and it is the first CAES commercial power station in China [20]. In the same month, the national demonstration project of 60 MW salt-cave CAES power generation system in Jintan, Jiangsu Province, successfully implemented a grid connection test [21].

As shown in Figure 1, the AA-CAES system is a typical non-combustion CAES system, and it includes two stages, namely energy storage and energy release [22]. In the energy-storage stage, air is compressed by compressors and stored in a storage tank. In the energy-release stage, compressed air enters a turbine from the gas storage tank, expands and releases energy to drive the synchronous generator to generate electricity, and then exhaust air is discharged into atmosphere [23].

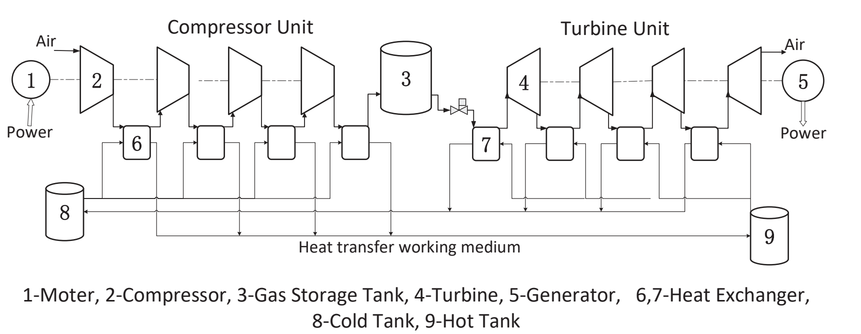


Figure 1. Advanced adiabatic compressed-air energy storage system.

At present, the research on the application of CAES system mainly focuses on two aspects, one is to participate in the source network coupling of grid load frequency regulation and improve the reliability of power supply on the grid side. Kamyar et al. [24] proposed a new design method of CAES system based on the performance requirements of the Ontario power grid by analyzing the actual operation data of a whole year. Based on the characteristics of the CAES system, Wen et al. [25] constructed its primary frequency modulation function and analyzed and set its dead zone, governor droop, clearance, and other parameters, which laid a foundation for CAES to participate in primary frequency modulation of power grid. Yang et al. [26] put forward a load rejection control strategy for an AA-CAES expansion generator by adding a shutoff valve between adjacent expansion units to effectively prevent speed rise. AmirReza. et al. [27] proposed a cogeneration system composed of CAES, organic Rankine cycle, and absorption–compression refrigeration cycle. Considering the multi-generation characteristics of the AA-CAES power plant, Li et al. [28] constructed the joint dispatching constraint model of cooling, heating, and power multi-energy flow for the AA-CAES power station. Hesamoddin et al. [29] proposed a two-stage mathematical optimization model for optimizing the day-ahead operation of generation units, as well as CAESs, in energy and reserve markets in a stochastic way.

The other aspect is to couple with the source storage of new energy power stations, such as wind energy and photovoltaic, on the generation side. Deng et al. [30] proposed a control strategy of a wind-storage combined system on the basis of power stabilization and verified its feasibility by simulation. Amirreza et al. [31] researched an absorption-recompression refrigeration system with CAES and wind turbines, which employed a CAES system to compensate the further energy consumption of the vapor compressor. Alirahmi et al. [32] proposed and thoroughly investigated a novel efficient and environmentally friendly hybrid energy production/storage system comprising a compressed-air energy storage, and the system has an exergy round trip efficiency of 60.4% and a total cost rate of 117.5 $/GJ. Li et al. [33] studied the operation optimization strategy of the wind-storage combined system, considering the dynamic characteristics and operation constraints of CAES. Li et al. [34] studied a grid-connected power-optimization strategy that integrates wind energy and a low-temperature CAES, which can balance the fluctuation of wind energy by reducing energy storage capacity and ensure continuous and stable output power to power grid.

As mentioned above, the current research studies of the AA-CAES system are all based on a single timescale operation, which can only meet the requirements of storing energy when new energy power generation is surplus and releasing energy in a short timescale when new energy power generation is not enough. Nevertheless, the long-timescale fluctuation of new energy and the extreme conditions of zero power generation have a growing impact on the power system with the rapid development of new energy and the increasing proportion of installed capacity. Although increasing the capacity of energy storage system is an important method to effectively solve this problem, it will lead to high system cost and low utilization.

In consideration of the demands of multiple timescales' power generation and economy, a novel AA-CAES system with two main components, namely an ejector and a burner, is proposed in this paper. This system has three operating modes, including a short-timescale mode with adiabatic non-supplementary combustion condition, a long-timescale mode with ejector-supplementary combustion condition, and a continuous-output mode with energy storage and energy release operating simultaneously, which can flexibly adapt to the requirements of multiple timescales' energy release. In addition, a simulation model of the new system was established based on APROS software, and the characteristics of the three operating modes are analyzed.

2. Architecture of Energy Storage System

2.1. Compressed-Air Energy Storage System with Ejector and Combustor

Based on the AA-CAES system, this paper proposes a new CAES system with an ejector and combustor which can release energy in multi-timescales. Its system architecture is shown in Figure 2. The system is composed of a compression energy storage subsystem, a gas storage subsystem, and an expansion energy-release subsystem. Among them, the compression energy storage subsystem is composed of multistage compressors and motors; the gas storage subsystem is a high-pressure gas storage tank; and the expansion energy-release subsystem is composed of an ejector, a combustor, multistage expanders, and a generator.

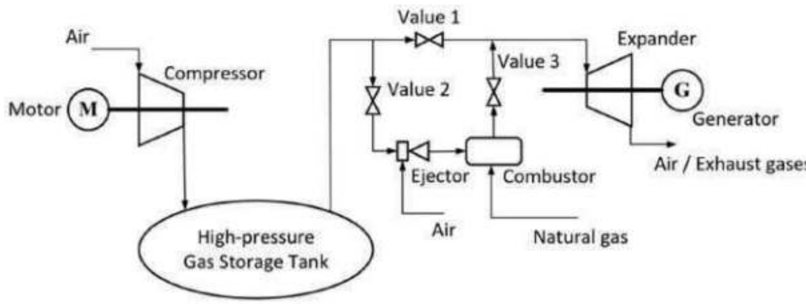


Figure 2. Diagram of compressed-air energy storage system with an ejector and combustor.

In this study, a simulation model of the new CAES system with an ejector and burner is established based on APROS software [35,36]. Considering the characteristics of the CAES system with supplemental combustion and the relevant operating parameters of a 10 MW advanced CAES system demonstration project in Bijie, China [36,37], the detailed design parameters [19,36] are shown in Table 2.

Table 2. Basic design parameters for the new CAES system.

Parameters	Units	Value
Energy-Release Power	MW	10
Energy-Release Pressure	MPa	7
Maximum Storage Pressure	MPa	10
Gas Storage Tank Volume	m^3	6000
Ambient Pressure	MPa	0.1
Ambient Temperature	K	298
10 MW Release Time	s	2880
Hot Tank Temperature	K	403
Hot Tank Pressure	MPa	0.4
Cold Tank Temperature	K	298
Cold Tank Pressure	MPa	0.1
Compressor Motor Power	MW	10
Burner Thermal Power	MW	20
Burner Fuel	/	Natural Gas
Fuel Calorific Value	MJ/Nm^3	36.22
Electric Power Consumed by Other Auxiliary Machines of Energy Storage Subsystem	MW	1.5
Electric Power Consumed by Other Auxiliary Machines of Energy-Release Subsystem	MW	1.5
Generator Power	MW	30

In this paper, a short-timescale simulation method is adopted, and the operation parameters of each expander are shown in Table 3.

Table 3. Operation parameters in the mode of short timescale.

Expander Stages	Unit	1	2	3	4
Inlet Pressure	MPa	7	2.46	0.853	0.283
Outlet Pressure	MPa	2.48	0.873	0.303	0.1
Inlet Temperature	°C	82	82	82	82
Outlet Temperature	°C	14.4	7.6	10.0	18.0
Pressure Loss in Heat Exchanger	MPa	0.02	0.02	0.02	0.02
Air Flow Rate	kg/s	34.7	34.7	34.7	34.7
Expansion Ratio	/	2.82	2.82	2.82	2.82
Isentropic Efficiency	/	0.88	0.88	0.88	0.88
Output power	MW	2.536	2.786	2.691	2.358
speed	r/min	3 000	3 000	3 000	3 000

2.2. Mathematic Model

2.2.1. Expander

The air-expansion process is generally regarded as a polytropic process in the expander, and the output power of the expander [38,39] is calculated as follows:

$$W_i = \frac{k}{k-1} m R_{\mathrm{g}} T_i^{\mathrm{in}} \eta_i \left(1 - \beta_i^{\frac{1-k}{k}}\right), i = 1, 2, \cdots M \quad (1)$$

where W_i is the output power of the ith expanders, kW; m is the air mass flow, kg/s; k is the air adiabatic index; R_{g} is the gas constant; T_i^{in} is the intake temperature of the ith stage expander, K; η_i is the adiabatic efficiency of the ith stage expander; and β_i is the expansion ratio of the ith stage expander.

2.2.2. Heat Exchanger

In the AA-CAES power generation system, heat exchangers and expanders are arranged in a series. After air flows into the heat exchanger, the heat exchange between the air and tube-wall [36] is as follows:

$$Q_{\mathrm{h}} = A_{\mathrm{h}}(T_{\mathrm{h}} - T_{\mathrm{w}})/(\delta/2K_{\mathrm{w}} + 1/\alpha_{\mathrm{h}}) \quad (2)$$

The heat exchange between heat-transfer medium and pipe-wall is as follows:

$$Q_{\mathrm{c}} = A_{\mathrm{c}}(T_{\mathrm{c}} - T_{\mathrm{w}})/(\delta/2K_{\mathrm{w}} + 1/\alpha_{\mathrm{c}}) \quad (3)$$

where δ is the thickness of the tube-wall, m; T_{w} is the average temperature of the tube-wall, K; T_{h} is the air temperature inside the tube-wall, K; T_{c} is the average temperature of the heat-transfer medium, K; K_{w} is the thermal conductivity of the tube-wall, W/(m·K); α_{c} and α_{h} are the convective heat-transfer coefficients of the inner and outer tube-walls, respectively, $\mathrm{W}/\left(\mathrm{m}^2\cdot\mathrm{K}\right)$; and A_{c} and A_{h} are the areas of the inner and outer tube-walls, respectively, m^2.

2.2.3. Gas Storage Tank

In the energy-release stage, the discharge process of the gas storage tank is a polytropic process. According to the mass and energy balance equations, the change rules of gas in the gas storage tank during the discharge process [40] can be obtained as follows:

$$\frac{\mathrm{d}p}{\mathrm{d}t} = \frac{R_{\mathrm{g}}}{V c_{\mathrm{v}}}\left[\frac{\mathrm{d}m}{\mathrm{d}t} c_{\mathrm{p}} T_{\mathrm{ac}} - h_{\mathrm{a}} A (T_{\mathrm{ac}} - T_{\mathrm{a}})\right], \quad (4)$$

$$\frac{\mathrm{d}T_{\mathrm{ac}}}{\mathrm{d}t} = \frac{T_{\mathrm{ac}}}{m}(k-1)\frac{\mathrm{d}m}{\mathrm{d}t} - \frac{1}{m c_{\mathrm{v}}} h_{\mathrm{a}} A (T_{\mathrm{ac}} - T_{\mathrm{a}}), \quad (5)$$

where p is the air pressure of the gas storage tank, Pa; V is the volume of the gas storage tank, m^3; c_p is the air-specific heat at constant pressure, kJ/kg·K; c_v is the air-specific heat at constant volume, $kJ/(kg \cdot K)$; h_a is the convective heat transfer coefficient of the gas storage tank, $kW/(m^2 \cdot K)$; A is the internal surface area of the gas storage tank, m^2; m is the air mass in the gas storage tank, kg; T_a is the wall temperature of the gas storage tank, K; and T_{ac} is the air temperature in the gas storage tank, K.

2.2.4. Regulating Valve

The flow equation of the regulating valve [41] is as follows:

$$m_s = \varepsilon C_s f(\mu) \sqrt{\rho \Delta p}, \tag{6}$$

where m_s is the outlet air flow rate of regulating valve, kg/s; ε is the fluid compressibility, C_s is the valve admittance, $f(\mu)$ is the characteristic function of the regulating valve, and ρ is the inlet air density of the regulating valve.

3. Energy Storage Condition

The compression energy-storage subsystem of CAES system has the same system architecture and operation mode under the conditions of energy release at different timescales.

In the stage of energy storage, the air is compressed into high-pressure compressed air by the compressor and stored in the high-pressure gas storage tank to realize the conversion of electric energy to internal energy. The judgment conditions for the completion of compressed-air preparation and the end of the energy-storage phase are as follows:

$$\left\{ \begin{array}{c} p = p_e \\ t = t_s \end{array} \right\}, \tag{7}$$

where p is the pressure of high-pressure gas storage tank, p_e is the rated pressure of gas storage tank, t is the operation time, and t_s is the end time of energy storage stage.

The end time t_s of energy storage stage is dispatched by the power grid, which is determined according to the peak regulation demand of power grid and dispatched through the day-ahead scheduling plan.

The power consumption of the ith compressor is defined as follows:

$$P_{ci} = G_{ci}(h_{ci,\text{out}} - h_{ci,\text{in}}) \tag{8}$$

where P_{ci} is the power consumption of the ith compressor, G_{ci} is the air mass flow of the ith compressor during energy storage process, and $h_{ci,\text{in}}$ and $h_{ci,\text{out}}$ are the specific enthalpies of inlet and outlet of the ith compressor.

The total power consumption of compressor motor can be expressed as follows:

$$P_m = \frac{P_c}{\eta_m \times \eta_c} = \frac{G_c \times \sum_{i=1}^{m} (h_{ci,\text{out}} - h_{ci,\text{in}})}{\eta_m \times \eta_c} \tag{9}$$

where P_m is the power consumption of compressor motor, P_c is the total power consumption of compressor motor, G_c is the air mass flow of compressor, η_m is the motor efficiency, η_c is the compressor efficiency, and m is the number of compressors.

4. Multi-Timescale Energy-Release Condition

In this paper, the short-timescale power-generation condition is defined as a power-generation time less than 6 h, the long-term scale power-generation condition is defined as a power-generation time between 6 and 30 h, and the continuous-output condition is defined as continuous uninterrupted power generation.

The compressed-air energy storage system with an ejector and combustor has three operation modes in the energy-release stage that can flexibly adapt to three power-generation conditions, namely short timescale, long timescale, and continuous output. The three operating conditions of the energy-release stage of the system are discussed below.

4.1. Short-Timescale Condition

Under the short-timescale condition, the CAES system operates in adiabatic non-supplementary combustion mode.

4.1.1. Energy-Release Stage

The CAES system starts the process of air expansion to generate electricity when electric energy in the power grid is in short supply. As shown in Figure 1, the compressed air enters the expander from the high-pressure gas storage tank to expand and drive the generator to generate electricity by opening Valve 1, closing Valve 2 and Valve 3, sequentially. The compressed air after the work of the upper stage enters the heat exchanger for heating, and then it enters the expander of the next stage for power generation and is discharged into atmosphere after multistage expansion.

The generating power, P_{ti}, of the ith expander is defined as follows:

$$P_{ti} = \frac{G_{ti} \times (h_{ti,in} - h_{ti,out})}{3600} \tag{10}$$

where G_{ti} is the air mass flow of the ith expander, and $h_{ti,in}$ and $h_{ti,out}$ are the specific enthalpies of inlet and outlet of the ith expander.

The total generating power, P_g, of the CAES system can be expressed as Equation (11):

$$P_g = P_t \times \eta_g \times \eta_t = G_g \sum_{i=1}^{m} (h_{ti,in} - h_{ti,out}) \times \eta_g \times \eta_t \tag{11}$$

where P_t is the total generating power of expanders, G_g is the air mass flow of expanders in short scale, η_g is the generator efficiency, η_t is the expander efficiency, and m is the number of expanders.

4.1.2. Timescale Determination

The determination of timescale is mainly related to parameters such as the volume, temperature, and pressure of the high-pressure gas storage tank.

To simplify the calculation, the following assumptions need to be made in this paper:

(1) The compressed air is an ideal gas;
(2) The gas loss at the pipes and valves are ignored;
(3) The temperature change in the process of compression and depressurization of gas storage device is not considered. (The energy storage system has a heat-exchange device, so the internal temperature change of gas storage device is ignored and regarded as an isothermal expansion process.)

According to the ideal gas law, the relationship between the state parameters of gas storage device at the initial time in the energy storage stage is as follows:

$$p_{c,s} V = m_{c,s} R T_{c,s} \tag{12}$$

where $p_{c,s}$ is the air pressure at the initial time, V is the volume of gas storage device, $m_{c,s}$ is the air mass at the initial time, and $T_{c,s}$ is the air temperature at the initial time.

The state parameters of gas storage device at the end time can be expressed as follows:

$$p_{c,f} V = m_{c,f} R T_{c,f} \tag{13}$$

where $p_{c,f}$ is the air pressure at the end time, $m_{c,f}$ is the air mass at the end time, and $T_{c,f}$ is the air temperature at the end time and is equal to $T_{c,s}$.

Equations (12) and (13) can be subtracted to obtain Equation (14):

$$V(p_{c,f} - p_{c,s}) = \Delta m R T_{c,s} \tag{14}$$

where Δm is the total mass of compressed air produced by compressor during energy storage process.

According to the total mass of compressed air and the mass flow of compressor under the rated working conditions, the time, t_c, of the energy storage process can be expressed as follows:

$$t_c = \frac{V(p_{c,f} - p_{c,s})}{RT_{c,s}G_c}, \tag{15}$$

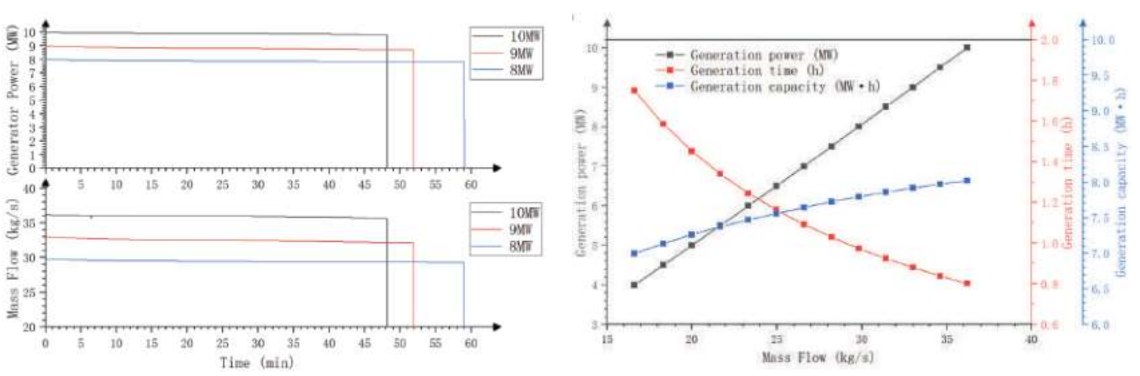
Similarly, we derive the working time, t_g, of the energy-release process as follows:

$$t_g = \frac{V\left(p_{g,s} - p_{g,f}\right)}{RT_{g,s}G_g}, \tag{16}$$

where $p_{g,s}$ is the air pressure at the initial time, $p_{g,f}$ is the air pressure at the end time, and $T_{c,s}$ is the air temperature at the initial time.

4.1.3. Simulation Analysis

As shown in Figure 3a, with the mass flow of compressed air decreased, the generator power decreases, while the working time of the energy-release process, t_g, is increased. Moreover, the relations of different mass flows of compressed air with the generator power, power-generation capacity, and power-generation time are shown in Figure 3b. The compressed air flow is inversely proportional to the power-generation time. Because the total quality of the workable working medium in the gas storage tank is constant, the power-generation capacity tends to decrease with the decrease of the compressed air flow.



(**a**) (**b**)

Figure 3. (**a**) Operating parameters under short-timescale conditions. (**b**) Relation diagram of different mass flow of compressed air with power-generation capacity and power-generation time.

4.2. Long-Timescale Condition

The working condition of the power consumption and energy-storage stage in the long timescale is the same as that in the short timescale.

4.2.1. Energy-Release Stage

As shown in Figure 2, the ambient air is inhaled into the ejector to increase the flow when the high-pressure compressed air passes through the injector by closing Valve 1 and opening Valve 2 and Valve 3, sequentially. By adding natural gas into the combustor for mixed combustion, the temperature, pressure, and expansion work capacity of the compressed air are increased. Finally, the pressurized and heated air enters the turbine and drives the generator to generate electricity, realizing the conversion of internal energy to electrical energy.

This model improves the power-generation capacity through two means. One is to increase the air-mass flow of the expander through the ejector, and the other is to increase the temperature and pressure of the working air through the external combustor. Air enthalpy is positively correlated with air temperature and pressure; that is, raising the enthalpy of working air inlet could increase the enthalpy drop of air inlet and outlet of expander.

4.2.2. Ejector Model

The structure diagram of the ejector is shown in Figure 4.

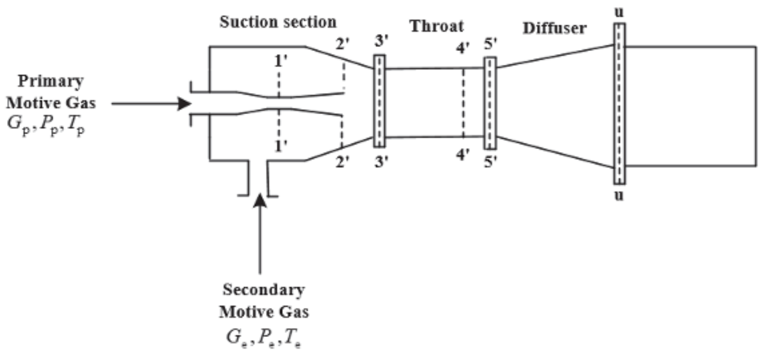


Figure 4. Structure diagram of the ejector.

According to the mass-balance equations, the total mass balance [42] is calculated as follows:

$$G'_g + G_e = G_t \tag{17}$$

The suction coefficient of the ejector is defined [42] as follows:

$$\gamma = \frac{G_e}{G'_g} \tag{18}$$

where G'_g is the mass flow rate of compressed air, and G_e is the mass flow rate of atmosphere air.

Therefore, after high-pressure compressed air passes through the ejector, its mass flow rate changes as follows:

$$G_t = G'_g(1 + \gamma) \tag{19}$$

4.2.3. Combustor Model

The structure diagram of the combustor is shown in Figure 5.

In this paper, the new system uses natural gas as the fuel of burner, and its calorific value is 36.22 MJ/Nm^3. The energy changes after burning with natural gas are as follows [43]:

$$Q_b = Q_{Fuel} \times \eta_b \tag{20}$$

where Q_b is the thermal power output of the combustor, Q_{Fuel} is the thermal power of burning natural gas, and η_b is the conversion efficiency of the combustor.

Therefore, under the long-timescale condition, the generating power of the energy-release generation stage is as follows:

$$P'_{\mathrm{g}} = \frac{\left(G'_{\mathrm{g}}(1+\gamma) \times \sum_{i=1}^{m} \left(h'_{\mathrm{ti,in}} - h'_{\mathrm{ti,out}}\right) + Q_b\right)}{3600} \times \eta_{\mathrm{g}} \times \eta_{\mathrm{t}} \tag{21}$$

where P'_{g} is the generating power of the energy-release generation stage.

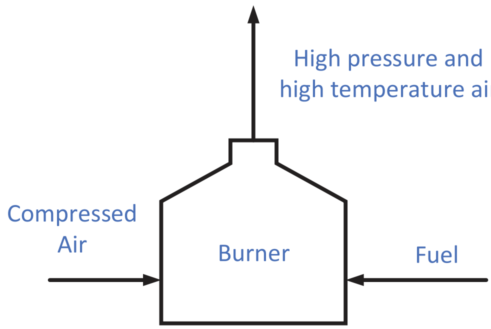


Figure 5. Structure diagram of combustor.

4.2.4. Timescale Determination

There are two aspects to prolonging the energy-release time in this mode: one is to increase the working air flow rate from G'_{g} to $G'_{\mathrm{g}}(1+\gamma)$ through the ejector, and the other is to increase its capacity for work through combustion. When the output power is constant at the rated power of the expansion generator, reducing the compressed-air consumption per unit output power can increase the time of the power generation:

$$P'_{\mathrm{g}} = P_{\mathrm{g}} \tag{22}$$

$$G'_{\mathrm{g}} = \frac{G_{\mathrm{g}} \sum_{i=1}^{m} (h_{\mathrm{ti,in}} - h_{\mathrm{ti,out}}) - Q_b}{(1+\gamma) \times \sum_{i=1}^{m} \left(h'_{\mathrm{ti,in}} - h'_{\mathrm{ti,out}}\right)} \tag{23}$$

Compressed-air energy that can perform expansion work in the high-pressure gas storage tank is constant, so we can obtain the following:

$$G'_{\mathrm{g}} t'_{\mathrm{g}} = G_{\mathrm{g}} t_{\mathrm{g}} \tag{24}$$

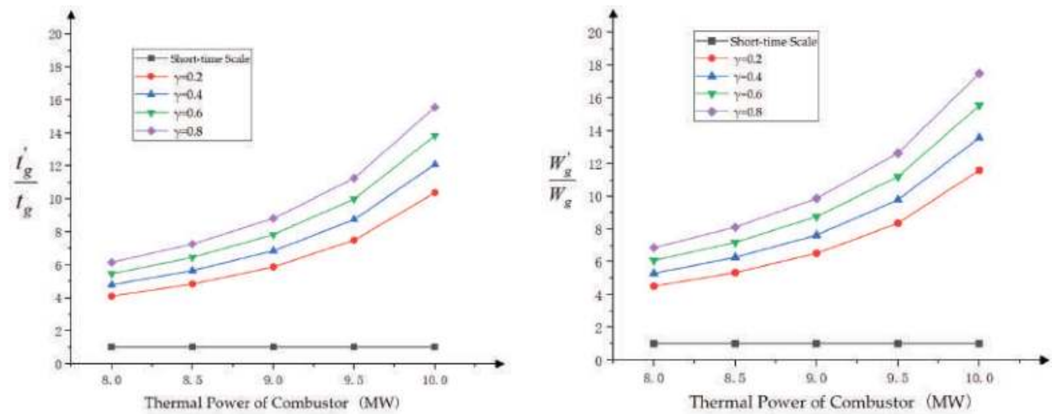
$$t'_{\mathrm{g}} = G_{\mathrm{g}} \frac{(1+\gamma) \times \sum_{i=1}^{m} \left(h'_{\mathrm{ti,in}} - h'_{\mathrm{ti,out}}\right)}{G_{\mathrm{g}} \sum_{i=1}^{m} (h_{\mathrm{ti,in}} - h_{\mathrm{ti,out}}) - Q_b} t_{\mathrm{g}} \tag{25}$$

$$W'_{\mathrm{g}} = (1+\gamma) \frac{h_{\mathrm{J,in}}}{h_{\mathrm{J,mix}}} W_{\mathrm{g}} + Q_{\mathrm{b}} t'_{\mathrm{g}} \tag{26}$$

where t'_{g} is the working time of the energy-release process under the long-timescale condition, and W'_{g} is the generation capacity during the energy-release process under the long-timescale condition.

4.2.5. Simulation Analysis

From Equations (25) and (26), it can be seen that the power-generation capacity and power-generation time in a long timescale are affected by the ejection coefficient of the ejector and the thermal power of the burner, and their relationships are shown in Figure 6a,b. It can be seen from Figure 6a,b that, when the ejector coefficient is 0.2 and the burner thermal power is 8 MW, the power-generation time is 3.28 h and the power-generation capacity is 36,114.3 kW·h, which are 4.1 and 4.5 times that of the short-timescale conditions, respectively. When the ejector coefficient and burner thermal power are increased to 0.8 and 10 MW, the power-generation time and power-generation capacity increase by 8.35 h and 103,937.7 kW·h, which are 15.6 and 17.5 times that of those of the short-timescale conditions, respectively.



(**a**) (**b**)

Figure 6. Relation diagram of different injection coefficients and burner thermal power with power-generation capacity and power-generation time: (**a**) power-generation time and (**b**) power-generation capacity.

Under the condition of the long timescale, increasing the ejection coefficient can increase the power-generation time and power-generation capacity of the system. When the compressed-air mass available in the gas storage tank and the flow rate after passing through the jet are constant, increasing the ejection coefficient will make the air flow into the jet decrease and prolong the power-generation time. The power-generation time and power-generation capacity will increase with the increase of the burner thermal power. This is because increasing the burner thermal power will lead to the specific enthalpy of the intake air of the first-stage expander increasing, thus reducing the compressed air flow required to achieve the rated power. Meanwhile, compared with the ejector coefficient, the change of burner thermal power has a more obvious influence on the power-generation time and power-generation capacity.

4.3. Continuous-Output Condition

The continuous-output condition is that the energy-storage stage and energy-release stage are carried out simultaneously. The energy-storage stage can provide high-pressure compressed air for the energy-release stage, while the generator provides electricity energy for the energy-storage subsystem. The power-generation capacity of the whole system is improved by increasing the air mass flow of the expander through the ejector and increasing the air enthalpy at the inlet of the expander by burning natural gas in the external burner.

4.3.1. Constraint Conditions

The construction of continuous-output conditions needs to meet the following constraints:

$$\begin{cases} P_g'' > P_u \\ G_g'' < G_c \\ p_p > p_{g,\lim} \end{cases} \tag{27}$$

where P_g'' is the generating power of the energy-release subsystem, P_u is the system power consumption, G_g'' is the air mass flow of compressed air into the jet, G_c is the air mass flow at the outlet of the last-stage compressor, P_p is the pressure of high-pressure gas storage tank, and $P_{g,lim}$ is the minimum working pressure of the ejector.

Through the above constraints, it is ensured that the mass flow rate of the working air consumed by the ejector is less than or equal to the mass flow rate at the outlet of the last-stage compressor, so that the system has working air that can meet the continuous operating conditions. Meanwhile, it ensures that the ejector can work normally and prevents the ejector from not forming an entrainment effect due to a too-low working pressure.

The generating power of the energy-release subsystem is the same as that under the long-timescale condition:

$$P''_g = P'_g = \frac{\left(G''_g(1+\gamma) \times \sum_{i=1}^{m}\left(h''_{ti,in} - h''_{ti,out}\right) + Q_b\right)}{3600} \times \eta_g \times \eta_t \quad (28)$$

The power consumption of the whole system is as follows:

$$P_u = P_m + \sum P_{o,m} + \sum P_{o,g} \quad (29)$$

where $P_{o,m}$ is the electric power consumed by other auxiliaries in the energy-storage subsystem, and $P_{o,g}$ is the electric power consumed by other auxiliaries in the energy-release subsystem.

The system's generating power is as follows:

$$P_s = P''_g - P_u \quad (30)$$

where P_u is the generating power of the whole system.

4.3.2. Simulation Analysis

Under continuous-output conditions, the compression energy-storage subsystem, gas-storage subsystem, and expansion energy-release subsystem are all put into operation, and their operating parameters are shown in Figure 7. During the duration of 0–12.45 h, the system operates under the long-timescale condition, and the system's output power is 27 MW. When the air pressure at the inlet of the ejector drops to 3 MPa, the compression energy-storage subsystem is put into operation, and gas storage tank's pressure rises. The system operates under continuous-output conditions, and the compressor's motor power is 10 MW, so that the system's output power drops to 17 MW. When the gas storage tank's pressure rises to 10 MPa again, the compression energy-storage subsystem stops running.

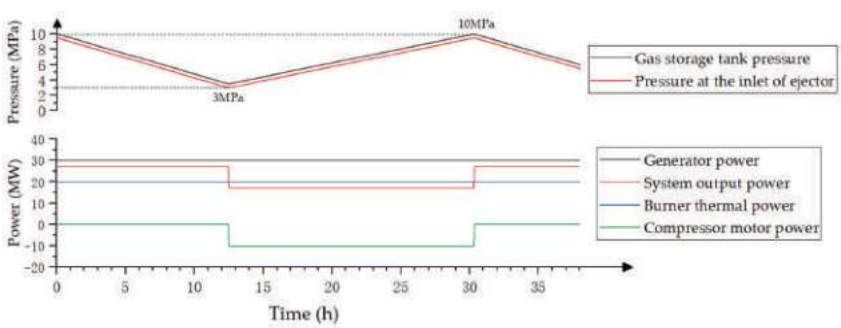


Figure 7. Operating parameters under continuous-output conditions.

5. Conclusions

The current literature has identified the CAES as a potentially important part of coupling renewable energy generation and low-carbon power grids. The extreme conditions of the long timescale's fluctuation and zero-power generation of the new energy power generation require that the CAES system has the ability to generate electricity on multiple timescales. The existing commercial CAES systems cannot meet the needs of multi-timescale power generation well compared to the system studied in this paper. The AA-CAES system can only meet the needs of short-timescale power generation. The supplementary combustion CAES system can meet the needs of short-timescale and long-timescale power generation, but it will cause pollution problems in the short-timescale power generation due to supplementary combustion compared to the system studied here.

In this paper, a new type of compressed-air energy-storage system with an injector and burner was established. The simulation analysis was carried out under short-timescale, long-timescale, and continuous-output conditions, sequentially. Several conclusions were obtained:

(1) The new system can meet the needs of multiple timescales under different operating conditions. The energy-storage stage of the new system is consistent with that of the AA-CAES system, and there are three operation modes in the energy-release stage which can flexibly adapt to three power-generation conditions: short timescale, long timescale, and continuous output.
(2) Under the short-timescale condition, the ejector and combustor are not put into operation, and the CAES system operates as adiabatic compression/expansion processes, which do not need to burn natural gas and are environmentally friendly.
(3) Under medium- and long-timescale conditions, the ejector and combustor are put into operation to prolong the duration of the power generation, so that the new system can meet the needs of long-timescale power generation. Moreover, the power-generation duration and capacity of the system improve with the increase of the ejector coefficient and burner thermal power. When the ejector coefficient is 0.8 and the burner thermal power is 10 MW, the power-generation time is 12.45 h, and the power-generation capacity is 140,052 kW·h, which are 15.6 and 17.5 times that of the short-timescale conditions, respectively.
(4) Under the continuous-output condition, the energy-storage system and energy-release system operate at the same time. The energy-storage subsystem provides high-pressure compressed air for the energy-release stage, while the generator provides electricity for the energy-storage subsystem. By selecting equipment parameters according to the constraint conditions, the new system can realize continuous power output.

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Nomenclature

AA-CAES	Advanced adiabatic compressed-air energy storage
CAES	Compressed-air energy storage
A	Internal surface area of gas storage tank $[\mathrm{m}^2]$
$A_\mathrm{c}, A_\mathrm{h}$	Area of inner and outer pipe-wall $[\mathrm{m}^2]$
c_p	Air specific heat at constant pressure $[\,\mathrm{kJ}/(\mathrm{kg}\cdot\mathrm{K})]$
c_v	Air specific heat at constant volume $[\mathrm{kJ}/(\mathrm{kg}\cdot\mathrm{K})$
C_s	Valve admittance

$f(\mu)$	Characteristic function of regulating valve
G_c	Air mass flow of compressors [kg/s]
G_g	Air mass flow of expander under the short-timescale condition [kg/s]
G'_g	Mass flow of compressed air entering the jet under the long-timescale condition [kg/s]
G''_g	Mass flow of compressed air entering the jet under continuous-output conditions [kg/s]
G_c	Air mass flow at the outlet of last-stage compressor [kg/s]
h_a	Convective heat transfer coefficient of gas storage tank $\left[kW/\left(m^2 \cdot K\right)\right]$
$h_{ci,in}, h_{ci,out}$	Specific enthalpy of inlet and outlet of stage i compressor [kJ/kg]
$h_{ti,in}, h_{ti,out}$	Specific enthalpy of inlet and outlet of stage i expander under short-timescale condition [kJ/kg]
$h'_{ti,in}, h'_{ti,out}$	Specific enthalpy of inlet and outlet of stage i expander under long-timescale condition [kJ/kg]
$h''_{ti,in}$, $h''_{ti,out}$	Specific enthalpy of inlet and outlet of stage i expander under continuous-output condition [kJ/kg]
$h_{J,in}$	Specific enthalpy of compressed air entering the jet under long-timescale condition [kJ/kg]
$h_{J,mix}$	Specific enthalpy of mixed air outflow from the jet in under long-timescale condition [kJ/kg]
k	Air adiabatic index
K_w	Thermal conductivity of pipe-wall [W/(m·K)]
m	Air mass flow of expanders [kg/s]
$m_{c,s}$	Air mass at the initial time [kg]
$m_{c,f}$	Air mass at the end time [kg]
p_e	Rated pressure of gas storage tank [Pa]
P_m	Power consumption of compressor motor
P_g	Generating power of expanders under short-timescale condition
$p_{c,s}$	Air pressure at the initial time [Pa]
$p_{c,f}$	Air pressure at the end time [Pa]
P'_g	Generating power under long-timescale condition [kW]
P''_g	Generating power in continuous-output condition [kW]
P_u	System power consumption [kW]
P_s	System generating power [kW]
P_p	Air pressure of high-pressure gas storage tank [Pa]
$P_{g,lim}$	Minimum working pressure of ejector [Pa]
Q_h	Heat exchange between air and pipe-wall [kJ]
Q_c	Heat exchange between heat − transfer medium and pipe-wall [kJ]
Q_b	Thermal power output of combustor [kW]
Q_{Fuel}	Thermal power of burning natural gas [kW]
R_g	Gas constant [kJ/(kg·K)]
T_i^{in}	Intake temperature of i stage expander [K]
T_h	Air temperature inside pipe-wall [K]
T_w	Average temperature of pipe-wall [K]
T_c	Average temperature of heat-transfer medium [K]
T_a	Wall temperature of gas storage tank [K]
T_{ac}	Air temperature in gas storage tank [K]
t_s	End time of energy storage stage
$T_{c,s}$	Air temperature at the initial time [K]
$T_{c,f}$	Air temperature at the end time [K]
t_c	Working time of energy storage process
t_g	Working time of energy-release process under short-timescale condition
t'_g	Working time of energy-release process under long-timescale condition
V	Gas storage tank volume $\left[m^3\right]$

W_i	Output power of expanders [kW]
W_g	Generation capacity during under short-timescale condition [kW·h]
W'_g	Generation capacity during under long-timescale condition [kW·h]
α_c, α_h	Convective heat transfer coefficient of inner and outer pipe-wall $\left[W/\left(m^2 \cdot K\right)\right]$
β_i	Expansion ratio of i stage expander
δ	Thickness of pipe-wall [m]
η_i	Adiabatic efficiency of i stage expander
ε	Fluid compressibility
ρ	Inlet air density of regulating valve $\left[kg/m^3\right]$
η_m	Motor efficiency
η_c	Compressor efficiency
η_g	Generator efficiency
η_t	Expander efficiency
γ	Suction coefficient of ejector
η_b	Conversion efficiency of combustor

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Article

Surrogate Model-Based Heat Sink Design for Energy Storage Converters

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Abstract: As forced-air cooling for heat sinks is widely used in the cooling design of electrical and electronic equipment, their thermal performance is of critical importance for maintaining excellent cooling capacity while reducing the size and weight of the heat sink and the equipment as a whole. This paper presents a method based on the combination of computational fluid dynamics (CFD) simulation and surrogate models to optimize heat sinks for high-end energy storage converters. The design takes the thermal resistance and mass of the heat sink as the optimization goals and looks for the best design for the fin height, thickness and spacing, as well as the base thickness. The analytical and numerical results show that the thermal resistance and mass of the heat sink are reduced by the proposed algorithms, as are the temperatures of the heating elements. Test results verify the effectiveness of the optimization method combining CFD simulation with surrogate models.

Keywords: computational fluid dynamics (CFD); energy storage; surrogate model; design optimization; heat sinks; power converters

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1. Introduction

In the face of global warming and the rapid depletion of fossil-fuel resources, the utilization of renewable energy has become a general trend across the world. With the large-scale and high penetration of renewable energy in power grids, the stable supply of electricity and energy-storage technologies have become technical concerns. Power Conversion Systems (PCS) can improve the utilization quality of clean energy and stabilize the load fluctuations of the power grid. In some cases, PCS can be used as an energy source where power grids are not available (i.e., power islands). Because of these advantages, PCS are gaining in popularity in industrial, as well as domestic, applications. In PCS power electronics converters, switching devices such as metal-oxide-semiconductor field-effect transistors (MOSFETs) and insulated gate bipolar transistors (IGBTs) are major heat sources and the heat is removed typically by heat sinks, which can be of natural cooling or forced-air cooling. The thermal performance of components such as IGBTs is directly related to the reliability of the whole converter. Recognizing that the main cause of electronic equipment failure is overheating [1], heat sink design for converters is a key focus of this work.

