Unified Control Scheme of Five-Phase Open-Winding Permanent-Magnet Synchronous Generator Systems for Aerospace Applications

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\begin{abstract}
In order to satisfy with the demand of high reliability in the aerospace field, a unified control scheme is proposed for five-phase open-winding permanent-magnet synchronous generator (FPOWPMSG) in this paper. Firstly, the mathematical model of FPOWPMSG in the phase coordinate system is established. Then, the proposed unified control scheme is presented in detail, which is mainly composed of the unified reference current generator, the current regulator, the synchronous electromotive force (EMF) compensator, and the piecewise nonlinear predictive compensator. In the unified reference current generator, normal operation, single-phase open-circuit, adjacent two-phase open-circuit, and non-adjacent two-phase open-circuit are taken into account by introducing a fault matrix. In the current regulator, the PI regulator is used. And the current tracking accuracy and the torque ripple are improved by both the synchronous EMF compensator and the piecewise nonlinear predictive compensator. Finally, the effectiveness of the proposed unified control scheme is validated by experimental results.
\end{abstract}

\begin{IEEEkeywords}
Five-phase permanent-magnet synchronous generator, open-winding, open circuit fault, fault-tolerant control.
\end{IEEEkeywords}

\section*{NOMENCLATURE}

\begin{itemize}
\item $L_A$ self-inductance of A-phase winding
\item $\psi_{fA}$ permanent magnet flux linkage of A-phase windings
\item $\psi_A$ flux linkage of A-phase winding
\item $L_s$ constant component of stator inductance
\item $L_{q2}$ amplitude of the alternating component of the stator inductance
\item $\omega$ electrical angular velocity
\item $\epsilon_{fA}$ no-load back EMF of the A-phase
\item $R$ resistance of winding
\item $u_A$ phase voltage of A-phase
\item $T_{ex}$ $x$-phase torque ($x = A, B, C, D, E$)
\item $T_e$ total torque of the studied OW-FPMSG
\item $p_n$ pole pairs
\item $\omega_m$ mechanical angular velocity
\item $I$ amplitude of the phase current
\item $F_x$ $x$-phase situation state ($x = A, B, C, D, E$)
\item $pCu$ copper loss
\item $L$ Lagrangian function
\item $I$ phase current vector
\item $I^*$ reference phase current vector
\item $k_p$ proportionality coefficient
\item $k_i$ integral coefficient
\item $T_s$ switch cycle
\item $u_{Ax}^*$ reference phase voltage of A-phase
\item $i_{Ax}^*$ reference phase current of A-phase
\item $i_A$ phase voltage of A-phase
\item $k\Delta$ compensation factor of piecewise nonlinear predictive
\item $\Delta u$ compensated voltage of harmonic inductance
\end{itemize}

\section{I. INTRODUCTION}
High-reliability is one major requirement of aerospace applications. Due to the good fault-tolerant performance, therefore, the multiphase motor has attracted more and more...
attentions recently [1]–[7], and the fault-tolerant performance can be enhanced if the number of motor phases is increased. However, more motor phases will make the control scheme complex [8]. As a compromise of the fault-tolerance and the control complexity, the five-phase motor has received more attentions recently [9]–[12].

So far, a lot of fault-tolerant control methods have been proposed for five-phase motors, they can be mainly divided into two categories: reduced-order decoupling matrix and optimal fault-tolerant current setting [13]–[17]. The former one constructs a decoupling matrix in the synchronous rotating coordinate system. However, different reduced-order decoupling matrices are required according to different fault locations and types [18]–[20]. In other words, unified fault tolerant control cannot be achieved to cope with different faults. The latter one needs to track the AC reference current in the natural coordinate system. The hysteresis current modulation is usually used. But its accuracy of current tracking is closely related to the hysteresis width. If the hysteresis loop width is too large, the average switching frequency and loss will be reduced. But the control performance will be degraded and the tracking error will increase. On the contrary, if the hysteresis loop width is small, the current tracking performance can be improved. However, the average switching frequency and loss will greatly increase [21], [22].

