Sensorless Control of Induction Motors Based on Fractional-Order Linear Super-Twisting Sliding Mode Observer With Flux Linkage Compensation

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ABSTRACT A robust fractional-order linear super-twisting sliding mode observer (LSTSMO) for an induction motor speed sensorless system is proposed in this study to solve the problem of balance between sliding mode chattering and estimation accuracy. First, the stability of the linear super-twisting algorithm is proven with the Lyapunov function, and the convergence and gain coefficient range of the algorithm are analyzed. Next, the fractional integral sliding mode surface based on the stator current error term is established, and the linear super-twisting algorithm is introduced to design the fractional LSTSMO. In addition, a flux correction link is added to enhance the robust performance of the observer. The results of the simulations and experiments verify that the observer has high flux linkage estimation accuracy, which can enhance transient performance rapidity and chattering suppression ability simultaneously, thereby improving robustness against parameter drifts. Moreover, the designed observer improves the dynamic and steady state performance of the motor in the full-speed range.

INDEX TERMS Flux linkage estimation, sliding mode observer, induction motor, fractional order.

I. INTRODUCTION

High-performance induction motor (IM) speed control systems are widely used in the field of renewable energy and industries [1], [2]. In the vector control system of an IM, a photoelectric encoder can detect speed signals, but speed sensors have faults, such as poor contact and signal loss, which can affect the reliability of the speed control system [3], [4]. Therefore, speed sensorless control has attracted researchers’ attention.

In IM speed sensorless technology, a motor observer can be established based on the motor model to estimate flux linkage and speed [5]. Classical flux linkage and speed estimation methods include the extended Kalman filter (EKF), the model reference adaptive system (MRAS), and the sliding mode observer (SMO) [6]. The MARS relies heavily on the accuracy of model parameters, and its dynamic stability can only be adjusted to the best in a certain speed area, and observer flux calculation at low speeds is susceptible to stator resistance and pure integration. The EKF has a complex structure and high system computing power requirements [7], [8]. Meanwhile, the SMO has a simple structure and involves convenient calculation, which can improve a system’s anti-noise ability and sensitivity to parameter drifts. Thus, the SMO has become a research hotspot in speed sensorless technology [9], [10].

In a traditional SMO, the switching control of the switching function and inertia of the switch generate the problem of sliding mode chattering. Excessively high sliding mode chattering will increase a system’s energy consumption and affect estimation accuracy [11], [13]. An improved SMO is designed in [14] using the sigmoid function instead of the sign function to reduce the chattering problem in the rotor position estimation process. In [15], a new SMO is proposed to estimate rotor position and speed. In addition, the observer uses an s-shaped switching function to improve the sign function to estimate the back electromotive force (EMF). Compared with the sign function-improved SMO, the high-order...
SMO can effectively eliminate the chattering problem without affecting system robustness. In [16], a stator flux observer based on the super-twisting algorithm is designed for a direct torque control system to estimate stator flux, thereby improving the accuracy of the flux estimation and the robustness of the observer. In [17], an observer that can estimate the back EMF and identify the rotation speed is designed by combining the MRAS and super-twisting algorithm. To enhance a system’s ability to resist external disturbances, a speed loop disturbance performance of the observer, and the rotor flux is introduced to improve the estimation accuracy and anti-chattering and estimation performance in the SMO. Observer encounters a problem of pure integration in the flux linkage, and stability is not analyzed. Moreover, feasibility experiments have yet to be conducted for the observer.

A fractional-order linear super-twisting sliding mode observer (LSTSMO) is designed to estimate rotor flux and speed in an IM, focusing on the balance between sliding mode chattering and estimation performance in the SMO. First, the fractional-order integral sliding mode surface is combined with the linear super-twisting algorithm (LSTA) to form the LSTSMO. Next, the flux linkage compensation link is introduced to improve the estimation accuracy and anti-disturbance performance of the observer, and the rotor flux linkage and speed estimation link of the IM are designed. Finally, the designed observer is applied to an IM high-precision vector control system.

