A Discrete Adaptive Field-Oriented Induction Motor Drive

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***Abstract*—In this paper, a discrete model reference adaptive controller is designed and implemented to let the performance of the field-oriented induction motor drives be insensitive to the parameter changes. Only the information of the reference model and the plant output are required for the control. No on-line explicit parameter identification is required. Hence, the proposed controller is easy to implement practically. For designing the proposed adaptive controller, the dynamic model of the drive system is estimated from the sampled input/output data using the stochastic modeling technique. Theoretical basis of the adaptive control is derived and simulation is made. Then the hardware of the drive system and the microprocessor-based adaptive controller are implemented. Some experimental results are given to demonstrate the effectiveness of the proposed controller.**

I. Introduction

LATELY, the field-oriented induction motor drive [1]–[3] can be applied for high-performance industrial applications where, traditionally, only the dc motors were used. However, its performance depends heavily on the motor parameters. The dominant parameter to be considered is the rotor time constant [2]–[8]. Thus, it is desirable to have a robust controller for the drive system to reduce parameter sensitivity. Adaptive control is an efficient technique for dealing with large parameter variations. Basically, adaptive control systems can be classified into two categories [9], [10]—the self-tuning regulators and the model reference adaptive control (MRAC) systems. The former is based on the explicit identification of the system transfer function parameters. It is not convenient for it to be applied to the induction motor drives using microprocessors. As to the MRAC systems, a control input is designed to drive the controlled plant to track the response produced by the reference model. Various control algorithms developed [9], [10] require the system states, thus they are not easy to implement.

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To overcome this problem, and to enhance the flexibility of changing control algorithms, a digital output following MRAC system for field-oriented induction motor drives is introduced and successfully implemented in this paper. Only the available information of the reference model and the plant output are required for the control. No on-line explicit parameter identification is required. Hence, the proposed controller is easy to implement practically. For designing the proposed controller, the dynamic model of the drive system is found based on the stochastic modeling technique. Theoretic derivation and implementation of the proposed controller are presented. Some experimental results are provided to demonstrate the effectiveness of the proposed controller.

II. Dynamic Modeling of the Field-Oriented Induction Motor Drive

A. Physical Modeling

The voltage equations of an induction motor in the synchronously rotating frame can be expressed as follows [2]:

$$\begin{pmatrix} V_{qs} \\ V_{ds} \\ 0 \\ 0 \end{pmatrix} = \begin{pmatrix} R_s + sL_s & \omega_e L_s & sL_m & \omega_e L_m \\ -\omega_e L_s & R_s + sL_s & -\omega_e L_m & sL_m \\ sL_m & \omega_{sl} L_m & R_r + sL_r & \omega_{sl} L_r \\ -\omega_{sl} L_m & sL_m & -\omega_{sl} L_r & R_r + sL_r \end{pmatrix} \begin{pmatrix} i_{qs} \\ i_{ds} \\ i_{qr} \\ i_{dr} \end{pmatrix} \tag{1}$$

$$T_e = (3P/4) L_m (i_{qs} i_{dr} - i_{ds} i_{qr}) \tag{2}$$

where

R_s = stator resistance per phase
L_m = magnetizing inductance per phase
R_r = rotor resistance per phase referred to stator
L_r = rotor inductance per phase referred to stator
L_s = stator inductance per phase
P = number of poles
ω_e = electrical angular speed
ω_{sl} = slip angular speed
$V_{qs}(V_{ds})$ = q-axis (d-axis) stator voltage
$i_{qs}(i_{ds})$ = q-axis (d-axis) stator current
$i_{qr}(i_{dr})$ = referred q-axis (d-axis) rotor current.

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Equations (1) and (2) can be rearranged to yield the following state equations:

$$\frac{d}{dt}\begin{bmatrix} i_{ds} \\ i_{qs} \\ \lambda_{dr} \\ \lambda_{qr} \end{bmatrix} = \begin{bmatrix} \frac{-R_s}{\sigma L_s} - \frac{R_r(1-\sigma)}{\sigma L_r} & \omega_e & \frac{L_m R_r}{\sigma L_s L_r^2} & \frac{P\omega_r L_m}{2\sigma L_s L_r} \\ -\omega_e & \frac{-R_s}{\sigma L_s} - \frac{R_r(1-\sigma)}{\sigma L_r} & \frac{-P\omega_r L_m}{2\sigma L_s L_r} & \frac{L_m R_r}{\sigma L_s L_r^2} \\ \frac{L_m R_r}{L_r} & 0 & -\frac{R_r}{L_r} & \omega_e - \frac{P}{2}\omega_r \\ 0 & \frac{L_m R_r}{L_r} & -\left(\omega_e - \frac{P}{2}\omega_r\right) & -\frac{R_r}{L_r} \end{bmatrix} \begin{bmatrix} i_{ds} \\ i_{qs} \\ \lambda_{dr} \\ \lambda_{qr} \end{bmatrix} + \frac{1}{\sigma L_s}\begin{bmatrix} v_{ds} \\ v_{qs} \\ 0 \\ 0 \end{bmatrix} \quad (3)$$

