Robust Fault-Tolerant Synergetic Control for Dual Three-Phase PMSM Drives Considering Speed Sensor Fault

LI XIAO1, LI YI ZHANG1, FENG GAO2, AND JIALIN QIAN1
1 College of Information Engineering, Tianjin University of Commerce, Tianjin 300134, China
2 Shenhua Tianjin Coal Terminal Company Ltd., Tianjin 300452, China
Corresponding author: Li Xiao (xiaoli060755@126.com)
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ABSTRACT Stability and reliability are the most important indicators of drive systems. In this article, a robust nonlinear control method is advanced for dual three-phase permanent-magnet synchronous machine (dual-PMSM) drives, based on the synergetic control theory. This control is implemented by a new synergetic controller (SGC) integrating with an extended sliding mode observer (ESMO), and an optimization method is applied to obtain the optimal parameters of the SGC. This control scheme not only improves the dynamic performance even with operating parameters change, but also can diagnose the speed sensor fault and realize the fault tolerant control (FTC). The proposed control method is compared with the PI control, the sliding mode control (SMC) and the predictive control, as well as the simulation and experimental results confirm the superior dynamic performance and the effectiveness on the speed sensor fault diagnosis (FD) and FTC.

INDEX TERMS Synergetic controller, fault diagnosis, fault-tolerant control, speed sensor fault, dual three-phase permanent-magnet synchronous machine (dual-PMSM).

I. INTRODUCTION
Permanent magnet synchronous machines (PMSMs) using in variable speed drives are acquired significant regard owing to their large power density. Compared with ordinary PMSMs, multiphase machines possess superior property such as smaller torque ripple, larger power rating and powerful fault tolerance [1], [2]. The dual three-phase machine drive (dual-PMSMs) as a popular multiphase machine drive, is received benefits from the collaboration of two groups of three-phase windings to vanish the torque harmonic pulsation. Meanwhile, the advanced control technology and reliability issue are the essential property of driving systems in various industrial and military applications [3]. At present, most research on dual-PMSM drive systems used in practical applications, that employs proportional–integral (PI) controllers owing to their simplicity of structure and principle [4]. However, PI controllers is difficult to achieve perfect response performances for dual-PMSM drive systems owing to their characteristics of high nonlinear and strong coupling, especially dual-PMSMs correspond with the sensor fault. With the development of predictive control technology, predictive controllers have applied to the dual-PMSM drives to attain superior performance [4], [5]. To improve the robustness against the periodic disturbances, a dual-loop predictive control structure for PMSMs is advanced [6]. Meanwhile, a hybrid direct torque control method is employed to improve operation performance for dual three-phase PMSM [7]. But this advanced control methods are irrelevant to the FTC for the sensor fault. To enhance stability and fault tolerance of dual-PMSM drive systems, several efforts have been contributed to exploit the reliable and effective FTC.

Previously, various of FD and FTC techniques have been reported in many literatures for the speed sensor fault. The general sensor fault detection strategies can be assorted into two categories: signals-based approaches and model-based approaches. For one thing, signal-based approaches are almost achieved by extracting particular indexes of faulty signals. In [8], the speed sensor fault is diagnosed and tolerated by approximately estimating the rotor angle speed to be the
rotating speed of stator flux. In [9], by combining binary Hall-effect sensors with high frequency signal injection, a low-cost sensor FD method is implemented for PMSM. For another, the main principle of model-based approaches is relied on models of electrical motor to obtain errors between measured signals and estimated signals used as fault indexes. Indeed, observer-based schemes are mostly used to acquire estimated signals. An adaptive observer and an extended Kalman filter are respectively introduced to detect the speed sensor fault for PMSM in [10] and [11]. In [12], [13], a sliding mode observer is applied to sensor fault diagnosis and system reconfiguration. In [14], based on extreme learning machine technology, the intelligent time-adaptive data-driven method to identify the sensor fault.

