Study on six-phase PMSM control strategy based on improved position sensorless detection technology

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Abstract
The six-phase motor control system has low torque ripple, low harmonic content, and high reliability; therefore, it is suitable for electric vehicles, aerospace and other applications requiring high power output and reliability. This study presents a superior sensorless control system for a six-phase permanent magnet synchronous motor (PMSM). The mathematical model of PMSM in stationary coordinate system is presented; its sliding mode observer (SMO) of back electromotive force (BEMF) acquires the motor speed and position information. As torque ripple and harmonic components affect the BEMF estimated value through the traditional SMO, the function of the frequency-variable tracker of the stator current (FVTSC) is used instead of the traditional switching function. By improving the SMO method, the BEMF is estimated independently, and its precision is maintained under startup or variable-speed states. To improve performance of the system, the auto disturbance rejection controller (ADRC) is used to suppress load disturbance and speed overshoot. To reduce model parameters, an ADRC with sliding mode algorithm is proposed, simplifying the structure and simultaneously solving the parameter tuning problem. Finally, comparing the ADRC with FVTSC SMO and the traditional PI controller shows that the new ADRC effectively improves the dynamic and static performance of the PMSM.

1 | INTRODUCTION

Because of the high output power [1], low torque ripple [2], low harmonic content [3] and high reliability of the multiphase PMSM [4, 5], it has broad application prospects in electric vehicles, ship propulsion, aerospace, and rail transit, all of which require high power and high reliability [6, 7]. The traditional position detection method needs position sensor, which increases the volume, cost and reduces the reliability; therefore, the sensorless position detection technology is a promising technical method [8, 9]. Generally, there are two types of sensorless position detection techniques. The first is a high-frequency injection (HFI) method, which is suitable for low speed and uses the salient effect of the rotor to estimate the position and speed [10, 11]. The estimate precision is irrelevant to the speed and insensitive to parameter variation; however, this method requires a certain saliency in PMSM. In addition, the amplitude of the HFI signal must be properly selected; otherwise, the electromagnetic noise will be generated. The other position estimation method depends on the BEMF and is suitable for medium and high speed. Examples include the extended Kalman filter method [12, 13], the model reference adaptive control [14, 15], and the sliding mode observer (SMO) method [16, 17].

Owing to the non-linearity, strong coupling, and time-varying parameters of the PMSM, operation performance is easily affected. Therefore, it is difficult to obtain ideal speed-adjusting performance with a traditional proportional-integral (PI) controller. The ADRC can meet the requirement for high-performance control strategies, not only controlling the state variables and derivatives of the system, but also dynamically suppressing the comprehensive disturbances of the system, thus improving its dynamic performance [18, 19]. The disadvantages are that the ADRC depends on non-linear theory, an efficient analysis method is yet to be developed, and the controller parameters need to be selected according to experience.

Gao [20] proposed a linearization method for non-linear ADRC and a method for determining controller
gain by adjusting the bandwidth. However, it significantly increased the complexity of the algorithm; meanwhile, this study only conducted simulation research without experimental verification, and the practicality still needs to be verified.

Dong and Li [21] suggested a fuzzy adaptive ADRC to improve current waveforms and suppress torque ripple for a brushless DC motor. However, the disadvantages of this method are a slower response and unsuitability for frequent speed regulation occasions.

Zuo et al. [22] proposed an ADRC with position feedback to solve contradictions between tracking and anti-interference performance. The study optimized the dynamic performance of the extended state observer (ESO) by adjusting the scale coefficient and bandwidth, thereby simplifying the parameter setting of the controller. However, because of the existence of a differential element, it is still difficult to achieve the decoupling effect and reduce the system stability, specifically when position information fluctuates substantially.

Deng and Guan [23] applied ADRC to the vector control system of the PMSM, designing the position, speed, and current loop, and then establishing the ESO to compensate for the system disturbance. Furthermore, the parameters of the ADRC were linearized, which reduced the number of adjustable parameters. However, owing to the phenomenon of integral saturation, tracking performance and system stability of this controller will decrease when the motor is in an abnormal working condition.

