Sliding Mode Predictive Current Control of Permanent Magnet Synchronous Motor with Cascaded Variable Rate Sliding Mode Speed controller

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This work was supported by the National Natural Science Foundation of China (52177047), Fund project of Liaoning Provincial Department of Education (LJ2019JL011) and Fund project of Liaoning Provincial Education Department Scientific Research (LJ2020QNL013).

ABSTRACT To solve the problem that the conventional vector control system of permanent magnet synchronous motor is susceptible to motor parameters and load disturbances, a sliding mode predictive current control strategy of permanent magnet synchronous motor with cascaded variable rate sliding mode speed controller based on extended state observer (ESO) is proposed, which improves the dynamic and static performance of the system. First, a sliding mode speed controller is designed based on the novel variable rate exponential reaching law, and the ESO is used to estimate and compensate for parameter mismatch and disturbance, which improves the system’s rapidity and robustness while reducing the chattering. Then, to solve the problem that the traditional deadbeat predictive control of current loop ignores the influence of cross-coupling electromotive force, based on the above reaching law, a sliding mode controller is designed to equate the back electromotive force to effectively compensate the deadbeat predictive current controller, which not only retains the rapidity of current but also enhances the robustness. Finally, simulation and experimental results show the effectiveness of the proposed control strategy.

INDEX TERMS Permanent magnet synchronous motor, deadbeat predictive current control, sliding mode control, extended state observer

I. INTRODUCTION

Permanent magnet synchronous motor (PMSM) has the advantages of simple structure, high torque-to-current ratio, and high efficiency, and has been widely used in industry, agriculture, aerospace and other fields [1], [2]. However, the PMSM is a multivariable, nonlinear and strongly coupled controlled object, the control system is vulnerable to uncertainties such as motor parameter change, external load disturbances, object unmodeled, and nonlinear dynamics [3], [4]. The traditional PI control method is often used in the speed controller design of PMSM due to its simple and stable characteristics. However, due to its low control accuracy and sensitivity to the mathematical model of the system, it is easily affected by external disturbances and parameter mismatch, so it is difficult to meet the high-performance control requirements of the system [5].

In order to realize the high-performance control of PMSM, many advanced control strategies such as deadbeat predictive control [6], model-free control [7], adaptive control [8], sliding mode control [9], [10] have been applied to PMSM control system. Among them, sliding mode control (SMC) does not require high model accuracy and has strong robustness, so it is widely used in the drive system of permanent magnet synchronous motors [11], [12]. Reference [13] proposes a method combining adaptive sliding mode control with model predictive torque control. Using the robustness of sliding mode control, it effectively solves the problem of relying on motor parameters and being sensitive to load torque changes in traditional model predictive torque control. But this makes the control scheme relatively complex and increases the computational burden. Reference [14] combines sliding mode control with active disturbance rejection control, and introduces simulated annealing particle
swarm optimization (SAPSO) to optimize the parameters of the sliding mode active disturbance rejection speed controller offline. Experiments show that this method can improve the dynamic performance and steady-state speed tracking accuracy of PMSM. However, it is difficult to choose the controller parameters of this control scheme.

However, the above advantages of sliding mode control are often at the cost of setting a larger switching gain, which leads to the inevitable chattering phenomenon [15], [16]. To solve this problem, many methods have been proposed, such as method of reaching law [17], [18], nonsingular terminal sliding mode [19] and fractional order sliding mode [20]. In [21], the generalized proportional integral observer (GPIO) is used to accurately estimate to concentrated disturbance, and combined with super-twisting sliding mode control (STSM) to form a composite controller. Experiments show that this scheme can achieve better tracking accuracy with smaller switching gain. However, the discontinuity in the integral process is still difficult to completely eliminate the chattering phenomenon. Reference [22] proposes a new type of super-twisting-like fractional order (STLF) controller, which optimizes the problem of control performance decline caused by integral discontinuity in STSM algorithm. Nevertheless, there are still some aspects that need further research to improve its practicability, such as the dynamic response of the system. As we all know, since the method of reaching law can directly act on the approach process, it can solve the chattering problem more effectively. Reference [23] designed an adaptive terminal sliding mode reaching law based on continuous fast terminal sliding mode control, which reduces the control input, dynamically achieves finite-time convergence, high-precision tracking, and reduces chattering to a certain extent. However, its design is too complex and the starting current is large. Reference [24] introduces the terminal arrival part on the basis of the traditional exponential reaching law. The simulation and experimental verification show that this method can effectively suppress chattering and shorten the reaching time. Reference [25] also used the hyperbolic tangent function to replace the sign function in the traditional sliding mode control, which also weakened the chattering to a certain extent, but the effect is general.

