Auto-Tuning Method for Vector Controlled Induction Motor Drives

Member	Masahiko Akiyama	(Meiji University)
Student Member	Kohji Kobayashi	(Meiji University)
Member	Ichiro Miki	(Meiji University)
Non-member	Mohamed A. El-Sharkawi	(University of Washington)

A new auto-tuning method for vector controlled induction motor drives for the start up is presented in this paper. The tuning algorithm is based on the rotor flux behavior of the induction motor for a stepwise torque current command. The inverse of the rotor time constant, which is important for vector control, is tuned using the transient terminal voltage response caused by undesirable variation of the rotor flux. Furthermore, the electrical motor parameters of the induction motor can also be calculated by this method. The validity of the proposed auto-tuning method is verified by experimental results.

Key words : *Induction motor, Vector control, Auto-tuning, Electrical motor parameters, Rotor flux*

1. Introduction

Vector control system of an induction motor is a high performance drive system to achieve the instantaneous torque control. To obtain the high performance, however, it is necessary to use accurate electrical motor parameters in the vector control system. In particular, the slip gain K_s defined as the inverse of the rotor time constant considerably affects the performance of the motor. These parameters can be computed by conventional motor tests, such as no-load and locked rotor tests. However, these values are different from the actual motor operating parameters. Therefore, it may be difficult to obtain a high performance when these parameters are used to constitute the vector control. Furthermore, the adjustment process of the parameters is time consuming. To overcome this problem, several methods were proposed(1), (2), (3), and (4).

This paper presents a new auto-tuning method for the vector controlled induction motor drive system for the start up. The tuning algorithm is based on the rotor flux behavior of the induction motor for a stepwise torque current command. During the tuning, the induction motor is continuously given a constant exciting current and a stepwise torque current commands. First, the slip gain K_s is tuned using the transient terminal voltage caused by the variation of the rotor flux. At the same time, the stator resistance is calculated. Next, the leakage inductance and the stator inductance are calculated using the linear relationship between the terminal voltage and the motor velocity when the torque current command has the non-zero constant value. Fi-

nally, the torque gain K_t and all the electrical motor parameters are directly computed. By this method, the speed control is not required for the parameter computations.

2. Constraints for Vector Control

The equation of an induction motor in the synchronously rotating d-q reference frame is given as:

$$\begin{bmatrix} V_{ds} \\ V_{qs} \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_s + pL_\sigma & -\omega_o L_\sigma & pM/L_r & -\omega_o M/L_r \\ \omega_o L_\sigma & R_s + pL_\sigma & \omega_o M/L_r & pM/L_r \\ -MR_r/L_r & 0 & R_r/L_r + p & -\omega_s \\ 0 & -MR_r/L_r & \omega_s & R_r/L_r + p \end{bmatrix} \begin{bmatrix} i_{ds} \\ i_{qs} \\ \lambda_{dr} \\ \lambda_{qr} \end{bmatrix} \quad \dots\dots\dots (1)$$

where

p	differential operator
i_{ds}, i_{qs}	d-axis and q-axis stator currents
$\lambda_{dr}, \lambda_{qr}$	d-axis and q-axis rotor interlinked fluxes
V_{ds}, V_{qs}	d-axis and q-axis stator voltages
R_s, R_r	stator and rotor resistances
L_s, L_r	stator and rotor self-inductances
M	mutual inductance
q	number of pole pairs
ω_o	electrical stator angular velocity
ω_r	mechanical rotor angular velocity
ω_s	slip angular velocity
$L_\sigma = L_s - M^2/L_r$	

Since, in this study, the vector control scheme uses a reference frame in which the rotor flux vector always remains in phase with the d-axis, the primary constraint is

$$\lambda_r = \lambda_{dr} = M I_{ds}^* \quad (2)$$

where I_{ds}^* is the exciting current command.

Furthermore, under this constraint, the slip angular velocity ω_s^* and the electromagnetic torque T can be expressed as

$$\omega_s^* = \frac{R_r}{L_r} \frac{i_{qs}^*}{I_{ds}^*} \quad (3)$$

$$T = q \frac{M^2}{L_r} I_{ds}^* i_{qs}^* \quad (4)$$

where i_{qs}^* is the torque current command.

