A Rotor Position/Speed Identification Method of Permanent Magnet Synchronous Motor for Low Speed Operating Condition and Parameter Perturbation

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Abstract

Accurate positioning of the motor requires high control performance for low-speed operation and motor’s parameter perturbation. In the AC motor vector control technology, the nonsingular higher-order terminal sliding mode observer of Insertion Permanent Magnet Synchronous Motor (IPMSM) is put forward based on the object current and voltage values to obtain the rotor position and speed signa. The accurate and reliable identification of the motor rotor position and speed is realized to provide the necessary information of rotor position and speed for high performance vector control system. In order to improve the dynamic response of the observer and overcome the parameter perturbation, the nonsingular terminal sliding mode control is used in the present study. Compared with the general sliding mode control, the chattering phenomenon is reduced by using the high-order sliding mode control technology features. The algorithm is applied to the theodolite micro shafting control system with sensorless IPMSM, which paves the way for theodolite precise target localization. The experiment results show that the method can accurately identify the motor rotor position and speed with strong robustness, good steady precision and dynamic performance.

Keywords: Insertion Permanent Magnet Synchronous Motor; Low Speed Control; The Rotor Position/Speed Identification; Higher-order Sliding Mode Control; Terminal Sliding Mode Control

1 Introduction

Insertion Permanent Magnet Synchronous Motor (IPMSM) has significantly properties of simple structure, compact volume and high torque/weight ratio, which is widely used in aerospace, NC machine tools, robotics and other fields. PMSM should be the important application of precision instrument, such as theodolite, in which the sensorless control technology is important to improve the control reliability and reduce the size of PMSM.

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It’s necessary to obtain the accurate rotor magnetic pole position of PMSM to adjust the size and direction of the stator current vector in the field-oriented vector control or direct torque control, and to produce the required electromagnetic torque. Rotor position and speed signal is usually detected by using rotating transformer mechanical position sensors and photoelectric encoder tachogenerator. It increases the motor size, the system cost and the moment of inertia, reduces the system reliability, and limits the application of transmission system in some specific conditions [1]. In sensorless control technology, some physical variables, such as the stator voltage and current, are measured, and the position sensor is replaced by the rotor position estimation using various control algorithms. It is commonly used to calculate the rotor speed and position by the stator terminal voltage and current directly, for example, the back electromotive forces method (back-EMF) [2]. The rotor position signal is gotten by detecting the zero crossing points of non-conduction phases. High frequency injection method is applied to the rotor initial position estimation of the surface permanent magnet synchronous motor in reference [3]; In reference [4], model reference and adaptive method is used to obtain the rotor position signal. The observer method is to structure the state observer using known information, such as the stator current and voltage, to estimate real-time rotor position [5]. Theoretically speaking, while the continuous rotor position information can be got in full speed range; practically there is a certain recognition effect in only high-speed area. In addition, in recent years, some intelligent methods, such as neural networks, are applied to the motor position observation [6], but they have some problems, for instance, complex computation or excessive motor parameters.

With the development of DSP, the sliding mode observer method [7-10] is widely used in the motor control. PMSM is a multi-variable, strong coupling, nonlinear systems, so it’s difficult to analyse the stability of the observer, for example, the stability analysis much relies on the motor parameters. The adaptive problem of motor parameter must be solved in the position and speed estimation, and the motor position signal is acquired based on sliding mode variable structure control strategy, with a good application prospect in the field of ac speed regulating system control. The traditional sliding mode control contains discontinuous high frequency switching signal and chattering, which can be filtered by the low pass filter. It usually causes phase shift, and the trial and error method is usually used since the selection of filter parameters is difficult. The high-order sliding mode control [11-12] will add the high frequency switch control to the higher derivative of the sliding mode variable, which not only effectively eliminates the chattering phenomenon, but also retains the excellent characteristics of traditional sliding mode control. Furthermore, the observer is not conducive to the low speed motor control, and limits its application in precision instruments. At present, the studies on the sensorless control system for PMSM mainly focus on high speed control. Therefore, the research on low speed and reliable control should be highlighted in motor control.

