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Position Sensorless Control of Interior Permanent Magnet Synchronous Motors using Unknown Input Observer and High-frequency Signal Injection

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Abstract: The position sensorless control method based on unknown input observer was proposed aiming at the Interior Permanent Magnet Synchronous Motor (IPMSM) in this study. Due to PWM inverter output voltage of the IPMSM inconsistent with input reference voltage and in order to eliminate the influence of input voltage to rotor position, this study use sampling current as another state variable of the observer to build the unknown input observer, thereby estimate the rotor position at the high-speed. Simulation results show that the composite control has good dynamic characteristics in the full speed range.

Key words: IPMSM, Unknown Input Observer, Position Sensorless Control, High-frequency Signal Injection

INTRODUCTION

Interior Permanent Magnet Synchronous Motor (IPMSM) was suitable as an electric vehicle drive motor due to the advantages of high power density and wide speed range (Consoli *et al*., 2010). Sensorless control technology is undoubtedly appropriate control strategy to electric vehicle drive motor consider for reducing costs as well as the space limited. IPMSM sensorless control method is mainly divided into two categories (Gu *et al*., 2009).

suitable for start and slow stage, suitable for high-speed. Initial position detection and low-speed stage is basically using a high-frequency signal injection method (Jansen and Lorenz, 1996; Jang *et al*., 2003; Ha *et al*., 2003; Jang *et al*., 2004; Bianchi *et al*., 2008) while can be used technology to high-speed have many choices, such as flux estimation method (Deng *et al*., 2007), Model Reference Adaptive System (MRAS) method (Qi *et al*., 2004), state observer method (Yang and Lorenz, 2012), sliding mode variable structure method (Parasiliti *et al*., 2001; Su *et al*., 2009), Kalman filtering method (Bolognani *et al*., 2003).

The higher the speed of IPMSM, the greater the counter Electromotive Force (EMF). EMF will be close to the DC bus voltage of the inverter when the speed is increased to a certain value. Then the torque response time will be prolonged if motor still high-speed running, so result in potential safety hazard. In order to solve this problem, overmodulation mode of the PWM inverter is often employed to enlarge inverter output voltage while keeping stator current waveform (Hasegawa *et al*., 2009). Meanwhile, the sensorless control technology are generally required use the inverter output voltage and current to estimate the rotor position of the motor, but PWM inverter output voltage is not always consistent with the input reference voltage which result in the sensorless strategy using the voltage information is unavailable under the overmodulation mode of the PWM inverter. Based on this point, the literature (Hasegawa *et al*., 2009) proposed a rotor position estimation control strategy without the voltage information-unknown input observer.

Although unknown input observer (Hasegawa *et al*., 2007; Hasegawa *et al*., 2009) estimate the rotor position does not require any voltage information and can used in overmodulation mode. But the method cannot be applied to drive the IPMSM at standstill. Therefore, initial rotor position detection and rotor position estimates at low speed also need a high-frequency signal injection method. So, this study proposed using high-frequency signal injection-unknown input observer to complete IPMSM sensorless control.

IPMSM UNKNOWN INPUT OBSERVER

Mathematical model of IPMSM: The IPMSM mathematical model in the rotating coordinate can be described as:

$$\begin{pmatrix} u_d(t) \\ u_q(t) \end{pmatrix} = \begin{pmatrix} R + pL_d & -\omega_r L_q \\ \omega_r L_d & R + pL_q \end{pmatrix} \begin{pmatrix} i_d(t) \\ i_q(t) \end{pmatrix} + \begin{pmatrix} 0 \\ \omega_r K_e \end{pmatrix} \tag{1}$$

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where, u_d (t), u_q (t), i_d (t) and i_q(t) be voltages on the rotating coordinate aligned with rotor position (d-q axis) and currents on d-q axis, respectively. L_d and L_q mean d-axis and q-axis inductances, respectively. Moreover, R and ω_r stand for winding resistance and rotor speed, respectively. p means the differential operator.