In the respect of FTC for machine driving system, the main principle common relies on substituting abnormal feedback signals with estimated signals to tolerate confused sensor signals [15]. In practice, observer-based approaches for estimating signals are reliably and effectively. In [16], the FTC method for the fault speed sensor is relied on compensating abnormal speed feedback values. In [17], an adaptive residual generator is applied to recognize irregular varies and an iterative tuning algorithm is advanced to optimize the system performance, which are achieved in the FTC. Further, FTC is implemented based on observer approaches reliably and effectively. In [11], sensor FTC using a voting scheme is developed for PMSMs, yet the voting scheme is too complex for the speed sensor fault diagnosis and two observers are employed in FTC. In [18] and [19], a simple adaptive observer-based scheme and a sliding mode observer (SMO) are involved in the FTC strategy, where the SMO contributes on stabilization even in present of the variable reference speed and external disturbances.

For multiphase machines, the advanced control methods involved in the FTC can be classified into three categories: direct torque control (DTC), field-oriented control (FOC) and model predictive control (MPC). Though DTC is not related to the coordinate rotation transformation and carried out easily, a stator flux observer is always constructed to achieve related position signals in [20], [21], which leads to the lower reliability. MPC as a multivariable control method, becomes more and more popular. In [22], MPC applied to the five-phase PMSM, that is successfully tolerated for the faulty phase winding. However, the main shortcoming of MPC-based FTC is that the scheme calculation is larger and the weighting factor is difficult to decide. FOC-based FTC method for the multiphase machine drive system correspond with one faulty phase are discussed in [23]– [27], and fault tolerance performances are effective. However, the dynamic response of FOC-based FTC is limited since the parameters of PI controller would overspend time by the designed experience. Thus, performances of FOC-based FTC for the speed sensor fault depend on two aspects. One aspect is that accuracy and effectiveness of estimated signals that replacing to abnormal feedback signals. The other is that dynamic response performances of the speed controller.

Recently, synergetic control theory is used to design the nonlinear control system, which is a state-space method originated from the combination of modern mathematics and synergy [28]. Since the synergetic control has inherent advantages such as the global stability and the robustness against modeling inaccuracy and internal parameter disturbances, it has already used in power system, yet not applied to the machine driving system. Similar to the SMC method, the synergetic control method can also reduce the order of the controlled system, but it is more suitable for digital control and can eliminate the chattering problem [29], [30]. Besides, the synergetic control theory is used in the FTC. An adaptive synergetic observer and a reconfigurable controller are built to compensate for pitch actuator fault in [31]. Thus, in this paper, an optimized SGC is designed to substitute to the PI controller in the closed-loop control system of the dual-PMSM to generate a superior response and control performances. Besides, in the SGC-based closed-loop control system, an observer is employed to provide estimated values to define manifold. By this way, a sensor FD and FTC methods can be developed based on the SGC integrating with an observer.

The rest of this paper is organized as follows: the model and principle control of dual-PMSM drive systems are depicted in Section II. Afterwards, the SGC controller is designed and optimized in Section III. In Section IV, a SGC integrating with EMSO is developed to diagnose and tolerate the speed sensor fault. Then, FD and FTC strategies for the speed sensor fault are proposed in Section V. Finally, the effectiveness of advanced strategies for dual-PMSM corresponding with variable torque and speed is confirmed by simulation and experiments in Section VI, as well as the comparison between the SGC and other types of controllers is also presented in this section.

II. DECOUPLED MODEL OF DUAL-PMSM

The dual-PMSM involves in two sets of Y-type three-phase stator windings diverted by 30° within separated neutral points. Two sets of stator windings that indicated by (A,B,C) and (X,Y,Z) lay in the stator but not fracture with each other, as shown in Fig.1.
According to the Fig.1, the model in natural coordinate system of dual-PMSM can be obtained as follows:

\[
\begin{align*}
\dot{u}_s &= R_s i_s + \frac{d\psi_s}{dt} \\
\dot{\psi}_s &= L_s i_s + \gamma_s \dot{\psi}_f
\end{align*}
\]

(1)

where \(u_s = [u_A u_B u_C u_X u_Y u_Z]^T\), \(i_s = [i_A i_B i_C i_X i_Y i_Z]^T\), \(\psi_s = [\psi_A \psi_B \psi_C \psi_X \psi_Y \psi_Z]^T\), \(R_s = \text{diag} [R_s R_s R_s \psi_s \psi_s \psi_s]^T\).

