Research Article

A High-Performance Indirect Torque Control Strategy for Switched Reluctance Motor Drives

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This paper proposes a high-performance indirect control scheme for torque ripple minimization in the switched reluctance motor (SRM) drive system. Firstly, based on the nonlinear torque-angle characteristic of SRM, a novel torque sharing function is developed to obtain the optimal current profiles such that the torque ripple is minimized with reduced copper losses. Secondly, in order to track current accurately and indirectly achieve high-performance torque control, a robust current controller is derived through the Lyapunov stability theory. The proposed robust current controller not only considers the motor parameter modeling errors but also realizes the fixed frequency current control by introducing the pulse width modulation method. Further, a disturbance-observer-based speed controller is derived to regulate the motor speed accurately, and the load torque is considered an unknown disturbance. The simulations and experiments on a 1.5 kW SRM prototype are carried out to demonstrate the effectiveness of the proposed high-performance indirect torque control strategy. Results verify the superiority of the proposed strategy with respect to the torque ripple suppression, system efficiency, and antidisturbance.

1. Introduction

Switched reluctance motor (SRM) has recently attracted much attention from the industrial and academic communities due to its own advantages, such as simple and strong structures, high reliability, wide speed range, no need for rare Earth permanent magnetic materials, and low manufacturing cost [1–4]. However, relatively higher torque ripples are viewed as a significant disadvantage in comparison with other types of motors [5, 6]. Therefore, it is of great significance to study the torque ripple suppression of SRM for its further popularization and application.

The literature review reveals that the torque ripple of SRM can be effectively reduced by means of an appropriate control algorithm [7]. At present, the torque control methods of SRM mainly include direct torque control (DTC) and indirect torque control (ITC). In DTC, the switching signal of the power converter is directly generated by combining the error between the command torque and the instantaneous torque of SRM with the appropriate commutation logic [8–10]. The instantaneous torque of the SRM is usually calculated from the measured rotor position and phase current [11]. DTC has the advantage of simple structure and easy implementation because it has no current loop [12]. However, the shortcomings of DTC strategy, such as high sampling rate, no over-current protection, and variable switching frequency, greatly limit its popularization and application.

Alternatively, ITC methods for SRM can partly overcome these disadvantages. In ITC, torque sharing function (TSF) is adopted to distribute the electromagnetic torque command to each phase of the motor, and then the optimized phase current reference is obtained by the nonlinear mapping of current-torque-angle i(T, θ). Torque control can be indirectly realized by regulating the current to track the optimized current reference [13–15]. Therefore, the key to determining the performance of the ITC algorithm is the TSF and the inner-loop current controller. In [16], the effects of different TSFs such as linear, cosine, quadratic, and exponential to reduce the torque ripple of SRMs are compared and evaluated. However, these TSFs will produce a larger peak current during commutation, which will increase the
torque ripple and reduce the driving efficiency of the SRM. An offline optimization method of TSF is used in [17] to minimize torque ripple, and its objective function combines both the phase current and the rate of change of flux-linkage. In [18], the objective function of offline TSF is further improved and simplified by using a single weight parameter. In [19], an online compensation of TSF method is developed to smooth the torque output by applying positive and negative compensation to the outgoing phase and incoming phase, respectively. In [20], a proportional-integral (PI) controller is adopted to compensate for the tracking error of the torque in real time. However, in [19, 20], a large memory is needed to store additional torque characteristics \( T(i, \theta) \) for the on-line estimation of torque.

For iTC, the performance of the current controller will also directly affect the torque ripples of the motor. Hysteresis controllers are widely used in current loops because of their simple structure, model independence, and fast dynamic response [21, 22]. However, when hysteresis current controller is used in the SRM drive system, the switching frequency of power converter is uncontrollable, which will produce some unpredictable acoustic noise. To solve this problem, a digital pulse-width-modulation (PWM) current controller for SRM is studied with constant switching frequency in [23]. However, the digital PWM current controller cannot guarantee that the tracking error tends to zero when the current reference is a time-varying signal. A model-based predictive current controller is investigated in [24]. The predictive controller achieves the fixed switching frequency. However, the performance of the predictive controller needs a large gain, which may degrade the performance of the controller. Additionally, the robustness of the predictive current controller was not investigated.

In order to simultaneously solve the aforementioned problems, a high-performance indirect torque control (HPITC) strategy for SRM drive is proposed in this paper. To the best of our knowledge, it is the first time in the literature that TSF optimization, accurate current tracking, and disturbance rejection are simultaneously dealt with in the SRM drive system. A novel TSF is developed to suppress the torque ripple while reducing the peak current during commutation. Then, based on the Lyapunov stability theory, a robust current controller is derived to achieve accurate current tracking. Moreover, a novel speed controller associated with a load torque observer has been developed to achieve accurate speed regulation and improve the antidisturbance performance of the SRM drive system. Due to the comprehensive improvement of the HPITC strategy, the proposed method offers the feasibility of effectively reducing the torque ripple, improving system efficiency, and enhancing antidisturbance ability. The effectiveness of the HPITC method is demonstrated by simulations and experiments.