II. LSTA THEORY

High order sliding mode control usually needs information of sliding mode surface $S$ and its derivatives, e.g., $S$, which is not always known. The super-twisting algorithm avoids the use of $S$ and requires only $S$ information. Therefore, this algorithm is widely used in sliding mode control [21], [22]. To further enhance the convergence speed and robustness of the super-twisting algorithm, a linear super-twisting algorithm is designed, and its basic form is as follows

$$y = h_1 |x_1|^{0.5} \text{sign}(x_1) + h_3 x_1 + x_2$$
$$\dot{x}_2 = h_2 \text{sign}(x_1) + h_4 x_1.$$

Based on (1), $x_1$ is the required sliding mode surface, and $h_i (i = 1, 2, 3, 4)$ represents the sliding mode gains. The stability of $y$ can be proved with the aid of Lyapunov stability theory. Consider a Lyapunov function candidate $v$ as

$$v = 2h_2 |x_1| + h_4 x_1^2 + \frac{1}{2} \left[ h_1 |x_1|^{0.5} \text{sign}(x_1) + h_3 x_1 - x_2 \right]^2 = \beta^T P \beta,$$

where $\beta^T = [\beta_1, \beta_2, \beta_3] = [\sqrt{x_1} \text{sign}(x_1), x_1, x_2]$, and $P$ is a positive definite matrix,

$$P = \frac{1}{2} \begin{bmatrix} 3h_2 + 2h_1^2 & h_1 h_2 & -h_1 \\ h_1 h_2 & 3h_4 + h_3^2 & -h_3 \\ -h_1 & -h_3 & 1 \end{bmatrix}.$$

This equation satisfies

$$\lambda_{\min} \{P\} \|\beta\|^2_2 \leq V \leq \lambda_{\max} \{P\} \|\beta\|^2_2,$$

where $\|\beta\|^2_2 = x_1^2 + x_2^2 + x_3^2$ is the Euclidean norm of $\beta$, $\lambda_{\min} \{P\} > 0$ and $\lambda_{\max} \{P\} > 0$ are the minimum and maximum eigenvalues of the matrix $P$, respectively. Next, $\dot{V}$ becomes

$$\dot{V} = -\frac{1}{\sqrt{|x_1|}} \beta^T M_1 \beta - \beta^T M_2 \beta,$$

with

$$M_1 = \frac{h_1}{2} \begin{bmatrix} 2h_2 + h_1^2 & 0 & -h_1 \\ 0 & 2h_4 + 3h_3^2 & -h_3 \\ -h_1 & -h_3 & 1 \end{bmatrix},$$

and

$$M_2 = h_3 \begin{bmatrix} h_2 + h_1^2 & 0 & 0 \\ 0 & h_4 + h_3^2 & -0.5h_3 \\ 0 & -0.5h_3 & 1 \end{bmatrix}.$$

From (4), imposing $\dot{V} < 0$, $M_1 > 0$, $M_2 > 0$, and $h_i > 0$, $i = 1, 2, 3, 4$, the following inequalities about coefficients $h_i$, $i = 1, 2, 3, 4$ can be derived

$$h_2 + h_1^2 > 0$$

$$h_2 + 4h_4 + 8h_3^2 > h_1^2 h_3^2$$

According to (3) and (4), we have

$$\dot{V} \leq -\frac{1}{|x_1|^{1/2}} \lambda_{\min} \{M_1\} \|\beta\|^2_2 - \lambda_{\min} \{M_2\} \|\beta\|^2_2,$$

where $\lambda_{\min} \{M_1\}$ and $\lambda_{\min} \{M_2\}$ are the minimum eigenvalues of $M_1$ and $M_2$, respectively. Based on (3), the upper bound and lower bound of the vector $\|\beta\|^2_2$ can be derived as

$$|x_1|^{1/2} \leq \|\beta\|_2 \leq \frac{\nu^{1/2}}{\lambda_{\min} \{P\}},$$

Finally, with (3), (6), and (7), $\dot{V}$ satisfies (8).

$$\dot{V} \leq -K_1 \nu^{1/2} - K_2 \nu,$$

with $K_1 = \frac{\lambda_{\min} \{P\}}{\lambda_{\min} \{M_1\}}$, and $K_2 = \frac{\lambda_{\min} \{M_2\}}{\lambda_{\max} \{P\}}$. Since $K_1$ and $K_2$ are greater than 0, the system state $y$ converges to $x_1 = 0$ in finite time, and the convergence time cost $T \leq \frac{2}{K_1} \ln \frac{K_1 \nu^{1/2} (x_0) + K_2}{K_2 \nu}$. When the function converges near the equilibrium point, the linear term $K_2 \nu$ can expedite the converging process. This algorithm guarantees the stability of a system near and far from the equilibrium point.
III. DESIGN OF FRACTIONAL-ORDER LSTSMO

