

Electromagnetic Performance Optimization of PMSM Based on Improved PSO and Super-Twisting Active Disturbance Rejection Control

Dehai Chen^{1,2} and Ruilong Liu^{1,*}

¹*School of Electrical Engineering and Automation, Jiangxi University of Science and Technology, Ganzhou 341000, Jiangxi, China*

²*Jiangxi Zhongke Electric Drive Co., Ltd., Ganzhou 341000, Jiangxi, China*

ABSTRACT: To improve the electromagnetic operating stability and dynamic response of a permanent magnet synchronous motor (PMSM) speed control system under load disturbances and speed variations, an improved particle swarm optimization-based super-twisting active disturbance rejection control (IPSO-ISTA-ADRC) method is proposed. Focusing on the speed control loop, high-order power terms are introduced into the super-twisting control law within the ADRC framework to accelerate error convergence. A dynamic boundary-layer function is employed to smooth the switching terms and reduce chattering. In addition, a sigmoid function is incorporated into the extended state observer to improve disturbance estimation and compensation, while an improved particle swarm optimization algorithm is used to optimize the controller parameters. The proposed method is further applied to a sensorless PMSM speed control system to evaluate its control performance under sensorless operating conditions. Simulation results show that, compared with PI, LADRC, and boundary-layer-based STA-ADRC, the proposed method effectively suppresses speed overshoot and fluctuations and shortens the recovery time under no-load startup, load disturbance, and speed variation conditions, thereby achieving improved disturbance rejection, faster dynamic response, and enhanced robustness.

1. INTRODUCTION

Permanent magnet synchronous motors (PMSMs) are widely used in new-energy vehicles, industrial robots, and high-performance servo systems because of their high power density, high efficiency, and fast dynamic response. Electromagnetic torque is generated through the interaction between the rotor permanent-magnet field and the rotating stator field. Load disturbances and abrupt speed changes can cause fluctuations in the stator current and electromagnetic torque. Conventional mechanical position sensors, however, increase both system cost and maintenance requirements. Sensorless control has therefore become an important research topic for PMSM drive systems.

Research on sensorless PMSM operation has primarily focused on high-frequency signal injection, back-electromotive-force estimation, and state-observer design. Methods that account for cross-coupling inductances or employ Luenberger observers and reduced-order disturbance observers have improved rotor position and speed estimation over different operating regions. Sensorless control methods for different speed ranges have also been systematically reviewed and compared [1–4]. Although these studies have laid the foundation for sensorless PMSM operation, system performance under load disturbances and abrupt speed changes also depends strongly on the dynamic response and disturbance-rejection capability of the speed controller.

Active disturbance rejection control (ADRC) treats parameter variations, unmodeled dynamics, and external load variations as a lumped disturbance, which is estimated and compensated for in real time by an extended state observer (ESO). This approach reduces the controller's dependence on an accurate mathematical model. In [5], ADRC was applied to a PMSM current loop to improve current tracking under parameter variations and load disturbances. To address the trade-off between command tracking and disturbance rejection in conventional linear active disturbance rejection control (LADRC), a two-degree-of-freedom structure was introduced in [6] to decouple the tracking and disturbance rejection channels, thereby providing greater flexibility in controller tuning. A two-stage ESO was subsequently developed in [7] to improve the estimation of aperiodic and periodic disturbances. Although these methods enhance the disturbance estimation and compensation capabilities of ADRC, LADRC relies mainly on linear error feedback. Its dynamic response and nonlinear regulation capability under rapid speed changes and severe load disturbances therefore can be further improved.

The super-twisting algorithm (STA) provides a continuous control output while retaining the robustness of sliding-mode control and has consequently been increasingly adopted in PMSM speed control. In [8], continuous adaptive fast terminal sliding-mode control was combined with an improved super-twisting observer to enhance speed response and disturbance estimation. Studies [9, 10] incorporated improved super-twisting algorithms into LADRC and introduced improvements

* Corresponding author: Ruilong Liu (1046973744@qq.com).

to the ESO, state-error feedback law, and smoothing function, thereby enhancing the dynamic performance of the PMSM speed loop and reducing chattering. In [11], the conventional sign function was replaced by a continuous quasi-sliding-mode function, and a proportional-integral term was incorporated into the super-twisting control law to improve speed recovery following load disturbances. These studies demonstrate that combining the STA with ADRC can improve the controller's response speed and robustness. Nevertheless, the multiple gain parameters in the control law are mutually coupled, making the resulting performance highly dependent on appropriate parameter coordination.

Closed-loop performance is also affected by the ESO's disturbance estimation accuracy and output smoothness. In [12], the nonlinear error function and sliding-mode ESO were improved to enhance disturbance estimation and current control. Refs. [13, 14] employed cascaded nonlinear ESOs or incorporated sliding-mode control into ADRC to improve the compensation of lumped disturbances and observation residuals. In [15], the controller switched between LADRC and nonlinear ADRC, according to the magnitude of the speed error, combining a rapid response in the large-error region with high control accuracy in the small-error region. Although these studies have improved PMSM control through observer design, error-feedback functions, and combinations of control laws, the large number of parameters involved still makes coordinated tuning of the controller and observer difficult.

To reduce reliance on manual tuning, intelligent optimization algorithms have been increasingly used to optimize motor-control parameters. In [16], a hybrid grey wolf optimization algorithm was used to tune the parameters of an improved super-twisting LADRC, improving speed tracking and disturbance rejection performance of a PMSM speed regulation system. In [17], an improved particle swarm optimization algorithm was applied to the LADRC parameter-tuning problem of a permanent-magnet linear synchronous motor, resulting in improved position tracking. These studies demonstrate the effectiveness of intelligent optimization algorithms for multi-parameter controller tuning and provide a basis for the parameter optimization considered in this work.

In summary, previous studies have improved PMSM speed regulation through super-twisting control, ESO enhancement, and intelligent parameter tuning. However, most existing improvements focus on the control law, disturbance observer, or parameter optimization separately, leaving scope for a more coordinated design of these components. Accordingly, rather than simply retuning an existing STA-ADRC, this study jointly considers the speed-loop control law, disturbance estimation, and parameter configuration.

To address these issues, this paper proposes an improved PSO-based, power-augmented super-twisting active disturbance rejection control method for a sensorless PMSM speed regulation system. The main contributions are summarized as follows:

(1) The conventional super-twisting control law is augmented with higher-power nonlinear terms. The standard super-twisting terms provide the required disturbance rejection

capability, while the higher-power terms strengthen the control action when the speed error is large, thereby improving the dynamic response to abrupt speed changes and load disturbances. A dynamic boundary-layer function is also employed to smooth switching terms and reduce chattering in control output.