2. SRM Model

Neglecting the effect of mutual inductance, the phase torque of SRM can be calculated as

\[
T_p = \frac{dW_p}{d\theta} \bigg|_{i_p=\text{const}} = \frac{dW_s}{d\theta} \bigg|_{\psi_p=\text{const}}, \tag{1}
\]

where \( W_c \) and \( W_s \) denote the magnetic coenergy and the magnetic energy storage, which can be calculated as

\[
\begin{aligned}
W_c &= \int_0^{i_p} \psi_p(\theta, i_p) di_p, \\
W_s &= \int_0^{\psi_p} i_p(\theta, \psi_p) d\psi_p,
\end{aligned} \tag{2}
\]

where \( T_p, i_p, \) and \( \psi_p \) are the \( p \)-th phase torque, the \( p \)-th phase current, and the \( p \)-th phase flux-linkage, respectively. \( \theta \) is the rotor position.

The total electromagnetic torque \( T_e \) of the studied three-phase SRM is described as follows:

\[
T_e = \sum_{p=1}^{3} T_p. \tag{3}
\]

In order to improve the energy conversion, the SRM always works in the magnetic saturation region. In this region, the electromagnetic characteristics of the SRM exhibit high nonlinearity. As a result, it is difficult to establish an accurate nonlinear model of SRM based on conventional electromagnetic and physical characteristics deduction. At present, most of the nonlinear models of SRMs are identified based on the sample data of electromagnetic characteristics obtained by finite element analysis (FEA) or experiment [25]. An analytical model [26] that is embedded in the SimPowerSystems toolbox of Matlab/Simulink software is adopted in this paper. As analyzed in [26], the flux-linkage profile of SRM is represented analytically as

\[
\psi_p(\theta_p, i_p) = L_q i_p + [L_{\text{dsat}} i_p + A(1 - e^{B_i} - L_q i_p)] f(\theta_p), \tag{4}
\]

where \( L_q \) denotes the inductance at unaligned rotor position \((q\text{-axis})\) and \( L_{\text{dsat}} \) denotes the saturated inductance at aligned rotor position \((d\text{-axis})\). \( A, B, \) and \( f(\theta_p) \) can be further represented as follows:

\[
\begin{aligned}
A &= \psi_m - L_{\text{dsat}} i_m, \\
B &= \frac{L_d - L_{\text{dsat}}}{\psi_m - L_{\text{dsat}} i_m}, \\
f(\theta_p) &= \begin{cases} 1024 \pi^3 \theta_p^3 - 192 \pi^2 \theta_p^2 + 1, & \theta_p \in [0, \frac{\pi}{8}], \\ f\left(\frac{\pi}{4} - \theta_p\right), & \theta_p \in \left[\frac{\pi}{8}, \frac{\pi}{4}\right], \end{cases}
\end{aligned} \tag{5}
\]

where \( L_d \) denotes the nonsaturated inductance in \( d \)-axis and \( i_m \) denotes the rated maximum current with corresponding flux-linkage \( \psi_m \).

Substituting (4) into (1), the stator phase torque \( T_p \) can be calculated as
The experimental measurement of the studied SRM was conducted using the method in [25]. The measured flux-linkage and torque profiles of the motor are shown in Figures 1(a) and 1(b), respectively. Based on the measured electromagnetic characteristics, as shown in Figure 1, the model coefficients \( \psi_m = 0.9 \), \( L_q = 0.0226 \), \( L_{\text{dissat}} = 0.0185 \), \( L_d = 0.3152 \), and \( f_m = 10 \) are obtained by the least-squares fitting. Thus, the nonlinear model of SRM is established.

### 3. High-Performance Indirect Torque Control

The schematic diagram of the HPITC system is shown in Figure 2, involving speed controller, TSF, torque-to-current module, current controller, 12/8 SRM, position and current sensors, and power converter.

As shown in Figure 2, the total torque command \( T_p^* \) is derived by the designed disturbance-observer-based speed controller. The inputs of disturbance observer are speed \( \omega \) and reference torque \( \tau_p^* \). In the torque control scheme, the torque reference \( T_p^* \) is distributed for every single phase by using a novel torque sharing function. Then the phase current command is generated from the current-torque sharing function. Hence, the nonlinear model of SRM is established.

