Closed-loop Parameter Identification of Permanent Magnet Synchronous Motor Considering Nonlinear Influence Factors*

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Abstract

This paper proposes a closed-loop identification method based on an AC servo control system for permanent magnet synchronous motor (PMSM) parameters. The parameters include the inductance of the quadrature-direct axis, resistance of the stator, and magnetic flux linkage of the rotor. First, the nonlinear factors influencing the parameter identification accuracy—the motor model error caused by non-sinusoidal flux, switch dead zone of the inverter, and rotor position feedback delay of the encoder—were analyzed to develop corresponding compensation methods. Second, an experimental system was developed, and the proposed method was implemented and evaluated. The analytical and experimental results showed that the identified inductance-current curve was consistent with the results of finite element analysis after compensation, while the estimated resistance and flux were closed to the measured values. These results confirm the validity of the proposed method.

Key words: AC Servo Control System, Permanent Magnet Synchronous Motor, Closed-loop Identification

1. Introduction

In AC servo control systems, the permanent magnet synchronous motor (PMSM) is universally controlled by the vector control method; it is based on the Park equation, which is available under several assumed conditions. However, the complexity of design and processing technology and the saturation characteristics of the permanent magnetic material introduce modeling errors and nonlinear variations in the PMSM parameters, thereby affecting the performance and precision of AC servo control systems. There are several ways to obtain a fairly accurate mathematical model and parameters of the motor. One is to estimate the parameters and analyze the motor harmonic by the finite element analysis method⁽¹⁾; this is complicated, but effective. This method is normally used in motor design and analysis. Another is the AC standstill method^(2,3), which is mainly used for the estimation of the inductance, but not for identification of other parameters and a variety of motor parameters being controlled. In addition, closed-loop identification^(4,5) and online parameter identification^(6,7) use adaptive algorithms that identify the equivalent parameter of motor. Both these methods are based on the AC servo control system and are applied in a closed-loop situation. Adaptive online identification is effective for parameter variation caused by system condition variation, and thus, improves the system control performance. However, there may still be accuracy failures because the reference model is actually a theoretical model that is affected by several factors of the closed-loop control system.

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By analyzing the d - q axis mathematical model of a permanent magnet synchronous motor (PMSM), we propose a closed-loop identification method for the inductance of the quadrature-direct axis, resistance of the stator, and magnetic flux linkage of the rotor. We also analyze the motor model error caused by non-sinusoidal flux, switch dead zone of the inverter, and rotor position feedback delay of the encoder and propose suitable compensation methods.

2. Parameter identification model of PMSM

To establish the mathematical model of the PMSM mode, we made the following assumptions:

- 1) The parameters of the motor are independent of the motor temperature.
- 2) The saturation level is fully determined by the permanent magnets and hence the value of the relative permeability of the iron does not significantly change with the stator current values.
- 3) Iron losses are neglected.
- 4) The cogging torque is neglected.
- 5) The variations of the flux induced in the stator phases by the rotor magnets are sinusoidal, which implies that the induced electromotive forces are sinusoidal.

The d - q axis mathematical model of the PMSM is

$$\begin{cases} u_d = R_s i_d + L_d P i_d - \omega_e L_q i_q \\ u_q = R_s i_q + L_q P i_q + \omega_e L_d i_d + \omega_e \psi_m \end{cases} \quad (1)$$

where u_d and u_q are the q and d axis voltages, R_s is the phase resistance, i_d and i_q are the d and q axis currents, L_d and L_q are the q and d axis inductances, P is the differential operator, ω_e is the electrical angular velocity, and ψ_m is the rotor flux.

For the PMSM used in this study, L_d and L_q were considered equal. By using vector control with a d axis current of $i_d = 0$, if the motor is running under a high speed and constant load, $R_s i_d$ and $\omega_e L_d i_d$ are close to zero and i_q is almost constant; thus, compared with $\omega_e L_q i_q$ and $\omega_e \psi_m$, the values of $L_d P i_d$, $L_q P i_q$, $R_s i_d$, and $\omega_e L_d i_d$ are small, and the mean value of each is zero. The model for parameter identification can thus be described as

$$\begin{cases} u_d = -\omega_e L_q i_q + \varepsilon_1 \\ u_q = R_s i_q + \omega_e \psi_m + \varepsilon_2 \end{cases} \quad (2)$$

where ε_1 and ε_2 are white noise.

