Research on the SCVM Flux Error MRAS Observer of Asynchronous Motor Based on SMC

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Abstract: Aiming at the sensorless control system of asynchronous motor, a MRAS observer method for asynchronous motor SCVM flux error based on sliding mode control (SMC) is proposed. This method uses the SCVM voltage model flux error to form a MRAS observer to identify the speed and estimate the rotor flux, and a sliding mode speed controller is designed to replace the traditional PI speed controller. The algorithm can reduce the overshoot of the dynamic response of the system speed at medium and high speeds, shorten the time to enter the steady state value, and enhance the robustness of the system. Simulation results show the effectiveness of the algorithm.

1. Introduction
Asynchronous motors have the characteristics of low manufacturing cost and high operating reliability, which occupy an important position in the AC motor frequency conversion speed control system. Speed closed-loop control is indispensable for high-performance asynchronous motor variable frequency speed control systems, but the installation of speed sensors such as photoelectric encoders will not only increase the cost of the system, but also be difficult to maintain due to the impact of the installation environment, so the speed sensorless technology has become the key technology of asynchronous motor control[1].

In recent years, with the continuous research on scholars at home and abroad, speed sensorless technology is divided into two categories: open-loop flux observer and closed-loop flux observer. The direct calculation method is used to estimate the speed, which is obtained by making the difference between the synchronous speed and the slip. It is simple and fast. The disadvantages are that the accuracy of the estimated speed cannot be guaranteed, it depends on the motor parameters, and the anti-interference ability is poor. The anti-interference ability of model reference adaptive system (MRAS) has been improved, but the reference model and the adjustable model are related to the motor parameters, and must be used together with parameter identification technology, which is not conducive to wide application [2]. Sliding mode observer (SMO) has good dynamic response and strong robustness, but the existing chattering will affect the steady-state accuracy of the system [3]. Extended Kalman filter (EKF) is computationally intensive and not suitable for practical application [4]. The full-order flux observer (AFO) uses the motor itself as a reference model, observes the stator current and rotor flux of the asynchronous motor according to the state estimation equation, and adds an error between the actual motor current and the observed current as a correction term in this equation. The matrix of correction terms is used to improve the dynamic performance and steady-state performance of the observer, thereby improving the accuracy of the estimation [5]. However, a single correction term matrix cannot guarantee

the requirements of all operating states, involving multi-matrix design increases the difficulty and requires a large amount of calculation.

The SCVM voltage model is simple, and the speed estimation accuracy is not high. In order to improve the accuracy of the SCVM voltage model to estimate the speed, a MRAS observer based on the flux error of the SCVM voltage model is designed. In order to improve the high-speed dynamic performance and steady-state performance of the MRAS observer, a sliding mode speed controller [6] is designed to replace the traditional PI speed controller. The effectiveness of this method can be seen by simulating the speed follow-up, speed abrupt change and load capacity through the model built.

2. Asynchronous motor model

According to the inverse τ steady-state equivalent circuit shown in Figure 1, the mathematical model under the two-phase static coordinate system can be established as shown in equations (1) and (2)

$$L_\sigma \frac{d i_s^s}{dt} = v_s^s - R_s i_s^s - \frac{d \psi_R^s}{dt}$$  \hspace{1cm} (1)

$$\frac{d \psi_R^s}{dt} = R_s i_s^s - (a - j \omega) \psi_R^s$$  \hspace{1cm} (2)

\(a = R_s / L_s = R_R / L_R\); \(a\) is rotor time constant; \(v_s^s\) is stator voltage; \(i_s^s\) is stator current; \(R_s\) is stator resistance; \(R_R\) is rotor resistance; \(\psi_R^s\) is rotor flux; \(\omega\) is rotor speed; \(L_M\) is magnetizing inductance; \(L_\sigma\) is leakage inductance.

The mathematical model of the two-phase rotating coordinate system obtained by the equations (1) and (2) through park transformation is shown in (3) and (4):

$$L_\sigma \frac{d i_s}{dt} = v_s - (R_s + j \omega L_\sigma) i_s - \left(j \omega \psi_R + \frac{d \psi_R}{dt}\right)$$  \hspace{1cm} (3)

$$\frac{d \psi_R}{dt} = R_s i_s - \left[a + j (\omega - \omega)\right] \psi_R$$  \hspace{1cm} (4)

3. Design of SCVM flux error MRAS observer

3.1. SCVM voltage model

In order to derive the SCVM voltage model, first establish a traditional VM voltage model, and replace the actual value of the motor parameters with the model estimation parameters in equation (1), the direct field orientation (DFO) implementation formulas such as equations (5) and (6) can be obtained as:

$$\dot{E}_R^s = v_R^s - \dot{R}_s i_s^s - \dot{L}_\sigma \frac{d i_s^s}{dt}$$  \hspace{1cm} (5)
\[
\frac{d\hat{\psi}_R}{dt} = \hat{E}^s
\]  
(6)

where \( \hat{\psi}_R = \hat{\psi}_\alpha + j\hat{\psi}_\beta \).

