Sensorless Vector Control of Closed-slot Induction Machines at Low Frequency

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Sensorless vector control for induction motor drives is widely used. However, in the case of mass-produced closed-slot motors, it is difficult to operate them at zero frequency with regenerative load. In this paper, a sensorless vector control method based on three-phase PWM carrier waves is applied to a mass-produced closed-slot induction motor. Further, methods to improve stability are proposed.

Keywords: induction machine, sensorless control, zero frequency, closed slot

1. Introduction

Sensorless vector control methods for induction motor drives have been proposed^{(1)–(7)}. Most of them are based on the information of the electromotive force, EMF^{(1)–(4)}. These methods are not able to drive induction machines at zero frequency continuously, because the EMF becomes zero.

Some methods are based on the saliency which is caused by the magnetic saturation for open slot or semi-closed slot induction machines^{(5)–(6)}. Usually, the saliency is detected by injecting high frequency voltages. However, it is difficult for mass-produced closed slot induction motor to detect the saliency. Only a few methods succeed to drive the closed slot motors⁽⁷⁾. Saturation effects in induction machines are associated with the main flux created by the magnetizing current and with leakage flux created by slot currents. With open and semi-closed rotor slot, the leakage flux is not significant. With closed rotor slot, rotor current will tend to drive the slot bridges much further into saturation under loaded conditions. These effects were described in detail in Ref. (6).

In this paper, a sensor-less vector control method based on three phase PWM carrier waves is applied to a mass-produced closed slot induction motor, and stability improvement methods are proposed.

2. Flux Position Estimation

For interior permanent magnet synchronous machines sensor-less vector control at low speeds, high frequency voltage injecting methods have been proposed. These methods utilize magnetic saliency. When induction machines have magnetic saliency because of saturation, these methods are applicable. However, mass-produced closed slot induction machines have weak saliency, so some methods fail to drive them.

In this paper, the method proposed in Ref. (8) based on three phase PWM carrier waves is applied. When the three

phase carrier waves for the PWM pulse generation are chosen as shown in Fig. 1, the carrier frequency components are injected automatically into the stator voltages.

The voltage equation for the PWM carrier frequency component is expressed by the following equation in the α - β stationary reference frame.

$$\begin{bmatrix} v_{\alpha h} \\ v_{\beta h} \end{bmatrix} = \begin{bmatrix} L_0 + L_1 \cos 2\theta & L_1 \sin 2\theta \\ L_1 \sin 2\theta & L_0 - L_1 \cos 2\theta \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{\alpha h} \\ i_{\beta h} \end{bmatrix} \quad \dots\dots\dots (1)$$

where $L_0 = (L_m + L_t)/2$, $L_1 = (L_m - L_t)/2$, L_m : flux axis self-inductance, L_t : torque axis self-inductance, θ : Flux position. The sub-script h means the PWM carrier frequency component.

When the three phase triangular PWM carrier waves are used, the PWM carrier frequency components of the armature voltage are expressed by the following equation.

$$\begin{bmatrix} v_{\alpha h} \\ v_{\beta h} \end{bmatrix} = V_h \begin{bmatrix} \cos \omega_h t \\ \sin \omega_h t \end{bmatrix} \quad \dots\dots\dots (2)$$

Substituting Eq. (2) into Eq. (1) and solving about the current under steady state condition, the PWM carrier frequency component of the armature current, $i_{\alpha h}$, is expressed by the following equation.

Fig. 1. Three phase triangular PWM carrier waves and current measurement points

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$$i_{\alpha h} = \frac{V_h \{(L_0 - L_1 \cos 2\theta) \sin \omega_h t + L_1 \sin 2\theta \cdot \cos \omega_h t\}}{\omega_h (L_0^2 - L_1^2)} \dots \dots \dots (3)$$

The stator current is measured at the points shown by arrows in Fig. 1. The u-phase current is detected twice in a cycle of the PWM carrier waves at $\omega_h t = 0$ and $\omega_h t = \pi$. Therefore, the u-phase current at these two points are expressed by the following equation.

$$i_u = i_{uf} \pm \frac{V_{h1}}{\omega_h (L_0^2 - L_1^2)} L_1 \sin 2\theta \dots \dots \dots (4)$$

where i_{uf} is the fundamental component of the u-phase current, V_{h1} is the transferred value of V_h into the u-v-w three phase axis.