Using the slip gain ($K_s = R_r / L_r$) and the torque gain ($K_t = M^2 / L_r$), Eqs.(3) and (4) can be rewritten as

$$\omega_s^* = K_s \frac{i_{qs}^*}{I_{ds}^*} \quad (5)$$

$$T = q K_t I_{ds}^* i_{qs}^* \quad (6)$$

Thus, Eq.(6) is applicable under the constraint given by Eq.(2). Accurate K_s is necessary to attain the instantaneous torque control expressed by Eq.(6).

3. Principle of Tuning Method

In this study, we assumed that the current control system is adjusted in advance.

Fig.1 shows the tuning signals. I_{ds}^* is found from the no-load test and is kept constant at all times. The torque current command I_{qs}^* in interval B is restricted according to the following equation,

$$I_{qs}^* \leq \sqrt{(\sqrt{3} I_n)^2 - I_{ds}^{*2}} \quad (7)$$

where I_n is the rated current of the induction motor.

The length of interval B shown in Fig.1 is set to meet the condition

$$\omega_r \leq \omega_{r_limit} \quad (8)$$

where ω_{r_limit} is an arbitrary angular velocity of the rotor smaller than the rated value. In interval B, $I_{qs}^* (\neq 0)$ satisfying Eq.(7) is kept constant. The length of interval A is about 5.0 s, which is considerably larger than the rotor time constant. Time

Fig.1. Tuning signals.

t_{2n-1} and t_{2n} shown in Fig.1 are the time to start interval B and A, respectively.

The stator d-axis voltage equation in interval A can be obtained from the first, third, and fourth rows of Eq.(1). The derivation of this equation is described in Appendix in detail.

$$\begin{aligned} V_{ds} &= R_s I_{ds}^* - \frac{M}{L_r} e^{-\frac{R_r}{L_r} t} \left\{ \frac{R_r}{L_r} (\lambda_{dr}(t_{2n}) - M I_{ds}^*) + q \omega_r \lambda_{qr}(t_{2n}) \right\} \\ &= R_s I_{ds}^* + \Delta V_{ds} \quad (9) \end{aligned}$$

where $\lambda_{dr}(t_{2n})$ and $\lambda_{qr}(t_{2n})$ are the initial values of the rotor flux during interval A.

In the above equation, ΔV_{ds} appears when K_s has an inaccurate value. If K_s has the accurate value, ΔV_{ds} becomes zero and V_{ds} always equals $R_s I_{ds}^*$ during interval A because the constraint given by Eq.(2) is satisfied.

Rotor flux and torque deviation caused by the variation of rotor time constant have already been reported in detail⁽¹⁾ and (5). In this paper, we examine the relationship among the rotor flux λ_{dr} , λ_{qr} , the slip gain K_s , and ΔV_{ds} for tuning signal.

From Eq.(9), ΔV_{ds} can be rewritten as

$$\Delta V_{ds} = -\frac{M}{L_r} e^{-\frac{R_r}{L_r} t} \left\{ \frac{R_r}{L_r} \lambda_{dr_err} + q \omega_r \lambda_{qr_err} \right\} \quad (10)$$

where

$$\lambda_{dr_err} = \lambda_{dr}(t_{2n}) - M I_{ds}^* \quad (11)$$

$$\lambda_{qr_err} = \lambda_{qr}(t_{2n}) \quad (12)$$

The sign of ΔV_{ds} depends on the sign of λ_{dr_err} and λ_{qr_err} . Therefore, the relationship between K_s and ΔV_{ds} can be derived by showing the relationship between K_s and λ_{dr_err} or λ_{qr_err} . As is obvious from Eqs.(11) and (12), λ_{dr_err} and λ_{qr_err} are determined by the initial values of the rotor flux during interval A.