Another reliable strategy to reduce chattering is to use disturbance observer technology. Its advantage is to improve the robustness by observing and compensating the integrated disturbance without sacrificing the performance of the closed-loop system. References [26], [27] all improved the anti-disturbance performance of the sliding mode controller by introducing the disturbance observer, but the design of the observer depends on the accurate model of the drive system. As the core part of active disturbance rejection control (ADRC), extended state observer (ESO) can effectively estimate the total disturbance of the system without an accurate system model [28], [29]. Therefore, it is widely used in various application scenarios, such as linear induction motors [30], test rocket control systems, etc. [31].

Deadbeat predictive current control originates from discrete linear state feedback control. It has the advantages of constant switching frequency, high bandwidth and easy realization. It is a lightweight algorithm that can make the feedback current track the given current in a short time. However, as a control strategy commonly used in the current loop, deadbeat predictive current control often ignores the back electromotive force generated by the mutual coupling of dq axis current. With the increase of the speed, the influence of the coupling effect gradually increases, and the proportion of the back EMF also increases gradually. And during the dynamic operation of the system, the change of parameters will make the estimation of back EMF inaccurate, resulting in a decrease in the current change rate and the increase of current tracking error [32]. In order to solve this problem and improve the dynamic performance of the system, reference [33] proposed a new discrete-time robust predictive current controller for permanent magnet synchronous motor. Based on deadbeat predictive current control, the discrete-time integral term was introduced to weaken the influence of parameter uncertainty on the back electromotive force term. However, the current error under this control scheme converges slowly. Reference [34] proposed a composite control method that combines deadbeat predictive current control and feedforward compensation, which solved the problem of low control accuracy caused by parameter mismatch to a certain extent. References [35] and [36] used the back EMF as a disturbance to estimation and compensation, which improved the tracking performance and robustness of the system. However, the latter used parameter online identification to estimate unknowns in real time, which was complex in calculation, and the compensation accuracy was easily affected by the parameter accuracy.

To sum up, this paper designs a sliding mode predictive current control strategy of PMSM with cascaded variable rate sliding mode speed controller. Based on the traditional exponential reaching law, the variable speed term with the sliding surface as the independent variable is introduced, which effectively improves the dynamic quality and chattering phenomenon in the reaching motion stage, a novel sliding mode speed controller based on ESO is constructed. The controller improves the robustness and rapidity of the system. Aiming at the problem that the traditional deadbeat predictive current control is not sensitive to the change of the back electromotive force, the discrete sliding mode controller is designed based on the above reaching law to compensate it equivalently. The main contributions of this paper can be summarized into four aspects.

1) The novel variable rate exponential reaching law is designed by introducing variable rate term with sliding surface as independent variable. The reaching speed is adaptively adjusted by the system state position, which speeds up the approach process and reduces the chattering.
phenomenon at the same time. It effectively solves the problems of poor parameter adaptability, contradiction between robustness and chattering degree of the traditional exponential reaching law.

2) The novel sliding mode controller is designed by using the variable rate exponential reaching law, and ESO is introduced to estimate and compensate the integrated disturbance term, which reduces the influence of parameter mismatch, load disturbance and other uncertain factors. This effectively improves the robustness of the system.

3) The novel discrete sliding mode deadbeat current controller is designed to compensate the back EMF term equivalently. It solves the problem that the traditional deadbeat predictive current control is easily affected by the change of the back EMF, resulting in excessive current tracking error. The control accuracy of the current loop is further improved.

4) Simulation and experimental results show that the proposed method is effective and superior to the existing theories.