\(u_s\) is the stator phase voltage vector, \(i_s\) is the stator phase current vector, \(\psi_s\) is the stator phase flux vector, \(\gamma_s\) is the magnetic coefficient matrix, \(\dot{\psi}_f\) is the flux amplitude. The motion equation of the dual-PMSM is given by

\[
T_e - T_L - B\omega_m = J \frac{d}{dt} \omega_m
\]

(2)

where \(T_e\) and \(T_L\) respectively denote the electromagnetic torque and load torque, \(J\) and \(B\) respectively express the moment of inertia and damping coefficient, \(\omega_m\) is mechanical angular velocity. Since the dual-PMSM is a nonlinear high-order system, it is difficult to analysis and control precisely. By using the vector space decomposition (VSD) method [2], the voltage and flux space vectors of the dual three-phase PMSM in the coordinate system that defined as \(D-Q-D_0-Q_0\), can be decoupled as follows:

\[
\begin{bmatrix}
u_D \\
\psi_D \\
\psi_{DD} \\
\psi_{D0}
\end{bmatrix}
=
\begin{bmatrix}
R_s & 0 & 0 & 0 & i_d \\
0 & R_s & 0 & 0 & i_q \\
0 & 0 & R_s & 0 & i_{dq} \\
0 & 0 & 0 & R_s & i_{d0}
\end{bmatrix}
\begin{bmatrix}
\dot{i}_d \\
\dot{i}_q \\
\dot{i}_{dq} \\
\dot{i}_{d0}
\end{bmatrix}
+ p
\begin{bmatrix}
\psi_D \\
\psi_{D0} \\
\psi_{DD} \\
\psi_{D0}
\end{bmatrix}
+ \omega
\begin{bmatrix}
-\psi_Q \\
-\psi_{Q0} \\
-\psi_{QQ} \\
-\psi_{Q0}
\end{bmatrix}
\]

(3)

\[
\begin{bmatrix}
\psi_D \\
\psi_Q \\
\psi_{D0} \\
\psi_{Q0}
\end{bmatrix}
=
\begin{bmatrix}
L_D & 0 & 0 & 0 & i_d \\
0 & L_Q & 0 & 0 & i_q \\
0 & 0 & L_{dQ} & 0 & i_{dq} \\
0 & 0 & 0 & L_{QQ} & i_{d0}
\end{bmatrix}
\begin{bmatrix}
\dot{i}_d \\
\dot{i}_q \\
\dot{i}_{dq} \\
\dot{i}_{d0}
\end{bmatrix}
+ \sqrt{2} \psi_f
\]

(4)

where \(L_D = L_d + L_{dd} = 3L_{aad} + L_{1s} = L_{d0}, L_Q = L_q + L_{qq} = 3L_{aqq} + L_{1s} = L_{q0}, L_d - L_{dd} = L_{1s}, L_q - L_{qq} = L_{1s}, L_{aad}\) and \(L_{aqq}\) are the main self-inductance of \(d\)-axis and \(q\)-axis, respectively. \(\omega\) is the electrical angular frequency. \(L_{1s}\) is the leakage inductance of the stator winding, \(\dot{\psi}_f\) is amplitude of rotor flux. \(T_e\) in the \(D-Q-D_0-Q_0\) coordinate system is denoted by

\[
T_e = \frac{3}{2} n_p \left[ (i_Q \psi_D - i_D \psi_Q) + (i_{d0} \psi_{D0} - i_{D0} \psi_{Q0}) \right]
= \frac{3}{2} n_p \left[ \sqrt{2} \psi f i_Q + (L_d - L_q) i_D i_Q \right]
\]

(5)

From the expression (3) and (4), it can be seen that the voltage, flux linkage and torque in the \(D-Q-D_0-Q_0\) coordinate system are all decoupled completely. Variables in the \(d_1-q_1-d_2-q_2\) coordinate system that participate in the electromechanical energy conversion are all projected to the \(D-Q\) subspace, yet variables in the \(d_1-q_1-d_2-q_2\) coordinate system that independent of the electromechanical energy conversion are all projected to the \(D_0-Q_0\) subspace. The components in the \(D-Q-D_0-Q_0\) coordinate system are presented in the Table 1.