By taking the Laplace transform of the third and fourth rows of Eq.(1) during interval B, the following equation can be given

$$\begin{bmatrix} s + \frac{R_r}{L_r} & -\omega_s^* \\ \omega_s^* & s + \frac{R_r}{L_r} \end{bmatrix} \begin{bmatrix} \lambda_{dr}(s) \\ \lambda_{qr}(s) \end{bmatrix} = \begin{bmatrix} \frac{MR_r}{L_r} \frac{I_{ds}^*}{s} + \lambda_{dr}(t_{2n-1}) \\ \frac{MR_r}{L_r} \frac{I_{qs}^*}{s} + \lambda_{qr}(t_{2n-1}) \end{bmatrix} \quad (13)$$

where $\lambda_{dr}(t_{2n-1})$ and $\lambda_{qr}(t_{2n-1})$ are the initial values of the rotor flux during interval B. Because λ_{dr} and λ_{qr} at t_{2n-1} are the initial values, we can replace t_{2n-1} and t_{2n} with zero and t_m , respectively. The initial values of the rotor flux during interval B equal the final values of that during interval A. Therefore, $\lambda_{dr}(t_{2n-1})$ and $\lambda_{qr}(t_{2n-1})$ can be given as follows:

$$\lambda_{dr}(t_{2n-1}) = \lambda_{dr}(0) = MI_{ds}^* \quad (14)$$

$$\lambda_{qr}(t_{2n-1}) = \lambda_{qr}(0) = 0 \quad (15)$$

By substituting Eqs.(14) and (15) into Eq.(13), the inverse Laplace transform of Eq.(13) is as follows:

$$\lambda_{dr}(t) = MI_{ds}^* + \Delta\lambda_{dr} - \Delta\lambda_{dr} e^{-\frac{R_r}{L_r}t} \cos \omega_s^* t - \Delta\lambda_{qr} e^{-\frac{R_r}{L_r}t} \sin \omega_s^* t \quad (16)$$

$$\lambda_{qr}(t) = \Delta\lambda_{qr} - \Delta\lambda_{qr} e^{-\frac{R_r}{L_r}t} \cos \omega_s^* t + \Delta\lambda_{dr} e^{-\frac{R_r}{L_r}t} \sin \omega_s^* t \quad (17)$$

where

$$\Delta\lambda_{dr} = \frac{\omega_s^* L_r^2 M}{R_r^2 + \omega_s^{*2} L_r^2} I_{qs}^* \left(\frac{R_r}{L_r} - K_s \right) \quad (18)$$

$$\Delta\lambda_{qr} = \frac{R_r M L_r}{R_r^2 + \omega_s^{*2} L_r^2} I_{qs}^* \left(\frac{R_r}{L_r} - K_s \right) \quad (19)$$

Accordingly, the trajectory of the rotor flux during interval B can be obtained as shown in Fig.2.

First, the relationship between K_s and λ_{dr_err} is explained. Differentiating Eq.(16), we can obtain

$$p\lambda_{dr}(t) = e^{-\frac{R_r}{L_r}t} \frac{R_r}{L_r} \Delta\lambda_{qr} \left(\frac{\omega_s^{*2} L_r^2}{R_r^2} + 1 \right) \sin \omega_s^* t \quad (20)$$

As is obvious from Eq.(20), $\lambda_{dr}(t)$ is a maximum value for $t = (0/\omega_s^*), (2\pi/\omega_s^*), \dots$ when $K_s > R_r/L_r$. Therefore, from Eqs.(16) and (11), the final value of the d-axis rotor flux during interval B and λ_{dr_err} can be represented as

Fig.2. Trajectory of rotor flux

$$\lambda_{dr}(t_m) \leq MI_{ds}^* + \Delta\lambda_{dr} \left(1 - e^{-\frac{R_r}{L_r}t} \right) \quad (21)$$

$$\lambda_{dr_err} = \lambda_{dr}(t_m) - MI_{ds}^* \leq \Delta\lambda_{dr} \left(1 - e^{-\frac{R_r}{L_r}t} \right) \quad (22)$$

Furthermore, from Eq.(18), $\Delta\lambda_{dr} < 0$ when $K_s > R_r/L_r$. Accordingly, λ_{dr_err} is always a negative value for $K_s > R_r/L_r$.