This study first considers the flux observer on α-β coordinates for position estimation. Consider the aforementioned voltage problem, hence, need to eliminate the voltage variable and only take current and a kind of flux as state variables to constructed linear flux model of IPMSM. The developed torque ô of IPMSM is written as:

$$T_e(t) = (L_d - L_q)i_d(t)i_q(t) + K_e i_q(t) = |?(t)| i_q(t) \quad (2)$$

where, $|\lambda(t)| = (L_d\text{-}L_q)i_d(t)+K_e$, definition the direction of $\lambda(t)$ aligns with the d-axis. Therefore, the rotor position θ_r can be obtained by estimating flux vector λ (t) on the stationary reference frame (α-β coordinates).

From the aforementioned definition, λ (t) is expressed on α-β coordinates by:

$$\lambda(t) = \begin{pmatrix} \lambda_\alpha(t) \\ \lambda_\beta(t) \end{pmatrix} = ((L_d - L_q)i_d(t) + K_e)\begin{pmatrix} \cos\theta_r \\ \sin\theta_r \end{pmatrix} \quad (3)$$

Its differential function is shown by:

$$p\begin{pmatrix} \lambda_\alpha(t) \\ \lambda_\beta(t) \end{pmatrix} = (L_d - L_q)\dot{i}_d(t)\begin{pmatrix} \cos\theta_r \\ \sin\theta_r \end{pmatrix} + \omega_r((L_d - L_q)i_d(t) + K_e)\begin{pmatrix} -\sin\theta_r \\ \cos\theta_r \end{pmatrix} \quad (4)$$

Combine Eq. 4 and 1 can be manipulated by the following:

$$\begin{pmatrix} u_d(t) \\ u_q(t) \end{pmatrix} = \begin{pmatrix} R + pL_d & -\omega_r L_q \\ \omega_r L_d & R + pL_q \end{pmatrix}\begin{pmatrix} i_d(t) \\ i_q(t) \end{pmatrix} + \begin{pmatrix} (L_d - L_q)\dot{i}_d(t) \\ 0 \end{pmatrix} + \begin{pmatrix} 0 \\ \omega_r((L_d - L_q)i_d(t) + K_e \end{pmatrix} \quad (5)$$

Transforming this equation onto α-β coordinates, the following equation can be obtained:

$$\begin{pmatrix} u_\alpha(t) \\ u_\beta(t) \end{pmatrix} = \begin{pmatrix} R + pL_d & 0 \\ 0 & R + pL_q \end{pmatrix}\begin{pmatrix} i_\alpha(t) \\ i_\beta(t) \end{pmatrix} + (L_d - L_q)\dot{i}_d(t)\begin{pmatrix} \cos\theta_r \\ \sin\theta_r \end{pmatrix} + (\omega_r((L_d - L_q)i_d(t) + K_e)\begin{pmatrix} -\sin\theta_r \\ \cos\theta_r \end{pmatrix} \quad (6)$$

According to Eq. 4, the above equation can be further rewritten as:

$$\begin{pmatrix} u_\alpha(t) \\ u_\beta(t) \end{pmatrix} = \begin{pmatrix} R + pL_d & 0 \\ 0 & R + pL_q \end{pmatrix}\begin{pmatrix} i_\alpha(t) \\ i_\beta(t) \end{pmatrix} + p\begin{pmatrix} \lambda_\alpha(t) \\ \lambda_\beta(t) \end{pmatrix} \quad (7)$$

It can be seen that the nondiagonal terms do not appear in impedance matrix and be linear equations to the flux. To realize position sensorless control system, the rotor position θ_r can be obtained by estimating flux vector λ, that is, $\theta_r = \tan^{-1}(\lambda_\beta/\lambda_\alpha)$.