II. MATHEMATICAL MODEL OF PMSM

In this paper, the surface permanent magnet synchronous motor is used as the controlled object (\( L_d = L_q = L \)), and the \( i_q = 0 \) control strategy is selected. In order to simplify the analysis, assuming that the conductivity of the rotor permanent magnet material is zero and there is no damping winding, eddy current and hysteresis losses are not considered, and the magnetic circuit is ignored, then the mathematical model of PMSM in the \( dq \) coordinate system is

\[
\begin{align*}
J \frac{d\omega_m}{dt} &= \frac{3}{2} p_n \psi_f i_q - B\omega_m - T_L \\
\frac{du_d}{dt} &= R_s i_d + L \frac{di_d}{dt} - p_n \omega_m L i_q \\
\frac{du_q}{dt} &= R_s i_q + L \frac{di_q}{dt} + p_n \omega_m L i_d + p_n \omega_n \psi_f
\end{align*}
\]  

(1)

where \( u_d, u_q, i_d \) and \( i_q \) are the \( dq \) axis components of the stator voltage and current, respectively. \( L \) is the stator inductance, \( \omega_m \) is the rotor mechanical angular velocity, \( R_s \) is the stator resistance, \( \psi_f \) is the permanent magnet flux linkage, \( p_n \) is the number of pole pairs, \( T_L \) is the load torque, \( B \) is the viscous coefficient and \( J \) is the moment of inertia.

III. DESIGN OF A NOVEL SLIDING MODE SPEED CONTROLLER BASED ON ESO

A. NOVEL VARIABLE RATE EXPONENTIAL REACHING LAW

The traditional exponential reaching law is as follows:

\[
\dot{s} = -\varepsilon \text{sgn}(s) - q_s, \varepsilon, q > 0
\]

(2)

Its essence is to combine the exponential reaching term with the constant reaching term. When the system is far away from the sliding surface and is in the approach motion stage, the exponential reaching term is mainly working. When \( s \) gradually decreases, the approach speed becomes \( \varepsilon \), which can ensure that it can be reach the sliding mode surface in a limited time. However, the existence of the constant reaching term means that chattering must also exist, which not only affects the control accuracy of the speed control system, but also stimulates the unmodeled high-frequency components in the system.

Therefore, this paper proposes a novel variable rate exponential reaching law:

\[
\begin{align*}
\dot{s} &= -\varepsilon |s|(1 - e^{-\mu|s|})\text{sat}(s) - (q + k|s|)s \\
\text{sat}(s) &= \begin{cases} 
-1, s < -\Delta \\
\frac{s}{\Delta}, |s| \leq \Delta \\
1, s > \Delta
\end{cases}
\end{align*}
\]

(3)

where \( \varepsilon, \mu, q \) and \( k \) are all constants greater than zero, and \( \Delta \) is the boundary layer thickness.

As shown in (3), the sliding mode surface is used as an independent variable to introduce the traditional exponential reaching law to realize the real-time update of the approach speed. When the system state variable is far away from the sliding mode surface, \(|s|\) is larger, \((1 - e^{-\mu|s|}) \to 1\), and the system approaches the sliding mode surface at variable rates according to the constant reaching term \(-\varepsilon |s|\text{sat}(s)\) and the exponential reaching term \(-(q + k|s|)s\), so that the approach speed is improved. When the system state variable is close to the sliding mode surface, \(|s|\) is small, \((1 - e^{-\mu|s|}) \to 0\). On the one hand, the constant reaching term approaches 0, reducing the chattering phenomenon caused by this. On the other hand, \((q + k|s|)\) gradually decreases, so that the system state variables reach the sliding surface as quickly and smoothly as possible. Moreover, replacing the original sign function with a saturation function will further reduce chattering.

In order to verify the performance of the novel reaching law mentioned above, it is analyzed and studied with a typical system.

\[
\frac{dx}{dt} = Ax + Bu
\]

(4)

Take the sliding surface as \( s = hx \), and the derivation of the sliding surface is

\[
\frac{ds}{dt} = h \frac{dx}{dt}
\]

(5)
Bring (3) and (4) into (5), and the controller output can be obtained as
\[
\begin{align*}
\mathbf{u} = (hB)^{T} & \left[ -c|s|(1-e^{-|s|}) \right] \text{sat}(s) - \\
& (q + k|s|)s - hA\mathbf{x}
\end{align*}
\] 
(6)
where \( \mathbf{x} \) is the system state variable, and \( \mathbf{x} = \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} \),
\[
A = \begin{bmatrix} 0 & 1 \\ 0 & -25 \end{bmatrix}, \quad B = \begin{bmatrix} 0 \\ 200 \end{bmatrix}, \quad h = [1.5 \quad 1].
\]
Let \( \varepsilon = 5 \), \( q = 60 \), \( \mu = 0.01 \), \( k = 400 \) and select the initial state of the system as \( \mathbf{x}(0) = \begin{bmatrix} 0.5 \\ 0.5 \end{bmatrix} \).