3. The influencing factors of closed-loop parameter identification and their compensations

The system diagram for closed-loop parameter identification is shown in Fig. 1. The servo controller consists of a control unit based on a Microprogrammed Control Unit (MCU), drive module based on an intelligent power module (IPM), and an angle feedback module based on a digital absolute magnetic encoder. The servo controller not only controls the operation of the motor but also samples related data from the motor operation and translates it to the PC. Estimation of the respective parameters using the least square method is calculated in the PC. The load is in terms of resistance. The power output of the synchronous generator can be adjusted by a voltage regulator.

Theoretically, the accuracy of each segment of the closed-loop system affects the accuracy of the parameter identification. As shown in Fig. 1, the IPM outputs the drive current of the motor. The angle and current feedback signal can be obtained by the encoder and current sensor, respectively. In this system, the PMSM model error, IPM switch dead zone, encoder rotor position feedback delay, and current sensor delay and resolution affect parameter identification.

Fig.1 System diagram for closed-loop parameter identification

To minimize the delay and maximize the accuracy of the current sensor, a state observer is used to observe the current. Because the control cycle of the servo controller is minimized to the ultimate $60\mu s$, the influence of the current sensor accuracy on parameter identification can be ignored.

3.1 Influence of Permanent Magnet Flux Harmonic and Compensation

Normally, the permanent magnet flux of the rotor is non-sinusoidal and has harmonics which cause modeling errors. Taking these harmonics into account, the d - q axis mathematical model of PMSM can be described as⁽⁸⁾

$$\begin{cases} u_d = R_s i_d + L_d P i_d - \omega_e L_q i_q + \Psi_{d6} \omega_e \sin 6\theta + \Psi_{d12} \omega_e \sin 12\theta + \dots \\ u_q = R_s i_q + L_q P i_q + \omega_e L_d i_d + \omega_e \Psi_m + \Psi_{q6} \omega_e \cos 6\theta + \Psi_{q12} \omega_e \cos 12\theta + \dots \end{cases} \quad (3)$$

where θ is the angle of the rotor and Ψ_{d6} , Ψ_{d12} , Ψ_{q6} , and Ψ_{q12} are the sixth and twelfth harmonic amplitudes of the permanent magnet flux harmonic for the d and q axes; $\Psi_{d6} = (-5\psi_5 - 7\psi_7)$, $\Psi_{d12} = (-11\psi_{11} - 13\psi_{13})$, $\Psi_{q6} = (-5\psi_5 + 7\psi_7)$, $\Psi_{q12} = (-11\psi_{11} + 13\psi_{13})$. ψ_i ($i=5, 7, 11, 13$) are the fifth, seventh, eleventh, and thirteenth harmonic amplitudes of the permanent magnet flux harmonic.

Considering the phase deviation angle φ between the harmonic and its subharmonic, the d - q axis mathematical model of PMSM can be described as

$$\begin{cases} u_d = R_s i_d + L_d P i_d - \omega_e L_q i_q + \Psi_{d6} \omega_e \sin(6\theta + \varphi) + \Psi_{d12} \omega_e \sin(12\theta + \varphi) + \dots \\ u_q = R_s i_q + L_q P i_q + \omega_e L_d i_d + \omega_e \psi_m + \Psi_{q6} \omega_e \cos(6\theta + \varphi) + \Psi_{q12} \omega_e \cos(12\theta + \varphi) + \dots \end{cases} \quad (4)$$

The sixth and twelfth harmonics are considered to be the main subharmonics; the higher harmonic can be ignored. Using equation (2), equation (4) can be simplified as follows:

$$\begin{cases} u_d = -\omega_e L_q i_q + \Psi_{d6} \omega_e \sin(6\theta + \varphi) + \Psi_{d12} \omega_e \sin(12\theta + \varphi) + \varepsilon_1 \\ u_q = R_s i_q + \omega_e \psi_m + \Psi_{q6} \omega_e \cos(6\theta + \varphi) + \Psi_{q12} \omega_e \cos(12\theta + \varphi) + \varepsilon_2 \end{cases} \quad (5)$$

Equation (5) shows that at the same speeds, the harmonic makes the identified inductance vary linearly and the identified resistance and flux vary nonlinearly. The smaller the current and the higher the speed are, the bigger their influences.