#### 3.1. Torque Sharing Function

At present, the most commonly used TSFs mainly include linear, cubic, exponential, and sinusoidal functions [16]. A typical torque sharing curve of the sinusoidal TSF is shown in Figure 3(a). The phase torque command \( T_p^* \) can be obtained by a predefined TSF as follows:

\[
T_p^* (\theta) = \begin{cases} 
0, & (0 \leq \theta \leq \theta_{\text{on}}), \\
T_c f_{\text{rise}} (\theta), & (\theta_{\text{on}} \leq \theta \leq \theta_{\text{on}} + \theta_{\text{ov}}), \\
T_c \theta_{\text{ov}}, & (\theta_{\text{on}} + \theta_{\text{ov}} \leq \theta \leq \theta_{\text{off}} - \theta_{\text{ov}}), \\
T_c f_{\text{fall}} (\theta), & (\theta_{\text{off}} - \theta_{\text{ov}} \leq \theta \leq \theta_{\text{off}}), \\
0, & (\theta_{\text{off}} \leq \theta \leq \theta_{\text{c}}),
\end{cases}
\]

(7)

where \( f_{\text{rise}} (\theta) \) and \( f_{\text{fall}} (\theta) \) denote the rising TSF and decreasing TSF, respectively. \( \theta_{\text{on}} \), \( \theta_{\text{off}} \), and \( \theta_{\text{ov}} \) represent turn-on angle, turn-off angle, and overlap angle, respectively. In the overlap region, the total electromagnetic torque command is equal to the sum of torque commands for incoming phase and outgoing phase. Hence, the relationship between \( f_{\text{rise}} \) and \( f_{\text{fall}} \) can be described as

\[
f_{\text{fall}} (\theta) = 1 - f_{\text{rise}} (\theta + \theta_{\text{on}} + \theta_{\text{ov}} - \theta_{\text{off}}).
\]

(8)

From Figures 1(b) and 3(a), it is observed that the torque curves in the rising and falling stages are quite different from conventional TSF curves. If the conventional TSF is used, the incoming phase will be allocated a large reference torque in the initial stage of excitation, which will increase the peak current of stator winding during commutation. As a result, the copper losses of the SRM will increase and the drive efficiency of the motor system will decrease.

In order to further reduce copper losses while minimizing torque ripples, a novel nonlinear TSF is presented as shown in Figure 3(b). The proposed TSF uses a power function to distribute the total torque reference in the overlap region of incoming and outgoing phases. The functions \( f_{\text{rise}} \) and \( f_{\text{fall}} \) of the proposed TSF can be described as

\[
f_{\text{rise}} (\theta) = \left( \frac{\theta - \theta_{\text{on}}}{\theta_{\text{ov}}} \right)^a,
\]

(9)

\[
f_{\text{fall}} (\theta) = 1 - \left( \frac{\theta + \theta_{\text{off}} - \theta_{\text{ov}}}{\theta_{\text{ov}}} \right)^a,
\]

(10)

where \( a \geq 2 \) is a positive constant. The larger the value of \( a \), the smaller the torque reference allocated to the incoming phase at the initial stage.

The TSF parameters \( \theta_{\text{on}} \), \( \theta_{\text{off}} \), and \( \theta_{\text{ov}} \) can be linked as

\[
\theta_{\text{off}} - \theta_{\text{on}} = \theta_{\text{ov}} + \epsilon,
\]

(10)

where \( \epsilon = 360/\text{(mN)} \).

#### 3.2. Robust Current Controller Design

For SRM, phase voltage \( u_p \) is described as follows:

\[
u_p = \frac{d\psi_p}{dt} + R i_p,
\]

(11)

where \( R \) denotes the resistance of each phase winding of the motor.

The phase flux-linkage \( \psi_p (i_p, \theta_p) \) is a function of the phase current \( i_p \) and rotor position \( \theta_p \). Thus, the differential of flux-linkage is derived as

\[
\frac{d\psi_p}{dt} = \frac{\partial \psi_p}{\partial i_p} \frac{di_p}{dt} + \frac{\partial \psi_p}{\partial \theta_p} \frac{d\theta_p}{dt}.
\]

(12)

Substituting (12) into (11), the derivative of phase current \( i_p \) of SRM can be represented as

\[
\frac{di_p}{dt} = \left( \frac{\partial \psi_p}{\partial i_p} \right)^{-1} \left( u_p - Ri_p - \frac{\partial \psi_p}{\partial \theta_p} \omega \right)
\]

\[
= \left( \frac{\partial \psi_p}{\partial i_p} \right)^{-1} \left( -i_p R - \frac{\partial \psi_p}{\partial \theta_p} \omega \right) + L_{\text{inc}}^{-1} u_p,
\]

(13)

where \( L_{\text{inc}} = (\partial \psi_p/\partial i_p) \) denotes the increment inductance and \( E_{\text{bmf}} = (\partial \psi_p/\partial \theta_p) \) is the back electromotive force constant. Further, according to the flux-linkage expression (4), the constituent terms in (13) like incremental inductance \( L_{\text{inc}} \) and back electromotive force constant \( E_{\text{bmf}} \) can be calculated as follows:
\[
L_{\text{inc}} = L_q + \left[ L_{\text{dsat}} + AB \theta_{\text{p}} - L_q \right] f(\theta_p),
\]
\[
E_{\text{bmf}} = \left[ (L_{\text{dsat}} - L_q) \theta_{\text{p}} + A(1 - e^{-B \theta_{\text{p}}}) \right] f(\theta_p).
\]
phase current $i_p$ can accurately track its reference $i_p^*$ in the presence of model uncertainty.

