Journal of the Korean Society of Marine Engineers, Vol.27, No.4, pp.503~510, 2003(ISSN 1226-9549)

Sensorless Speed Control of Induction Motor using Current Compensation

Sae-Gin Oh†, Jong-Su Kim*, Sung-Hwan Kim**

(Manuscript : Received MAR 16, 2003 ; Revised JUN 24, 2003)

Key words : sensorless speed control, current compensation, numerical model.

Abstract

A new method of induction motor drive, which requires neither shaft encoder nor speed estimator, is presented. The proposed scheme is based on decreasing current gap between a numerical model and an actual motor. By supplying the identical instantaneous voltage to both model and motor in the direction of reducing the current difference, the rotor approaches to the model speed, that is, reference value. The indirect field orientation algorithm is employed for tracking the model currents.

The performance of induction motor drives without speed sensor is generally characteristic of poorness at very low speed. However, in this system, it is possible to obtain good speed response in the extreme low speed range.

Nomenclature

v_{ds}, v_{qs} : d, q component of stator voltage

i_{ds}, i_{qs} : d, q component of stator current

i_{dr}, i_{qr} : d, q component of rotor current

$\lambda_{dr}, \lambda_{qr}$: d, q component of rotor flux linkage

i_{dref} : reference value of rotor flux current

$v_{\alpha s}, v_{\beta s}$: α, β component of stator voltage

$i_{\alpha s}, i_{\beta s}$: α, β component of stator current

$i_{\alpha r}, i_{\beta r}$: α, β component of rotor current

ω_e : angular speed of $d-q$ reference frame

ω_r : rotor angular speed

ω_{sl} : slip angular speed [$= \omega_e - \omega_r$]

θ_e : angle of rotor flux(d axis) in stationary reference frame

T_e : electromagnetic torque

L_s : stator self-inductance

L_r : rotor self-inductance

† Corresponding Author(Division of Marine Engineering, Graduate School of Korea Maritime University), E-mail : osgengen@hanmail.net, Tel : 051) 410-4193

* Research Institute of Maritime Industry of Korea Maritime University

** Division of Mechatronics Engineering of Korea Maritime University

L_m : mutual inductance
σ : total leakage inductance [$= L_s(1 - L_m^2/L_sL_r)$]
R_s, R_r : stator, rotor resistance
T_r : time constant of rotor circuit [$= L_r/R_r$]
P : pole number
ϕ : angular deviation between motor flux axis and model flux axis
p : derivative operator [$= d/dt$]
m : numerical model variable
$*$, ref : command or setting value
Δ : increment of each quantity
k : constant [$= \frac{3}{2}\frac{P}{2}\frac{L_m^2}{L_r}$]
$K_{mp}, K_{mi}, K_{kp}, K_{ki}$: gain constants

1. INTRODUCTION

In the industrial speed and torque controlled drive system, the closed loop control is usually based on the measurement of speed or position of motor.

However, the mechanical sensor requires a mounting space on the shaft of motor, reduces the reliability and increases the cost of drive system. Therefore, various control algorithms have been proposed for eliminating the speed sensor of induction motor. But, the characteristics of those methods deteriorate as the speed gets lower because of difficulty in estimating the rotor speed.

This paper describes the high performance drive that can be obtained in low speed range from an induction motor without shaft encoder.

The proposed method is on the basis of compensating current difference between the induction motor and its numerical model, in which the identical stator voltage is supplied for both the actual motor and the model so that the gaps between stator currents of the two can be forced to decay to zero as time proceeds.

Consequently, rotor speed approaches to the model speed, namely, setting value and the system can be controlled precisely in low speed range.

These features are verified by the theoretical examination and the experimental results.