Considering the difficulty of adjusting speed with a PI controller, parameter identification was introduced into ADRC in the literature [24]. This method can identify the stator resistance, stator inductance and torque inertia with model reference adaptive system and Landau iterative algorithms, which suppress external disturbance adequately. The disadvantage of this method is that it is easily affected by the parameter variation of the motor itself, and the robustness is poor.

In this study, the dual Y shift 30° six-phase PMSM with an isolated neutral point was investigated, a new sliding mode sensorless detection technology was adopted, and an ADRC method with sliding model control was proposed. First, the frequency-variable tracker of the stator current (FVTSC) was added to the traditional SMO to solve the problem of BEMF dynamic estimation. Especially when the current frequency of stator windings changes, the new method can improve the estimation precision of BEMF and the anti-chattering ability of SMO. Second, the speed controller with sliding mode ADRC was adopted to limit overshoot during start-up process, simultaneously, improve tracking performance and anti-interference performance of the control system. Finally, in order to verify the effectiveness of the method, a system simulation model and experimental platform were built, and experimental research was carried out. The results showed that the control system with ADRC and SMO FVTSC had fast dynamic response and strong anti-interference ability.

2 | MATHEMATICAL MODEL OF THE SIX-PHASE PMSM

The inverter topology is shown in Figures 1 and 2 shows the distribution structure.

The six-phase PMSM is a six-dimensional system with six independent variables, and its mathematical model in the natural coordinate system has the disadvantages of high-order, non-linearity, and strong coupling. Flux is not only related to the phase current but also to the rotor position angle \( \theta \), thereby making it difficult to realize effective control [25, 26].

\[
C_{6s/2s} = \frac{1}{\sqrt{3}} \begin{bmatrix}
1 & \sqrt{3}/2 & -1/2 & -\sqrt{3}/2 & -1/2 & 0 \\
0 & 1/2 & \sqrt{3}/2 & 1/2 & -\sqrt{3}/2 & -1 \\
1 & -\sqrt{3}/2 & -1/2 & \sqrt{3}/2 & -1/2 & 0 \\
0 & 1/2 & -\sqrt{3}/2 & 1/2 & \sqrt{3}/2 & -1 \\
1 & 0 & 1 & 0 & 1 & 0 \\
0 & 1 & 0 & 1 & 0 & 1
\end{bmatrix}
\]
Through the transformation matrix in Equation (1), the electrical elements in the double three-phase motor are mapped to the fundamental plane ($\alpha\beta$ space) and harmonic plane ($z_1-z_2$ space). In the matrix, the first and second vectors correspond to the $\alpha\beta$ plane, the third and fourth vectors correspond to the $z_1-z_2$ plane, and the fifth and sixth vectors correspond to the zero order plane ($o_1-o_2$ space). The conversion of electromechanical energy is only connected to the $\alpha\beta$ subspace rather than $z_1-z_2$ and $o_1-o_2$ subspaces; thus, these two spaces are called harmonic subspaces.

As the neutral point of the dual Y shift 30° six-phase PMSM is not connected, the $o_1-o_2$ space has a zero sequence variable, which is also called the zero sequence space. The rotation coordinate transformation of the $\alpha\beta$ space is performed while the $z_1-z_2$ and $o_1-o_2$ subspaces are multiplied by the unit matrix and remain unchanged.

Thus, the transformation matrix is expressed as:

$$C_{2s/2r} = \begin{bmatrix} \cos \theta & \sin \theta & 0 & 0 & 0 & 0 \\ -\sin \theta & \cos \theta & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 1 \end{bmatrix}$$

(2)

where $\theta$ denotes rotor electric degree. Using the transformation matrix, we can determine the flux, voltage, electromagnetic torque, and motion equations as follows.