In the literature, quantitative analysis of the key parameters is considered useful. Typically, the analysis is based on calculating the thermal resistance of the heat sink, which is related to the thickness of the substrate, and the number and thickness of the heat sink fins. However, this does not give an optimization method for the structural design of the heat sink [2]. In some cases, the geometric parameters that affect the thermal resistance of the heat sink are optimized by using a genetic algorithm method [3]. Ref. [4] studies the influence of a single parameter of the heat sink on the performance and then develops the orthogonal method to assess the influence of different parameters on the performance of the heat sink. However, the design cycle is tedious, and the experimental cost is relatively high. Generally, when studying the heat dissipation performance of a heat sink, it is necessary to consider not only the thermal resistance but also the volume and the cost of the heat sink. There are several parameters that affect the heat dissipation performance of the heat sink, such as the thickness of the base, and the length, thickness and number of fins. The optimization in this paper considers the influence of multiple parameters for multiple optimization targets. In this paper, the heat sink and forced air cooling are used in an IGBT-based energy storage converter where IGBTs are considered the main heat sources for cooling design. CFD simulation and surrogate model algorithms are combined to realize the double objective optimization (the thermal resistance and mass of the heat sink). The optimal parameter design can be obtained by bringing the optimized results into the CFD model to verify the correctness of the designs.

2. Mathematical Model

This work carries out a thermal analysis of a high-end PCS converter which involves an extensive understanding of heat transfer within a heat sink.

2.1. Design Variables

In this work, there are four independent variables: the thickness of the base, and fin height, thickness and spacing. The structural diagram of the optimized forced-air-cooled heat sink is shown in Figure 1.

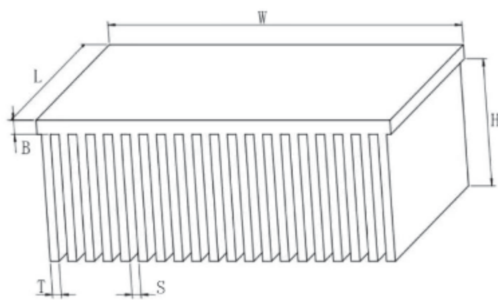

Figure 1. Structural diagram of the forced-air-cooled heat sink.

Where B is the thickness of the base, and H, T and X are the height, thickness and spacing of the fins, respectively, all of which taken together are defined as x_1, x_2, x_3, x_4, W is the width of the heat sink, L is the length of the heat sink and H is the height of the heat sink. Based on these variables, the thermal resistance and mass of the heat sink can be obtained and used as the objective functions of the optimization design.

2.1.1. Boundary Conditions

The optimization variables of the heat sink are associated with IGBT size, fan size, model and airflow rate. For effective heat transfer, the height of the fins shall not be greater than that of the fan, otherwise, airflow misses part of the fins and reduces the heat dissipation performance. Similarly, the spacing between fins shall be greater than 1 mm. The thickness of the heat sink fin shall be greater than 1 mm for easy manufacture. In summary, the constraints of design optimization are given by:

$$\begin{cases} 1\text{mm} \le x_1 \le 20\text{mm} \\ 50\text{mm} \le x_2 \le 100\text{mm} \\ 1\text{mm} \le x_3 \le 4\text{mm} \\ 1\text{mm} \le x_4 \le 8\text{mm} \end{cases} \tag{1}$$

2.1.2. Objective Functions

Considering the heat dissipation in the heat sink, the optimization objectives of the heat sink are low thermal resistance, low mass, high heat transfer efficiency and overall cost. In this paper, the thermal resistance and mass of the heat sink are selected as the design targets, and the entropy weight method [5] and linear weighting method [6] are combined to process the numerical data. This processing method can not only solve the problem of different variable units but also eliminate the influence of different variable values on the optimization results, making the data dimensionless for comparison [7]. The specific process of data processing has five steps:

- Standardize the data;

$$T_{ij} = \frac{\max(x_{1j}, \ldots, x_{nj}) - x_{ij}}{\max(x_{1j}, \ldots, x_{nj}) - \min(x_{1j}, \ldots, x_{nj})} + 0.0001 \tag{2}$$

- Calculate the entropy value of the jth index, which ranges between 0 and 1;

$$e_j = -\left(\frac{1}{\ln n}\right)\sum_{i=1}^{n} \frac{T_{ij}}{\sum\limits_{i=1}^{n} T_{ij}} \ln \frac{T_{ij}}{\sum\limits_{i=1}^{n} T_{ij}} \tag{3}$$

- Calculate the weighting of each data set (I = 1,2 … ,n; j = 1,2 … ,m);

$$w_j = \frac{1 - e_j}{\sum\limits_{j=1}^{m} (1 - e_j)} \tag{4}$$

- Establish the priority function of the targets, based on the linear weighting method;

$$P_i = \sum_{j=1}^{m} w_j x_{ij} \tag{5}$$

- Obtain the multi-objective optimization function of the heat sink;

$$\min F(x) = \min[w_1 R(x) + w_2 M(x)] \tag{6}$$

From the above calculations, the weightings of the thermal resistance and the mass are:

$$w_1 = 0.588,$$
$$w_2 = 0.412.$$

2.2. Thermal Resistance Model

The thermal resistance of the heat sink consists of four parts [8], the thermal resistance between the junction and the device case, the thermal resistance between the shell and the environment, the thermal resistance between the shell and the heat sink, and the thermal resistance between the heat sink and the environment [9]. Figure 2 shows the equivalent thermal resistance network diagram of the heat sink.

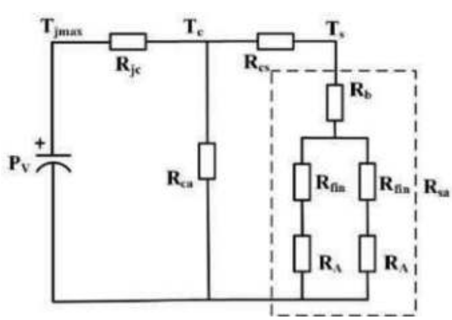

Figure 2. Thermal resistance model.

In Figure 2, P_V is the total heat loss of the IGBT, R_{jc} is the thermal resistance between the junction and shell, R_{ca} is the thermal resistance between the shell and the environment, R_{cs} is the thermal resistance between the shell and heat sink and R_{sa} and is the thermal resistance between the heat sink and the environment. R_b is the heat conduction resistance of the heat sink base, R_{fin} is the heat conduction resistance of the fins and R_A is the heat convection resistance of the fins. Similar to electrical resistances in electrical circuits, the thermal resistance can be calculated by adding up the component resistances according to their series or parallel connections. The heat transfer in the forced-air-cooled heat sink follows Ohm's law for thermal circuits. Compared with R_{cs} and R_{sa}, R_{ca} is very large and can be ignored in parallel connection; then:

$$\Delta T = P\left(R_{jc} + R_{cs} + R_{sa}\right) \tag{7}$$

According to Equation (7), the temperature rise of the device is determined by the loss and thermal resistance of the power device. The losses of power devices and the thermal resistance of the junction–shell are mainly determined by the production process, packaging materials and performance index. These two parameters are constant values [10]. The thermal resistance between the shell and heat sink is very small and can be ignored. The thermal resistance between the heat sink and the environment can be reduced by optimizing the geometric structure, materials and external environment of the heat sink. Compared to other parameters, the thermal resistance between the heat sink and the environment is also relatively simple and convenient to optimize. For this reason, the thermal resistance between the heat sink and the environment is the objective that majorly corresponds to the total thermal resistance.

2.3. Optimization Model

The thermal resistance of the heat sink in this design is mainly composed of the heat convection thermal resistance and the heat conduction resistance of the heat sink base and fins, which can be calculated by Equations (8)–(10). The total thermal resistance of the heat sink can be calculated by Equation (11).

$$R_b = \frac{B}{k_s WL} \tag{8}$$

$$R_{fin} = \frac{H}{k_s TL} \tag{9}$$

$$R_A = \frac{1 - 0.152(vL)^{-\frac{1}{10}}}{5.12(vL)^{\frac{4}{5}}(H+S)\cdot 2N} \tag{10}$$

$$R_{sa} = R_b + \frac{R_{fin}}{2N} + R_A \tag{11}$$

where N is the number of air ducts, ks is the thermal conductivity of the heat sink material and v is the airflow velocity.

In this design, the size of the heat sink is 230 × 420 mm and the height is variable. According to Equations (8)–(11), the optimization model of the heat sink resistance is:

$$\min F_1 = \frac{x_1}{k_s * 420 * 230} + \frac{x_2}{k_s * x_3 * 230}8 + \frac{1 - 0.152(230v)^{-\frac{1}{10}}}{5.12(230v)^{\frac{4}{5}}(x_2 + x_4) * 2 * \left\{\left[\frac{420 - x_3}{x_3 + x_4}\right] + 1\right\}} \tag{12}$$

From Figure 2, the heat sink mass optimization model is:

$$\min F_2 = 420 * 230 * x_1\rho_m + 230 * x_2x_3x_5\rho_m \tag{13}$$

where ρ_m is the density of the heat sink material, which is aluminum in this case.

3. The Proposed Optimization Methods

In the literature, surrogate model algorithms are adopted, which do not rely on actual models, thus reducing computational time and cost. The surrogate model algorithms use a Kriging algorithm to establish a simplified model by using basic parameters such as independent variables, objective functions and limiting conditions [11]. The success of the surrogate model depends on its sampling strategy and the number and location of sampling points [12]. The generated sample points should accurately reflect the distribution characteristics of sampling points [13,14]. A Kriging algorithm is then applied to search for the optimization point.

3.1. Kriging Models

At present, the methods of establishing models are divided into two categories: one is a parametric model, and the other is a non-parametric model. The former is based on the assumption that the known parameters obey a given population distribution. The latter model does not assume any particular distribution of the population. In the case of a given sample, it is calculated according to non-parametric statistics [15].

In the optimization, a Kriging model based on a parametric model is selected, which is also an alternative model widely used at present. The Kriging model is given by [16–18]:

$$y(x) = \mu + z(x) \tag{14}$$

where $y(x)$ is the response function, μ is the constant, $z(x)$ is the random process, respectively. Its expectation, variance and covariance are:

$$\begin{cases} E[z(x)] = 0 \\ Var[z(x)] = \sigma^2 \\ cov[z(x^j), z(x^k)] = \sigma^2 R[R(x^j, x^k)] \end{cases} \tag{15}$$

where R is the symmetric correlation matrix on the diagonal, $R(x^i, x^j)$ and is the correlation parameter between sample points x^j, x^k. In this optimization, a Gaussian correlation function is selected, then

$$R(x^j, x^k) = \exp[-\sum_{i=1}^{n} \theta_i \left| x_i^j - x_i^k \right|^2] \tag{16}$$

where n is the number of groups of design variables. In this optimization, $n = 50$. $\left| x_i^j - x_i^k \right|$ is the distance between the kth components of sample points x^j and x^k, and θ_i is the unknown related parameter used to fit the model, which can be solved by Equation (17).

$$\max F(\theta_i) = -\frac{[n_s \ln(\widehat{\sigma}^2) + \ln|R|]}{2} \tag{17}$$

Then, the estimated value at sample point x is:

$$\widehat{y}(x) = \widehat{\mu} + r^T(x) R^{-1} (y - \widehat{\mu} m) \tag{18}$$

$$r^T(x) = [R(x, x^1), R(x, x^2), \ldots R(x, x^{n_s})]^T \tag{19}$$

where y is the column vector with length n_s and m is the unit vector.

Figure 3 shows six Kriging models constructed with different variables and target parameters. It can be seen that the Kriging models differ from their 2D parameters and objectives. Considering different two-dimensional parameters and the minimum value of the objective function, a set of optimization values can be generated. This design needs to consider that the target value is the lowest in the case of four-dimensional parameters, which requires a search algorithm to find the target minimum value [19].

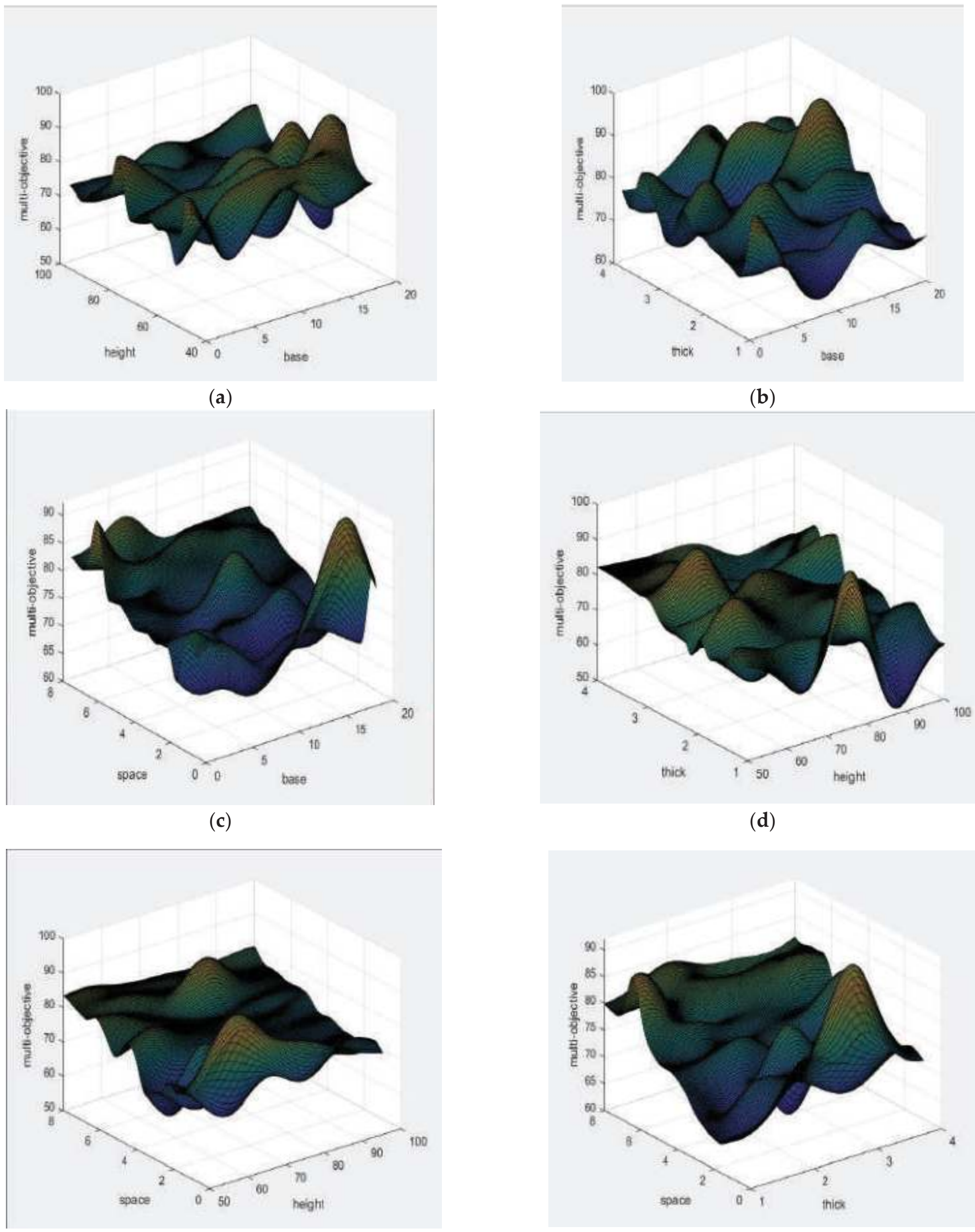

Figure 3. Kriging model constructed with different variables and target parameters. (**a**) Kriging model constructed with B, H and multi-objective. (**b**) Kriging model constructed with B, T and multi-objective. (**c**) Kriging model constructed with B, S and multi-objective. (**d**) Kriging model constructed with H, T and multi-objective. (**e**) Kriging model constructed with H, S and multi-objective. (**f**) Kriging model constructed with T, S and multi-objective.

3.2. Particle Swarm Optimization (PSO) Algorithm

In the literature, particle swarm optimization (PSO) algorithms are favored due to their advantages of easy programming, high efficiency and fast convergence [20], the optimization in this paper selects PSO as the search algorithm. Each particle in the particle swarm itself can find the optimal solution of its current position and record it as an individual extreme value. Particles can share an individual extremum with other particles, and the optimal individual extremum is regarded as the global optimal solution of the whole particle swarm [21]. Then, the particle adjusts its speed and position according to its individual extreme value and the current global optimal solution, and constantly approaches the optimal solution [22]. The PSO algorithm is shown in Equation (20).

$$\begin{cases} V_i^{k+1} = wV_i^k + C_1 r_1 (P_i^k - X_i^k) + C_2 r_2 (P_g^k - X_i^k) \\ X_i^{k+1} = X_i^k + V_i^{k+1} \end{cases} \tag{20}$$

where V_i^k and X_i^k are the velocity and position of the ith particle at the kth iteration, P_i^k represents the individual extreme value of the ith particle at the kth iteration, P_g^k represents the global optimal solution of the ith particle at the kth iteration, C_1 and C_2 are individual learning factors and global learning factors of particles, r_1 and r_2 are random constants between $[0,1]$ and w is the inertia weight, respectively. In this design, the value of w decreases linearly according to the number of iterations k. The equation is given by [23]:

$$w(k) = w_{max} - \frac{w_{max} - w_{min}}{k_{max}} \times k \tag{21}$$

where w_{max}, w_{min} are the upper and lower limits of the inertia weight and k_{max} is the maximum number of iterations, respectively.

3.3. Process of Surrogate Algorithms

The whole optimization process is shown in Figure 4. First, design the variables and determine the range of independent variables, and then take Latin hypercube sampling [24]. After obtaining a limited number of sample points, choose whether to supplement points in some areas or redesign the design according to the sampling conditions of the sample points [25]. The sampling condition of this design is good, so the supplement of sample points is not considered. The second step is the selection of target parameters. This can choose single objective optimization or multi-objective optimization according to the design requirements. In general, the objectives of multi-objective optimization are in conflict with each other. The improvement of one sub-objective may cause the reduction of another sub-objective, and the solution result cannot make all sub-objectives optimal. This requires that, when carrying out multi-objective optimization, the weight proportion between different objectives should be considered [26]. In the multi-objective optimization problem, the solution is not unique, which requires specific analysis of specific problems in the optimization design to find solutions that meet more sub-objectives as much as possible. In this process, the optimal design should meet the requirements of independent variables, constraints and target parameters at the same time. The third step is to establish a finite element model and use the finite element model to obtain relevant parameters. Then, the Kriging model is established using 80% of the sampled data for an approximate model, which is used to replace the established finite element model. The remaining 20% is used for error detection. If the error between the data obtained from the approximate model and the simulation is within the 20% threshold, the surrogate model is considered successful. After error detection, the PSO algorithm is used to obtain the optimal solution, which is finally brought into the finite element model to verify the results, to complete the optimization design [27].

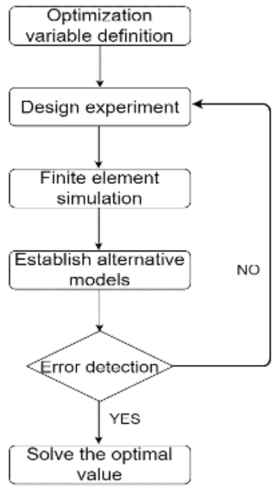

Figure 4. Flowchart of the process of surrogate algorithms.

4. Simulation Results

The structure of the PCS converter is complex for 3D thermal modeling and analysis. In this work, a simplified model is developed for design optimization, and the 3D finite element model only includes key components, as shown in Figure 5.

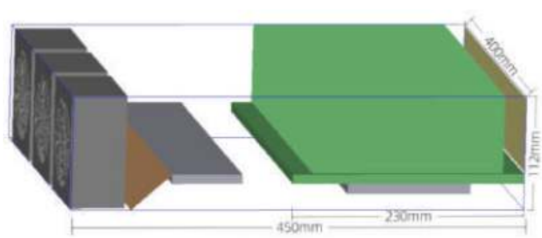

Figure 5. 3D finite element model.

In this converter, four heating element IGBTs are installed on the base of the heat sink, and the three fans blow air to the fins of the heat sink through the air deflector. In the model, the size of the cabinet is 450 mm × 400 mm × 114 mm, the radius of the fan is 44 mm, the size of the IGBT is 122 mm × 62 mm × 14 mm, and the spacing between IGBTs is 38 mm. The length and width of the heat sink base are 420 mm and 230 mm, the height of the base is 12 mm, the height and thickness of the fins are 64 mm and 3.4 mm, and the spacing between heat sinks is 7 mm, respectively. The power loss of each IGBT is 800 W, and the total power loss is 3200 W. The ambient temperature is 50 °C. The simulation results are plotted in Figures 6–9. It can be seen that the maximum temperature of the heating element can reach 133.43 °C. The calculated results are brought into the finite element model for verification. The finite element simulation and optimization results are shown in Figures 6–9, respectively.

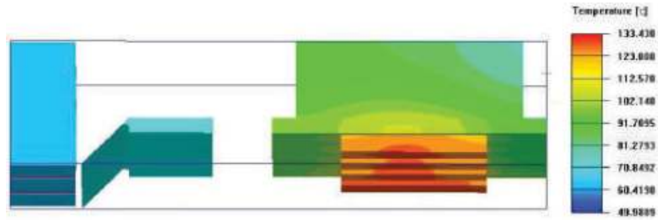

Figure 6. 3D Finite element simulation results before optimization.

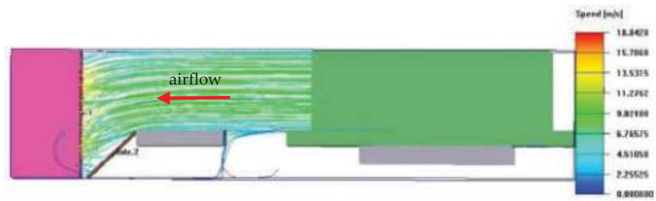

Figure 7. Airflow speed vector results before optimization.

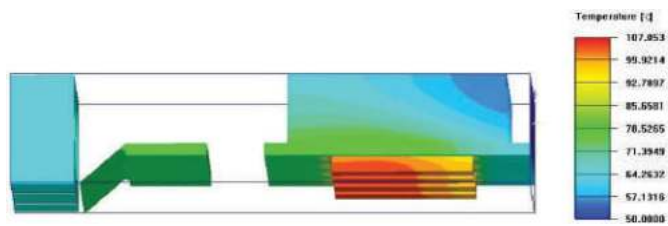

Figure 8. 3D Finite element simulation results after optimization.

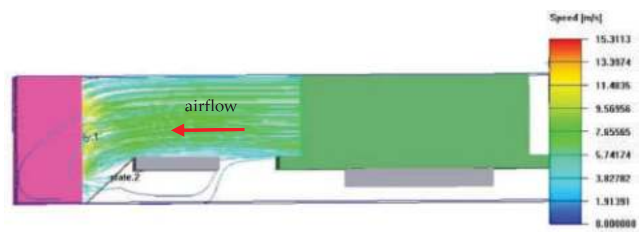

Figure 9. Airflow speed vector results after optimization.

It can be seen that the maximum airflow velocity after optimization is significantly reduced. The changes in the heat sink ducts and the relative positions of the heat sink and fans can vary the airflow path and contact surface of the heatsink, which is confirmed by Figures 6 and 8.

In contrast, a traditional method with response surface optimization in ANSYS adopts an optimization algorithm, as illustrated in Table 1. The proposed surrogate optimization results with the surrogate model are better. After the optimization design, the thermal resistance and mass are reduced from 0.025 °C/W and 7.562 kg to 0.017 °C/W and 7.312 kg, respectively. From the temperature cloud diagrams in the 3D finite element simulation, it can also be seen that the hot-spot temperature is reduced from the 133.430 °C value to 107.053 °C, proving the effectiveness of the proposed surrogate optimization algorithms.

Table 1. Optimum design of the heat sink.

	B	H	T	S	Thermal	Mass	Temperature
Base design	12.0 mm	64.0 mm	3.4 mm	7.0 mm	0.0253 °C/W	7.562 kg	133.430 °C
Traditional design	8.0 mm	80.0 mm	2.0 mm	4.4 mm	0.0194 °C/W	7.423 kg	114.733 °C
Optimization design	10.8 mm	65.6 mm	1.1 mm	2.3 mm	0.0171 °C/W	7.312 kg	107.053 °C
Reduction (%)	10.0	−2.5	67.6	67.1	32.4	3.3	19.8

5. Gray Correlation Analysis

Gray correlation analysis can evaluate the importance of input parameters, calculate the impact of different parameters on multiple objectives, and verify the selection of parameters in the design [28]. After all data are made dimensionless [29], the correlation coefficient is calculated by Equation (22). Then the average value is calculated to obtain the gray correlation degree of each parameter with the target [30].

$$\begin{cases} r_i^k = \frac{\Delta_{min}+\lambda\Delta_{max}}{\Delta_i^k+\lambda\Delta_{max}} \\ \Delta_i^k = \left|y_0^k - y_i^k\right| \end{cases} \tag{22}$$

where r_i^k represents the correlation coefficient of the kth component of the ith parameter, y_0^k and y_i^k represents the results of dimensionless processing of the reference sequence and the comparison sequence, Δ_i^k represents the deviation sequence, Δ_{min} and Δ_{max} represents the minimum and maximum deviation, and λ is the resolution coefficient, respectively. In this design, the resolution coefficient is taken as 0.5. The gray correlation degrees of the four variables relative to thermal resistance, mass and weighted multi-objective are shown in Table 2.

Table 2. Summary of Gray correlation analysis.

Dimensionless	Base	Height	Thick	Space
Temperature	0.637699	0.590268	0.668855	0.697728
Mass	0.622044	0.673455	0.698375	0.605095
Multi-objective	0.639337	0.586499	0.672616	0.726296

Within the given range of variables, the four design variables have a high degree of correlation with the target parameters, and the correlation degree can basically reach 0.6, which also proves the rationality of the design variables selected. Of these, for the temperature single target, the spacing of the heat sink fins has the greatest influence, and the correlation degree is as high as 0.7. The correlation degree of the thickness of the fins is the second, which is 0.67. Compared with other parameters, the correlation between the thickness of the fins and the temperature is the smallest, but it can also reach 0.59. It can also be seen that the airflow path is very crucial for the heat dissipation optimization design, while the influence of the surface area of the fins is less obvious. For the mass in the target parameters, the parameter with the greatest correlation is the thickness of the fins, and the correlation reaches 0.7. For the dual objectives of thermal resistance and mass, the most relevant parameter in this optimization is fin spacing, which is as high as 0.73, and the second important parameter is fin thickness, with a correlation degree of 0.67. By comparison, fin height has the smallest correlation degree with the double targets, with a correlation coefficient of 0.59. In this double objective optimization, fin spacing and thickness are the most significant factors affecting the performance of the heat sink.

6. Conclusions

This paper has presented an optimization method for heat sinks based on the CFD method and surrogate models. Firstly, the heat dissipation process is simulated by using the 3D finite element model. Next, using the sampling points selected by Latin hypercube sampling and the simulation results of CFD, a surrogate model is established to assess the correlation of four design variables (B, H, T, S) and the two design targets (thermal resistance and mass), according to Kriging models. Then a PSO algorithm is used to obtain the optimal solution. The test results show that the optimized thickness of the base and the height, thickness and spacing of the fins can effectively reduce the thermal resistance of the heat sink and the mass of the heat sink. In turn, the proposed method can reduce hotspots (such as an IGBT junction) and improve the thermal efficiency of the converters. The developed technology can provide guidance for heat sink designs and optimize the energy efficiency of industrial products.

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Article

Comparative Study of the Transmission Capacity of Grid-Forming Converters and Grid-Following Converters

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Abstract: The development trend of high shares of renewables and power electronics has increased the demand for new energy converters in the power system, but there is a lack of systematic research on the stability of different types of converters when transmitting power, which is worth exploring in depth. In this study, the power transfer capabilities of grid-forming and grid-following converters are investigated separately through an equivalent circuit diagram and phasor diagram when connected to the grid, and a quantitative relationship between converters' power transmission limit and short circuit ratio under static stability conditions is obtained, leading to the conclusion that, in terms of power transmission, grid-forming converters are more suitable for weak grids with high damping and low inertia, whereas grid-following converters are more suitable for strong grids with high inertia. The conclusions are further verified by constructing the converter grid-connected models for different grid strengths through the PLECS simulation platform and the real-time simulation RTBOX1 and F28379D launchpad platform.

Keywords: grid-forming converter; grid-following converter; static power transmission limit; power coupling; short circuit ratio

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1. Introduction

The penetration rate of renewable energy and power electronic equipment has increased, forming a "double high" development trend [1]. The high transmission capacity requirement of converters also leads to problems in grid stability [2].

When a new energy converter is connected to the power grid as a power source, there may be an imbalance between its actual output power and its input power on the DC side when disturbance occurs, which will affect the capacitance voltage on the DC side of the converter, then change its output current through the control system; finally, its actual output power changes to adopt a new balance. If one part fails during this circulation, different kinds of power grid instability will occur. Among them, the static stability can initially be judged according to whether the converter has a stable static operating point [3,4]. However, power electronic converters have different response characteristics compared with traditional equipment, and there is still scant research on the stability of different types of converters when transmitting power and quantitative analyses of their power transmission limit, which are highly related to the power grid stability of power-angle, voltage, and frequency [5,6].

Among them, power-angle stability refers to the fact that the output power of the generator to the grid can still be maintained within a constant range after the grid is subjected to minimal interference (interference close to zero). Considering the synchronization consistency between converters and synchronous machines, the stability problem can be divided into static stability, small disturbance stability, and transient stability [7–9]. However, static stability is the basis for the study of dynamic small disturbance, transient stability, and other problems; the transient instability problem presents the characteristics of time-varying nonlinearity and needs to take into account all elements constituting the grid, which is relatively complex. Therefore, this paper mainly studies the influence of converter power transmission on the static stability of the grid.

Power electronic converters, as non-synchronous sources, can be divided into grid-forming (GFM) and grid-following (GFL) converters [10]. Among them, GFM converters can be seen as voltage sources, while GFL converters can be seen as current sources [11]. Power electronic converters should not only meet the power-angle stability like traditional converters when transmitting power, but also need to prevent PLL loss [12], delay [13], or other effects causing static or small interference instability, which need further research.

In this study, the equivalent circuit models of GFM and GFL converters are established, the power transmission limits of the two converters under static stability are obtained, and the conclusions and conditions that they are equal are determined. Then, the short circuit ratio (SCR) is used to measure the power grid strength of the system under static stability [14], utilizing the control strategy block diagram of both converters; we conclude that the GFM converter is more suitable for weak grids, and the GFL converter is more suitable for strong grids in terms of power transmission and smooth regulation. Finally, the PLECS platform is used to construct the grid-connected system suitable for different power grid strengths, and the validity of this theoretical analysis is verified using the simulation platform PLECS, RTBOX1, and the F28379 launchpad.

2. Static Stability Limit of the Converter's Power Transmission

The main focus of this section is the power transmission limit of the converter under static stability. The converter is connected to the grid as a power source; therefore, in addition to providing active power, a large amount of reactive power and harmonics may be output to the grid, complicating analyses of the stability. Thus, it seems reasonable to simplify this by considering static power transmission limits when converters operate in unity power factor first [15]. Equivalent circuits and phasor diagrams of converters are constructed to facilitate further analyses of power transmission limits and the static stability.

2.1. Equivalent Circuit of the GFM Converter

In Figure 1a, U_c is the RMS voltage value of the converter, and U_g is the RMS value of the grid side voltage. Assuming that the voltage at the point of common coupling (PCC) is the reference phase, the angle between the grid side voltage and the voltage at the PCC is the power-angle δ, and the line impedance $R_g + jX_g = Z_g\angle\varphi$. To facilitate control and improve stability, all the reactive power is provided by reactive power compensators. The power factor of the converter is one, expressed as $I_t\angle 0^\circ$.

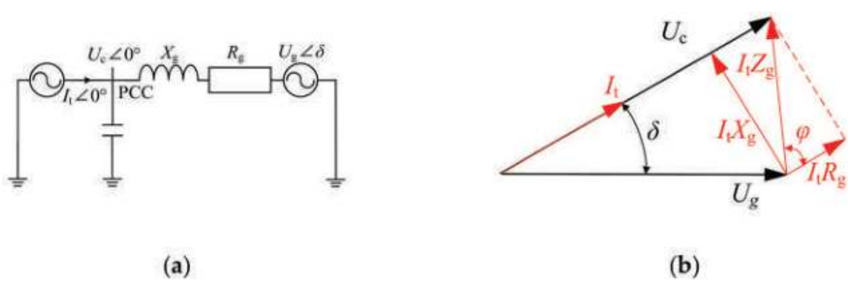


Figure 1. (**a**) Equivalent circuit diagram of the GFM converter; (**b**) phasor diagram of the GFM converter with unity power factor.

The phase diagram shown in Figure 1b was drawn by combining the relationships of the variables in Figure 1a, which, in turn, gives U_c and the active power, P, transmitted by the system in three symmetrical phases:

$$U_c = \sqrt{U_g^2 - I_t^2 X_g^2} + I_t R_g \tag{1}$$

$$P = 3U_c I_t = 3\left(\sqrt{U_g^2 - I_t^2 X_g^2} + I_t R_g\right) \cdot I_t \tag{2}$$

As shown in Figure 2a, the P–I_t curve can be plotted for different line impedance ratios after the transmission of active power P and current I_t has been normalized. When I_t increases, P exhibits an overall trend of increasing and then decreasing; thus, there is an extreme point which is the maximum value, P_{max}, referring to the maximum power that the converter can transmit in the unity power factor state. As the line impedance angle, φ, increases, the resistive component of the line impedance decreases, and the maximum power, P_{max}, that can be transmitted by a GFM converter will also decrease. As shown in Figure 2a, when the converter is commanded to transmit power P_{set}, for R_g/X_g = 1 and R_g/X_g = 0.5 grids, there is a stable static operating point, i.e., the left intersection of curves; for the R_g/X_g = 0 grid, there is no intersection, so static instability and other severe phenomena will occur.

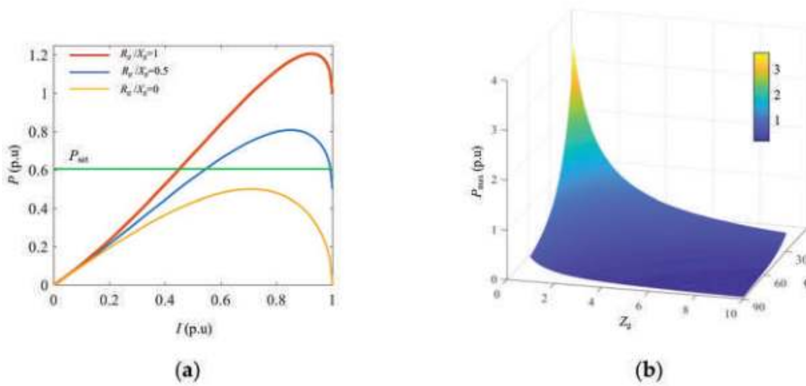


Figure 2. (**a**) P–I_t curves for different line impedance ratios; (**b**) plot of P_{max} and Z_g, φ.

Replacing X_g and R_g in Equation (2) with the impedance modulus Z_g and the impedance angle φ by equivalents $X_g = \tan\varphi/(1 + \tan^2\varphi)^{(1/2)}$ and $R_g = (1 + \tan^2\varphi)^{(-1/2)}$, Equation (3) can be derived, and with the aid of the mathematical application Wolfram Alpha, the explicit analytical solution for P_{max} can be displayed as in Equation (4):

$$P = 3\left(I_t\sqrt{U_g^2 - I_t^2 Z_g^2 \frac{\tan^2\varphi}{1+\tan^2\varphi}} + \frac{Z_g}{\sqrt{1+\tan^2\varphi}} I_t^2\right) \tag{3}$$

$$P_{max} = P(I_t)\Big|_{I_t = I_{pmax}} = \frac{3U_g^2}{2Z_g} \cdot \frac{\sqrt{1+\tan^2\varphi}}{\sqrt{1+\tan^2\varphi} - 1},\ I_{pmax} \approx \frac{U_g}{\sqrt{2}Z_g} \tag{4}$$

From Equation (4), we can derive Figure 2b; when Z_g and φ increase, P_{max} will be reduced, and the specific analysis will be combined with the SCR in Section 3.1.

However, in real operation conditions, especially in weak grids, converters may be called upon to provide the necessary reactive power to support the power system and compensate the voltage drop of the original transmission line using locally installed reactive power compensators [14]; notably, the required reactive power from compensators can be considerably high which makes the installation of them costly.

As shown in Figure 3a, the phasor diagram illustrates how GFM converters provide approximately 0.7P reactive power, which is required by grids with θ which means a power factor angle of I_t. Similarly, we can derive P in the variable power factor in Equation (5).

$$\begin{aligned} P &= U_c I_{tp} = \left(\sqrt{U_g^2 - (I_{tp}X_g - I_{tq}R_g)^2} + I_{tp}R_g + I_{tq}X_g\right) I_{tp} \\ &= I_t \cos\theta \sqrt{U_g^2 - I_t^2(X_g\cos\theta - R_g\sin\theta)^2} + I_t^2\left(R_g\cos^2\theta + \frac{X_g}{2}\sin 2\theta\right) \end{aligned} \quad (5)$$

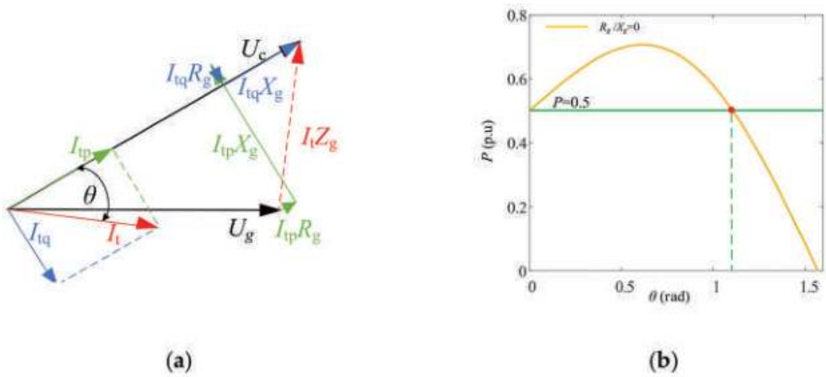


Figure 3. (**a**) Phasor diagram of the GFM converter provides the required reactive power; (**b**) plot of P and θ.

As shown in Figure 3b, for the R_g/X_g = 0 grid, P reached the highest at 0.5 (p.u.) when I was near 0.7 (p.u.). Keep the I and change the converter's power factor, and it turns out that P reached the highest value when the θ neared 0.6 rad and was higher than 0.5 (p.u.) when the θ value was smaller than 1.0 rad. By decreasing the power factor in a large scale, GFM converters can provide an even higher active power. Therefore, although power transmission with variable power factors is very complicated, some aspects of which are beyond the scope of this study, it can be concluded that converters can meet the requirements of rated active power with or without providing sufficient reactive power; this paper focuses more on the state of unity power factor to simplify the question.

2.2. Equivalent Circuit of the GFL Converter

As shown in Figure 4a, compared with Figure 1a, due to the presence of the reactive power compensation capacitor, X_c [16], the output current is not equal to the output current, Ig, from the PCC to the grid side when the GFL converter is considered as a current source; the output voltage is still equal to the voltage at the PCC when the GFL converter is considered as a voltage source, so the model of the GFL converter is slightly more complex. The GFL converter is regarded as a current source whose output current is decoupled by dq through Park transformation and the PLL control strategy in the GFL converter [10]. The dq decoupling analysis method of currents is more suitable for GFL than GFM converters, not only because the GFL converter is a current source which should be focused more in its output current, but also due to its current inner loop control, which will be further discussed in Section 3.2. The quantitative relationship is $I_{gd} = I_{td}$, $I_{gq} = U_{td}/X_c$.

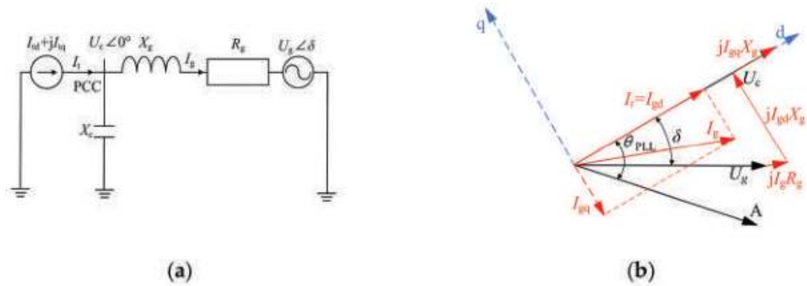


Figure 4. (**a**) Equivalent circuit diagram of a GFL converter; (**b**) phasor diagram of GFL converter.

Combining the relationship and the diagram in Figure 4b, the voltage at the PCC can be derived as shown in Equation (6). Under static stability conditions, PLL accurately tracks the current; thus, $U_{\mathrm{td}} = U_{\mathrm{t}}$ can be substituted into Equation (6) to obtain Equation (7). In general, the reactance to ground at the PCC X_{c} is much greater than the inductive reactance in the X_{g} grid, i.e., $X_{\mathrm{c}}/(X_{\mathrm{c}} - X_{\mathrm{g}}) = 1$; thus, Equation (7) is formally identical to Equation (2). Similarly, the maximum transmission power of a GFL converter can be presented as Equation (8).

$$P_{\max} = \sqrt{U_{\mathrm{g}}^2 - I_{\mathrm{gd}}^2 X_{\mathrm{g}}^2} + I_{\mathrm{gd}} R_{\mathrm{g}} + I_{\mathrm{gq}} X_{\mathrm{g}} = \sqrt{U_{\mathrm{g}}^2 - I_{\mathrm{td}}^2 X_{\mathrm{g}}^2} + I_{\mathrm{td}} R_{\mathrm{g}} + \frac{U_{\mathrm{td}}}{X_{\mathrm{c}}} X_{\mathrm{g}} \tag{6}$$

$$U_{\mathrm{c}} = \frac{X_{\mathrm{c}}}{X_{\mathrm{c}} - X_{\mathrm{g}}} \left(\sqrt{U_{\mathrm{g}}^2 - I_{\mathrm{td}}^2 X_{\mathrm{g}}^2} + I_{\mathrm{td}} R_{\mathrm{g}} \right) \tag{7}$$

$$P_{\max} = \frac{X_{\mathrm{c}}}{X_{\mathrm{c}} - X_{g}} \cdot \frac{3U_{\mathrm{g}}^2}{2Z_{\mathrm{g}}} \cdot \frac{\sqrt{1 + \tan^2 \varphi}}{\sqrt{1 + \tan^2 \varphi} - 1} \tag{8}$$

The form of Equation (8) is the same as Equation (4); judging from the equivalent circuit diagram, the ultimate transmission capacity in the unity power factor state is the same for both GFM and GFL converters under static stability conditions. It is a useful conclusion and means that the type of converter can be ignored when considering the ultimate transmission capacity of the converter alone.