Conversely, when $K_s < R_r/L_r$, $\lambda_{dr}(t)$ is a minimum value for $t = (0/\omega_s^*), (2\pi/\omega_s^*), \dots$. Similarly, the following equation can be obtained

$$\lambda_{dr_err} \geq \Delta\lambda_{dr} \left(1 - e^{-\frac{R_r}{L_r}t} \right) \quad (23)$$

Therefore, from Eq.(18), $\Delta\lambda_{dr} > 0$ when $K_s < R_r/L_r$, and λ_{dr_err} is always a positive value for $K_s < R_r/L_r$. The relationship between K_s and λ_{dr_err} can be shown in Table 1.

Next, the relationship between K_s and λ_{qr_err} is explained. Differentiating Eq.(17), we can obtain

$$p\lambda_{qr}(t) = e^{-\frac{R_r}{L_r}t} \frac{R_r}{L_r} \Delta\lambda_{qr} \left(\frac{\omega_s^{*2} L_r^2}{R_r^2} + 1 \right) \cos \omega_s^* t \quad (24)$$

From the above equation, $\lambda_{qr}(t)$ is a maximum value for $t = (3\pi/2\omega_s^*)$ when $K_s > R_r/L_r$. Therefore, from Eqs.(17) and (12), the final value of the q-axis rotor flux during interval B and λ_{qr_err} can be represented as

$$\begin{aligned} \lambda_{qr}(t_m) &= \lambda_{qr_err} \leq \Delta\lambda_{qr} - \Delta\lambda_{dr} e^{-\frac{R_r}{L_r}t} \frac{3\pi}{2\omega_s^*} \\ &= \Delta\lambda_{qr} \left(1 - \frac{\Delta\lambda_{dr}}{\Delta\lambda_{qr}} e^{-\frac{R_r}{L_r}t} \frac{3\pi}{2\omega_s^*} \right) \end{aligned}$$

$$= \Delta\lambda_{qr} \left(1 - \frac{L_r}{R_r} \omega_s^* e^{-\frac{R_r}{L_r} \omega_s^* \frac{3}{2}\pi} \right) \dots\dots\dots (25)$$

From Eq.(19), $\Delta\lambda_{qr} < 0$ when $K_s > R_r/L_r$. If the following equation is satisfied, λ_{qr_err} is always a negative value for $K_s > R_r/L_r$.

$$1 - \frac{L_r}{R_r} \omega_s^* e^{-\frac{R_r}{L_r} \omega_s^* \frac{3}{2}\pi} > 0 \dots\dots\dots (26)$$

Conversely, when $K_s < R_r/L_r$, $\lambda_{qr}(t)$ is a minimum value for $t = (3\pi/2\omega_s^*)$. Similarly, the following equation can be obtained

$$\lambda_{qr_err} \geq \Delta\lambda_{qr} \left(1 - \frac{L_r}{R_r} \omega_s^* e^{-\frac{R_r}{L_r} \omega_s^* \frac{3}{2}\pi} \right) \dots\dots\dots (27)$$

From Eq.(19), $\Delta\lambda_{qr} > 0$ when $K_s < R_r/L_r$. If Eq.(26) is satisfied, λ_{qr_err} is always a positive value for $K_s < R_r/L_r$. The relationship between K_s and λ_{qr_err} is also shown in Table 1.

Accordingly, from Eq.(10), the relationship between K_s and ΔV_{ds} can also be obtained as shown in Table 1.

Eq.(26) must be satisfied so that the relation shown in Table 1 can be valid⁽⁶⁾. Therefore, the following equation can be obtained from Eq.(26).

$$\frac{L_r}{R_r} \omega_s^* = \frac{L_r}{R_r} \frac{I_{qs}^*}{I_{ds}^*} K_s < 3.6442\dots = L_{limit} \dots\dots\dots (28)$$

Fig.3 shows the responses of V_{ds} which verifies the relation given by Table 1. In Fig.3 (a), ΔV_{ds} does not become zero in spite of $K_s = R_r/L_r$. This is caused by neglecting the influence of stator core loss. However, we can tune K_s using the relation between K_s and the behavior of ΔV_{ds} because the influence of it is comparatively small.