In this paper, a unified control scheme is proposed for five-phase open-winding permanent-magnet synchronous generator (FPOWPMSG) systems in aerospace applications. By using the proposed unified control scheme, the modification of the control structure is not required in no matter healthy or post-fault operations, which is the first contribution of this paper. It should be emphasized that only single-phase and two-phase open-circuit faults are taken into account. Considering the rich harmonic back-EMF and inductance components, both the synchronous EMF compensator and the piecewise nonlinear predictive compensator are proposed to balance the closed-loop current control and the parameter dependence, which is the second contribution of this paper. The rest of this paper is organized as follows. The mathematical model of FPOWPMSG is introduced in Section II. The proposed unified control scheme is presented in Section III. Some experiments are carried out to verify the effectiveness of the proposed unified control scheme in Section IV. Finally, the conclusion is drawn in Section V.

II. MATHEMATICAL MODEL OF FPOWPMSG

The design principle of FPOWPMSG is to achieve electrical, magnetic, thermal isolation, etc., between phases, as shown in FIGURE 1. The FPOWPMSG adopts an open-winding structure to achieve electrical isolation, and the introduction of insulation boards to achieve the thermal isolation. In order to enhance the magnetic isolation, fault-tolerant teeth are adopted in the structure of FPOWPMSG. As a result, all mutual inductances are nearly zero. Therefore, the studied FPOWPMSG can be modelled as five independent phase windings instead of one entire generator.

Take A-phase winding as an example. The flux linkage $\psi_A$ of A-phase winding can be given as

$$\psi_A = L_A i_A + \psi_f$$  \hspace{1cm} (1)

For a non-salient electrical machine, the self-inductance of each phase winding should be a constant value. However, FPOWPMSG is a salient machine, and its self-inductance contains a secondary pulsation component. The self-inductance curve is shown in FIGURE 2, and it can be modelled as

$$L_A = L_s + L_s^2 \cos(2\omega t)$$  \hspace{1cm} (2)

The voltage equation can be given as

$$u_A = R_i_A + L_A \frac{d i_A}{dt} - 2\omega L_s i_A \sin(2\omega t) + e_f$$  \hspace{1cm} (3)

With

$$e_f = -\omega \psi_f \cos \omega t$$  \hspace{1cm} (4)

The torque equation can be given

$$T_e = -2p_s L_s^2 \tan(2\omega t) + \frac{e_f}{\omega_m} i_A$$  \hspace{1cm} (5)

Similarly, the other four phase torques can be obtained, and the total torque of the studied OW-FPMSG can be obtained as

$$T_e = T_eA + T_eB + T_eC + T_eD + T_eE$$  \hspace{1cm} (6)
III. UNIFIED FAULT-TOLERANCE CONTROL

In this section, the proposed unified control scheme is presented in FIGURE 3, which mainly contains four modules: the unified reference current generator, the current regulator, the synchronous EMF compensator and the piecewise non-linear predictive compensator.

A. UNIFIED REFERENCE CURRENT GENERATOR

In this module, three typical open-circuit fault types are taken into account: single-phase open-circuit, adjacent two-phase open-circuit, and non-adjacent two-phase open-circuit. As listed in TABLE 1, there are 15 open-circuit fault candidates in this paper, including 5 single-phase open-circuit faults, 5 adjacent two-phase open-circuit faults and 5 non-adjacent two-phase open-circuit faults. Considering the normal operation, there are 16 different operation situations in this paper.

When an open-circuit fault occurs, the rotating magnetic motive force (MMF) should remain unchanged. In other words, all healthy phase windings should be able to generate the same rotating MMF no matter in which situation. The corresponding constraint condition can be given as

\[ \mathbf{T} \mathbf{I} = \mathbf{C} \]  

(7)

With

\[ \mathbf{T} = \begin{bmatrix} 1 & \cos \frac{2\pi}{5} & \cos \frac{4\pi}{5} & \cos \frac{6\pi}{5} & \cos \frac{8\pi}{5} \\ 0 & \sin \frac{2\pi}{5} & \sin \frac{4\pi}{5} & \sin \frac{6\pi}{5} & \sin \frac{8\pi}{5} \end{bmatrix} \]

\[ \mathbf{C} = \begin{bmatrix} \frac{5}{2} I \cos (\omega t) \\ \frac{5}{2} I \sin (\omega t) \end{bmatrix}^T \]

\[ \mathbf{I} = \begin{bmatrix} i_A(t) & i_B(t) & i_C(t) & i_D(t) & i_E(t) \end{bmatrix}^T \]

In order to simplify the fault-tolerant control, a fault matrix \( \mathbf{F} \) is introduced as follows

\[ \mathbf{F} = \begin{bmatrix} F_A & 0 & 0 & 0 & 0 \\ 0 & F_B & 0 & 0 & 0 \\ 0 & 0 & F_C & 0 & 0 \\ 0 & 0 & 0 & F_D & 0 \\ 0 & 0 & 0 & 0 & F_E \end{bmatrix} \]

(8)

In normal operation, \( F_x \) is set as 1; otherwise, \( F_x \) is set as 0.