3. Relationship between Static Power Transmission of the Converter and Grid Strength

3.1. Influence of Power Grid Strength Variation on the Steady Working Area of the System

According to the relative definition of SCR used to measure power grid strength in traditional power systems, the power grid strength of a grid-connected system with GFM and GFL converters can also be expressed by SCR [15]: the larger its value, the higher the strength of the system.

Equation (9) defines the SCR in renewable energy power grids: P_{sc} is the short circuit capacity of the grid, $P_{\mathrm{sc}} = 3U_{\mathrm{g}}^2/Z_{\mathrm{g}}$; U_{g} is the RMS value of the grid phase voltage; Z_{g} is the impedance modulus value of the grid; and P_{N} is the rated transmission power of the converters, which is equal to the P_{set} value in the GFM and GFL converters mentioned above. Thus, the strength of the power grid is not only related to the power grid itself, but also to the converter, which is jointly determined by both.

It seems that the definition of the SCR requires analyses of both converters and grids; however, because the transmission power limit of GFM converters is similar to that of GFL converters, it can reasonably be assumed that the rated transmission power of the two converters is equal, to determine the steady working area by only discussing the characteristics of the power grid itself using the SCR theory.

According to Equation (8) and Figure 2b, as Z_{g} increases continuously, SCR decreases, and $P_{\max}$ also decreases rapidly. However, when φ increases, the power grid strength increases, and P_{max} increases. Moreover, power grids connected to the converter are mostly of high voltage, whose φ is almost 90°; in this case, the line impedance angle has little effect on the improvement in P_{max}. In summary, as Z_g increases, the grid's stiffness reduces, the SCR decreases, and P_{max} decreases.

Furthermore, the steady working area requires $P_{set} \leq P_N$; according to Equations (8) and (9), the constraint relationship between the transferable power (steady working area) of converters and power grid strengths under static stability can be calculated as Equation (10).

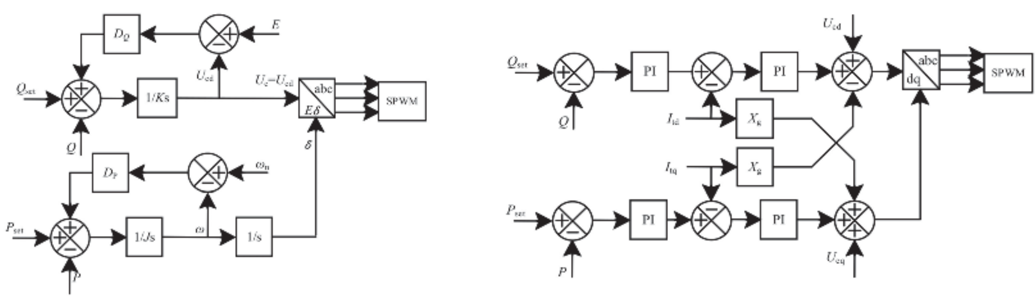
Notably, although the minimum SCR increases with the increase in impedance angle, it always exists at $SCR_{min} \leq 2$, which is consistent with the conclusion that the SCR of a very weak power grid is less than 2, and further verifies the accuracy of the conclusion.

$$\mathrm{SCR} = \frac{P_{sc}}{P_N} = \frac{3U_g^2}{Z_g \cdot P_{set}} \tag{9}$$

$$\begin{cases} P_{set} \leq \frac{3U_g^2}{2Z_g} \cdot \frac{\sqrt{1+\tan^2\varphi}+\tan^2\varphi+1}{\tan^2\varphi} \\ \mathrm{SCR} \geq 2 - \frac{2(\sqrt{1+\tan^2\varphi}+1)}{\sqrt{1+\tan^2\varphi}+\tan^2\varphi+1} \end{cases} \tag{10}$$

3.2. Influence of Power Grid Strength on the Power Transmission of Converters

As shown in Figure 5a, the block diagram of a GFM converter was used to further analyze the influencing factors of its stability during transmission power, which is not only conducive to exploring whether the converter is stable when it reaches P_{max}, but also when exploring the stability characteristics of the converter when it transmits arbitrary power under a certain strength grid. GFM converters regulate the active and reactive power output, P and Q, to the system by controlling the output voltage, U_c, and power-angle, δ, having $P = 3U_cU_g\sin\delta/X_g$ and $Q = 3U_c(U_c - U_g\cos\delta)/X_g$.



(a) **(b)**

Figure 5. (**a**) Block diagram of a GFM converter; (**b**) block diagram of a GFL converter.

Thus, by taking the partial derivative with respect to P and Q, with $\partial P/\partial U_c = 3U_g\sin\delta/X_g$, $\partial Q/\partial\delta = 3U_cU_g\sin\delta/X_g$, U_c and δ are expected to be taken as control quantities, P and Q are taken as controlled quantities, and the control GFM converter is a two independent single-input single-output system; thus, power coupling is not conducive to GFM converter control, which demonstrably does not meet the objective of the independent control of P and Q.

When the power grid strength decreases, SCR decreases, X_g increases, and $\partial P/\partial U_c$ decreases; therefore, the coupling relationship between P and Q decreases, and the stability of the GFM converter when transmitting P_{max} is further guaranteed. Notably, when P changes, the coupling relationship between P and Q does not change. Therefore, the characteristics of a GFM converter are applicable to weak grids in the steady working area of the system.

At the same time, because GFM converters continue the traditional control strategy of virtual synchronous generators (VSGs) and introduce gains, such as D_P and D_Q, when they are set to a certain value, the damping and standby inertia of the system is sufficient, and the power transmission is smoothly regulated within the limit range because the impedance of the weak grid is large. However, when the grid strength increases, the impedance decreases. Due to the limitation of VSG parameters, high-frequency oscillations may occur during power transmission due to parameter mismatch, which is not conducive to system stability. This further verifies that GFM converters are not suitable for strong grids.

As shown in Figure 5b, GFL converters regulate P and Q by controlling the d- and q-axis components of the output current, I_{td} and I_{tq}, respectively. As shown in Figure 4b, the converter is in a state of unity power factor; thus, $I_t = I_{td}$, and its P–I_{td} curve is similar to the curve in Figure 2a. When I_{td} increases, the operation is stable if P increases monotonically, and may be statically unstable if the monotonicity of P changes. The P–I_{td} curve is a convex function; therefore, it is concluded that the ultimate transmission power of GFL converters is the same as P_{max} in Equation (4), and as such, GFL converters are more suitable for higher-strength grids. Additionally, when the grid strength is high, due to the small impedance of the system, the time constant requirement is not high, and the power can be smoothly adjusted within the specified time; when the grid strength decreases, due to the rising impedance of the system, the power may not be adjusted within the specified time. If PI parameters are not set reasonably and the converter itself is adjusted too fast, it will produce an overshoot and a shock on the grid, which further verifies that GFL converters are more suitable for strong grids.

4. Simulation Waveform Analysis

Simulated waveforms are used to further verify the characteristics of GFM and GFL converters in transmitting power at different grid strengths. Assuming that the systems connected to the converter are all high-voltage grids, the grid strength is changed by varying the inductive reactance parameters in the line. More details about the simulation in PLECS are shown in Tables 1–3.

Table 1. Parameters of systems of three grid strengths.

Phase Voltage U_g (V)	R_g (Ω)	X_g (Ω)	SCR	P_{max} (W)
220	0.1	0.50	29.04	145,200
		3.00	4.84	25,000
		10.00	1.45	7300

Table 2. Parameters of systems of the GFM converter.

J/(kg·m^2)	D_p	D_q	K
0.057	5	321	7.1

Table 3. Parameters of systems of the GFL converter.

K_{p_PLL}	K_{i_PLL}	K_{p_i}	K_{i_i}
0.7978	99.0138	0.0043	0.7143

As shown in Table 1, it is assumed that the rated transmission power of both GFM and GFL converters is constant, and since the SCR is generally considered to be stronger than 20 for a strong grid and less than 2 for a very weak grid, it was reasonable to set the SCR to 29.04, 4.84, and 1.45 for this simulation.

Tables 2 and 3 show the parameters of GFM and GFL converter systems; the simulation was performed to verify the transmission capacity rather than modify converters, so these parameters were kept the same during the simulation.

4.1. Simulation of GFM Converter Power Transfer When SCR = 4.84

The main purpose of the grid-connected simulation model with an SCR of 4.84 is to reduce the influence of the grid strength on both converters, and thus, to observe the influence of the systems own power transmission limit on the converters' transmission power. As shown in Figure 6a, the actual value of $P_{\max}$ is slightly lower than the theoretical value (25,000 W) due to the limitations of GFM converter parameter adjustments. When the commanded power transmission is 22,000 W, the converter can deliver a rapid response within 1 s and reach the static stable equilibrium, as expected.

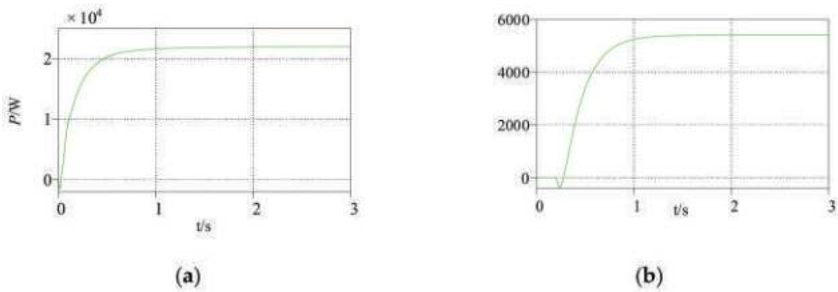


Figure 6. (**a**) Active power of a GFM converter (P_{set} = 22,000 W, SCR = 4.84); (**b**) reactive power of a GFM converter (P_{set} = 25,000 W, SCR = 4.84).

As shown in Figure 7, the GFM converter has high output current sinusoidality, low harmonics, and high immunity to interference at a statically stable transmission of 22,000 W, further verifying its good static stability performance. When the converter command rose from 22,000 W to 25,000 W, the actual power transmitted by the converter will rise steadily to 25,000 W; however, at this time, the converter cannot work in the unity power factor state, so it will output a certain amount of reactive power steadily to the outside. At this time, the converter output current is stable, as shown in Figure 7.

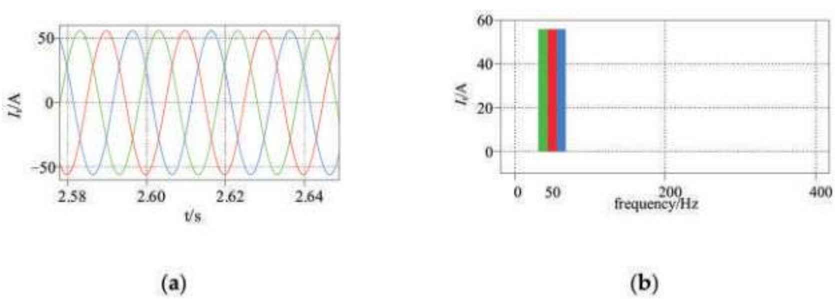


Figure 7. (**a**) Current curve of a GFM converter (P_{set} = 25,000 W, SCR = 4.84); (**b**) current THD of a GFM converter (P_{set} = 25,000 W, SCR = 4.84). (Green for phase A, red for phase B, blue for phase C).

When the active power command continues to increase, the converter cannot transmit the active power evenly when compensated by infinite reactive power. However, as shown in Figure 8a, when the active power command increases from 25,000 W to 50,000 W, the converter can only transmit about 40,000 W with high reactive power, so the system will be in a state of static instability and the output active power will be in a state of low-frequency oscillation (at a frequency of approximately 1 Hz). The reason is that the change in power-angle, δ, at this time cannot meet the requirements of active power instruction, so the power angle oscillates at low frequency, and the reactive power oscillates synchronously with the active power. In this case, although the converter's output current oscillates, it has a good sinusoidal degree and no obvious harmonics, as shown in Figure 8b.

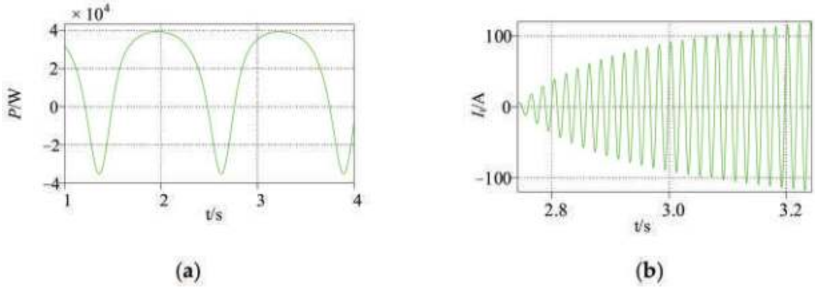


Figure 8. (**a**) Active power of a GFM converter (P_{set} = 50,000 W, SCR = 4.84); (**b**) current of a GFM converter (P_{set} = 50,000 W, SCR = 4.84).

4.2. Simulation of GFL Converter Power Transfer When SCR = 4.84

Similarly, when a GFL converter is commanded to transmit at 22,000 W, it still responds quickly and has little harmonics. When P_{set} is raised to 25,000 W, P slowly rises to slightly above the command, then becomes out of control and quickly rises to another stable operating point; the value of P is determined by the converter itself as long as P_{set} exceeds the limit power. The reason is shown in Figure 9b, as Q first operates out of control and cannot be stabilized in a smaller range, resulting in a sharp increase in Q, which further leads to a runaway P value, but at this point, Q can be as high as 40,000 var and a certain degree of harmonics is injected into the grid, as shown in Figure 10. The fifth and seventh harmonics indicate that Q is so high that it exceeds the GFL converter's transmission limit and causes instability. Due to the power coupling of a GFM converter, the reactive power does not become out of control; only a stable offset occurs, as shown in Figure 6.

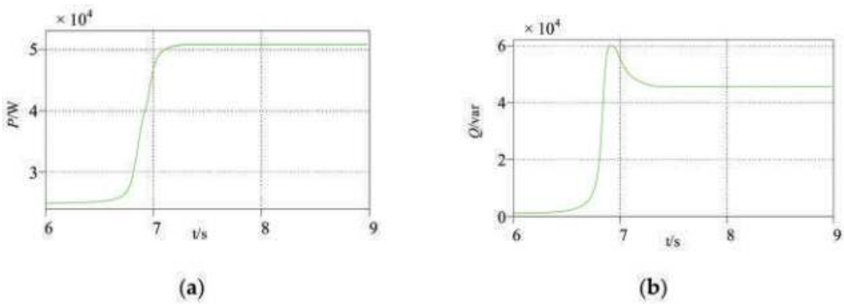


Figure 9. (**a**) Active power of a GFL converter (P_{set} = 25,000 W, SCR = 4.84); (**b**) reactive power of a GFL converter (P_{set} = 25,000 W, SCR = 4.84).

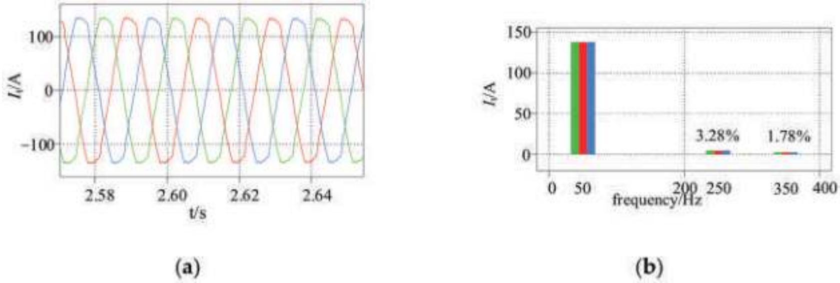


Figure 10. (**a**) Current of a GFL converter (P_{set} = 25,000 W, SCR = 4.84); (**b**) current THD of a GFL converter (P_{set} = 25,000 W, SCR = 4.84). (Green for phase A, red for phase B, blue for phase C).

4.3. Comparison of the Power Transfer Capability of Converters at Strong and Weak Grid Strengths

A simulation model for a strong grid with a SCR greater than 20 and a weak grid with SCR less than 2 was established to compare the influence of the grid strength on the transmission of power within the limit range of GFM and GFL converters.

As shown in Figure 11, when the P_{set} of the very weak grid was 7300 W, it could reach the target requirement within 1–2 s, whereas I_t and U_c have little harmonic similarity to that depicted in Figure 7: when the P_{set} of the strong grid reached 140,000 W (slightly below the ideal limit), it could respond rapidly in about 0.1 s, whereas U_c had high fifth and seventh harmonics, as shown in Figure 12, which is due to the large influence of power coupling to make the control quantity U_{cd} exceed normal levels and the PWM of the GFM converter to become far from sinusoidal. High harmonics from both GFM and GFL converters are caused by the runaway of control quantities, although specific reasons are different. GFM converters set a fixed virtual impedance in the face of a strong grid with a small impedance, which does not match the grid, resulting in a high frequency oscillation of U_c with a frequency of approximately 100 Hz; in contrast, the weaker grid impedance is larger, which offsets the virtual impedance setting problem and smooths the power output of the converter.

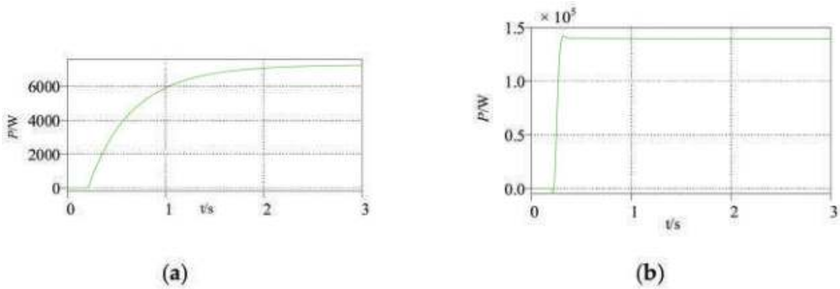


Figure 11. (**a**) Active power of a GFM converter (P_{set} = 7000 W, SCR = 1.45); (**b**) active power of a GFM converter (P_{set} = 140,000 W, SCR = 29.04).

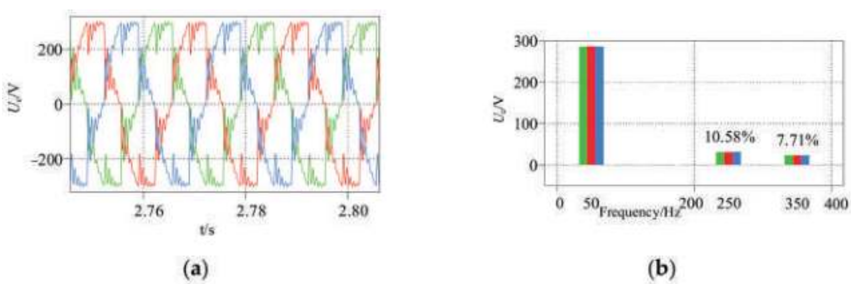


Figure 12. (**a**) Voltage of a GFM converter (P_{set} = 140,000 W, SCR = 29.04); (**b**) THD of a GFM converter (P_{set} = 140,000 W, SCR = 29.04). (Green for phase A, red for phase B, blue for phase C).

As shown in Figure 13, GFL converters can output P_{max} relatively more rapidly under both weak and strong grid conditions, with I_t and U_c having small harmonics. The PI controller time constant here is fixed at approximately 0.5 s; when the GFL converter runs on the weak grid, its regulation speed is higher than the grid, so P has obvious oscillation overshoot. At the same time, there is no virtual impedance setting in the GFL converter, so the regulation time is not too fast but relatively smooth in strong grids.

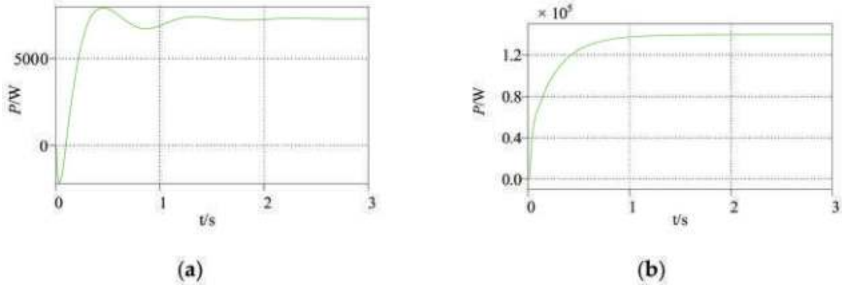


Figure 13. (**a**) Active power of the GFL converter (P_{set} = 7000 W, SCR = 4.84); (**b**) active power of the GFL converter (P_{set} = 140,000 W, SCR = 4.84).

4.4. Hardware in-Loop Simulation Using RTBOX1

Figure 14 shows the real-time simulation devices RTBOX1 and F28379D launchpad. Using this platform facilitates the verification of the above conclusions. Figure 15 shows two states of transmitting power by GFM or GFL converters; the simulation waveform characteristics, such as harmonics, are similar to those in PLECS, which further verifies that their transmission capacity could be realized in actual power grids.

Figure 14. RTBOX1 and F28379D launchpad.

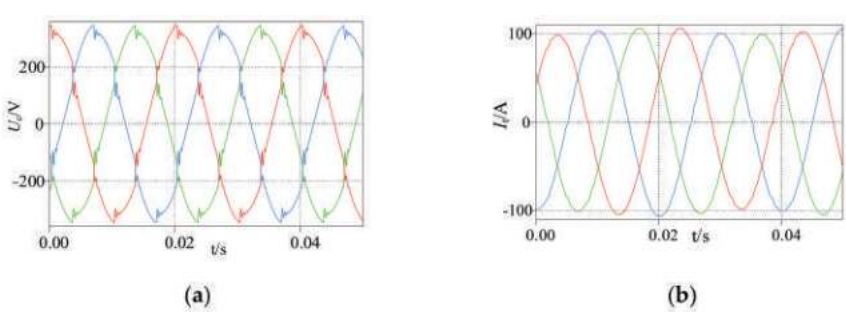


Figure 15. (**a**) Voltage of the GFM converter (P_{set} = 140,000 W, SCR = 29.04); (**b**) current of the GFL converter (P_{set} = 22,000 W, SCR = 4.84). (Green for phase A, red for phase B, blue for phase C).

5. Conclusions

This study investigated the transfer capabilities of the two main types of non-synchronous machine power sources connected to GFM and GFL grids, focusing on the transfer capability of converters at different grid strengths based on their static and stable operation.

1. There is a quantitative relationship between the statically stable power transmission limit and the line impedance of GFM and GFL converters at a unity power factor. At variable power factors, the converters can even transmit active rated power sufficiently. For high-voltage grids, the reactance, X_g, increases; its impedance modulus, Z_g, increases. SCR decreases, P_{max} decreases, and the operating range is reduced. However, when X_g increases, its impedance angle, φ, increases, and P_{max} increases, but the impact of φ on the transmission capacity is much smaller than that of Z_g. Therefore, the larger the impedance, the smaller the P_{max}; thus, the two are approximately inversely related.
2. The grid strength of a system is determined by SCR, which is jointly influenced by Z_g and the rated transmission power of the converter. When the reactance of PCC to ground is much greater than X_g, the P_{max} values of GFM and GFL converters are equal, and it can be considered that the grid strength is only related to X_g, i.e., the greater the grid strength, the greater the P_{max}.
3. Within P_{max} (in a steady working area), GFM converters are more suitable for weak grids because of the power coupling problem, which is more obvious in strong grids, whereas GFL converters are more suitable for strong grids because of the monotonic nature of the current in the P–I_t curve. Moreover, in smoothing power transfer, the parameters of GFM converters, which require virtual impedance settings, make them more suitable for weak grid regulation because they are prone to high-frequency oscillations under strong grids. GFL converters, however, with their parameters of time constants and PI controllers, are more suitable for the regulation of strong grids because they are prone to overshooting oscillations in weak grids.

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A Model Independent Predictive Control of PMSG Wind Turbine Systems with a New Mechanism to Update Variables

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Abstract: Permanent magnet synchronous generator (PMSG) wind power system with full power rating converter configuration is especially suitable for wind energy applications. Direct model predictive control (DMPC) has led to more possibilities in terms of choice because of its straightforward concept for PMSG wind turbine systems in *high-power* off-shore wind farms. However, due to complete dependence on the model knowledge, parameter mismatches will seriously deteriorate the system control performances. This work presents a model/parameter-independent predictive control method with a novel mechanism to update current/power variations online. The proposed method makes use of only *two* measurements from the former intervals and the selected control vectors to estimate all variations of the candidate vectors in the present interval. Benefiting from this updating mechanism, the proposed method is completely independent of the model parameters in the state prediction. However, it still has a very low calculating requirement and smooth current/power variation waveforms. The proposed method is compared with classical DMPC. The results validate that the proposed solution outperforms the classical DMPC with model deviations, with considerably improved robustness.

Keywords: PMSG wind turbines; back-to-back power converters; model-independent predictive control; robust control

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1. Introduction

Electricity generated by renewable energy has grown significantly since the end of the 20th century. Among the variations, high-power offshore wind energy has become increasingly competitive with other energy sources in terms of rich reserves, long generation time, and "cost-per-generated kilowatt hour" [1,2]. Large-capacity wind turbines have become essential in offshore wind energy installation in the last few years [3]. Permanent magnet synchronous generator (PMSG) with direct-drive configuration has considerable advantages in such a system in terms of higher energy density, gearbox elimination/reduction, and less maintenance [4,5]. As one such configuration, a very potent system based on full power rating back-to-back converter is spreading rapidly for high-power offshore wind energy generation [5]. A simplified circuit diagram of such a PMSG system is introduced in Figure 1. Considering both the nonlinear characters and switching nature of the power converters, the direct model predictive control (DMPC, also called finite-control-set model predictive control, FCS-MPC) is a promising method to implement with multiple-target optimization in a single step [6]. Therefore, DMPC makes the time-averaged modulation stage unnecessary in the control of power converters [7]. It offers fast dynamic performance and a straightforward design. The literature shows that the DMPC has already emerged as a widely-used control option for the system shown in Figure 1 [8–10].

Due to the reliance on model-based prediction of system states, DMPC suffers from system model deviations (caused by, e.g., inaccurate modeling and incorrect component parameters) [6,11]. In practice, the considerable *mismatches* between the values in the actual equipment and the values used in the controller often occur on *unavailable* (e.g., stator flux) and *time-varying* parameters (e.g., coil inductance in generator or filter) of the wind energy system. Inaccurate prediction of the system's future behaviors would result in this situation and induce performance deterioration (e.g., large steady-state tracking bias and increased control variable ripples [12]).

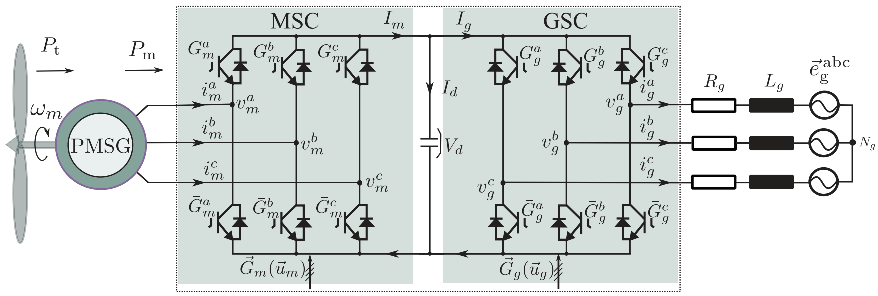


Figure 1. Simplified electrical circuit of a two-level back-to-back power converter-based PMSG wind turbine system with equivalent RL-filter.

Researchers have proposed several methods to improve the model predictive control's robustness. Three methods can be grouped as three different concepts. The first concept is *observer-based methods* [13,14]. In [15,16], an observer for perturbations caused by mismatch was employed in a deadbeat control solution. Reference [17] presents a full parameter disturbance and load observer to simultaneously estimate both electrical and mechanical parameters of PMSG for model predictive speed control. However, observer-based solutions are usually complicated, requiring higher tuning efforts and compromising system stability by adding redundant control loops. The second concept is *error compensation* [18]. In [19,20], the prediction model compensates for the state tracking bias caused by parameter mismatches, assuming that the prediction error for a given switching state remains unchanged within a few control intervals. In [21], a new cost function in proportional integral (PI) form is designed to eliminate steady-state errors. The improved parameter stability region is theoretically derived. However, since the prediction stage still uses the mismatched parameters, the control performance improvement is not satisfactory under large parameter deviation conditions. The third group is *model-free or model-independent predictive control (MIPC)* [22]. Reference [23] first presented such a solution for a PMSG drive. Different from the previous two groups of solutions, it designed look-up tables (LUTs) to maintain status variation for a previous control interval as information for the prediction of the system behaviors in the future interval. This eliminates the model information and, therefore, the controller gains greater robustness to significant parameter deviation. Nevertheless, if certain specific voltage vectors have not been applied for multiple control intervals, most of the previously stored current variation information will be outdated and unreliable for future prediction [23,24], resulting in large *updating lag*. This will significantly affect the control performance, causing unacceptable ripples in the output waveforms and even posing a significant risk to the stability of the system during transient states. Authors of the recent contributions [25] proposed a solution in which previously applied voltage vectors and current measurements are used to estimate the current variation under other voltage vectors to form a smoother updating waveform with less lag. However, such an update mechanism can only be activated when *three* previously applied voltage vectors differ. Moreover, a total number of 210 switching sequences have to be categorized to seek the optimal one, which significantly increases the computational burden of the controller.

To conquer the problems analyzed above, this paper proposes a new MIPC method based on the estimation of all possible variations to quickly update the LUTs. We validate this method on the control of both the generator and grid sides of a two-level back-to-back converter-based PMSG wind power system. It requires only *two* measurements from the former intervals and the relevant information regarding control vectors. Compared to the method that only updates the relevant variation of the selected vector, it estimates the necessary state variations for all possible voltage vectors in the present interval. The main contributions of this work include:

1. A new effective model-independent predictive control for both the generator and the grid side of high-power PMSG wind turbine systems is presented (see Section 4). The proposed method is immune to both generator- and grid-side parameter mismatches and model deviations. Robustness improvement of the proposed method outperforms the classical model-based MPC technique (see Section 5.2);
2. A new state variable variation updating mechanism is proposed, which assures smooth current/power variation waveforms and non-lag updating. The proposed solution is analytically developed and requires many fewer online calculations in comparison with the recently reported approaches (see Section 4);
3. The proposed solution is tested in various scenarios (see Section 5), which show promising results in enhanced robustness.

The contents of this article are organized as follows. In Section 2, the basic modeling of the grid-tied two-level back-to-back converter system is presented. Section 3 reviews the classical model-based and model-free methods for such a system, and in Section 4, we introduce the detailed proposed mechanism and the design of the controller. Section 5 reports on the verification and analysis of the proposed method. Finally, Section 6 concludes this paper.

2. System Description and Modeling

A two-level voltage source back-to-back power converter PMSG wind turbine system with direct-drive configuration is presented in Figure 1. A machine-side converter (MSC) and grid-side converter (GSC) are connected by the DC-link capacitor. During normal operation phases, MSC is used as the power interface to control the generator, while GSC aims to regulate the DC-link and grid-side power. The aerodynamics and turbine are modeled in, e.g., ref. [6]. These are not repeated here. Meanwhile, we assume a speed reference ω_{m}^{*} has been obtained as the maximum power point tracking (MPPT) requirement by another external controller which is not the focus of this study. In the following, the generator side, grid side, MSC, and GSC are modeled. Note that variables $\boldsymbol{x}^{\alpha\beta}$ in the stationary frame and $\boldsymbol{x}^{\mathrm{dq}}$ in the rotary reference frame are derived, invoking the corresponding (power invariant) Clarke and Park transformations, respectively.

The current dynamics of the surface-mounted permanent magnet generator (SPMSG) model in rotary reference coordinates can be expressed as

$$\begin{pmatrix} g_{i_{\mathrm{m}}^{\mathrm{d}}} \\ g_{i_{\mathrm{m}}^{\mathrm{q}}} \end{pmatrix} = \frac{1}{L_{\mathrm{s}}} \begin{pmatrix} v_{\mathrm{m}}^{\mathrm{d}} \\ v_{\mathrm{m}}^{\mathrm{q}} \end{pmatrix} - \frac{R_{\mathrm{s}}}{L_{\mathrm{s}}} \begin{pmatrix} i_{\mathrm{m}}^{\mathrm{d}} \\ i_{\mathrm{m}}^{\mathrm{q}} \end{pmatrix} - \omega_{\mathrm{e}} \begin{pmatrix} -i_{\mathrm{m}}^{\mathrm{q}} \\ i_{\mathrm{m}}^{\mathrm{d}} \end{pmatrix} - \frac{1}{L_{\mathrm{s}}} \begin{pmatrix} 0 \\ \omega_{\mathrm{e}} \cdot \psi_{\mathrm{pm}} \end{pmatrix}, \tag{1}$$

where $\boldsymbol{G}_{\mathrm{m}} = (g_{i_{\mathrm{m}}^{\mathrm{d}}}, g_{i_{\mathrm{m}}^{\mathrm{q}}})^{\top}$ is the gradient of the PMSG stator currents and $\boldsymbol{v}_{\mathrm{m}}^{\mathrm{dq}}$ and $\boldsymbol{i}_{\mathrm{m}}^{\mathrm{dq}}$ denote MSC output voltage and PMSG stator currents in the dq frame, respectively.

Using measured values of the voltage and current from the *point of coupling*, the instantaneous power is calculated as [26]

$$\begin{pmatrix} P \\ Q \end{pmatrix} = \underbrace{\begin{pmatrix} e_{\mathrm{g}}^{\alpha} & e_{\mathrm{g}}^{\beta} \\ e_{\mathrm{g}}^{\beta} & -e_{\mathrm{g}}^{\alpha} \end{pmatrix}}_{\boldsymbol{E}_g} \begin{pmatrix} i_{\mathrm{g}}^{\alpha} \\ i_{\mathrm{g}}^{\beta} \end{pmatrix}. \tag{2}$$

Power dynamics in stationary coordinates for a balanced grid are

$$\begin{pmatrix} g_\mathrm{P} \\ g_\mathrm{Q} \end{pmatrix} = \frac{1}{L_\mathrm{g}} \boldsymbol{E}_g \begin{pmatrix} v_\mathrm{g}^\alpha - e_\mathrm{g}^\alpha \\ v_\mathrm{g}^\beta - e_\mathrm{g}^\beta \end{pmatrix} - \begin{pmatrix} \frac{R_\mathrm{g}}{L_\mathrm{g}} P + \omega_\mathrm{g} Q \\ \frac{R_\mathrm{g}}{L_\mathrm{g}} Q - \omega_\mathrm{g} P \end{pmatrix}, \tag{3}$$

where $\boldsymbol{G}_\mathrm{g} = (g_\mathrm{P}, g_\mathrm{Q})^\top$ represents the grid-side power gradient and $v_\mathrm{g}^\alpha, v_\mathrm{g}^\beta$, and $e_\mathrm{g}^\alpha, e_\mathrm{g}^\beta$ represent GSC output voltage vectors and grid voltages in the $\alpha\beta$ frame, respectively.

Introducing G_y^x as the switch signal for the IGBTs in Figure 1, where $\mathrm{y} \in \{\mathrm{m}, \mathrm{g}\}$ and $x \in \{\mathrm{a}, \mathrm{b}, \mathrm{c}\}$, the complementary signal for the opposite IGBTs in the same converter leg can be written as $\bar{G}_\mathrm{y}^\mathrm{x}$. The switching state u_y^x can be defined accordingly as

$$u_\mathrm{y}^\mathrm{x} := \mathcal{G}(G_\mathrm{y}^\mathrm{x}) = \begin{cases} P & \text{if} : G_\mathrm{y}^{\mathrm{x}1} = 1 \\ N & \text{if} : G_\mathrm{y}^{\mathrm{x}1} = 0 \end{cases} \tag{4}$$

for phase x. The 3-phase has 2^3 vector options for each side converter to meet the control requirements, presented as

$$\boldsymbol{u}_\mathrm{y}^\mathrm{abc} = (u_\mathrm{y}^\mathrm{a}, u_\mathrm{y}^\mathrm{b}, u_\mathrm{y}^\mathrm{c})^\top \in \mathcal{S}_8 := \{NNN, NNP, \cdots, PPN, PPP\}. \tag{5}$$

Taking switching states and DC-link voltage V_d into consideration, the phase voltages of the converter can be obtained as [27]

$$\boldsymbol{v}_\mathrm{y}^\mathrm{abc} = \begin{pmatrix} v_\mathrm{y}^\mathrm{a} \\ v_\mathrm{y}^\mathrm{b} \\ v_\mathrm{y}^\mathrm{c} \end{pmatrix} = \frac{V_\mathrm{d}}{3} \begin{pmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{pmatrix} \boldsymbol{u}_\mathrm{y}^\mathrm{abc}. \tag{6}$$

Known from Figure 1, DC-link voltage depends on the current flow of the MSC and GSC and can be modeled as

$$\frac{\mathrm{d}V_\mathrm{d}(t)}{\mathrm{d}t} = \frac{1}{C} I_\mathrm{d}(t) = \frac{1}{C}\big(I_\mathrm{g}(t) - I_\mathrm{m}(t)\big), \tag{7}$$

where $I_\mathrm{g} = \boldsymbol{i}_\mathrm{g}^{\mathrm{abc}\top} \cdot \boldsymbol{u}_\mathrm{g}^\mathrm{abc}$ and $I_\mathrm{m} = \boldsymbol{i}_\mathrm{m}^{\mathrm{abc}\top} \cdot \boldsymbol{u}_\mathrm{m}^\mathrm{abc}$ present the DC-link current components of the grid and generator side, respectively.

3. Classical Direct Model Predictive Control Methods

3.1. System Requirements and Cost Function Design

The control objectives for the GSC and MSC in a normal back-to-back converter-based wind turbine system, shown in Figure 1, are listed as (see, e.g., [28]):

- *Torque/current control:* The primary torque/current control must have a promising dynamic performance to generate proper reference considering: (a) MPPT of the wind turbine system or (b) suitable torque generation for supercritical wind speed. Stress on the mechanical components needs to be reduced by minimizing torque ripples and current THDs;
- *Complex power control:* The GSC must regulate the grid-side complex power quickly and dynamically to reduce fluctuations in the DC link caused by the intermittent feed-in of wind power. To meet the requirements of the Grid Code, a low-current distortion factor should be ensured;
- *DC-link voltage control:* A stable DC-link voltage is required for the proper operation of the system;
- *Switching frequency regulation:* As the wind turbine power level rises, needed improvement of efficiency at any point requires reducing switching losses by low switching frequencies.

Closely related to these requirements, state variables, i.e., the generator stator current $\boldsymbol{x}_\mathrm{m} = (i_\mathrm{m}^\mathrm{d}, i_\mathrm{m}^\mathrm{q})^\top$ and grid power $\boldsymbol{x}_\mathrm{g} = (P, Q)^\top$ are taken into consideration in two cost functions ($\mathrm{y} = \mathrm{m}$ for generator side, $\mathrm{y} = \mathrm{g}$ for grid side), as

$$J_y = \left\| \boldsymbol{x}_y(k+2) - \boldsymbol{x}_y^* \right\|_2^2 + \lambda_y \Delta \boldsymbol{u}_y(k+1), \tag{8}$$

where $\|\boldsymbol{x}\|_2$ is the status of the system $\boldsymbol{x}$ and $\Delta\boldsymbol{u}_y(k+1) = \boldsymbol{u}_y(k+1) - \boldsymbol{u}_y(k)$ represents the switching change in this control period. The weighting factor, λ_y, is designed for the optimization of multiple control objectives, which consist of the desired status and switching frequency here. For a surface-mounted PMSG wind turbine control, i_m^{d*} shall be zero to ensure the maximum torque per ampere (MTPA) control of PMSG. A unity power factor can be achieved by setting Q^* as zero in the grid-side control. The outer PI loops calculate the reference i_m^{q*} and P^* to meet the above requirements of the generator's speed and DC-link voltage. The delay of the sampling process in the real digital controller can be compensated by introducing the state variables at $k+2$, instead of variables at $k+1$, during the traversal of all possible control vectors [29].