Let i (t), λ (t) be current, flux, respectively. The state equation of the IPMSM on a continuous-time system based on complex notation is expressed as the following:

$$\begin{cases} \dfrac{d}{dt}\begin{pmatrix} i(t) \\ ?(t) \end{pmatrix} = \begin{pmatrix} -\dfrac{R}{L_q} & -\dfrac{j\omega_r}{L_q} \\ 0 & j\omega_r \end{pmatrix}\begin{pmatrix} i(t) \\ ?(t) \end{pmatrix} + \begin{pmatrix} \dfrac{1}{L_q} \\ 0 \end{pmatrix} u(t) \\ i(t) = (1 \quad 0)\begin{pmatrix} i(t) \\ \lambda(t) \end{pmatrix} = C\begin{pmatrix} i(t) \\ ?(t) \end{pmatrix} \end{cases} \quad (8)$$

Discrete the above equation with sampling period T, can be obtained:

$$\begin{aligned} \begin{pmatrix} i(k+1) \\ \lambda(k+1) \end{pmatrix} &= \begin{pmatrix} \exp\left(\dfrac{-R}{L_q}T\right) & A_{12} \\ 0 & \exp(j\omega_r T) \end{pmatrix}\begin{pmatrix} i(k) \\ \lambda(k) \end{pmatrix} \\ &+ \begin{pmatrix} \dfrac{1}{R}\left(1 - \exp\left(\dfrac{-R}{L_q}T\right)\right) \\ 0 \end{pmatrix} u(k) \\ &= A_d\begin{pmatrix} i(k) \\ \lambda(k) \end{pmatrix} + B_d u(k) \end{aligned} \quad (9)$$

Where:

$$A_{12} = \frac{\omega_r}{L_q}\,\frac{\omega_r \exp\left(\dfrac{-R}{L_q}T\right) - \omega_r\cos(\omega_r T) + \dfrac{R}{L_q}\sin(\omega_r T)}{\dfrac{R^2}{L_q^2} + \omega_r^2} + j\frac{\omega_r}{L_q}\,\frac{\dfrac{R}{L_q}\exp\left(\dfrac{-R}{L_q}T\right) - \dfrac{R}{L_q}\cos(\omega_r T) - \omega_r\sin(\omega_r T)}{\dfrac{R^2}{L_q^2} + \omega_r^2}$$

Design of IPMSM unknown input observer: In general, constructing IPMSM rotor position observer required voltage state variable. In order to eliminate the voltage, PWM carrier frequency increase twice and add once the current sampling at the moment of T/2 as a second current

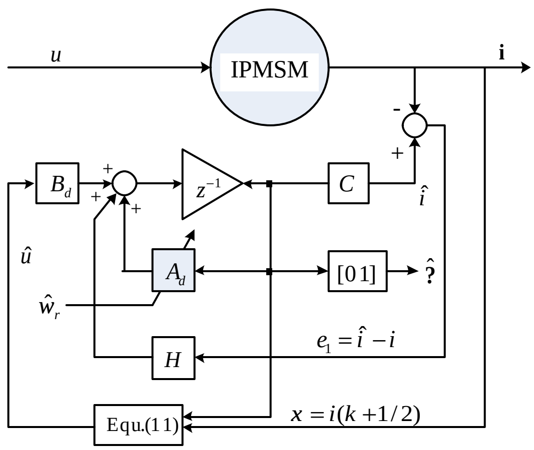

Fig. 1: Block diagram of the unknown input observer

state variable. References literature (Hasegawa *et al*., 2009; Mita and Chida, 1988) for details, unknown input observer can be obtained:

$$\begin{pmatrix}\hat{i}(k+1)\\ \hat{\lambda}(k+1)\end{pmatrix} = (A_d - B_d(C\tilde{B}_d)^{-1}C\tilde{A}_d)\begin{pmatrix}\hat{i}(k)\\ \hat{\lambda}(k)\end{pmatrix} + B_d(C\tilde{B}_d)^{-1}\xi + H(\hat{i}(k) - i(k)) \tag{10}$$

Finally, estimated rotor position $\hat{\theta}_r$ can be given by $\hat{\theta}_r(k) = \angle ?(k)$ and estimated speed is obtained by differential operation of $\hat{\omega}_r = d\hat{\theta}_r/dt$. Figure 1 shows the block diagram of the unknown input observer.

This system becomes observable if $\omega_r \neq 0$, in other words, the proposed method cannot be applied to drive the IPMSM at standstill. The motor starting also needs high-frequency signal injection method.