2. PRINCIPLES OF PROPOSED METHOD

The voltage equation of induction motor in the stationary reference frame is given by (1).

$$\begin{bmatrix} v_{\alpha s} \\ v_{\beta s} \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_s + pL_s & 0 & pL_m & 0 \\ 0 & R_s + pL_s & 0 & pL_m \\ pL_m & \omega_r L_m & R_r + pL_r & \omega_r L_r \\ -\omega_r L_m & pL_m & -\omega_r L_r & R_r + pL_r \end{bmatrix} \begin{bmatrix} i_{\alpha s} \\ i_{\beta s} \\ i_{\alpha r} \\ i_{\beta r} \end{bmatrix} \tag{1}$$

Using matrix inversion, the voltage equation (1) can be rearranged into a current equation.

$$p\begin{bmatrix} i_{\alpha s} \\ i_{\beta s} \\ i_{\alpha r} \\ i_{\beta r} \end{bmatrix} = \frac{1}{D}\left\{ \begin{bmatrix} -L_rR_s & L_m^2\omega_r & L_mR_r & L_mL_r\omega_r \\ -L_m^2\omega_r & -L_rR_s & -L_mL_r\omega_r & L_mR_r \\ L_mR_s & -L_sL_m\omega_r & -L_sR_r & -L_sL_r\omega_r \\ L_sL_m\omega_r & L_mR_s & L_sL_r\omega_r & -L_sR_r \end{bmatrix} \begin{bmatrix} i_{\alpha s} \\ i_{\beta s} \\ i_{\alpha r} \\ i_{\beta r} \end{bmatrix} + \begin{bmatrix} L_r & 0 & -L_m & 0 \\ 0 & L_r & 0 & -L_m \\ -L_m & 0 & L_s & 0 \\ 0 & -L_m & 0 & L_s \end{bmatrix} \begin{bmatrix} v_{\alpha s} \\ v_{\beta s} \\ v_{\alpha r} \\ v_{\beta r} \end{bmatrix} \right\} \tag{2}$$

where, $D = L_s L_r - L_m^2$

Transforming the equation (1) into the synchronously rotating reference frame, it can be expressed as follows.

$$\begin{bmatrix} v_{ds} \\ v_{qs} \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_s + pL_s & -\omega_e L_s & pL_m & -\omega_e L_m \\ \omega_e L_s & R_s + pL_s & \omega_e L_m & pL_m \\ pL_m & -\omega_{sl} L_m & R_r + pL_r & -\omega_{sl} L_r \\ \omega_{sl} L_m & pL_m & \omega_{sl} L_r & R_r + pL_r \end{bmatrix} \begin{bmatrix} i_{ds} \\ i_{qs} \\ i_{dr} \\ i_{qr} \end{bmatrix} \quad (3)$$

Also, the electromagnetic torque equation is written by (4).

$$T_e = \frac{3}{2}\frac{P}{2}\frac{L_m}{L_r}(i_{qs}\lambda_{dr} - i_{ds}\lambda_{qr}) \quad (4)$$

Under the steady state, rotor flux linkage, slip angular speed and electromagnetic torque on the rotor flux oriented frame are obtained by equations (3) and (4) as below.

$$\lambda_{dr} = L_m i_{ds} \quad (5)$$

$$\omega_{sl} = \frac{1}{T_r}\frac{i_{qs}}{i_{ds}} \quad (6)$$

$$T_e = \frac{3}{2}\frac{P}{2}\frac{L_m^2}{L_r} i_{ds} i_{qs} \quad (7)$$

According to equation (2), induction motor can be regarded as multiple inputs -multiple outputs system, in which inputs are stator voltages and outputs are stator currents and rotor speed, as shown in Fig. 1.

Fig. 1 Input and output variables of induction motor

Here, let's assume a numerical model represented by equation (2), in which inputs are stator voltages, reference rotor speed and outputs are stator currents as Fig. 2.

Fig. 2 Input and output variables of numerical model

When the model describes the actual motor precisely, in case that stator voltages and currents of an induction motor are same as those of its numerical model, the rotor speed would become equal to the model speed, namely, the setting value.