The difference between the u-phase currents at $\omega_h t = 0$ and $\omega_h t = \pi$ becomes

$$\Delta i_{uh} = i_u(u+) - i_u(u-) = \frac{2V_{h1}}{\omega_h (L_0^2 - L_1^2)} L_1 \sin 2\theta \dots \dots (5)$$

where $i_u(u+)$ and $i_u(u-)$ are the u phase current at the points u+ and u- shown in Fig. 1.

The following equations are obtained about the phase v and w with the same procedures as the phase u.

$$\Delta i_{vh} = \frac{2V_{h1}}{\omega_h (L_0^2 - L_1^2)} L_1 \sin 2 \left(\theta - \frac{2\pi}{3} \right) \dots \dots \dots (6)$$

$$\Delta i_{wh} = \frac{2V_{h1}}{\omega_h (L_0^2 - L_1^2)} L_1 \sin 2 \left(\theta + \frac{2\pi}{3} \right) \dots \dots \dots (7)$$

Fig. 2. Block diagram of angular velocity calculation

From Δi_{uh} , Δi_{vh} , and Δi_{wh} , the following equation is obtained.

$$\begin{bmatrix} \Delta i_{\alpha h} \\ \Delta i_{\beta h} \end{bmatrix} = \begin{bmatrix} 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} \Delta i_{uh} \\ \Delta i_{vh} \\ \Delta i_{wh} \end{bmatrix} = \frac{3}{2} \frac{V_{h1}}{\omega_h (L_0^2 - L_1^2)} \begin{bmatrix} \sin 2\theta \\ \cos 2\theta \end{bmatrix} \dots \dots \dots (8)$$

Therefore, the estimated flux position, $\hat{\theta}$, can be calculated through the ratio of $\Delta i_{\alpha h}$ and $\Delta i_{\beta h}$ of Eq. (8) and the arctan function as follows.

$$\hat{\theta} = \frac{1}{2} \tan^{-1} \left(\frac{\Delta i_{\alpha h}}{\Delta i_{\beta h}} \right) \dots \dots \dots (9)$$

The motor angular velocity, $\hat{\omega}_r^c$, is calculated as shown in Fig. 2. The stator angular frequency, ω_1^c , is obtained from the flux position, $\hat{\theta}$, by using the PLL technique. The motor angular velocity is obtained by subtracting the slip angular frequency, ω_{sl} , as follows.

$$\hat{\omega}_r^c = \omega_1^c - \omega_{sl} \dots \dots \dots (10)$$

The slip angular frequency is calculated by using the torque current component, i_t^* , and the magnetizing current component, i_m^* , as follows.

$$\omega_{sl} = \frac{i_t^*}{\tau_2 \left(\frac{1}{1 + \tau_2 p} \right) i_m^*} \dots \dots \dots (11)$$

where $\tau_2 = \frac{L_2}{R_2}$, $p = \frac{d}{dt}$.

The block diagram of a control system is shown in Fig. 3.

3. Flux Strengthening Control

Ratings of the tested induction machine are shown in Table 1. The tested machine is a mass produce closed slot motor made by Fuji Electric Co.

Fig. 4 shows experimental results of speed control with the proposed method. The torque of the load machine connected with the tested induction machine is controlled. The speed command is constant at 10 rad/s (3.3% of the rated speed).

Fig. 3. Block diagram of proposed sensorless vector control system

Table 1. Ratings of tested induction machine

Output Power	2.2 kW	Stator Voltage	200 V
Stator Current	9.2 A	Poles	4
Frequency	50 Hz	Revolution	1430 min ⁻¹

(a) powering mode

(b) regenerating mode

Fig. 4. Experimental result for sensorless speed control
Speed command: 10 rad/s (Electric), Load torque: From
60% to 65%, Step change

The PWM carrier frequency is set at 1,200 Hz, which is relatively low for precise measurements of the phase current differential. The system is operating stably when the load torque is 60% of the rated value. However, it becomes unstable when the load torque increases to 65%. The reason of the instability is thought that the flux position estimation is affected by the torque current component and the estimated position becomes incorrect. Fig. 5 shows the loci of $\Delta i_{\alpha h}$ and $\Delta i_{\beta h}$ under no load and 60% load conditions. The radius of the loci for the 60% load torque is much larger than that for no load. It means that the torque current component affects the inductance value.