### III. DESIGN AND PARAMETERS OPTIMIZATION OF SGC

#### A. SYNERGETIC CONTROL THEORY

The essential of the SGC is to define a control manifold that contains state variables, which is the control target of the designed controller, and the control system will eventually reach to the steady state along with the designed manifold. The first step of designing SGC is to define macro-variables based on characteristics of the system, and then to search the control discipline according to the manifold convergence equation and the system state space model. The SGC has a good robustness against modeling inaccuracies and parameter disturbances. For complex nonlinear systems, it can eliminate the chattering problem to robust against the noise or disturbances since it implements at a fixed switching frequency. The detailed derivation process has been presented in [30].

Assuming that the control system is a \(N\)-dimensional nonlinear system, which is defined by

\[
\dot{x} = f(x, \Delta u, t)
\]

(6)

where \(x\) is the system state vector, \(\Delta u\) is the control input, and \(f(\cdot)\) is a continuous nonlinear function. To achieve control objectives that make the state of the control system astringents from any point in state space to the designed manifold, the macro-variable is denoted by

\[
\psi(x) = 0
\]

(7)

The dynamic equation is defined to describe the convergence process that the manifold gradually approaches to the steady point as follows:

\[
T \dot{\psi} + \psi = 0, T > 0
\]

(8)

where \(T\) denotes a time parameter that the system converges from the initial state to the manifold. Since macro variables are functions of state variables, the (7) is derivative as

\[
\dot{\psi} = \frac{d\psi}{dx} \dot{x}
\]

(9)
Substitute equations (6) and (9) into the equation (8), that is
\[ T \frac{d^2 \psi}{dx^2} f(x, \Delta u, t) + \psi = 0 \] (10)

Due to introduce the manifold equation, the order of dual-PMSM driving systems can be reduced. The solution of the variable \( \Delta u \) from the equation (10) can emerge predictive control performances. The motion discipline of the nonlinear system controlled by the SGC converges to the manifold along with the dynamic equation (8), thus the evolutionary discipline of the \( i \)-th component in macro-variables is given by
\[ T_i \dot{\psi}_i + \psi_i = 0, \quad T_i > 0 \] (11)

Supposing at the time \( t = t_0 \), the value of \( \psi_i \) is \( \psi_i(t_0) \), then the motion discipline can be found as follow
\[ \psi_i(t) = \psi_i(t_0) e^{-t(t-t_0)/T_i} \] (12)

Due to \( T_i > 0 \), if \( t \to \infty \), then \( \psi_i \to 0 \) and \( \psi_i \) weakens from any initial point by the discipline within the index \( (t-t_0)/T_i \), finally converges to the manifold. As well as the smaller the value of \( T_i \), the faster the system converges to the manifold. The convergence process of other components in macro-variables is similar. Thus, the SGC are directly applied to nonlinear systems without simplifying the system model, and the solution process of the control discipline is simple. Even when the system deviates from the equilibrium state, the control effect based on SGC is still better.

**B. DESIGNING FOR SGC**

On the basis of the tradition closed loop control, the control accuracy of the speed loop is affected by the control error produced from the inside current loop, the self-parameters uncertainty of the machine and the load torque disturbance. To improve the system stability, a new controller is designed and subjoined into the speed loop base on the SGC.

For convenient calculation, the difference between the reference speed \( \omega_{ref} \) and the measured speed \( \omega_e \) is defined as the speed error by
\[ \dot{\eta} = \omega_{ref} - \omega_e \] (13)

The control target is the rotor speed, which is mainly related to the feedback speed error and the rotor position. Therefore, both the speed and the rotor position are selected to construct the macro variable, which is denoted by
\[ \psi_1 = k_1 (\omega_{ref} - \omega_e) + k_2 \int (\omega_{ref} - \omega_e) dt \] (14)
where \( k_1 \) and \( k_2 \) are control parameters. Combining (8), (13) and (14), the follows can be obtained
\[ T (k_1 \dot{\eta} + k_2 \eta) + k_1 \eta + k_2 \int \eta dt = 0 \] (15)