In this paper, combining the high-order sliding mode control with the Nonsingular Terminal Sliding Mode (NTSM) [13-14] can effectively eliminate system chattering and convergence in a limited time. A new rotor position/speed observer based on high-order nonsingular terminal sliding mode is put forward. The present observer under low speed running and parameter perturbation is more suitable for engineering application. At last, the algorithm is applied in the theodolite micro shaft control system, and the experimental results testify the effectiveness of the algorithm.
2 Rotor Position Estimation Based on Conventional Sliding Mode Observer

Rotor position estimation method of PMSM using sliding mode observer is based on the motor mathematical model. Taking the stator current as the variable, the voltage balance equation [15] in the motor $\alpha$, $\beta$ stationary coordinate system can be described as:

$$
\begin{align*}
\frac{di_\alpha}{dt} &= \frac{1}{L_\alpha} (-R_s i_\alpha + u_\alpha - e_\alpha) \\
\frac{di_\beta}{dt} &= \frac{1}{L_\beta} (-R_s i_\beta + u_\beta - e_\beta)
\end{align*}
$$

(1)

where all the physical variables are defined in $\alpha$, $\beta$ coordinates, $u_\alpha$ and $u_\beta$ are the stator voltages, $i_\alpha$ and $i_\beta$ represent the stator currents, $L_\alpha$ and $L_\beta$ are the equivalent inductances of stator winding $\alpha$, $\beta$ axis, $R_s$ is the stator resistance, and $e_\alpha$ and $e_\beta$ are the back-EMF. Then, the back-EMF equation can be written as:

$$
\begin{align*}
e_\alpha &= -n_p \psi_r \omega_m \sin \theta_r \\
e_\beta &= n_p \psi_r \omega_m \cos \theta_r
\end{align*}
$$

(2)

where $n_p$ is the number of pole pairs, $\psi_r$ is the permanent magnet flux linkage, $\omega_m$ is the motor mechanical angular velocity satisfying $\omega_m = \omega_r / n_p$, and $\theta_r$ is the rotor position electrical angle value. From equation (2), it is clear that the back-EMF of PMSM is a sine wave, whose amplitude is proportional to the speed and phase, which is only related to the rotor position. So the back-EMF of the motor contains the information of the rotor position and speed, by which the motor speed and angle can be calculated.

2.1 The Conventional Sliding Mode Observer Design

Fig. 1 shows the block diagram of sensorless PMSM control system based on high-order terminal sliding mode observer.

![Block diagram of sensorless PMSM control system](image)

Fig. 1: Sensorless control system of PMSM based on high-order terminal sliding mode observer
In this system, the mechanical position sensor is replaced with the speed/position observer. The inputs of this observer are stator voltage and current, with which can estimate the accurate rotor position and speed with a particular observation algorithm. As the feedback signal and the coordinate transformation link, the rotor position and speed is accurately estimated using certain observation algorithm. The motor speed servo controller contains speed loop, direct axis current loop and quadrature axis current loop. PID control method is used in three-loop control. Therefore, the key of the sensorless vector control is to improve the robustness and accuracy of the observer algorithm.

The back-EMF is solved on the basis of voltage balance equation of PMSM. The conventional sliding mode observer is designed as follows:

\[
\left\{\begin{array}{l}
\frac{d\hat{i}_\alpha}{dt} = \frac{1}{L_\alpha} (-R_s\hat{i}_\alpha + u_\alpha + u_1) \\
\frac{d\hat{i}_\beta}{dt} = \frac{1}{L_\beta} (-R_s\hat{i}_\beta + u_\beta + u_2)
\end{array}\right.
\]  

(3)

where \(\hat{i}_\alpha\) and \(\hat{i}_\beta\) are the observations of the motor’s current, and \(u = [u_1, u_2]^T\) is the sliding mode control law as well as the input of the observer. The stator voltage \(u_\alpha\) and \(u_\beta\) can be acquired by the control system directly or by the control system and SVPWM switching signal regardless of the nonlinear perturbation of the inverter. \(i_\alpha\) and \(i_\beta\) are the Clark transforming values of three-phase stator current received from the sensor. It shows that observer (3) cannot be used to observe the back-EMF directly, but to observe the back-EMF based on tracking the stator current.