According to (8), (7) can be rewritten as

\[ \mathbf{T} \mathbf{F} \mathbf{I} = \mathbf{C} \]

(9)

The challenge has become how to determine the reference phase current vector \( \mathbf{I}^* \). Because \( \mathbf{T} \mathbf{F} \mathbf{I} \) is a \( 2 \times 5 \) matrix, the reference phase current vector \( \mathbf{I}^* \) cannot be directly determined according to (9). In this paper, a minimum copper loss principle is proposed to determine the reference phase current vector \( \mathbf{I}^* \). The copper loss of the studied FPOWPMSG can be represented as

\[ P_{Cu} = R \mathbf{I}^T \mathbf{I} \]

(10)

A Lagrangian function \( L \) constructed with the copper loss is selected as the objective function

\[ L = R \mathbf{I}^T \mathbf{I} + \lambda^T (\mathbf{T} \mathbf{F} \mathbf{I} - \mathbf{C}) \]

(11)
where $\lambda = \begin{bmatrix} \lambda_1 \lambda_2 \end{bmatrix}$. Making the derivative of each variable in the Lagrangian function (11) zero, the reference phase current vector $I^*$ can be obtained as

$$I^* = (TF)^T \left[ TF (TF)^T \right]^{-1} C$$  \hspace{1cm} (12)

**B. CURRENT REGULATOR**

In this module, the PI regulator is used to follow the AC current in the natural coordinate system. Taking A-phase as an example, when there is only a current regulator, the expression of the phase voltage reference is shown in equation (13).

$$u_A^* = k_p (i_A^* - i_A) + k_i \int (i_A^* - i_A) \, dt$$  \hspace{1cm} (13)

Theoretically, when the switching frequency is high enough, the current in each switching cycle can be regarded as a constant value. However, the switching frequency of the medium and high-power devices is limited at present. The phase current electric frequency is high for the FPOWPMSG,
and there are fewer control commands applied per electric cycle. It is difficult to achieve better tracking control performance only through the current regulator.

C. SYNCHRONOUS EMF COMPENSATOR

In this module, synchronous EMF compensation is added to compensate for the influence of EMF and delay on control performance. Firstly, the model of FPOWMSG in the phase coordinate system is analyzed. From the phase voltage equation (3), it can be known that the phase voltage is composed of resistance voltage drop, inductance voltage drops and back EMF. The back EMF is the derivative of the permanent magnet flux linkage with respect to time, and the permanent magnet flux linkage is a function of position. Therefore, the back EMF must be related to the rotational speed and the derivative of the permanent magnet flux linkage with respect to position, and only related to these two variables. Therefore, in the case of known rotational speed and permanent magnet flux linkage, the back EMF can be obtained. By directly compensating the calculated back EMF in the control variable, the reference voltage that needs to be tracked by PI controller can be reduced, and better tracking performance under limited PI parameters can be achieved.

Due to the limitation of the manufacturing process, there is a certain deviation between the actual and rated parameters of FPOWMSG. In order to accurately compensate the back EMF, the measured no-load back EMF can be fitted to obtain the value of each harmonic flux linkage. The measured back EMF waveform is shown in FIGURE 4 at 3500r/min.

After Fourier decomposition of FIGURE 4, it can be seen from FIGURE 5 that the back-EMF waveform contains not only the fundamental and third harmonic components, but also the fifth, seventh, ninth, and eleventh harmonic components.