(2) A sigmoid function is incorporated into the ESO's nonlinear error-feedback channel to improve the smoothness of the disturbance estimate while preserving its error-correction capability. The resulting estimate of the lumped disturbance is then fed forward to the speed controller for compensation.

(3) The introduction of higher-order power terms increases the number of controller parameters, strengthens parameter coupling, and makes manual tuning more difficult. An improved particle swarm optimization algorithm is therefore employed to jointly optimize the key speed-controller parameters. A composite objective function is constructed using speed-tracking error, steady-state error, overshoot, load-disturbance error, and q -axis current fluctuation so that the optimized parameters account for dynamic response, disturbance rejection, and current smoothness.

(4) The proposed speed controller is integrated with an LESO-IPLL sensorless estimation structure to form a sensorless PMSM speed regulation system. Under identical operating conditions, the proposed method is compared with PI, LADRC, and STA-ADRC in terms of speed response, recovery from load disturbances, and electromagnetic torque, thereby verifying its effectiveness.

2. MATHEMATICAL MODEL OF THE PMSM

A surface-mounted permanent magnet synchronous motor (SPMSM) is considered in this study. Under ideal conditions, the effects of magnetic saturation, iron losses, temperature variations, and other factors are neglected. The PMSM's mathematical model is established in the synchronously rotating d - q reference frame. The stator-voltage equations are given by

$$\begin{cases} u_d = R_s i_d + L_d \frac{di_d}{dt} - \omega_e L_q i_q \\ u_q = R_s i_q + L_q \frac{di_q}{dt} + \omega_e L_d i_d + \omega_e \psi_f \end{cases} \quad (1)$$

where u_d and u_q are the d - and q -axis stator voltages, respectively; i_d and i_q are the d - and q -axis stator currents, respectively; R_s is the stator resistance; L_d and L_q are the d - and q -axis inductances, respectively; ω_e is the electrical angular velocity; and ψ_f is the rotor permanent-magnet flux linkage.

The PMSM's electromagnetic torque is expressed as

$$T_e = \frac{3}{2} p_n [\psi_f i_q + (L_d - L_q) i_d i_q] \quad (2)$$

where T_e is the electromagnetic torque, and p_n is the number of pole pairs.

The mechanical equation of motion is given by

$$J \frac{d\omega_m}{dt} = T_e - T_L - B\omega_m \quad (3)$$

where T_L is the load torque, J the moment of inertia, B the viscous friction coefficient, and ω_m the mechanical angular velocity. The relationship between the electrical and mechanical angular velocities is

$$\omega_e = p_n \omega_m \quad (4)$$

For an SPMSM, it can be approximated that $L_d = L_q$. With the $i_d = 0$ vector-control strategy adopted in this study, the electromagnetic torque can be simplified as

$$T_e = \frac{3}{2} p_n \psi_f i_q \quad (5)$$

Substituting Eq. (5) into Eq. (3) gives

$$\frac{d\omega_m}{dt} = \frac{3p_n \psi_f}{2J} i_q - \frac{T_L + B\omega_m}{J} \quad (6)$$

Define

$$u = i_q \quad b_0 = \frac{3p_n \psi_f}{2J} \quad f = -\frac{T_L + B\omega_m}{J} \quad (7)$$

The speed subsystem can then be written as

$$\dot{\omega}_m = b_0 u + f \quad (8)$$

where u is the equivalent control input of the speed loop, namely the q -axis current; b_0 is the system input gain; and f represents the total system disturbance.

3. DESIGN OF THE IMPROVED SPEED CONTROLLER

As shown in Eq. (8), the PMSM speed subsystem can be represented as a first-order uncertain system subject to a lumped disturbance. Conventional LADRC employs linear error feedback and features a simple structure and straightforward parameter tuning. However, its dynamic response and disturbance rejection capability are limited under abrupt speed changes and load disturbances. The super-twisting algorithm drives the sliding variable to zero through a continuous control action, thereby reducing high-frequency chattering while preserving system robustness. Accordingly, an improved super-twisting algorithm is incorporated into the LADRC speed loop in this section. The reference-signal processing and disturbance-observation stages are also modified to improve the speed controller's tracking and disturbance-rejection performance.

3.1. Smoothing of the Speed Reference

Abrupt changes in the reference speed can cause a transient increase in the speed error and fluctuations in the control output. To address this issue, a second-order tracking differentiator is employed to smooth the reference speed. Let ω_{ref} denote the reference speed, v_1 the smoothed speed-reference signal, and v_2 its derivative. The tracking differentiator is designed as

$$\begin{cases} e_0 = v_1 - \omega_{\text{ref}} \\ \dot{v}_1 = v_2 \\ \dot{v}_2 = -r^2 e_0 - 2rv_2 \end{cases} \quad (9)$$

where r is the tracking-rate factor. Equation (9) smooths the transition of the reference-speed signal, thereby reducing the effect of abrupt speed-loop input changes on the control system.

3.2. Design of the Improved Super-Twisting Control Law

The speed-tracking error is defined as

$$s = v_1 - z_1 \quad (10)$$

where s denotes the sliding-mode variable of the speed loop.

The conventional super-twisting control law is expressed as

$$\begin{cases} u_0 = u_p(s) + u_s \\ \dot{u}_s = u_i(s) \end{cases} \quad (11)$$

where

$$\begin{cases} u_p(s) = k_1 |s|^{1/2} \text{sgn}(s) \\ u_i(s) = k_3 \text{sgn}(s) \end{cases} \quad (12)$$

Here, u_0 denotes the virtual control input before disturbance compensation; u_s is the internal state of the super-twisting controller; and k_1 and k_3 are the controller gains.

The conventional super-twisting controller provides good continuity and disturbance rejection performance. However, its switching terms employ the ideal sign function, which is discontinuous at the origin and may cause high-frequency chattering in practical control systems. To improve the continuity of the switching terms, a dynamic boundary-layer function is constructed as

$$H_s(s) = \frac{s}{|s| + \delta(s)} \quad (13)$$

where the dynamic boundary-layer thickness is defined as

$$\delta(s) = \delta_{\min} + (\delta_{\max} - \delta_{\min}) \exp(-\rho|s|) \quad (14)$$

Here, $\delta_{\min}$ and $\delta_{\max}$ denote the minimum and maximum boundary-layer thicknesses, respectively, and ρ is an adjustment coefficient.