When analyzing flux, it is generally assumed that the response time of the current loop is shorter than
the response time of the flux, i.e., \( \frac{di}{dt} = 0 \). Use park changes with \( \hat{E}^s = e^{j\theta} \hat{E} \) and \( \hat{\psi}_R = e^{j\theta} \hat{\psi}_R \). Change equations (5) and (6) to a two-phase rotating coordinate system to obtain the realization formula of indirect magnetic field orientation (IFO) as shown in (7) and (8):

\[
\frac{d\hat{\psi}_R}{dt} + j \frac{d\theta}{dt} \hat{\psi}_R = \hat{E}
\]  
(7)

\[
\hat{E} = v_i - (\hat{R}_s + j\omega_l \hat{L}_s) i_i
\]  
(8)

Expand (7) and (8) to get:

\[
\frac{d\hat{\psi}_R}{dt} = \hat{E}_d, \quad \frac{d\theta}{dt} = \omega_l = \frac{\hat{E}_q}{\hat{\psi}_R}
\]  
(9)

where \( \hat{E}_d = v_i - \hat{R}_s i_i + \alpha_l \hat{L}_q i_q \), \( i_i = i_d + j i_q \), \( \hat{E}_q = v_i - \hat{R}_s i_q + \alpha_l \hat{L}_d i_d \).

The above traditional VM voltage model cannot be stable in all the operating frequency range of the motor, and the formula (9) can be changed to obtain the SCVM voltage model as shown in equation (10):

\[
\frac{d\hat{\psi}_R}{dt} = \gamma \hat{E}_d, \quad \frac{d\theta}{dt} = \omega_l = \frac{\hat{E}_q - \lambda \hat{E}_d}{\hat{\psi}_R}
\]  
(10)

3.2. stability analysis

For simple analysis, the actual parameters of the motor are used to estimate the parameters of the model. Substitute equation (4) into equation (3) with \( i_d = \psi_{\text{ref}} / L_M \).

\[
\hat{E}_d = a (\psi_{\text{ref}} - \psi_d) - \omega_l \psi_q
\]  
(11)

\[
\hat{E}_q = R_s i_q - a \psi_q + \omega \psi_d
\]  
(12)

Define \( x = [\psi_d \; \psi_q \; \psi_R]^T \); \( x^* = [\psi_{\text{ref}} \; 0 \; \psi_R]^T \). Bring equations (11) and (12) into equation (10):

\[
\dot{x} = f(x)
\]  
(13)

where \( f(x) = [f_1(x), f_2(x), f_3(x)]^T \)

\[
f_1(x) = a \left( 1 - \frac{\lambda \psi_R}{\psi_R} \right) (\psi_{\text{ref}} - \psi_d) + (\lambda, \omega_l - a) \frac{\psi_d}{\psi_R} + (\psi_{\text{ref}} - \psi_R + R_s i_q) \frac{\psi_q}{\psi_R}
\]

\[
f_2(x) = (R_s i_q + \omega_l \psi_d) \left( 1 - \frac{\psi_d}{\psi_R} \right) + \frac{\lambda a \psi_d}{\psi_R} (\psi_{\text{ref}} - \psi_d) - \left[ a + \frac{\psi_d}{\psi_R} (\lambda, \omega_l - a) \right] \psi_q
\]

\[
f_3(x) = \gamma \left[ a (\psi_{\text{ref}} - \psi_d) - \omega_l \psi_q \right]
\]

After linearization, we get:

\[
\dot{x} = A(x - x^*)
\]  
(14)
where

\[
A = \begin{bmatrix}
\frac{\partial f_1(x)}{\partial \psi_d} & \frac{\partial f_1(x)}{\partial \psi_q} & \frac{\partial f_1(x)}{\partial \psi_R} \\
\frac{\partial f_2(x)}{\partial \psi_d} & \frac{\partial f_2(x)}{\partial \psi_q} & \frac{\partial f_2(x)}{\partial \psi_R} \\
\frac{\partial f_3(x)}{\partial \psi_d} & \frac{\partial f_3(x)}{\partial \psi_q} & \frac{\partial f_3(x)}{\partial \psi_R}
\end{bmatrix}_{x=x_0^*}
\]

The characteristic polynomial is obtained as shown in equation (15), and the stability is determined by the Rolls criterion.

\[
\det(sI - A) = s^3 + c_2 s^2 + c_1 s + c_0
\]

where \(c_2 = a + \lambda \omega_r\), \(c_0 = a \gamma \omega^2\), \(c_1 = \omega_0 \left[ \omega_1 + \lambda \omega_r + (\gamma - 1) \omega_r \right]\)

\[\text{(15)}\]