Therefore, the fitting formula of the back EMF can be obtained as shown in (14).

\[

e_{r} = 193 \sin (700\pi t + 0.9\pi) + 21.616 \sin (2100\pi t + 0.687\pi) + 3 \sin (3500\pi t - 0.494\pi) + 5.964 \sin (4900\pi t + 1.278\pi)
\]

According to (14), the harmonics of the flux linkage can be obtained, as shown in TABLE 2.

| Flux linkage | Amplitude/Wb |
|--------------|--------------|
| \(\psi_f\)   | 0.088        |
| \(\psi_f\)   | 3.276×10³   |
| \(\psi_f\)   | 2.728×10⁴   |
| \(\psi_f\)   | 3.874×10³   |
| \(\psi_f\)   | 2.425×10⁴   |
| \(\psi_f\)   | 8.536×10⁵   |

The expression of back EMF in equation (4) is extended to equation (15).

\[
e_{r} = -\omega \psi f_1 \sin \left( \omega t - k_n \frac{2\pi}{5} \right) - 3 \omega \psi f_3 \sin \left( 3\omega t - k_n \frac{6\pi}{5} \right)
+ 5 \omega \psi f_5 \sin \left( 5\omega t - k_n \frac{10\pi}{5} \right)
+ 7 \omega \psi f_3 \sin \left( 7\omega t - k_n \frac{14\pi}{5} \right)
+ 9 \omega \psi f_1 \sin \left( 9\omega t - k_n \frac{18\pi}{5} \right)
+ 11 \omega \psi f_3 \sin \left( 11\omega t - k_n \frac{22\pi}{5} \right)
\]

FIGURE 6 and FIGURE 7 are the back-EMF waveforms at 1500r/min and 2500r/min fitted by equation (19), and both waveforms have a good fitting effect.

After introducing back-EMF compensation, (13) can be modified as

\[
u^*_A = k_p (i^*_A - i_A) + k_i \int (i^*_A - i_A) dt + e_{fA}
\]
signal used at $kT_s$ is calculated at $(k-1)T_s$. That means there is a delay of $T_s$ in sampling. In addition, in the case of a known modulation signal, it takes half a cycle to convert it into a PWM signal and add it to the phase windings through the inverter, which means there is a $0.5T_s$ delay in the PWM output. Therefore, when calculating the angle of the back EMF, it is necessary to compensate the delay. The delay problem cannot be ignored. It is necessary to compensate the delay of the control part. The phase voltage reference after adding delay compensation is shown in equation (17).

$$
u_A^* (\{(k-1) T_s\} ) = k_p (\nu_A^* (\{(k-1) T_s\} ) - i_A [\{(k-1) T_s\} ] )$$
$$+ k_i T_s (i_A^* (\{(k-1) T_s\} ) - i_A [\{(k-1) T_s\} ] )$$
$$+ v_A^* [\{(k-2) T_s\} ]$$
$$+ e_f A [\theta (\{(k-1) T_s\} ) + 1.5\omega T_s]$$

(17)

D. PIECEWISE NONLINEAR PREDICTIVE COMPENSATOR

In this module, the impact of dead time is analyzed. Theoretically, the upper and lower IGBT are complementary conduction, but in the actual circuit, the dead time is usually added to prevent the bridge arm from passing through. In order to analyze the influence of the dead time, take the winding of A-phase as an example to analyse its working principle. As shown in FIGURE 8, in the power generation mode, when the current is positive, it output energy through a+-D1-R-D4-a-, or energy storage through T2 and D4; when the current is negative, it output energy through a–D3-R-D2-a+, or energy storage through T4 and D2.

It can be obtained that regardless of the positive half cycle or the negative half cycle, the two states of output energy and stored energy alternate. Only when the energy is stored, the current flows through the switch tube, and when the energy is output, the current flows through the diode. Therefore, the dead time of the upper and lower IGBT will not affect the alternation of the two states in a half cycle. However, when the current is from a positive half cycle to a negative half cycle, or from a negative half cycle to a positive half cycle, due to the low switching frequency and the large current change rate, the reference will instantly reverse. Therefore, the two energy storage states usually alternate, and the dead time of the upper and lower IGBT will cause the phase currents to be distorted.

A piecewise nonlinear predictive compensation strategy is proposed to reduce the phase current distortion caused by the dead time. The given phase voltage is shown in equation (18).

$$u_A^* (\{(k-1) T_s\} ) = k_p (i_A^* (\{(k-1) T_s\} ) - i_A [\{(k-1) T_s\} ] )$$
$$+ k_i T_s (i_A^* (\{(k-1) T_s\} ) - i_A [\{(k-1) T_s\} ] )$$
$$+ u_A^* [\{(k-2) T_s\} ]$$
$$+ e_f A [\theta (\{(k-1) T_s\} ) + 1.5\omega T_s] + \Delta u$$

(18)