Flux equation:

$$\begin{bmatrix} \psi_d \\ \psi_q \end{bmatrix} = \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} \psi_f \\ 1 \end{bmatrix}. $$

(3)

Voltage equation:

$$\begin{bmatrix} \varepsilon_d \\ \varepsilon_q \end{bmatrix} = \begin{bmatrix} R & 0 \\ 0 & R \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \psi_d \\ \psi_q \end{bmatrix} + \omega_e \begin{bmatrix} -\psi_q \\ \psi_d \end{bmatrix}. $$

(4)

Electromagnetic torque equation:

$$T_e = 3p[\psi_f i_q + (L_d - L_q)i_d i_q]. $$

(5)

Motion equation:

$$T_e - T_L - B\omega = J\frac{d\omega}{dt}, $$

(6)

where $L_d$ and $L_q$ are the inductance; $i_d$ and $i_q$ are the currents; $\psi_d$ and $\psi_q$ are stator flux; and $u_d$ and $u_q$ are stator voltage of the $d$ and $q$ axes, respectively. Further, $T_e$, $T_L$, $\psi_f$, $B$, $\omega$, $J$, $R$, and $p$ represent electromagnetic torque, load torque, rotor permanent magnet flux, damping coefficient, mechanical angular velocity, electric angular velocity, moment of inertia of the motor, stator resistance, and pole pairs, respectively.

3 | SENSORLESS CONTROL OF PMSM

3.1 | Traditional sliding mode observer

In the six-phase PMSM system, the mathematical model in the $\alpha\beta$ coordinate system is:

$$\begin{bmatrix} \varepsilon_d \\ \varepsilon_q \\ \varepsilon_f \end{bmatrix} = \begin{bmatrix} R & 0 & 1 \\ 0 & R & 1 \end{bmatrix} \begin{bmatrix} i_d \\ i_q \\ \frac{d}{dt} \begin{bmatrix} i_d \\ i_q \end{bmatrix} \end{bmatrix} + \begin{bmatrix} \varepsilon_d \\ \varepsilon_q \\ \varepsilon_f \end{bmatrix}, $$

(7)

where $\varepsilon_d$, $\varepsilon_q$ are components of BEMF and can be expressed as:

$$\begin{bmatrix} \varepsilon_d \\ \varepsilon_q \end{bmatrix} = \omega \psi_m \begin{bmatrix} -\sin \theta \\ \cos \theta \end{bmatrix}. $$

(8)

The SMO can make the state variable trajectory move along the ideal sliding mode surface by changing the system structure dynamically. Taking the stator current components as the state variable, from Equation (7), we obtain:

$$\begin{align*}
\frac{d\hat{i}_d}{dt} &= -\frac{R}{L_s} i_d + \frac{1}{L_s} \frac{u_d}{L_s} - \frac{1}{L_s} \varepsilon_d \\
\frac{d\hat{i}_q}{dt} &= -\frac{R}{L_s} i_q + \frac{1}{L_s} \frac{u_q}{L_s} - \frac{1}{L_s} \varepsilon_q,
\end{align*}$$

(9)

According to the theory of sliding mode variable structure, take $\varepsilon_d \sim \hat{i}_d - i_d$, $\varepsilon_q \sim \hat{i}_q - i_q$ as the sliding mode surface; the SMO of winding currents are:

$$\begin{align*}
\frac{d\tilde{i}_d}{dt} &= -\frac{R}{L_s} \tilde{i}_d + \frac{1}{L_s} \frac{u_d}{L_s} - \frac{K_s}{L_s} \text{sign} (\hat{i}_d - i_d) \\
\frac{d\tilde{i}_q}{dt} &= -\frac{R}{L_s} \tilde{i}_q + \frac{1}{L_s} \frac{u_q}{L_s} - \frac{K_s}{L_s} \text{sign} (\hat{i}_q - i_q),
\end{align*}$$

(10)

where $\hat{i}_d$ and $\hat{i}_q$ are the estimated currents. $K_s$ is the sliding gain, and its selection must satisfy the requirements of existence, stability and reachability of the SMO.