Based on equation (3), the motor model can be modified by feed forward, in which the sixth and twelfth harmonics are fed to the q and d axis voltages.

The phases current and power spectrum of a single phase before and after harmonic compensation with the q axis current set to $i_q = 3$ A and motor speed set to $\omega_e = 2000$ rpm are illustrated in Fig. 2(a) and (b), Fig. 3(a) and (b). Table 1 shows the power spectral density contrast of them based on the equation below.

$$A_i = 10 \log \left(\frac{R_i^2 + I_i^2}{\gamma} \right) \quad (6)$$

where A_i is the power spectral density of the i -th harmonic, R_i is the real part of the i -th harmonic, and I_i is the imaginary part of the i -th harmonic.

The gliding property and correctness of the latter current and the latter power spectral

density of the power spectrum are better than their previous ones. Moreover, the amplitude and frequency of the d and q axis currents are also improved after harmonic compensation, as shown in Figs. 4 and 5.

(a) Before harmonic compensation

(b) After harmonic compensation

Fig.2 Current wave curves of three phases

(a) Before harmonic compensation

(b) After harmonic compensation

Fig.3 Power spectrum of single phase

Table.1 Power Spectral Density Contrast Before and After Harmonic Compensation

i_f	Before Harmonic Compensation	After Harmonic Compensation
5th harmonic ($A^2 \cdot s$)	-17.61 dB	-27.81 dB
7th harmonic ($A^2 \cdot s$)	-25.63 dB	-42.80 dB
11th harmonic ($A^2 \cdot s$)	-27.21 dB	-28.41 dB
13th harmonic ($A^2 \cdot s$)	-29.58 dB	-36.20 dB

(a) Before harmonic compensation

(b) After harmonic compensation

Fig.4 Current wave curves of i_d

(a) Before harmonic compensation

(b) After harmonic compensation

Fig.5 Current wave curves of i_q

3.2 Influence of Switch Dead Zone of Inverter and Compensation

Because of the switch dead zone and physical delay of the inverter, there is a voltage difference Δu between the actual phase voltages of the inverter outputs and the ideal voltages⁽⁹⁾. The three actual phase voltages can be transformed to the d - q coordinate system:

$$\begin{cases} \begin{bmatrix} u_d \\ u_q \end{bmatrix} = \begin{bmatrix} u_{d0} \\ u_{q0} \end{bmatrix} + \begin{bmatrix} \Delta u_d \\ \Delta u_q \end{bmatrix} \\ \Delta u_d = \sqrt{\frac{2}{3}} [\Delta u_u \cos \theta + \Delta u_v \cos(\theta - \frac{2}{3}\pi) + \Delta u_w \cos(\theta + \frac{2}{3}\pi)] \\ \Delta u_q = -\sqrt{\frac{2}{3}} [\Delta u_u \sin \theta + \Delta u_v \sin(\theta - \frac{2}{3}\pi) + \Delta u_w \sin(\theta + \frac{2}{3}\pi)] \end{cases} \quad (7)$$

where u_d and u_q are actual voltages of the d and q axes, u_{d0} and u_{q0} are the ideal voltages of the d and q axes, Δu_d and Δu_q are the voltage errors of the d and q axes, and Δu_u , Δu_v , and Δu_w are voltage errors of the three phases.

After introducing the voltage errors caused by the switch dead zone and physical delay of the inverter, the motor model in the d - q coordinate system can be described as

$$\begin{cases} u_d = u_{d0} + \Delta u_d = R_s i_d + L_d p i_d - \omega_e L_q i_q \\ u_q = u_{q0} + \Delta u_q = R_s i_q + L_q p i_q + \omega_e L_d i_d + \omega_e \psi_m \end{cases} \quad (8)$$

The model for parameter identification shown in equation (2) can be rewritten as

$$\begin{cases} u_{d0} + \Delta u_d = -\omega_e L_q i_q + \varepsilon_1 \\ u_{q0} + \Delta u_q = R_s i_q + \omega_e \psi_m + \varepsilon_2 \end{cases} \quad (9)$$

In equation (9), Δu_d and Δu_q are constant values that are independent of the current and speed. Δu_d makes the identified inductance vary nonlinearly; the smaller the speed ω_e and current i_q are, the greater the difference between the identified and true inductance values. Similarly for Δu_q , the smaller the speed ω_e and current i_q , the larger is the difference between the identified resistance value R_s and rotor flux value ψ_m and their respective true values.