The compensation calculation needs to differentiate the current, and the actual current has high-frequency signals such as measurement noise, which will be amplified during the differentiation. In order to avoid the amplification of high-frequency interference signals in the differential process, the current reference is used to predict the compensation amount. The expression is shown in formula (19).

$$\Delta u = \begin{cases} 
  k_L (\theta) \frac{di_A^*}{dt} & 0 \leq \theta \leq \frac{\pi}{3} \\
  0 & \frac{\pi}{3} \leq \theta \leq \pi \\
  k_L (\theta) \frac{di_A^*}{dt} & \pi \leq \theta \leq \frac{4\pi}{3} \\
  0 & \frac{4\pi}{3} \leq \theta \leq 2\pi 
\end{cases}$$

(19)
E. COMPARISON

In this subsection, the proposed unified control scheme is compared with the deadbeat predictive control.

Because the PI current regulator is eliminated in the deadbeat predictive control, the closed-loop current control depends on the accurate mathematical model of the electrical machine. Considering the parameter variation, the closed-loop current control cannot be well achieved by using the deadbeat predictive control. On the other hand, the back-EMF of FPOWPMSG contains significant third harmonic component and its self-inductance contains a secondary pulsation component, which makes the deadbeat predictive control difficult.

Because the generator parameters are also used in both the synchronous EMF compensator and the piecewise nonlinear predictive compensator, the performances of the proposed unified control scheme also affected by the parameter variations. However, due to the existence of PI current regulators, the closed-loop current control can be achieved by using the proposed unified control scheme, which reduce the parameter dependence. By adding both the synchronous EMF compensator and the piecewise nonlinear predictive compensator, the performances of the FPOWPMSG system can be improved.

IV. EXPERIMENTAL VALIDATION

As shown in FIGURE 9, a FPOWPMSG system is built to verify the effectiveness of the proposed unified control scheme. The experimental platform includes a three-phase permanent magnet synchronous motor (PMSM) as a traction motor, a 20kW FPOWPMSG, a high-power electronic load that simulates electrical loads, a three-phase motor controller, a five-phase generator controller and other necessary equipment. The three-phase PMSM and the FPOWPMSG are coaxially connected by a coupling. During the experiment, the three-phase PMSM is used to drive the FPOWPMSG at a constant speed of 3500r/min. The DC bus voltage is set as 270V. The structure diagram of FPOWPMSG is further shown in FIGURE 10, and the specific parameters of the motor are given in TABLE 3.

In this section, three different control schemes (Scheme 1, Scheme 2, and Scheme 3) are defined as follows:

Scheme 1—the proposed unified control scheme without the synchronous EMF compensator and the piecewise nonlinear predictive compensator;
A. NORMAL OPERATION

In this subsection, three control schemes are compared in normal operation. Hence, \( F_A \sim F_E \) are 1. The experimental results are shown in FIGURE 11 and FIGURE 12. The load resistance is 10 \( \Omega \) in FIGURE 11 while 5 \( \Omega \) in FIGURE 12. It can be found that the current tracking accuracy of Scheme 3 is best while those of Scheme 1 is worst.

B. SINGLE-PHASE OPEN-CIRCUIT OPERATION

In this experiment, A-phase is opened, and the load is 10 \( \Omega \). Hence, \( F_A \) is 0 while \( F_B \sim F_E \) are 1. According to (12) and TABLE 1, the reference phase currents can be obtained as

\[
\begin{align*}
    i_A^*(t) &= 0 \\
    i_B^*(t) &= 1.082I \cos(\omega t - 0.342\pi) \\
    i_C^*(t) &= 1.471I \cos(\omega t - 0.869\pi) \\
    i_D^*(t) &= 1.471I \cos(\omega t + 0.869\pi) \\
    i_E^*(t) &= 1.082I \cos(\omega t + 0.342\pi)
\end{align*}
\]

(20)

Three schemes are compared and the experimental results are illustrated in FIGURE 13. Similarly, Scheme 3 also shows best current tracking accuracy. In FIGURE 13, the fundamental wave amplitude of the four healthy phase currents are 16.8A, 23.64A, 22.68A and 17.02A, respectively. They are 1.084 times, 1.525 times, 1.463 times and 1.098 times of 15.5A, which satisfy with the theoretical analysis.

