We propose a novel speed sensorless vector control method for induction motors, which is applicable to industrial drive systems. This method calculates the flux axis angular velocity and rotor speed of an induction motor from the induced voltages, without using any delay components. According to this method, the system can directly estimate an accurate motor speed, and precisely generate torque.

In addition, this system can smoothly restart operation after an instantaneous power interruption.

We show the validity of the proposed control method through several experiments.

key words: speed sensorless control, vector control, speed estimation, induction motor

1. Introduction

In recent years, several methods for speed sensorless vector control have been proposed (1)-(7). Some of them have been used in industrial drives.

There are three major requirements in applying these methods to motor control in industrial plants. First, these controllers must provide high control performance for accuracy of torque and speed. Second, when applied to a process line, they need to keep the speed of the line as constant as possible. A constant speed is demanded even after an instantaneous power interruption. Third, it must be taken into account that some motors are mechanically connected, and so it is necessary for the motor to start smoothly from the condition in which the motor is passively driven by other motors.

As for the first requirement, various solutions have been proposed and put into practical use. On the other hand, it seems to be difficult to meet the second and the third requirements. This difficulty is mainly caused by delay components used in the speed estimation.

In this paper, we propose a novel method for speed sensorless vector control applicable to industrial drive systems. We focus on a method for immediate and accurate estimation of the flux axis angular velocity and the rotor speed.

We first introduce the estimation scheme based on the induced voltages. The velocity is mainly estimated with a torque component of the induced voltages, and is corrected with a magnetizing component of them. Then, we show the validity of this proposed method through several experiments.

2. Control Method

<2-1> Configuration of Speed Sensorless Controller

Fig. 1 shows a control block diagram of the speed sensorless vector control method proposed in this paper.

The reference value of flux axis angular velocity $\omega^*_f$ is calculated from the induced voltages. The reference value of the flux angle $\theta^*$ is obtained by integrating $\omega^*_f$. The vector rotators transform coordinates based on $\theta^*$.

The symbols in Fig. 1 are defined as follows.

$v_M$ : magnetizing component of output voltage,
$v_T$ : torque component of output voltage,
$i_M$ : magnetizing component of stator current,
$i_T$ : torque component of stator current,
$e_M$ : magnetizing component of rotor induced voltage,
$e_T$ : torque component of rotor induced voltage,
$\omega^*_r$ : reference value of rotor speed,
$\hat{\omega}_r$ : estimated value of rotor speed,
$\omega^*_f$ : reference value of flux axis angular velocity,
$\omega_f$ : real value of flux axis angular velocity,
This control system employs the output voltage estimator instead of using output voltage sensors. The estimator calculates the output voltages from the reference values of the output voltages by compensating the errors caused by the inverter.

With this control method, low speed operation causes a problem with control accuracy, because the output voltage is very low. But since the main theme of this paper is the restarting operation, we do not mention the low voltage problems here.

(2.2) Instantaneous velocity calculation

Fig. 2 shows a vector chart of current, induced voltage, and rotor flux. M-axis corresponds to the flux axis, and T-axis the torque axis. α-β coordinates are based on stator axes.

The rotor induced voltage, \( e = (e_M, e_T)^T \), is described by the flux axis angular velocity, \( \omega_1 \), and the rotor flux, \( \phi_2 \), as (1),

\[
\begin{align*}
\epsilon_M &= p\phi_2M - \omega_1\phi_T \\
\epsilon_T &= \omega_1\phi_2M + p\phi_2T
\end{align*}
\]

where \( p \) is a differential operator.

On the other hand, the induced voltage, \( e \), is calculated from voltages and currents by (2),

\[
\begin{align*}
e_M &= v_M - (R_1 + pL_\sigma)i_M + \omega_1L_\sigma i_T \\
e_T &= v_T - (R_1 + pL_\sigma)i_T - \omega_1L_\sigma i_M
\end{align*}
\]

where

\[
\begin{align*}
R_1 & : \text{primary resistance}, \\
L_\sigma & : \text{leakage inductance}.
\end{align*}
\]

When the direction of M-axis coincides with the direction of the rotor flux \( \phi_2 \), the magnetizing component of the induced voltage \( e_M \) equals 0, and the rotor flux components satisfy \( \phi_2T = 0 \), and \( \phi_2M = \phi_2 \).