3.2. Classical Direct Model Predictive Control

To get the state variables at the $k+1$ instant, the classical DMPC calculates the future stator currents and power for a finite set of voltage vectors by utilizing the system model, as

$$\boldsymbol{x}_y(k+1) = \boldsymbol{x}_y(k) + T_s \cdot \boldsymbol{G}_y(k), \tag{9}$$

where $y \in \{m, g\}$, $\boldsymbol{x}_m$ is the stator current vector $\boldsymbol{i}$ and $\boldsymbol{x}_g$ is the grid power vector $\boldsymbol{S}$. $T_s \cdot \boldsymbol{G}_y(k)$ denotes the state variation caused by this control interval. The same idea gives the prediction at the $k+2$ instant of the state variables as

$$\boldsymbol{x}_y(k+2) = \boldsymbol{x}_y(k+1) + T_s \cdot \boldsymbol{G}_y(k+1). \tag{10}$$

The controller selects and records the optimal vector from all control vectors, which minimizes the cost function (8). In the next control interval, this will be applied, and relevant delay compensation will be achieved as mentioned above. This procedure is repeated while new measurements arrive continuously [30]. However, it is observed that prediction of state variables in Equation (10) requires accurate information on the system parameters. The controller will derive the wrong prediction of the system utilizing mismatched parameters, resulting in the selection of the non-optimal control vectors. This has serious effects on the control of the stator current or the grid power until the system no longer functions properly and stably. To this end, we will present model-independent solutions in the following part.

3.3. A Latest Model-Independent Predictive Control Solution

Reference [24] presented an efficient MIPC algorithm for synchronous reluctance motor control. In this work, its principles are extended to the underlying PMSG wind turbine system as a benchmark. The dq-axis stator currents (of PMSG) and complex power (of GSC) variations under different voltage vectors are calculated and stored in four LUTs. The LUTs are updated online with new measurements. The future behavior of the system is calculated using the LUTs' data as

$$\boldsymbol{x}_y(k+1) = \boldsymbol{x}_y(k) + \Delta\boldsymbol{x}_y(k), \tag{11}$$

where $\Delta\boldsymbol{x}_y(k)$ denote the currents (for $y = m$) and power variations (for $y = g$) caused by the application of the voltage vector at time step k, which is determined in the $k-1$ control interval. $\Delta\boldsymbol{x}_y(k)$ is considered the same as the last ones stored in the LUTs. Then the possible status at time step $k+2$ is derived as

$$\boldsymbol{x}_y(k+2) = \boldsymbol{x}_y(k+1) + \Delta\boldsymbol{x}_y(k+1). \tag{12}$$

The current and power prediction of MIPC is hence accomplished using $\Delta\boldsymbol{x}_y$ stored in the LUTs, avoiding the use of system parameters.

This is easy to understand because, for a voltage vector that has not been selected during a long period, the stored information will be obsolete, which will influence the prediction accuracy and the control performance. The MIPC in [24] forces the output of this vector that has not been applied for a given time period, regardless of the cost function minimization optimization principle, which results in increased ripple and degraded steady-state performance. An improved MIPC with a new state variable update principle is proposed in the next section.

4. Proposed Model-Independent Predictive Control Solution

This section introduces an improved model-independent predictive control method with fast updating look-up tables (LUTs). The overall diagram is given in Figure 2. A grid-side power analysis based on instantaneous power clearly describes the idea of the method. The basis for this begins with a small signal model analysis.

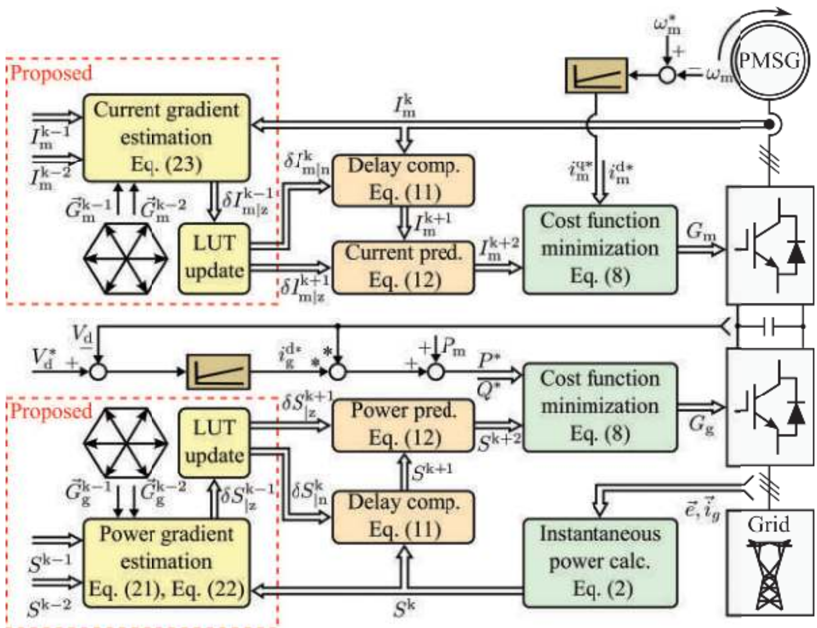


Figure 2. Control scheme of the proposed model-independent solution for back-to-back power converter based PMSG wind turbine system.

4.1. Small-Signal Modeling of the System

Grid active and reactive power in steady-state is described as

$$\boldsymbol{G}_{\mathrm{m}} = \begin{pmatrix} 0 \\ 0 \end{pmatrix} = \frac{1}{L_{\mathrm{g}}} \boldsymbol{E}_g \begin{pmatrix} v_{\mathrm{g}}^{\alpha} - e_{\mathrm{g}}^{\alpha} \\ v_{\mathrm{g}}^{\beta} - e_{\mathrm{g}}^{\beta} \end{pmatrix} - \begin{pmatrix} \frac{R_{\mathrm{g}}}{L_{\mathrm{g}}} P + \omega_{\mathrm{g}} Q \\ \frac{R_{\mathrm{g}}}{L_{\mathrm{g}}} Q - \omega_{\mathrm{g}} P \end{pmatrix}. \tag{13}$$

Equation (3) subtracts Equation (13), deriving the small-signal model of the grid side as

$$\begin{pmatrix} g_{\delta P} \\ g_{\delta Q} \end{pmatrix} = \frac{1}{L_{\mathrm{g}}} \boldsymbol{E}_g \begin{pmatrix} \delta v_{\mathrm{g}}^{\alpha} \\ \delta v_{\mathrm{g}}^{\beta} \end{pmatrix} - \begin{pmatrix} \frac{R_{\mathrm{g}}}{L_{\mathrm{g}}} \delta P + \omega_{\mathrm{g}} \delta Q \\ \frac{R_{\mathrm{g}}}{L_{\mathrm{g}}} \delta Q - \omega_{\mathrm{g}} \delta P \end{pmatrix}, \tag{14}$$

where $(\delta P, \delta Q)^{\top} = \delta \boldsymbol{S}$ represents the complex power variation and $\delta v_{\mathrm{g}} = v_{\mathrm{g}} - v_{\mathrm{g}}^{\mathrm{s}}$ indicates the relationship between control voltage vectors and the steady-state voltage vector. For instance, $\delta v_{\mathrm{g|i}} = v_{\mathrm{g|i}} - v_{\mathrm{g}}^{\mathrm{s}}$ represents the vector between the steady-state voltage and the ith converter voltage vector.

Converting to discrete form, Equation (14) yields

$$\begin{pmatrix} \delta P \\ \delta Q \end{pmatrix} = \frac{T_\mathrm{s}}{L_\mathrm{g}} \boldsymbol{E}_g \begin{pmatrix} \delta v_\mathrm{g}^\alpha \\ \delta v_\mathrm{g}^\beta \end{pmatrix}, \tag{15}$$

where T_s denotes the sampling interval. Now the small-signal model can be introduced into the controller with measurements of the grid-side voltages. Note that $T_\mathrm{s} \cdot \delta \boldsymbol{S}$ are small enough to be neglected. Similarly, the discrete small-signal model of PMSG can be derived as

$$\begin{pmatrix} \delta i_\mathrm{m}^\mathrm{d} \\ \delta i_\mathrm{m}^\mathrm{q} \end{pmatrix} = \frac{T_\mathrm{s}}{L_\mathrm{s}} \begin{pmatrix} \delta v_\mathrm{m}^\mathrm{d} \\ \delta v_\mathrm{m}^\mathrm{q} \end{pmatrix}. \tag{16}$$

The small-signal equations illustrate the relationship between the state variables' variation and the relevant control vectors. With these equations, the LUTs have the opportunity to update variation caused by all control vectors at the same interval.

4.2. Look-Up Table Update Principle

The steady-state vector v_g^s can not easily be obtained by the sampling process. The elimination of this vector from the small-signal model (see Equation (15)) relies on the iterative calculation among the converter voltage vectors. Based on Equation (15), two former voltage vectors ($\delta v_{g|k-1}$ and $\delta v_{g|k-2}$) with their relevant power variation ($\delta P_{|k-1}, \delta Q_{|k-1}$ and $\delta P_{|k-2}, \delta Q_{|k-2}$) can be derived as

$$\begin{pmatrix} \delta P_{|k-1} - \delta P_{|k-2} \\ \delta Q_{|k-1} - \delta Q_{|k-2} \end{pmatrix} = \frac{T_\mathrm{s}}{L_\mathrm{g}} \boldsymbol{E}_g \begin{pmatrix} \delta v_{\mathrm{g|i}}^\alpha - \delta v_{\mathrm{g|j}}^\alpha \\ \delta v_{\mathrm{g|i}}^\beta - \delta v_{\mathrm{g|j}}^\beta \end{pmatrix}. \tag{17}$$

Note that, given $\delta v_{g|k-1} = v_{\mathrm{g|k-1}} - v_\mathrm{g}^\mathrm{s}$, $\delta v_{g|k-2} = v_{g|k-2} - v_\mathrm{g}^\mathrm{s}$, we obtain $\delta v_{g|k-1} - \delta v_{g|k-2} = v_{g|k-1} - v_{g|k-2}$.

Hence, the steady-state vector v_g^s can be eliminated from Equation (17), and Equation (17) can be rewritten as

$$\begin{pmatrix} \delta P_{|k-1} - \delta P_{|k-2} \\ \delta Q_{|k-1} - \delta Q_{|k-2} \end{pmatrix} = \frac{T_\mathrm{s}}{L_\mathrm{g}} \boldsymbol{E}_g \begin{pmatrix} v_{\mathrm{g|i}}^\alpha - v_{\mathrm{g|j}}^\alpha \\ v_{\mathrm{g|i}}^\beta - v_{\mathrm{g|j}}^\beta \end{pmatrix}. \tag{18}$$

Based on Equation (18), the state variations can be derived using information from interval $k-1$ and $k-2$ as

$$\begin{pmatrix} \Delta P_{|\mathrm{i}}(k-1) - \Delta P_{|\mathrm{j}}(k-2) \\ \Delta Q_{|\mathrm{i}}(k-1) - \Delta Q_{|\mathrm{j}}(k-2) \end{pmatrix} = \frac{T_\mathrm{s}}{L_\mathrm{g}} \boldsymbol{E}_g \times \begin{pmatrix} v_{\mathrm{g|i}}^\alpha(k-1) - v_{\mathrm{g|j}}^\alpha(k-2) \\ v_{\mathrm{g|i}}^\beta(k-1) - v_{\mathrm{g|j}}^\beta(k-2) \end{pmatrix}, \tag{19}$$

where $\boldsymbol{v}_{\mathrm{g|i}}(k-1)$ and $\boldsymbol{v}_{\mathrm{g|j}}(k-2)$ are the ith and jth converter vector, which are applied at the $k-1$ and $k-2$ instants, respectively; $\Delta \boldsymbol{S}_{|\mathrm{i}}(k-1) = \boldsymbol{S}(k) - \boldsymbol{S}(k-1)$, $\Delta \boldsymbol{S}_{|\mathrm{j}}(k-2) = \boldsymbol{S}(k-1) - \boldsymbol{S}(k-2)$ represent state variation calculated with sampling values.

We then change the special ith vector at $k-1$ instant to all vectors (zth, $z \in \{1, 2, \cdots, 8\}$) that may be selected, and their relevant power variation can be estimated as

$$\begin{pmatrix} \Delta P_{|\mathrm{z}}(k-1) - \Delta P_{|\mathrm{j}}(k-2) \\ \Delta Q_{|\mathrm{z}}(k-1) - \Delta Q_{|\mathrm{j}}(k-2) \end{pmatrix} = \frac{T_\mathrm{s}}{L_\mathrm{g}} \boldsymbol{E}_g \times \begin{pmatrix} v_{\mathrm{g|z}}^\alpha(k-1) - v_{\mathrm{g|j}}^\alpha(k-2) \\ v_{\mathrm{g|z}}^\beta(k-1) - v_{\mathrm{g|j}}^\beta(k-2) \end{pmatrix}. \tag{20}$$

So far, a relationship has been found between the selected voltage vector at the instant $k-2$, with its relevant change in complex power, and the possible complex power variation with all candidate control vectors at instant $k-1$.

Note that the calculation of Equation (20) still relies on the inductance parameter. To remove this parameter from status estimation, Equations (19) and (20) can be combined and Equations (21) and (22) can be derived, only introducing the control vector applied at instant $k-1$ and the corresponding status variation, as

$$\Delta P_{|\mathrm{z}}(k-1) = \frac{\left(v^{\alpha}_{\mathrm{g|z}}(k-1) - v^{\alpha}_{\mathrm{g|j}}(k-2)\right)e^{\alpha}_{\mathrm{g}} + \left(v^{\beta}_{\mathrm{g|z}}(k-1) - v^{\beta}_{\mathrm{g|j}}(k-2)\right)e^{\beta}_{\mathrm{g}}}{\left(v^{\alpha}_{\mathrm{g|i}}(k-1) - v^{\alpha}_{\mathrm{g|j}}(k-2)\right)e^{\alpha}_{\mathrm{g}} + \left(v^{\beta}_{\mathrm{g|i}}(k-1) - v^{\beta}_{\mathrm{g|j}}(k-2)\right)e^{\beta}_{\mathrm{g}}} \times \left(\Delta P_{|\mathrm{i}}(k-1) - \Delta P_{|\mathrm{j}}(k-2)\right) + \Delta P_{|\mathrm{j}}(k-2), \quad (21)$$

$$\Delta Q_{|\mathrm{z}}(k-1) = \frac{\left(v^{\alpha}_{\mathrm{g|z}}(k-1) - v^{\alpha}_{\mathrm{g|j}}(k-2)\right)e^{\beta}_{\mathrm{g}} - \left(v^{\beta}_{\mathrm{g|z}}(k-1) - v^{\beta}_{\mathrm{g|j}}(k-2)\right)e^{\alpha}_{\mathrm{g}}}{\left(v^{\alpha}_{\mathrm{g|i}}(k-1) - v^{\alpha}_{\mathrm{g|j}}(k-2)\right)e^{\beta}_{\mathrm{g}} - \left(v^{\beta}_{\mathrm{g|i}}(k-1) - v^{\beta}_{\mathrm{g|j}}(k-2)\right)e^{\alpha}_{\mathrm{g}}} \times \left(\Delta Q_{|\mathrm{i}}(k-1) - \Delta Q_{|\mathrm{j}}(k-2)\right) + \Delta Q_{|\mathrm{j}}(k-2). \quad (22)$$

For the generator side, following a similar theoretical calculation process, stator current variation update equations are estimated using the last two current measurements as

$$\Delta \boldsymbol{I}_{\mathrm{m|z}}(k-1) = \frac{\left(\boldsymbol{v}^{\mathrm{dq}}_{\mathrm{m}|z}(k-1) - \boldsymbol{v}^{\mathrm{dq}}_{\mathrm{m}|j}(k-2)\right)}{\left(\boldsymbol{v}^{\mathrm{dq}}_{\mathrm{m}|i}(k-1) - \boldsymbol{v}^{\mathrm{dq}}_{\mathrm{m}|j}(k-2)\right)} \times \left(\Delta \boldsymbol{I}_{\mathrm{m|i}}(k-1) - \Delta \boldsymbol{I}_{\mathrm{m|j}}(k-2)\right) + \Delta \boldsymbol{I}_{\mathrm{m|j}}(k-2), \quad (23)$$

where $\Delta \boldsymbol{I}_{\mathrm{m|i}}(k-1), \Delta \boldsymbol{I}_{\mathrm{m|j}}(k-2)$ are the calculated current variation corresponding to applied vectors $\boldsymbol{v}^{\mathrm{dq}}_{\mathrm{m}|i}(k-1), \boldsymbol{v}^{\mathrm{dq}}_{\mathrm{m}|j}(k-2)$; and where $\Delta \boldsymbol{I}_{\mathrm{m|i}}(k-1) = \boldsymbol{I}_{\mathrm{m}}(k) - \boldsymbol{I}_{\mathrm{m}}(k-1), \Delta \boldsymbol{I}_{\mathrm{m|j}}(k-2) = \boldsymbol{I}_{\mathrm{m}}(k-1) - \boldsymbol{I}_{\mathrm{m}}(k-2)$. $\Delta \boldsymbol{I}_{\mathrm{m|z}}(k-1)(z \in \{1, 2, \cdots, 8\})$ denote estimated current variation under all candidate MSC output vectors at instant $k-1$.

Furthermore, the denominators in Equations (21)–(23) need to remain non-zero while estimating the status variation update (Equations (21)–(23)). In this work, the LUTs are updated when the denominators in these equations outweigh a certain value. Otherwise, the LUTs maintain as their values the previous interval. By obtaining the variation updating mechanism under all control vectors of back-to-back power converters, the controller guarantees the LUTs' update frequency and skips *updating-lag* completely.

The process of this method is summarized as follows. Firstly, the controller obtains the current state value of the system by sampling and calculates the variation between instant $k-2$ and $k-1$ by reading the value of the previous moment. Secondly, it uses Equations (21)–(23) to update all candidate control vectors with their possible variation in one control interval. Finally, the controller uses the generator side and grid side LUTs instead of the system model to calculate the future stator currents and to power-predict the system variables for the finite set of voltage vectors, completing the compensation and prediction and selecting the optimal vector that minimizes the cost function.

5. Verification

This section investigates the control performances of the proposed MIPC and the classical DMPC, which validate the effectiveness of the proposed method. The system parameters are listed in Table 1.

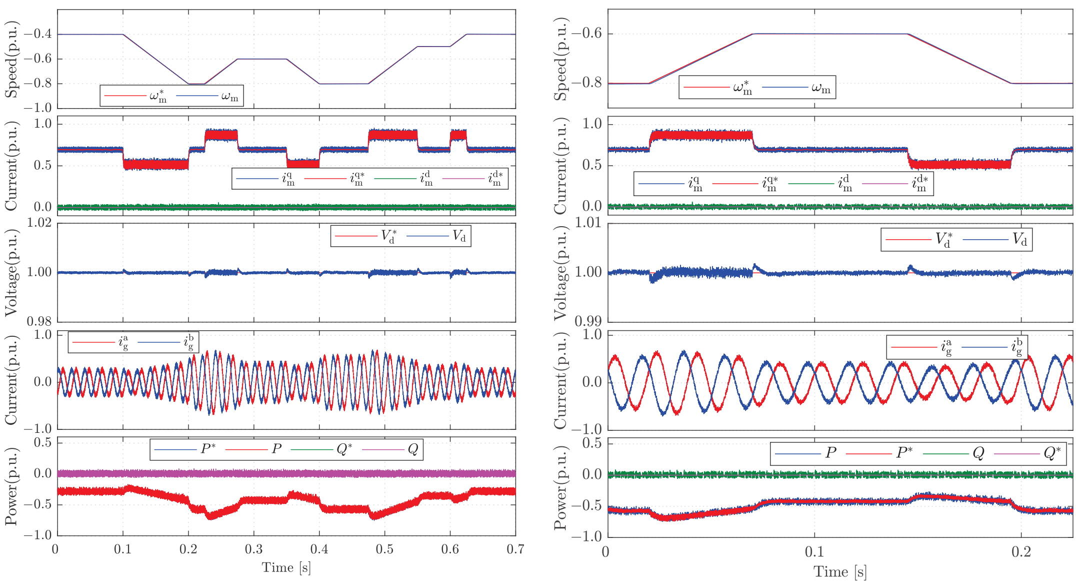
Table 1. System configuration.

Parameter	Value	Parameter	Value
DC-link Voltage V_{d}(V)	600	DC Capacitance C (F)	1100×10^{-6}
PMSG Inductance $L_{\mathrm{s}}^{\mathrm{d}} = L_{\mathrm{s}}^{\mathrm{q}}$ (H)	19.43×10^{-3}	PMSG Resistance R_s (Ω)	0.14
Nominal Torque $T_{\mathrm{e}}^{\mathrm{n}}$ (N · m)	29	Nominal Power P^{n} (kVA)	3.475
PM Flux ψ_{pm} (Wb)	0.43	PMSG Pole Pairs $\mathrm{N_p}$ (-)	3
PMSG Inertia J (kg · m^2)	0.01	Grid Voltage $\boldsymbol{e}_{\mathrm{g}}^{\mathrm{abc}}$ (V)	$210/\sqrt{2}$
Grid Frequency ω_{g} (rad/s)	100π	Filter Resistance R_{g} (Ω)	1.56×10^{-3}
Filter Inductance L_{g} (H)	16×10^{-3}	Sampling Time $\mathrm{T_s}$ (µs)	50

5.1. Overall Validation of the Proposed Method

The test at various operating points is conducted in order to confirm the effectiveness of the system's overall control. Depending on the wind speed, an optimal torque reference T_e^* or speed reference ω_{m}^* is determined by a proper "maximum power point tracking" (MPPT) control. While the reference generation techniques are not the central focus of this work, for simplicity, the test scenario is designed as follows: the "MPPT speed reference" ω_{m}^* has multiple changes, with a particularly steep slope to test the roughest conditions; the DC-link voltage reference V_{d}^* remains 600 V during the whole process; reactive power reference is set at 0 var to achieve unity power factor control.

Figure 3a shows the overall performances of the proposed method, and the zoomed performances are given in Figure 3b. The waveforms show that the proposed MIPC achieved good steady and transient state performances globally. The smooth and good tracking of speed, current, and power for both sides of the back-to-back converter is obtained, and the DC-link voltage remains stable in both the steady state and the transient state.



(**a**) (**b**)

Figure 3. Performance of the proposed MIPC method. (**a**) Overall control performance. (**b**) Zoomed control performance. From top to bottom are the PMSG mechanical speed, stator dq-axis currents, DC-link voltage, grid-side currents, active and reactive power, and their references, respectively. The speed base is 125 r/min. The current base is 15 A.

5.2. Robustness Comparison

Under various parameter mismatch conditions, the control performances of the proposed MIPC and the classical DPMC method are compared in this section. For a fair evaluation, the same test scenarios were created for both control methods in each condition.

In the first test, the permanent-magnet flux linkage in the controller is varied to 50% and 200% of the actual value (ψ_{pm}) to investigate the influence of flux variation. As can be seen from Figure 4, flux mismatch will mainly lead to torque tracking bias. For 50% flux error, the torque is 4.5% larger than the reference. For 200% flux error, the torque is 5.8% smaller than the reference. This phenomenon is in accordance with the analytical analysis presented in [16].The control performance of the proposed MIPC is unaffected for the permanent-magnet flux linkage variations, which is in line with the principle that ψ_{pm} is not introduced throughout the control process of this method.

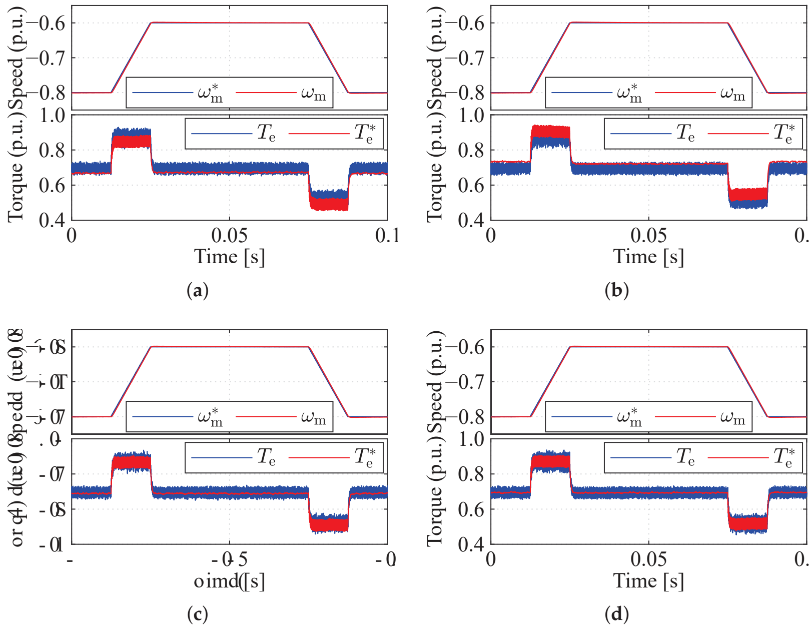


Figure 4. Performances under permanent-magnet flux linkage variations: (**a**) Classical DMPC (50% ψ_{pm}); (**b**) Classical DMPC (200% ψ_{pm}); (**c**) Classical DMPC (100% ψ_{pm}); (**d**) Proposed MIPC. For all sub-figures, from top to bottom are the generator speed (base value 125 [rad/s]) and the generator torque (base value 29 [N · m]), respectively.

In the controller of the classical DMPC solution, we set the inductance of the filter and PMSG stator to vary from 50% to 200% compared to the actual value in the wind turbine system plant. The results are shown in Figure 5 (generator side) and Figure 6 (grid side). Obviously, inaccurate inductance parameters will cause increased ripples (both dq-axis currents and active and reactive power) and enlarge current THDs. The stator current THD increased from 2.14% with the nominal parameter to 3.321% with 0.5 L_s and to 2.968% with 2 L_s. The grid-side performances in terms of current THD both exceed 4% with 0.5 L_g and with 2 L_g, which represents deterioration in power quality compared to the performance with the nominal parameter (3.712%). The control performances under various parameters' mismatches are all collected in Table 2. The results are in accordance with the theoretical analysis, i.e., the controller will derive the wrong prediction of the system utilizing mismatched parameters relating to the selection of the non-optimal control vectors. This will seriously affect the control performance. The control performance of the proposed MIPC is unaffected for all parameter variations, which is in line with the principle that no parameters are introduced throughout the control process of this method. In this section, we verify the good robustness of the proposed model-independent approach compared to the classical DMPC.

Table 2. Comparative test data of classical DMPC and proposed MIPC.

Control Method	Maximum Torque Error	Generator Current THD	Grid Current THD
Classical DMPC	4.50% (50% ψ_{pm})	3.32% (50% L_s)	4.05% (50% L_g)
	5.80% (200% ψ_{pm})	2.97% (200% L_s)	6.28% (200% L_g)
	0.73% (100% ψ_{pm})	2.15% (100% L_s)	3.71% (100% L_g)
Proposed MIPC	0.75%	2.09%	3.66%

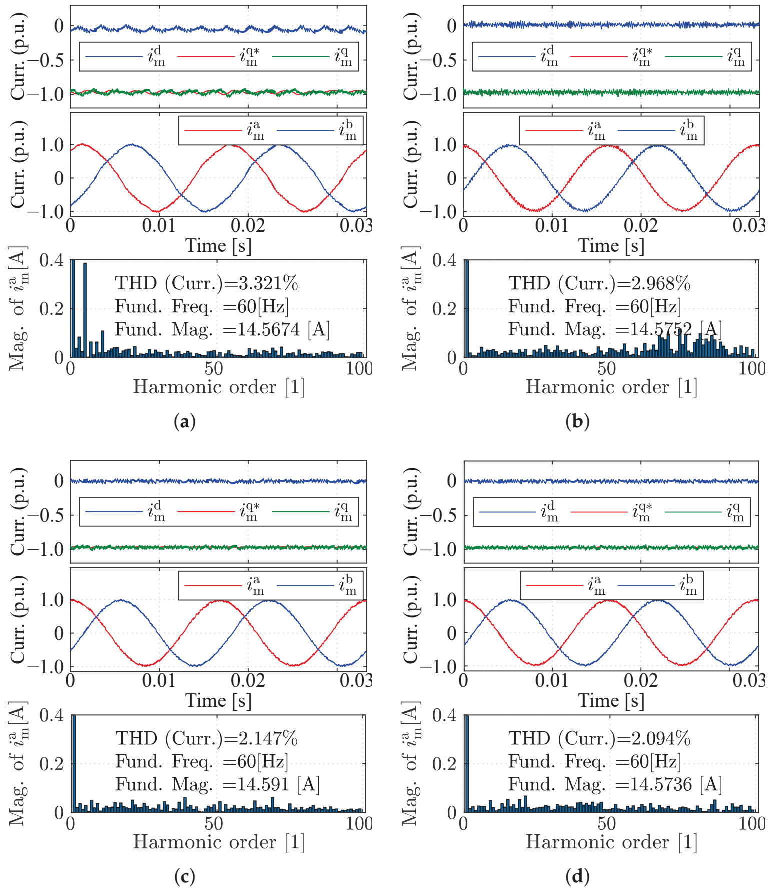


Figure 5. Generator side: (**a**) Classical DMPC (50% L_s); (**b**) Classical DMPC (200% L_s); (**c**) Classical DMPC (100% L_s); (**d**) Proposed MIPC. For all sub-figures, from top to bottom are the dq-axis stator current and their references (base value 15 [A]), phase currents, and the current spectrum, respectively.

In addition, the Classical DMPC and proposed MIPC under several simultaneous parameter mismatches are also tested. In the DMPC controller, the permanent-magnet flux linkage and the inductance of the PMSG stator are set to 50% as the actual value, while the filter inductance is set to 200% as the actual value. The performances are given in Figure 7. The tracking of torque and power as well as the current distortion show that the control performance of the proposed MIPC does not deteriorate even in the face of multiple mismatch at the same time and still shows higher robustness than the classical DMPC.

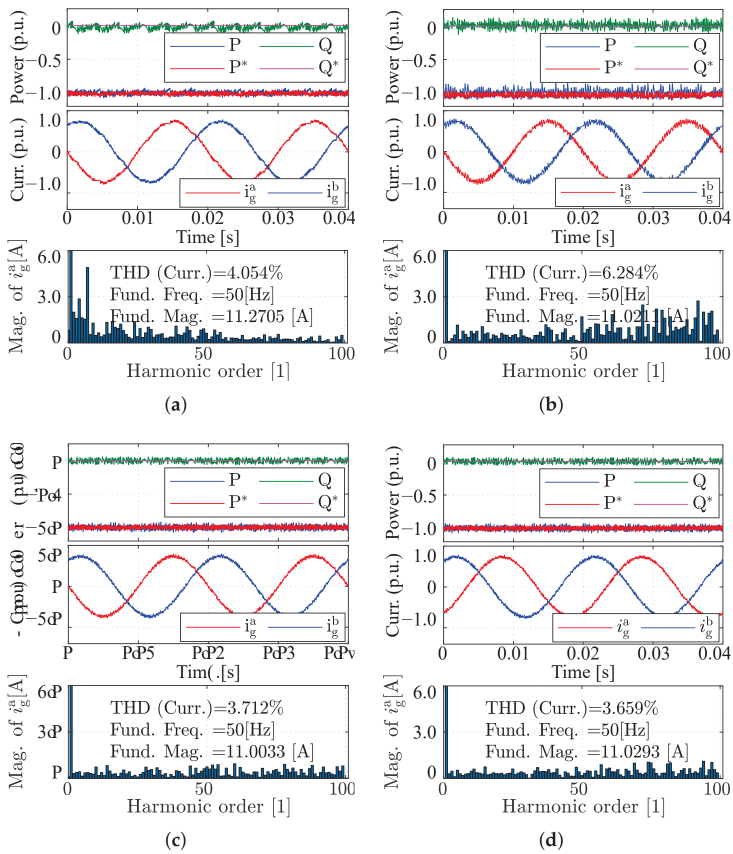


Figure 6. Grid side: (**a**) Classical DMPC (50% L_g); (**b**) Classical DMPC (200% L_g); (**c**) Classical DMPC (100% L_g); (**d**) Proposed MIPC. For all sub-figures, from top to bottom are the active and reactive power (base value 3475 [W]) and their references, phase currents, and the current spectrum (base value 15 [A]), respectively.

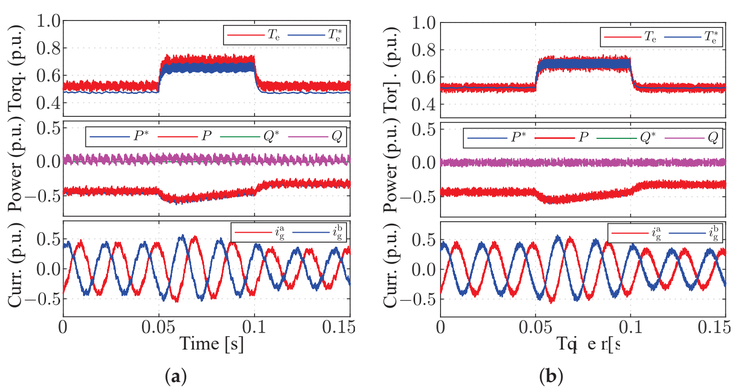


Figure 7. The control performance in the face of multiple mismatches: (**a**) Classical DMPC; (**b**) Proposed MIPC. For all sub-figures, from top to bottom are the generator torque (base value 30 [Nm]), the active and reactive power (base value 3475 [W]) and their references, and the phase currents on the grid side (base value 15 [A]), respectively.

5.3. Current/Power Update Mechanism Comparison between the Proposed and the Classical MIPC

System prediction is the key to the control performance of the predictive controller. In the proposed MIPC, the fast and accurate state variable variation estimation contributes to the accurate prediction of the system (see Figure 8). The proposed MIPC updates current and power variation for all possible voltage vectors during one control interval by means of the measured $k-1$ and $k-2$ instant values. Fast variable variation update frequency can be assured; see Figure 9a. The classical MIPC scheme in [24] updates the current and power variation only once for one voltage vector during the whole control interval. *Stagnant* current and power variation appear when one voltage vector is not applied for long consecutive control intervals; see Figure 9b. Comparing the sampled values of the system with the predicted values calculated using the LUTs in Figure 8, the two overlap, indicating that the proposed model-independent predictive control accurately predicts the trajectory of the system at future moments, which guarantees the control performance.

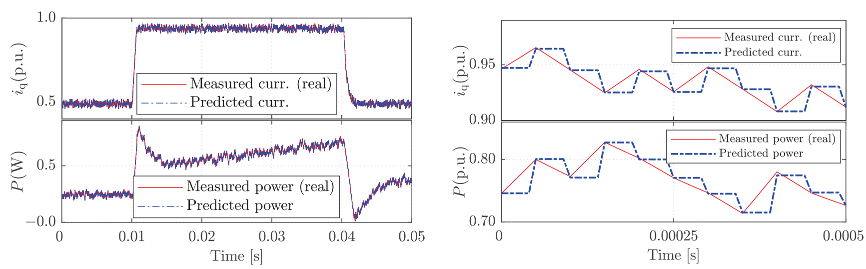


Figure 8. Prediction accuracy of the proposed MIPC. (**a**) Q-axis stator current and grid power prediction validation. (**b**) Zoomed comparison between measurement and prediction.

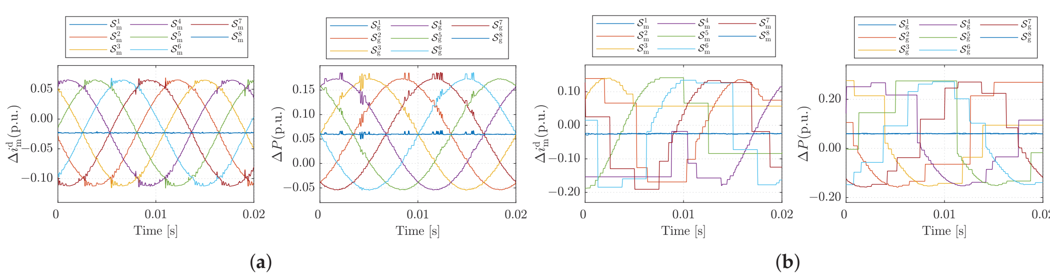


Figure 9. Estimated d-axis current variation and active power variation caused by different voltage vectors using (**a**) the proposed MIPC and (**b**) classical MIPC [24]. S_x^1–S_x^8 denote the available voltage vectors of the machine side ($x = m$) and the grid side ($x = g$), respectively.

6. Conclusions

Constrained by its complete dependence on the model, conventional direct model predictive control easily exhibits deterioration in control performance when the model parameters are mismatched. State-of-the-art model-independent predictive control (MIPC) introduces historical operation data in the prediction of future statuses. Nevertheless, it suffers from low look-up table (LUT) update frequency, unsteady state variable changing rate, and extensive computational burden. This work proposed an improved MIPC with a new look-up table update method that only introduces the information from the former two instants to estimate all needed variations in the same period. Compared with the traditional finite-set model predictive control and the existing MIPC, the proposed solution achieves robustness to unmeasurable and time-varying parameters without sacrificing control performance. The proposed method can be applied to other power converter topologies with minor modifications. Future work will focus on addressing measurement robustness and extending the proposed methods to multilevel power conversion systems.

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Annual Energy Production Design Optimization for PM Generators Considering Maximum Power Point Trajectory of Wind Turbines

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Abstract: Efficiency optimization is an important goal in the design of permanent magnet generators. However, traditional design optimization methods only focus on improving the rated efficiency without considering the annual cycle for overall efficiency improvement. To overcome this drawback, this paper presents a design optimization method for improving annual energy production (AEP) of wind direct-drive permanent magnet generators. Unlike the conventional efficiency optimization method that only improves the rated point efficiency, the proposed method improves the overall efficiency of the generator during the operating cycle by matching the maximum power point trajectory of the wind turbine. The periodic loss model of the permanent magnet generator is established and further constituted as the objective function to perform the optimization search using a genetic algorithm. Through simulation and experimental verification, the proposed method can obtain a higher AEP compared with the conventional design optimization method, and the proposed method can be extended to other variable speed power generation fields.

Keywords: annual energy production; permanent magnet synchronous generator; design optimization; maximum power point tracking; wind turbine

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1. Introduction

As a clean energy resource, wind power has received great research attention and achieved continuous development [1]. Because of their low failure rate and high-power-density, permanent magnet synchronous generators (PMSGs) are increasingly used in wind energy conversion systems (WECSs) [2]. In order to obtain high efficiency, reliable operation, and low cost-effectiveness of WECSs, the generator design optimization is an important aspect [3]. Power electronic converters allow synchronous generators to operate with variable speed while guaranteeing a constant electricity generation frequency [4]. In the period of variable speed operation, the PMSG needs to match the maximum energy trajectory of the turbine by regulating the mechanical speed under different operating conditions. Therefore, for PMSGs of WECSs, it is a new challenge to realize high efficiency operation under variable speed and load during the working cycle.

Currently, intelligent stochastic algorithms such as the genetic algorithm (GA) [5], evolutionary algorithm [6], and other algorithms [7] are commonly utilized to design and optimize the performances of PMSGs. They are mainly based on establishing an analytical or numerical link between structural parameters and generator performance, and then evolving the optimal combination of structural parameters to maximize or minimize the target value by random search and successive iterations.

The optimal design of a PMSG considering the uncertainty of raw material cost is proposed in [8] using a GA, and a hybrid process of GA and pattern search was used to optimize a 6 MW direct-drive PMSG to reduce the cost of energy [9]. In [10], a sensitivity analysis of structural parameters on the performance of a hybrid-excited dual-PM wind generator was carried out, and some leading parameters were identified and further optimized using a combination of finite elements and GA. High efficiency is a valuable optimization goal for generators to increase system energy production and reduce the temperature rise due to operating losses, which has been investigated in many studies [11,12].

However, most of the existing papers are only optimized for rated efficiency, i.e., the objective function only includes the generator efficiency at the rated point, without considering the improvement of the cycle operation efficiency. This may not be applicable to WECSs, where wind turbines are less likely to operate near the rated conditions due to climatic and seasonal influences. Therefore, the classical approach of design optimization with rated efficiency may instead lead to a lower AEP.

In [13], a generator's operation range is divided into four parts, these four parts are equated into four feature points, and the efficiencies of all four points are optimized collaboratively, allowing the overall efficiency to be enhanced. However, its operating conditions are relatively simple, with a fixed rotational speed of 4200 rpm. For variable speed WECS, the PMG speed needs to rise with increasing input power to improve the energy conversion efficiency, which increases the difficulty of design optimization.

In this paper, a new design optimization method is proposed to maximize the AEP of the PMSG considering the maximum power point trajectory (MPPT) of the wind turbine. First, the aerodynamic model of the wind turbine is introduced to develop the relationship between its MPPT and optimal shaft speed, and the wind speed distribution is investigated to obtain the feature operating points with multiple wind speeds. Secondly, the loss models of the PMSG with multiple feature points are formulated as the objective function, and the GA is utilized to improve the overall efficiency. Finally, the optimization scheme to improve the AEP is obtained and verified by the simulation and experiment.

2. Direct-Drive WECS

2.1. Wind Speed Distribution

With the elimination of gearboxes in direct-drive WECS, the efficiency and reliability of the system can be improved greatly. The system features a PMSG driven directly by the wind turbine and synchronizes the power generation with the grid frequency through a full-size AC/DC/AC converter, and is capable of variable speed operation in the full speed range, the structure of which is shown in Figure 1.