Under the same parameters, the two reaching laws are compared by simulation. The simulation results are shown in Figure 1.

From the simulation results in Fig. 1, it can be seen that the output accuracy of the controller with the novel variable rate exponential reaching law is more accurate and has almost no chattering, while the output response of the traditional exponential reaching law is relatively slow and has obvious chattering phenomenon. The sliding mode surface convergence time of the former is only 0.02s, which is faster than 0.08s of the latter. Therefore, the simulation results show that the proposed novel variable rate exponential reaching law can effectively improve the dynamic quality of the approaching motion, reduce the time required for approaching and weaken the chattering compared with the traditional exponential reaching law. In this way, the sliding surface can be reached quickly and smoothly, and the stability of the system can be improved.

B. DESIGN OF NOVEL SLIDING MODE SPEED CONTROLLER

The design of the speed controller should ensure that the actual speed strictly follows the given reference \( \omega_{mr} \) under the condition of parameter mismatch and load disturbance. In this paper, a novel sliding mode speed controller is designed on the basis of the reaching law mentioned above. The specific design steps are as follows.

For the mechanical motion equation in (1), considering the influence of load torque and parameter mismatch, it can be written as
\[
\begin{align*}
J & \frac{d^3\omega_m}{dt^3} = \frac{3}{2} p_n \psi_f i_q - B \omega_m - T_L - \Delta B \omega_m + \\
& \frac{3}{2} p_n \Delta \psi_f i_q - \Delta J \frac{d \omega_m}{dt} - \Delta T + \sigma
\end{align*}
\] 
(7)
where \( \Delta \psi_f \), \( \Delta B \) and \( \Delta J \) are the variations of permanent magnet flux linkage, viscous coefficient and moment of inertia, respectively, \( \Delta T \) is the variation of load torque and \( \sigma \) is other unknown disturbances.

Let \( F = \frac{1}{J}(-B \omega_m - T_L + \frac{3}{2} p_n \Delta \psi_f i_q - \Delta B \omega_m - \Delta J \frac{d \omega_m}{dt} - \Delta T + \sigma) \) be the integrated disturbance term, then we can get
\[
\frac{d \omega_m}{dt} = \frac{3 p_n \psi_f}{2 J} i_q + F
\] 
(8)
Take the state variable of PMSM as
\[
\begin{align*}
\dot{x}_1 &= \omega_m - \omega_m \\
\dot{x}_2 &= \int \omega_m
\end{align*}
\] 
(9)
where \( \omega_m \) is the given value of the motor speed and \( \omega_m \) is the actual value of the mechanical angular velocity of the rotor.
The sliding surface is designed as
\[ s = x_1 + cx_2 \]  
(10)
where \( c \) is the integral sliding mode gain, which satisfies \( c > 0 \).

Taking the derivative of the sliding mode surface function and substituting (3) into
\[ \dot{i}_f = \frac{-2J}{3p_s\omega_f} \left[ e[1 - e^{-\alpha t}](s) + \left(q + k|s|\right)s - F + cx_i \right] \]  
(11)
\[ \beta \] is the integral sliding mode gain, which satisfies \( \beta > 0 \).

In order to prove the stability of the designed controller, the Lyapunov function is selected as \( V = \frac{1}{2}s^2 \), and its derivative can be obtained
\[ \dot{V} = s\dot{s} = s\left(\dot{x}_1 + cx_2\right) \]
\[ = \left(3p_s\omega_f\right)\left[ \frac{2J}{3p_s\omega_f} \left[ -F + cx_i \right] \right] \]
\[ = s\left[ -e[1 - e^{-\alpha t}](s) + \left(q + k|s|\right)s - F + cx_i \right] \]
\[ = s\left[ -e[1 - e^{-\alpha t}](s) + \left(q + k|s|\right)s \right] \]
\[ = -e[1 - e^{-\alpha t}](s) + \left(q + k|s|\right)s \]
(12)
The designed controller parameters satisfy \( e, \mu, q, k > 0 \). According to the Lyapunov stability theorem, at this time \( \dot{V} = s\dot{s} \leq 0 \), the system satisfies the sliding mode reaching condition, which indicates that the sliding mode can still converge to equilibrium state from any time when the system is under external load disturbance and parameter mismatch.