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

where

$$\sigma = 1 - L_m^2/(L_s L_r)$$

$$\lambda_{qr} = L_m i_{qs} + L_r i_{qr}$$

$$\lambda_{dr} = L_m i_{ds} + L_r i_{dr}.$$

The ideal decoupling between the d and q axes can be achieved by letting the rotor flux linkage be aligned in the d axis, i.e.,

$$\lambda_{qr} = 0 \quad \text{and} \quad \frac{d}{dt}\lambda_{qr} = 0. \quad (5)$$

Using (5), the desired rotor flux linkage $\tilde{\lambda}_r = \lambda_{dr}$ in terms of i_{ds} can be found from the third row of (3) as

$$\lambda_{dr} = (i_{ds}L_m)\Big/\left(1 + \frac{L_r}{R_r}s\right). \quad (6)$$

Compared with the dynamic characteristics of the mechanical system, the dynamic characteristic of (6) is neglected and i_{ds} is set constant for the desired constant rated rotor flux and the torque equation (4) becomes

$$T_e = K_t i_{qs} \quad (7)$$

where

$$K_t = (3P/4)(L_m^2/L_r)i_{ds}. \quad (8)$$

The generated torque and rotor angular speed are related by

$$\omega_r(s) = \left(\frac{1/J}{s + B/J}\right)(T_e(s) - T_L(s)) \quad (9)$$

where B and J denote the system damping ratio and inertia constant, respectively.

In a commonly used hysteresis current-controlled PWM inverter drive, the current commands denoted by i_{ds}^* and i_{qs}^* must be transformed into the stationary reference frame to yield the reference phase currents for the inverter. For the indirect field orientation, the unit vector in the transformation matrix is generated by using the rotor position and an estimated feedforward slip speed signal, which can be derived from the fourth row of (3) as

$$\omega_{sl} = L_m R_r/(L_r|\tilde{\lambda}_r|)i_{qs}^*. \quad (10)$$

An indirect field-oriented induction motor speed drive described above is shown in Fig. 1 where the torque current command i_{qs}^* is generated from the speed error through a PI controller, which has the following structure:

$$G_c(s) = K_p + K_I/s. \quad (11)$$

The control system block diagram corresponding to Fig. 1 is shown in Fig. 2. The response of $\omega_r(s)$ can be found from Fig. 2 as

$$\omega_r(s) = H_{\omega c}(s)\omega_r^*(s) + H_{\omega d}(s)T_L(s) \quad (12)$$

where

$$H_{\omega c}(s) = \frac{s(K_p K_t i_{ds}) + (K_I K_t i_{ds})}{s^2 J + (B + K_p K_t i_{ds})s + K_I K_t i_{ds}} \quad (13)$$

$$H_{\omega d}(s) = \frac{-s}{s^2 J + (B + K_p K_t i_{ds})s + K_I K_t i_{ds}}. \quad (14)$$

The induction motor used in this paper is a three-phase Δ-connected four-pole 1-Hp 60-Hz 220-V induction motor having the following parameters:

$R_s = 3.20\ \Omega$ $\quad R_r = 2.349\ \Omega$ $\quad L_s = 129.4$ mH

$L_r = 132.9$ mH

$L_m = 126.7$ mH $\quad J = 0.009$ kg-m^2.

The transfer function model $H_{\omega c}(s)$ in (13) will be used as the plant model for the adaptive control system design. However, the system damping ratio and inertia constant of the system are rather difficult to be measured with reasonable accuracy. Hence, the plant model of $H_{\omega c}(s)$ is not easy to find analytically. Fortunately, the stochastic modeling technique described as follows can be applied as an alternative.

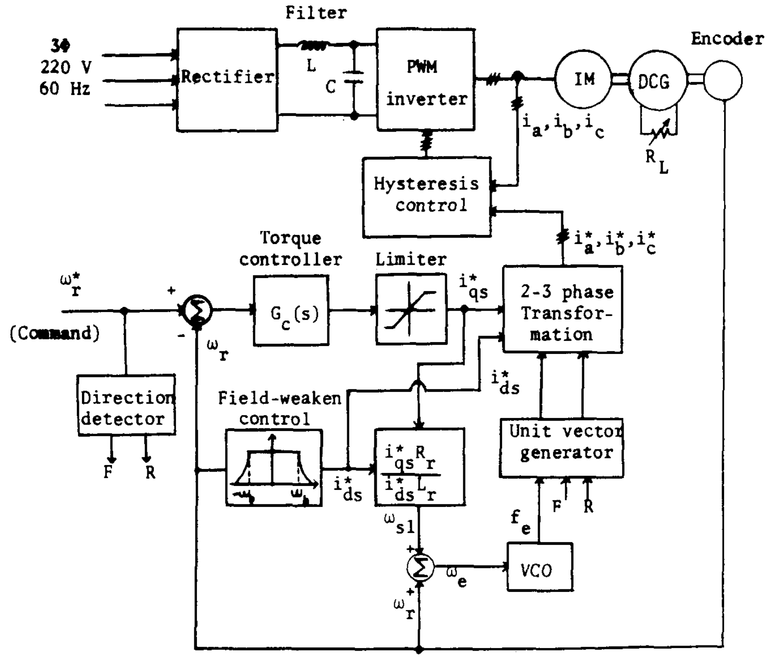

Fig. 1. The configuration of an indirect field-oriented induction motor drive.