From Equations (9) and (10), the current deviation equation can be expressed as:

$$\begin{align*}
\frac{d\tilde{i}_d}{dt} &= -\frac{R}{L_s} \tilde{i}_d + \frac{1}{L_s} \frac{u_d}{L_s} - \frac{K_s}{L_s} \text{sign} (\tilde{i}_d) \\
\frac{d\tilde{i}_q}{dt} &= -\frac{R}{L_s} \tilde{i}_q + \frac{1}{L_s} \frac{u_q}{L_s} - \frac{K_s}{L_s} \text{sign} (\tilde{i}_q)
\end{align*}$$

(11)

where $\tilde{i}_d = \hat{i}_d - i_d$, $\tilde{i}_q = \hat{i}_q - i_q$, represent the current estimation error of the observer.
3.2 Frequency-variable tracker of the stator current (FVTSC)

An FVTSC method is introduced to estimate BEMF to address the torque pulsation problem caused by the SMO. FVTSC is an improvement for the traditional SMO. By optimizing the symbol function in traditional SMO, the chattering problem of traditional SMO is solved. Compared with traditional SMO, the BEMF estimation of FVTSC is more accurate by tracking the stator current frequency in real time, simultaneously, the problems of phase lag and amplitude attenuation in traditional SMO are solved. The main principle is to use the F function instead of the switch sign function. To solve this problem, this study introduces FVTSC to replace the original switch sign function.

The disadvantage of the traditional proportional resonance control (PR) is that it can only track the specific frequency and is not suitable for the variable frequency, but FVTSC can adapt to the frequency change by the modified PR, as shown in Figure 3.

\[
\Delta i = -\hat{\theta_c} \cos \hat{\theta_c} - \hat{\theta_p} \sin \hat{\theta_p},
\]
\[
\omega_c = 2\pi f,
\]
\[
f = \frac{(K_p + K_c) \cdot \Delta i}{2\pi},
\]

Here, \( f \) is the operating frequency, \( K_p \) is the proportional gain, and \( \omega_c \) is the angular frequency.

Let \( u = [i_x, i_y] \) and \( y = [e_x, e_y] \), then the FVTSC function can be expressed as:
\[
\frac{Y(s)}{U(s)} = \frac{K_f [s^2 + 2f s + (2\pi f)^2] + K_r f}{s^2 + 2f s + (2\pi f)^2},
\]
where \( K_f \) and \( \omega_c \) denote the resonance gain and cutoff frequency, respectively.

Equation (16) shows that the main parameters \( K_f, K_r, \omega_c \) that affect the performance of the controller. \( K_f, K_r \) are the proportional gain and the peak gain at resonance frequency, respectively, and \( \omega_c \) affects the resonance gain and bandwidth. The principle of parameter adjustment is to adjust \( K_r \) to eliminate steady-state error and \( \omega_c \) to suppress frequency fluctuation. In the experiment, each parameter needs to be tuned online.

In order to facilitate calculation and combine the parameters of motor and observer, the frequency range of \( \omega_c \) is \( 2000–4000 \), \( K_f \) and \( K_r \) parameters can be adjusted according to the real-time response and frequency characteristics of the observer, and the appropriate parameters can be determined according to the expected experimental results.

To adapt to the digital implementation, the transfer function in the time domain must be converted to the Z domain; the expected experimental results.

The state space equation can be obtained by deriving Equation (19) as:
\[
\begin{align*}
\dot{x}_1 &= x_2, \\
\dot{x}_2 &= -(2\omega_c x_2 + \omega_c^2 x_1) + u.
\end{align*}
\]

Equation (20) provides the FVTSC function as follows:
\[
\begin{align*}
x &= \begin{pmatrix} 0 & 1 \\ -\omega_c & 0 \end{pmatrix} \cdot x + \begin{pmatrix} 0 \\ 1 \end{pmatrix} \cdot u \\
y &= \begin{pmatrix} 0 & K_f \end{pmatrix} \cdot x + \begin{pmatrix} 0 \\ K_r \end{pmatrix} \cdot u.
\end{align*}
\]
Based on the above analysis, the structure diagram of the FVTSC is demonstrated in Figure 4.