Without considering the modeling errors, the ideal control law can be derived as

$$u = L_{inc} \left[ i_p^* + L_{inc}^{-1} (i_p R + \omega E_{bmf}) + ye \right],$$

(16)

where $y > 0$ is the controller parameter.

Lyapunov stability theory is adopted to design the control law $u(t)$ for current tracking control of SRM. Select the following Lyapunov function candidate:

$$V = \frac{1}{2} e^2.$$  

(17)

The time derivative of (17) is

$$\dot{V} = ee.$$  

(18)

By substituting (13), (15), and (16) into (18), (18) can be rewritten as

$$\dot{V} = e \left[ i_p^* + L_{inc}^{-1} (i_p R + \omega E_{bmf} \omega) - L_{inc}^{-1} u_p \right]$$

$$- ye^2 - 0.$$  

(19)

It should be noted that $\dot{V} < 0$ for all the conditions except $e = 0$. Equation (19) indicates that the phase current $i_p$ can track the current reference $i_p^*$ asymptotically. However, in the real application, the terms $L_{inc}$ and $E_{bmf}$ cannot be calculated accurately because of the modeling errors in equation (4). Accordingly, the real values of the increment inductance $L_{inc}$ and back electromotive force constant $E_{bmf}$ can be divided into the estimated values and their modeling errors. Thus, equation (14) is rewritten as

$$\begin{cases} L_{inc} = \tilde{L}_{inc} + L_{inc} \Delta L_{inc}, \\ E_{bmf} = \tilde{E}_{bmf} + E_{bmf} \Delta E_{bmf}, \end{cases}$$

(20)

A robust control design method is used to overcome the above model uncertainty. The robust current controller allows the use of a simplified nonlinear model of SRM in a control system design, which is readily available for magnetization curves at two extreme positions. Using the estimated values $\tilde{L}_{inc}$ and $\tilde{E}_{bmf}$ instead of the actual values $L_{inc}$ and $E_{bmf}$, the control law (16) can be reformulated as

$$\tilde{u}(t) = \tilde{L}_{inc} \left[ i_p^* + L_{inc}^{-1} (i_p R + \omega \tilde{E}_{bmf}) + ye \right].$$

(21)

Substituting (21) into (19) results in

$$\dot{V} = -\tilde{L}_{inc} L_{inc}^{-1} ye^2 - e \left[ \omega L_{inc}^{-1} (\tilde{E}_{bmf} - E_{bmf}) - i_p^* (1 - \tilde{L}_{inc} L_{inc}^{-1}) \right]$$

$$= -\tilde{L}_{inc} L_{inc}^{-1} ye^2 - e g(\cdot),$$

(22)

where

$$g(\cdot) = \omega L_{inc}^{-1} (\tilde{E}_{bmf} - E_{bmf}) - i_p^* (1 - \tilde{L}_{inc} L_{inc}^{-1}).$$

(23)

According to the Lyapunov stability condition, equation (22) needs to satisfy the following condition:

$$|\tilde{L}_{inc} L_{inc}^{-1} ye^2| > |e g(\cdot)|.$$  

(24)

Let the absolute value of current tracking error converge to error bound $e_p$, which is very close to zero, and inequality (24) can be rewritten as

$$\gamma > \frac{|g(\cdot)|}{\tilde{L}_{inc} L_{inc}^{-1} e_p}.$$  

(25)

It can be seen from equation (25) that the controller parameter $\gamma$ can be adjusted adaptively according to the value of $g(\cdot)$. By substituting (20) in (23), equation (23) can be rewritten as

$$g(\cdot) = -\left( \tilde{L}_{inc} + \Delta L_{inc} \right)^{-1} \left( \omega \Delta E_{bmf} + i_p^* \Delta L_{inc} \right).$$

(26)

Substituting (20) and (26) into (25), the condition in (25) can be further simplified as follows:

$$\gamma > \frac{\left| \tilde{L}_{inc} + \Delta L_{inc} \right|^{-1} \left( \omega \Delta E_{bmf} + i_p^* \Delta L_{inc} \right)}{\tilde{L}_{inc} L_{inc}^{-1} e_p}$$

$$= \frac{\omega \Delta E_{bmf} + i_p^* \Delta L_{inc}}{\tilde{L}_{inc} e_p}.$$  

(27)

In the practical applications, excessive feedback gain will result in oscillatory response near unaligned rotor position. Therefore, a feedback gain $\gamma$ varying according to (27) is adopted to achieve better current control performance. For a specific SRM, the values of $\Delta L_{inc}$ and $\Delta E_{bmf}$ are usually fixed and can be obtained by calculating the error between the real and simplified model parameters [27]. In this paper, $\Delta L_{inc}$ and $\Delta E_{bmf}$ are taken to be constant as 50% of the nominal value over the entire range of operation.