As mentioned previously, we assumed that the current control system is adjusted in advance. However, the effect of the current control loop on tuning is important in practice, and this problem remains to be solved after this.

Table 1. Relationship between K_s and ΔV_{ds} .

	λ_{dr_err}	λ_{qr_err}	ΔV_{ds}
$K_s < R_r/L_r$	+	+	-
$K_s = R_r/L_r$	0	0	0
$K_s > R_r/L_r$	-	-	+

Fig.3. Responses of V_{ds} .

4. Auto-Tuning Procedure

To implement the auto-tuning, the current commands I_{ds}^* and i_{qs}^* , the motor angular velocity ω_r , and the terminal voltages V_{ds} and V_{qs} are required. The method is described below.

4.1 Calculation of Stator Resistance R_s

The stator resistance R_s is always calculated at the end of interval A because V_{ds} in the steady state equals $R_s I_{ds}^*$ as shown in Fig.3.

4.2 Tuning of Slip Gain K_s

K_s can be tuned by examining the relationship shown in Table 1. Since ΔV_{ds} has ripples as observed in Fig.3, we can tune K_s using the characteristic value F which is the integration of ΔV_{ds}

$$F = \int_{t_{2n}}^{t_{2n}+\tau} \Delta V_{ds} dt \dots\dots\dots (29)$$

where τ is the time of integration.

Next, we define the following equation to tune K_s :

$$K_s(n+1) = \alpha \cdot K_s(n) \quad (30)$$

where the initial slip gain $K_s(0)$ is set to 2.0 because the rotor time constant of induction motor generally ranges from 1.0 ms to 1.0 s. α is the gain factor defined as

$$\alpha = N^{-F} \quad (31)$$

where $N > 1$. From Eq.(31), the relationship between F and α can be obtained as shown in Table 2. Therefore, K_s can be tuned to the accurate value using the gain α defined by Eq.(31), and α is restricted as follows:

$$\alpha_{\text{limit}} \leq \frac{\frac{L_{\text{limit}}}{I_{qs}^*}}{\frac{I_{ds}^*}{I_{ds}^*}} \left(< \frac{\frac{L_{\text{limit}}}{L_r K_s(n) \frac{I_{qs}^*}{I_{ds}^*}}}{R_r \frac{I_{qs}^*}{I_{ds}^*}} \right) \quad (32)$$

This equation can be derived from Eqs.(28), (30), and (31). Finally, the necessary condition to stop tuning K_s is

$$\frac{|K_s(n+1) - K_s(n)|}{K_s(n)} < \varepsilon \quad (33)$$

where ε is the final margin of tuning.

Consequently, K_s can be tuned using Eqs.(29), (30) and (31) under the restrictions described in Eqs.(32) and (33).

Table 2. Relationship between F and α .

	F	α
$K_s < R_r / L_r$	—	$\alpha > 1$
$K_s = R_r / L_r$	0	$\alpha = 1$
$K_s > R_r / L_r$	+	$0 < \alpha < 1$

4.3 Calculation of Torque Gain K_t and Electrical Motor Parameters

As mentioned previously, Eq.(2) is satisfied when accurate value of K_s is used in the vector control system. From the first and second rows of Eq.(1), the voltage equations during interval B can be obtained as follows:

$$\begin{aligned} V_{ds} &= (R_s I_{ds}^* - \omega_s^* L_\sigma I_{qs}^*) - q\omega_r L_\sigma I_{qs}^* \\ &= (R_s I_{ds}^* - \omega_s^* L_\sigma I_{qs}^*) + \omega_r G(V_{ds}) \quad (34) \end{aligned}$$

$$\begin{aligned} V_{qs} &= (R_s I_{qs}^* + \omega_s^* L_s I_{ds}^*) + q\omega_r L_s I_{ds}^* \\ &= (R_s I_{qs}^* + \omega_s^* L_s I_{ds}^*) + \omega_r G(V_{qs}) \quad (35) \end{aligned}$$