When the moving point converges on the sliding mode surface, the BEMF estimated by the SMO is equivalent to:

\[
\begin{align*}
\hat{e}_\alpha &= K_s F (\tilde{i}_\alpha) \\
\hat{e}_\beta &= K_s F (\tilde{i}_\beta)
\end{align*}
\]

(22)

From Equations (22) and (8), the estimated rotation angle can be obtained:

\[
\hat{\theta}_e = -\arctan \frac{\hat{e}_\alpha}{\hat{e}_\beta}.
\]

(23)

Figure 5 presents the structure diagram of SMO position detection based on FVTSC. The estimated stator currents \(\hat{i}_\alpha, \hat{i}_\beta\) are obtained by the current observer and compared with the measured values \(\tilde{i}_\alpha, \tilde{i}_\beta\) of the stator current, the deviation \(\tilde{i}_\alpha, \tilde{i}_\beta\) is obtained, and the stator current frequency \(f\) obtained from the PLL is input into FVTSC. This method can accurately estimate the BEMF \(\hat{e}_e\) and the rotor position angle \(\hat{\theta}_e\) can be obtained with Equation (23).

**4 | DESIGN OF ADRC FOR SIX-PHASE PMSM SYSTEM**

**4.1 | Design of traditional ADRC**

In this study, a six-phase PMSM system utilizes \(\dot{\xi}_l = 0\) of the vector control strategy, and ADRC is adopted for the speed loop. The state equation of the speed loop can be derived from the mathematical model of the PMSM, as shown in Equation (24).

\[
\dot{\omega} = 3 p \psi_f \frac{B}{J} \omega - \frac{T_L}{J}.
\]

(24)

The comprehensive disturbance term and compensation factor of the speed loop are defined as:

\[
\begin{align*}
a(\omega, T_L) &= -\frac{B}{J} \omega - \frac{T_L}{J} \\
b &= \frac{3 p \psi_f}{J}
\end{align*}
\]

(25)

The modified state equation of the speed loop of the PMSM control system is as follows:

\[
\dot{\omega} = a(\omega, T_L) + bi_q.
\]

(26)

It can be seen from Equation (26) that the speed loop of the PMSM is a first-order system. Since the design of ADRC does not depend on the model, the corresponding ADRC can be designed according to the order of the model. Combining the design principles of ADRC and state equation of the speed loop, the standard ADRC of the speed loop consists of three parts, as shown in Figure 6.

The first-order tracking derivative controller (TD) is modelled as follows:

\[
\begin{align*}
\dot{x}_1 &= x_2 \\
x_2 &= -r f a l (x_1 - v_\alpha, \delta_0)
\end{align*}
\]

(27)
The second-order (ESO) is given by:

\[
\begin{align*}
\dot{\varepsilon} &= x_1 - y \\
\dot{z}_1 &= x_2 - \beta_{01} \, \text{fal} \left( \varepsilon, \alpha, \delta_1 \right) \\
\dot{z}_2 &= x_3 - \beta_{02} \, \text{fal} \left( \varepsilon, \alpha, \delta_2 \right) + bu \\
\dot{z}_3 &= -\beta_{03} \, \text{fal} \left( \varepsilon, \alpha, \delta_3 \right)
\end{align*}
\]  

(28)

The non-linear state error feedback control rate (NLSEF) can be expressed as:

\[
\begin{align*}
u_0 &= \beta_1 \, \text{fal} \left( x_1 - z_1, \alpha, \delta \right) + \beta_2 \, \text{fal} \left( x_2 - z_2, \alpha, \delta \right) \\
u &= u_0 - \frac{z_3}{b}
\end{align*}
\]  