C. DESIGN OF EXTENDED STATE OBSERVER

ESO essentially belongs to state reconstruction. Compared with ordinary state observers, there is no need to know the specific mathematical model of the controlled object. It only needs to expand the internal and external disturbances of the system into a new system state. By adjusting the observer parameters, the total disturbance can be detected and fed back to the controller in real time. In this way, dynamic estimation and compensation can be realized, and the influence of disturbance on the control effect can be reduced.

In this paper, a second-order ESO is designed to estimate the total disturbance F, and (8) is rewritten as
\[ \begin{align*}
\dot{x} &= au + F \\
y &= x
\end{align*} \]  
(13)
where \( x = \omega_m \), \( a = \frac{3p_s\omega_f}{2J} \), and \( u = i_q \).

Thus, ESO is constructed as
\[ \begin{align*}
\dot{z}_1 &= au - \beta_1 f_a \left( \gamma, \alpha, \delta \right) + z_2 \\
\dot{z}_2 &= -\beta_2 f_a \left( \gamma, \alpha, \delta \right)
\end{align*} \]  
(14)
where \( z_1 \) and \( \gamma \) are the estimated and actual mechanical angular velocities, respectively, \( z_2 \) is the estimated matching disturbance, \( \beta_1 \) and \( \beta_2 \) are the observer parameters, the \( f_a \) function is an error nonlinear function, and the specific expression is as follows
\[ f_a \left( \gamma, \alpha, \delta \right) = \begin{cases} 
\frac{\gamma}{\delta^\alpha} & |\gamma| > \delta \\
0 & |\gamma| \leq \delta
\end{cases} \]  
(15)
where \( \delta \) is the filter factor and \( \alpha \) is the nonlinear factor. When \( 0 < \alpha < 1 \) is satisfied, the \( f_a \) function has the characteristics of large gain with small error and small gain with large error. By choosing appropriate observer parameters, the total disturbance \( F \) can be effectively estimated.

IV. DESIGN OF DISCRETE SLIDING MODE DEADBEAT CUTTENT CONTROLLER

The traditional deadbeat predictive current control ignores the cross-coupling electromotive force. When the given current of one axis changes, the current of the other axis will produce instantaneous error, which will affect the current control accuracy. Therefore, a discrete sliding mode deadbeat current controller \( u_{\text{DF}} = u_1 + u_2 \) is designed. Among them, \( u_1 \) is the deadbeat current controller of PMSM. The predictive control has the advantages of good dynamic performance and can realize the fast tracking control of the current. \( u_2 \) is the novel discrete integral sliding mode controller, which can effectively realize the equivalent compensation of the back electromotive force and improve the robustness of the current loop.

A. DESIGN OF DEADBEAT CUTTENT CONTROLLER

Let
\[ u_1 = \begin{bmatrix} u_{d1} \\ u_{q1} \end{bmatrix} = \begin{bmatrix} R_i j_d + L \frac{di_d}{dt} \\ R_i j_q + L \frac{di_q}{dt} \end{bmatrix} \]  
(16)
From (16), the PMSM current state equation is
\[ \begin{align*}
\frac{di_d}{dt} &= -\frac{R}{L} i_d + \frac{u_{d1}}{L} \\
\frac{di_q}{dt} &= -\frac{R}{L} i_q + \frac{u_{q1}}{L}
\end{align*} \]  
(17)
The first-order Euler method is used to discretize (17), and the discrete state space equation is obtained as
\[ i(k+1) = Mi(k) + Nu_i(k) \]  
(18)
where \( i(k) = \begin{bmatrix} i_d(k) \\ i_q(k) \end{bmatrix} \), \( u_i(k) = \begin{bmatrix} u_{d1}(k) \\ u_{q1}(k) \end{bmatrix} \),

\[
M = \begin{bmatrix} 1 - \frac{R_s}{L} T_s & 0 \\ 0 & 1 - \frac{R_q}{L} T_s \end{bmatrix}, \quad N = \begin{bmatrix} T_s & 0 \\ 0 & T_s \end{bmatrix}
\]

and \( T_s \) is the sampling period.