We compensate for the drive signal of the inverter by the time-compensation method. This method compensates the upper bridge when the current runs out of the inverter and compensates the lower bridge when the current runs in. The compensation time is $T_{cmp} = T_d + T_{on} - T_{off}$, where T_d is the dead time and T_{on} and T_{off} are the physical open and closed times of the power transistor for the inverter. The key to this method is the detection of the current direction. Since the zero point of the current is difficult to detect, we ignore it and propose the following method: if $-RG_1 < i < RG_1$, no compensation is made; if $RG_1 < i < RG_2$ or $-RG_2 < i < -RG_1$, compensate linearly; and if $i > RG_2$ or $i < -RG_2$, use the value T_{cmp} as a constant compensation, as shown in Fig. 6.

Fig.6 Compensation for switch dead zone of inverter

3.3 Influence of Angle Feedback Error and Compensation

The rotor angle feedback sensor for the proposed parameter identification system is a digital absolute magnetic encoder⁽¹⁰⁾ that converts the magnetic angle into a voltage signal through a Hall sensor⁽¹¹⁾. After receiving an instruction signal from the main control unit through SPI, the encoder enables the sampling process; after digital processing, it transmits the digital angle signal to the central control unit.

In this process, the feedback angle that the controller receives is delayed. Suppose that the communication cycle has a length T_s ; the encoder samples are in $(n)T_s$, get to the buffer of the encoder in $(n+1)T_s$, and are transmitted to the central control unit in $(n+2)T_s$. Suppose the communication time is t_c , and the time this value is put into use after the end of communication is t_e . In total, the angle delay time is thus $T_{delay} = 2T_s - (t_c + t_e)$, and the angle error is $\Delta\theta = \omega_e T_{delay}$.

Fig.7 Relationship between the current and electrical rotor angle error

As shown in Fig. 7, the actual electrical angle of the rotor is $\theta = \theta_0 + \Delta\theta$, where θ_0 is the angle that the central control unit receives. The actual d and q currents i_d and i_q and voltages u_d and u_q can be described as

$$\begin{cases} i_d = i_{d0} \cos \Delta\theta - i_{q0} \sin \Delta\theta \\ i_q = i_{q0} \cos \Delta\theta + i_{d0} \sin \Delta\theta \end{cases} \quad (10)$$

$$\begin{cases} u_d = u_{d0} \cos \Delta\theta + u_{q0} \sin \Delta\theta \\ u_q = u_{q0} \cos \Delta\theta - u_{d0} \sin \Delta\theta \end{cases} \quad (11)$$

where i_{d0} and i_{q0} are ideal currents and u_{d0} and u_{q0} are ideal voltages.

If the motor speed is $\omega_e = 3000$ rpm, $\Delta\theta = 0.028274$ rad, $\cos \Delta\theta = 0.9996 \approx 1$, and $\sin \Delta\theta = 0.028271 \approx \Delta\theta$, then the motor model in the d - q axis coordinate system can be described as

$$\begin{cases} u_{d0} + \Delta\theta u_{q0} = R_s(i_{d0} - i_{q0}\Delta\theta) + L_d P(i_{d0} - i_{q0}\Delta\theta) - \omega_e L_q(i_{q0} + i_{d0}\Delta\theta) \\ u_{q0} - \Delta\theta u_{d0} = R_s(i_{q0} + i_{d0}\Delta\theta) + L_q P(i_{q0} + i_{d0}\Delta\theta) + \omega_e L_d(i_{d0} - i_{q0}\Delta\theta) + \omega_e \psi_m \end{cases} \quad (12)$$

Equation (2) can be rewritten as

$$\begin{cases} u_{d0} + \Delta\theta u_{q0} = -R_s i_{q0} \Delta\theta - \omega_e L_q i_{q0} + \varepsilon_1 \\ u_{q0} - \Delta\theta u_{d0} = R_s i_{q0} - \omega_e L_d i_{q0} \Delta\theta + \omega_e \psi_m + \varepsilon_2 \end{cases} \quad (13)$$

In equation (13), $\Delta\theta$ is proportional to the speed. At the same current value, the higher the speed is, the greater the influence of $\Delta\theta$ on the identified parameter. Inversely, at the same speed, the smaller the current is, the greater the influence of $\Delta\theta$ on the identified parameter.