Under identical controller parameters and operating conditions, the ideal sign function, the fixed boundary-layer function corresponding to $\rho = 0$, and the dynamic boundary-layer function corresponding to $\rho = 3$ are employed to compare the q -axis reference current responses. The results are shown in Fig. 1.

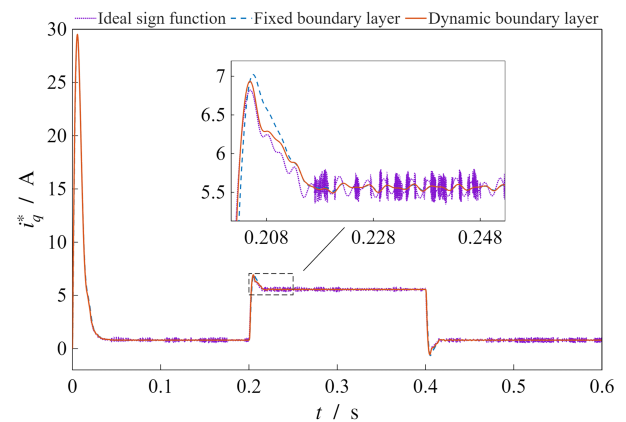


FIGURE 1. Comparison of q -axis reference current responses using different switching functions

To quantitatively evaluate high-frequency chattering, the steady-state interval from 0.25 to 0.39 s is selected. The standard deviation of the q -axis reference current is adopted as the

chattering evaluation index, defined as

$$J_{ch} = \sqrt{\frac{1}{N-1} \sum_{k=1}^N \left(i_{q,k}^* - \bar{i}_q^* \right)^2} \quad (15)$$

where N is the number of samples in the selected steady-state interval; $i_{q,k}^*$ is the q -axis reference current at the k -th sampling instant; and $\bar{i}_q^*$ is its mean value over this interval.

According to Eq. (15), the values of J_{ch} obtained using the ideal sign function, fixed boundary-layer function, and dynamic boundary-layer function are 0.072, 0.011, and 0.013 A, respectively. Compared with the ideal sign function, the dynamic boundary-layer function reduces this index by approximately 81.9%, quantitatively demonstrating its effectiveness in suppressing high-frequency chattering. The indices obtained using the dynamic and fixed boundary-layer functions are close, indicating that both functions provide effective steady-state chattering suppression. In the large-error region, the thickness of the dynamic boundary layer decreases, causing $H_s(s)$ to approach the ideal sign function and thereby strengthening the transient control action. Therefore, compared with the fixed boundary-layer function, the dynamic boundary-layer function maintains a comparably low steady-state fluctuation level while providing a faster transient response and a smaller response lag.

The lower-order nonlinear terms in the conventional super-twisting control law grow relatively slowly as the tracking error increases, thereby limiting the convergence rate when the system is far from the equilibrium point. Therefore, the dynamic boundary-layer function $H_s(s)$ in Eq. (13) is introduced into the switching terms in Eq. (12), and higher-order power terms are added to the original lower-order terms. The resulting improved super-twisting control law is given by

$$\begin{cases} u_p(s) = [k_1|s|^{1/2} + k_2|s|^\mu] H_s(s) \\ u_i(s) = [k_3 + k_4|s|^\mu] H_s(s) \end{cases} \quad (16)$$

Here, k_2 and k_4 are the gains of the higher-order power terms, and $\mu > 1$ is the corresponding exponent. The terms $k_1|s|^{1/2}H_s(s)$ and $k_3H_s(s)$ are lower-order terms, which mainly improve control accuracy and smoothness near the origin. $k_2|s|^\mu H_s(s)$ and $k_4|s|^\mu H_s(s)$ are higher-order power terms, which strengthen the control action when the tracking error is large and accelerate convergence process when the system is far from the equilibrium point.

3.3. Disturbance Compensation Based on the Improved ESO

To suppress high-frequency chattering during disturbance estimation in the speed loop and simplify the tuning of the nonlinear error-correction function, a Sigmoid smoothing function is incorporated into the conventional ESO to construct an improved ESO. The sigmoid function is continuous and bounded and contains only one slope-adjustment parameter, making its parameter structure simple and easy to tune. The function is approximately linear near the origin, which reduces the high-frequency chattering caused by discontinuous switching. For

large inputs, its output gradually approaches ± 1 , thereby preserving the nonlinear error-correction capability while improving the smoothness of the observer output. The sigmoid function is expressed as

$$\text{sigmoid}(x) = \frac{2}{1 + e^{-ax}} - 1 \quad (17)$$

where $a > 0$ is the slope-adjustment coefficient. The resulting Sigmoid-based improved ESO is designed as

$$\begin{cases} e_1 = z_1 - \omega_m \\ \dot{z}_1 = b_0 u + z_2 - \beta_1 e_1 \\ \dot{z}_2 = -\beta_2 \text{sigmoid}(e_1) \end{cases} \quad (18)$$

where e_1 is the speed-estimation error; z_1 is the estimated speed; z_2 is the estimate of the lumped disturbance in the speed loop; and β_1 and β_2 are the ESO gains.

Using z_2 as an estimate of the lumped disturbance f , the final output of the speed controller is given by

$$u = \frac{u_0 - z_2}{b_0} \quad (19)$$

where u is the equivalent control signal generated by the speed controller and corresponds to the q -axis current reference i_q^* .

3.4. Stability Analysis

According to Eqs. (10), (11), (16), and (19), the dynamics of the speed-tracking error s satisfy

$$\begin{cases} \dot{s} = v_2 - (b_0 u + z_2 - \beta_1 e_1) \\ b_0 u = u_0 - z_2 \\ u_0 = u_p(s) + u_s \end{cases} \quad (20)$$

Eliminating the intermediate variables and rearranging the resulting expression yields

$$\dot{s} = -u_p(s) - u_s + v_2 + \beta_1 e_1 \quad (21)$$

Define the ESO disturbance-estimation residual e_f and the lumped equivalent term d_s in the speed loop as

$$\begin{cases} e_f(t) = z_2(t) - f(t) \\ d_s(t) = v_2(t) + \beta_1 e_1(t) \end{cases} \quad (22)$$

Substituting Eq. (22) into Eq. (21) gives the augmented error system of the speed controller as

$$\begin{cases} \dot{s} = -u_p(s) - u_s + d_s(t) \\ \dot{u}_s = u_i(s) \end{cases} \quad (23)$$

Combining Eqs. (8) and (18), the ESO error dynamics are obtained as

$$\begin{cases} \dot{e}_1 = e_f - \beta_1 e_1 \\ \dot{e}_f = -\beta_2 \phi(e_1) - \dot{f} \end{cases} \quad (24)$$

where $\phi(e_1) = \text{sigmoid}(e_1)$. Under the constant lumped-disturbance condition considered here, $\dot{f} = 0$. The following Lyapunov function is selected, and its derivative along the trajectories of Eq. (24) is given by

$$\begin{cases} W(e_1, e_f) = \beta_2 \int_0^{e_1} \phi(\sigma) d\sigma + \frac{1}{2} e_f^2 \\ \dot{W} = -\beta_1 \beta_2 e_1 \phi(e_1) \leq 0 \end{cases} \quad (25)$$

Because the sigmoid function is continuous, odd, and strictly monotonically increasing, $\phi(e_1)$ always has the same sign as e_1 . Therefore, for any $(e_1, e_f) \neq (0, 0)$, $W(e_1, e_f) > W(0, 0) = 0$, and $W(e_1, e_f)$ is radially unbounded.