Deforming (15)
\[ T k_1 \dot{\eta} + (T k_2 + k_1) \eta + k_2 \int \eta dt = 0 \] (16)

Further derivation on both sides of the (16)
\[ T k_1 \ddot{\eta} + (T k_2 + k_1) \dot{\eta} + k_2 \eta = 0 \] (17)

Eigenvalue equation of (16) is expressed by
\[ T k_1 \dot{\eta}^2 + (T k_2 + k_1) \dot{\eta} + k_2 \eta = 0 \] (18)

Eigenvalues of (17) can be obtained as follows
\[ P_1 = -\frac{1}{T} \quad P_2 = -\frac{k_2}{k_1} \] (19)

Solution of the second order differential equation
\[ \eta = K_1 e^{-t/T} + K_2 e^{-k_2 t/t_1} \] (20)
where \( K_1 \) and \( K_2 \) are all constants. It can be found from (20) that the speed error converges to zero regardless of \( K_1 \) and \( K_2 \). The control discipline is achieved as follows
\[ i_q = \frac{J}{3 n_p \psi_f} (\frac{1}{J} (B \omega + T_L) + \frac{1}{T} + \frac{k_2}{k_1}) \eta + \frac{k_2}{T k_1} \int \eta dt \] (21)

where the estimated parameter \( T \) is uncertainty, which is determined by
\[ T = \sqrt{T_{\text{min}}, T_{\text{max}}} \] (22)
where \( T_{\text{min}} \) and \( T_{\text{max}} \) are respectively the lower and upper limits of parameter \( T \).

**C. PARAMETERS OPTIMIZATION FOR SGC**

Since the parameter \( T \) is must smaller than the system dynamic corresponding time, as well as parameters \( k_1 \) and \( k_2 \) reflect the convergence rate when the system convergences to the equilibrium point, these parameters are crucial to the system control property. Thus, the genetic algorithm (GA) is applied to optimize parameters of the designed SGC. For the conventional damping controller, parameters are optimized repeatedly corresponding with a variety of conditions until the system has higher robustness. But for the SGC, only a typical operation condition is applied to optimize parameters since the SGC has an inherent robustness to keep the system converge to a stable manifold and even has a lobal stability after the system deviates from the equilibrium point. The objective function is defined as follows
\[ \min J = \int_{t_0}^{t_s} \left[ \alpha (\omega_{ref} - \omega_e)^2 + \beta (\theta_{ref} - \theta)^2 \right] dt \]
\[ k_{\text{min}} < k_1 < k_{\text{max}}, \quad k_{\text{min}} < k_2 < k_{\text{max}} \]
\[ T_{\text{min}} < T < T_{\text{max}} \] (23)

where \( t_s \) is the simulated time, \( k_{\text{min}} \) and \( k_{\text{max}} \) are respectively the lower and upper limits of parameters \( k_1 \) and \( k_2 \). Coefficients \( \alpha \) and \( \beta \) are used to adjust the weight between the speed deviation and rotor position deviation in the objective function. Additionally, the process for solving parameters based on the GA is as follows:

**Step1** Setting feasible domains of the control parameters and the maximum number of iterations in the optimization process.
where $l > 0$ and $C_i$ is a constant. Actually, the ESMO contains two adjustable parameters of $l$ and $k_i$, which are selected appropriately to adjust response speed and improve the identity accuracy. The control structure of the ESMO is given by Fig. 2.

The solution of the above equation is as follows:

$$e^2 = \dot{e}^2 + \omega^2 e^2$$

(27)

Based on sliding mode control and Lyapunov theory, the following (28) must be ensured to keep the error converge to zero.

$$\dot{e}^2 + \omega^2 e^2 \leq 0$$

(28)

Thus, the detection criterion is set as follows:

$$g = \begin{cases} 1, & \text{if } \omega_i(\infty) > \omega_i(\infty) \text{ or } \omega_i(\infty) \leq \omega_i(\infty) \\ 0, & \text{otherwise} \end{cases}$$

(32)

The diagram of the above equation is as follows:

(29)

where $e$ is the estimated error of the speed, which is regarded as sliding mode switching surface. $\dot{e}$ is the estimated error of the torque.