3.3. Flux error MRAS observer

When the direct magnetic field orientation (DFO) of the SCVM voltage model is implemented with \(\gamma = 1 + \lambda \omega^2\). The accuracy of the rotation speed estimation at this time is not high, and there are large burrs. In order to improve its accuracy, using the flux error, the following speed adaptation rate is obtained as shown in equations (16) and (17).

\[
\dot{\omega}_{\text{MRAS}} = \left( \frac{k_p}{s} + \frac{k_i}{s} \right) \psi_{\text{qMRAS}} \quad \text{(16)}
\]

\[
\psi_{\text{qMRAS}} = \psi_{\text{SCVM}} \cos \theta_{\text{MRAS}} - \psi_{\text{aSCVM}} \sin \theta_{\text{MRAS}} \quad \text{(17)}
\]

4. Design of sliding mode speed controller

To compare with the traditional PI speed controller, the current state of the system is set as a deviation, that is, the difference between the given speed value and the estimated speed value. Define the sliding mode surface function:

\[
z_{\omega} = \omega_{\hat{}} - \omega_r \quad \text{(18)}
\]

Compared with the traditional PI speed controller, the first two items remain unchanged, and the sliding mode speed controller is designed as follows:

\[
i_{q \hat{}} = K_i z_{\omega} + K_1 \int z_{\omega} dt + K_2 \text{sgn}(z_{\omega}) \quad \text{(19)}
\]

As long as the rational design can satisfy Lyapunov’s stability theorem, the system is gradually stable, and the estimated speed is consistent with the given speed.

5. System simulation and result analysis

Build a model in MATLAB/Simulink environment to simulate and verify the designed sliding mode speed controller. The simulation model uses vector control based on rotor magnetic field orientation, and the asynchronous motor speed sensorless vector control block diagram is shown in Figure 2.

The given speed is 750r/min, that is, 0.5pu, and the pre-excitation setting is performed 1s before. After 1s, the motor starts without load, and the load torque is 7.35N·m suddenly at 2s. The dynamic response of the system speed based on traditional PI and sliding mode PI control is shown in Figure 3 and Figure 4.
It can be seen from Figure 3 and Figure 4 that the speed overshoot of traditional PI control is 34.9%, the speed overshoot of sliding mode PI control is 32.8%, and the overshoot is reduced. The time for the latter to enter the steady state value is 1.35s, and the time for the former is 1.6s, so there is a certain amount of overshoot. When half load is suddenly added at 2s, the speed drop of sliding mode PI control is smaller than that of traditional PI control, and it returns to the steady state value at 2.26s. The time is shorter than that of traditional PI control, and it has a strong load capacity.

The given speed is 1200r/min, which is 0.8pu, and the pre-excitation setting is performed 1s before. After 1s, the motor starts without load, and the load torque is 7.35N·m suddenly at 2s. The dynamic response of the system speed based on traditional PI and sliding mode PI control is shown in Figure 5 and Figure 6.

It can be seen from Figure 5 and Figure 6 that the speed overshoot of traditional PI control is 22.6%, the speed overshoot of sliding mode PI control is 21.1%, and the overshoot is reduced. The time for the latter to enter the steady state value is 1.4s, and the time for the former is 1.6s, so there is a certain overshoot. Sudden at 2s with half load, the speed drop of sliding mode PI control is smaller than that of traditional PI control, and it returns to the steady state value at 2.26s. The time is shorter than that of traditional PI control, and it has a strong load capacity.

The given speed is 1200r/min and -750r/min, namely 0.8pu and -0.5pu. Set the pre-excitation before 1s. After 1s, the motor starts without load, and the load torque is 7.35N·m suddenly at 2s. The speed changes suddenly at 3s, and it is switched from forward to reverse. It has been running with half load after 2s. The dynamic response of system speed based on traditional PI and sliding mode PI control is shown in Figure 7 and Figure 8.
As can be seen from Figures 7 and 8, the speed response before 3s, the sliding mode PI control, whether it is the overshoot or the time to enter the steady state value, are less than the traditional PI control. The time for sliding mode PI control to enter -0.5pu is 3.44s, and the time for traditional PI control to enter -0.5pu is 3.7s, indicating that the method has good adaptive ability.

6. Conclusion
Aiming at the problems of low speed accuracy, poor mid-to-high-speed dynamic performance and too long steady-state time brought by the SCVM voltage model, this paper proposes an MRAS observer for the SCVM flux error of an asynchronous motor based on sliding mode control. This method uses SCVM flux error MRAS to identify the speed and estimate the flux, and replaces the traditional PI speed controller with a sliding mode PI speed controller, which can reduce the overshoot of the speed response at medium and high speeds and shorten the time to enter the steady state value. The simulation results show that the designed MRAS observer of asynchronous motor SCVM flux error based on sliding mode control has good steady-state performance and dynamic performance.

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