The above equations show that the relation between V_{ds} (or V_{qs}) and ω_r is linear. The terms $qL_\sigma I_{qs}^*$ and $qL_s I_{ds}^*$ in Eqs.(34) and (35) can be regarded as the slope G and are calculated by the least squares method. Therefore, L_s and L_σ can be calculated as follows:

$$L_\sigma = -G(V_{ds})/qI_{qs}^* \quad (36)$$

$$L_s = G(V_{qs})/qI_{ds}^* \quad (37)$$

Furthermore, K_t can also be obtained by

$$K_t = L_s - L_\sigma \quad (38)$$

Assuming that L_r equals L_s , M and R_r can be calculated as follows:

$$M = \sqrt{K_t L_r} \quad (39)$$

$$R_r = K_s L_r \quad (40)$$

Thus, two gains K_s and K_t , and the electrical motor parameters can be obtained. The algorithm of this tuning procedure is shown in Fig.4. After the auto-tuning, we can immediately implement the vector control by simply adding the speed control system to the auto-tuning system.

5. Auto-Tuning System

Fig.5 shows the experimental system. A DSP (Digital Signal Processor, NEC : μ PD77240) is used for the auto-tuning and the vector control drive system. One instruction is executed in 90 ns in this processor. Only the auto-tuning system and the voltage sensor are added to the normal vector control system. The current control system is also achieved by DSP and directly sends the ON/OFF signal to the PWM inverter. The sampling time T_s is 256 μ s. The stator current and the stator voltage are detected with HCT and LEM Module, respectively. The cut-off frequency of LPF is 1 kHz. The rotor speed is detected with the pulse encoder (PE), and the feedback signals such as currents and voltages are transformed into the digital signals through a 12 bit A/D converter. As is obvious from Eq.(29), the accuracy of the tuning depends on that of measurement of V_{ds} . We think that the accuracy enough to detect V_{ds} can be obtained because the resolution of A/D converter in this system is 20.5 mV. In practice, Eq.(29) is calculated as follows:

$$F = \sum_{k=1}^k V_{ds} T_s - \tau R_s I_{ds}^* \quad (41)$$

where k is the number of sampling.

Fig.4. Algorithm of auto-tuning.

6. Experimental Results

The rated parameters of the induction motor are shown in Table 3. The variables and coefficients of the test motor are as follows.

$$I_{ds}^* = 9.1 \text{ A}, \quad I_{qs}^* = 9.0 \text{ A}, \quad \omega_{r_limit} = 31.4 \text{ rad/s}, \\ \tau = 2.048 \text{ s}, \quad \alpha_{limit} = 3.0, \quad N = 1.5, \quad \epsilon = 0.03$$

It takes about 2 minutes to complete the auto-tuning. The gains and motor parameters obtained by this auto-tuning are shown in Table 4. For reference, the values obtained by conventional tests such as the no-load and locked rotor tests are also shown in this table. R_r is smaller than that obtained from the conventional tests. In order to verify the accuracy of the tuned K_s , we discuss a linear relationship between input and output torque. Fig.6 shows the input-output torque curves. Load generator shown in Fig.5 is locked to obtain these curves. The output torque is measured on strain gage without ASR by directly varying the torque current command i_{qs}^* . The rated torque of the tested motor is 5.12 Nm. The relation between input and output is linear when the tuned value of K_s ($K_s = 10.68$) is used. The

Fig.5. Experimental system.

Table 3. Induction motor parameters.