From the second equation in (1), we have the flux axis angular velocity as (3),

\[
\omega_1 = \frac{e_T}{\phi_2}
\]

So we can calculate the reference value of flux axis angular velocity \( \omega^*_1 \) by (4).
where, $\phi_2$ is an estimated value of the rotor flux given by the flux calculator in Fig. 1. When the speed is relatively high, the flux calculator obtains $\phi_2$ from $e_T$ and $\omega_1$, but when the speed is low, it uses $\phi_2^*$ as $\phi_2$.

Then, the flux angle of controller $\theta^*$ is obtained by (5).

$$\theta^* = \int \omega_1^* dt$$

Next, we consider the case in which there is an error between the flux angle of the controller and that of the induction motor. Fig. 3 shows a vector chart in this case. M–T coordinates are based on $\theta^*$, and M'–T' coordinates are based on $\theta$, where $\theta$ is the flux angle of motor.

From Fig. 3, when $\theta^*$ advances more than $\theta$, $e_M$ is positive, and when $\theta$ advances more than $\theta^*$, $e_M$ is negative.

The sign of $\omega_1^*$ coincides with the sign of $e_T$ because of (3). So, using $e_M$, we can compensate $\omega_1^*$ error as (6),

$$\omega_1^* = \text{sgn}(e_T) \left[ \frac{|e_T|}{\phi_2} - K_{PEM} \frac{\phi_2^*}{\phi_{20}} e_M \right]$$

where $K_{PEM}$ is the compensation gain by $e_M$.

The first term in the brackets in (6) corresponds to the main component of estimated value of the flux axis angular velocity. The second term compensates for an error between the flux angle of the controller $\theta^*$ and that of the induction motor $\theta$. The normalized flux reference ($\phi_2^* / \phi_{20}$) maintains the stability of the system even in the field-weakening operation range.

Fig. 4 shows a block diagram of the flux axis angular velocity reference calculator. It should be noted that this part contains no delay components such as integrators. Therefore, $\omega_1^*$ can be estimated without delay.

<2.3> Effects of the Variation of the Secondary Resistance

The flux axis angular velocity reference calculator in Fig. 4 can yield true $\omega_1$ independent of the variation of the secondary resistance.

$$\omega_1 = \omega_1^* - \omega_r = \omega_1^* - \frac{R_2}{\phi_2} e_T$$

$$\omega_r = \omega_1 - \omega_r = \omega_1 - \frac{R_2}{\phi_2}$$

where $R_2^*$ and $R_2$ are setting and actual value of secondary resistance, respectively.

In the steady state, $\phi_2^*$ converges to $\phi_2$, and $\omega_1^*$ converges to $\omega_1$. So subtracting (8) from (7) gives an estimation error caused by the secondary resistance variation as shown by (9).

$$\omega_1^* - \omega_r = \frac{R_2 - R_2^*}{R_2^*} \omega_1^* + (\omega_1^* - \omega_1)$$

In the case of an induction motor, the slip angular velocity is much smaller than the rated rotor speed. So the estimation error due to $R_2$ variation is very small, when compared with the base speed.

3. Experimental Results

<3.1> Experimental Equipment

Experiments have been performed with an industrial-use transistor inverter system and an induction motor (2.2 kW, 160 V, 4 poles, base speed:1500 r/min, top speed:3000 r/min, rated slip:3%). The carrier frequency of the inverter is 2 kHz. Arithmetic operations for current regulation and vector control are executed by a digital signal processor (DSP).

<3.2> Four-quadrant operation Characteristics

Fig. 5 shows the four-quadrant operation characteristics. Both acceleration and deceleration time to and from a top speed are 2 seconds. In the field-weakening operation range from 1500 r/min to 3000 r/min, the flux reference $\phi_2^*$ is inversely proportional to estimated value of rotor speed $\omega_r$. This method exhibits stability in the field-weakening operation area as well as smoothness at the crossing of zero speed.
<3-3> Transient Characteristics

Fig. 6 shows the response to a speed step. This system demonstrates a high speed response of about 50 rad/s. It shows that the estimated speed follows the actual speed without delay.