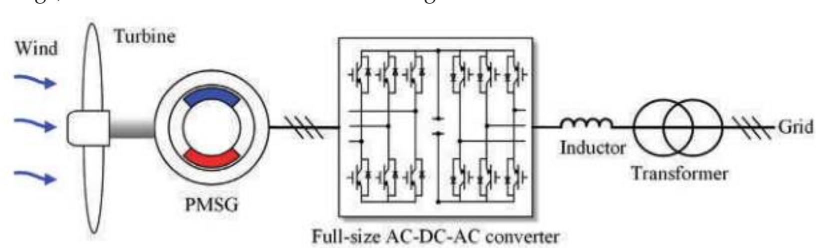

Figure 1. Structure of direct-drive WECS.

The wind speed of the wind farm conforms to the Rayleigh distribution [14], and the probability distribution is as follows:

$$p(V_\omega) = \frac{\pi}{2}\frac{V_\omega}{V_{\text{avg}}}\, e^{-\frac{\pi}{4}\left(\frac{V_\omega}{V_{\text{avg}}}\right)^2} \tag{1}$$

where V_ω is the wind speed, V_{avg} is the average value of the wind speed, and the annual average wind speed of the case study is about 8.5 m/s. The cut-in speed is 3 m/s, and the cut-out speed is 12 m/s. The cut-in and cut-out wind speeds are made according to the comprehensive situation of site conditions, the mechanical strength of the generation unit, operating condition characteristics, etc. On the basis of lightweight and large blade design, the cut-out wind speed is therefore relatively lower in this study.

Five feature points are selected to represent the overall distribution of wind speed for design optimization trade-offs, since too many feature points would lead to convergence failure and too few would not be representative. According to the calculation of the definite integral, the probability at each typical wind speed can be obtained as

$$\alpha_i = \frac{\int_{i_{\min}}^{i_{\max}} p(V_\omega)dV_\omega}{\int_3^{12} p(V_\omega)dV_\omega} \qquad i = 1,\ 2,\ \ldots,\ 5 \tag{2}$$

It can be calculated that α_1 is equal to 0.1017 (corresponding to the wind speed range of 3–4 m/s); α_2 is 0.2468, (4–6 m/s); α_3 is 0.2580 (6–8 m/s); α_4 is 0.2247 (8–10 m/s); and α_5 is 0.1688 (10–12 m/s). The plotted wind speed curve and the probability fitting results are shown in Figure 2.

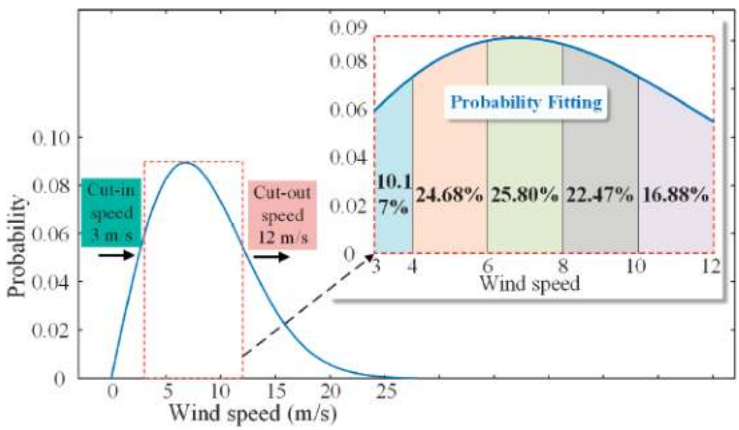

Figure 2. Wind speed Rayleigh distribution and the probability fitting of representative wind speeds.

2.2. Aerodynamic Model of Wind Turbines

The output power (mechanical power P_{mech}) of the wind turbine is given by [14]

$$P_{\text{mech}} = \frac{1}{2}\rho\pi R^2 V_\omega^3 C_p \tag{3}$$

where ρ is the air density, and R is the turbine radius. C_p is the wind turbine power coefficient, which is a function of the tip speed ratio λ and the pitch angle β, and it can be expressed as

$$\begin{gathered} C_p(\lambda,\ \beta) = 0.35\left(\tfrac{151}{\lambda_i} - 0.58\beta - 0.002\beta^{2.14} - 13.2\right)^{-\left(\frac{18.4}{\lambda_i}\right)} \\ \text{with } \tfrac{1}{\lambda_i} = \tfrac{1}{\lambda+0.08\beta} - \tfrac{0.035}{\beta^3+1} \text{ and } \lambda = R\tfrac{\omega_r}{V_\omega} \end{gathered} \tag{4}$$

where ω_r is the turbine rotational speed. The energy available to the generator is determined by both the wind speed as well as the turbine speed, and their relationship is shown in Figure 3.

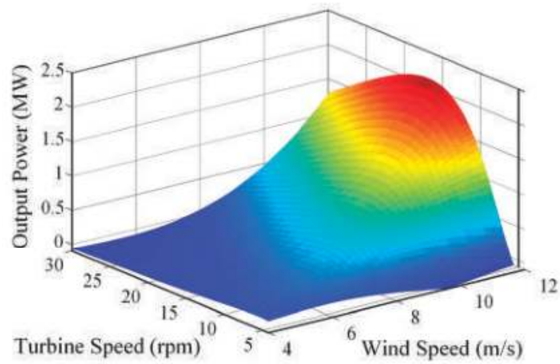

Figure 3. The three-dimensional characteristics of wind turbine energy conversion.

The variable-speed operation of the wind turbine is beneficial to improving the energy conversion efficiency of the system, from which the MPPT curve of the turbine can be obtained, as shown in Figure 4a, and the optimal turbine shaft speed corresponding to the five feature points can also be derived, as shown in Figure 4b.

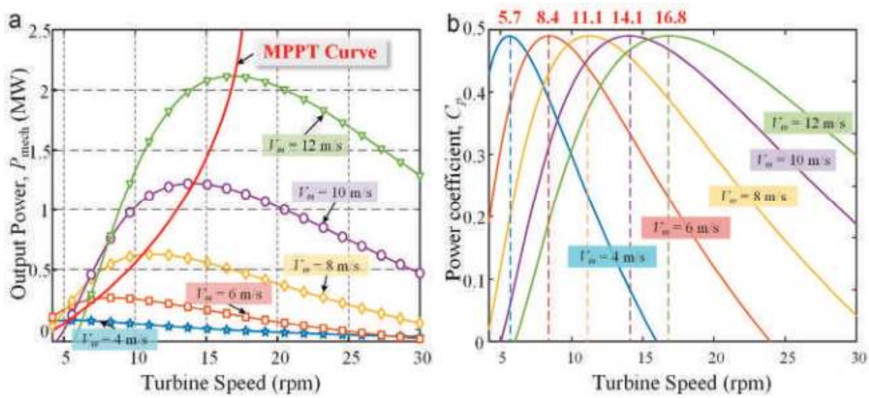

Figure 4. Operating characteristics of the wind turbine at representative wind speeds. (**a**) Output power vs. turbine speed. (**b**) Power coefficient vs. turbine speed.

Therefore, the actual operating conditions of the generator during the annual cycle can be predicted (including shaft speed, mechanical power, and operating hours) and are concisely represented as five feature operating points, as shown in Table 1. It should be noted that these characteristic points are the basis of the generator's cycle loss modeling and are closely related to the objective function in the optimization procedure.

Table 1. Feature operating points of the designed PMSG.

Point-i	Wind Speed-$V_{\omega.i}$ (m/s)	Optimal Shaft Speed-$\omega_{r.i}$ (rpm)	Input Power-$P_{int.i}$ (MW)	Weights-α_i (%)
1	4	5.7	0.1	10.17
2	6	8.4	0.3	24.68
3	8	11.1	0.6	25.80
4	10	14.1	1.2	22.47
5	12	16.8	2.1	16.88

3. Design Optimization

3.1. Optimization Process

Design optimization is performed by constructing a mathematical correlation between generator structural parameters and losses, and using GA to find the optimal weighted minimum value of losses at multiple feature points so as to obtain the structural parameters of PMSG that achieve the maximum cycle efficiency. The optimization process is shown in Figure 5.

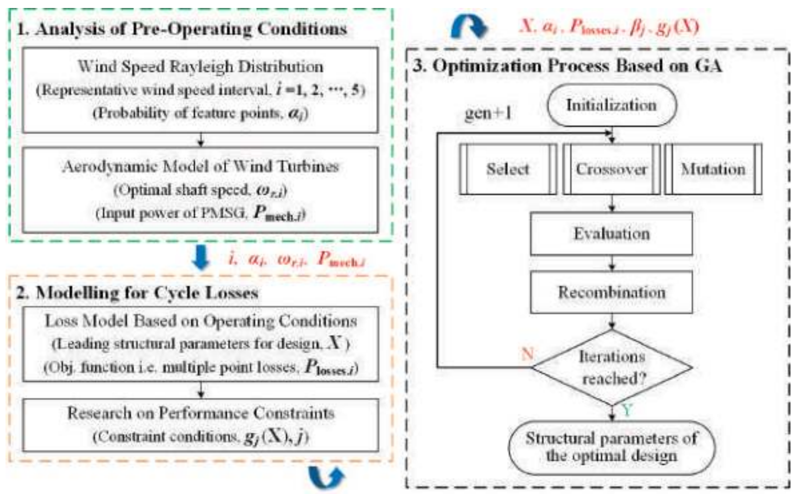

Figure 5. Flowchart of cycle efficiency design optimization.

As seen from the workflow, the optimization is divided into: (1) operating conditions pre-analysis, (2) cycle loss calculation, and (3) the stochastic algorithm optimization process, of which the operating conditions pre-analysis has been introduced in Section 2. The rest will be introduced in this section, and the optimization process is completed when the number of iterations reaches a set value.

3.2. Modelling for Cycle Losses

The accuracy of the loss model of the PM generators is crucial, and directly determines the validity of the optimization results. The model can provide a straightforward link between structural parameters, operating conditions, and losses.

The losses of permanent magnet synchronous generators come from many sources, and the total losses are expressed as follows [15]

$$P_{\text{losses}} = P_{\text{cu}} + P_{\text{iron}} + P_{\text{edPM}} + P_{\text{wind}} + P_{\text{fr}} \tag{5}$$

where P_{cu} is the copper Joule loss of stator windings, P_{iron} is the iron core loss including the stator and rotor core, P_{edPM} is the eddy current loss of PMs, P_{wind} is the windage loss, and P_{fr} is the friction loss.

First, the Joule loss of copper windings is the main component of PM machine loss, which can be expressed as [16]:

$$P_{\text{cu}} = \int_{V_{\text{w}}}^{\text{Tot vol}} \rho \boldsymbol{J}^2 dV \tag{6}$$

where $\boldsymbol{J}$ is the current density, indicating the winding current density per unit area, and is affected by the skin effect and proximity effect; the current density in the winding is not uniformly distributed, especially under high frequency current; V_{w} is the copper area; and ρ is the electrical resistivity, which is related to the material properties and proportional to the temperature.

$$\rho = \rho_{20}(1 + \alpha_T(T_{\mathrm{w}} - 20)) \tag{7}$$

where ρ_{20} is the resistivity at 20 °C, about 1.724×10^{-8} [Ohm·m] for pure copper, α_T is the temperature coefficient, which equals about 0.003862 per degree C, T_{w} is the winding temperature, and can be solved analytically by the thermal equivalent circuit method or numerically, where the analytical method has a significant computational speed advantage and the numerical method has a more reliable computational accuracy. In order to speed up the optimization process, the thermal network method of ANSYS Motor-CAD is used in this paper. During the iteration of the algorithm, electromagnetic and thermal calculations are required in each design case, and an example of the thermal network calculation diagram for the maximum operating condition of the generator under the natural cooling method is shown in Figure 6.

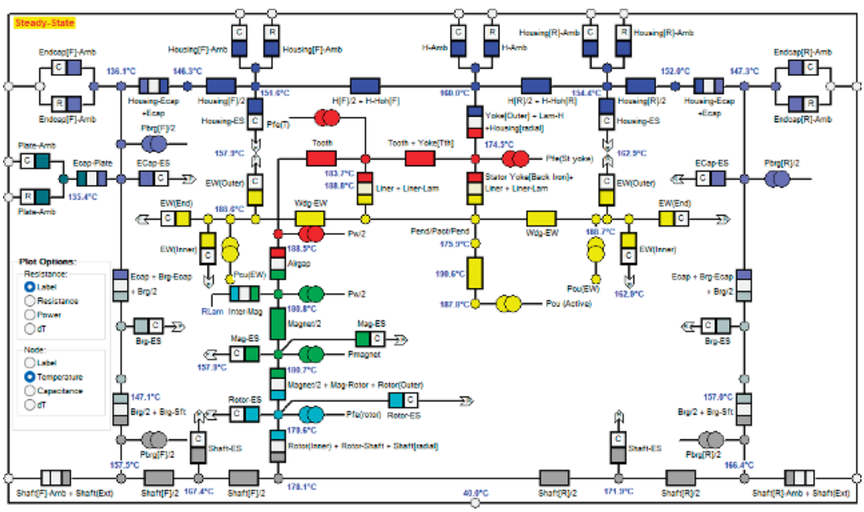

Figure 6. Diagram of thermal network circuit solving.

Next, the core loss is also the main component of the PM motor loss, especially in the case of high electrical speed and high magnetic saturation. This accounts for a significant proportion, and the core loss is expressed as follows:

$$P_{\mathrm{iron}} = P_{\mathrm{ironH}} + P_{\mathrm{ironE}} \tag{8}$$

where P_{ironH}, P_{ironE} are the iron hysteresis and eddy-current loss, respectively, and they are generated in the stator and rotor iron core, and dominated in the stator. They can be expressed as [16].

$$P_{\mathrm{ironH}} = \int_{V_{\mathrm{ic}}}^{\mathrm{Tot\ vol}} \left(\sum_{n=1}^{N} k_h n \omega_r B_n(n)^2 \right) dV \tag{9}$$

$$P_{\mathrm{ironE}} = \int_{V_{\mathrm{ic}}}^{\mathrm{Tot\ vol}} \left(\sum_{n=1}^{N} k_c (n\omega_r)^2 B_n(n)^2 \right) dV \tag{10}$$

where V_{ic} is the volume of the iron core, k_h and k_c are the coefficients of hysteresis and eddy-current losses, n is the harmonic order of iron flux density, and $B_n(n)$ is the amplitude of the nth harmonic order of flux density. It can be seen that the iron core loss is mainly related to the parameters such as core flux density, operational speed, and iron core volume, etc.

In addition, as a surface-mounted PM machine structure, the eddy current loss of the PMs is also not negligible, and it can be calculated as [17].

$$P_{\mathrm{edPM}} = \frac{n\omega_r}{\pi} \int_{0}^{\frac{2\pi}{\omega_r}} \int_{R_r}^{R_m} \int_{-\frac{\alpha p}{2}}^{\frac{\alpha p}{2}} \rho_{PM} J_m^2 r dr d\theta dt \tag{11}$$

where ρ_{PM} is the electrical resistivity of PMs, R_r, R_m are the radius of the inner and outer PM, respectively, and J_m is the induced eddy current in PMs due to the time-varying armature reaction field, which can be calculated by the 2D finite-element method.

Furthermore, P_{wind} and P_{fr} are the windage and friction losses of the PM generator, respectively, and they are related to the generator structure and the operating speed, etc. Their calculation formula are as follows [18]:

$$P_{\mathrm{wind}} = 2k_{\mathrm{wd}} D_{ro}^2 l_{\mathrm{ef}} \omega_r^3 10^{-6} \tag{12}$$

$$P_{\mathrm{fr}} = k_{\mathrm{fb}} m_r \omega_r 10^{-3} \tag{13}$$

where k_{wd}, k_{fb} are the factors of the windage and friction loss, respectively. D_{ro} is the rotor diameter; l_{ef} is the effective length of the core; m_r is the rotor mass. It can be seen that almost all types of losses are closely related to the generator speed ω_r and the input power of the generator.

According to the investigation of the wind turbine operating conditions in Section 2.1, the loss model with multiple feature points can be developed, and the cycle loss function $P_{\mathrm{cyc}}(X)$ is as follows:

$$P_{\mathrm{cyc}}(X) = \sum_{i=1}^{5} \alpha_i P_{\mathrm{losses}.i}(X) \tag{14}$$

where X is the set of design variables, and $P_{\mathrm{losses}.i}$ is the total loss for each feature point.

3.3. Relevant Parameters for Optimization

The fixed parameters of the PMSG to be designed through the preliminary research are shown in Table 2. The material used for the stator and rotor is a silicon steel sheet with 50 mm thickness, and the material used for the PMs is NdFeB of the brand N38UH and has a residual magnetic density of 1.26 T at 20 °C.

Table 2. Fixed parameters of the designed PMSG.

Symbol	Parameter	Value
P	Rated power (kW)	2100
p/Q_{S}	Poles/slots	60/288
ω_{M}	Rated speed (rpm)	16.8
V_{R}	Rated voltage (V)	660
V_{dc}	DC bus voltage (V)	1100

In order to improve the operating efficiency of the PMSG at multiple wind speeds, a stochastic algorithm is used to find the optimal solution between multiple efficiency objective functions. Furthermore, multiple constraints are considered to rationalize the optimization process. The processing of the objectives and constraints is as follows:

$$F(X) = \begin{cases} P_{\mathrm{cyc}}(X) & \text{Meet the constraints} \\ P_{\mathrm{cyc}}(X) + \sum_{j=1}^{6} \beta_j g_j(X) & \text{Does not meet the constraints} \end{cases} \tag{15}$$

where $g_j(X)$ is the constraint function derived by the design constraints, including the flux density, current density, power density, and slot filling factor; and β_j is the penalty coefficient of the constraint conditions. Table 3 shows the parameters of the constraint conditions.

Table 3. Parameters of constraint conditions.

Parameter	Range	Parameter	Range
Power density (kW/kg)	$\geq$0.1	Current density (A/mm^2)	$\leq$3.5
Slot filling factor (%)	$\leq$80	Airgap flux density (T)	$\leq$0.8
Tooth flux density (T)	$\leq$1.85	Yoke flux density (T)	$\leq$1.65

Eight design variables with specific ranges are selected in this paper, as shown in Table 4. Figure 7 shows the PMSG structure and the geometric description of the design variables.

Table 4. Parameters of the design variables.

Symbol	Parameter	Range
D_{i1}	Stator inner diameter (mm)	3500–4500
L_1	Core length (mm)	1200–1800
h_S	Slot depth (mm)	80–120
w_T	Tooth width (mm)	15–30
h_Y	Stator yoke thickness (mm)	30–70
α_P	Pole-arc coefficient	0.5–1
h_M	PM thickness (mm)	15–30
h_δ	Air-gap thickness (mm)	4–10

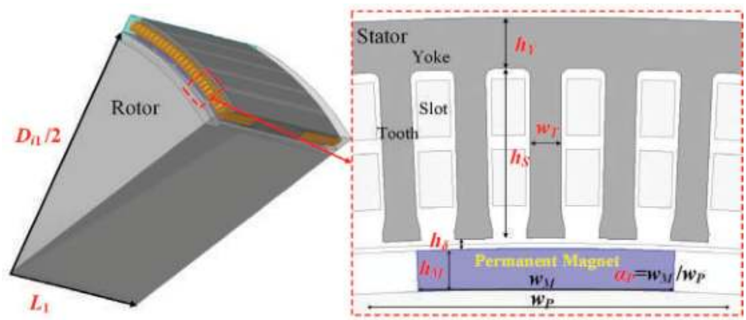

Figure 7. Generator structure and its design parameters.

4. Results and Verification

4.1. Results from Algorithm Calculation

In order to demonstrate the effectiveness of the proposed method for cycle efficiency improvement, a comparison experiment based on rated efficiency optimization (REO) is provided in this paper. The parameters of the two optimization schemes are the same except for the difference of the objective function (the objective function of the comparison experiment is the rated loss of the generator). The genetic algorithm was used to optimize the two optimization schemes in turn. For the trade-off between convergence and computational time, the number of iterations is set to 200 in this paper, and it can be seen in Figure 8 that multiple efficiencies have converged. The two optimization schemes are optimized by GA in turn. Each population of GA contains 500 individuals, and the number of generations is 200. After the iterative calculations, the variation of the fitness values for each feature operating condition is shown in Figure 8. The optimized design parameters of the proposed method are shown in Table 5.

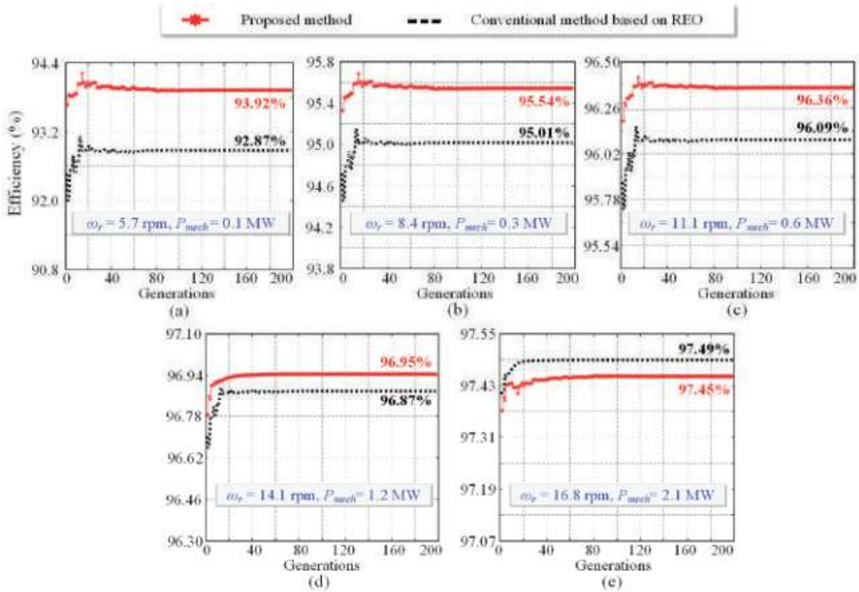

Figure 8. Iterative optimization of the proposed method and the conventional method based on REO. (**a**) Generator efficiency at the wind speed of 4 m/s. (**b**) Generator efficiency at the wind speed of 6 m/s. (**c**) Generator efficiency at the wind speed of 8 m/s. (**d**) Generator efficiency at the wind speed of 10 m/s. (**e**) Generator efficiency at the wind speed of 12 m/s.

Table 5. Optimized design parameters.

Symbol	Value	Symbol	Value
D_{i1}	3778.2 mm	h_Y	40.6 mm
L_1	1261.8 mm	α_P	0.779
h_S	96.7 mm	h_M	19.8 mm
w_T	20.5 mm	h_δ	6.2 mm

From the comparison in Figure 8, it can be seen that the conventional optimization method is effective in optimizing the rated point efficiency, which is 97.49% for a wind speed of 12 m/s, 96.87% for 10 m/s, 96.09% for 8 m/s, 95.01% for 6 m/s, and 92.87% for a wind speed of 4 m/s, while its cycle-weighted efficiency is 95.91%. On the other hand, the proposed optimization method has a higher cycle efficiency, which is 97.45% for a wind speed of 12 m/s, 96.95% for 10 m/s, 96.36% for 8 m/s, 95.54% for 6 m/s, 93.92% for a cycle-efficiency of 4 m/s, and a cycle-weighted efficiency of 96.23%. In summary, the proposed method has a higher annual cycle efficiency than the conventional REO-based method, with a weighted average efficiency of 0.32% higher.

4.2. Simulation Verification

The optimized design parameters were brought into the finite-element electromagnetic Ansys Maxwell software, and the simulated performances of the designed PMSG under various operating conditions is shown in Table 6. The results show that the performance indicators are well within the constraint conditions, and the generator has high operating efficiency under different operating conditions.

Table 6. Simulation performance of the optimized PMSG.

Performance	Value	Efficiency	Value
Power density	0.1 kW/kg	At wind speed of 4 m/s	94.32%
Current density	2.3 A/mm^2	At wind speed of 6 m/s	95.62%
Slot fill. factor	79.8%	At wind speed of 8 m/s	96.29%
Flux density in airgap/tooth/yoke	0.79 T/1.82 T/1.58 T	At wind speed of 10 m/s	96.84%
Power factor	0.97	At wind speed of 12 m/s	97.41%

4.3. Experimental Verification

The designed PMSG was fabricated and assembled, and its performance was tested on a dragging experimental platform. The photos of the test site and the performance demonstration are shown in Figure 9. The efficiencies obtained from the tests and model calculations at different speeds and powers are shown in Table 7. The results show that the experimental data are closer to the model calculations, and the effectiveness of the proposed method can be verified.

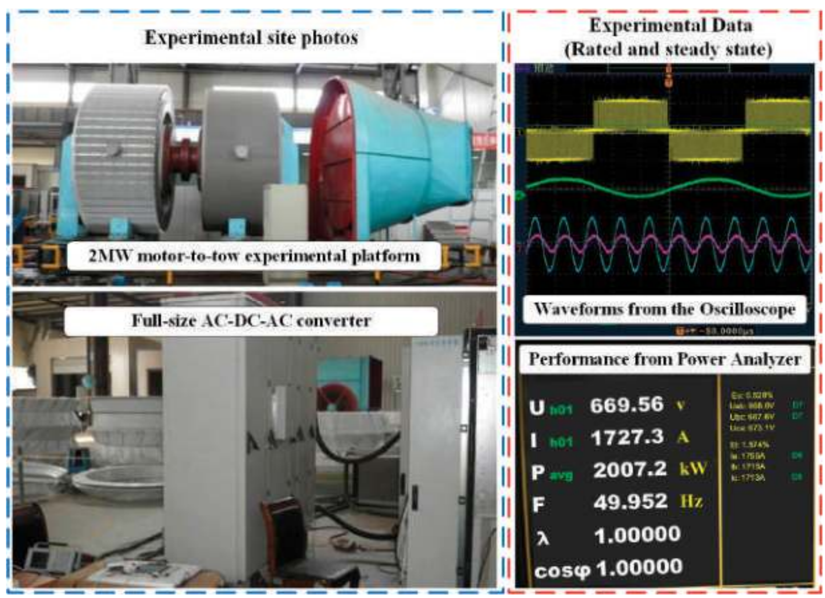

Figure 9. Experimental site photos and performance demonstration.

Table 7. Comparison of the efficiency data between model calculations and experiments.

	ω_r=5.7, P_{mech}=0.1	ω_r=8.4, P_{mech}=0.3	ω_r=11.1, P_{mech}=0.6	ω_r=14.1, P_{mech}=1.2	ω_r=16.8, P_{mech}=2.1
Model	93.92%	95.54%	96.36%	96.95%	97.45%
Test	94.01%	95.81%	96.64%	97.13%	97.67%

5. Conclusions

This paper presented a design optimization method for direct-drive wind PMSGs to improve the AEP of WECSs. The proposed method considered the optimal trajectory of the wind turbine and optimized the efficiency of several feature operating points in an integrated manner, and the method was compared with the conventional design optimization method based on the rated point efficiency. The computational results of the algorithm showed that the proposed method effectively improved the annual weighted efficiency by 0.32%, and the designed generator was validated by simulation and an experimental platform.

The proposed method is not only applicable to the design of wind turbines, but also has implications for the design and optimization of other electric machines with variable speeds and variable operating conditions.

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Permanent Magnet Flux Linkage Analysis and Maximum Torque per Ampere (MTPA) Control of High Saturation IPMSM

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Abstract: The maximum torque per ampere (MTPA) control is significant for improving the efficiency of the interior permanent magnet synchronous motor (IPMSM). However, for the high saturation IPMSM, the change of the permanent magnet (PM) flux linkage is more complicated, which can cause the MTPA control to deviate from the optimal solution. Therefore, an improved MTPA control method for the high saturation IPMSM is proposed in this paper. Compared with other methods, the proposed method improves the conventional models of flux linkage and torque by analyzing the nonlinear variation of the PM flux linkage with the *dq*-axis currents. Subsequently, an expression suitable for the MTPA control of high saturated IPMSM is derived based on the improved models. The proposed parameter fitting models are then fitted using data from 11 operating points and incorporated into the MTPA optimization algorithm to obtain the MTPA curve. Finally, the effectiveness of the proposed method in enhancing the control accuracy of the MTPA angle is verified through simulations and experiments.

Keywords: maximum torque per ampere; high saturation; permanent magnet flux linkage; interior permanent magnet synchronous motor; nonlinear fitting model

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1. Introduction

Interior permanent magnet synchronous motor (IPMSM) has been an attractive choice in practical applications due to its excellent features of high efficiency and high power density, such as robotics, elevators, air conditioners, and compressors [1,2]. Meanwhile, due to the asymmetry of the *dq*-axis magnetic circuits, the reluctance torque and the magnet torque exist simultaneously in the electromagnetic torque of the IPMSM. For a given output torque, there are different combinations of the *dq*-axis currents [3]. In order to fully utilize the reluctance torque through the optimal combination of *dq*-axis currents, the maximum torque per ampere (MTPA) control has become a preferred control strategy of the IPMSM. The purpose of the MTPA control is to trace an optimal current angle under a given output torque to minimize the stator current amplitude, and the optimal current angle is known as the MTPA angle [4–24].

The conventional method regards the parameters of the motor as constant values in the *dq*-axis model. An optimized equation is then used to calculate the MTPA angle for the different torque [4]. However, the inductances will change nonlinearly with the current, which affects the reluctance torque. The permanent magnet (PM) flux linkage will change with the temperature, which affects the magnet torque [5–7]. Hence, when the parameters of the motor are regarded as constants to implement the MTPA control, the real-time optimal distribution of the *dq*-axis currents cannot be achieved. For the problem of parameter uncertainty in MTPA control, many scholars have proposed various methods to identify the changing parameters under different operating loads. For instance, in Ref. [8], the finite element method is used to obtain the variation of parameters with current. In

Ref. [9], the high frequency injection method is employed to identify the parameters during motor operation. Based on different parameter identification methods, MTPA control can be categorized into offline and online methods.

The offline methods are used to obtain the MTPA curve before the motor is operated in an actual environment. In Ref. [10], the finite element method is used to obtain the MTPA curve after the motor design is completed. Although the finite element method is accurate in theory, and the MTPA angle can be found, it requires knowing the detailed mechanical dimensions of the motor. Furthermore, some technical and assembly errors usually exist in the prototype manufacturing process, resulting in a mismatch between the actual and simulated values of the motor parameters [11]. The look-up table (LUT) method can solve the shortcomings of the finite element method [12–14]. In Ref. [14], the optimal *dq*-axis currents are measured corresponding to different working conditions through experiments; the data are saved to the controller storage space, and the optimal given values are accurately indexed according to different working conditions during the IPMSM operation. However, the LUT method requires a time-consuming experimental process and a large number of experimental data.

To compensate for the shortcomings of offline methods, online methods can usually be used. Refs. [15–17] present the search-based methods to online search the optimal current angle corresponding to the minimum stator current by injecting a signal. Although the search-based methods do not depend on the *dq*-axis model, the response performance of the system is unsatisfactory, and the algorithm may fail in the process of torque variation, thus affecting the stability of the system. In Refs. [18–22], the parameters of the motor are estimated by the high frequency signal injection or advanced algorithm, and the estimated parameters are brought into the MTPA algorithm to obtain the MTPA angle. However, Ref. [23] has proved that although the parameter estimation method can compensate for the problem of parameter inaccuracy, each optimization iteration needs to update the parameters rather than directly consider the changes of parameters in the optimization algorithm. This problem will still lead to deviation in the calculated MTPA angle. In Refs. [23–26], the MTPA methods based on surface fitting are proposed. The proposed fitting models of the flux linkage and the inductance are integrated into the MTPA algorithm, which avoids updating parameters outside the MTPA algorithm.

However, the above methods all ignore the effect of the high saturation on the PM flux linkage. For the high saturation IPMSM, due to the restriction of weight and volume, its power density and saturation degree are much higher than those of commercial motors. Hence, in the actual operation process, the change of the motor parameters is more complicated than that of the low saturation motor [27–29]. Ref. [28] presents that the vector direction of the PM flux linkage will change with the saturation degree of the motor. Ref. [29] presents that the amplitude of the PM flux linkage will change nonlinearly with the *dq*-axis currents. These variations of the PM flux linkage will also cause the MTPA control to deviate from the optimal solution, while the above methods ignore these variations. Hence, the above methods are not applicable to the high saturation IPMSM.

To avoid these problems and develop an accurate MTPA control for the high saturation IPMSM, this paper proposes an improved MTPA control method based on a modified *dq*-axis model, and the organization is as follows. The conventional *dq*-axis model and the MTPA control method are described in Section 2. The nonlinear change of PM flux linkage with the current and the influence of PM flux linkage on torque are analyzed, and the proposed MTPA is explicated in Section 3. In Sections 4 and 5, the simulation and experimental results are presented to verify the proposed method. Finally, the conclusions are given in Section 6.

2. Conventional IPMSM Model and MTPA Control

In the conventional *dq*-axis model of the IPMSM, the PM flux linkage and the inductance are usually regarded as constants, and the flux linkage equations are denoted as

$$\begin{bmatrix} \psi_d \\ \psi_q \end{bmatrix} = \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} \psi_f \\ 0 \end{bmatrix} \tag{1}$$

where ψ_d, ψ_q, i_d, i_q, L_d, and L_q are the *d*-axis and *q*-axis flux linkages, currents, and inductances, respectively; ψ_f is the PM flux linkage. The total torque of the motor is expressed by

$$Te_{total} = \frac{3}{2}p(\psi_d i_q - \psi_q i_d) \tag{2}$$

where p is the pole pair; Te_{total} is the total torque. Substituting (1) into (2), the total torque can then be divided into the magnet torque caused by the PM flux linkage and the reluctance torque caused by the inductance difference.

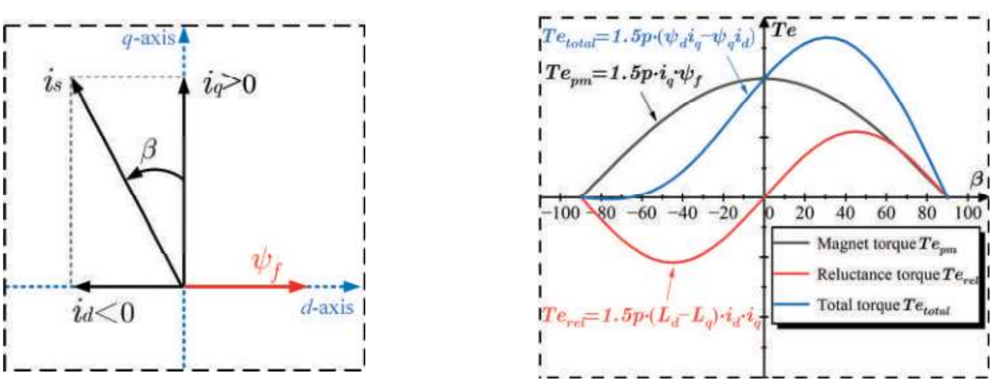
$$Te_{total} = Te_{pm} + Te_{rel} = \frac{3}{2}pi_q\psi_f + \frac{3}{2}p(L_d - L_q)i_d i_q \tag{3}$$

where Te_{pm} is the magnet torque; Te_{rel} is the reluctance torque.

In addition, the relationships between the phase current and the *dq*-axis currents are shown in Figure 1a, and the equations are expressed as

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} -i_s & 0 \\ 0 & i_s \end{bmatrix} \begin{bmatrix} \sin\beta \\ \cos\beta \end{bmatrix} \tag{4}$$

where i_s is the maximum phase current; β is the current angle between the phase current vector and the *q*-axis. The relationships between each torque and the current angle are shown in Figure 1b.



(**a**) (**b**)

Figure 1. (**a**) Relationships between phase current and *dq*-axis currents. (**b**) Relationships between each torque and current angle.

Substituting (4) into (3), the torque Equation (2) can be expressed as

$$Te_{total} = \frac{3}{2}p(\psi_f i_s \cos\beta - \frac{1}{2}L_\Delta i_s^2 sin2\beta),\ L_\Delta = L_d - L_q \tag{5}$$

The purpose of the MTPA control is to trace a current angle to make the ratio of total torque and current maximize, which is expressed as

$$\max_{\beta} \frac{Te_{total}}{i_s} \tag{6}$$

To find an appropriate current angle β from (6), make the partial derivative of the total torque expression (5) with respect to the current angle be zero $\partial Te_{total}/\partial\beta = 0$.

$$\frac{\partial \psi_f}{\partial \beta} cos\beta - \psi_f \sin\beta - \frac{1}{2}\frac{\partial L_\Delta}{\partial \beta} i_s \sin 2\beta - L_\Delta i_s \cos 2\beta = 0 \tag{7}$$

The conventional MTPA control method regards the inductance and the PM flux linkage as constants. Hence, the first and third terms on the left side of (7) are zero, and the current angle can be obtained by

$$\beta = \sin^{-1} \frac{\psi_f \sqrt{\psi_f^2 + 8L_\Delta^2 i_s^2}}{4L\Delta i_s} \tag{8}$$

However, the inductance and the PM flux linkage will change with different operating conditions of the motor, resulting in a significant error between the MTPA angle calculated by Equation (8) and the actual MTPA angle. Moreover, in the conventional *dq*-axis model, the change of the PM flux linkage with saturation is not considered; this will further increase the error.

3. PM Flux Linkage Analysis and Proposed MTPA Control

3.1. Proposed IPMSM Model and Fitting Models

For the high saturation IPMSM, the magnetic density of the motor is high, which will lead to the PM flux linkage change nonlinearly with the *dq*-axis currents. When i_d remains unchanged, the core is gradually magnetized with the increase of i_q, which will cause the vector direction of the PM flux linkage ψ'_f to gradually shift to the negative direction of the *q*-axis, and the *q*-axis component of the PM flux linkage ψ'_{fq} will be generated, as shown in Figure 2 (superscript$'$ indicates that magnetic saturation is considered).

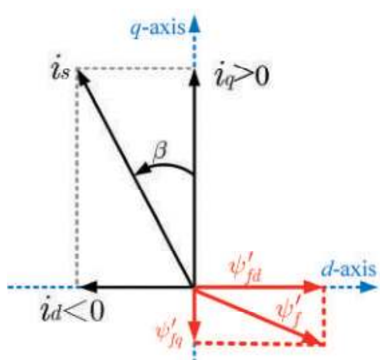
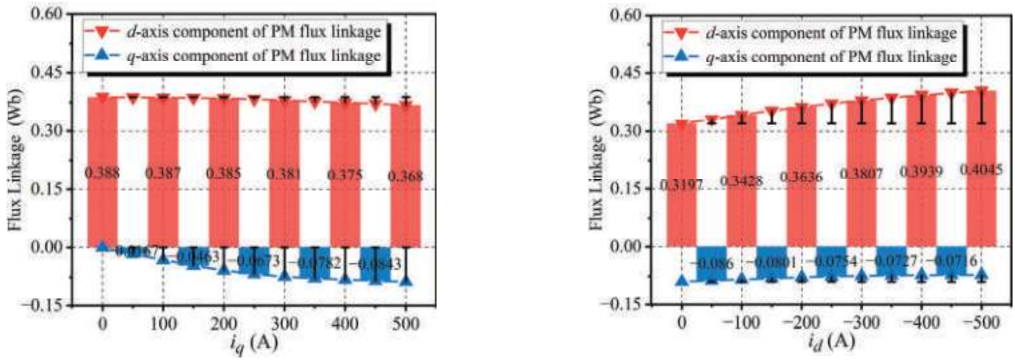


Figure 2. Vector direction of PM flux linkage.

With the increase of i_q, the saturation of the core increases, and the inclination degree of PM flux linkage also increases, which causes the *d*-axis component of the PM flux linkage ψ'_{fd} to decrease and ψ'_{fq} to increase along the negative direction of the *q*-axis, as shown in Figure 3a. Meanwhile, when i_q remains unchanged, with the decrease of i_d, due to the influence of the demagnetization current, desaturation will occur in the *d*-axis direction, which causes ψ'_{fd} to increase and ψ'_{fq} to decrease along the negative direction of the q-axis, as shown in Figure 3b.



(**a**) (**b**)

Figure 3. Variation of PM flux linkage: (**a**) Increase i_q when i_d remains unchanged. (**b**) Decrease i_d when i_q remains unchanged.

Therefore, not only the inductance changes but also the change of PM flux linkage should be considered in the flux linkage equations, and (1) should be rewritten as

$$\begin{bmatrix} \psi'_d \\ \psi'_q \end{bmatrix} = \begin{bmatrix} L'_d & 0 \\ 0 & L'_q \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} \psi'_{fd} \\ \psi'_{fq} \end{bmatrix} \tag{9}$$

where ψ'_{fd} and ψ'_{fq} are the dq-axis components of the PM flux linkage, respectively. Then, the magnet torque can also be divided into the dq-axis components, and (3) is rewritten as

$$\begin{aligned} Te'_{total} &= Te'_{pm} + Te'_{rel} = Te'_{pmd} + Te'_{pmq} + Te'_{rel} \\ &= \tfrac{3}{2} p \psi'_{fd} i_q - \tfrac{3}{2} p \psi'_{fq} i_d + \tfrac{3}{2} p L'_{\Delta} i_d i_q \end{aligned} \tag{10}$$

where Te'_{pmd} and Te'_{pmq} are the dq-axis components of the magnet torque, respectively. The relationships between each torque and current angle are shown in Figure 4.