(29)

where \( r \) is the tracking gain of the differentiator; \( v \) is the given speed of the PMSM; \( x_1 \) and \( x_2 \) are the intermediate state variables of TD output and their difference, respectively; \( \beta_{01}, \beta_{02}, \beta_{03} \) and \( \beta_{03} \) are the order gains of the extended state controller, and \( \varepsilon \) is the error between the estimated value and the actual value. The selection of parameters directly affects the observation accuracy and dynamic and static characteristics of the control system; \( y \) is the actual output value of the control object; \( z_1 \) is the estimated value of the output variable; \( z_2 \) is the differential signal of \( z_1 \); \( z_3 \) is the estimated value of the total disturbance of the system; \( b \) is the disturbance compensation factor; \( u \) is the control output of the system; \( \alpha, \delta_1, \delta_2, \delta_3 \) are non-linear factors; \( \delta_1, \delta_2, \delta_3 \) are the filtering factors of the non-linear functions; \( \beta_1 \) and \( \beta_2 \) are the gains of NLSEF. The function \( \text{fal} () \) is a non-linear function, and its specific form is as follows:

\[
\begin{align*}
\text{fal} \left( \varepsilon, \alpha, \delta \right) &= \begin{cases} 
\varepsilon/\delta^{1-\alpha}, & |\varepsilon| < \delta \\
|\varepsilon|^\alpha \, \text{sign}(\varepsilon), & |\varepsilon| > \delta
\end{cases}
\end{align*}
\]  

(30)

where \( \alpha \) and \( \delta \) denote the non-linear factor and filtering factor of the non-linear function, respectively.

As a non-linear control function is used in ADRC, the adjustable parameters include \( r, \beta_{01}, \beta_{02}, \beta_{03}, \beta_1, \beta_2, \alpha, \delta, \) and \( b \), which increase the difficulty in determining parameters. Therefore, this paper introduces a sliding mode control algorithm to replace the original non-linear function, thereby simplifying the parameter tuning process of the controller.

4.2 Design of sliding mode ADRC

From Equation (29), \( x_1 \) is the state variable of TD, and \( z_1 \) is the estimated value of the output variable. Both \( x_1 \) and \( z_1 \) reflect the actual speed reference and feedback; thus, the error \( e_1 = x_1 - z_1 \) is set as the state error of the speed loop, and \( u_0 = e_1 \) is regarded as the control output.

\[
\begin{align*}
e_1 &= x_1 - z_1 \\
e_1' &= e_1 = x_1 - z_1 \\
u &= u_0 - \frac{z_3}{b}
\end{align*}
\]  

(31)

For the state error feedback control system, the sliding mode surface is:

\[
s = e_1.
\]  

(32)

From Equation (32), we can get:

\[
\dot{s} = e_1 = u_0.
\]  

(33)

In order to shorten the time to reach the switching surface, the appropriate approaching velocity should be chosen. The exponential approaching law is adopted in this paper, that is \( \dot{s} = -\varepsilon \, \text{sign}(s) - k \, s \), where \( \varepsilon > 0 \) and \( k > 0 \) are both adjustable, thus, the output of sliding mode control is:

\[
u_0 = -\varepsilon \, \text{sign} \left( x_1 - z_1 \right) - k \left( x_1 - z_1 \right) .
\]  

(34)

From Equations (32) and (33), the following equation can be obtained:

\[
s \cdot \dot{s} = s \left( -\varepsilon \, \text{sign}(s) - k \, s \right) = -\varepsilon \, |s| - k \, s^2.
\]  

(35)

Because \( \varepsilon \) and \( k \) are greater than zero, the condition \( s \cdot \dot{s} < 0 \) is tenable; that is, it satisfies the existence condition of the sliding mode. Under this condition, system states can reach the sliding mode surface and the zero point of the sliding mode within a finite time. Based on the above analysis, the non-linear error feedback control law with sliding mode control proved to be stable.