From the observer (3) and voltage balance equation (1), we can obtain the stator current deviation system equation as follows:

\[
\left\{\begin{array}{l}
\frac{d\bar{i}_\alpha}{dt} = \frac{1}{L_\alpha} (-R_s\bar{i}_\alpha + e_\alpha + u_1) \\
\frac{d\bar{i}_\beta}{dt} = \frac{1}{L_\beta} (-R_s\bar{i}_\beta + e_\beta + u_2)
\end{array}\right.
\]  

(4)

where \(\bar{i}_\alpha = \hat{i}_\alpha - i_\alpha\) and \(\bar{i}_\beta = \hat{i}_\beta - i_\beta\) are defined as the current observation error.

Based on the theory of sliding mode control, the sliding mode surface is chosen as:

\[
s = [s_1, s_2]^T = [\bar{i}_\alpha, \bar{i}_\beta]^T
\]  

(5)

The sliding mode control law is designed as:

\[
u = [u_1, u_2]^T = [-k \text{sgn}(\bar{i}_\alpha), -k \text{sgn}(\bar{i}_\beta)]
\]  

(6)

where \(k>0\) is the sliding gain, and the sgn is sign function. When \(k > max(|e_\alpha|, |e_\beta|)\), the condition for Lyapunov stability is satisfied. The system (4) states can reach the sliding mode \(s = 0\) within finite time.

As with other motors, the parameters of IPMSM, such as resistors, inductors, can cause perturbation as the change of operation conditions, then it will affect motor control performance. Considering motor parameters change, the new definition is as follows:

\[
\Delta R_s = \hat{R}_s - R_s, \quad \Delta L_\alpha = \hat{L}_\alpha - L_\alpha, \quad \Delta L_\beta = \hat{L}_\beta - L_\beta
\]

So, the sliding mode observer (3) can be expressed as:

\[
\left\{\begin{array}{l}
\frac{d\hat{i}_\alpha}{dt} = \frac{1}{L_\alpha} (-R_s\hat{i}_\alpha + u_\alpha + u_1) + H_\alpha \\
\frac{d\hat{i}_\beta}{dt} = \frac{1}{L_\beta} (-R_s\hat{i}_\beta + u_\beta + u_2) + H_\beta
\end{array}\right.
\]  

(7)
where $H = [H_\alpha, H_\beta]$ is parameter error input matrix. Combine equation (7) with equation (3) to attain:

\[
\begin{cases}
H_\alpha = \frac{L_\alpha R_s - L_\alpha R_\alpha}{L_\alpha} \dot{i}_\alpha - \frac{L_\alpha - L_\alpha}{L_\alpha} (u_\alpha + u_1) \\
H_\beta = \frac{L_\beta R_s - L_\beta R_\beta}{L_\beta} \dot{i}_\beta - \frac{L_\beta - L_\beta}{L_\beta} (u_\beta + u_2)
\end{cases}
\]

From equation (3) and equation (7), the motor stator current deviation equation under the condition of parameter perturbation is as follows:

\[
\begin{align*}
\frac{di_\alpha}{dt} &= \frac{1}{L_\alpha} (-R_s \ddot{i}_\alpha + e_\alpha + u_1) + H_\alpha \\
\frac{di_\beta}{dt} &= \frac{1}{L_\beta} (-R_s \ddot{i}_\beta + e_\beta + u_2) + H_\beta
\end{align*}
\]

Design the suitable sliding mode surface and control law to make observer (3) reach the sliding mode state, when it satisfies $i_{\alpha}^T \dot{i}_{\alpha} < 0$, $i_{\beta}^T \dot{i}_{\beta} < 0$, namely:

\[
i_\alpha = \ddot{i}_\alpha = 0, \quad i_\beta = \ddot{i}_\beta = 0
\]

The sliding mode control law is as follows:

\[
u = [u_1, u_2]^T = [-k \text{sgn}(\ddot{i}_\alpha), -k \text{sgn}(\ddot{i}_\beta)]
\]

Substitute the control into equation (8) to get:

\[
[\dot{e}_\alpha, \dot{e}_\beta]^T = [-u_1, -u_2]^T = [k \text{sgn}(\ddot{i}_\alpha), k \text{sgn}(\ddot{i}_\beta)]^T
\]

### 2.2 Design of the Sliding Mode Gain $k$ and Analysis of the Observer

The sliding mode gain $k$ must satisfy the reaching condition of the sliding mode, but the general value is so large that the system chattering is increased, so it is important to choose appropriate $k$.