0.75[kW]	100[V]	7.6[A]
50[Hz]	1400[rpm]	4 poles

Table 4. Tuning results.

parameters	proposed method	conventional tests
R_s [Ω]	0.8259	0.8401
L_s [H]	0.03630	0.03461
R_r [Ω]	0.3877	0.4039
L_r [H]	0.03630	0.03461
M [H]	0.03198	0.03244
K_s	10.68	11.68
K_t	0.02818	0.03041

Fig.6. Output torque versus torque command.

output torque is non-zero at $i_{qs}^* = 0$. According to reference (3), the non-zero torque at zero command current is caused by the nonlinear behavior of the inverter near zero frequency. The torque curve is approximately straight even when K_s obtained from the conventional test results is used. Therefore, it is found that this value of K_s ($K_s = 11.68$) is also valid in the tested motor. Fig.7 shows instantaneous torque waveforms for $K_s = 11.68$ and $K_s = 10.68$. These torque waveforms are obtained for the induction motor without ASR when the stepwise torque current command shown in Fig.1 is applied. As is obvious from measured instantaneous torque waveforms, the torque waveform is slightly distorted when K_s equals 11.68.

Consequently, the slip gain obtained by the auto-tuning method are accurate, and the effective performance of this method is confirmed.

(a) $K_s = 11.68$ (Conventional tests)

(b) $K_s = 10.68$ (Proposed method)

Fig.7. Instantaneous torque waveform

7. Conclusion

Accurate tuning of the slip gain K_s , the torque gain K_t , and the electrical motor parameters is very important for the vector control. A new auto-tuning method has been developed based on the rotor flux behavior of the induction motor for a stepwise torque current command. The slip gain K_s can be tuned by examining the transient terminal voltage response of the motor.

According to the experimental results, the gains and most of

the electrical motor parameters obtained by the proposed method disagree with those obtained from the results of the conventional motor tests. The linearity between the torque current command and the output torque and the measured instantaneous torque waveform show the accuracy of K_s obtained by auto-tuning method.

Acknowledgment

The authors would like to express their thanks to Dr. Yamada of Meidensha Corporation, Nagoya Factory for his support of devices and discussion.

(Manuscript received August 1, '95

and revised April 16, '96)

References

- (1) R.D.Lorenz, "Tuning of field oriented induction motor controllers for high performance application", IEEE Trans. Ind. Appl., pp.293-297, (Mar./Apr. 1986)
- (2) R.D.Lorenz, D.B.Lawson, "Flux and Torque Decoupling Control for Field-Weakened Operation of Field-Oriented Induction Machines", IEEE Trans. Ind. Appl, pp290-295, (Mar./Apr. 1990)
- (3) Chu Wang, Donald W. Novotny, Thomas A. Lipo, "An automated rotor time constant measurement system for indirect field-oriented drives", IEEE Trans. Ind. Appl, pp151-159, (Jan./Feb. 1988)
- (4) S.K.Sul, "A novel technique of rotor resistance estimation considering variation of mutual inductance", IEEE Trans. Ind. Appl, pp578-587, (Jul./Aug. 1989)
- (5) M. Iwasaki, N. Matsui, "DSP-Based Vector Controlled IM Drive System with Identification of Secondary Time Constant", Trans. IEE Japan, Vol.108-D, No. 9, pp845-852, (1988)
- (6) M.Akiyama, I.Miki, "Auto-tuning method for vector control of induction motor with emphasis on variation of secondary flux", speedam conference. Taormina Italy, pp.209-214, (Jun. 1994)

APPENDIX

Derivation of Eq.(9)

During interval A, $\omega_s = 0$, $pI_{ds}^* = 0$, and $I_{qs}^* = 0$. From the fourth row of Eq.(1),

$$0 = \left(\frac{R_r}{L_r} + p \right) \lambda_{qr}(t) \dots \dots \dots (A1)$$

where p is the differential operator. Taking Laplace transform of above equation and solving for $\lambda_{qr}(s)$, we have

$$\lambda_{qr}(s) = \frac{\lambda_{qr}(t_{2n})}{\frac{R_r}{L_r} + s} \dots \dots \dots (A2)$$

where $\lambda_{qr}(t_{2n})$ is an initial value of the q-axis rotor flux during interval A. The inverse Laplace transform of Eq.(A2) is

$$\lambda_{qr}(t) = \lambda_{qr}(t_{2n})e^{-\frac{R_r}{L_r}t} \dots\dots\dots (A3)$$