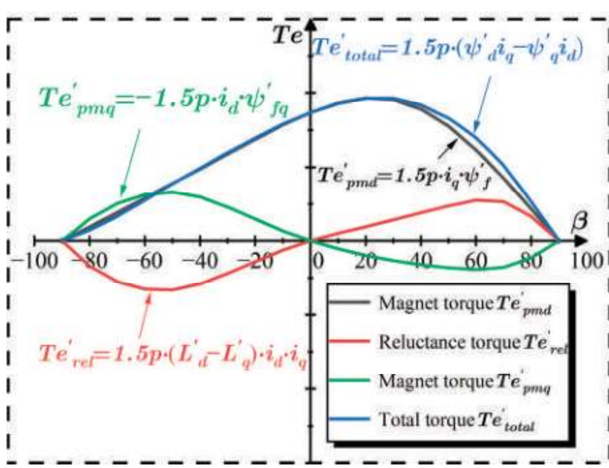


Figure 4. Relationships between each torque and current angle.

Figure 4 presents that when the current angle is 0 to 90 degrees ($i_d \leq 0$, $i_q \geq 0$), Te'_{pmq} will cause Te'_{total} to decrease, while the MTPA control needs to trace the appropriate current angle according to the change of Te'_{total} (6). Therefore, in the actual MTPA control, the influence of Te'_{pmq} on Te'_{total} cannot be ignored. It means that for the MTPA control of high saturation IPMSM, it is necessary to consider the changes of the PM flux linkage amplitude and vector direction.

Substituting (4) into (10), make the partial derivative of Te'_{total} with respect to β be zero.

$$\begin{aligned}\frac{\partial Te'_{total}}{\partial \beta} &= \frac{\partial \psi'_{fd}}{\partial \beta} cos\beta - \psi'_{fd}\sin\beta - \frac{1}{2}\frac{\partial L'_{\Delta}}{\partial \beta} i_s \sin 2\beta \\ &-L'_{\Delta} i_s \cos 2\beta + \frac{\partial \psi'_{fq}}{\partial \beta}\sin\beta + \psi'_{fq}\cos\beta = 0\end{aligned} \tag{11}$$

According to ref [23], $\partial\psi'_{fd}/\partial\beta$, $\partial\psi'_{fq}/\partial\beta$, and $\partial L'_{\Delta}/\partial\beta$ cannot be ignored in the MTPA algorithm (11). Thus, this paper proposes the fitting models of ψ'_{fd}, ψ'_{fq}, and L'_{Δ} in regard to the dq-axis currents, respectively.

$$\begin{aligned}&L'_{\Delta} = a_0 + a_1 i_d + a_2 i_q + a_3 i_d i_q \\ &\psi'_{fd} = b_0 + b_1 i_d + b_2 i_q + b_3 i_d i_q + b_4 i_q^2,\ \psi'_{fq} = c_0 i_q + c_1 i_d i_q + c_2 i_q^2\end{aligned} \tag{12}$$

where a_i, b_i, and c_i are constant coefficients (i = 0, 1, 2, 3, 4).

3.2. Determination of the Coefficients and Implementation of the MTPA Control

This paper only considers the motoring mode of the IPMSM. Assuming that the dq-axis currents ranges are $-i \leq i_d \leq 0$ and $i \geq i_q \geq 0$, plot four quarter circles with amplitudes of 1/4, 1/2, 3/4, and 1 of the current i, respectively. Divide the quarter circle with radius i into four sectors equally, and select points from 1 to 11. Point 1 is on a circle with a radius of 0.25i; points 2 and 3 are on a circle with a radius of 0.5i; points 4, 5, and 6 are on a circle with a radius of 0.75i; points 7, 8, 9, 10 and 11 are on a circle with a radius of i; and each point is on the boundary of the sector. The specific coordinates of each point are shown in Figure 5.

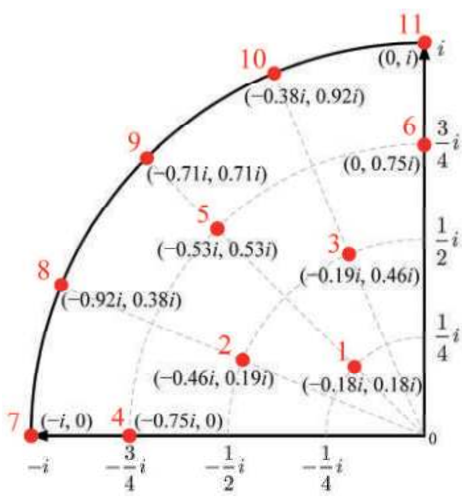


Figure 5. The specific coordinates of each point.

Points 4, 7, and 8 are selected to occupy more weight in the desaturation area of the motor, and points 6, 10, and 11 are selected to occupy more weight in the saturation area of the motor. The inductance parameters of 11 points can be firstly obtained by parameter estimation. Then, ψ'_{fd} and ψ'_{fq} at each point can be calculated by using the steady-state voltage Equation (13) according to the monitored dq-axis voltage.

$$\psi'_{fd} = \frac{v_q - R_s i_q}{w_e} - L'_d i_d \,, \psi'_{fq} = \frac{R_s i_d - v_d}{w_e} - L'_q i_q \tag{13}$$

where R_s is the stator resistance; v_d and v_q are the dq-axis voltage, respectively; w_e is the electrical speed.

Finally, the coefficients in (12) can be determined based on the data of the 11 operating points in Figure 5 and the proposed optimization algorithm (14).

$$\begin{gathered}\text{Minimize :}\\ \sum_{i=1}^{11} (L^*_\Delta - L^\wedge_\Delta)^2 + \sum_{i=1}^{11} (\psi^*_{fd} - \psi^\wedge_{fd})^2 + \sum_{i=1}^{11} (\psi^*_{fq} - \psi^\wedge_{fq})^2\end{gathered} \tag{14}$$

where L^*_Δ, ψ^*_{fd}, and ψ^*_{fq} are the actual value; $L^\wedge_\Delta$, $\psi^\wedge_{fd}$ and $\psi^\wedge_{fq}$ are the fitted value.

After determining the coefficients of the fitting models, the total torque expression without the PM flux linkage and the inductance can be expressed as

$$Te'_{total} = \frac{3}{2} p(b_0 i_q + k i_d i_q + b_2 i_q^2 + m i_d i_q^2 + b_4 i_q^3 + n i_d^2 i_q + a_3 i_d^2 i_q^2) \tag{15}$$

where

$$k = b_1 + a_0 + c_0 \,, m = b_3 + a_2 + c_2 \,, n = a_1 + c_1 \tag{16}$$

Substituting (4) into (15), make the partial derivative of Te'_{total} with respect to β be zero.

$$\begin{aligned}\frac{\partial Te'_{total}}{\partial \beta} &= k i_s \sin^2\beta - b_0 \sin\beta - k i_s \cos^2\beta - 2b_2 i_s \cos\beta \sin\beta \\ &- m i_s^2 \cos^3\beta + 2m i_s^2 \sin^2\beta \cos\beta - 3b_4 i_s^2 \cos^2\beta \sin\beta \\ &+ 2n i_s^2 \cos^2\beta \sin\beta - n i_s^2 \sin^3\beta + 2a_3 i_s^3 \sin\beta \cos^3\beta \\ &- 2a_3 i_s^3 \sin^3\beta \cos\beta = 0\end{aligned} \tag{17}$$

It can be seen from (17) that the equation has only two variables (i_s, β), which can be solved by numerical solution or nonlinear equation solver. For real-time applications, the proposed algorithm can also be utilized to generate LUT in advance to improve the dynamics of the control system. Compared with the conventional LUT method, the proposed method can generate the LUT offline only by using a PC without sampling the entire working range of the motor, which significantly reduces the workload. Eventually, the dq-axis currents trajectories can be obtained by substituting the current angle β into (4). This method not only considers the changes of the inductance in the optimization process but also considers the changes of the PM flux linkage in the high saturation state. The control block diagram is shown in Figure 6.

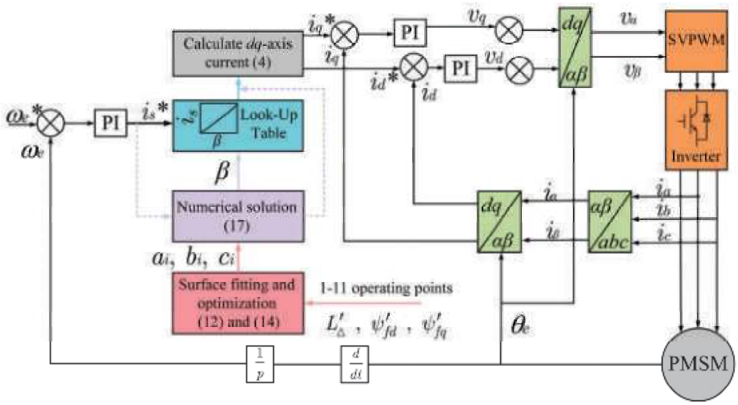


Figure 6. The control block diagram(* represents the corresponding set value).

4. Simulation Analysis and Verification

To verify the accuracy of the proposed method, this paper establishes a finite element model (FEM) of the high saturation IPMSM, as shown in Figure 7. The main parameters of the model are listed in Table 1, including the *dq*-axis inductance and the PM flux linkage when the *dq*-axis currents are zero.

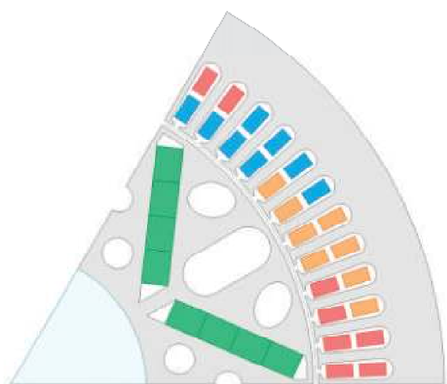
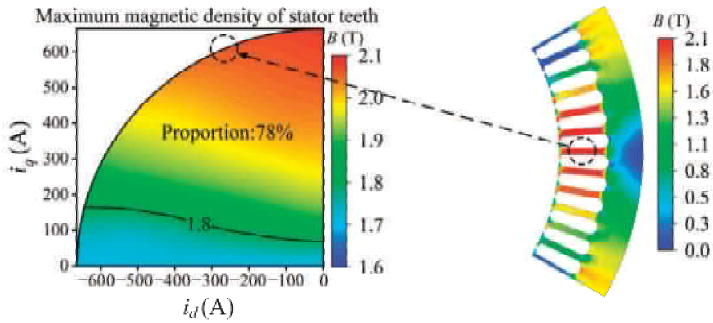


Figure 7. The FEM of IPMSM.

Table 1. Key parameters of IPMSM.

Parameters	Quantity	Parameters	Quantity
Rated power	300 kW	Pole pairs	3
Rated current	470 A	PM flux linkage	0.332 Wb
Rated speed	3000 r/min	Stator resistance	0.005 Ω
Rated torque	955 Nm	*d*-axis inductance	0.4723 mH
Rated frequency	150 Hz	*q*-axis inductance	1.0228 mH

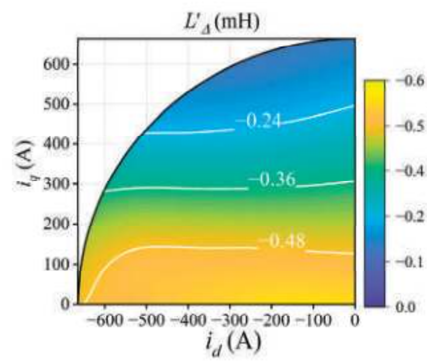
For the high saturation IPMSM at the different load conditions, the stator core is usually in the high saturation state. Especially, the maximum magnetic density (MMD) of stator teeth is generally greater than 1.8 T. To verify the saturation degree of stator teeth, the motor speed is set to the rated speed. Then, the MMD of the stator teeth is simulated at the different load conditions, as shown in Figure 8a.



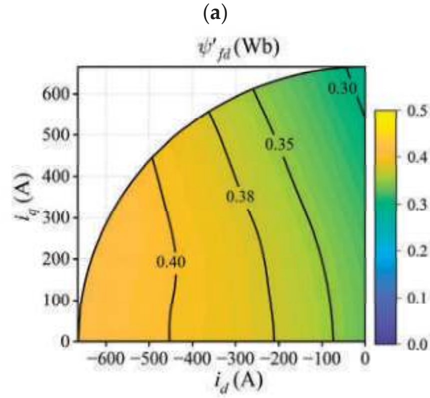
(**a**) (**b**)

Figure 8. Saturation degree of IPMSM. (**a**) MMD of stator teeth at different load conditions. (**b**) Distribution of stator magnetic density.

Figure 8a presents that the MMD of the stator teeth exceeds 1.8 T for 78% of the rated current range. When $i_d = -275$ A and $i_q = 605$ A, the IPMSM can operate at the rated torque with the minimum current. Meanwhile, the MMD of the stator teeth is 2.05 T, as shown in Figure 8b. Thus, Figure 8 indicates that the saturation of the IPMSM model is high. In the high saturation state, the inductance and the PM flux linkage will change nonlinearly with the dq-axis currents, as shown in Figure 9.

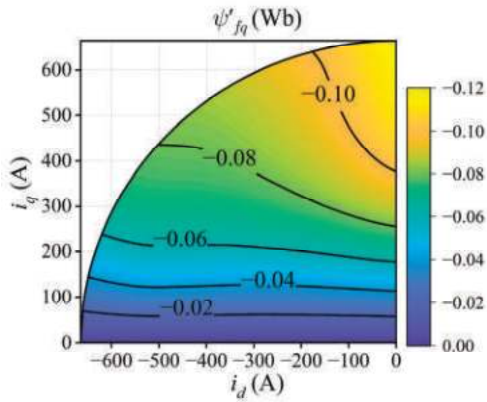


(**a**)



(**b**)

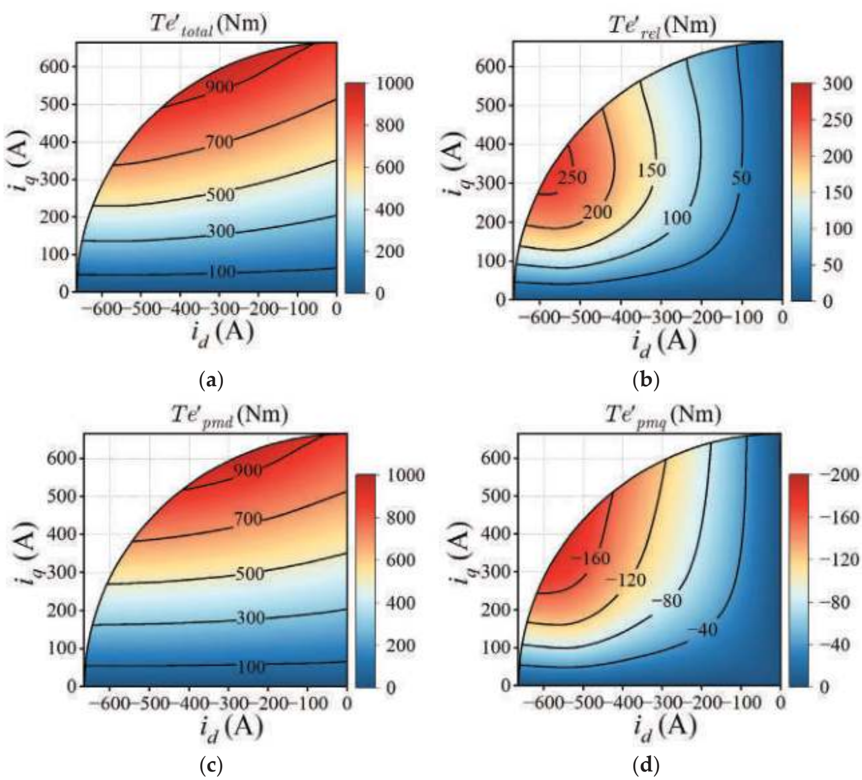
Figure 9. *Cont.*



(**c**)

Figure 9. Variation of parameters. (**a**) Inductance difference. (**b**) *d*-axis component of PM flux linkage. (**c**) *q*-axis component of PM flux linkage.

According to the analysis in Section 3, the variation of the inductance and PM flux linkage will directly affect the amplitude of the reluctance and magnet torque, respectively, as shown in Figure 10.



(**a**) (**b**) (**c**) (**d**)

Figure 10. Variation of each torque within rated current range. (**a**) Total torque. (**b**) Reluctance torque. (**c**) *d*-axis component of magnet torque. (**d**) *q*-axis component of magnet torque.

Figure 10 presents that for the high saturation IPMSM, the total torque is composed of three components. Each torque component significantly impacts the total torque, and all of them change nonlinearly with the change of the *dq*-axis current. For this reason, the variation of these three torque components should be considered simultaneously in the MTPA optimization algorithm. According to the coordinates of 11 working points in Figure 5, the simulated inductance and the PM flux linkage are substituted into (14), and the coefficients of fitting models are obtained, as shown in Table 2.

Table 2. Results of coefficient fitting based on simulation.

Coefficient	a_i	b_i	c_i
$i = 0$	-5.55×10^{-4}	0.342	-3.91×10^{-4}
$i = 1$	-1.12×10^{-8}	-1.32×10^{-4}	-1.23×10^{-7}
$i = 2$	6.23×10^{-7}	-4.37×10^{-5}	3.43×10^{-7}
$i = 3$	-1.46×10^{-10}	-1.34×10^{-7}	0
$i = 4$	0	-5.41×10^{-8}	0

Then, the coefficients are substituted into (17), and the equation results are directly substituted into (4) by numerical solution, as shown in Figure 6. Finally, compared with other methods, (1) the conventional method substitutes the constant parameters at no-load condition into (8); (2) the parameter estimation method [22] substitutes the variable parameters at different conditions into (8); (3) the conventional fitting method [23]. The MTPA curves obtained by different methods are shown in Figure 11. It is obvious from Figure 11 that the proposed method is closer to the actual MTPA curve than other methods, which verifies the validity of the analysis and the method proposed in this paper.

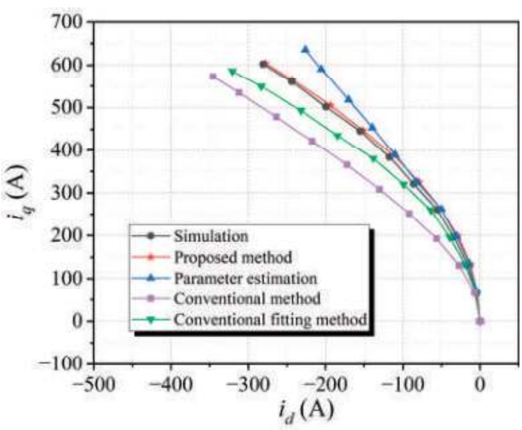


Figure 11. Comparison of proposed method and other methods.

5. Experimental Results

A 300 kW high saturation IPMSM has been prototyped to verify the proposed approach, and the design parameters of the prototype are given in Table 1. The experiment platform is shown in Figure 12a. The prototype is driven by a traction motor at 3000 rpm and operates in the torque control mode. Meanwhile, a torque sensor (ATESTEO: DF3) is also installed to measure the torque at different load conditions. In addition, in order to reduce the influence of temperature on parameters, so as to better verify the influence of high saturation on parameter, the cooling mode of the prototype is water-cooling, the temperature of cooling water is 30 °C, and the water flow is 60 L/min. The Hi-Techniques' high-speed logger is utilized to observe and reserve the experimental information in real time, and the PC software is used to export and plot the experimental data, as shown in Figure 12b.

Figure 12. Experimental platform. (**a**) Prototype and driving system. (**b**) Instrument for recording data.

Firstly, the prototype is tested to find the actual MTPA curve. The detailed test steps are as follows. Set i_d to make the prototype operate at no-load condition. Then, as the load gradually increases, the voltage, current, and torque are recorded at different load conditions. Finally, gradually decrease i_d and repeat the previous steps. The measured torque within the rated current range is shown in Figure 13, and the MTPA curve is *OB*.

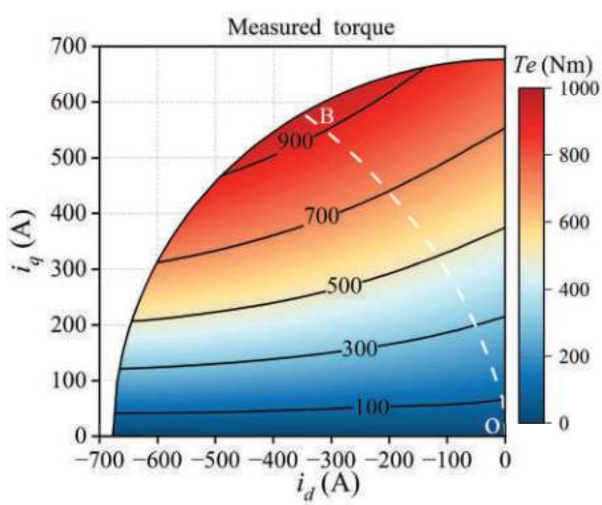


Figure 13. Measured torque and MTPA curve.

Due to certain technical and assembly errors in the prototype manufacturing process, the actual value and the simulated value are different. For instance, the minimum current at the rated torque is different from the simulation. When the IPMSM is operated at rated torque, the dq-axis currents are i_d = −350 A and i_q = 580 A, respectively. The recorded waveforms of the DC bus voltage, three-phase voltage, current, and torque are shown in Figure 14.

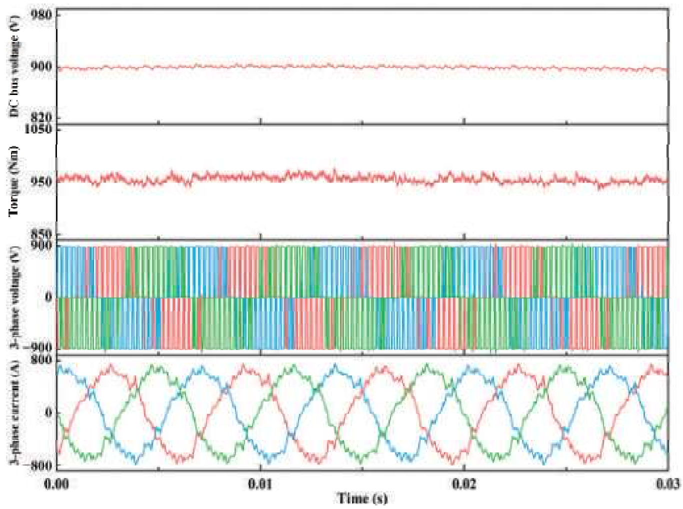


Figure 14. Test waveforms at rated torque.

Then, the parameter estimation method is used to estimate the inductance parameters of the 11 operating points, and the *dq*-axis components of PM flux linkage can be calculated according to the monitored *dq*-axis voltage, as shown in Figure 15. According to the algorithm (14), the fitting coefficients of the prototype are listed in Table 3.

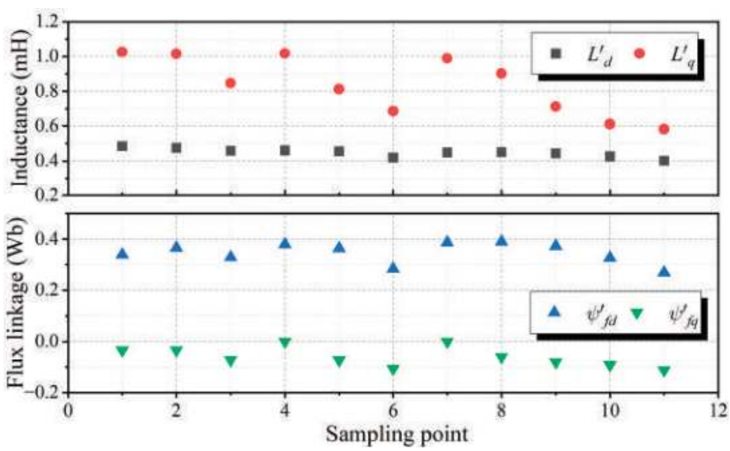
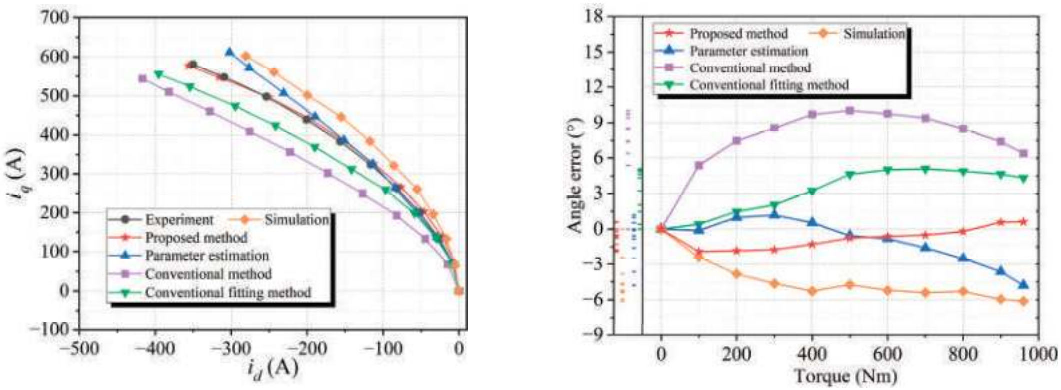


Figure 15. Parameters of 11 operating points.

Table 3. Results of coefficient fitting based on experiments.

Coefficient	a_i	b_i	c_i
$i = 0$	-6.28×10^{-4}	0.317	-3.34×10^{-4}
$i = 1$	-1.13×10^{-7}	-6.19×10^{-5}	-9.24×10^{-8}
$i = 2$	6.97×10^{-7}	-1.1×10^{-4}	2.54×10^{-7}
$i = 3$	1.92×10^{-10}	-2.3×10^{-7}	0
$i = 4$	0	-8.23×10^{-9}	0

Based on the fitting coefficients of the prototype, the offline numerical calculation is used to obtain the relationship between i_s and β, and generates a LUT for real-time applications, as shown in Figure 6. Eventually, the MTPA curves and MTPA angle errors of the different methods are shown in Figure 16.



(**a**) (**b**)

Figure 16. Experimental results. (**a**) MTPA curves; (**b**) MTPA angle errors.

Figure 16b presents that the error of the proposed method is relatively stable, and the error does not increase with the increase of the saturation degree. For example, when the motor is operated at the rated load, although the saturation degree of the motor is large, the error of the proposed method is still small. The errors of different methods at the rated load are listed in Table 4.

Table 4. Comparison of different methods.

Methods	Errors at Rated Load
Proposed	0.6°
Simulation	−6.1°
Conventional	6.4°
Conventional fitting	4.3°
Parameter estimation	−4.8°

In the rated load range, the maximum errors caused by the parameter estimation and conventional fitting method are −4.8° and 5.1°, respectively. The main reason for these errors is that the influence of the high saturation on the PM flux linkage is not considered. Furthermore, the maximum error of the conventional method and the simulation are 10° and −6.1°, respectively. The maximum error of the proposed method is −1.8°. Therefore, the proposed method can achieve better accuracy compared with other methods. Meanwhile, these experimental results indicate that for the MTPA control of high saturation IPMSM, the nonlinear change of the PM flux linkage with the dq-axis current cannot be ignored.

6. Conclusions

The MTPA control is a regular method for the IPMSM to find an optimal current angle to minimize the stator current amplitude for given output torque. However, for the high saturation IPMSM, the change of the motor parameters is more complicated than that of the low saturation motor, and the nonlinear relationship between the PM flux linkage and the dq-axis currents becomes more distinct. When the conventional MTPA control method is used, the accuracy of the MTPA angle often fails to meet the requirements.

Thus, an improved MTPA control method is proposed in this paper. By analyzing the change of the PM flux linkage and the influence of the PM flux linkage on the total torque, the conventional flux linkage and torque models are improved. Based on the improved *dq*-axis models, an expression more suitable for MTPA control of high saturation IPMSM is derived. Compared to other methods, the proposed method takes into account the nonlinear variation of the PM flux linkage with magnetic saturation. Additionally, this method also considers the *q*-axis component generated by the PM flux linkage and the partial derivatives of the parameters in the MTPA optimization algorithm. Simulation and experimental results demonstrate that the proposed method can better follow the MTPA angle of the high saturation IPMSM at different load conditions.

Author Contributions: Methodology, C.L.; software, W.Z.; validation, J.G.; formal analysis, S.H. All authors have read and agreed to the published version of the manuscript.

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Adaptive Control of a Virtual Synchronous Generator with Multiparameter Coordination

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Abstract: This paper proposes an adaptive strategy of co-regulating the three parameters—P/ω droop coefficient, virtual inertia, and damping coefficient—for the virtual synchronous generator (VSG). This approach is able to solve the uncoordinated performance between the virtual inertia and the damping using the conventional adaptive control in which the system may experience serious frequency fluctuations. Through the mathematical modeling of the VSG grid-connected system, the segmental analysis of the VSG transient process is carried out, and the parameter adjustment law of each stage is obtained. The VSG angular velocity change and the angular velocity instantaneous change rate are associated with the inertia to realize the adaptive adjustment of the inertia, and the adaptive adjustment of the P/ω droop coefficient is carried out in real time according to the VSG angular velocity change. A functional relationship is established between the P/ω droop coefficient, virtual inertia, and damping coefficient so that the P/ω droop coefficient, virtual inertia, and damping coefficient are coordinated to keep the system in the best damping ratio state all the time. Finally, the superiority of the proposed strategy is proved by simulation comparison.

Keywords: virtual synchronous generator grid-connected system; adaptive co-regulation strategy; optimal damping ratio

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1. Introduction

With the increasingly serious environmental pollution and the exhaustion of traditional fossil energy, clean energy, such as wind and light, has become an indispensable alternative [1]. New energy power generation is generally connected to the AC microgrid through the inverter device, but since the inverter device does not have the inertia and damping of the synchronous generator when it is connected in a large proportion, the inertia and damping of the system will be insufficient. When disturbed, its ability to suppress interference becomes weak and even causes the system frequency to collapse in severe cases [2,3].

The virtual synchronous generator control [4] simulates the inertia and damping characteristics of the synchronous generator, while the inverter equipment can also provide inertia and damping support for the system. Virtual inertia and damping are the core control parameters of VSG, which are flexible and adjustable, and proper adjustment of these parameters can effectively improve the control performance of VSG. With the increase in the proportion of new energy connected to the power system, VSG technology has received more and more attention from researchers. Ref. [5] proposes a virtual inertial control strategy for the microgrid system with energy storage, which effectively improves the frequency characteristics of the microgrid system. However, when the proportion of new energy inside the system changes, its inertia takes a fixed value, and the frequency response characteristics of the system deteriorate. To solve this problem, adjustment techniques based on adaptive strategies have been proposed by researchers [6–13]. Ref. [8]

proposed a VSG virtual inertia adaptive control algorithm based on stick control. When the rate of change of the angular frequency is less than a certain threshold, the inertia takes a smaller value; otherwise, it takes a larger value. However, stage adjustment and the optimal tracking of inertia to frequency changes cannot be achieved. Ref. [9] proposes a virtual inertia adaptive control strategy that is jointly determined by the VSG rotor angular frequency change rate and deviation, which solves the adaptive effective tracking to a certain extent, but the selection range of virtual inertia and the basis for the selection of key parameters are not given in the paper. Refs. [11,12] proposed a coordinated adaptive control strategy of inertia and damping. When the VSG angular frequency change rate increases, the inertia value is higher, and when the angular frequency deviation increases, the damping value is higher, which effectively improves the dynamic and static state of the system frequency performance; however, only the adaptive expression is given, and the selection basis of the correlation coefficient in the expression cannot be given. Ref. [13] proposed a cooperative adaptive control strategy for VSG parameters. The strategy uses an exponential adaptive algorithm to determine the virtual inertia, which reduces the sensitivity of the relevant control parameters in the adaptive algorithm and combines the performance index constraints to achieve the coordination of the damping coefficient, which realizes the optimal tracking of inertia and damping for frequency changes. However, the paper ignores the effect of the P/ω droop coefficient. The literature [14] proposes an adaptive inertial damping integrated control (SA-RIDC) method, which decides whether to adjust the virtual inertia or the damping coefficient according to the derivative of the angular frequency differential during frequency oscillations in order to achieve alternating control of the virtual inertia and the damping coefficient. The literature [15] proposes a control strategy based on a joint adaptive virtual rotational inertia and damping coefficient with an optimal damping ratio. The literature [16] proposes an adaptive VSG control strategy for battery energy storage systems to ensure the stability of the power system. The literature [17] draws on the work angle and angular frequency curves of synchronous generators to design a refined virtual inertia fuzzy regulation law, while considering the four performance indicators of active overshoot, the frequency rate of change, regulation time, and rise time to select a suitable damping ratio, and the virtual damping is coordinated and adaptively adjusted according to the selected damping ratio with the change of virtual inertia. In the paper [18], virtual synchronous generator technology was introduced in the IIDG control system to optimize the active-frequency control, the reactive-voltage control, and the voltage-current control, respectively. The impact of the damping coefficient on the output of the microgrid system is analyzed by means of a small signal model, and a self-adaptive damping control strategy is proposed.

In this paper, an improved VSG multiparameter optimal cooperative control strategy is proposed. Based on the optimal damping ratio, the initial value of the virtual inertia and the adaptive control of the P/ω droop coefficient are set, and combined with the performance index constraints, the coordination between the P/ω droop coefficient, inertia coefficient, and damping coefficient is realized. This not only realizes the optimal tracking of the P/ω droop coefficient, inertia coefficient, and damping coefficient for frequency changes but also avoids the influence of parameter incoordination on the quality and stability of the system.

2. Topology and Mathematical Model of VSG

2.1. Topology of VSG

A traditional VSG system is shown in Figure 1. The main circuit of the system consists of a DC source, voltage-type converter, filter circuit, load, and power grid. The control loop collects the output voltage, current, active power, reactive power, and other signals of the main circuit; generates modulation signals through the virtual governor, excitation controller, voltage and current double-loop controller; and finally generates the pulse signal to control the converter [19,20]. Topology and control block diagram of Figure 1: L_{abc} is the filter inductor; C_{abc} is the filter capacitor; P_e and Q_e are the active and reactive power output by the converter, respectively; i_{abc_inv} and u_{abc_inv} are the output current and voltage, K is the integral loop coefficient, U_n is the amplitude of the given voltage, U_{inv} is the RMS value of the three-phase voltage output from the converter, R_r is the stator resistance; L_r is the inductance, R_g is the line resistance, L_g is the line inductance; u_g is the grid voltage, i_{Labc} is the inductor current, u_{abc_ref} is the modulation voltage reference, g_{PWM} is the modulation signal.

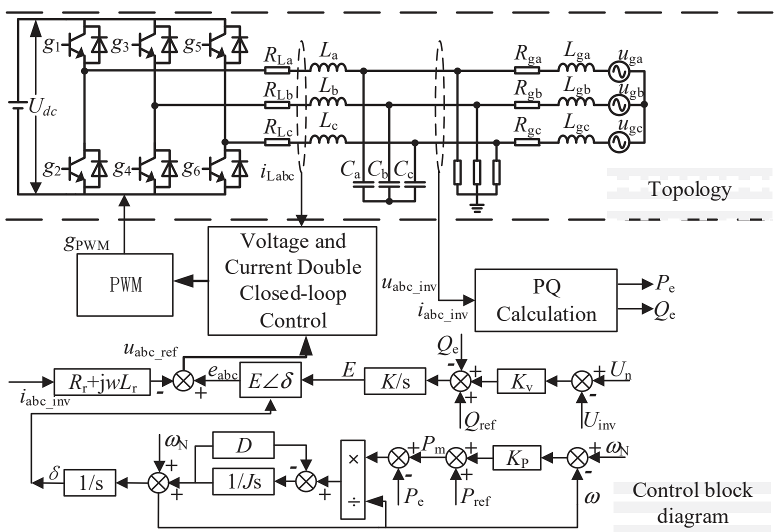

Figure 1. Topology and control block diagram of VSG.

2.2. Mathematical Model of VSG

The stator electrical equation and the typical second-order rotor motion equation of the synchronous generator are [6]:

$$\begin{cases} e = u + i(R + \mathrm{j}\omega L) \\ T_\mathrm{m} - T_\mathrm{e} = \frac{P_\mathrm{m} - P_\mathrm{e}}{\omega} = J\frac{d\omega}{dt} + D\Delta\omega \\ \Delta\omega = \omega - \omega_\mathrm{N} \\ \frac{d\delta}{dt} = \omega \end{cases} \tag{1}$$

Among them, R is the stator resistance; L is the inductance; u is the armature terminal voltage; T_m is the mechanical torque; T_e is the electromagnetic torque; J is the moment of inertia; D is the damping coefficient; ω is the mechanical angular velocity; ω_N is the rated angular velocity of the system; δ is the output power angle.

As shown in Figure 1, drawing on the principles of the synchronous generator governor and excitation regulator, the active power regulation and reactive power regulation equations are designed so that the entire converter system can truly simulate the characteristics of the synchronous generator.

$$\begin{cases} \omega = 2\pi f \\ E = \frac{K}{s}[Q_\mathrm{ref} + K_\mathrm{V}(U_\mathrm{n} - U_\mathrm{inv}) - Q_\mathrm{e}] \\ \Delta T = [P_\mathrm{ref} + K_\mathrm{P}(\omega_\mathrm{N} - \omega) - P_\mathrm{e}]/\omega \end{cases} \tag{2}$$

In (2), P_{ref} and Q_{ref} are the system active and reactive commands respectively; K_V is the voltage regulation coefficient; K_P is the active power droop coefficient; K is the integral loop coefficient; E is the virtual potential command; U_n is the amplitude of the given voltage; U_{inv} is the effective value of the three-phase voltage output by the converter; f is the frequency of the terminal voltage of the virtual synchronous generator.

Finally, after combining the above-obtained voltages in the voltage synthesis link reference value and phase angle, the output voltage of VSG can be obtained as

$$e = \begin{cases} \sqrt{2}E\sin\delta \\ \sqrt{2}E\sin\left(\delta - \frac{2\pi}{3}\right) \\ \sqrt{2}E\sin\left(\delta + \frac{2\pi}{3}\right) \end{cases} \quad (3)$$

3. Multiparameter Cooperative Adaptive Control of VSG

3.1. The Influence of VSG Parameters on the System

According to the equivalent diagram in Figure 2, the output active power of the grid-connected inverter of VSG can be expressed as:

$$P_{\mathrm{e}} = \frac{EU}{X}\sin\delta \quad (4)$$

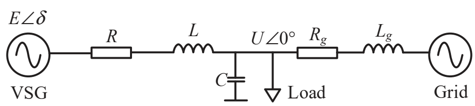

Figure 2. Equivalent diagram of VSG-connected to the grid.

Combining Formulas (1) and (4), we can obtain:

$$\frac{\Delta\omega}{\Delta P} = \frac{\omega_{\mathrm{N}} - \omega}{P_{\mathrm{ref}} - P_{\mathrm{e}}} = -\frac{1}{J\omega_{\mathrm{N}}s + D\omega_{\mathrm{N}} + K_{\mathrm{P}}} \quad (5)$$

Therefore, the closed-loop transfer function of the VSG active loop can be deduced as:

$$\varphi(s) = \frac{P_{\mathrm{e}}}{P_{\mathrm{ref}}} = \frac{\frac{EU}{J\omega_{\mathrm{N}}X}}{s^2 + \frac{D\omega_{\mathrm{N}}+K_{\mathrm{P}}}{J\omega_{\mathrm{N}}}s + \frac{EU}{J\omega_{\mathrm{N}}X}} \quad (6)$$

The corresponding natural oscillation angular frequency and damping ratio are:

$$\begin{cases} \xi = \frac{0.5(D\omega_{\mathrm{N}}+K_{\mathrm{P}})}{\sqrt{\frac{J\omega_{\mathrm{N}}EU}{X}}} \\ \omega_n = \sqrt{\frac{EU}{J\omega_{\mathrm{N}}X}} \end{cases} \quad (7)$$

When $0 < \xi < 1$, the power-frequency system is an underdamped system; when $\xi = 1$, the power-frequency system is a critically damped system; When $\xi > 1$, the power-frequency system is an overdamped system. Considering the two dynamic indicators of response speed and overshoot, the "Siemens second-order optimal system" control strategy is adopted, that is, the damping ratio is set to 0.707. Among them, in the underdamped state, within a certain allowable error, the adjustment time t_{s} and $\sigma\%$ are:

$$\begin{cases} t_s = \frac{4}{\xi\omega_n} = \frac{8J\omega_{\mathrm{N}}}{D\omega_{\mathrm{N}}+K_{\mathrm{P}}} \\ \sigma\% = e^{-\frac{\pi\xi}{\sqrt{1-\xi^2}}} \times 100\% \end{cases} \quad (8)$$

When the droop coefficient is constant, it can be seen from Figure 3a that the damping ratio ξ of the VSG system is proportional to the damping coefficient D and inversely proportional to the virtual inertia J.

While the system adjustment time in Figure 3b is the opposite, it is proportional to the virtual inertia J and inversely proportional to the damping coefficient D.

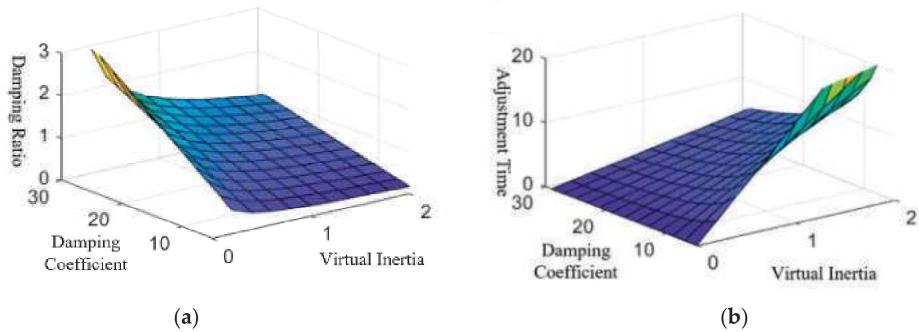

Figure 3. (**a**) Plot of damping ratio versus virtual inertia and damping coefficient. (**b**) Plot of adjustment time versus virtual inertia and damping coefficient.