**Proof:** Consider the Lyapunov function candidate as follows:

\[
V = \frac{1}{2} \ddot{i}^T \ddot{i}
\]

Differentiate $V$ with respect to time along the trajectory of the system equation (8) to yield

\[
\dot{V} = \ddot{i}^T \ddot{i} = \ddot{i}_\alpha^T \ddot{i}_\alpha + \ddot{i}_\beta^T \ddot{i}_\beta = \ddot{i}_\alpha^T \left( -R_s \ddot{i}_\alpha + e_\alpha + u_1 \right) + H_\alpha + \ddot{i}_\beta^T \left( -R_s \ddot{i}_\beta + e_\beta + u_2 \right) + H_\beta
\]

\[
= -\frac{R_s}{L_\alpha} \ddot{i}_\alpha^2 + k |\dot{i}_\alpha| + e_\alpha \dot{i}_\alpha + H_\alpha \ddot{i}_\alpha + \left( -\frac{R_s}{L_\beta} \ddot{i}_\beta^2 - k |\dot{i}_\beta| + e_\beta \dot{i}_\beta + H_\beta \ddot{i}_\beta \right)
\]

The sliding gain can be given,

\[
k = \max_{i=\alpha, \beta} \{(e_i + H_i) \text{sgn}(\dot{i}_i)\} + \eta
\]

where $\eta$ is a small constant.
Substitute $k$ into equation (11), it will make $\dot{V} < 0$, which satisfies the reaching condition. And the back-EMF can be written as:

$$[\hat{e}_\alpha, \hat{e}_\beta]^T = [-u_1, -u_2]^T = [k\text{sgn}(\bar{i}_\alpha), k\text{sgn}(\bar{i}_\beta)]^T$$

(12)

It shows that the back-EMF can be got from the input of observer controller. However, because of the sliding mode control, in fact, the equivalent control is just the switching signal for current observation error, in which there is a large amount of high frequency and discontinuous switching signal. In the application, there need to be a low pass filter, which can effectively extract continuous electromotive force estimation from the switching signal:

$$[\hat{e}_\alpha, \hat{e}_\beta]^T = \left[\left(-\hat{e}_\alpha + k\text{sgn}(\bar{i}_\alpha)\right)/\tau_0, \left(-\hat{e}_\beta + k\text{sgn}(\bar{i}_\beta)\right)/\tau_0\right]^T$$

(13)

where $\tau_0$ is the filter time constant. The selection of the time constant needs to both eliminate higher harmonic very well and fully keep inherent continuous signal contained in the original signal. Then, according to equation (2), the rotor position $\theta_r$ and speed $\omega_m$ can be obtained.

The rotor position and speed estimation method based on conventional sliding mode observer has a certain robustness, but the low pass filter (13) will inevitably bring in phase delay. Currently, for this problem, the experts put forward several methods of phase compensation and adaptive filtering [16, 17], but it is difficult to completely avoid the error from signal extracting. In this paper, the high-order sliding mode control is used to redesign the control strategy of sliding mode observer (7), resulting in a continuous back-EMF, which can avoid the estimation error caused by the extra low pass filter.

3 Rotor Position Estimation of PMSM Based on Higher-order Terminal Sliding Mode Observer

3.1 Design of High-order Non-singular Terminal Sliding Mode Observer

The sliding mode output of the system (8) can be expressed as follows:

$$\bar{i} = [\bar{i}_\alpha, \bar{i}_\beta] = e$$

(14)

According to the high-order sliding mode control theory, the system (8) is the first-order differential coefficient to the sliding mode $s$. The second-order sliding mode control can realize the $u = [u_1, u_2]$ smooth control without chattering and make the state variables reach the second-order sliding mode, just as:

$$\bar{i} = \ddot{i} = 0$$

(15)