To achieve the fixed switching frequency control, PWM control is embedded in the proposed HPTTC algorithm. The duty cycle of the PWM control is described as

$$\lambda_p = \frac{u_p}{u_{dc}},$$

(28)

where $\lambda_p$ is the duty cycle of the $p$th phase.

3.3. Speed Controller Design. In this subsection, the objective is to explore a disturbance-observer-based speed controller to provide the total electromagnetic torque reference for the middle loop (torque loop). Assuming that the torque loop bandwidth is sufficiently large, the equation of motion is given by

$$\frac{d\omega}{dt} = -A_1 \omega + A_2 T_e^* - A_2 T_L,$$  

(29)

where $A_1 = k_\omega / J$, $A_2 = 1 / J$, $k_\omega$ is the friction coefficient, $J$ is the moment of inertia, and the load torque $T_L$ is taken as an unknown external disturbance.

Let the idea control law of speed loop be

$$T_e^* = A_3 \omega + \omega^* + k_1 e - A_2 T_L.$$  

(30)
where \( e = \omega^r - \omega \) is speed error and \( k_1 \) is a positive parameter.

Combining with (29) and (30), we get
\[
\dot{e} + k_1 e = 0. \tag{31}
\]

Equation (31) shows that the real speed \( \omega \) can track the speed command \( \omega^r \) asymptotically. However, the desired control law (30) cannot be implemented because the load disturbance \( TL \) of the SRM drive system is unknown. Since the external load disturbance of the motor system is constantly changing and has finite energy, the external load acting on the motor can be seen as the unknown, time-varying yet bounded signals with the finite change rates [28]. Therefore, the following assumption can be made. The load disturbance \( TL \) is unknown and time-varying yet bounded and there exists an unknown positive constant \( \sigma \) such that \( |TL| \leq \sigma \).

A disturbance observer can be derived to estimate the lamped disturbance as follows [29]:
\[
\begin{align*}
\dot{\eta} &= -\tau + \frac{1}{A_2} ( -A_1 + I ) \omega + IT_L^*, \\
\hat{TL} &= \tau - \frac{1}{A_2} \omega,
\end{align*}
\tag{32}
\]
where the disturbance observer gain \( l > 0 \).

Define the disturbance estimation error \( \eta \) of the observer as
\[
\eta = \hat{TL} - TL. \tag{33}
\]

From (29) and (32), we have
\[
\dot{\eta} = -\tau - \frac{1}{A_2} ( A_1 - I ) \omega + IT_L^* + \frac{1}{A_2} [ A_1 \omega - A_2 ( T_L^* - TL ) ]
= -\left( \tau - \frac{1}{A_2} \omega - TL \right) - \left( \hat{TL} - TL \right) . \tag{34}
\]
The time derivative of (33) along (34) is
\[
\dot{\eta} = -\dot{\eta} - \hat{TL} . \tag{35}
\]
Consider the Lyapunov function candidate as
\[
V = \frac{1}{2} \eta^2 . \tag{36}
\]
Using (34) and Young’s inequality, the time derivative of (36) is
\[
\dot{V} = \eta \dot{\eta} = \eta ( -\dot{\eta} - \hat{TL} ) \leq -\left( l - \frac{1}{2} \right) \eta^2 + \frac{1}{2} \sigma^2 = -aV + C , \tag{37}
\]
where \( a = l - (1/2) \), \( C = (1/2)\sigma^2 \), and \( l \) satisfies \( l > (1/2) \).

From (37), we have
\[
0 \leq V \leq \frac{C}{a} + \left[ V (0) - \frac{C}{a} \right] e^{-at} . \tag{38}
\]
Thus, \( V(t) \) is globally uniformly ultimately bounded. From (36) and (38), we have
\[
|\eta| \leq \frac{2C}{a} + 2 \left( V (0) - \frac{C}{a} \right) e^{-at} . \tag{39}
\]

Distinctly, \( \eta \) is globally uniformly ultimately bounded. For any positive constant \( k > \sqrt{2C/a} \), there exists a time constant \( \tau > 0 \) such that \( \eta \leq k \) for all \( t > \tau \). Therefore, \( \eta \) settles within \( \Omega_\eta = \{ \eta \in R || \eta | \leq k \} \). Since \( 2C/a = 2\sigma^2/(1 - l) \), it is obvious that \( \Omega_\eta \) can be made arbitrarily small by appropriately selecting \( l \) such that \( l > 1/2 \) is satisfied.