It follows from Eq. (25) that $\dot{W} = 0$ if and only if $e_1 = 0$. When $e_1 = 0$, Eq. (24) gives $\dot{e}_1 = e_f$. For a system trajectory

to remain in the set $\{\dot{W} = 0\}$, it must satisfy $\dot{e}_1 = 0$, and hence $e_f = 0$. Therefore, the largest invariant set contained in $\{\dot{W} = 0\}$ consists only of the origin $(e_1, e_f) = (0, 0)$. According to LaSalle's invariance principle, the origin of the ESO error subsystem is globally asymptotically stable; that is, both e_1 and e_f converge asymptotically to zero.

At the equilibrium operating point corresponding to a constant speed reference and a constant lumped disturbance, $v_2 = 0$. Combining Eq. (22) with the equilibrium conditions $e_1 = e_f = 0$ of the ESO error subsystem gives $d_s = 0$. The corresponding speed-loop error system is therefore expressed as

$$\begin{cases} \dot{s} = -u_p(s) - u_s \\ \dot{u}_s = u_i(s) \end{cases} \quad (26)$$

Consider the Lyapunov function

$$\begin{cases} V(s, u_s) = \int_0^s u_i(\sigma) d\sigma + \frac{1}{2} u_s^2 \\ u_i(\sigma) = k_3 H_s(\sigma) + k_4 |\sigma|^\mu H_s(\sigma) \end{cases} \quad (27)$$

Since $\delta(s) > 0$, $H_s(s)$, $u_p(s)$, and $u_i(s)$ always have the same sign as s . Consequently, the integral term in Eq. (27) is strictly positive for any $s \neq 0$. In addition, when $|s| \geq \delta_{\max}$,

$$|H_s(s)| = \frac{|s|}{|s| + \delta(s)} \geq \frac{|s|}{|s| + \delta_{\max}} \geq \frac{1}{2} \quad (28)$$

Therefore, the integral term in Eq. (27) tends to infinity as $|s|$ increases, and $V(s, u_s)$ is positive definite and radially unbounded. Differentiating V along the trajectories of Eq. (26) and using Eq. (16) gives

$$\begin{cases} u_p(s) = [k_1 |s|^{1/2} + k_2 |s|^\mu] H_s(s) \\ u_i(s) = [k_3 + k_4 |s|^\mu] H_s(s) \\ \dot{V} = -u_p(s) u_i(s) \leq 0 \end{cases} \quad (29)$$

It follows from Eq. (29) that $\dot{V} = 0$ if and only if $s = 0$. On the set $s = 0$, Eq. (26) gives $\dot{s} = -u_s$. For the system trajectory to remain in the set $\{\dot{V} = 0\}$, it is necessary that $\dot{s} = 0$, which implies $u_s = 0$. Hence, the largest invariant set contained in $\{\dot{V} = 0\}$ consists only of the origin. According to LaSalle's invariance principle, the origin of the system described by Eq. (26) is globally asymptotically stable, that is,

$$\begin{cases} \lim_{t \rightarrow \infty} s(t) = 0 \\ \lim_{t \rightarrow \infty} u_s(t) = 0 \end{cases} \quad (30)$$

In summary, according to LaSalle's invariance principle, the origins of the designed Sigmoid-ESO error subsystem and the speed-loop error system described by Eq. (26) corresponding to the improved super-twisting control law are both globally asymptotically stable.

3.5. Parameter Optimization Based on Improved Particle Swarm Optimization

As shown in Eq. (16), after introducing higher-order power terms, the parameters to be tuned in the speed controller are expanded to k_1 , k_2 , k_3 , and k_4 . These control gains jointly affect the system's dynamic response, disturbance recovery, and control-output smoothness. To reduce the dependence on empirical manual tuning and obtain a better combination of controller parameters, an improved particle swarm optimization algorithm is employed to optimize speed-loop controller parameters.

The parameter vector to be optimized is defined as

$$x = [k_1, k_2, k_3, k_4] \quad (31)$$

Tuning speed-loop parameters needs to account for tracking accuracy, transient performance, disturbance-rejection capability, and control smoothness of the closed-loop system. The integral of the speed tracking error reflects the accumulated deviation over the entire response process, while the steady-state error and overshoot are used to evaluate steady-state accuracy and transient response, respectively. The error over the load-disturbance interval is introduced to assess the system's recovery capability after a disturbance. According to the electromagnetic torque equation of the PMSM, when $\dot{i}_d = 0$, the electromagnetic torque is mainly determined by the q -axis current. Therefore, a q -axis current chattering term is included to constrain torque fluctuations and improve the smoothness of the control action. In addition, a parameter penalty term is introduced to prevent the parameters from excessively deviating from or approaching the search boundaries. Based on these performance requirements, the composite objective function is constructed as

$$J = w_1 J_e + w_2 J_s + w_3 J_0 + w_4 J_d + w_5 J_{iq} + w_6 J_p \quad (32)$$

Here, J_e denotes the speed tracking error term, J_s the steady-state error term, J_0 the overshoot penalty term, J_d the error term over the load-disturbance interval, J_{iq} the q -axis current chattering penalty term, and J_p the parameter-deviation or boundary penalty term. The coefficients w_1 – w_6 are the correspond-

ing weights. To reduce differences in dimensions and numerical ranges of the performance indices, each term is normalized using its corresponding reference quantity. The weights are assigned with emphasis on steady-state tracking accuracy and parameter feasibility, while also considering dynamic response, disturbance recovery, and current smoothness. The weights remain unchanged throughout the parameter-tuning process and are set as follows:

$$\begin{cases} w_1 = 1000, & w_2 = 3000, & w_3 = 2000 \\ w_4 = 1000, & w_5 = 1, & w_6 = 1 \end{cases} \quad (33)$$

The q -axis current-chattering term and parameter-penalty term are further defined as

$$\begin{cases} J_{iq} = 5J_{im} + 120J_f + 250J_l \\ J_p = 40J_{pd} + 3000J_{pb} \end{cases} \quad (34)$$

where J_{im} , J_f , and J_l denote the current-amplitude term, the overall current-chattering term, and the current-chattering term over the load-disturbance interval, respectively; J_{pd} and J_{pb} denote the parameter-deviation and boundary-penalty terms, respectively.