DESIGN OF IPMSM HIGH-FREQUENCY SIGNAL INJECTION METHOD

High frequency injection is based on the principle of the rotor salient pole track, the basic idea is to inject an extra voltage or current excitation signal and determine the salient pole position of the rotor by detecting the corresponding response signal, then achieve the estimate of the rotor position and speed. High frequency injection including rotating high-frequency voltage signal injection method and pulsating high-frequency voltage signal injection method. In this study, use the pulsating voltage injection method.

Assumed θ_r be unknown to be estimated rotor position, $\hat{\theta}_r$ is known, injection the high-frequency pulsating voltage signal to the d-axis of rotor estimated coordinates, vector form as:

$$u_c^s = u_c \cos\omega_c t \tag{11}$$

where, ω_c be AC voltage frequency, u_c be voltage amplitude, ω_c much larger than the rotor rotational speed ω_r.

The Eq. 11 is transformed to the d-q coordinates, can be obtained:

$$u_c^d = u_c \cos(\omega_c t)e^{j(\hat{\theta}_r - \theta_r)} \tag{12}$$

This voltage vector modulated by the rotor saliency and inevitably be reflected in the stator current response. Now, study the current response.

Stator voltage equation is:

$$u_s^{dq} = L_{dq}\frac{di_s^{dq}}{dt} \tag{13}$$

Injection voltage u_{dc} is applied in the Eq. 13 and solved the current response i^{dq}_c:

$$i_c^{dq} = \frac{u_c}{\omega_c}\sin\omega_c t\left(\frac{1}{L_d}\cos(\hat{\theta}_r - \theta_r) + j\frac{1}{L_q}\sin(\hat{\theta}_r - \theta_r)\right) \tag{14}$$

The Eq. 14 is the current response of the injected signal. When $\theta_r \to \hat{\theta}_r$ mean that the rotor estimated value gradually approaches the actual value while the harmonics torque generated by the quadrature-axis current in Eq. 14 is gradually reduced until it reaches zero. Because only can detect the stator three-phase current, so need transformed the Eq. 14 into the stationary ABC coordinates, obtained:

$$\begin{aligned} {}_c^s &= i_c^{dq}e^{j\theta_r} \\ &= \frac{-ju_c}{4\omega_c L_d L_q}\Big\{\Big((L_d + L_q)e^{j(\omega_c t + \hat{\theta}_r)} - (L_d - L_q)e^{j(\omega_c t + \hat{\theta}_r - 2\gamma)}\Big) \\ &\quad + \Big(-(L_d + L_q)e^{j(-\omega_c t + \hat{\theta}_r)} + (L_d - L_q)e^{j(-\omega_c t + \hat{\theta}_r - 2\gamma)}\Big)\Big\} \\ &= i_p + i_n \end{aligned} \tag{15}$$

where, γ is the angular difference between rotor estimated coordinates and actual coordinates. In i_p contains rotor speed information ω_r, the rotor position information can be obtained by demodulating.

High-frequency carrier current demodulation: Take the Eq. 15 transformed into the forward rotation coordinates which rotating at $i\omega_c t + \hat{\theta}_r$ from the stationary ABC coordinates, equivalent to the Eq. 15 multiplied by $e^{-j(\omega_c t + \hat{\theta}_r)}$, can be written as:

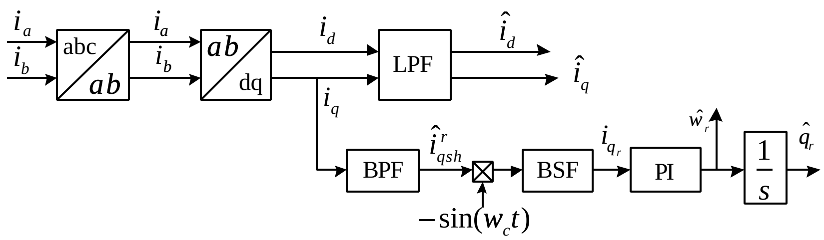

Fig. 2: Position and speed estimated schematic diagram

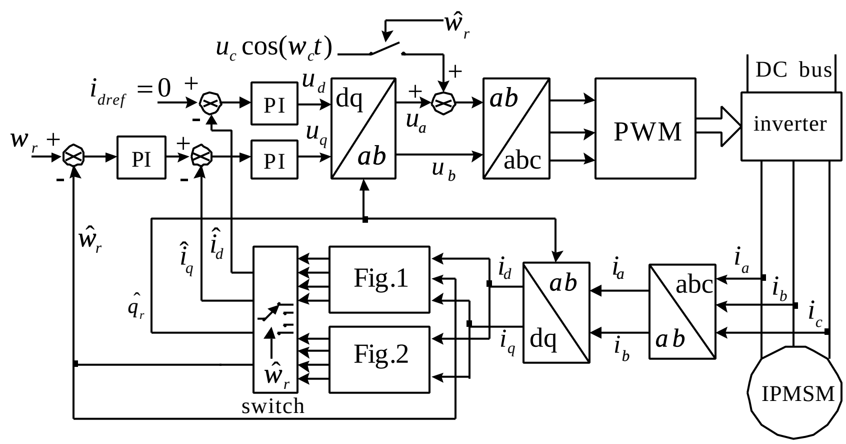

Fig. 3: Sensorless control system block diagram

$$i_p e^{-j(\omega_c t+\hat{\theta}_r)} = \frac{-ju_c}{4\omega_c L_d L_q}\left((L_d+L_q)-(L_d-L_q)e^{j(2\theta_r-2\hat{\theta}_r)}\right) \quad (16)$$

$$i_n e^{-j(\omega_c t+\hat{\theta}_r)} = \frac{ju_c}{4\omega_c L_d L_q}\left((L_d-L_q)e^{j(-2\omega_c t-2\gamma)}-(L_d+L_q)e^{j(-2\omega_c t)}\right) \quad (17)$$

Equation 17 is the high-frequency current component that can be filtered off through low pass filter LPF. The Eq. 16 contains position information $2\Delta\theta_r = 2(\theta_r - \hat{\theta}_r)$, can be rewritten according to the Euler equation and taken the real part as:

$$i_{\theta_r} = \frac{-u_c}{4\omega_c L_d L_q}\sin(2\Delta\theta_r) \approx -\frac{1}{2}\frac{u_c}{\omega_c}\frac{L_d-L_q}{L_d L_q}\Delta\theta_r \quad (18)$$

It can be seen from the Eq. 18 that i_{θ_r} is proportional to the rotor position estimation error $\Delta\theta_r$ when $\Delta\theta_r$ is small. If adjust the q-axis high-frequency current component i_{θ_r} to zero by a certain method, then the rotor position estimation error to zero, that is, the estimated rotor position is the true rotor position.

The position and speed estimated schematic diagram shown in Figure 2 by the method of filtering and PI regulator.

CONTROL SYSTEM SIMULATION RESULTS

In order to verify the performance of the proposed control method, simulation experiments was done in the Matlab7.0 Simulink platform. In simulation, take the idref = 0 control strategy, the entire control system block diagram shown in Fig. 3. While the system is running, start and low-speed stage the switch turn on to the high-frequency signal injection method, high-speed stages the switch is turned on to unknown input observer. The conversion between the switch controlled by estimating the speed $\hat{\omega}_r$, the injection of the high-frequency signal or not was controlled also by estimating the speed $\hat{\omega}_r$.