In the synchronously rotating reference frame, if the $d-q$ component voltages and currents, frequencies and rotor flux angles of both the motor and the model are identical simultaneously, the same result as above mentioned can be expected.

Neglecting the transient state, the $d-q$ voltage equations on both sides can be written by (3) and (6) as follows.

$$v_{ds} = R_s i_{ds} - \sigma\omega_e i_{qs} \quad (8)$$

$$v_{qs} = R_s i_{qs} + \omega_e L_s i_{ds}$$
$$= R_s i_{qs} + L_s i_{ds}(\omega_r + \frac{1}{T_r}\frac{i_{qs}}{i_{ds}}) \quad (9)$$

$$v_{dsm} = R_s i_{dsm} - \sigma\omega_{em} i_{qsm} \quad (10)$$

$$v_{qsm} = R_s i_{qsm} + \omega_{em} L_s i_{dsm}$$
$$= R_s i_{qsm} + L_s i_{dsm}(\omega_{rm} + \frac{1}{T_r}\frac{i_{qsm}}{i_{dsm}}) \quad (11)$$

If the rotor flux angle of the motor is

consistent with that of the model, the following equations can be expressed by supplying the identical instantaneous voltages.

$$\sigma(\omega_e i_{qs} - \omega_{em} i_{qsm}) = R_s(i_{ds} - i_{dsm}) \tag{12}$$

$$L_s(\omega_{rm} i_{dsm} - \omega_r i_{ds}) = (R_s + \frac{L_s}{T_r})(i_{qs} - i_{qsm}) \tag{13}$$

In the above equation (12) ω_e is equal to ω_{em} due to the same stator voltages.

From (12) and (13), we know that ω_r agrees with its setting value ω_{rm} and i_{ds} with i_{dsm} if i_{qs} can be controlled in accordance with i_{qsm}.

Then, how can i_{qs} approach to i_{qsm} as supplying the same voltages for both actual motor and numerical model?

At first, it is essential to define the control method for the numerical model in order to calculate model currents.

In this paper, we employ indirect field orientation algorithm characterized by the speedy performance.

Since the drive gets started, the model flux current i_{dsm} remains at the same quantity as the setting value i_{dref} by controlling d component voltage.

Here, let's assume that the model speed becomes higher than the actual rotor speed. In this case slip speed, torque current, flux current and electromagnetic torque of the motor surpass those of the numerical model as shown in the above equations (6), (7) and (12). Raising the q component voltage by Δv_{qs} $(= \Delta v_{qsm})$ for both sides, the increment of the model q component current is expressed as (14) under the steady state with i_{dsm} unchanged by controlling the model flux current constantly by equation (11).

$$\Delta i_{qsm} = \frac{1}{R_s + \frac{L_s}{T_r}} \Delta v_{qsm} \tag{14}$$

In the actual motor, the following equation (15) is derived from (9).

$$\Delta i_{qs} = (R_s + \frac{L_s}{T_r})\Delta i_{qs} + \omega_r L_s \Delta i_{ds} + L_s \Delta i_{ds} \Delta\omega_r + L_s i_{ds} \Delta\omega_r \tag{15}$$

Investigating (15) sequentially, the instant that the q component voltage is stepped up, the rotor speed remains unchanged. But, the torque current rises in a moment, which increases the electromagnetic torque and the rotor speed goes up gradually. Then, the rotor speed has access to the reference value, and i_{qs}, i_{ds} are settled at new steady point.

After all, the increment of the motor torque current is rewritten as (16) with neglecting infinitesimal term.

$$\Delta i_{qs} = \frac{1}{R_s + \frac{L_s}{T_r}} (\Delta v_{qs} - \omega_r L_s \Delta i_{ds} - L_s i_{ds} \Delta\omega_r) \tag{16}$$

As we know from equations (10) and (8), Δv_{ds} $(= \Delta v_{dsm})$ to make the model flux current constant is decreased by the increase of ω_{em} and i_{qsm}, then i_{ds} is also reduced.