In the static $\alpha - \beta$ reference frame, an IM model with stator current and rotor flux linkages stated as can be as

\[
\begin{align*}
\frac{d\psi_{\alpha}}{dt} & = k_1 \left( \frac{1}{\eta_r} \psi_{\alpha} + \omega_r \psi_{\beta} - \frac{L_m}{\eta_r} i_{sa} \right) \\
\frac{d\psi_{\beta}}{dt} & = k_1 \left( \frac{1}{\eta_r} \psi_{\beta} - \omega_r \psi_{\alpha} - \frac{L_m}{\eta_r} i_{sb} \right) \\
\frac{d\psi_{\alpha}}{dt} & = k_1 \left( \frac{1}{\eta_r} \psi_{\alpha} - \omega_r \psi_{\beta} - \frac{L_m}{\eta_r} i_{sa} \right) \\
\frac{d\psi_{\beta}}{dt} & = k_1 \left( \frac{1}{\eta_r} \psi_{\beta} - \omega_r \psi_{\alpha} - \frac{L_m}{\eta_r} i_{sb} \right)
\end{align*}
\]  

(9)

where $k_1 = (k_3 L_m)/L_r$, $k_2 = R_s/(\sigma L_a)$, $k_3 = 1/(\sigma L_a)$, and $u_{sa}$, $u_{sb}$, $i_{sa}$, $i_{sb}$, $\omega_r$, and $\psi_{\alpha \beta}$ are the components of the stator voltage, stator current, and rotor flux linkage, respectively; $R_s$ is the rotor resistance, $R_r$ the stator resistance, $\omega_r$ the rotor electrical angular velocity, $\sigma = 1 - (L_2^2/L_0 L_a)$ the leakage coefficient, $p$ the differential operator, $L_s$ the stator inductance, $L_r$ the rotor inductance, $L_m$ the magnetizing inductance, $\eta_r = L_r/R_r$ the rotor electromagnetic time constant.

In (9), the current and flux linkage equations have the same coupling term, and the coupling term is replaced by the variable $\psi_t = [\psi_{\alpha \beta}]^T$, which can be described as

\[
\begin{bmatrix}
\psi_{\alpha} \\
\psi_{\beta}
\end{bmatrix}
= \begin{bmatrix}
\frac{1}{\eta_r} & \frac{1}{\eta_r} \\
\frac{1}{\eta_r} & \frac{1}{\eta_r}
\end{bmatrix}
\begin{bmatrix}
\frac{\psi_{\alpha}}{\eta_r} \\
\frac{\psi_{\beta}}{\eta_r}
\end{bmatrix}
- \frac{L_m}{\eta_r} \begin{bmatrix}
i_{sa} \\
i_{sb}
\end{bmatrix}.
\]  

(10)

By substituting (10) into (9), the matrix equation of the stator current and rotor flux estimation equations can be presented as

\[
\begin{align*}
\frac{d\psi_{\alpha}}{dt} & = k_1 \left( \psi_{\alpha} \psi_{\beta} \right) - k_2 \left( i_{sa} \right) + k_3 \left( u_{sa} \right) \\
\frac{d\psi_{\beta}}{dt} & = k_1 \left( \psi_{\alpha} \psi_{\beta} \right) - k_2 \left( i_{sb} \right) + k_3 \left( u_{sb} \right)
\end{align*}
\]

(11)

Based on the stator current error term, by increasing the fractional-order integral term to enhance the flexibility of the sliding mode surface, the chattering problem can be further suppressed, and the sliding mode convergence speed can be accelerated [20], [23]. The observer sliding mode surface $S_t$ are defined as

\[
S_t = \begin{bmatrix}
S_{\alpha a} \\
S_{\alpha b}
\end{bmatrix} = \begin{bmatrix}
c_1 i_{sa} + c_2 D^{-\alpha} i_{sa} \\
c_1 i_{sb} + c_2 D^{-\alpha} i_{sb}
\end{bmatrix},
\]