The control idea of deadbeat predictive current control is similar to PI control. By taking the given signal \( i_{dgr} = \begin{bmatrix} i_{dgr} \\ i_{qgr} \end{bmatrix} \) as the input \( i(k+1) = \begin{bmatrix} i_d(k+1) \\ i_q(k+1) \end{bmatrix} \) at the next moment, the current can keep up with the given current after one cycle, so as to realize the fast tracking of the current loop. When the sampling period \( T_s \) is small enough, the system can be treated as Euler.

\[
\text{Push (18) forward one sampling time step to obtain}
\]

\[
u_i(k) = N^{-1} \left[ i_{dgr}(k) - Mi(k) \right]
\]

**B. DESIGN OF THE NOVEL DISCRETE SLIDING MODE CONTROLLER**

Due to the strong robustness, this paper adopts the sliding mode control strategy to solve the problem that deadbeat predictive current control is easily affected by back EMF. Compared with the traditional exponential reaching law, the proposed reaching law not only speeds up the reaching process, but also weakens the chattering. Therefore, the discrete sliding mode controller is designed by using the above novel variable rate exponential reaching law to effectively compensate the back EMF.

Let \( u_2 \) be the back EMF term, that is

\[
u_2 = \begin{bmatrix} u_{d2} \\ u_{q2} \end{bmatrix} = \begin{bmatrix} -p_s \omega_m L_q \\ p_s \omega_m L_d + p_n \omega_m \gamma_f \end{bmatrix}
\]

(20)

The PMSM state variable is selected as

\[
e = \begin{bmatrix} c_1 \\ c_2 \end{bmatrix} = \begin{bmatrix} i_{qgr} - i_q \\ i_{dgr} - i_d \end{bmatrix}
\]

(21)

In this paper, the control strategy of \( i_d = 0 \) is selected, so it is considered that \( i_{dgr} = 0 \) and it can be obtained from (21)

\[
\dot{e} = \begin{bmatrix} i_{qgr} - i_q \\ -i_d \end{bmatrix} = \begin{bmatrix} i_{qgr} + \frac{R_s}{L} i_q + \frac{u_{d2}}{L} - \frac{u_q}{L} \\ \frac{R_q}{L} i_d + \frac{u_{q2}}{L} - \frac{u_d}{L} \end{bmatrix}
\]

(22)

In order to simplify the calculation, \( i_{qgr} \) is treated by Euler method, namely \( i_{qgr} = \frac{i_{qgr}(k+1) - i_{qgr}(k)}{T_s} \), where \( i_{qgr}(k+1) \) is the given value of the current in the next cycle. When the sampling period \( T_s \) is small enough, the linear extrapolation method can be used to obtain

\[
i_{qgr}(k+1) = 2i_{qgr}(k) - i_{qgr}(k-1)
\]

(23)

Thus, (22) is discretized as

\[
e(k+1) - e(k) = \begin{bmatrix} i_{qgr}(k) - i_{qgr}(k-1) \\ 0 \end{bmatrix}
\]

\[
e(k+1) - e(k) = \frac{1}{T_s} \begin{bmatrix} R_e i_q(k) + u_{q2}(k) - u_q(k) \\ \frac{R_q}{L} i_d(k) + u_{q2}(k) - u_q(k) \end{bmatrix}
\]

(24)

The discrete integral sliding mode surface is selected as

\[
s(k) = e(k) + c \tau(k)
\]

(25)

where \( c = \begin{bmatrix} c_1 \\ c_2 \end{bmatrix} \) is the discrete integral term coefficient whose terms are all non-negative and \( s(k) = \begin{bmatrix} s_1(k) \\ s_2(k) \end{bmatrix} \).

\[
\tau(k) = \tau_1(k) + \tau_2(k)
\]

(26)

By discretizing the new variable rate exponential reaching law (3) proposed above, we can get

\[
s(k+1) = \begin{bmatrix} 1 & T_s(q + k |s(k)|) \end{bmatrix} s(k) - T_s e s(k) \begin{bmatrix} 1 & -e^{-\mu(k)} \end{bmatrix} \text{sat}(s(k))
\]

(27)

Push (25) forward one sampling time step to obtain

\[
s(k+1) = e(k+1) + c \tau(k+1)
\]

(28)

Combining (24), (27) and (28), we can get

\[
u_2(k) = \begin{bmatrix} u_{d2}(k) \\ u_{q2}(k) \end{bmatrix} = u(k) - R_s i(k) + \begin{bmatrix} \tau_1(k) \\ \tau_2(k) \end{bmatrix}
\]