According to above equations (12), (14) and (16), as the following inequality (17) is effective under the steady state after current compensation, $\Delta i_{qs} < \Delta i_{qsm}$, which is considered evident in low and medium speed range with a relative great flux current, would make two torque currents approachable.

$$-\frac{\Delta\omega_r}{\omega_r} < \frac{\Delta i_{ds}}{i_{ds}} < 0 \qquad (17)$$

On the contrary, as the rotor speed gets higher than the setting speed, the reverse process in contrast to what above mentioned can be carried out by lowering $\Delta v_{qs}\,(=\Delta v_{qsm} < 0)$.

Now, let's be concerned about the possibility of the case that the rotor flux angle of the actual motor deviates from that of the model. Under the steady state, $i_{drm}=0$ and $i_{dr}=0$ simultaneously by the result of equation(5). If the axis of the motor can go in advance that of the model by "ϕ", the following equations are effectuated by the same stator voltages and the current compensation with reference to "model" q axis.

$$\begin{bmatrix} v_{ds} \\ v_{qs} \end{bmatrix} = \begin{bmatrix} \cos\phi & \sin\phi \\ -\sin\phi & \cos\phi \end{bmatrix} \begin{bmatrix} v_{dsm} \\ v_{qsm} \end{bmatrix} \qquad (18)$$

$$i_{qs} = [\,-\sin\phi \quad \cos\phi\,][\,i_{dsm} \quad i_{qsm}\,]^T \qquad (19)$$

Applying above equations to (8) ~(11), the motor flux current is as (20).

$$i_{ds} = [\,\cos\phi \quad \sin\phi\,][\,i_{dsm} \quad i_{qsm}\,]^T \qquad (20)$$

Then, both currents with reference to the "model" d axis also become equal to reach the result as Fig. 3.

In this case, rotor currents should be also maintained as (21) by the first and second rows of equation (3).

$$\begin{bmatrix} i_{dr} \\ i_{qr} \end{bmatrix} = \begin{bmatrix} \cos\phi & \sin\phi \\ -\sin\phi & \cos\phi \end{bmatrix} \begin{bmatrix} i_{drm} \\ i_{qrm} \end{bmatrix} \qquad (21)$$

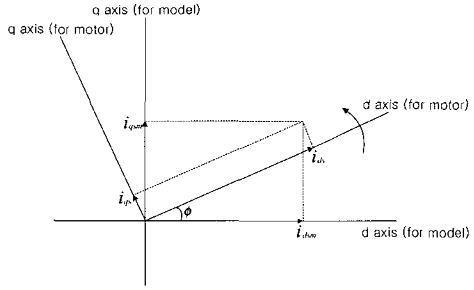

Fig. 3 Hypothetical deviation of motor flux axis from model flux axis

Where, $i_{drm}=0$, however, if $\phi \neq 0$ under the steady state, $i_{dr} \neq 0$ and that would be inconsistent with above mentioned condition. Consequently, the assumption that the rotor flux angle of the actual motor could deviate from that of the model under the steady state is considered to be an incorrect notion.

Therefore, when both currents with reference to "model" q axis are controlled to be equalized stably, all values of both sides also become equal.

In conclusion, it is practicable to construct a simple control algorithm for both the numerical model and the actual motor as below.

$$\begin{bmatrix} v^*_{ds} \\ v^*_{qs} \end{bmatrix} = \begin{bmatrix} K_{mp}(i_{ref} - i_{dsm}) + K_{mi}\int(i_{ref} - i_{dsm})dt \\ K_{tp}(i_{qs} - i_{qsm}) + K_{ti}\int(i_{qs} - i_{qsm})dt \end{bmatrix} \qquad (22)$$

$$\begin{bmatrix} v^*_{\alpha s} \\ v^*_{\beta s} \end{bmatrix} = \begin{bmatrix} \cos\theta_{em} & -\sin\theta_{em} \\ \sin\theta_{em} & \cos\theta_{em} \end{bmatrix} \begin{bmatrix} v^*_{ds} \\ v^*_{qs} \end{bmatrix} \qquad (23)$$

$$\theta_{em} = \int(\omega_{rm} + \frac{1}{T_r}\frac{i_{qsm}}{i_{ref}})dt \qquad (24)$$

Fig. 4 is a configuration of the proposed system, which consists of a digital controller, a current-controlled transistor inverter and a squirrel cage induction motor.