In summary, the mathematical model of the non-linear state error feedback control law based on sliding mode control is:

\[
\begin{align*}
u_0 &= -\varepsilon \, \text{sign} \left( x_1 - z_1 \right) - k \left( x_1 - z_1 \right) \\
u &= u_0 - \frac{z_3}{b}
\end{align*}
\]  

(36)

Equation (36) only contains three adjustable parameters: \( \varepsilon, k \) and \( b \). Compared with the former standard ADRC solution, the controller parameters are reduced and the parameter tuning problem is solved.

5 SIMULATION RESULTS

To prove the effectiveness and feasibility of the new high-efficiency control technology and sensorless position detection technology proposed in this study, a simulation platform of the dual Y shift 30° six-phase PMSM was built, as shown in Figure 7, and simulation studies were conducted. The parameters of the six-phase PMSM were set as follows: \( R = 0.102 \, \Omega, L_d = 0.82 \, mH, L_q = 0.82 \, mH, \psi = 0.072 \, Wb, J = 0.001 \, kg \cdot m^2, P = 3 \).

5.1 Simulation at startup process

The motor started with a 10 N-m load torque, and the given rotating speed was set to 500 r/min. Figures 8 and 9 showed
As demonstrated in Figures 8 and 9, the speed through the ADRC with sliding mode control was fairly consistent with the actual speed during the startup process; that was, the observer could accurately track the rotor position angle. From Figure 10, the speed overshoot with the novel ADRC in this study was almost negligible and reached a steady state at 0.03 s. In comparison, the starting waveform under the traditional PI controller had an evident overshoot, reached a steady state after 0.08 s, and the response time was slightly longer.

In summary, the simulation results showed that SMO could accurately track the rotor position angle. The sensorless control method of six-phase PMSM with sliding mode ARDC had the advantages of fast response speed and good steady-state characteristics.

5.2 Simulation of sudden load changes

In order to verify the feasibility of the new control strategy, the following experimental study was carried out. The load torque was increased to 15 N·m suddenly at 0.4 s and decreased to 10 N·m at 0.8 s at 500 r/min. Figures 11 and 12 showed speed and rotor position angle comparison waveforms, respectively. The speed comparison waveforms between PI controller and ADRC with SMO under sudden load change were shown in Figure 13. Figure 11(a) showed that, when the load changed suddenly, the estimated speed with ADRC and SMO was consistent with the actual speed, the observer could accurately observe the actual speed when the load changes suddenly. Figure 11(b,c) showed the enlarged waveforms of the load increase and decrease, respectively. The enlarged waveforms shown in Figure 12(b,c) revealed that the rotor position was slightly changed due to the speed fluctuation at the moment of 0.4 s sudden loading and 0.8 s unloading, the estimated angle accurately tracked the actual rotor position. Figure 13 demonstrated that the speed drop with PI was approximately 18 r/min at 0.4 s on sudden loading, and the speed increase was approximately 28 r/min at 0.8 s on sudden unloading; thereafter, the speed restored to the target speed after 0.48 s. With the sliding mode ADRC mentioned in this paper, the speed drop was only 6 r/min at 0.4 s on sudden loading, and the speed increase was 12 r/min on
unloading with velocity being restored to the target speed after 0.02 s. The proposed system demonstrated superior dynamic properties.

5.3 Robustness analysis

According to the working principle of the observer, if the prediction model is not accurate, the robustness of the system will also deteriorate accordingly. In the traditional SMO, the variation of resistance, inductance and flux will lead to the error of the predicted current, which will affect the control performance of the traditional SMO. If the model observer can be improved, the influence of the parameters on the predicted values of the system can be avoided, and the robustness of the system can be improved accordingly. Through tracking the stator current frequency, FVTSC can detect and compensate the predicted current in real time. When the actual model parameters change, it can make the predicted current track the actual current to increase the accuracy of the prediction.
Figure 13 Speed waveforms (a) Speed comparison of ADRC and PI controllers, (b) enlargement of speed at load addition, and (c) enlargement of speed at load removal.