(29)

where

\[
\tau_i(k) = \begin{bmatrix} T_s e s_i(k) \begin{bmatrix} 1 & -e^{-\mu(k)} \end{bmatrix} \text{sat}(s_i(k)) + T_s (q + k |s_i(k)|) s_i(k) + c_i e_i(k) \end{bmatrix} \end{bmatrix} \begin{bmatrix} -L \\ T_s (1 + c_i) \end{bmatrix}, i = 1, 2
\]
Assuming that the sampling time \( T_s \) is extremely short, it can be known from the discrete sliding mode approach condition that it needs to meet

\[
\begin{align*}
\left[ s(k+1) - s(k) \right] \text{sgn} \{ s(k) \} & < 0 \\
\left[ s(k+1) + s(k) \right] \text{sgn} \{ s(k) \} & > 0
\end{align*}
\]

(30)

In summary, the discrete sliding mode deadbeat current controller can be obtained as

\[
u_{\text{ref}} = u_1 + u_2
\]

(31)

V. SIMULATION AND EXPERIMENTAL RESEARCH

The principle block diagram of PMSM speed control system designed in this paper is shown in Figure 2.

![Principle Block Diagram of PMSM Speed Control System](image)

In order to verify the performance of the above speed control system, the simulation research was carried out under the simulation environment of Matlab/Simulink. The motor parameters in the hardware-in-the-loop simulation are consistent with those in the simulation. The parameter values of the permanent magnet synchronous motor used in the experiment are shown in Table I.

### TABLE I

| Parameters                        | Value             |
|-----------------------------------|-------------------|
| Stator resistance                 | 0.346 \( \Omega \) |
| Stator inductance                 | 7.8mH             |
| Viscous coefficient               | 0.005N.m.s        |
| Moment of inertia                 | 0.089kg.m.s       |
| Permanent magnet flux linkage     | 0.51825V.s        |
| Number of pole pairs              | 2                 |
| Rated power                       | 10kw              |
| Rated voltage                     | 260V              |

In Simulink simulation and hardware-in-the-loop simulation, three schemes are used for comparative analysis. Scheme 1 adopts traditional PI control scheme. Scheme 2 is the sliding mode predictive current control strategy of PMSM with cascaded variable rate sliding mode speed controller without considering the disturbance (no ESO). Scheme 3 is the method proposed in this paper. Table II shows the controller parameters of the three schemes, in which the parameters of the novel variable rate exponential reaching law are obtained by combing the trial and error method with the simulation comparison.

### TABLE II

| Control plan | Speed controller | Current controller |
|--------------|------------------|--------------------|
| Scheme 1     | \( k_p = 0.5 \), \( k_i = 8 \) | \( k_p = 30 \), \( k_i = 500 \) |
|              | \( \epsilon = 5 \), \( \mu = 0.01 \) | \( \epsilon = 10 \), \( \mu = 0.001 \) |
|              | \( q = 0.2 \) | \( q = 0.1 \) |
| Scheme 2     | \( k = 0.1 \), \( c = 125 \) | \( k = 0.1 \), \( c = 150 \) |
|              | \( \epsilon = 10 \), \( \mu = 0.01 \) | \( c_1 = 50 \), \( c_2 = 50 \) |
|              | \( q = 0.1 \) | |
| Scheme 3     | \( \beta_2 = 8500 \), \( \alpha = 0.5 \) | Same as scheme 2 |
|              | \( \delta = 0.01 \) | |

A. SYSTEM SIMULATION RESEARCH

In order to verify the superiority of the method proposed in this paper, the simulation and experimental verification are carried out under three working conditions of loading and unloading, speed up and down, forward and reverse rotation.

1) LOADING AND UNLOADING SIMULATION

At the beginning of the simulation, the given speed is 500r/min, and the load was suddenly increased by 10N.m at 1s, and the load was suddenly reduced to 0 at 2.5s. Figure 3 shows speed and current simulation results of three control schemes.

![Simulation Results](image)
2) SPEED UP AND DOWN SIMULATION

The given speed is 300r/min at the beginning of the simulation. Adjust it to 500r/min at 1s, and then adjust it to 300r/min again at 2.5s. Speed curves of the three control schemes are shown in Figure 4.