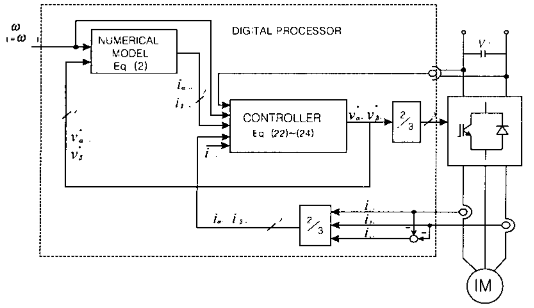

Fig. 4 The block diagram of the proposed system

3. EXPERIMENTAL RESULTS

The system was implemented in a 2.2 Kw PWM inverter fed induction motor drive. The machine data : 380V, 5.2A, 4 poles, 1720rpm. A dynamometer was used as the load.

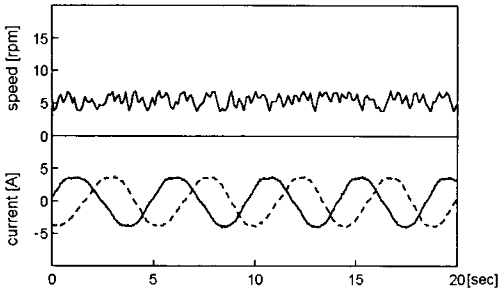

Fig. 5 Constant speed operation at 5rpm(0.16Hz) at 50% nominal load : speed and phase currents

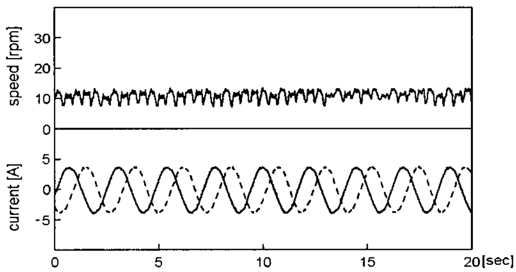

Fig. 6 Constant speed operation at 10rpm(0.33Hz) at 50% nominal load : speed and phase currents

Fig. 5 shows 5[rpm] speed operation in the steady state at 50% nominal load. The stator currents are exactly sinusoidal and the commanded speed is maintained.

Fig. 6 shows the stator currents and the speed response in the steady state at 10[rpm], 50% nominal load. The smooth speed operation is achieved.

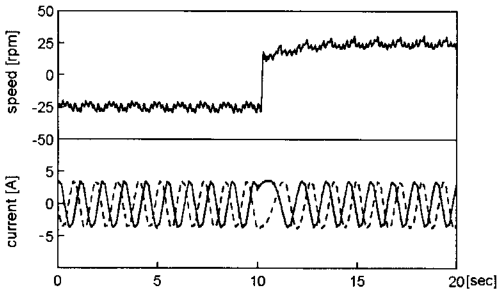

Fig. 7 Reversal of speed between the setpoint values ± 25rpm at 50% constant nominal load : speed and phase currents

Fig. 7 shows the speed performance in a speed reversal process between the set values ± 25[rpm]. The torque is held constant at 50% nominal value while the speed is negative. The dynamic change in the phases of the stator currents is represented.

4. SUMMARY

The performance of vector controlled induction motor drive without speed sensor is generally poor at very low speed. Noise, offset, drift and parameter mismatch lead to speed oscillations and instabilities.

A new algorithm which requires neither shaft encoder nor speed estimator is introduced in this paper. The proposed method is based on decreasing differences of stator currents between a numerical model and an actual motor.