(13)

where $\begin{bmatrix} i_{sa} \ i_{sb} \end{bmatrix}^T = \begin{bmatrix} i_{sa} \ i_{sb} \ i_{sb} \end{bmatrix}^T$, and $c_1 > 0$, $c_2 > 0$, $D^{-\alpha}$ is the fractional integration factor, $0 < \alpha < 1$ [24], [25]. The Lyapunov function $V = S_t^T S_t/2$ greater than 0 is selected to verify the stability of the sliding mode surface; thus, $dV/dr$ can be derived as

\[
\dot{V} = S_t^T S_t = S_t^T \left\{ c_1 k_1 \begin{bmatrix} \psi_{\alpha} \\ \psi_{\beta}
\end{bmatrix} + c_1 k_1 \begin{bmatrix} A \\ B
\end{bmatrix} + \left( c_2 2^{-\alpha + 1} - c_1 k_2 \right) \begin{bmatrix} \dot{i}_{sa} \\ \dot{i}_{sb}
\end{bmatrix} \right\}.
\]

(14)

where $A = -\eta_r \psi_{\alpha} \omega_r \psi_{\beta} + \eta_r L_m i_{sa}$, and $B = \omega_r \psi_{\alpha} \eta_r \psi_{\beta} + \eta_r L_m i_{sb}$. According to (1), $y$ and $S_t$ satisfy the following inequalities

\[
\begin{align*}
S_{\alpha a}^* y_a & = S_{\alpha a}^* \left[ \frac{1}{2} \frac{1}{\eta_r} \left| \psi_{\alpha} \right|^2 \left| \psi_{\beta} \right|^2 \right] + \frac{1}{2} \frac{1}{\eta_r} \left| \psi_{\alpha} \right|^2 \left| \psi_{\beta} \right|^2 \\
S_{\alpha b}^* y_b & = S_{\alpha b}^* \left[ \frac{1}{2} \frac{1}{\eta_r} \left| \psi_{\alpha} \right|^2 \left| \psi_{\beta} \right|^2 \right] + \frac{1}{2} \frac{1}{\eta_r} \left| \psi_{\alpha} \right|^2 \left| \psi_{\beta} \right|^2
\end{align*}
\]

(15)

Based on (15), it is assumed that

\[
\begin{align*}
S_{\alpha a}^* & \geq 0 \\
S_{\alpha b}^* & \geq 0
\end{align*}
\]

(16)

Substituting (16) into (14) yields

\[
\dot{V} = S_t^T S_t < 0.
\]

(19)

From the above deduction, $S_t$ meets the stability conditions. According to (18), the dynamics of the rotor flux linkage estimate can be derived as follows.

\[
\begin{align*}
\frac{d\dot{\psi}_{\alpha}}{dt} & = - \frac{1}{c_1 k_1} \left\{ \left( c_2 k_2 - c_2 D^{-\alpha + 1} \right) \left[ \frac{i_{sa}}{\eta_r} \right] - \frac{\dot{y}_a}{\eta_r} \right\} \\
\frac{d\dot{\psi}_{\beta}}{dt} & = - \frac{1}{c_1 k_1} \left\{ \left( c_2 k_2 - c_2 D^{-\alpha + 1} \right) \left[ \frac{i_{sb}}{\eta_r} \right] - \frac{\dot{y}_b}{\eta_r} \right\}
\end{align*}
\]

(18)

Substituting (18) into (17) yields

\[
\dot{V} = S_t^T S_t < 0.
\]

(20)

where $\dot{\psi}_{\alpha \beta}$ and $\dot{\psi}_{\beta}$ are the estimated values of flux linkage before flux correction. Then, the rotor flux angle $\theta$ used in vector control is obtained as follows.

\[
\theta = \arctan \left( \frac{\dot{\psi}_{\beta}}{\dot{\psi}_{\alpha \beta}} \right).
\]

(21)
In (20), problems such as motor parameter deviation and DC offset affect observer’s estimation accuracy when passing through the pure integration term [21]. Therefore, a correction term can be added to the flux observer dynamics to improve its performance. The new fractional-order LSTA flux observer equation proposed as follows.