According to Figure 3(a), when the given speed is 500 r/min, scheme 1 produces about 6.34% overshoot in the start-up stage. And scheme 2 and scheme 3 start smoothly without obvious overshoot. When t=1s, the increase of the load makes scheme 1 produce about 2.2% speed change, and it takes 0.2s to adjust to the original speed. Scheme 2 produces about 1.8% speed change, and the recovery time is long. The speed does not rise to the original 500r/min within 1.5s. Scheme 3 produces only about 1.1% of the speed drop, and the recovery time is only 0.1s. Combined with Figure 3(b), it can be seen more intuitively that scheme 3 can track the change of current faster at the moment of startup and load change, and the current fluctuation is relatively small. It can be seen from Figure 3(c), (d) and (e) that the current change in scheme 3 is smoother. The above simulation results show that scheme 3 has better dynamic performance, anti-disturbance ability and robustness.

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B. HARDWARE-IN-THE-LOOP SIMULATION RESEARCH

In order to further verify the effectiveness of the proposed scheme, the hardware-in-the-loop simulation verification is carried out using the comprehensive experimental platform of motor speed regulation and loading. The actual platform is shown in Figure 6. The system mainly includes NI virtual controller, PWM inverter, motor speed regulation and loading mechanical platform (including PMSM, magnetic powder brake and torque and speed measurer), monitoring host PC and distribution lines, etc., which can realize the
experiment of adding and reducing load and speed regulation of permanent magnet synchronous motor.

FIGURE 6. Comprehensive experimental platform of motor speed control and loading.

During the experiment, install VeriStand on the PXI controller, then call the FPGA board through VeriStand and run the corresponding program. The modified control algorithm program in the Simulink environment is downloaded to the NI controller through the NI MAX software, and the experimental comparison and research steps are consistent with the simulation.

1) LOADING AND UNLOADING EXPERIMENT

The speed curves of the motor loading and unloading experiment are shown in Figure 7. The motor starts with no load at a speed of 300 r/min, the load is suddenly increased to 10N·m at about 30s, and the motor load was reduced to 0N·m at about 60s.

It can be seen from Figure 7 that when the given speed is 500r/min and no-load start-up, scheme 1 produces about 3.6% speed overshoot. Scheme 2 and scheme 3 reach the given value quickly and smoothly in the start-up stage without obvious overshoot. When the same load is applied, scheme 1 produces about 3% speed change, and scheme 2 produces about 0.5% speed change, but the speed recovery is slow, and it still does not return to the given value after running for 30s. Scheme 3 only produces a speed drop of about 0.5r/min, and the corresponding recovery time is also shorter. It can also be seen from Figure 7 that the speed curve chattering degree of scheme 3 is significantly smaller than scheme 1 and scheme 2, reflecting the excellent dynamic and static performance of scheme 3.
It can be seen from Figure 8 that the current fluctuation of scheme 3 is small, and the current response is fast and stable during loading and unloading, which can realize no static error control. Compared with scheme 2, it shows that the designed extended state observer can effectively estimate the influence of external disturbance and parameter mismatch in real time, and can reduce the chattering amplitude to a certain extent. The conclusions of the analysis and research are basically consistent with the simulation results.

2) SPEED UP AND DOWN EXPERIMENT

Figure 9 shows the speed curves of the motor speed up and down experiment. The initial speed is 300 r/min, when it runs for about 30s, it is adjusted to 500 r/min, and it returns to 300 r/min after 60s.

It can be seen from Figure 9 that scheme 1 produces an overshoot of about 15.1 r/min at the startup, and an overshoot of about 8.75 r/min occurs when the given speed rises to 500 r/min. However, the speed of Scheme 2 and Scheme 3 changes smoothly, and there is basically no overshoot or drop, and Scheme 3 has a shorter adjustment time. The experimental analysis results are basically consistent with the simulation results.

3) FORWARD AND REVERSE EXPERIMENT

Figure 10 is the speed curve of the motor forward and reverse experiment. At the beginning of the experiment, the speed was set to 300 r/min, and then adjusted to -300 r/min after 45s.
dynamic response caused by excessive switching gain in

be realized, which avoids the problem of unsatisfactory

Combined with ESO, the fast and stable control of speed can

reaching law is used to design the speed controller.

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VI. CONCLUSION

It can be seen from Figure 10 that when the motor changes from forward to reverse, scheme 1 produces a speed drop of about 14.3 r/min, scheme 2 and scheme 3 achieve a smooth transition without obvious speed drop, and scheme 3 has a small degree of speed chattering. While the scheme 2 produces about 1.25r/min steady-state static error when the system is stable. The conclusions of the hardware-in-the-loop simulation experiment are basically consistent with the simulation results.

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