Figure 14 was a comparison between the estimated current and the actual current when the winding resistance is rated at $0.102 \, \Omega$.

Figures 15–18 showed comparison between the estimated and the actual current waveform when the rated resistance changes by 65%, 75%, 125% and 135%, respectively. From Figures 14–18, the current estimation values based on the observer was not affected when the motor parameters changed, so the robustness of the observer was verified.

6 EXPERIMENTAL RESULTS

To prove the accuracy and reliability of the six-phase PMSM control strategy in this study, a dual Y shift $30^\circ$ six-phase PMSM with an isolated neutral point was taken as the research object.
The six-phase motor parameters were shown in Table 1. And the experimental platform can be seen in Figure 19.

The motor started with a 3 N·m load torque, and the given speed was set to 500 r/min. As demonstrated in Figure 20, the starting speed with PI had an evident overshoot, and it took longer to reach the steady state. However, the motor with ADRC started smoothly, and the speed rose without significant overshoot. When the PMSM was running at 500 r/min and 3 N·m load torque, a load torque of 3 N·m was added suddenly, and then reduced suddenly. The speed response with the different controllers were shown in Figures 21 and 22. When the

### TABLE 1  Six-phase PMSM parameters

| Parameters         | Value          |
|--------------------|----------------|
| Rated power        | 1.5 kW         |
| Rated speed        | 2000 r/min     |
| Rated voltage AC242V | 100 Hz        |
| Rated current      | 4 A            |
| Stator resistance  | 0.102 Ω        |
| $L_d, L_q$         | 0.82 mH        |
| $\psi$             | 0.072 Wb       |
motor was running steadily under 1000 r/min and 3 N·m load, adding load to 6 N·m and then reducing load to 3 N·m, the system response were shown in Figure 23.

It was clearly seen that, in the case of sudden increasing load torque, the motor phase current increased from 2 to 4 A, the speed ripple under ADRC was smaller, which was reduced by 40 r/min and PMSM can quickly restore the steady state after 0.2 s. In comparison, the torque ripple under the traditional PI controller was large, which was reduced by 120 r/min and the recovery time was about 0.3 s. In the case of sudden reducing load torque, the motor phase current decreased from 4 to 2 A, the speed ripple under ADRC was still small, increasing by 55 r/min and the recovery time was about 0.2 s, however the torque ripple under the traditional PI controller was relatively large, increasing 120 r/min and the recovery time was about 0.3 s. Figure 23 illustrated that the motor had a small current pulsation and a torque fluctuation at a high speed of 1000 r/min.

Figure 24 showed the ADRC current waveforms as the speed from 500 r/min to 1000 r/min with 3 N·m load. When the speed changed abruptly, there was no overshoot in the speed rise, the current ripple was small.

Figure 25 showed rotor position at startup process, and the reference speed was 500 r/min. From Figure 25, we can see that the estimated angle can quickly and accurately track the rotor position, which demonstrated the feasibility of the SMO in this study.

7 | CONCLUSION

In this paper, the mathematical model of the six-phase PMSM was established in the static coordinate system, and the FVTSC technology was proposed to improve the traditional SMO, which solved the problems of chattering, BEMF estimation and
phases. At the same time, ARDC were used to replace the traditional speed PI controller, which had the advantages of suppressing load disturbance, high reliability and no overshoot.

Finally, simulation and experimental researches were carried out, and results showed the new detection and control strategies proposed in this paper could accurately estimate the rotor position, ensuring that the motor had good dynamic performance, and the system had strong robustness when the motor parameters changed.

However, this paper only studied the speed-loop ARDC, and the current loop still adopted the traditional PI. Therefore, current-loop ARDC will be one of the hot spots in the future. In addition, considering the multidimensional characteristics of six-phase motor system, the future research directions also include sensorless detection and control in the case of failure.

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