The velocity and position of each particle are updated according to

$$\begin{cases} v_{i,d}^{t+1} = w^t v_{i,d}^t + c_1^t r_1^t (p_{i,d}^t - x_{i,d}^t) + c_2^t r_2^t (g_d^t - x_{i,d}^t) \\ x_{i,d}^{t+1} = x_{i,d}^t + v_{i,d}^{t+1} \end{cases} \quad (35)$$

Here, $x_{i,d}^t$ and $v_{i,d}^t$ denote the position and velocity, respectively, of particle i in dimension d at iteration t ; $p_{i,d}^t$ is the personal best position of particle i ; g_d^t is the global best position of the swarm; and r_1^t and r_2^t are random numbers in the interval $[0, 1]$.

To improve the distribution uniformity of the initial particles, a Logistic chaotic map is employed to generate the initial particle swarm:

$$\begin{cases} z_{i,d}^{m+1} = 4z_{i,d}^m (1 - z_{i,d}^m) \\ x_{i,d}^0 = L_d + z_{i,d}^0 (U_d - L_d) \end{cases} \quad (36)$$

Here, L_d and U_d denote the lower and upper bounds, respectively, of the parameter in the d th dimension. The inertia weight and learning factors are adjusted during the iterative process according to

$$\begin{cases} w^t = w_{\min} + \frac{w_{\max} - w_{\min}}{2} [1 + \cos(\pi\tau)] \\ c_1^t = 2.5 - 2\tau \\ c_2^t = 0.5 + 2\tau \\ \tau = \frac{t-1}{T-1} \end{cases} \quad (37)$$

Here, T denotes the maximum number of iterations.

IPSO is an improved algorithm developed within the conventional PSO framework. Other intelligent optimization algorithms employ different search mechanisms and parameter configurations, and their comparative results may also be affected by algorithm structures and parameter settings. Therefore, conventional PSO is selected as the direct baseline. Both algorithms use the same objective function, weighting coefficients, and parameter search ranges. The swarm size and maximum number of iterations are set to 20 and 100, respectively, for both algorithms. The resulting parameter-tuning convergence curves are shown in Fig. 2.

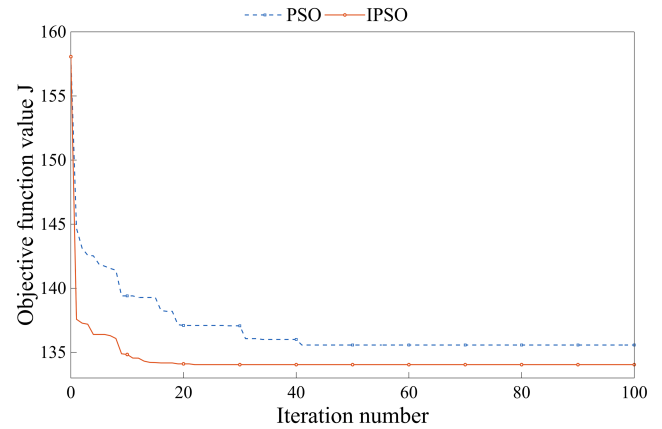


FIGURE 2. Convergence curves of speed-loop controller parameter tuning using PSO and IPSO.

As shown in Fig. 2, both algorithms gradually reduce the objective-function value. IPSO exhibits a faster decrease during the early iterations and stabilizes after approximately 22 iterations, whereas conventional PSO stabilizes after approximately 41 iterations. Compared with conventional PSO, IPSO achieves a faster convergence rate and a lower final objective-function value. These results demonstrate that IPSO can effectively tune the parameters of the proposed speed-loop controller.

The final structure of the speed-loop controller is shown in Fig. 3.

4. DESIGN OF THE SENSORLESS CONTROL SYSTEM

To reduce cost and alleviate reliability issues associated with mechanical position sensors, a sensorless control structure based on the LESO-IPLL is employed to estimate the rotor position and speed. First, an LESO is constructed in the stationary α - β reference frame to estimate the PMSM's back electromotive force. Its formulation is given by

$$\begin{cases} e_\gamma = z_{\gamma 1} - i_\gamma \\ \dot{z}_{\gamma 1} = z_{\gamma 2} - \beta_1 e_\gamma + \frac{1}{L_s} u_\gamma - \frac{R_s}{L_s} i_\gamma \\ \dot{z}_{\gamma 2} = -\beta_2 e_\gamma \end{cases} \quad \gamma = \alpha, \beta \quad (38)$$

where i_γ and u_γ denote the current and voltage components, respectively, along the γ -axis ($\gamma = \alpha, \beta$); $z_{\gamma 1}$ is the estimated

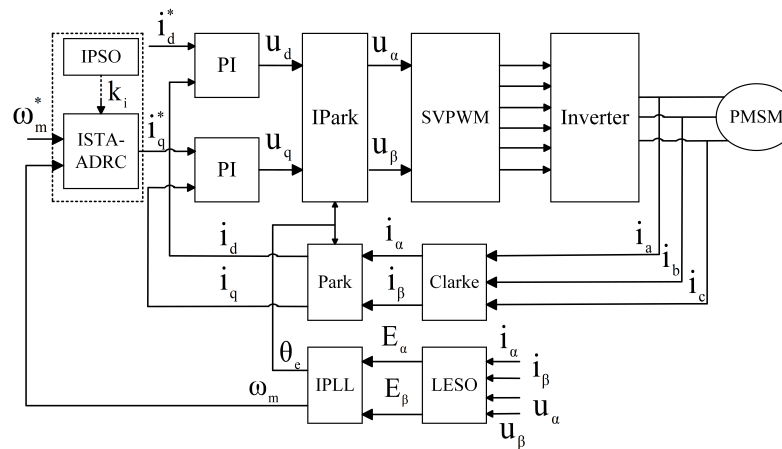


FIGURE 5. Block diagram of PMSM sensorless vector control based on IPSO-ISTA-ADRC.