To solve the instability, we propose the flux strengthening control. Fig. 6 shows the mutual inductance and the rotor flux linkage calculated from no-load test and locked rotor test. The rated (100%) points are marked in the figures. By strengthening the flux level, the saturation level becomes deeper. So the influence of the torque current component becomes weak. Considering that the system is stable at 60% load torque with 100% magnetizing current, we tried to increase the magnetizing current to 150%, marked in Fig. 6, for full load torque operation.

Fig. 7 shows the experimental results with the flux strengthening control. The magnetizing current component is increased to 150% of the rated value. The speed command

(a) No Load

(b) 60% Load Torque (regenerative)

Fig. 5. $\Delta i_{\alpha h}$ and $\Delta i_{\beta h}$ at 100% Magnetizing Current, 10 rad/s

(a) Mutual Inductance

(b) Rotor Flux Linkage

Fig. 6. Mutual inductance and flux linkage vs. magnetizing current

is 10 rad/s and the load torque is 100%. The system is operating stably under powering condition with full load torque.

4. Regenerating Operation

When the induction machines are coupled with regenerative load, the synchronous speed becomes lower than the motor speed. So, the stator frequency might be zero. Electromotive force based methods are not applicable to zero frequency condition with regenerative load continuously.

Fig. 7. Experimental results of speed control with flux strengthening control, powering mode, Speed command: 10 rad/s (Electric) Load torque: 100%

Fig. 8. Experimental results of speed control with flux strengthening control, regenerating mode, Speed command: 10 rad/s (Electric) Load torque: 100%

Fig. 8 shows experimental results for speed control. The speed command is 10 rad/s and the load torque is 100%, which is regenerative torque. The motor speed fluctuates. The synchronous speed or the stator frequency is zero for a while. At zero frequency, the torque current component is increasing, and the actual speed reaches the speed command (around 10 s, and 30 s). However, estimated speed is decreasing during zero frequency. From this experiment, it is found that the rated torque can be generated, but the estimated speed is incorrect at zero frequency with regenerative load. So it is difficult to control the motor speed precisely around zero frequency.

Considering above phenomenon, we propose variable gain PI speed controller. Fig. 8 indicates that the integral action of the PI speed controller triggers the fluctuation. So, the gain of the integral component of the speed controller is reduced around zero frequency as shown in Fig. 9. The threshold value, $\Delta\omega$, is set at 3 rad/s, which is decided experimentally.

Fig. 10 shows experimental results with the variable PI gain. The speed command and the load torque are the same as those of Fig. 8. Although there is small steady state error between the speed command and the actual value, the system is operating stably. The proposed system generates rated torque at zero frequency successfully.

Fig. 11 shows experimental results for speed step responses. At first, the speed command is set at 10 rad/s, the stator frequency and the integral gain are about zero. After the speed command is changed, the integral gain increases quickly.

Fig. 9. Integral gain of speed controller around zero frequency

Fig. 10. Experimental results of speed control with variable gain, Speed reference: 10 rad/s Load torque: 100%

(a) angular frequency and speed

(b) integral gain

Fig. 11. Experimental results of speed control with variable gain, Speed reference: 10 rad/s to 15 rad/s Load torque: 85%

5. Conclusion

Sensorless vector controls for induction machines injecting high frequency components have been proposed. These methods are easily applied to open slot and semi-closed machines. However, it is difficult to apply them to the closed slot machine, because the torque current components heavily

affect the magnetic saliencies.

In this paper, a sensorless vector control method utilizing high frequency components based on the three phase PWM carrier waves is applied to the closed slot induction machine. To solve some instabilities, the flux strengthening and variable integral gain of the speed controller are introduced. The validity of the proposed method is verified experimentally.

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