C. ADJACENT TWO-PHASE OPEN-CIRCUIT OPERATION

In this experiment, A-phase and B-phase are opened, and the load is 10 \( \Omega \). Hence, \( F_A \) and \( F_B \) are 0 while \( F_C \sim F_E \) are 1. According to (12) and TABLE 1, the reference phase currents

Scheme 2-the proposed unified control scheme without the piecewise nonlinear predictive compensator;

Scheme 3-the proposed unified control scheme.

FIGURE 12. Experimental results in normal operation (load = 5\( \Omega \)). (a) Scheme 1; (b) Scheme 2; (c) Scheme 3.

FIGURE 13. Experimental results in single-phase open-circuit operation (load = 10\( \Omega \)). (a) Scheme 1; (b) Scheme 2; (c) Scheme 3.

FIGURE 14. Experimental results in adjacent two-phase open-circuit operation (load = 10\( \Omega \)). (a) Scheme 1; (b) Scheme 2; (c) Scheme 3.
can be obtained as

\[
\begin{align*}
    i_A^*(t) &= 0 \\
    i_B^*(t) &= 0 \\
    i_C^*(t) &= 1.466I \cos(\omega t - 0.845\pi) \\
    i_D^*(t) &= 2.099I \cos(\omega t + 0.8\pi) \\
    i_E^*(t) &= 1.466I \cos(\omega t + 0.446\pi)
\end{align*}
\]  

(21)

Three schemes are compared and the experimental results are illustrated in FIGURE 14. In this operation condition, the current tracking accuracy order of three control schemes are Scheme 1 < Scheme 2 < Scheme 3. In FIGURE 14, the fundamental wave amplitude of the three healthy phase currents are 23.1A, 33.24A and 22.59A, respectively. They are 1.490 times, 2.145 times and 1.457 times of 15.5A, which also satisfy with the theoretical analysis.

D. NON-ADJACENT TWO-PHASE OPEN-CIRCUIT OPERATION

In this experiment, A-phase and C-phase are opened, and the load is 10 \( \Omega \). Hence, \( F_A \) and \( F_C \) are 0 while \( F_B \), \( F_D \) and \( F_E \) are 1. According to (12) and TABLE 1, the reference phase currents can be obtained as

\[
\begin{align*}
    i_A^*(t) &= 0 \\
    i_B^*(t) &= 1.083I \cos(\omega t - 0.4\pi) \\
    i_C^*(t) &= 0 \\
    i_D^*(t) &= 2.3I \cos(\omega t + 0.976\pi) \\
    i_E^*(t) &= 2.3I \cos(\omega t + 0.224\pi)
\end{align*}
\]  

(22)

Three schemes are compared and the experimental results are illustrated in FIGURE 15. Scheme 3 still shows the best current tracking accuracy. In FIGURE 15, the fundamental wave amplitude of the three healthy phase currents are 17.45A, 34.63A and 38.35A, respectively. They are
torque ripple can be improved while the control structure can control scheme, both the current tracking accuracy and the affection of the dead-time. By using the proposed unified nonlinear predictive compensator is designed to reduce required. In order to improve the current tracking accuracy, currents can be easily calculated by introducing a fault matrix, proposed for aerospace applications. The reference phase is calculated according to (23) and they are illustrated in FIGURE 16∼18. The torque ripple before and after the fault tolerance are listed in TABLE 4. It can be obtained that Scheme 3 has smallest torque ripple in post-fault operation.

\[ T_e = 2.5p_n \left[ \psi_f i_d t_1 + 3\psi_f i_q t_3 + 2L_s i_d i_q t_1 \right] \]  

E. PERFORMANCE COMPARISON

In order to comprehensively evaluate the performances of three control schemes under different operation conditions, the torque is calculated according to (23) and they are illustrated in FIGURE 16∼18. The torque ripple before and after the fault tolerance are listed in TABLE 4. It can be obtained that Scheme 3 has smallest torque ripple in post-fault operation.

V. CONCLUSION

In this paper, a unified control scheme of FPOWPMSG is proposed for aerospace applications. The reference phase currents can be easily calculated by introducing a fault matrix, in which the modification of the control structure is not required. In order to improve the current tracking accuracy, the back EMF is synchronously compensated while a piecewise nonlinear predictive compensator is designed to reduce the affection of the dead-time. By using the proposed unified control scheme, both the current tracking accuracy and the torque ripple can be improved while the control structure can be simplified.

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