\[
\begin{align*}
\dot{\psi}_{ra} &= -\int (\psi_{ra} + H_1)\,dt \\
\dot{\psi}_{rb} &= -\int (\psi_{rb} + H_2)\,dt
\end{align*}
\]  

where \( H_1 \) and \( H_2 \) are flux correction items, and the correction term can be defined as

\[
\begin{align*}
H_1 &= K^\ast (\hat{\phi}_{ra} - \hat{\phi}_{ra}) \\
H_2 &= K^\ast (\hat{\phi}_{rb} - \hat{\phi}_{rb})
\end{align*}
\]  

where \( K^\ast \) is a positive constant. \( \hat{\phi}_{ra} \) and \( \hat{\phi}_{rb} \) can be obtained through the synchronous rotation orthogonal coordinate system, as follows

\[
\begin{align*}
\frac{d\hat{\phi}_r}{dt} &= \left( \frac{R_s L_m}{L_r} \right) i_d - \frac{R_r}{L_r} \hat{\phi}_r, \\
\hat{\phi}_{ra} &= \hat{\phi}_r \cos \theta, \\
\hat{\phi}_{rb} &= \hat{\phi}_r \sin \theta,
\end{align*}
\]

where \( i_d \) represents the d-axis flux linkage current. The combination of (10) and (22) can be rewritten as

\[
\begin{bmatrix}
\hat{\psi}_{rb} \\
\hat{\psi}_{ra}
\end{bmatrix} =
\begin{bmatrix}
\eta \hat{\psi}_{ra} \hat{\phi}_{rb} + \alpha^2 \hat{\psi}_{rb}^2 - \frac{L_m i_{sb} \hat{\psi}_{ra}}{\eta_r} \\
\eta \hat{\psi}_{ra} \hat{\phi}_{rb} + \alpha^2 \hat{\psi}_{rb}^2 - \frac{L_m i_{sb} \hat{\psi}_{ra}}{\eta_r}
\end{bmatrix} \cdot
\]

where (26), the speed estimate \( \hat{\omega}_r \) can be calculated as

\[
\hat{\omega}_r = \frac{\hat{\phi}_{rb} \psi_{ra} - \hat{\phi}_{ra} \psi_{rb} - \frac{L_m}{\eta_r} (\hat{i}_{sb} \hat{\psi}_{ra} - \hat{i}_{sb} \hat{\psi}_{rb})}{\hat{\phi}_{rb}^2 + \hat{\phi}_{ra}^2}.
\]  

The fractional-order LSTSMO does not use speed information during flux linkage estimation. Moreover, the observer avoids using the low-pass filter, which reduces the complexity of the system.

**IV. SIMULATION RESULTS**

To verify the effectiveness of the designed observer, it is compared with the STSMO. The IM vector control system is shown in Figure 2, which is built using MATLAB/Simulink. The inverter uses space-vector pulse width modulation to generate three phase voltages, and the current loop is closed with feedback from the current sensors while the speed loop is closed by the estimated speed from (27). The motor parameters used in simulation and experiment are listed in Table 1.

---

**TABLE 1. Detailed parameters for IM.**

| Parameter name          | Symbols | Parameter value |
|-------------------------|---------|-----------------|
| Stator resistance (Ω)   | \( R_s \) | 1.725           |
| Rotor resistance (Ω)    | \( R_r \) | 2.310           |
| Stator inductance (H)   | \( L_s \) | 0.240           |
| Rotor inductance (H)    | \( L_r \) | 0.240           |
| Mutual inductance (H)   | \( L_m \) | 0.228           |
| Rotor flux (Wb)         | \( \phi_r \) | 0.800           |

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**A. PARAMETER SELECTION AND OBSERVER ANALYSIS**

The parameters that need to be adjusted in fractional order LSTSMO are \( c_1, c_2, h_1, h_4, \) and \( K. \) \( c_1 \) and \( c_2 \) determine the sliding mode surface, \( K \) is the feedback gain, among which, \( h_1 — h_4 \) need to satisfy the inequalities in (5). By selecting different values for each parameter for parameter analysis, and then comparing the rotational speed with the estimated value of the flux linkage, a feasible adjustable parameter value can be finally obtained.