The non-singular terminal sliding mode control is selected to achieve the second-order sliding mode control as follows:

$$l = \bar{i} + \gamma \dot{i}^{p/q}$$

(16)

where $l \in R^2$, $\gamma = \text{diag}(\gamma_\alpha, \gamma_\beta)$, $\gamma_\alpha$ and $\gamma_\beta$ are the positive constants, and $p$ and $q$ are odd numbers which satisfy $1 < p/q < 2$. So it can be defined as:

$$\dot{i}^{p/q} = [\dot{i}_\alpha^{p/q}, \dot{i}_\beta^{p/q}]^T$$
By designing a suitable sliding mode control law, the non-singular terminal sliding mode surface (16) can converge to zero within a limited time. If $t_r$ is the convergence time, which means $l(t) = 0, \forall t \geq t_r$, then the sliding mode variable will get into the terminal sliding mode motion state, which can be described as:

$$\ddot{i} + \gamma \ddot{p/q} = 0$$

Suppose the total attaining time is $t_s$ from $e(0) \neq 0$ to $e(t_s) = 0$. And $t_s$ can be expressed as follows:

$$t_s = t_r + \frac{p}{(p - q) \min_i (\gamma i)} \max_i (\ddot{i}(t_r) \ddot{p/q})$$

(17)

So far, the second-order sliding mode state is realized: $\ddot{i} = 0$.

### 3.2 Analysis of Sliding Mode Observer and Calculation of Rotor Position and Speed

**Theorem 1.** For the current observation error system (8), if the sliding surface (14) and non-singular terminal sliding surface (16) are selected, the system will converge in a limited time and the control law is designed as follows:

$$u = u_{eq} + u_a$$

(18)

$$u_{eq} = R_s e$$

(19)

$$u_a = -\int_0^t \left[ L \gamma^{-1}(q/p) \ddot{i}^{2-p/q} + (k + \eta) \sgn(l) + \mu l \right] dt$$

(20)

where $L = \text{diag}(L_\alpha, L_\beta), k = \text{diag}(k_\alpha, k_\beta), k_\alpha > 0$ and $k_\beta > 0$, and $k > \| \dot{\xi} + \dot{H} \|$ should be satisfied, $\eta = \text{diag}(\eta_\alpha, \eta_\beta)$ and $\mu = \text{diag}(\mu_\alpha, \mu_\beta)$, among of them, $\eta_\alpha > 0, \eta_\beta > 0, \mu_\alpha > 0$ and $\mu_\beta > 0$ are all the designing parameters.

**Proof.** Select the Lyapunov function as proof:

$$V = \frac{1}{2} l^T l$$

(21)

Differentiate $V$ with respect to time to yield

$$\dot{V} = l^T \dot{l} = l^T (\dot{i} + \gamma (p/q) \text{diag}(\dot{i}^{p/q-1} \ddot{i}))$$

$$= l^T \gamma (p/q) \text{diag}(\dot{i}^{p/q-1})[\ddot{i}^{2-p/q} + \gamma^{-1}(q/p) \ddot{i}^{2-p/q}]$$

(22)

Substitute equation (19) and (20) into the equation (8) to get:

$$\begin{cases}
\frac{d\xi}{dt} = \frac{1}{L_{\alpha}} (u_{\alpha} + \varepsilon_{\alpha}) + H_{\alpha} \\
\frac{d\xi}{dt} = \frac{1}{L_{\beta}} (u_{\beta} + \varepsilon_{\beta}) + H_{\beta}
\end{cases}$$

(23)

Thus,

$$\dot{V} = l^T \gamma (p/q) \text{diag}(\dot{i}^{p/q-1})[L^{-1} \ddot{u}_{\alpha} + \dot{\xi} + H + \gamma^{-1}(q/p) \ddot{i}^{2-p/q}]$$

$$= l^T \gamma (L^{-1})(p/q) \text{diag}(\dot{i}^{p/q-1})[\ddot{i}^{2-p/q} - (k + \eta) \sgn(l) - \mu l]$$

(24)
where $\xi = [e_\alpha, e_\beta]^T$ is the back-EMF.

As the design parameter satisfies $k > \|H\|$, $\dot{V} \leq -l^T \gamma L^{-1}(p/q)\text{diag}(\bar{i}^{p/q-1})\{\eta \text{sgn}(l) + \mu l\}$, then, $\dot{V} \leq -(p/q)\lambda_\min(\gamma L^{-1}\text{diag}(\bar{i}^{p/q-1}))\{\min(\eta_\alpha, \eta_\beta)\|l\| + \min(\mu_\alpha, \mu_\beta)\|l\|^2\}$.