Since the dual PMSM driving is often affected by external noise or parameter uncertainty, the sensor has a certain error and speed. Generally, a small threshold is good for fast and sensitive diagnosis. But, if the threshold is too small, some fault can not be detected. For the proposed threshold, the probability of misdagnosis will be reduced and the diagnosis factor $g$ is defined as follows:

$$g(\omega) = \begin{cases} 1, & \text{if } \omega_i(\infty) > \omega_i(\infty) \text{ or } \omega_i(\infty) \leq \omega_i(\infty) \\ 0, & \text{otherwise} \end{cases}$$

(33)

The diagram of the above equation is as follows:

(30)

Therefore, the difference between $\omega_i$ and $\omega_i$ is selected as the fault diagnose index, as shown in (31).

$$y(t) = y(t) + \Delta y(t)$$

(31)

V. DIAGNOSIS AND FTC FOR SPEED SENSOR FAULT

A. MODELING FOR SENSOR FAULT

In the SGC system, an external disturbance also must be considered. In this part, ESMO is designed to not only achieve high performance but also eliminate the external disturbance. Thus, the sensor gain drop [15]. In fact, no matter which sensor fault category can be regarded as the additive perturbations.

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(31)
The main steps involved in the FTC are as follows:

i) In the limited time, the ESMO is regarded as a virtual sensor to estimate the machine speed.

ii) The estimated speed is sent to the sensor diagnosis module to execute diagnostic process.

iii) According diagnostic results, a switching mechanism chooses between the estimated and actual measured speed. If the sensor has no fault, the measured speed is chosen. Others, the estimated speed is chosen.

iv) The switching speed is sent to the SGC to track the speed objective.

The whole flow chart is shown in Fig.3. As the machine is running, the measured speed \( \omega_e \) is feedbacked and the estimated speed \( \tilde{\omega}_e \) is obtained by the ESMO. Then, the speed diagnosis index \( \omega_e(\text{in}) \) is calculated according to (31), and compared with the speed threshold \( \omega_e(\text{th}) \) to obtain the diagnostic factor \( g \) according to (32). If the diagnosis factor \( g \) is equal to 0, then the speed sensor is normal. The measured speed \( \omega_e \) continues to send to the SGC. If the diagnosis factor \( g \) is equal to 1, then the speed sensor is faulty. Simultaneously, the faulty speed sensor is isolated and is replaced by the ESMO, which is treated as a virtual sensor. Then the estimated speed \( \tilde{\omega}_e \) is sent to the SGC to implement the fault-tolerant control.

VI. SIMULATIONS AND EXPERIMENTS

A. SIMULATION ANALYSIS

The dual three-phase PMSM is configured with two isolated neutral points. The main parameters are depicted in Table 2.

![FIGURE 4. The structural schematic diagram.](image)

![FIGURE 5. Performances of the driving system when the dual-PMSM operates with a variable load.](image)

The structural schematic diagram of the SGC applied to the speed control loop for dual-PMSM is shown in Fig.4.

To verify the effectiveness and superiority of the proposed control method, the control system of dual-PMSM is set up in the Matlab/Simulink. In view of parameters optimization process for the controller in the section III, the optimal values can be obtained as \( T = 0.005 \text{s}, k_1 = 1, k_2 = 100 \).

1) THE DUAL-PMSM OPERATES NORMALLY

It is considered that the dual-PMSM operates with the load variable from 5N·m to 10N·m at 1.5s in the first test. Fig.5 presents output performances of the dual-PMSM driving system. Fig.5(a) displays the corresponding change in the output torque. Fig.5(b) shows the corresponding increase in the six-phase current, which of amplitudes smoothly rise from about 7A to 11A at 1.5s. Meanwhile, the measured speed drops from 1000r/min to 985r/min at 1.5s, then returns to

![FIGURE 3. The fault diagnosis and FTC flow chart.](image)
the reference speed after 15ms. In other words, the speed adjustment time is about three quarters of a period.