We are going to take one example to show the selection of parameters’ effect on observer performance. Consider the linear gain \( h_3, \) and let \( h_3 \) take different values (i.e., 0, 85, and 106.3), the consequent speed chattering and flux linkage estimation error values are shown in Figure 3. In Figure 3, other parameters are kept invariant: \( c_1 = 0.02, c_2 = 4.1, \)
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When $h_3 = 0$, the estimation error of the speed chattering and flux linkage is 1.7 rpm and $1 \times 10^{-4}$ Wb, respectively, and the performance of the observer is poor. As shown in Figure 3, as $h_3$ increases, the estimated speed chattering and flux linkage error gradually decrease, which can suppress speed chattering while enhancing the estimation accuracy of the observer. Owing to the existence of the parameter rule of (5), an excessively large $h_3$ value will cause system instability. Table 2 shows the specific values of the speed chattering and flux estimation error when $h_3$ changes.

The influence of the adjustable parameters in the designed observer on system performance is analyzed from the three following aspects.

1) For the adjustable parameters of the fractional integral sliding mode surface, when the value of $c_1$ is small, the response speed of the proportional link in the sliding mode surface will be slow, thereby reducing estimation performance. When the value of $c_2$ is too large, the integral link will lag, but the $\alpha$ will increase the flexibility of the sliding mode surface. By reducing $\alpha$, the integral saturation problem and increase in speed deviation can be well suppressed.

2) For the LSTA, $h_1$ and $h_2$ are directly related to the sign function. Attention should be paid to the balance between speed chattering and estimation accuracy when selecting values for $h_1$ and $h_2$. The addition of the linear links of $h_3$ and $h_4$ to the algorithm enhances the response speed of the observer and solves the balance problem in the selection of the $h_1$ and $h_2$ parameters.

3) According to (23), by increasing the flux linkage compensation coefficient $K$, the flux correction time can be shortened, and the problem caused by the pure integral link in (20) can be solved.

### B. MOTOR HIGH-SPEED OPERATION SIMULATION

In order to show the effectiveness of the proposed fractional LSTSMO in alleviating speed chattering and facilitating transient performance, the transient waveforms of the fractional-order LSTSMO are compared to that of STSMO, as shown in Figure 4. The motor accelerates from 100 rpm to 1500 rpm in 0.3 s. Figure 4 shows that the estimation speed of the STSMO reaches the given value at 1.261 s, and the speed chatter at the steady state reaches 1.5 rpm. The estimated rotational speed of the fractional-order LSTSMO reaches the given value at 1.25 s, and the steady state running speed chattering range is 0.1 rpm, thereby demonstrating improved transient performance and rotational speed chattering suppression ability.
C. MOTOR VARIABLE LOAD SIMULATION OPERATION

The simulated operation curve of the variable load torque of the IM is represented in Figure 5. The motor runs smoothly at 1000 rpm, and a load torque of 7.5 Nm is added in 1.5 s, and the motor maintains a no-load operation for 1.8 s. Figure 5 shows that when the load torque increases or decreases, the estimated rotational speed of the STSMO changes by 3 rpm, and the estimated chattering of the rotational speed is ±2 rpm. However, the rotational speed of the fractional LSTSMO changes by 1 rpm, and the rotational speed estimation error is ±0.1 rpm.

D. SIMULATION OPERATION WHEN RR DRIFT

According to the IM speed control system, rotor resistance $R_r$ is susceptible to temperature and the skin effect, and the rotor resistance deviation can seriously affect the estimation accuracy of the observer [26]. To verify the effect of $R_r$ on the observer, rotor resistance is drifted to 3.01 Ω for 2 s, and the simulation results are shown in Figure 6.

Figure 6 demonstrates the speed and flux linkage curve when rotor resistance drift. The speed chatter of the STSMO is 2.2 rpm, and the rotor flux linkage amplitude error is 0.3%. The speed chatter of the fractional-order LSTSMO is 0.1 rpm, the rotor flux linkage amplitude error is 0.015%, and the flux linkage accuracy is significantly better.

E. POSITIVE INVERSION SIMULATION RUNNING AT LOW SPEED

To verify the effectiveness of the fractional LSTSMO in the low-speed state of the motor, motor speed is changed from 50 rpm to −50 rpm in 1 s, and the simulation experiment of the
positive inversion rotation of the motor is shown in Figure 7. The speed chattering of the STSMO is 1.5 rpm during the positive inversion rotation, and the speed fluctuation is relatively large. The speed chattering of the fractional LSTSMO is 0.06 rpm, and the response speed is fast during the transient operation.