Therefore, the control law of speed loop is rewritten as follows
\[
T_e = A_2 \omega + \dot{\omega}^* + k_1 e + \hat{TL} . \tag{40}
\]

4. Simulation Results

In order to demonstrate the effectiveness of the HPITC method, the SRM drive system is simulated under different operating conditions using the Matlab/Simulink software. In this paper, a 1.5 kW three-phase 12/8-pole SRM is selected as an example. The measured flux-linkage \( \psi (i, \theta) \) and torque \( T (i, \theta) \) characteristics are shown in Figures 1(a) and 1(b), respectively. The sampling time in all simulations is 10 \( \mu \)s.

To quantitatively evaluate the performance of the proposed HPITC method, its performance indexes are defined as follows:
\[
T_{\text{err}} = \frac{T_{\text{e, max}} - T_{\text{e, min}}}{T_{\text{e, av}}} , \tag{41}
\]
\[
I_{\text{peak}} = \max_{k=1}^{N} \left| i_p (k) \right| , \tag{42}
\]
\[
P_{\text{loss}} = \frac{1}{N} \sum_{p=1}^{3} \sum_{k=1}^{N} i_p (k)^2 R , \tag{43}
\]
\[
I_{\text{RMSE}} = \frac{1}{N} \sum_{k=1}^{N} \left( i_p^* (k) - i_p (k) \right)^2 , \tag{44}
\]
where \( T_{\text{err}} \) represents the torque ripple factor and \( T_{\text{e, max}}, T_{\text{e, min}}, \) and \( T_{\text{e, av}} \) denote the maximum torque, the minimum torque, and the average torque, respectively. \( I_{\text{peak}} \) is the peak value of phase current. \( P_{\text{loss}} \) denotes the copper losses of SRM. \( I_{\text{RMSE}} \)
represents the root mean square error of phase current tracking. \( N \) is the number of samples.

In the first simulation, the comparisons between the proposed power TSF and traditional sinusoidal TSF are carried out to verify their steady-state performance. The proposed robust controller is applied to the current control loop. The turn-on angle \( \theta_{on} \) and turn-off angle \( \theta_{off} \) of the proposed TSF and traditional TSF are set to 22.5° and 37.5°, respectively. Figures 4(a) and 4(b) exhibit the simulation results of the proposed TSF and traditional linear TSF at low speed of 100 r/min, respectively. Figures 5(a) and 5(b) present the results of the proposed TSF and traditional sinusoidal TSF at high speed of 1500 r/min, respectively. The waveforms of the motor speed, total electromagnetic torque, phase torque, and stator phase current are presented in each subfigure. From Figures 4 and 5, we can see that both proposed TSF and conventional sinusoidal TSF can control the torque ripple within a certain range and have excellent torque ripple suppression ability. But the peak current of the proposed power TSF is much smaller than that of the traditional sinusoidal TSF. The current waveform of the proposed TSF is smoother than that of the conventional TSF. Moreover, Figures 6 and 7 show the comparison results of phase torque waveform between conventional and proposed TSFs at low speed of 100 r/min and high speed of 1500 r/min, respectively. As shown in Figures 6 and 7, the reference phase torque of the proposed TSF is much lower than that of the traditional sinusoidal TSF in the initial stage of inductance rise, which makes the peak current of the proposed TSF smaller than that of the traditional TSF. Correspondingly, in the second half of the inductance rise, the reference phase torque of the proposed TSF is higher than that of the conventional TSF to keep the total torque constant, but the phase current required to generate the same torque is smaller than that in the initial stage of inductance rise. As a result, the copper losses of the proposed TSF will be lower than those of the conventional TSF.

Further, the comparison of torque ripple factors \( T_{rf} \), peak current \( I_{peak} \), and copper losses \( P_{loss} \) between the proposed TSF and the traditional sinusoidal TSF is summarized in Table 2. From Table 2, it is seen that the torque ripple factors \( T_{rf} \) of the proposed TSF and the conventional sinusoidal TSF are very similar in a wide speed range. However, the peak current \( I_{peak} \) is reduced from 8.51 A to 7.48 A in conventional TSF to 5.42 A and 5.12 A at the speeds of 100 r/min and 1500 r/min, respectively. As a result, the conventional TSF has higher copper losses \( P_{loss} \) of 4.9887 and 5.5328 at speed commands of 100 and 1500 r/min, respectively. On the contrary, the copper losses \( P_{loss} \) of the proposed TSF are reduced to 4.2083 and 5.1164. Thus, the copper losses \( P_{loss} \) for the proposed TSF are lower by 15.6% and 10.1% compared with the traditional sinusoidal TSF. The simulation result demonstrates that the efficiency of the proposed TSF is higher than that of the traditional TSF.