Figure 4 shows the rotor speed and stator current simulation curve of permanent magnet synchronous motor from non-load starting to settings; Figure 5 shows the torque response curve and the dq-axis current simulation curve of the stator of permanent magnet

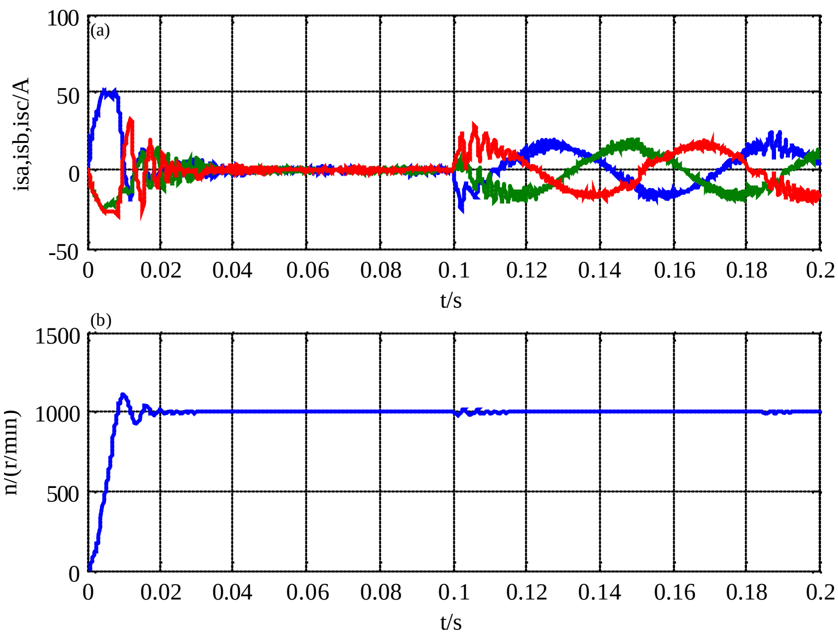

Fig. 4(a-b): Stator current curves and speed response curve (a) Stator current curves and (b) Speed response curve

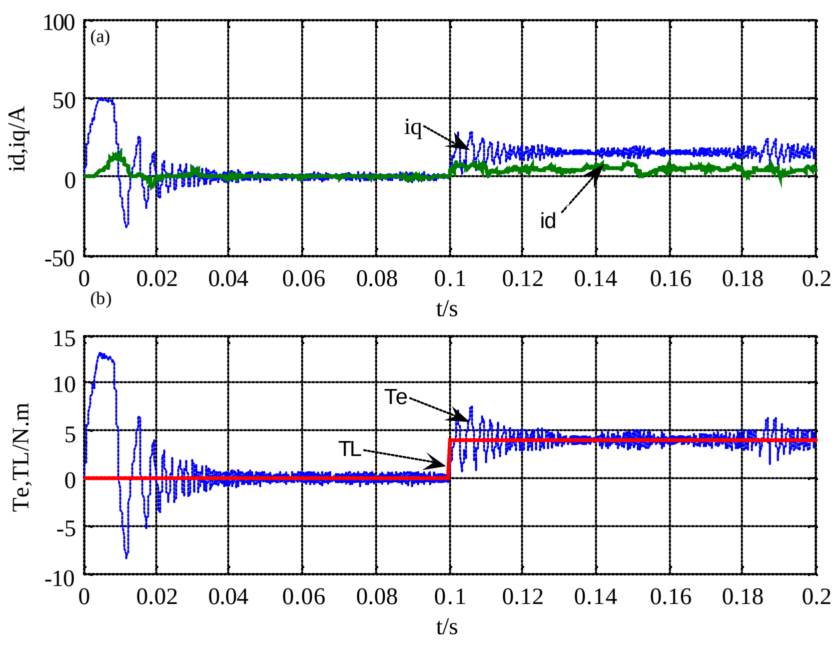

Fig. 5(a-b): dq-axis current and the torque response curve (a) dq-axis current curve and (b) Torque response curve

synchronous motor from non-load starting to settings. When simulation, added 4 N·m load to the motor in 0.1 sec suddenly.

As can be seen from the simulation, the entire system has very good dynamic performance and steady-state performance.

CONCLUSION

This study studied the position sensorless vector control system of permanent magnet synchronous motor based on the pulsating high-frequency voltage signal injection method and unknown input observer. The simulation results show that this method can effectively estimate the position and speed in full speed range of permanent magnet synchronous motor rotor. There are torque ripple in the conversion process, ie, need an algorithm to effectively solve the fluctuations in the switch between the two methods so that, minimizing the impact of this torque ripple.

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