The above analysis does not consider the influence of the droop coefficient on the dynamic response of the system. Combining Formulas (7) and (8), it can be known that when the virtual inertia and damping coefficient are constant, the damping ratio increases with the increase of the droop coefficient, and the adjustment time decreases with the increase of the droop coefficient. Relative to the damping ratio, the system adjustment time is more deeply affected by the droop factor.

It can be seen from the above analysis that the virtual inertia J, damping coefficient D, and droop coefficient K_P in the traditional virtual synchronous machine control strategy remain unchanged. If any one of the values is changed alone, although the transient characteristics of the VSG grid connection can be significantly improved, it cannot take into account the stability and rapidity of the transient process after the system is disturbed.

3.2. Cooperative Adaptive Selection Strategy of Control Parameters

In this paper, the frequency oscillation process of VSG is divided into four different stages, as shown in Figure 4. The characteristics of the different phases are analyzed, the virtual inertia J and the droop coefficient K_P are adjusted in real time, and the damping coefficient D is coordinated according to the relationship between the three to achieve stable control of the grid-connected VSG transient process.

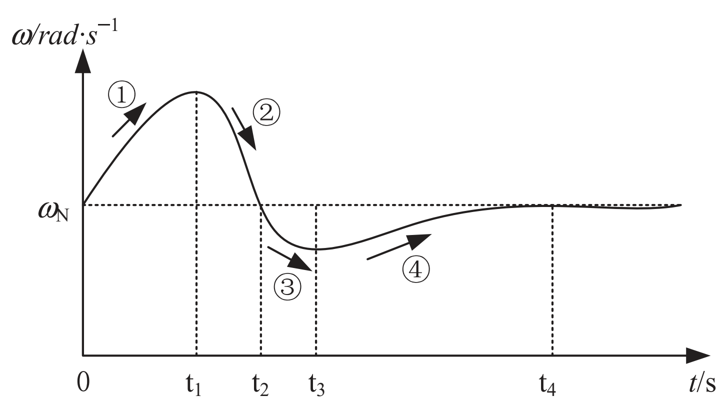

Figure 4. Angular velocity fluctuation graph.

$\Delta\omega \cdot d\omega/dt > 0$ exists in both stage 1 and stage 3 and $|\Delta\omega|$ gradually increases in both stages. These two stages are defined as the acceleration stage of rotor angular velocity. This stage requires larger virtual inertia J and larger droop coefficient K_P to reduce the amplitude of rotor angular velocity offset.

$\Delta\omega \cdot d\omega/dt < 0$ exists in both stage 2 and stage 4. Since $|\Delta\omega|$ gradually decreases, the two stages are defined as the deceleration stage of rotor angular velocity. In this stage, the virtual inertia J and the droop coefficient K_P need to be reduced to make the rotor angular velocity return to a stable value as soon as possible.

However, in the acceleration stage of the rotor, although the increase in inertia can improve the anti-interference, it will reduce the response speed. In this stage, the damping method can be used to improve the response speed.

In the deceleration stage, the virtual inertia J and the droop coefficient K_P are reduced, and the system's suppression of the fluctuation of the rotor angular velocity is weakened to speed up the decay rate of the rotor angular velocity; nevertheless, it will cause an increase in the amplitude of the fluctuation of the rotor angular velocity. Therefore, at this stage, the damping coefficient can be increased to reduce the overshoot of the system and make the frequency return to stability as soon as possible.

The selection of J is determined by $\Delta\omega$ and $d\omega/dt$ at the same time. In order to avoid the complicated control strategy, the change rule is set as:

- When $\Delta\omega$ and $d\omega/dt$ change in the same direction, J needs to be increased;
- When $\Delta\omega$ and $d\omega/dt$ change in opposite directions, J should be kept unchanged.

The virtual inertia J and the droop coefficient K_P are associated with the VSG angular velocity and the instantaneous value of the angular velocity to obtain the control parameter adaptive strategy, as shown in Formulas (9) and (10).

$$J = \begin{cases} J_0 + k_J \left| \Delta\omega \frac{d\omega}{dt} \right|, \Delta\omega \cdot \frac{d\omega}{dt} > 0 \\ J_0, \Delta\omega \cdot \frac{d\omega}{dt} \leq 0 \end{cases} \tag{9}$$

$$K_P = K_{P0} + k_\omega \cdot |\Delta\omega| \tag{10}$$

In the formulas, J_0 and K_{P0} are the virtual inertia and droop coefficient of VSG fixed parameters, respectively; k_J is the inertia adjustment coefficient and k_ω is the adjustment coefficient of the droop coefficient.

The coordinated control design of the droop coefficient, virtual inertia, and damping coefficient is carried out. Combining Formulas (7), (9) and (10), the damping coefficient D design under the correlation can be obtained:

$$D = 2\xi \sqrt{J \frac{EU}{\omega_N X}} - \frac{1}{\omega_N} K_P \tag{11}$$

Based on automatic control theory, in order to keep the system in the optimal control operation state, ξ can be set to 0.707. It can be known from Formula (11) that when other parameters in the system are constant, the droop coefficient, virtual inertia, and damping coefficient can be jointly designed according to the requirements of system characteristics.

3.3. The Setting of Parameter Value Range

The adaptive adjustment coefficient of VSG inertia can be selected according to the value range of inertia. According to the setting principle of the virtual inertia value of the VSG scheme of the University of Leuven [21], the maximum value of the virtual inertia must satisfy:

$$J_{\max} < \frac{P_{\max}}{\max\left\{ \omega \left(\frac{d\omega}{dt} \right) \right\}} \tag{12}$$

In Formula (12), $P_{\max}$ is the maximum power that the system can withstand. In order to ensure the stable operation of the system, the angular frequency of the system is limited with reference to the current national standard power system frequency deviation (50 ± 0.2 Hz).

The maximum and minimum frequencies are $\omega_{\max}$ and $\omega_{\min}$, respectively; then, the damping coefficient selection in Formula (10) should satisfy Formula (13):

$$0 \leq \frac{1}{D\omega_{\mathrm{N}} + K_{\mathrm{P}}} \leq \frac{\omega_{\max} - \omega_{\min}}{P_{\max} - P_{\min}} \tag{13}$$

In Formula (13), $P_{\min}$ is the minimum output power of VSG. Therefore, the minimum value of the damping coefficient D is:

$$\left(\frac{P_{\max} - P_{\min}}{\omega_{\max} - \omega_{\min}} - K_P \right) / \omega \mathrm{N} \leq D \tag{14}$$

And the droop factor shall also be taken to satisfy Formula (15):

$$K_{\mathrm{P}} \leq P_{\max} / \Delta\omega_{\max} \tag{15}$$

In addition, in order to make the system have good response rapidity, each transient component of its response should have a large decay factor—that is, the closed-loop pole of the system should be far away from the imaginary axis, and the closed-loop pole should satisfy [22–25]:

$$\mathrm{Re}(s_{\mathrm{i}}) = -\omega_{\mathrm{n}}\xi = -\frac{D + K_{\mathrm{P}}/\omega_{\mathrm{N}}}{2J} \leq -10 \tag{16}$$

In VSG control, moderate damping, a fast response, and a small overshoot are usually desired in the control system. The damping ratio ξ is therefore chosen to be in the range (0.7, 1). In summary, the range of virtual inertia J and droop factor K_{P} can be obtained as shown in Figure 5.

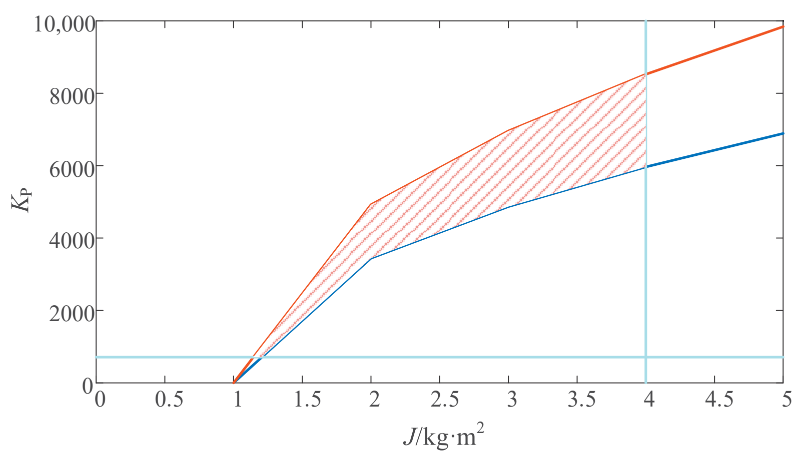

Figure 5. The range of virtual inertia J and droop factor K_{P}.

To sum up, the flow chart of the core algorithm in this paper is shown in Figure 6, where N_{ω} is the threshold for setting a triggering adaptive function.

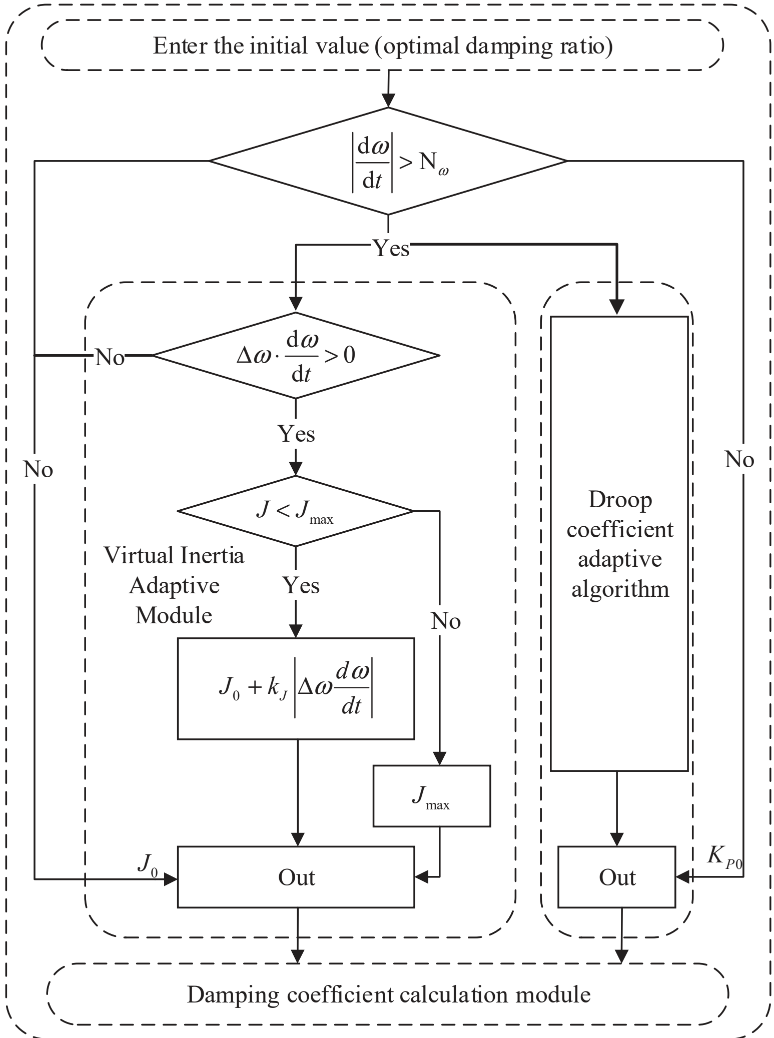

Figure 6. Core algorithm flow chart.

Based on the above design, the bode diagram of the whole system is shown in Figure 7. In the figure, the amplitude margin as well as the phase margin of the control strategy proposed in this paper is the largest, so that the error value is the smallest and not prone to damped oscillations, resulting in the highest stability and the best system performance. Fuzzy adaptive optimal control is the next most effective control strategy. Traditional adaptive control is the least effective.

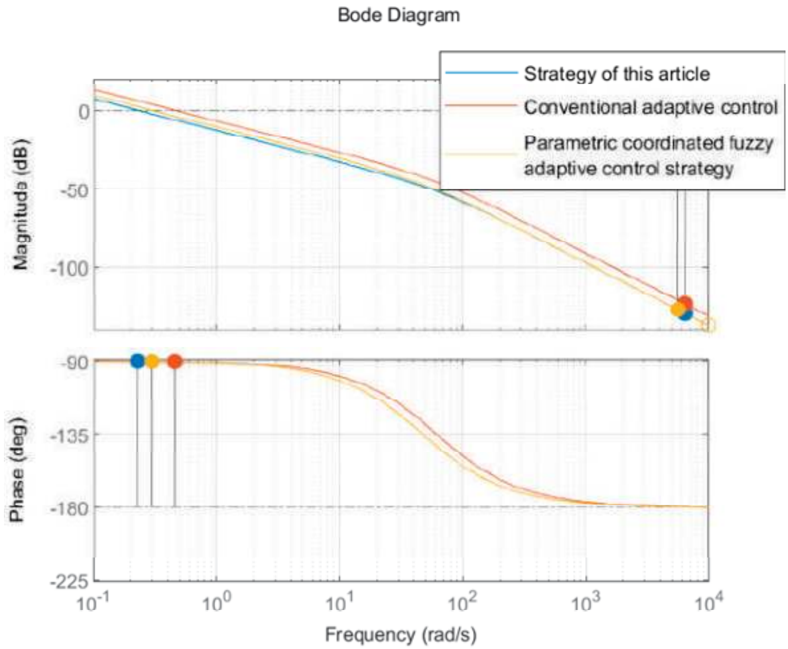

Figure 7. Bode diagram of the system.

4. Simulation Analysis

In order to verify the control strategy proposed in this paper, an improved VSG system simulation model on the MATLAB/Simulink platform is built into this work, and the relevant control parameters used are shown in Table 1.

The system is in a grid-connected operation state, the simulation duration is set to 2 s, the initial steady state is assumed, and the grid frequency is equal to the rated frequency. In order to simulate the change in load power, the load power is set to increase by 10 kW at 1 s, and the reactive power is constant at 0 kVar during this period.

Table 1. Simulation parameters of VSG.

Parameter	Numerical Value	Parameter	Numerical Value
Rated voltage on AC grid u_g (V)	380	Filter inductor L_{abc}(mH)	0.8
Rated voltage on DC side Udc (V)	800	Filter capacitor C_{abc} (uF)	10
Rated active power (W)	50,000	Initial value of virtual inertia J_0 (kg·m^2)	1.127
Rated reactive power (Var)	0	Initial value of droop coefficient K_{P0}	5000

4.1. Effect of Changes in System Parameters on System Frequency

Figures 8–10 show the frequency variation curve of the system with different inertia, damping, and droop coefficient. When the virtual damping D and droop factor K_P are fixed, the greater the virtual inertia J is, the longer the dynamic adjustment time is, but the frequency overshoot is reduced. The system transitions from an overdamped state to an underdamped state, as shown in Figure 8.

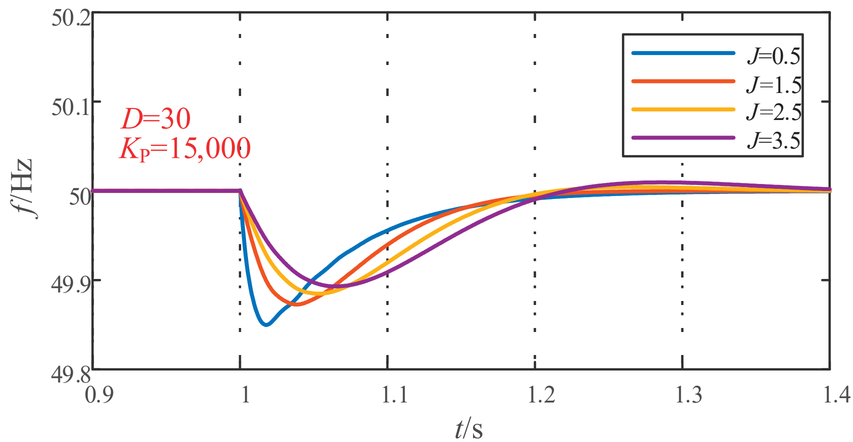

Figure 8. Frequency waveform under J change.

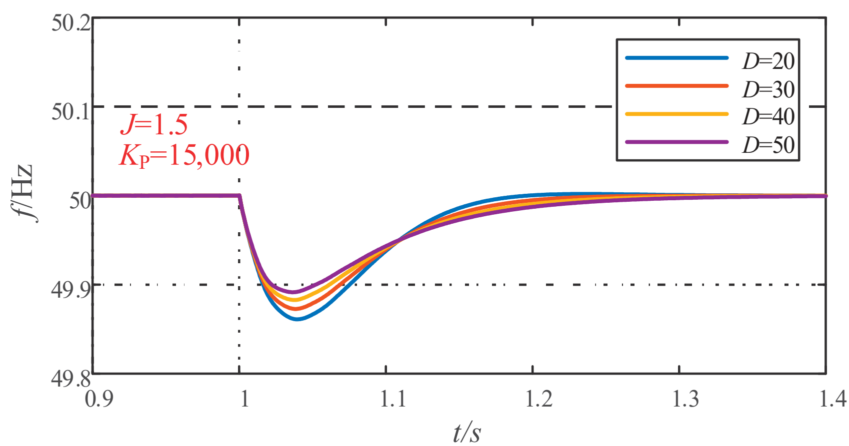

Figure 9. Frequency waveform under D change.

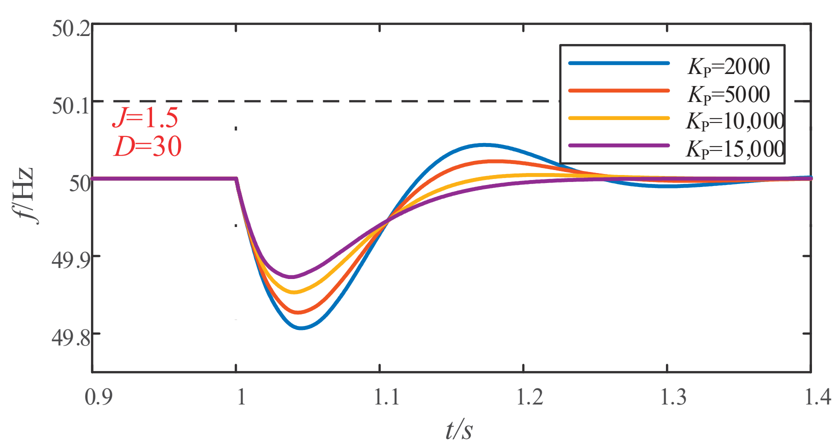

Figure 10. Frequency waveform under K_P change.

When the virtual inertia J and sag factor K_P are fixed, the smaller the virtual damping D is, the longer the dynamic adjustment time is, and the greater the frequency overshoot is, the system transitions from an overdamped state to an underdamped state, as shown in Figure 9.

When the virtual inertia J and virtual damping D are fixed, the larger the sag coefficient K_P is, the smaller the frequency overshoot is, as shown in Figure 10. The correctness of the theoretical analysis in this paper is verified.

4.2. Simulation under Varying System Loads

4.2.1. Simulation Comparison of Different Control Strategies

In order to verify the superiority of this control strategy in grid-connected mode, it is compared with coordinated inertia damping (ξ = 0.707), J and D coordinated adaptive control and parametric coordinated fuzzy adaptive control strategy, respectively. The frequency change is shown in Figure 11.

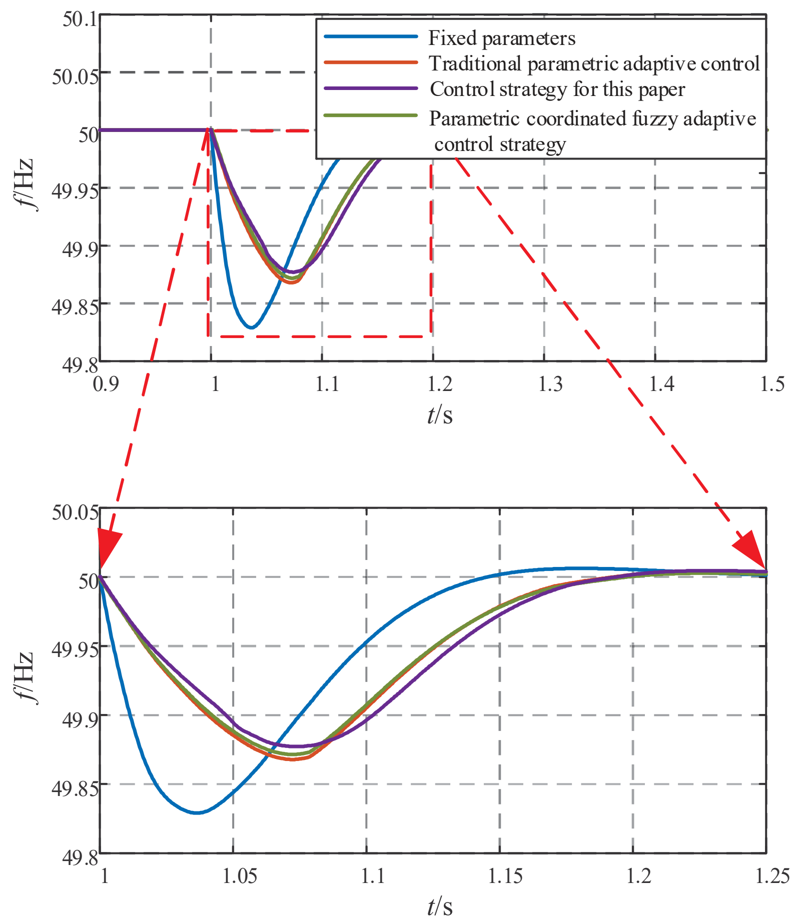

Figure 11. Frequency variation under different control strategies.

The frequency change curve and a partially enlarged view can be seen in Figure 11. It can be seen from Figure 11 that the coordinated inertia damping (ξ = 0.707) control strategy is adopted—when the parameters are fixed, the maximum offset of the frequency fluctuation reaches 0.17 Hz, and it takes about 0.3 s to reach the steady state. When using

J and *D* coordinated adaptive control, the maximum frequency offset is reduced to 0.133 Hz, and the adjustment time is about 0.22 s. When using a parametric coordinated fuzzy adaptive control strategy, the maximum frequency offset is reduced to 0.132 Hz, and the adjustment time remains the same. When the multiparameter cooperative adaptive control strategy is adopted, the maximum frequency offset is 0.123 Hz, and the adjustment time is 0.22 s. The specific performance indicators are shown in Table 2.

The maximum deviation of the multiparameter cooperative adaptive control is optimized by 7.52% compared with the *J* and *D* coordinated adaptive control and is optimized by 27.65% compared with the coordinated inertia damping (ξ = 0.707) control strategy. The adjustment time of the multiparameter cooperative adaptive control is equivalent to that of the *J* and *D* coordinated adaptive control and is 26.67% higher than that of the coordinated inertia damping control.

Table 2. Control strategy performance indicators.

Control Strategy	Nadir (Hz)	Deviation (Hz)	Adjustment Time (s)	ROCOF
Fixed parameters	49.83	0.17	0.3	maximum
J, *D* adaptive control	49.867	0.133	0.22	medium
Parametric coordinated fuzzy adaptive control strategy	49.868	0.132	0.22	medium
Multiparameter cooperative adaptive control	49.877	0.123	0.22	minimum

4.2.2. Comparison of Parameter Changes

Figures 12 and 13 show the variation of the *J* and *D* coordinated adaptive control parameters and the variation of the multiparameter collaborative adaptive control parameters in this paper, respectively.

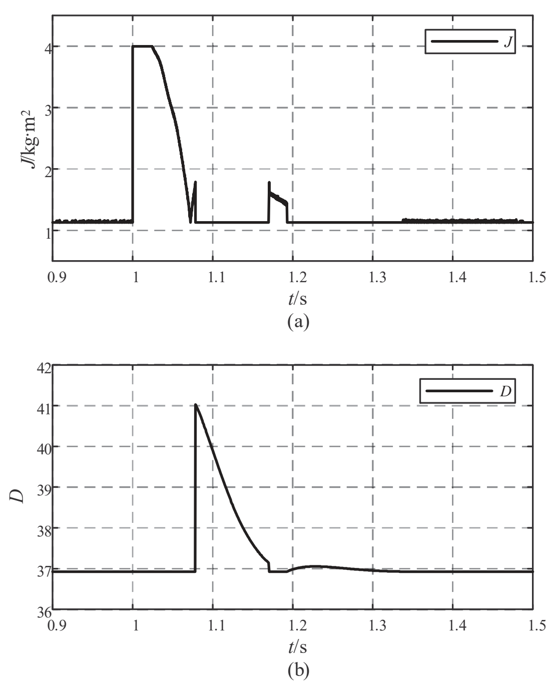

Figure 12. *J* and *D* coordinated adaptive control parameter variation: (**a**) virtual inertia change curve; (**b**) damping coefficient change curve.

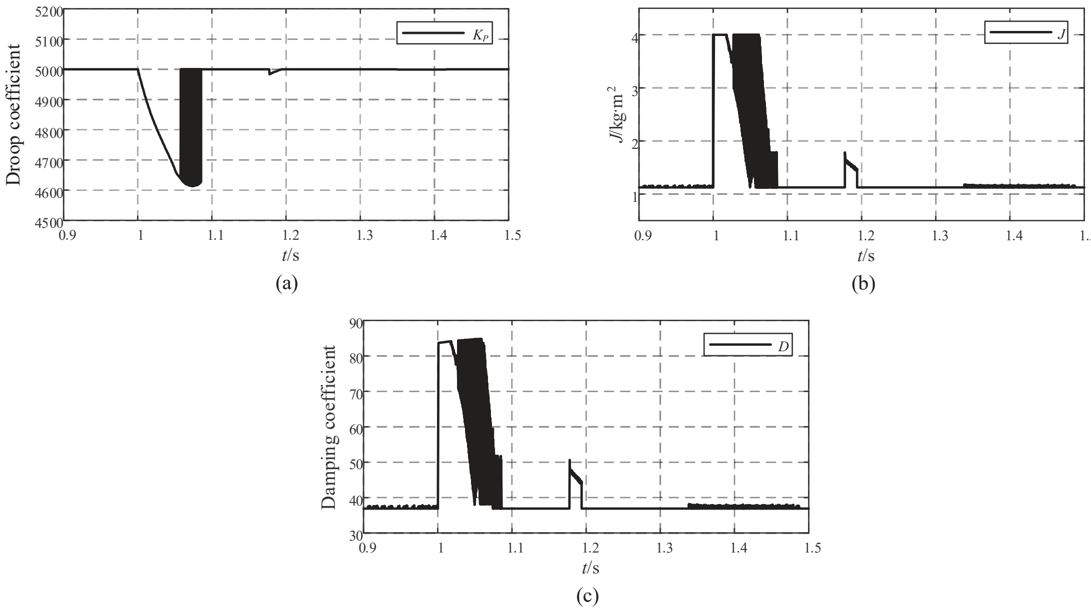

Figure 13. (**a**) Droop coefficient change curve. (**b**) Virtual inertia change curve. (**c**) Damping coefficient change curve.

Figure 12, as well as Figure 13b,c, shows the frequency change caused by the load change at 1 s; the angular frequency rate of change and frequency offset are both negative, and in the angular frequency acceleration phase, the virtual inertia and damping coefficient increase rapidly. During the frequency fluctuation, when the angular frequency rate of change and frequency offset have different signs, the virtual inertia and damping coefficient decrease rapidly in the angular frequency deceleration stage.

However, when J and D adaptive control is in a perturbation cycle of the change process, during the same moment only one quantity changes (that is, the control for alternate), other quantities are not affected. This control system is simple, ignoring the influence of the droop coefficient.

It also can be seen from Figure 13a that the droop coefficient will have a corresponding adaptive change when the frequency changes caused by the load change, to further synergistically optimize the damping coefficient. During dynamic changes, the droop coefficient varies continuously in the range of 4600 to 5000, and the virtual inertia varies in the range of 1 to 4, while the damping coefficient varies in the range of 38 to 85 under the combined effect of both.

4.3. Simulation under Grid Frequency Variation

To further verify the performance of the proposed control strategy, the grid frequency was simulated: at 0.5 s, the grid frequency increased by 0.2 Hz, and after 1 s, the frequency returned to 50 Hz.

4.3.1. System Frequency Performance

The frequency change curve and a partially enlarged view can be seen in Figure 14. As can be seen from the graph, frequency overshoot exists using the fixed parameter control strategy; when using conventional adaptive control, the overshoot is reduced. The rate of change in frequency is greater for both of these approaches. When using fuzzy adaptive control and the strategy proposed in this paper, there is no overshoot in the frequency, and the frequency variation rate of the system is minimal under the control strategy proposed in this paper.

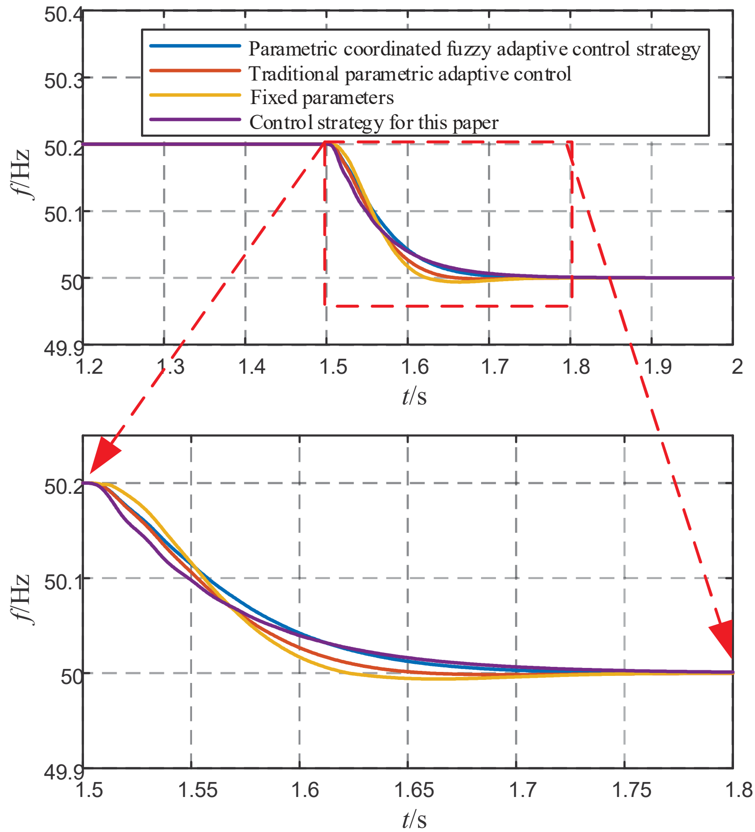

Figure 14. Frequency curves.

4.3.2. Variation of the Three Parameters

It can also be seen in Figure 15a that the VSG needs to adjust its output when the grid frequency rises by 0.2 Hz throughout; therefore, it needs to adjust the droop coefficient, which varies dynamically between 5000 and 5600. The virtual inertia also varies at this point between around 1.1 and 3.9, as shown in Figure 15b. The damping coefficient is adjusted in real time according to the mathematical logic between the three and varies between 36 and 82, as shown in Figure 15c. Unlike the load variation, the sag coefficient, virtual inertia, and the damping coefficient keep interacting with each other throughout the process, without stopping to vary when the frequency is stable.

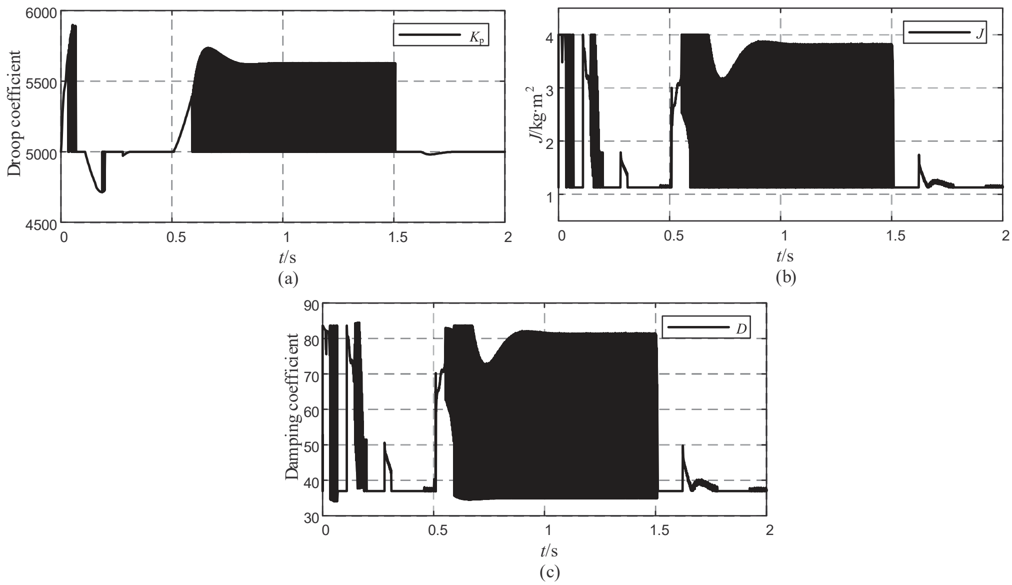

Figure 15. (**a**) Droop coefficient change curve. (**b**) Virtual inertia change curve. (**c**) Damping coefficient change curve.

5. Conclusions

Although the traditional virtual synchronous generator technology can improve the stability of the system by simulating the operating characteristics of the synchronous generator, it has problems, such as the inability to take into account multiple performance indicators for parameter design, long adjustment time, and large transient overshoot. To address these problems, this paper uses an improved VSG parameter optimization cooperative control strategy, and the effectiveness of this control strategy is verified by simulation through simulated load variation and frequency variation. The main work and contributions are as follows:

(1) Analyze the effect of droop factor, virtual inertia, and damping factor on the VSG system and determine the range of values for the system parameters.
(2) The existing J and D coordinated adaptive control is optimized. The droop coefficient and virtual inertia can be adjusted in real time according to the system frequency state, and the damping coefficient can be changed cooperatively according to the corresponding relationship. The three are always coordinated and adjusted during the change process, which effectively improves the dynamic performance of the system frequency. The effectiveness and reliability of the proposed multiparameter cooperative adaptive control strategy have been verified by simulation.

Author Contributions: Conceptualization, B.R. and Z.F.; methodology, Q.L.; software, Y.S.; validation, B.R., Z.F. and Y.S.; formal analysis, Q.L.; investigation, B.R.; resources, Q.L.; data curation, Q.L.; writing—original draft preparation, B.R.; writing—review and editing, Y.S.; visualization, B.R.; supervision, Y.S.; project administration, Y.S.; funding acquisition, Y.S. All authors have read and agreed to the published version of the manuscript.

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Article

Joint Detection and State Estimate with GSAs in PMU-Based Smart Grids

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Abstract: The Phasor Measurement Unit (PMU) with a GPS signal receiver is a synchronized sensor widely used for power system state estimation. While the GPS receiver ensures time accuracy, it is vulnerable to network attacks. GPS spoofing attacks can alter the phase angle of PMU measurement signals and manipulate system states. This paper derives a power system state model based on PMUs under GPS spoofing attacks, according to the characteristics of changes in bus voltages and branch currents after GSA. Based on the characteristics of this model, a detection and correction algorithm for attacked data is proposed to detect GSA and correct attacked measurements. The corrected measurements can be used for power system state estimation. Simulation results on the IEEE 14-bus system show that the proposed algorithm improves the accuracy of state estimation under one or multiple GSAs, especially when multiple GSAs are present, compared to classical Weighted Least Squares Estimation (WLSE) and Alternating Minimization (AM) algorithms. Further research indicates that this algorithm is also applicable to large-scale networks.

Keywords: phasor measurements unit (PMU); GPS spoofing attacks (GSAs); detection; correction; state estimate

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1. Introduction

Phasor Measurement Units (PMUs) based on global positioning system have been widely used in wide area measurement system [1,2]. PMU is an advanced digital instrument used for real-time monitoring of grid operation in smart grids. By installing a PMU on one bus, it can obtain the bus voltage phasor and all current phasors of the branches connected to that bus [3]. PMU data is an important data source for situation awareness, dispatching control, and early warning of modern power systems. Many advanced applications have been developed based on PMU data, such as state estimation, wide area damping control, generator state monitoring, static voltage stability evaluation, etc. Accurate PMU data is a prerequisite for advanced applications.

The PMU uses a common time source for synchronization and can measure the electrical waves on the grid. The PMU does this by converting the analog signals of voltage and current to digital signals and applying anti-alias filters and discrete Fourier transforms to isolate the fundamental frequency components and compute their phase representation [4,5]. This allows obtaining the fundamental frequency amplitude and phase information of the voltage and current signals at each bus in the power system, instead of the actual measurements in conventional Supervisory Control and Data Acquisition (SCADA) systems. When the number of PMUs deployed in the system is sufficient to make the system globally observable, state estimation is only required based on the PMU data [6,7]. In this case, the state of the power system can be expressed in the form of a linear equation that can be solved in a single iteration, which improves the processing speed [8,9]. Replacing nonlinear state estimation with linear state estimation using PMU measurements allows direct manipulation of the Jacobi matrix. The reliability of the simplified model and

PMU state variables was explained in [10]. The solution of the linear model is direct and non-iterative.

In recent years, with the intelligent development of power grid and the importance of the power grid security, PMUs have become increasingly important in power system state estimation [10–15]. The low reporting rate and complex nonlinear state estimation of traditional SCADA systems for power grids make it difficult to meet high-precision state analysis and real-time safety monitoring. However, PMU can provide synchronized phasor measurements, generating a linear model for state estimation. Their sampling rate is much higher than that of the SCADA system, and it can estimate the state of power system in real-time and quickly respond to abnormal conditions.

Each PMU is equipped with a GPS receiver, which is synchronized using GPS civilian signals, which are not encrypted like the military signals [11,16–18]. GPS provides sub-microsecond precision timing [16], which plays a crucial role in the time synchronization of PMU measurements. Wireless communication between civilian receivers and satellites is then vulnerable to cyber attacks. GPS spoofing is caused by transmitters mimicking GPS signals with the intent of altering the GPS time estimated by the PMU's GPS receiver [19]. These attacks maliciously introduce incorrect timestamps, which leads to incorrect phase angles in PMU measurements [20], making the system state estimation problem nonlinear, counteracting the original motivation for the introduction of PMUs as well as posing a serious threat to the correct state estimation of the power system. The issue of time delay in PMU measurements and its compensation methods have been discussed in [21], but these methods have limited effectiveness compared to GSAs. As a result, GSAs detection and countermeasures are widely studied and methods are proposed to prevent serious destructions to the power system.

2. Prior Work

In recent years, there have been much research on GSA, and the research directions were mostly divided into three categories. They were respectively the feasibility of GSA [22–25], the impact of GSA on power grid state [19,26], and some solutions for GSA [21,27–30].

The feasibility details of GSAs were shown in [22–25]. Shepard et al. [24] have performed field tests of GPS spoofing on PMUs, exposing the vulnerability of PMUs to this malicious attack; by spoofing the GPS signal receiver inside the PMU, the attacker can introduce timing errors, which result in a consequent shift in the phase angle of the PMU, which posed a serious problem for real-time monitoring of the smart grid. Humphreys et al. [22] first implemented GPS spoofing attack in the laboratory and gave a preliminary mitigation method for non-encrypted GSA. Jiang et al. [25] manipulated the acquired GPS signal data, and they found that this operation could change the phase angle of the PMU-measured signal.

However, more research were on the countermeasures of GSA [19,26–30]. Liang et al. [26] comprehensively discussed the theoretical basis of false data injection attacks (FDIA) and gave the most basic defense strategy for FDIA. After measuring and collecting the data provided by the PMUs, a cross-layer detection method was proposed in [19]. Its principle is to use the angle of arrival of GPS as the initial guess, and then detect whether GSA has occurred through the state estimation of the system. Mahapatra et al. [27] proposed a method to detect bad data in PMU measurements during interference. Which was based on the principal component analysis (PCA) method to distinguish the safety data caused by bad data from the manipulation data caused by interference. In particular, Zhang et al. [28] proposed a novel distribution system identification and correction algorithm for simultaneous GSA with multiple PMU positions. The algorithm first analyzed the sensitivity of the residual of the phase angle state estimation under a single GSA, and then used the proposed detection technology identification algorithm to locate the attacked PMU and the shift range of the phase angle after the attack. Finally, the phase shift was determined by minimizing the offset of measurement and system state estimation. PMU is vulnerable to attack because it receives unencrypted civil signals. To solve this problem, Mina et al. [29] proposed a wide-area spoofing detection algorithm for PMU, which used the hybrid communication architecture of NAPSInet. They created conditional signal fragments containing military P (Y) signals, whose precise code sequences are not available to civilian users, thus protecting PMUs from attack from the source. Unlike adding military encrypted signal fragments, Bhamidipati et al. [30] proposed a new algorithm for jointly spoofer location and GPS time using multiple receivers direct time estimation (MRDTE). The experimental results show that it can locate the attacked PMU within a certain range and estimate the time change. For the error synchronization in PMU measurement data, Zhang et al. [21] gave the source of this error, and proposed the method of using the Kalman filter to compensate for this error, which effectively alleviates the error caused by different.