Fig. 7 shows the response to a load step. In this experiment, the induction motor runs in a speed-control mode (speed reference value: 1500 r/min), and a load motor gives 100% rated torque for 15 seconds. This system can operate in a stable manner under such a torque variation. In this case again, the estimated speed follows the actual speed without delay.
<3-4> Restarting Operation

Fig. 8 shows the experimental result for the restarting operation after an instantaneous interruption.

In Fig. 8, the test motor is running at speed-control operation (speed reference value: 1000 r/min, torque limit value: 100%) and the load motor is also running at speed-control operation (speed reference value: 800 r/min, torque limit value: 80%). The torque limit value of the load motor is lower than that of the test motor. The speed can be accurately estimated without any fluctuations in the actual speed.

The experiment for the restarting operation of the system assumed that several motors are mechanically connected together and only some of them suffer a power interruption and restart operation.

The control sequence of restarting after an instantaneous power interruption is described as follows.

First, after an instantaneous power interruption is detected, the inverter stops its operation. In this experiment, we evaluate the restarting performance in the most severe condition where the rotor flux extinguishes completely. So the inverter stops for 2 seconds, which is long enough for the rotor flux to be extinguished.

After that, the inverter starts to supply the magnetizing current. At that time, the controller sets $\omega_r$ equal to $\omega_r$ so that the inverter can restart smoothly.

We know that the resonant frequency of the secondary circuit of the induction motor equals $\omega_r$. Therefore supplying the stepwise magnetizing current induces a voltage that contains the dominant frequency component equal to $\omega_r$. Therefore this system calculates $\omega_r$ and $\omega_r$ from the induced voltage directly.

After supplying the magnetizing current, the inverter starts to supply the torque current. Then, the controller restarts its speed control operation.
Equation (6) shows that the flux axis angular velocity $\omega_i^*$ and the flux angle reference $\theta^*$ are completely insensitive to the variation in secondary resistance. Consequently this controller can generate a constant torque independent of the secondary resistance variation.

Fig. 9 shows the torque characteristics at speed references of 75, 1500 and 3000 r/min. At 75 and 1500 r/min, the flux reference is 100%. At 3000 r/min, the system is in the field-weakening operation, and the flux reference is 50%. Torque-control precision with an error of less than 5% has been achieved over this speed range. The torque linearity of this controller is also satisfactory.

<3.6> Speed Characteristics

Fig. 10 shows the speed characteristics at speed references of 75, 1500 and 3000 r/min.

Fig. 10 tells us that a speed-control precision with an error less than 0.3% has been achieved over this speed range. In the case of 3000 r/min, the flux reference is 50%, and the slip is two times larger than in the case of 75 and 1500 r/min. Therefore, the speed error becomes twice as large.

<3.7> Effects of Temperature

Fig. 11 shows the setup of experimental equipment to measure the effect of temperature.

The tested induction motor is connected to a dc generator, and the induction motor runs in a torque-control mode (torque current reference value : 100%). The rotation is maintained at a constant speed by the load dc generator.

Fig. 12 shows that the torque and speed characteristics under the temperature change. The value of the secondary resistance varies according to the temperature. In this experiment, the temperature of the motor rises from 30°C to 80°C, and the secondary resistance increases about 20%. Assuming that the rated slip is equal to 3%, we can estimate the speed variation is about ±0.6%, and the torque variation is zero.

Fig. 12 shows that the torque variation is almost zero. The speed variation is within ±0.6%.
4. Conclusion

We proposed a speed sensorless vector controller for an induction motor. The flux axis angular velocity and the rotor speed of the motor are directly estimated from the rotor induced voltage. The controller can also estimate the accurate speed even when restarting after an instantaneous power interruption, because it calculates the flux axis angular velocity of the motor without using any delay components. The control method proposed in this paper can provide good torque-control precision and speed-control precision, even under temperature variation.

(Manuscript received January 22, '96, revised May 29, '96)

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