5.1. Startup Response Analysis

To evaluate the dynamic response performance of different control strategies during motor startup, the PMSM's reference speed is set to 1000 r/min, and a startup simulation is conducted under no-load conditions. The motor speed response curves are shown in Fig. 6.

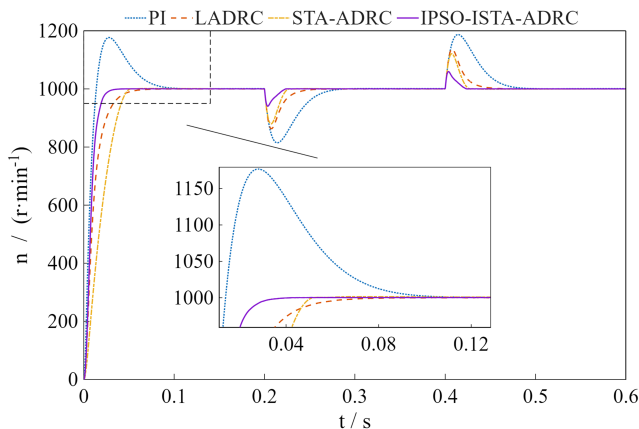


FIGURE 6. Motor speed response curves under no-load conditions.

As shown in Fig. 6, during no-load startup, the conventional PI controller provides a fast speed rise but exhibits a pronounced overshoot. Its peak speed is approximately 1176.8 r/min, corresponding to an overshoot of 17.68%. The LADRC and STA-ADRC strategies effectively reduce the startup overshoot, with settling times of 0.081 s and 0.052 s, respectively. In comparison, the proposed IPSO-ISTA-ADRC strategy tracks the reference speed more rapidly during startup, with a settling time of 0.039 s and no noticeable overshoot, demonstrating good startup response and stability.

5.2. Load Disturbance Response Analysis

To evaluate the disturbance-rejection capability of different control strategies under abrupt load changes, a load-disturbance simulation is conducted after the motor reaches steady-state operation. The reference speed is set to 1000 r/min. A load torque of 10 N · m is applied at 0.2 s and removed at 0.4 s, while all other simulation parameters remain unchanged. The motor

speed response curves under the different control strategies are shown in Fig. 7.

As shown in Fig. 7 and Table 3, during both load application and load removal, the proposed method produces the smallest speed deviation and the shortest recovery time. Compared with STA-ADRC, the proposed method reduces the speed deviation by approximately 50% and shortens the recovery time by approximately 14.3%, demonstrating good disturbance rejection capability under load variations.

TABLE 3. Comparison of performance indices under load disturbance.

Control method	Extreme speed (r/min)	Speed deviation (r/min)	Recovery time (s)
Stage 1: Application of a 10 N · m load at 0.2 s			
PI	815	185	0.100
LADRC	863	137	0.074
STA-ADRC	880	120	0.028
IPSO-ISTA-ADRC	940	60	0.024
Stage 2: Removal of the 10 N · m load at 0.4 s			
PI	1187	187	0.100
LADRC	1137	137	0.076
STA-ADRC	1121	121	0.028
IPSO-ISTA-ADRC	1059	59	0.024

Under the above load-disturbance condition, the electromagnetic torque response curves of different control strategies are shown in Fig. 8. As shown in the figure, all control strategies exhibit a certain degree of torque fluctuation during load application and removal. In comparison, IPSO-ISTA-ADRC provides a smoother electromagnetic torque response and a faster recovery following load variation, indicating that the proposed method improves the PMSM's electromagnetic operating stability under load disturbance.

5.3. Speed Response Analysis Under Composite Operating Conditions

To further evaluate the dynamic tracking and disturbance rejection performance of the proposed IPSO-ISTA-ADRC strat-

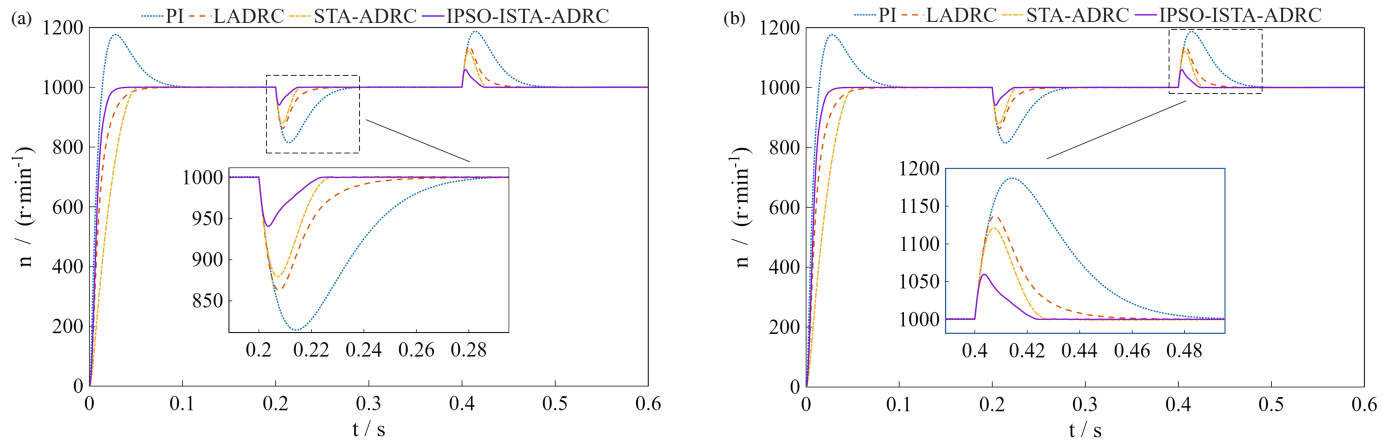


FIGURE 7. Motor speed response curves under load disturbance. (a) Speed response during load application at 0.2 s and (b) speed response during load removal at 0.4 s.