The impact of a GPS spoofing attack on the secure operation of the power grid system was shown in [25]. We can find that once the PMU is subject to GSA, it is impossible to conduct real-time monitoring and correct state estimation of the system, which will bring huge hidden dangers.

For system state estimation, the static estimation method was used to detect attacks of PMU in the power grid [31]. More PMUs were installed in the power grid to make the measurements of the bus redundant, and then various PMU signals are compared mathematically to obtain the GPS attack detection formula [32]. Most of the above literature treats the system as static. The dynamic changes of the system were not considered. Phase shift caused by PMU delay under system dynamic characteristics was considered in [33]. However, this phase shift is not caused by spoofing attacks, but the deployed PMUs come from different standards, protocols, and designs. Siamak et al. [34] also described the attacked power system under the dynamic framework, which was an extension of the dynamic model of the methods in [13,26], and then proposed a method to detect multiple non-constant attacks on the system. By using Kalman filter, this method can estimate the data and measurements of spoofing attack with higher accuracy, and determine the phase shift of spoofing attack. In addition, the traditional measurement model assumed that the incomplete synchronization of PMU was modeled as additive noise [27,35–37]. However, according to the numerical example given in [33], if the synchronization error and/or the time between successive synchronizations increase, the traditional estimation method will deteriorate significantly.

In this paper, we consider the introduced multiplicative noise by GSAs in the proposed model, which results in the attacked residuals being greater than the residuals of the nominal system. We propose a spoofing algorithm for attacked data that classifies the attacked data and secure data, and then inputs the attacked data into the proposed data correction algorithm to obtain measurement data under secure residuals. Finally, a static estimation method is employed to estimate the system state using all the data. The main contributions of this paper can be summarized as follows:

- The PMU's measurement model under GSA attack is derived by comparing changes in bus voltage and branch current before and after the attack.
- According to the characteristic that the measurement residual of PMU data will change after being attacked, an attack detection method is proposed, which can effectively detect one or more GSAs.
- Through the particularity of the matrix in the measurement model, a bad data correction algorithm is proposed to restore the corrected attacked to the security measurements residual range. This method effectively avoids the challenge of coupling two unknown parameters in the model, which is difficult to estimate.

The rest of this paper is organized as follows. Section 3 introduces the security measurement model of PMU and derives the measurement model of PMU after being attacked by GPS spoofing. Section 4 describes the proposed algorithm and explains the simulation environment and implementation. Section 5 presents the simulation results, and our conclusions are peserented in Section 6.

3. System Model

This section summarizes the measurement model for the presence of GPS spoofing attacks in the network. The PMU measurements are correlated by derivation to the network state and the phase angle shift caused by GPS spoofing attacks.

3.1. Measurement Model

We consider a power network with N buses connected via l transmission lines, e.g., the IEEE 14-bus model, that is observed by M PMUs installed on several buses. This collection of measured quantities (in rectangular corrdinate) at bus k is concatenated in a vector $\boldsymbol{z}_k \in \mathbb{R}^{(2+2l)\times 1}$. The PMU measurements at bus k, which is connected to l different buses, are given by

$$\boldsymbol{z}_k = [U_k^r, U_k^i, I_{k1}^r, I_{k1}^i, \ldots, I_{kl}^r, I_{kl}^i]^T \tag{1}$$

where U_k^r, U_k^i denote the real and imaginary parts of the complex voltage at bus k, respectively. I_{kl}^r, I_{kl}^i are the real and imaginary parts of the complex current injected into line (k, l), T is the transpose operator. The system state $\mathbf{x} \in \mathbb{R}^{2\mathrm{N}\times 1}$ can be written as

$$\mathbf{x} = [U_1^r, U_1^i, \ldots, U_k^r, U_k^i, \ldots, U_N^r, U_N^i]^T \tag{2}$$

Thus, using the bus admittance matrix of the network, $\boldsymbol{z}_k$ can be written as a linear function of the system state $\mathbf{x}$:

$$\boldsymbol{z}_k = \boldsymbol{H}_k \mathbf{x} + \boldsymbol{e}_k \tag{3}$$

where $\boldsymbol{z}_k$ denotes the PMU measurements at the bus k, $\boldsymbol{H}_k$ is the admittance matrix associated with a bus at bus k, and $\boldsymbol{e}_k$ denotes gauss measurement noise. Thus, the overall PMUs' measurements are given by

$$\boldsymbol{z} = \boldsymbol{H}\mathrm{x} + \boldsymbol{e} \tag{4}$$

where $\boldsymbol{z} = [\boldsymbol{z}_1, \ldots, \boldsymbol{z}_M]^T$, $\boldsymbol{H} = [\boldsymbol{H}_1, \ldots, \boldsymbol{H}_M]^T$, $\boldsymbol{e} = [\boldsymbol{e}_1, \ldots, \boldsymbol{e}_M]^T$. In this case, the estimation of the state variable $\mathbf{x}$ can be obtained from the least square estimation (LSE), which is expressed as

$$\hat{\mathbf{x}} = (\boldsymbol{H}^T\boldsymbol{H})^{-1}\boldsymbol{H}^T\boldsymbol{z} \tag{5}$$

In the network, the continuous voltage signal on the bus k when the time t is defined as

$$\bar{U}_k(t) = U_k(t)\cos(2\pi f_c t + \varphi_k(t)) \tag{6}$$

where f_c is the frequency, $\bar{U}_k$ can be expressed in the form of phasor $U_k(t)e^{j\varphi_k(t)}$, where $U_k(t)$ denotes magnitude and $\varphi_k(t)$ denotes the phase at time t. Since the subsequent analysis of the GPS spoofing attack is based on the data collected at a certain time point, we omit the symbol t in the subsequent formula for simplicity of notation. Therefore, according to its phasor form, it can be obtained that the real and imaginary parts of the complex voltage are

$$U_k^r = U_k \cos\varphi_k \tag{7}$$

$$U_k^i = U_k \sin\varphi_k \tag{8}$$

Here, a branch line is approximated using a π model, as illustrated in Figure 1. The admittance matrix relates the complex current flowing in a line with the complex voltages at the buses of the π model.

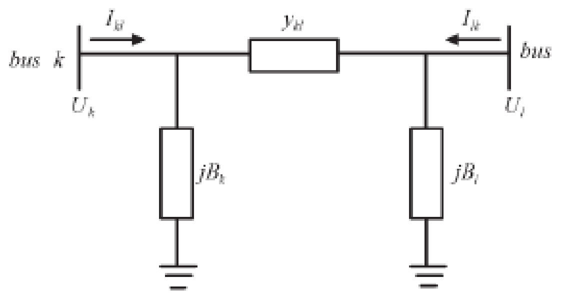

Figure 1. Bus branch model.

In Figure 1, we denote the susceptance at bus k as B_k and the admittance at branch (k, l) as y_{kl} with

$$y_{kl} = g_{kl} + jb_{kl} \tag{9}$$

where g_{kl} is the conductance and b_{kl} is the susceptance. In this paper, we assume these parameters are known and constant. Therefore, we can calculate the real and imaginary parts of the branch current as follows:

$$I^r_{kl} = (U^r_k - U^r_l)g_{kl} - (U^i_k - U^i_l)b_{kl} - B_k U^i_k \tag{10}$$

$$I^i_{kl} = (U^r_k - U^r_l)b_{kl} + (U^i_k - U^i_l)g_{kl} + B_k U^r_k. \tag{11}$$

3.2. Spoofing Attack Model

Consider attackers can manipulate the synchronization of PMUs through GPS spoofing, such that the time reference of an attacked PMU is delayed or advanced. For each attacked measurement, we consider GPS spoofing attack will shift the phase angle of the phasor z_n by an angle α_n while the phasor magnitude is unchanged [25]. Thus the attacked measurement are defined as

$$z^{spf}_n = z_n e^{j\alpha_n} \tag{12}$$

where z^{spf}_n denotes the change of measurement under attack. The resulting measurement vector $\boldsymbol{z}^{spf}$ is given by

$$\boldsymbol{z}^{spf} = (z^{spf}_1, \dots, z^{spf}_k, \dots z^{spf}_M)^T = \boldsymbol{z}^{spf} \odot \boldsymbol{w} \tag{13}$$

where $\boldsymbol{w} = (e^{j\alpha_1}, \dots, e^{j\alpha_k}, \dots, e^{j\alpha_M})^T$ is the attack vector, $\odot$ is the Hadamard product and if a PMU is security, the shift angle $\alpha_n = 0$, i.e., $e^{j\alpha_n} = 1$ can be obtained. Thus, the measurments are still z_n. However, the PMU installed on bus k is attacked, and the voltage measurements are

$$\bar{U}^{spf}_k = \bar{U}_k \cdot e^{j\alpha_k} = U_k \cdot e^{j(\varphi_k + \alpha_k)} \tag{14}$$

where α_k is the phase shift angle caused by the attack, and

$$\alpha_k = 2\pi f_c t_k \tag{15}$$

where t_k is the time delay of the kth bus due to a spoofing attack. Thus, by transforming (14) in the rectangular coordinates, we can obtain the real and imaginary parts of attacked voltage can be expressed as

$$\bar{U}^{spf,r}_k = U_k \cos(\alpha_k + \varphi_k) = U^r_k \cos\alpha_k - U^i_k \sin\alpha_k \tag{16}$$

and

$$\bar{U}_k^{spf,i} = U_k \sin(\alpha_k + \varphi_k) = U_k^r \sin\alpha_k + U_k^i \cos\alpha_k \tag{17}$$

Note that trigonometric identities $\cos(x+y) = \cos(x)\cos(y) - \sin(x)\sin(y)$ and $\sin(x+y) = \sin(x)\cos(y) + \cos(x)\sin(y)$ are introduced in the operation of the above formula. In addition, the delay expression of the current after being attacked are given by

$$\begin{aligned}
I_{kl}^{spf,r} &= (U_k \cos(\alpha_k + \varphi_k) - U_l \cos(\alpha_k + \varphi_k))g_{kl} - (U_k \sin(\alpha_k + \varphi_k) \\
&\qquad - U_l \sin(\alpha_k + \varphi_k))b_{kl} - B_k U_k \sin(\alpha_k + \varphi_k) \\
&= U_k^r(g_{kl}\cos\alpha_k - b_{kl}\sin\alpha_k - B_k \sin\alpha_k) + U_l^r(-g_{kl}\cos\alpha_k + b_{kl}\sin\alpha_k) \\
&\quad + U_k^i(-g_{kl}\sin\alpha_k - b_{kl}\cos\alpha_k - B_k\cos\alpha_k) + U_l^i(g_{kl}\sin\alpha_k + b_{kl}\cos\alpha_k)
\end{aligned} \tag{18}$$

$$\begin{aligned}
I_{kl}^{spf,i} &= (U_k \cos(\alpha_k + \varphi_k) - U_l \cos(\alpha_k + \varphi_k))b_{kl} + (U_k \sin(\alpha_k + \varphi_k) \\
&\qquad - U_l \sin(\alpha_k + \varphi_k))g_{kl} + B_k U_k \cos(\alpha_k + \varphi_k) \\
&= U_k^r(b_{kl}\cos\alpha_k + g_{kl}\sin\alpha_k + B_k \cos\alpha_k) + U_l^r(-b_{kl}\cos\alpha_k - g_{kl}\sin\alpha_k) \\
&\quad + U_k^i(-b_{kl}\sin\alpha_k + g_{kl}\cos\alpha_k - B_k\sin\alpha_k) + U_l^i(b_{kl}\sin\alpha_k - g_{kl}\cos\alpha_k)
\end{aligned} \tag{19}$$

Note that since GSA has the same effect on all signals measured by PMU at any time, phase angles of voltage and currents shift are identical [38]. Thus, the measurement of PMU installed at bus k after the spoofing attack is given by

$$\boldsymbol{z}_k^{spf} = \boldsymbol{\tau}_k \boldsymbol{H}_k \mathbf{x} + \boldsymbol{e}_k \tag{20}$$

where $\boldsymbol{\tau}_k$ is a block diagonal matrix with following submatrix

$$\boldsymbol{G} = \begin{bmatrix} \cos\alpha_k & -\sin\alpha_k \\ \sin\alpha_k & \cos\alpha_k \end{bmatrix} \tag{21}$$

and the matrix $\boldsymbol{H}_k$ can be written as

$$\boldsymbol{H}_k = \begin{bmatrix} \boldsymbol{I} & 0 & \cdots & 0 \\ \boldsymbol{\Psi}_{k1} & \tilde{\boldsymbol{\Psi}}_{k1} & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ \boldsymbol{\Psi}_{kl} & 0 & \cdots & \tilde{\boldsymbol{\Psi}}_{kl} \end{bmatrix} \tag{22}$$

where

$$\boldsymbol{I} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \tag{23}$$

$$\boldsymbol{\Psi}_{kl} = \begin{bmatrix} g_{kl} & B_k + b_{kl} \\ -B_k - b_{kl} & g_{kl} \end{bmatrix} \tag{24}$$

$$\tilde{\boldsymbol{\Psi}}_{kl} = \begin{bmatrix} -g_{kl} & b_{kl} \\ -b_{kl} & -g_{kl} \end{bmatrix} \tag{25}$$

Stacking (20) for all into one large model yields the measurements of all PMUs under attack as

$$\boldsymbol{z}^{spf} = \boldsymbol{\phi} \boldsymbol{H} \mathbf{x} + \boldsymbol{e} \tag{26}$$

where the measurement error vector $\boldsymbol{e}$ is assumed to be Gaussian $\boldsymbol{e} \sim N(0, \delta^2 \boldsymbol{I})$, and $\boldsymbol{\phi}$ is given by

$$\boldsymbol{\phi} = \begin{bmatrix} \boldsymbol{I}_1 & & & & 0 \\ & \ddots & & & \\ & & \boldsymbol{\tau}_k & & \\ & & & \ddots & \\ 0 & & & & \boldsymbol{I}_M \end{bmatrix} \tag{27}$$

where $\boldsymbol{I}$ denotes an identity matrix and the presence of $\boldsymbol{\tau}_k$ in the diagonal element of the matrix indicates that the kth PMU is attacked. Note that $\boldsymbol{\tau}_k$ depends on α_k, when $\alpha_k = 0$ (no spoofing attack), identity matrix $\boldsymbol{\tau}_k = \boldsymbol{I}_{2l+2}$ can be obtained. Without given $\boldsymbol{\phi}$, the solution of the least squares of the state variable $\mathbf{x}$ in (26) is

$$\hat{\mathbf{x}}^{spf} = (\boldsymbol{H}^T\boldsymbol{H})^{-1}\boldsymbol{H}^T\boldsymbol{z}^{spf} \tag{28}$$

4. Detection of Attacked Data and State Estimate of System

After collecting all the PMU measurement data, we propose a residual-based detection and correction method for attacked data. The specific steps of the entire algorithm are as follows:

(1) Initialization: Use the measurements of all PMUs to estimate the power system state $\hat{\mathbf{x}}^{spf}$ and calculate the measurement residuals $\boldsymbol{r}^{spf}$.

(2) Detection of Attacked Data: Compare the residuals $\boldsymbol{r}^{spf}$ calculated in step 1 with the predetermined threshold ε. If the norm of the measurement residuals $\boldsymbol{r}^{spf}$ is greater than the predetermined threshold, we consider the PMU's measurements to be generated under attack.

(3) Correction of Attacked Data: Correct the attacked data in step 2. We first estimate the attack angle and then use the special property of the PMU measurement model under attack to easily correct the attacked data.

(4) State Estimate: Use the corrected data and the non-attacked data to estimate the power system state $\mathbf{x}$. The detection and correction of attacked data are mainly composed of two sub-algorithms: the detection of attacked data and the correction of attacked data. We will provide specific details in the following sections.

4.1. Detection of Attacked Data

In Section 3.2, we theoretically derived the measurement model of PMU under GSAs. Equation (26) show that the attack model introduces multiplicative attacked data into the security measurement model. The more simplified relationship between the measurements before and after the attack is

$$\boldsymbol{z}^{spf} = \boldsymbol{\phi}\boldsymbol{z} \tag{29}$$

Given the security power grid, the measurement residual of PMU is expressed as

$$\boldsymbol{r} = \boldsymbol{z} - \boldsymbol{H}\hat{\mathbf{x}} = \boldsymbol{z} - \boldsymbol{H}(\boldsymbol{H}^T\boldsymbol{H})^{-1}\boldsymbol{H}^T\boldsymbol{z} = (\boldsymbol{I} - \boldsymbol{H}(\boldsymbol{H}^T\boldsymbol{H})^{-1}\boldsymbol{H}^T)\boldsymbol{z} \tag{30}$$

However, when PMU in the power grid is affected by GSAs, the residual becomes

$$\boldsymbol{r}^{spf} = \boldsymbol{z}^{spf} - \boldsymbol{H}\hat{\mathbf{x}}^{spf} = \boldsymbol{z}^{spf} - \boldsymbol{H}(\boldsymbol{H}^T\boldsymbol{H})^{-1}\boldsymbol{H}^T\boldsymbol{z}^{spf} = (\boldsymbol{I} - \boldsymbol{H}(\boldsymbol{H}^T\boldsymbol{H})^{-1}\boldsymbol{H}^T)\boldsymbol{z}^{spf} \tag{31}$$

Chauhan et al. [39] revealed that when GSAs occur in the power grid, the deviation of the residual error will be caused, which is embodied as $||\boldsymbol{r}^{spf}||^2 \geq ||\boldsymbol{r}||^2$. However, a necessary condition to be met is that the matrix $(\boldsymbol{I} - \boldsymbol{H}(\boldsymbol{H}^T\boldsymbol{H})^{-1}\boldsymbol{H}^T)$ must be Semi-positive definite.

Therefore, we can set a residual threshold ε for the measurement data to identify whether certain data is affected by GSA. When the residual error of a measurement is greater than a given threshold, the algorithm considers the PMU of the measured data as being attacked, otherwise the data is secure. For setting the predetermined threshold, as we mentioned in the previous section, the measurement residual increases under GSA attack. Based on this change, we perform Monte Carlo simulations under secure power system conditions and take the maximum measurement residual as the threshold to distinguish between secure and attacked scenarios. The specific simulation process is as follows: we use MATPOWER to simulate a power system, generate PMU measurements using Equation (4), and estimate the power system state using the least squares estimate. We perform multiple simulations and record the maximum residual norm as the threshold ε.

4.2. Attacked Data Correction and System State Estimate

In Section 4.1, we used the attack detection algorithm to determine the measurements. If the norm of the measurement residuals is greater than the selected threshold, we correct the measurements in the measurement correction algorithm; otherwise, we can directly use the measurements to estimate the power system state. In this section, specific algorithm is proposed to correct the attacked data so that measurements can also be used for state estimation normally.

Equation (29) shows the relationship between the measurements of a PMU under GSA attack and the secure condition. For simplicity, we only consider an attack on one measurements.

$$\boldsymbol{\phi} = \begin{bmatrix} \cos\alpha & -\sin\alpha \\ \sin\alpha & \cos\alpha \end{bmatrix} \tag{32}$$

where α is the shift angle generated by the attack. According to the property $\cos^2\theta + \sin^2\theta = 1$ of trigonometric function and the particularity of matrix $\boldsymbol{\phi}$, (33) can be derived.

$$\boldsymbol{\phi}^T\boldsymbol{\phi} = \begin{bmatrix} \cos\alpha & \sin\alpha \\ -\sin\alpha & \cos\alpha \end{bmatrix} \begin{bmatrix} \cos\alpha & -\sin\alpha \\ \sin\alpha & \cos\alpha \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \tag{33}$$

Therefore, for the attack angles matrix introduced by the attacked measurements, we can obtain

$$\boldsymbol{\phi}^T\boldsymbol{\phi} = \boldsymbol{I} \tag{34}$$

Inspired by (34), the correction method of the attacked data can be given by

$$\boldsymbol{z}_{new} = \boldsymbol{\phi}^T\boldsymbol{z}^{spf} \tag{35}$$

where $\boldsymbol{z}_{new}$ represents a new measurement $\boldsymbol{z}_{new}$ that can be used for normal state estimation obtained by correcting the measurements of the attacked PMU(s). With these data, we can use LSE to estimate the state of the system, that is

$$\hat{\mathbf{x}} = (\boldsymbol{H}^T\boldsymbol{H})^{-1}\boldsymbol{H}^T\boldsymbol{z}_{new} \tag{36}$$

When a PMU is detected to be under attack, i.e., the residual of its measurements exceeds the given threshold, we use an iterative method to minimize these residuals $||\boldsymbol{r}_i||$ and bring this set of data back to the level where safe state estimation can be performed. To this end, the goal of the question is

$$\underset{\alpha}{minimize}\, ||\boldsymbol{r}^{spf}_{max}|| \tag{37}$$

The specific steps of the correction and safe state estimation process are as follows:

(1) If the calculated residual norm is less than the predetermined threshold, i.e., the PMU is not under attack, the algorithm outputs the same measurements, which can be directly used for state estimation.

(2) When the measurements are under attack, we estimate the attack angle by solving the objective function (37). The specific method for minimizing the objective function is as follows:

- Firstly, we select the PMU with the largest residual norm, calculate its residual.
- Then give a priori state estimation $\mathbf{x}$, and minimize the objective function with respect to the angle to estimate the attack angle.
- Finally, use the estimated attack angle to correct the phase angle of the PMU's measurements using equation (29) and update the power system state. Repeat this process until the estimation converges.

(3) Repeat from step 1 until the residual satisfies $||\boldsymbol{r}^{spf}|| < \varepsilon$.

In this way, the difficult problem that the objective function (37) under the coupling of two unknown parameters is a non-convex becomes relatively simple.

We have presented the principles and implementation of the entire algorithm process for attack detection, correction of attacked data, and safe state estimation. The process is summarized in the flowchart shown in Figure 2.

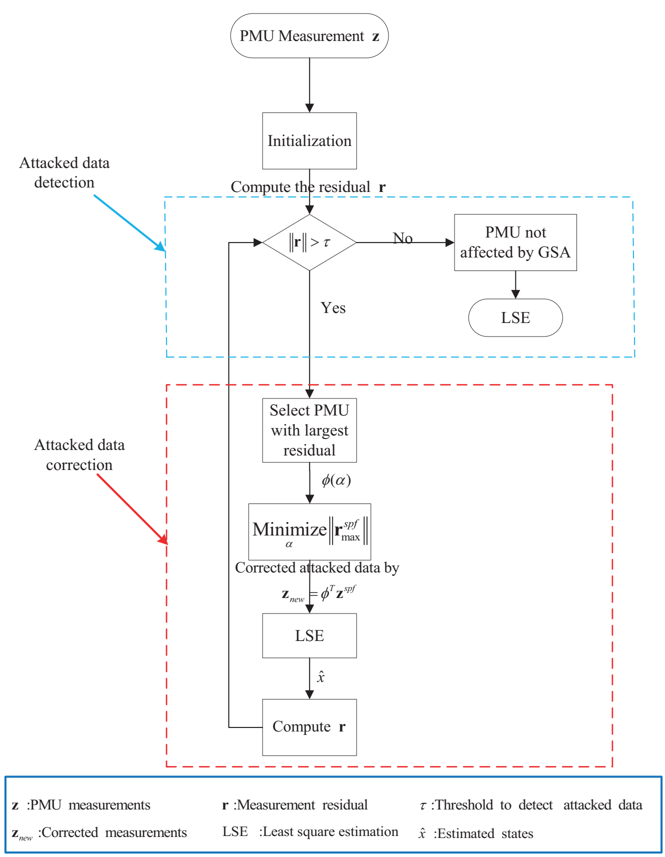

Figure 2. Flow chat of the combined attack detection, attacked data correction, and power system state estimation method.

5. Simulation Results

In this section, we conducted numerical experiments on the GSA detection, data correction, and system state estimation methods proposed in Section 3. The main simulation in this paper were conducted on the IEEE 14-bus system. The IEEE 14-bus test system (i.e., Figure 3) is a commonly used standard test system for evaluating the performance of power system load flow calculation and optimization algorithms. It is based on the IEEE standard 14-bus reference model and is widely used in research and applications in the field of power systems. The bus data, line data, and generator data of this test system all conform to the definitions and specifications of the IEEE standard, making it considered as a standard test system. It is extensively used for validating and comparing different load flow calculation algorithms, optimization algorithms, and other power system analysis methods. Using the IEEE standard test system as a benchmark ensures comparability of research and application results and provides a common platform for the academic and industrial communities. MATPOWER is a MATLAB toolbox used for power system load flow calculation and optimization. For the IEEE 14-bus test system, MATPOWER can provide the required buse data (bus number, bus types), line data (line resistance and reactance, etc.), and load flow calculation results (bus voltage magnitude and phase angle, etc.) as requested in this paper. With this data, the system model described in the paper has a standardized and realistic representation.

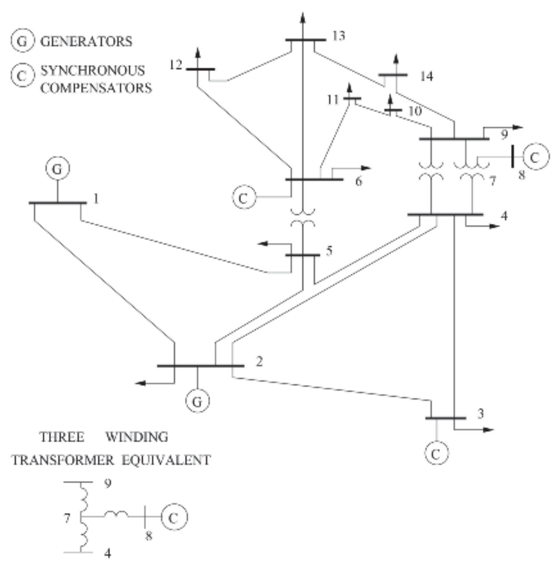

Figure 3. IEEE 14-bus system.

In addition, in the IEEE 14-bus system attack simulation, the threshold value ε that distinguishes the nominal scenario from the spoofing scenario is set to 0.0215. In all our simulations, we assume that the system bus network are observable. Table 1 provides the placement locations of PMUs for different test cases. And in this paper the measurement noise covariance of the PMUs is a diagonal matrix, with a standard deviation of 0.01 for each measurement.

Table 1. PMU buses for different IEEE bus test case.

Test Case	Number of PMUs	PMU Buses
IEEE 14	6	2,4,6,7,10,14
IEEE 30	12	1,3,5,7,9,10,12 18,24,25,27,28
IEEE 118	54	1,3,4,5,6,8,9 11,12,15,17,19,21 23,25,26,28,30,34 35,37,40,43,45,46 49,52,54,56,59,62 63,65,68,70,71,75 76,77,78,80,83,85 86,89,90,92,94,96 100,105,108,110,114

5.1. Attacked Data Detection

In this paper, attacked data detection algorithm uses the residual norm of the measurements as the threshold to distinguish whether data is attacked data generated by GSA. Therefore, the selection of this threshold is very important. We conducted 2000 Monte Carlo simulations for nominal and spoofing scenarios. In the nominal scenario, no modifications were made to the PMU measurements. The highest residual values obtained from static state estimation in these nominal scenarios were utilized as the threshold. Remarkably, during over 1000 simulations of the nominal scenarios, the occurrence rate of instances reaching the maximum residual value is estimated to be around 2 to 3 thousandths of a percent. This suggests that using this threshold to distinguish between secure and attacked data exhibits exceptional robustness. Figure 4 shows the residual norm distributions under different simulation scenarios. The results in Figure 4 clearly demonstrate that the presence of GSA causes an increase in the measurement residuals for PMUs, thus further validating the feasibility of using a set residual threshold to differentiate between attacked and non-attacked systems.

Through the Monte Carlo simulation results, we found that the maximum value of the measurement residual for PMUs under secure conditions is 0.025, which serves as the threshold for differentiating between attacked and non-attacked PMUs based on residual values. The specific method is to generate a series of PMU measurements according to the security measurement Equation (4), then use the traditional static state estimation (5) to estimate the state of the power grid, and finally calculate the measurement residual $\boldsymbol{r}$ under each simulation condition through (30), taking $r_{max} = \varepsilon = 0.0215$.

Since the GSA process requires special equipment and a long time to achieve, and the actual geographical span of the grid is huge, especially the distance between adjacent substations is long. Therefore, it is difficult to coordinate the generation of GSA in multiple locations or PMUs. Therefore, we first consider that only one PMU is subject to GSA on the system. For a single GSA, we consider that the measured phase of bus 2 has shifted by 12° due to the GSA. We record the measurements of all PMUs of the system under the security state and the attack of a single GSA in Figure 5. For multiple GSAs, we consider that the PMU measurement phase on bus 2 and bus 7 has a phase shift of 12° and 60°, respectively. Figure 6 shows a record of the measurements of all PMUs when the above attack occurs on PMUs on bus 2 and bus 7.

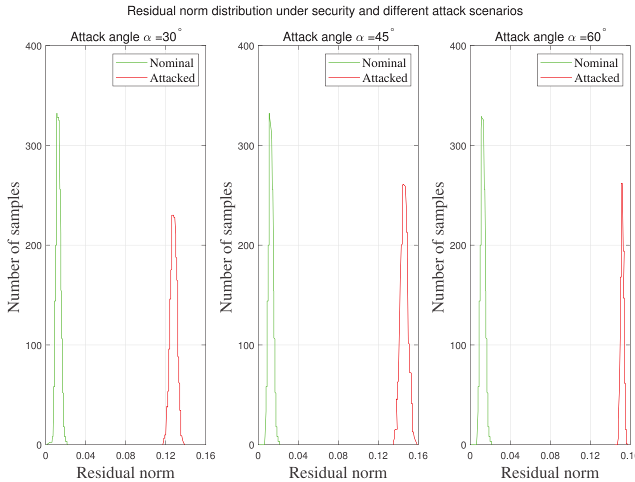

Figure 4. Residual norm distributions for both nominal and attacked scenarios.

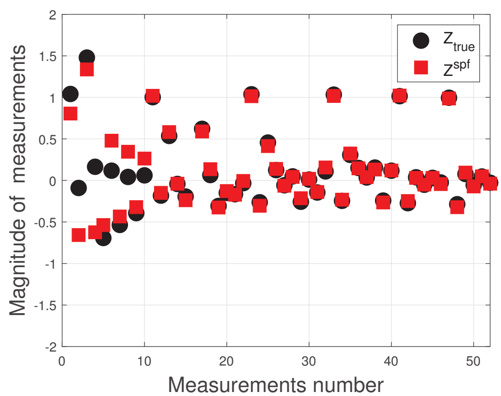

Figure 5. PMU measurements under security conditions and after the PMU on bus 2 has a 12° phase shift due to GSA.

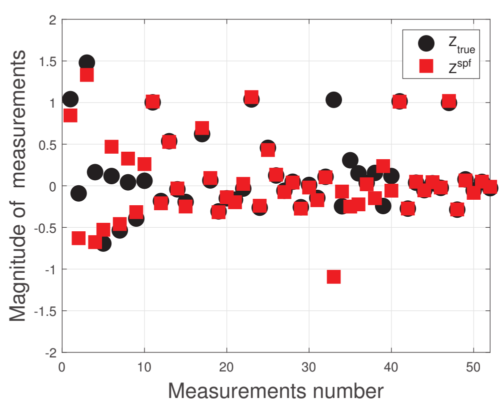

Figure 6. PMU measurements under security conditions and after the PMU on bus 2 and bus 7 have a phase shift of 12° and 60°, respectively.

Figures 5 and 6 show the measurements of the system PMU under a single GSA and multiple GSAs, and the measured results of each PMU under the security state. And the scatter diagram shows that when a PMU was attacked, the measurements will have a large deviation from the actual value, which is enough to significantly deteriorate the system state estimation. On the other hand, from the location of the data with obvious deviation, we can also accurately locate the location of the PMUs under attack. At the same time, this also verifies that it is reasonable for us to use the maximum deviation norm threshold to classify the security and attacked data within a certain range.

5.2. Correction of Attacked Data under Different Number of GSA

To observe the effects after correction, for a single GSA, we still take the PMU on bus 2 as an example, which is attacked to produce a 12° phase shift. In the measurement data detection algorithm, we can obtain a dataset of PMU measurements containing attacked data. After correcting these data with the data correction algorithm, a new set of effective measurements $\boldsymbol{z}_{new}$ can be obtained, satisfying $||\boldsymbol{r}^{new}|| < \boldsymbol{\varepsilon}$. The data correction process is the same for multiple GSAs. Therefore, the measurement data correction results of two different attack scenarios in Section 5.1 are shown in Figure 6 and Figure 7, respectively. The results in Figures 8 and 9 illustrated that the corrected measurements are almost identical to the true values with very small RMSE, indicating their suitability for power system state estimation for secure operation.

In addition, we apply 1–3 GSAs to the IEEE 14-bus system to study the performance of the algorithm. We take the root mean square error (RMSE) of corrected measurements as the performance metric, and RMSE is defined as $\sqrt{(\hat{\mathbf{x}} - \mathbf{x})^2/n}$, where n represents the number of data samples, $\hat{\mathbf{x}}$ represents estimated value and $\mathbf{x}$ is actual value.

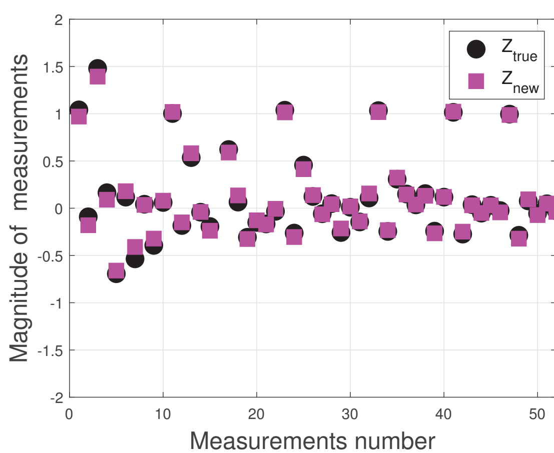

Figure 7. Scatter diagram of the measurements of 12° shift under a single GSA after data correction vs the security measurements.

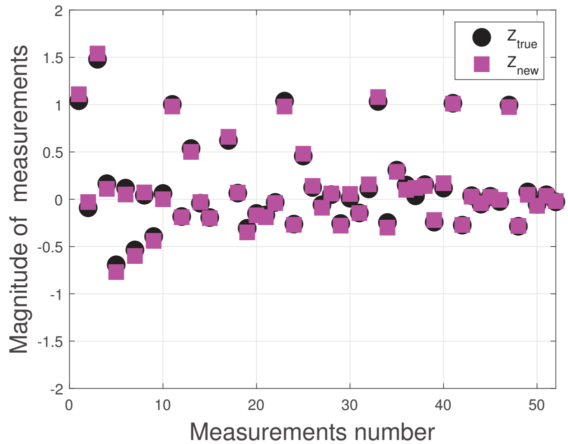

Figure 8. Scatter diagram of the measurements of 12° and 60° shift under multiple GSAs after data correction vs the security measurements.

Figure 9 shows that for the IEEE 14-bus system, the corrected measurements in this paper have a lower RMSE, and the algorithm also has good robustness with the increase of GSAs. Even when the system is subject to three GSAs at the same time, the RMSE of the corrected measurements remains below 0.02.

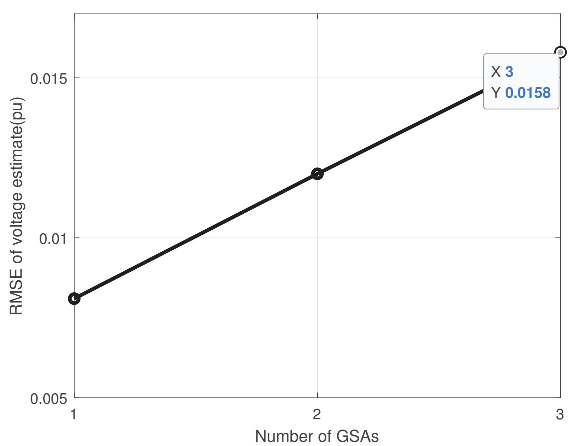

Figure 9. RMSE of corrected measurements at different GSAs.

On the other hand, in our simulation study, we examined the applicability of the algorithm to larger scale networks to showcase its performance on larger networks. We proposed algorithm on the IEEE 14, IEEE 57, and IEEE 118-bus test cases. We performed Monte Carlo simulations in which the number of GSAs are varied from 1 to 3. For each GSA, we perform 100 Monte Carlo simulations in which we randomly spoofed a given number of PMU buses with the attack angles. Table 2 presents the RMSE in the corrected measurements obtained from AM and proposed algorithms. The RMSE of proposed algorithm corrected measurements is smaller than AM for all test cases under multiple GSAs. In addition, Table 2 gave the computation time of all algorithms for 100 Monte Carlo simulations. The computation time of the proposed algorithm is less than AM in all the test cases. The computation time of the proposed algorithm increases with the increase of

GSA as it is an iterative estimator that mitigates one GSA at a time. For the scenario with 3 GSAs, the maximum computation time observed in the IEEE 118-bus test case is 0.0068 s. It is worth noting that all simulation tests were conducted in real-time calculations. This demonstrates that the proposed algorithm performs effectively when applied to large-scale networks and real-time monitoring.

Table 2. Comparison of the proposed algorithm, the computation time of the AM algorithm, and the RMSE of the measurements corrected for different GSAs under the IEEE 14, IEEE 30, and IEEE 118-bus test cases.

Test Case	Scenario	Corrected Measurements RMSE (pu) of Proposed Algorithm	Computation Time (s) of Proposed Algorithm	Corrected Measurements RMSE of AM Algorithm	Computation Time (s) of AM Algorithm
IEEE 14	1 GSAs	0.0093	0.0294	0.0413	8.4909
	2 GSAs	0.0120	0.1144	0.0429	14.6323
	3 GSAs	0.0158	0.2830	0.0522	21.3417
IEEE 30	1 GSAs	0.0026	0.1426	0.0193	87.9859
	2 GSAs	0.0219	0.2421	0.0396	94.3241
	3 GSAs	0.0413	0.3029	0.0641	107.4123
IEEE 118	1 GSAs	0.0033	0.1487	0.0137	91.3233
	2 GSAs	0.0057	0.4023	0.0366	122.4132
	3 GSAs	0.0068	0.7011	0.0539	129.5605

5.3. State Estimate of System with Different Number of GSAs

In the previous section, we used the attacked data correction algorithm to correct the data damaged by GSA into secure data within the specified threshold range. Then we can use these data to estimate the state of the system through (36). We compared the proposed algorithm in this paper with the Alternating minimization (AM) [20] algorithm and the traditional weighted least square (WLS) algorithm under the influence of 1–3 GSAs. The results are shown in Figures 10 and 11.

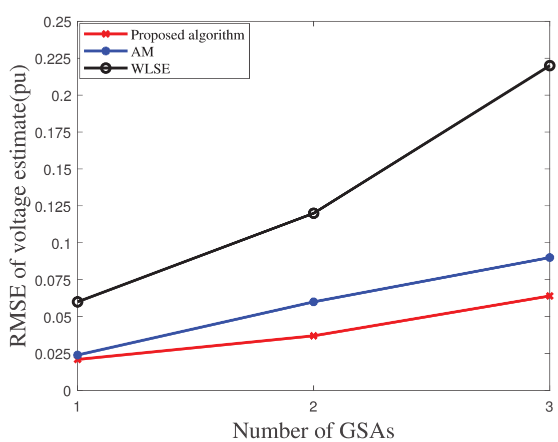

Figure 10. Proposed algorithm vs. AM and WLSE voltage estimate for different GSAs.

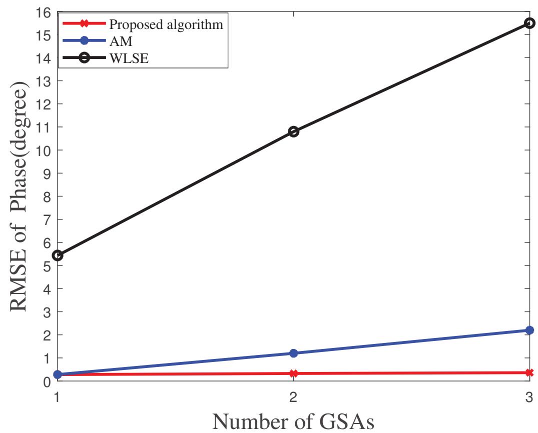

Figure 11. Proposed algorithm vs. AM and WLSE phase estimate for different GSAs.

The results in Figures 10 and 11 show that the proposed algorithm in this paper has a small RMSE for different GSAs. Even under three GSAs, the RMSE of voltage amplitude is below the same order of magnitude. The proposed algorithm can be well used in the process of system state estimation under GSAs.

6. Conclusions

This paper proposes a detection and data correction algorithm for the measurement data deviation caused by the GSAs of PMUs in the power grid, and uses the corrected data to estimate the system state. Through the simulation test on IEEE 14-bus system, we observed that the algorithm can detect the location of GSAs and the corrected data are very close to the original real measurements. And the root mean square error of the system state estimation for the corrected data is also very small, which greatly improves the estimation accuracy. In the generalized simulation, we also found that the algorithm is also applicable to larger scale networks.

Subsequent research will consider the study of the problem under different PMU placements, and develop a new joint estimation algorithm to estimate the two coupled unknown parameters in the model more accurately.

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