In the next simulation, the comparative study of the proposed robust current controller and traditional hysteresis current controller is carried out at low speed of 100 r/min and high speed of 1500 r/min with load torque \( T_L = 5 \) Nm to test their current tracking capability. The switching frequency of the proposed robust current controller is set to 50 kHz. The current band of the hysteresis controller is set to ±0.1 A, and its switching frequency is between 1 kHz and 100 kHz depending on the speed and current. The waveforms of the robust current controller and hysteresis controller at low speed of 100 r/min are shown in Figures 8 and 9. Figures 10 and 11 exhibit the current and electromagnetic torque waveforms of the two controllers at a high speed of 1500 r/min, respectively. As shown in Figures 8 and 9, the proposed robust current controller has higher current ripple than the conventional hysteresis controller. However, the robust current controller has a lower current ripple than the conventional hysteresis controller. It can be seen from Figures 9 and 11 that the reduced current ripple by the proposed controller also results in reduced torque ripple. Table 3 gives a quantitative comparison of current tracking performance between the proposed controller and the hysteresis controller. According to Table 3, it can be seen that the RMSE of the hysteresis controller is 0.1554 and 0.8076, while that of the proposed controller is 0.01 kg m² and 0.005 Nms. As shown in Table 3, the current ripple is reduced from 0.3762 to 0.5484 in hysteresis controller to 0.2503 and 0.3995 at the speed of 100 r/min and 1500 r/min, respectively. These results indicate that the performance of the proposed current controller is superior to that of the traditional hysteresis controller.

In the final simulation, the tracking performance and robustness of the proposed speed controller are investigated by controlling the SRM to work in the states of starting, acceleration, deceleration, and constant speed. Figure 12 shows the speed command changing from 0 r/min to 1500 r/min with load torque of 5 Nm. It can be seen that the motor speed can track the reference speed very well during starting, accelerating, and decelerating. In the whole tracking process, the maximum dynamic tracking error of motor speed is less than 3 r/min, and the steady-state error is less than 1 r/min. Further, in order to test the robustness of the proposed controller, a step load torque is added to the SRM drive system, which is taken into account as a disturbance. In the simulations, the load torque \( T_f \) is step-changed from 1 Nm to 6 Nm at a constant speed of 1000 r/min. Moreover, the proposed disturbance-observer-based
speed control is compared with conventional PI controller. The parameters ($k_p$ and $k_i$) of PI speed controller are derived by pole placement [30]. The parameters $k_p$ and $k_i$ are 2.25 and 6.25, respectively. The responses to external load disturbance are illustrated in Figures 13(a) and 13(b) for the proposed disturbance-observer-based speed control and conventional PI methods, respectively. As shown in Figures 13(a) and 13(b), although the speed drops are

![Image of graphs showing speed control responses](image-url)
Figure 6: Comparison results of phase torque waveform between conventional and proposed TSFs at low speed of 100 r/min. (a) Reference phase torque. (b) Actual phase torque.

Figure 7: Comparison results of phase torque waveform between conventional and proposed TSFs at low speed of 1500 r/min. (a) Reference phase torque. (b) Actual phase torque.

Table 2: Comparison between the proposed TSF and sinusoidal TSF at different speeds.

| Index   | Sinusoidal TSF | Proposed TSF |
|---------|----------------|--------------|
|         | 100 r/min      | 1500 r/min   | 100 r/min   | 1500 r/min   |
| $T_{rf}$ | 0.2519         | 0.4839       | 0.2506      | 0.4390       |
| $I_{peak}$ | 8.51          | 7.48         | 5.42        | 5.12         |
| $P_{loss}$ | 4.9887        | 5.5328       | 4.2083      | 5.1164       |

Figure 8: Current simulation results under speed reference $\omega^* = 100$ r/min and load torque $T_L = 5$ Nm. (a) Hysteresis current controller. (b) Robust current controller.
Figure 9: Torque simulation waveforms under speed reference $\omega^* = 100 \text{ r/min}$ and load torque $T_L = 5 \text{ Nm}$.

Figure 10: Current simulation results under speed reference $\omega^* = 1500 \text{ r/min}$ and load torque $T_L = 5 \text{ Nm}$. (a) Hysteresis current controller. (b) Robust current controller.

Figure 11: Torque simulation waveforms under speed reference $\omega^* = 1500 \text{ r/min}$ and load torque $T_L = 5 \text{ Nm}$.