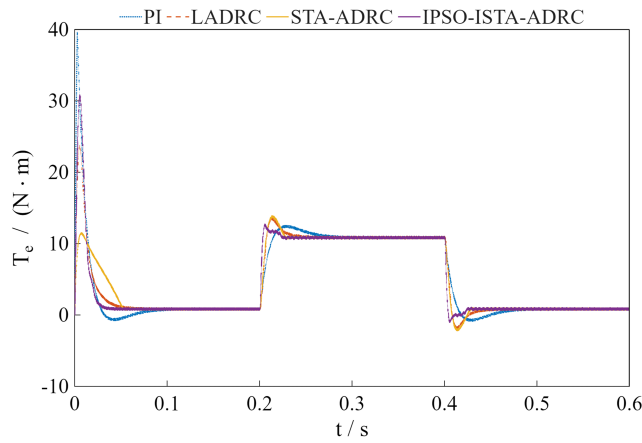


FIGURE 8. Electromagnetic torque response curves under load disturbance

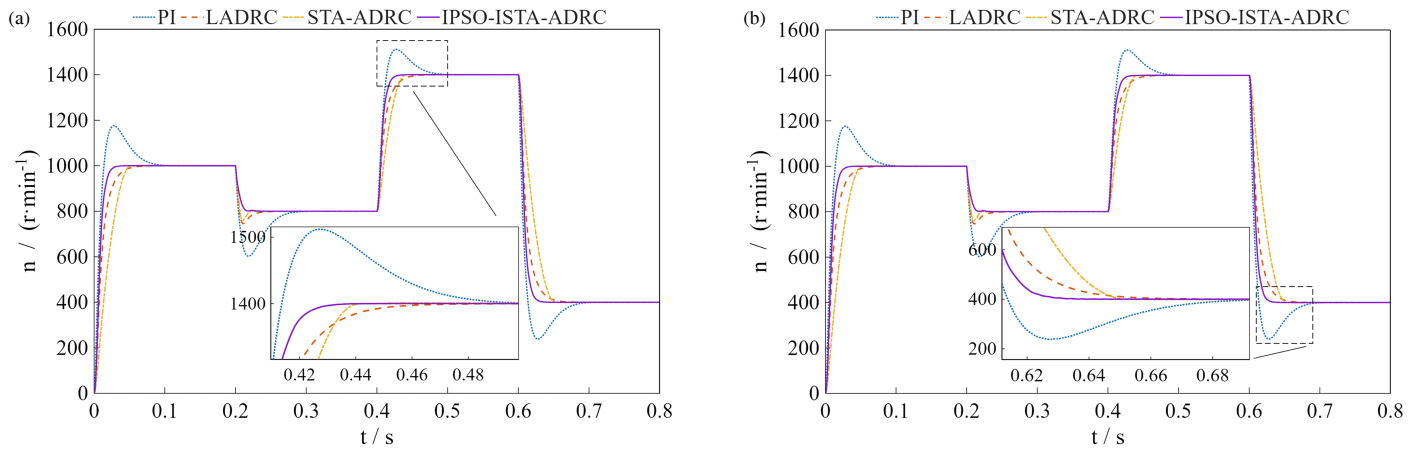


FIGURE 9. Motor speed response curves under load disturbance and speed variations. (a) Speed response during the increase from 800 r/min to 1400 r/min at 0.4 s. Speed response during the decrease from 1400 r/min to 400 r/min at 0.6 s.

egy under composite operating conditions, a simulation combining a load disturbance and speed variations is conducted. The simulation time is set to 0.8 s, and the initial reference speed is 1000 r/min. At 0.2 s, a load torque of 10 N · m is applied, while the reference speed is simultaneously reduced from

1000 r/min to 800 r/min. At 0.4 s, the reference speed is increased to 1400 r/min, and at 0.6 s, it is reduced to 400 r/min. The motor speed response curves under different control strategies are shown in Fig. 9.

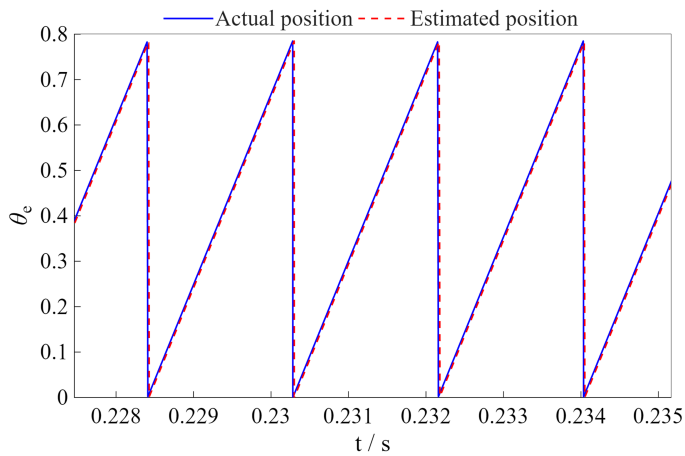


FIGURE 10. Comparison of actual and estimated rotor positions.

TABLE 4. Comparison of dynamic performance indices under speed variations.

Control method	Peak deviation (%)	Settling time (s)
Stage 1: Speed increase from 800 r/min to 1400 r/min at 0.4 s		
PI	7.99	0.098
LADRC	0	0.073
STA-ADRC	0	0.041
IPSO-ISTA-ADRC	0	0.038
Stage 2: Speed decrease from 1400 r/min to 400 r/min at 0.6 s		
PI	40.5	0.107
LADRC	0	0.082
STA-ADRC	0	0.052
IPSO-ISTA-ADRC	0	0.042

As shown in Fig. 9 and Table 4, the proposed method provides the shortest settling time during both speed increase and decrease, with no noticeable overshoot during the speed increase. Compared with STA-ADRC, the proposed method shortens the settling time by approximately 7.3% and 19.2% during speed increase and decrease, respectively, demonstrating good dynamic tracking performance.

5.4. Sensorless Estimation Performance Analysis

To verify the effectiveness of the sensorless control scheme adopted in this paper, the rotor-position and speed-estimation performance of the PMSM is evaluated through simulation. The LESO is used to estimate the back electromotive force, from which the rotor position and speed information is extracted using the improved phase-locked loop. Fig. 10 compares the actual and estimated rotor positions, while Fig. 11 shows speed-estimation error.

As shown in Figs. 10 and 11, the estimated rotor position closely tracks the actual rotor position, and the overall speed-estimation error remains small. During load disturbances and speed step changes, the estimation error exhibits only brief fluctuations

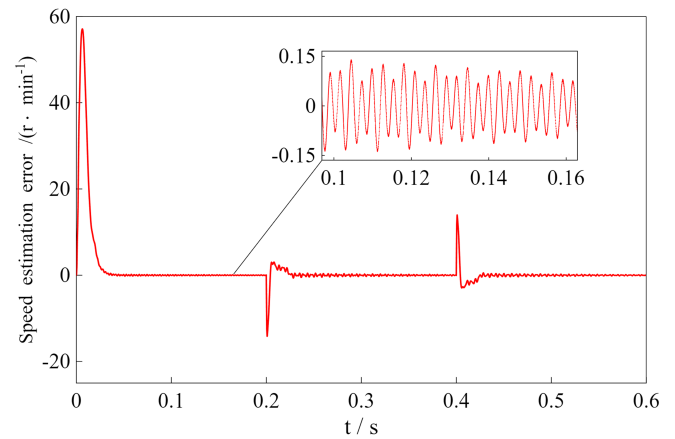


FIGURE 11. Speed-estimation error.

tuations and rapidly converges to a neighborhood of zero. Under steady-state operation, the speed-estimation error remains approximately within ± 0.15 r/min. These results demonstrate that the LESO-IPLL sensorless structure adopted in this paper provides good rotor-position and speed-estimation performance.