While controlling the motor, we compensate $\Delta\theta = \omega_e T_{delay}$ for the calculation angle. Based on the analysis above, the identified parameter is independent of the speed, and the test results shown in Fig. 8 confirmed the validity of the compensation.

Fig.8 Relationship between identified inductance and the speed before and after angle compensation

4. Experimental results

The experimental setup for practical application is shown in Fig. 9. The components include an AC servo controller, an 8-channel oscilloscope, data acquisition system, a voltage regulator, a synchronous generator, a PMSM, and some power resistances.

Fig.9 Experiment setting of parameter identification in this paper

The relationship between the inductance of the d and q axes and the motor current was analyzed by the finite element analysis method. The static resistance of the testing motor was tested by a resistance measuring instrument and found to be 1.613Ω . The rotor flux was calculated by back emf, which was tested while the rotor was driven by another motor.

Based on the closed-loop parameter identification system mentioned above, under the condition of speed $\omega_e = 3000$ rpm and current $i_q = 0.5-9$ A, we identified the d and q axis inductance, stator resistance, and rotor flux. Fig.10 shows the contrast of inductance analyzed by the finite element analysis method, the identified inductance after using all compensations and the identified inductance without using one of the compensation (while the other two compensations were used). As shown in the figure, the identified inductance value after using all compensations matches the inductance analyzed by the finite element analysis method.

Fig.10 Contrast in identified and analyzed inductance values without and after compensation

The contrast in the static resistance value tested by the resistance measuring instrument and the identified resistance value is shown in Fig. 11. The contrast in the rotor flux calculated by back EMF and the identified rotor flux is shown in Fig. 12. Both of the identified stator resistance and rotor flux values match the measurement results.

Fig. 11 Contrast in identified and measured resistance values without and after compensation

Fig. 12 Contrast in identified and measured rotor flux values without and after compensation

Table 2 lists the maximum error rate E_{L_max} between the identified inductance and inductance analyzed by the finite element analysis method, maximum error rate E_{R_max} between the identified and measured resistance values, and maximum error rate E_{ψ_max} between the identified and measured rotor flux values. The maximum error rate is defined by the equation below.

$$E_{\max} = \frac{\max|\lambda_e - \lambda_r|}{\lambda_r} \times 100\% \quad (14)$$

where λ_e is the identified value and λ_r is the true value.

Table 2 Maximum error rates between identified parameters and their measured values

Maximum Error Rates	Without Harmonic Compensation	Without Dead Zone Compensation	Without Angle Compensation	After Using All Compensations
E_{L_max}	19.16%	133.27%	43.78%	6.85%
E_{R_max}	17.54%	60.94%	55.72%	4.66%
E_{ψ_max}	2.35%	24.87%	1.5%	0.84%

As shown in the table, after using all compensations, the maximum error rate for each item was less than 7%, which confirms the validity of the proposed identification method.

5. Conclusion

This paper proposes a closed-loop identification method based on a mathematical model for a permanent magnet synchronous motor in an AC servo control system to identify the inductance of a quadrature-direct axis, stator resistance, and magnetic flux linkage of a rotor in a motor operation. By analyzing and compensating for each factor influencing the accuracy of the identified parameter, we can make the identified parameters match their

actual value. The proposed method provides an academic and experimental basis to the identification used in self-adaptive control. For the experiments, we used a 400-W motor. The experimental results showed that the maximum error rate between the identified inductance and inductance analyzed by the finite element analysis method was 6.85%, The maximum error rate between the identified and measured resistance values was 4.66%, and the maximum error rate between the identified and measured rotor flux values was 0.84%. These results confirm the validity of the proposed identification method and compensation method for nonlinear factors.

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