When $l \neq 0$, since $p$ and $q$ are odd numbers, $1 < p/q < 2$ and $i_{\alpha, \beta}^{p/q-1} \geq 0$, we can get $\dot{V} \leq 0$. If and only if $s_{\alpha, \beta} = 0$, $\dot{V} = 0$. However, $s_{\alpha, \beta} = 0$ and $s_{\alpha, \beta} \neq 0$ can be proved the unstable state, and $\dot{V} \leq 0$ can’t last. According to Lyapunov stability theorem, the system will reach and keep nonsingular terminal sliding mode state $l = 0$ in limited time, and finally the system will reach the second-order sliding mode (15).

Prove up.

**Remark 1**: The terminal sliding mode control law is made up of the measurable variables. The high frequency switching signal is added to the integral function of the control variable $u_n$, whose essence is equivalent to the low pass filter for the symbolic function signal to restrain the chattering phenomenon.

**Remark 2**: The purpose of remark 1 is not for observing the motor stator current, but obtaining the current observer control law $u$. Then we can gain the observed value of the back-EMF, which can be used to identify the rotor position and speed easily.

**Remark 3**: When the uncertainty and disturbance of the model is so significant that it causes severer chattering due to its big gain. To further restrain chattering, the controller $u$ replaces the symbolic function $\text{sgn}(l)$ with the saturation function $\text{sat}(l)$, which can be described as:

$$\text{sat}(l) = \left\{ \begin{array}{ll} 1 & l > \Delta \\ \lambda l & |l| \leq \Delta \\ -1 & l < -\Delta \end{array} \right. \quad (25)$$

where $\Delta$ is the boundary layer.

So, the above design of second-order sliding mode control strategy (14), (16), (18), (19), (20) makes the sliding mode observer (8) converge, and puts the current error system (9) into the second-order sliding state: $\dot{i} = \ddot{i} = 0$. According to the principle of equivalent control, the back-EMF can be represented as:

$$\dot{\xi} = -u \quad (26)$$

The rotor position and speed estimation of PMSM can be calculated based on the back-EMF estimation obtained from equation (26).

$$\dot{\omega}_m = \frac{\sqrt{\dot{e}_\alpha^2 + \dot{e}_\beta^2}}{v_r} \text{sgn}(\dot{e}_\beta \cos\hat{\theta}_r - \dot{e}_\alpha \sin\hat{\theta}_r)$$

$$\dot{\theta}_r = \arctan\left(-\frac{\dot{e}_\alpha}{\dot{e}_\beta}\right) \quad (27)$$

The sketch map of rotor position/speed estimation based on second-order terminal sliding mode observer is shown in Fig. 2. Since there is no chattering control characteristic in the high-order sliding mode, this method can filter high frequency switching signal in the control loop. Smooth control output can be directly taken as the back-EMF estimation to avoid the low-pass filtering outside the closed control loop, which may cause the phase lag in the back-EMF estimation.
4 Experiments and Analysis of Algorithm

To validate the feasibility of the sensorless algorithm, it will be applied to the theodolite micro shafting control system with sensorless PMSM. The platform for experiment system is as shown in Fig. 3, which includes PMSM experimental control system and micro precision inspection device.

The major parameters of the motor are written as: rated speed $\omega_{rs}=874$ r/min, resistance $R=2.5 \, \Omega$, the stator inductance $L_d=13.5 \, mH$ and $L_q=18.0 \, mH$, rotational inertia $J = 2.5 \times 10^{-4} \, kg/m^2$, damping coefficient $B=0.05 \, Nms$, permanent magnet flux linkage $\psi_r=0.477 \, Wb$, pole pairs $n_p=4$.

The servo controller parameters in the PMSM control system and the parameters of sliding mode observer are shown in Table 1.