6. CONCLUSION

In this paper, a super-twisting active disturbance rejection control method augmented with higher-order power terms and optimized using improved particle swarm optimization is applied to the PMSM speed loop. Simulation results show that the proposed method reduces speed overshoot and settling time during no-load startup, thereby improving the system's dynamic response.

Under load disturbance and speed variation conditions, the proposed method effectively suppresses speed fluctuations and shortens the system recovery process. Compared with boundary-layer-based STA-ADRC, the proposed method shortens the recovery time under load disturbance by approximately 14.3% and the settling time during the speed decrease by approximately 19.2%. Meanwhile, the electromagnetic torque responses under load disturbance show that the proposed method reduces torque fluctuations caused by load variations and improves the electromagnetic operational stability of the PMSM.

The LESO-IPLL sensorless structure enables rotor position and speed estimation, with the speed-estimation error remaining within approximately ± 0.15 r/min. Simulations under multiple operating conditions demonstrate that the proposed method improves the dynamic response of the PMSM and the stability of sensorless operation. Future work will include experimental validation on a PMSM platform. Since the LESO-IPLL relies on back-EMF information, the proposed method is mainly applicable to medium- and high-speed operation. Future work will investigate a composite sensorless method combining auxiliary low-speed position estimation with back-EMF observation, as well as online adaptive optimization of the controller and observer parameters.

ACKNOWLEDGEMENT

This work was supported by the National Natural Science Foundation of China under Grant 52067008 and the Science and Technology Project of Jiangxi Provincial Department of Education under Grant GJJ170554.

REFERENCES

- [1] Yang, Z., C. Qian, X. Sun, and K. Wang, "Enhanced linear ADRC strategy for sensorless control of IPMSM considering cross-coupling factors," *IEEE Transactions on Transportation Electrification*, Vol. 9, No. 3, 4437–4446, 2023.
- [2] Zhang, P., Z. Shi, B. Yu, and H. Qi, "Research on a sensorless adrc vector control method for a permanent magnet synchronous motor based on the luenberger observer," *Processes*, Vol. 12, No. 5, 906, 2024.
- [3] Kang, E.-L., B.-D. Wu, and C. Yu, "Research on sensorless of permanent magnet synchronous motor based on LADRC control," *Electric Machines and Control*, Vol. 27, No. 2, 69–78, 2023.
- [4] Zhang, G. Q. and J. H. Du, "Review of position sensorless control technology for permanent magnet synchronous motors," *Electric Machines & Control Application*, Vol. 51, No. 1, 1–13, 2024.
- [5] Liu, X., G. Zhang, and Z. Shi, "Improved current control for PMSM via an active disturbance rejection controller," *European Journal of Control*, Vol. 78, 101005, 2024.
- [6] Wang, C., J. Yan, W. Li, L. Shan, and L. Sun, "Disturbances rejection optimization based on improved two-degree-of-freedom LADRC for permanent magnet synchronous motor systems," *Defence Technology*, Vol. 33, 518–531, 2024.
- [7] Wu, H., C. Gan, H. Wang, S. Wang, R. Qu, and X. Liu, "Active disturbance rejection speed control with double-stage-ESO considering aperiodic and periodic disturbances for PMSM drives," *IEEE Transactions on Industrial Electronics*, 2025.
- [8] Chen, L., H. Zhang, H. Wang, K. Shao, G. Wang, and A. Yazdani, "Continuous adaptive fast terminal sliding mode-based speed regulation control of PMSM drive via improved super-twisting observer," *IEEE Transactions on Industrial Electronics*, Vol. 71, No. 5, 5105–5115, 2024.
- [9] Ma, Y. X., Z. Q. Lei, P. P. Gu, *et al.*, "Research on PMSM speed control based on improved super-twisting sliding mode linear active disturbance rejection," *Electric Machines & Control Application*, Vol. 51, No. 9, 103–112, 2024.
- [10] Zhang, Y. H., J. J. Huang, Y. T. Li, *et al.*, "Research on PMSM control strategy based on improved super-twisting sliding mode linear active disturbance rejection," *Power Electronics*, Vol. 59, No. 11, 120–126, 2025.
- [11] Chen, H., X. W. Bi, L. Yuan, *et al.*, "Improved super-twisting sliding-mode control applied to PMSM speed control system," *Power Electronics*, Vol. 58, No. 6, 37–39, 2024.
- [12] Chen, T., Q. Jiang, B. Jin, *et al.*, "Sliding mode active disturbance rejection control for PMSM based on improved ESO," *Micro-motors*, Vol. 57, No. 12, 40–46, 2024.
- [13] Jian, W., W. T. Hu, G. S. Peng, *et al.*, "Active disturbance rejection speed control strategy of PMSM based on improved extended state observer," *Modular Machine Tool & Automatic Manufacturing Technique*, No. 11, 142–146, 151, 2024.
- [14] Zhang, J. W., Z. X. Yang, G. L. Qi, *et al.*, "Improved active disturbance rejection control of permanent magnet synchronous motor," *Power Electronics*, Vol. 58, No. 11, 30–33, 2024.
- [15] Liu, X. D., K. C. Wang, Z. B. Wei, *et al.*, "Research on linear/nonlinear active disturbance rejection switching control algorithm for permanent magnet synchronous motor," *Journal of Chongqing University of Posts and Telecommunications(Natural Science Edition)*, Vol. 37, No. 4, 576–585, 2025.
- [16] Du, W. S. and Z. G. Wang, "Improved super-twisting sliding mode linear active disturbance rejection control of PMSM based on mixed grey wolf optimizer," *Large Electric Machine and Hydraulic Turbine*, No. 2, 27–36, 2026.
- [17] Ma, R. C., Q. Tan, X. R. Liu, *et al.*, "Active disturbance rejection control for permanent magnet linear synchronous motors using an enhanced particle swarm optimization algorithm," *Electric Machines and Control*, Vol. 30, No. 2, 36–48, 2026.