Moreover, output performances of the dual-PMSM driving system controlled by the PI controller, the SMC and the predictive controller are also introduced to compare with those by the SGC. The comparisons of the transient responses with the mutative load are presented in Fig.5(a) and (c). With the PI controller, the torque amplitude has a larger overshoot and is stable at the value of 10N·m through 0.11s, which is larger than five periods, yet the speed amplitude drops from the reference speed to nearly 970r/min. With the SMC, not only the torque has no the overshoot, but also performances of the torque response and the speed oscillation are all better than those with the PI controller. Obviously, the responding time of speed corresponding with the reference speed after 15ms. In other words, the speed adjustment time is about three quarters of a period.

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2) SPEED SENSOR FAILURE

In this simulation, Fig.7 displays the performance of the proposed method corresponding with an abrupt speed sensor fault. The speed sensor fault is applied at 1.42s and the reference speed is 1000r/min. Due to the sensor fault, the measured speed $\omega_e$ gradually starts to decrease at 1.42s until reaches zero after 6ms, which is described in Fig.7(a). As well as the waveform of the reference speed $\omega_{ref}$ and $\omega_e$ are also presented in this figure. Furthermore, Fig.7(b) depicts the speed threshold $\omega_{e(th)}$ and the fault speed index $\omega_{e(in)}$. The diagnosis factor $g$ is also depicted in Fig.7(c). According to (31) and (32), at 1.42s the amplitude of the $\omega_{e(in)}$ rises from 0 to 1000r/min, which results to the amplitude of the $\omega_{e(th)}$ and the diagnosis factor $g$ changes from 0 to 1. Thus, the speed sensor fault is promptly diagnosed and the faulty sensor is isolated. Meanwhile, the sensor fault also leads that amplitudes of six-phase currents all increase, which are shown in Fig.7(d).

As soon as the FTC scheme are carried out. The ESMO regarded as a virtual sensor is applied to calculate $\omega_e$, and the value of $\omega_e$ is sent to the SGC to generate control signals. Then, the $\omega_{e(in)}$ returns to zero and the value of the factor $g$ changes to 1, as well as amplitudes of six-phase currents all return to pre-fault state again owing to the FTC. In this figure, the instants for fault diagnosis and fault tolerance are respectively denoted by $t_D$ and $t_C$, which are total required 23ms.

B. EXPERIMENTAL VERIFICATION

To confirm the effectiveness of the proposed method, three representative that can sufficiently reveal performances of the dual-PMSM driving system controlled by the SGC are carried out. The experiments include the machine operates with the variable speed, the variable load torque and the sensor failure.
1) THE SPEED SENSOR IS HEATHLY

Fig. 8 displays performances of the dual-PMSM driving system corresponding with the variable load torque. At 1.05s, the load torque increases from 5N·m to 10N·m. The measured torque gradually and smoothly rises to 10N·m after 14ms. As well as the speed emerges weak oscillations which of amplitude drops from the reference value 1000r/min to 985r/min, then return to the reference value as soon as possible owing to the developed SGC. Meanwhile, due to the load torque arise, the six-phase currents also rise steadily until reaching steady state. The sign $t_A$ represents the response instant which is required almost 20ms.

Fig. 9 displays performances of the dual-PMSM driving system corresponding with the variable reference speed.

At 1.058s, the reference speed increases from 600r/min to 1000r/min. The measured speed smoothly rises till reaching to the reference speed, which results in the measured torque rises suddenly. But the measured torque not only returns to the value 5N·m again, but also has no additional torque ripple owing to the developed SGC. Furthermore, due to the variable speed, at 1.058s the six-phase currents begin to generate weak oscillations, yet restore to steady state as soon as possible, and the period of the six-phase current also changes from 33ms to 20ms. In this experiment, the response instant $t_A$ is also 20ms.
2) THE SPEED SENSOR IS FAULTY