Table 3: Comparison between the proposed current controller and the hysteresis controller.

| Index  | Hysteresis current controller | Robust current controller |
|--------|------------------------------|---------------------------|
|        | 100 r/min                    | 1500 r/min                | 100 r/min                  | 1500 r/min                |
| $I_{\text{RMSE}}$ | 0.1554                      | 0.8076                    | 0.1410                      | 0.7745                    |
| $T_{\text{rr}}$  | 0.3762                       | 0.5484                    | 0.2503                      | 0.3995                    |
roughly similar in both cases, the proposed controller can be returned to the original speed reference more quickly than the conventional PI controller. It is also observed that the motor torque increases rapidly to maintain the load torque. In Figure 13, the maximum speed dips are 24 and 26 r/min with respect to the proposed disturbance-observer-based speed controller and conventional PI control. The speed error of the proposed method back to the range of ±1 r/min only through 0.055 s, and the speed error of the traditional PI method recovers to the range of ±1 r/min through 0.35 s. The proposed disturbance-observer-based speed controller exhibits stronger antidisturbance ability than the traditional PI algorithm. This is due to the fact that the external load disturbance is compensated accurately by the disturbance observer in the proposed control scheme.

In general, the HPITC algorithm has an excellent performance in efficiency, minimization of torque ripple, and robustness for the SRM drive system.

5. Experimental Results

In this section, the availability of the HPITC algorithm is further verified by experiments. A three-phase 12/8-pole 1.5 kW SRM is selected as the experimental prototype. The hardware schematic diagram of the SRM drive system is depicted in Figure 14(a). The photograph of the experimental platform is shown in Figure 14(b). The experimental platform consists of two parts: the mechanical system and electrical system. The mechanical system comprises the SRM prototype, torque and speed sensor, and magnetic particle brake, allowing adjusting the load torque. The electrical system mainly includes control unit (dSPACE DS1103), measurement device (digital oscilloscope), and asymmetric half-bridge power electronic converter. The sampling periods of speed and current control in the experimental study are set as 1 ms and 100 us, respectively. The following experiments were carried out to test the performance of the HPITC method.
5.1. Current Tracking Performance. The HPITC method was implemented at different speeds (200 r/min, 1000 r/min, and 1500 r/min) with load torque $T_L = 5\, \text{Nm}$ to confirm the current tracking performances. Figures 15(a)–15(c) show the experimental results of the speeds of 200 r/min, 1000 r/min, and 1500 r/min, respectively. The stator phase current $i_{ph}$ and its reference $i_{ph}^*$ are observed, respectively, through channel 1 (CH1) and channel 2 (CH2) of the oscilloscope. It is observed that the stator phase current can quickly track the phase current reference at different speeds. In Figure 15(a), the maximum value of the stator phase current is 7.8 A, and the copper loss $P_{\text{loss}}$ is 5.43. It can be calculated from Figure 15(b) that the peak current is 7.5 A and the copper loss $P_{\text{loss}}$ is 5.92. In Figure 15(c), the maximum value of the stator phase current is 7.9 A, and the copper loss $P_{\text{loss}}$ is 6.37. Figure 15 shows that the proposed controller has good current tracking performance.

5.2. Torque Ripple Suppression Performance. A speed command changing from 200 to 600 r/min with external load torque $T_L = 1\, \text{Nm}$ is implemented to test the torque ripple suppression capability of the HPITC method. Figure 16 presents the ability of the proposed HPITC strategy to minimize the torque ripple of SRM. As shown in Figure 16, the total torque (CH2) can always maintain a low torque ripple in both steady and dynamic states. According to equation (42), the torque ripple factor $T_{rf}$ can be calculated to be 0.5614 in this case. It can be noted that lower torque ripple ensures that the motor speed (CH1) can track command speed well. From these results, it can be concluded that the HPITC algorithm has an excellent capability for torque ripple suppression.

5.3. Robustness against External Load Disturbance. In order to test the antidisturbance ability of the HPITC algorithm, the experiments are carried out with a step change of external load torque during the steady operation. Figure 17 shows the experimental results. In this test, the motor speed is 400 r/min and the external load torque is changed from 0 Nm to 6 Nm. In Figure 17, the maximum speed drop of the SRM is 39 r/min, and it only takes 60 ms for the motor to recover to the initial speed reference. It is seen that under a load torque step variation the total electromagnetic torque follows the load torque value closely and always sustains low
torque ripples. These results show that the HPITC method has an excellent antidisturbance ability.

In sum, the experimental results verify the effectiveness of the proposed control scheme.

6. Conclusion

In this paper, a high-performance control technology was developed to reduce the torque ripple of the SRM drive system. Based on the nonlinear electromagnetic
characteristics of SRM, a more efficient TSF is used to ensure that the copper loss is reduced, while the torque ripple is minimized. Then, a robust current controller with variable feedback gain is designed to accurately track the current reference. The stability of the robust current controller has been proved by the Lyapunov stability analysis. Moreover, a novel speed controller is designed to improve the speed tracking accuracy and antidisturbance performance of the SRM drive system. The simulation and experimental results demonstrated that the HPITC method has the advantages of suppressing the torque ripple, improving system efficiency, and enhancing antidisturbance ability.

Data Availability
The data that support the findings of this study are included within the article.

Conflicts of Interest
The authors declare no conflicts of interest regarding the publication of this paper.

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