Similarly, from the third row of Eq.(1), we can obtain

$$\lambda_{dr}(s) = \frac{MR_r I_{ds}^*}{s\left(\frac{R_r}{L_r} + s\right)} + \frac{\lambda_{dr}(t_{2n})}{\frac{R_r}{L_r} + s} \dots\dots\dots (A4)$$

$$\lambda_{dr}(t) = MI_{ds}^* + (\lambda_{dr}(t_{2n}) - MI_{ds}^*)e^{-\frac{R_r}{L_r}t} \dots\dots\dots (A5)$$

where $\lambda_{dr}(t_{2n})$ is an initial value of the d-axis rotor flux during interval A. Differentiating Eq.(A5) gives

$$p\lambda_{dr}(t) = -\frac{R_r}{L_r}(\lambda_{dr}(t_{2n}) - MI_{ds}^*)e^{-\frac{R_r}{L_r}t} \dots\dots\dots (A6)$$

From the first row of Eq.(1), the stator d-axis voltage equation during interval A is

$$V_{ds} = R_s I_{ds}^* + \frac{M}{L_r} \{p\lambda_{dr}(t) - q\omega_r \lambda_{qr}(t)\} \dots\dots\dots (A7)$$

By substituting Eqs.(A3) and (A6) into (A7), V_{ds} can be obtained.

$$\begin{aligned} V_{ds} &= R_s I_{ds}^* - \frac{M}{L_r} e^{-\frac{R_r}{L_r}t} \left\{ \frac{R_r}{L_r} (\lambda_{dr}(t_{2n}) - MI_{ds}^*) + q\omega_r \lambda_{qr}(t_{2n}) \right\} \\ &= R_s I_{ds}^* + \Delta V_{ds} \dots\dots\dots (9) \end{aligned}$$

Masahiko Akiyama (Member)

He was born in Chiba Prefecture, Japan 1970. He received the B.S. and M.S. degrees in Electrical Engineering from Meiji University, Japan, in 1993 and 1995 respectively. He joined Oriental Motor Co. Ltd as an engineer in 1995. He was assigned to the Kashiwa Plant and has been engaged in developing control of the servo motor.

Kohji Kobayashi (Student Member)

He was born in Nagano Prefecture, Japan 1970. He received the B.S. degree in Electrical Engineering from Meiji University, Japan, in 1994. He is at present a graduate student at Meiji University and has been engaged in control of motor.

Ichiro Miki (Member)

He received the B.S., M.S., and Ph.D. degrees in Electrical Engineering from Meiji University, Tokyo, Japan in 1973, 1975, and 1981, respectively.

Since 1978, he has been with the Department of Electrical Engineering at Meiji University, Kawasaki, Japan, and he is at present a Professor. His primary research interests are motor control, fuzzy control, and neural network applications to robot system.

Dr. Miki is a member of IEEE.

Mohamed A. El-Sharkawi (Non-member)

He received the B.Sc. in Electrical Engineering in 1971 from Cairo High Institute of Technology, Egypt. His M.A.Sc and Ph.D. in Electrical Engineering were received from the University of British Columbia in 1977 and 1980 respectively. In 1980, he joined the University of Washington as a faculty member where he is presently a professor.

He is the founder of the international conference on the Application of Neural Networks to Power System (ANNPS) which was later merged with the Expert Systems conference and renamed Intelligent Systems Applications to Power (ISAP). He is a member of the administrative committee of the IEEE Neural Networks Council representing the Power Engineering Society. He is a member of the editorial board and/or associate editor of several journals including the IEEE Transactions on Neural Networks. He is the video tutorial committee chair of the neural network council and the IEEE Continuing Education Committee.

His major areas of funded research include power electronics applications to power systems, neural network applications to power systems, high performance drives, and large scale power system dynamics and control.

Dr. El-Sharkawi is a Fellow of IEEE.