Fig. 10 depicts the fault diagnosis and tolerant performance when the machine operates at the reference speed 1000r/min. To evaluate the performance of the fault diagnosis algorithm, a sudden speed sensor fault is applied at 1.08s, which results in the rotor position varies to zero and the measured speed starts to decrease from 1000r/min to zero after 12ms, as shown in Fig.10(a). At this time, according to (31) and (32), the corresponding diagnosis index $\omega_{e(in)}$ exceeds the speed threshold $\omega_{e(th)}$ and the diagnosis factor changes from 0 to 1. Thus, the speed sensor fault is diagnosed. To confirm the accuracy of the diagnosis results, the six-phase currents are displayed in Fig.10(b). When the speed sensor failure, the amplitudes of the six-phase currents all increase to nearly 9A. Then, the FTC scheme are immediately carried on. The estimated speed is supplied as the feedback to reach the continuous operation. The estimated rotor position and the estimated speed are also shown in Fig.10(a). As soon as the index $\omega_{e(in)}$ and factor g all return to zero, and the six-phase current restore the stable state despite appearing a little oscillation in the initial FTC stage. In this experiment, the signals $t_D$ and $t_C$ are total required 27ms.

3) COMPARISONS WITH OTHER CONTROLLERS

To further verify the superiority of the SGC, performances of the dual-PMSM driving system controlled by PI controller and SMC are introduced in this part. Since it is found in the simulations that dynamic performances produced by using the SGC and the predictive controller are similar, comparisons of performances corresponding with those controllers are no longer performed in experiments. Fig.11(a) displays the output speed and torque of the machine operates in the variable reference speed. At 1.048s, the reference speed changes from 600r/min to 1000r/min. With the PI controller, the measured speed increases slowly and is stable at 1000r/min, yet its waveform has a large amount of overshoot at 1.12s. Meanwhile, the torque suddenly increases to 16N·m, which is much bigger than that controlled by the SGC. The response instant $t_{A(PI)}$ is also 152ms. With the SMC, the measured speed has no overshoot, but it reaches 1000r/min after 32ms, as well as the torque also appears a larger increase. The response instant $t_{A(PR)}$ is 32ms. Fig.11(b) displays the output speed and torque of the machine operates in the variable load torque. It can be seen that the load torque varies from 5N·m to 10N·m at 1.048s. Regardless with the PI controller and the SMC, the additional torque ripple is brought in the output performance and an obvious oscillation occurs in the measured speed. As well as the response instants $t_{A(PI)}$ and $t_{A(PR)}$ are 152ms and 32ms, respectively. Therefore, the output performances controlled by the PI controller and the SMC are all weaker than those by the SGC, and the response instants are also much longer than that by the SGC.

VII. CONCLUSION

In this article, a new optimized parameters-based SGC is designed for dual-PMSM driving systems. The SGC integrating an ESMO is applied to diagnose and fault tolerance control the speed sensor fault. When the dual-PMSM driving system is healthy, the EMSO only supplies the estimated speed, which contrasts with the actual speed to create the speed error. Particularly, the SGC can immediately reduce
the steady-state error to generate precise control values. Once the speed sensor is faulty, the EMSO is treated as a virtual sensor to estimate the machine speed, as well as the SGC is contributed to lower errors on the EMSO inputs, which can further enhance the precise and robustness of estimated speeds. No additional hardware is demanded to reach the advanced fault diagnosis and FTC strategies. The simulation and experimental results are all presented to prove the effectiveness and robustness of the advance strategies. Besides, dynamic performances corresponding with the PI controller and the SMC are compared with those with the SGC, that are used to verify the superiority of the SGC.

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LIYI ZHANG received the M.S. and Ph.D. degrees from Tianjin University, Tianjin, China. He is currently a member of the Circuits and Systems Branch, Chinese Institute of Electronics, and the Director of the Tianjin Institute of Communications. He is also an Executive Director of the China Logistics Association. He has published more than 50 published/accepted journal articles and conference papers. He holds six patents and patent applications in this area. His research interests include signal detection and information processing, and intelligent computing and information processing.

FENG GAO was born in Tianjin, China. He received the B.S. and M.S. degrees in control theory and control engineering from the Hebei University of Technology, Tianjin, China, in 2009 and 2012, respectively. He is currently a Research Fellow with Shenhua Tianjin Coal Terminal Company Ltd., Tianjin. His research interests include power electronics and electric drives, intelligent control theory and application, and fault diagnosis.

JIALIN QIAN is currently pursuing the degree with the College of Information Engineering, Tianjin University of Commerce, Tianjin, China. His research interests include power electronics and electric drives.