By supplying the same stator voltage to both model and motor toward eliminating currents gaps, the motor speed gets closer to the model speed, reference value. Comparing with other sensorless methods, the scheme has the advantage of no estimating the motor speed and it excludes problems attendant upon calculating speed.

The effectiveness of this method is demonstrated by experiments. Excellent steady state and dynamic performance is achieved even at extreme low speed.

5. REFERENCES

[1] R. Joetten and G. Maeder, Control methods for good dynamic performance induction motor drives based on current and voltage as measured quantities, *IEEE Trans. Indus. Appli.*, vol. 19, No. 3, pp. 356-363, 1983

[2] T. Ohtani, N. Takeda and K. Tanaka, Vector control of induction motor without shaft encoder, *IEEE Trans. Indus. Appli.*, vol. 28, No. 1, pp. 157-164, 1992

[3] H. Tajima, Y. Matsumoto, H. Umida and M. Kawano, Speed sensorless vector control method for an Industrial drive system, *Proc. Intl. Power Elec. Conf.* pp. 1034-1039, 1995

[4] B. K. Bose and M. G. Simoes, Speed sensorless hybrid vector controlled induction motor drive, *IEEE IAS Ann Mtg.*, pp. 137-143, 1995

[5] T. Kanmachi and I. Takahashi, Sensorless speed control of an induction motor with no influence of secondary resistance variation, *IEEE IAS Ann. Mtg.*, pp. 408-413, 1993.

[6] H. Tajima and Y. Hori, Speed sensorless field-orientation control of the induction machine, *IEEE Trans. Indus. Appli.*, vol. 29, No. 1, pp. 175-180, 1993.

[7] S. Tamai, H. Sugimoto and M. Yano, Speed-sensorless vector control of induction motor with model reference adaptive system, *IEEE IAS*. pp. 189-195, 1987.

[8] C. Schauder, Adaptive speed identification for vector control of induction motors without rotational transducers, *IEEE Trans. Indus. Appli.*, vol. 28, No. 5, pp. 1054-1061, 1992.

[9] H. Kubota, K. Matsuse and T. Nakano, DSP-based speed adaptive flux observer of induction motor, *IEEE Trans. Indus. Appli.*, vol. 29, No. 2, pp. 344-348, 1993.

[10] Y. R. Kim, S. K. Sul and M. H. Park, Speed-sensorless vector control of induction motor using extended Kalman filter, *IEEE Trans. Indus. Appli.*, vol 30, No. 5, pp. 1225-1233, 1994.

[11] L. Salvatore, S. Stasi and L. Tarchioni, A new EKF-based algori- thm for flux estimation in induction machines, *IEEE Trans. Indus. Elec.*, vol 40, No. 5, pp. 496-504, 1993.

[12] H. W. Kim and S. K. Sul, A new motor speed estimator using Kalman filter in low-speed range, *IEEE Trans Indus. Electron.*, vol. 43, No. 4, pp. 498-504, 1996.

[13] K. D. Hurst, T. G. Habetler, G. Griva and F. Profumo, Speed sensorless fi-

eld-oriented control of induction machines using current harmonics spectral estimation, *IEEE Ann. Mtg.*, pp. 601-607, 1994.

[14] K. D. Hurst and T. G. Habetler, Sensorless speed measurement using current harmonic spectral estimation in induction machine drives, *IEEE Trans. Power Electron.*, vol. 11, No. 1, pp. 66-73, 1996.

[15] S. I. Yong, J. W. Choi and S. K. Sul, Sensorless vector control of induction machine using high frequency current injection, *IEEE IAS Ann. Mtg.*, pp. 503-508, 1994.

[16] M. G. Simoes and B. K. Bose, Network based estimation of feedback signals for a vector controlled induction motor drive, *IEEE Trans. Indus. Appli .*, vol. 31, No. 3, pp. 620-629, 1993