V. EXPERIMENTAL RESULTS

The experimental platform of the IM speed control system based on dSPACE is shown in Figure 8. The experimental motor parameters are as follows: rated voltage $U_s$ is 380V, the number of pole-pairs is 2, the rated speed is 1450 rpm. The experimental motor uses a magnetic powder brake to adjust the load torque, and the actual speed is measured with an E6B2-CWZ6C photoelectric encoder. The estimation curve of the motor speed control system is obtained with dSPACE and an oscilloscope.

A. MOTOR RUNNING AT FULL SPEED

Speed estimation based on the STSMO and fractional-order LSTSMO at full speed is represented in Figure 9. The given speed of the motor is set to six stages, that is, 300 rpm, 700 rpm, 1410 rpm, 800 rpm, 500 rpm, and 0 rpm. The estimated speed $\hat{n}$ of the two observers can accurately track the actual speed $n$ when the motor is running in the steady state. The fractional-order LSTSMO converges quickly in the transient state, which reduces the time for the estimated speed to reach the given speed.

B. POSITIVE INVERSION EXPERIMENT VERIFICATION

Figure 10 shows the performance of the two observers when the motor is running from 700 rpm to –700 rpm. The STSMO
observes the oscillation in the speed of the motor when it crosses the zero point. The estimation error $\Delta n$ of the speed of the observer reaches 4 rpm, and the A-phase stator current $i_a$ of the motor is contaminated by noises. The estimation speed of the fractional-order LSTSMO motor can run smoothly when crossing zero, the maximum value of the speed error is 1.5 rpm during speed transients, and $i_a$ is close to the sine wave.

C. MOTOR EXPERIMENT OF STATOR RESISTANCE

In the low speed state of the motor, the drift of the stator resistance $R_s$ will cause system instability and oscillation [27], [28]. To verify the robustness of the observer against stator resistance deviation, a variable stator resistance experiment is conducted. Figure 11 presents the estimated speed and stator current curve when stator resistance drift. The motor first maintains stable operation at 20 rpm; when the $R_s$ drifts, the maximum fluctuations of the rotational speed of the STSMO and fractional LSTSMO reach 8.4 rpm and 2.6 rpm, respectively. Compared with the STSMO, the stator current fluctuation of the fractional-order LSTSMO is smaller, and the waveform is better. Thus, the fractional-order LSTSMO is highly robust against stator resistance.

D. EXPERIMENTAL VERIFICATION AT VARIABLE LOAD TORQUE

Figure 12 illustrates the performance of the motor speed response, speed error, and stator current $i_a$ when the motor changes from a no-load to a rated-load torque, and the motor keeps running at 100 rpm. When the load changes, the estimated speed of the STSMO drops by 8.3 rpm, and the maximum speed error is 6 rpm. In the same condition, the speed of the fractional-order LSTSMO drops by 7 rpm, and the maximum error of the speed is 3 rpm. Furthermore, the estimation accuracy is high when the motor is running in a steady state, and $i_a$ is smooth.

As shown in Figure 13, to verify the variable load operation capability of the motor at a low speed, the motor is maintained...
at 15 rpm, and an experiment to apply and remove load torque is performed. Figure 13 shows that the estimated speed of the two observers will return to the given value after the load torque change. The estimated speed of the STSMO is reduced by 6.6 rpm, and the maximum error of the speed estimation is 6 rpm. However, the fractional LSTSMO is reduced by 5 rpm, the maximum error of the speed estimation is 2 rpm, and the fluctuation of the stator current \( i_a \) waveform is minor to Figure 10. Both observers can operate stably under variable load operations at low motor speeds, but the estimation performance of the fractional-order LSTSMO is better.

VI. CONCLUSION

In this study, an SMO combining a fractional-order integral sliding mode surface and LSTA is designed in the \( \alpha-\beta \) reference frame. Moreover, the performance of the observer is improved by adding the flux compensation term. The proposed scheme is applied to the speed sensorless vector control system of an IM. The results of the simulations and experiments verify the feasibility and effectiveness of the scheme. The main conclusions are as follows:

1) The designed observer provides chatter reduction ability of the motor at full speed, and the response speed of the motor during transient acceleration and deceleration is improved.

2) The observer eliminates the low-pass filter and improves the estimation accuracy of the flux linkage and speed. Moreover, it is robust against motor parameter drifts.

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