<table>
<thead>
<tr>
<th></th>
<th>Speed loop</th>
<th>Direct axis current loop</th>
<th>Quadrature axis current loop</th>
<th>Observers</th>
</tr>
</thead>
<tbody>
<tr>
<td>$K_{P1}$</td>
<td>0.07</td>
<td>$K_{P2}$ = 0.07</td>
<td>$K_{P3}$ = 0.45</td>
<td>$\gamma_{\alpha,\beta} = 0.001, p = 5$</td>
</tr>
<tr>
<td>$T_{I1}$</td>
<td>$5 \times 10^{-4}$</td>
<td>$T_{I2}$ = $5 \times 10^{-4}$</td>
<td>$T_{I3}$ = $7.5 \times 10^{-3}$</td>
<td>$q = 3, (k + \eta)_{\alpha,\beta} = 5000$</td>
</tr>
<tr>
<td>$T_{D1}$</td>
<td>$1 \times 10^{-4}$</td>
<td>$T_{D2}$ = $1 \times 10^{-4}$</td>
<td>$T_{D3}$ = $1 \times 10^{-3}$</td>
<td>$\mu = 200, \Delta = 0.25$</td>
</tr>
</tbody>
</table>

In the experiment process, the motor is started by using the conventional method of V/F control. When the speed reaches 10 r/min, we make the motor stable for a period of time, then switch to the sensorless control.
Fig. 4 shows the changing curves of rotor position estimation and actual value, when the motor stably operates in 0.03 times of rated speed $\omega_{\text{rs}}$ without load. The results show that the error of the rotor position estimation decreases gradually. After a period of stable operation, without position angle compensation, the error of actual and estimated values is about 0.12 rad = 6.88°, which is negligible compared with the actual value. The experimental results prove that the non-singular terminal high-order sliding mode observer algorithm has good control performance and avoids the phase lag caused by low pass filter of the traditional sliding mode observer.

![Fig. 4: Estimated curves of rotor position (a) PMSM experimental control system; (b) Micro precision inspection device (pitch)](image)

In the sensorless closed-loop control mode, the motor’s speed increases from 0.05 times of rated speed to 0.1 times and then the changing curve is shown in Fig. 5. In the process of speed changing, the tracking error is a little bigger, but it decreases quickly. And the speed fluctuation is slight within ±1.2 r/min in the process of stable operation. It proves that the algorithm can better dynamic performance without motor parameter perturbation.

Fig. 6 shows the changing curve of the sliding mode control law $u_1$ and $u_2$. From the figure, we can attain that when the speed is changing, $u_1$ and $u_2$ can track the actual value in real time,

![Fig. 5: The curve of estimated and actual speed](image)
just as

\[ u_1 \approx -n_p \psi_r \omega_m \sin \theta_r \]

and

\[ u_2 \approx n_p \psi_r \omega_m \cos \theta_r. \]

The change of motor parameter and speed is considered. The values of \( R_s, L_d \) and \( L_q \) are all reducing or increasing 20\%, and the motor speed is increasing from 0.05 times of rated speed to 0.1 times. Fig. 7 shows the curve of estimated speed value and actual speed value, and Fig. 8 shows the sliding mode control law.

As we can see from the figure, when the motor parameters happen to change, the actual speed value will produce larger fluctuation. It mainly because that the speed servo controller is based on the ordinary PID control. However, for the sliding mode observer, the speed estimation and the sliding mode control law can track the change of the actual value. It proves that the higher-order nonsingular terminal sliding mode observer has stronger robustness.

Fig. 7: The curve of estimated and actual speed while the parameters change (a) \( R_s, L_d \) and \( L_q \) all reduce 20\%; (b) \( R_s, L_d \) and \( L_q \) all increase 20\%
Fig. 8: Sliding mode control law $u_1$ and $u_2$ while the parameters change (a) $R_s$, $L_d$ and $L_q$ all reduce 20%; (b) $R_s$, $L_d$ and $L_q$ all increase 20%

5 Conclusion

When the motor condition is in low speed running and parameter perturbation, the sliding mode surface is built according to the error between the measured current and observed current, and the estimated speed/position is calculated based on the observations of the back-EMF. A PMSM sensorless control technology is put forward in this paper based on the singular higher-order terminal sliding mode observer. Experiment results show the dynamic response speed and robustness of the observer is improved by using nonsingular terminal sliding mode. The chattering phenomenon of sliding mode is effectively suppressed by the properties of the high-order sliding mode control technology. The motor rotor position/speed identification algorithm is applied in the theodolite precision micro system, which has good dynamic and static performance.

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