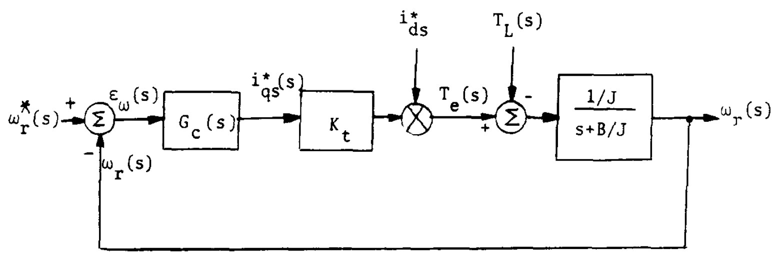

Fig. 2. The control system block diagram of field-oriented induction motor drive.

B. Stochastic Modeling

A linear stable discrete-time system can be expressed by an ARMA model $G(B)$ as

$$G(B) \triangleq \frac{b_0 + b_1 B + \cdots + b_{n-1} B^{n-1}}{1 + a_1 B + a_2 B^2 + \cdots + a_n B^n}. \quad (15)$$

The relationship between the output y_k and the input u_k can be expressed by

$$(1 + a_1 B + a_2 B^2 + \cdots + a_n B^n)(y_k - v_k) = (b_0 + b_1 B + \cdots + b_{n-1} B^{n-1}) u_k \quad (16)$$

where v_k denotes the output noise or disturbance. The maximum likelihood method [11], [12] is used to find parameters of a_i and b_i from the sampled input/output data $\{u_k\}$ and $\{y_k\}$.

Model reduction is necessary for simplifying the controller design. Before the model reduction is performed the most important and difficult issue is the selection of proper reduced order. A practical method of reduced-order determination is introduced as follows. The estimated transfer function $G(B)$ of (15) can be expressed as

$$G(B) = \sum_{i=1}^{n} \frac{g_i}{1 - \lambda_i B} = \sum_{j=0}^{\infty} \left(\sum_{i=1}^{n} g_i \lambda_i^j \right) B^j = \sum_{j=0}^{\infty} G_j B^j \qquad |\lambda_i| < 1. \quad (17)$$

G_j is the sequence of the impulse responses. Using G_j, the variance function of $x_k (x_k = y_k - v_k)$ can be found as [11]

$$\gamma(0) = \sigma_z^2 \sum_{j=0}^{\infty} G_j^2 = \sum_{j=1}^{n} d_j \quad (18)$$

where

$$d_j = \sigma_z^2 \sum_{i=1}^{n} \frac{g_i g_j}{1 - \lambda_i \lambda_j}. \quad (19)$$

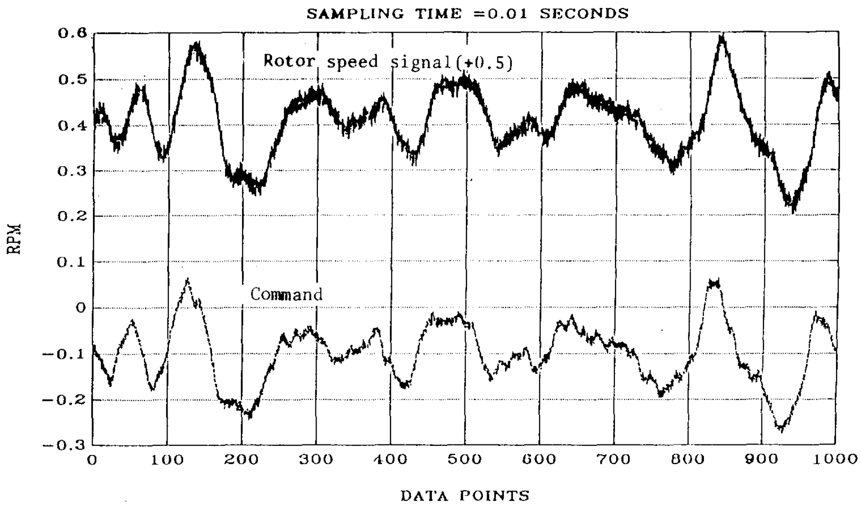

Fig. 3. The waveforms of the recorded command signal and rotor speed signal.

The energy dispersion of dynamic mode λ_j, defined as

$$D_j = \frac{d_j}{\gamma(0)} = d_j \bigg/ \left(\sum_{i=1}^{n} d_i \right) \tag{20}$$

can be used as a measurement of the importance of each dynamic mode. Accordingly, if the dynamic modes have large dispersions at $\lambda_1, \lambda_2, \cdots, \lambda_m$, then the order of the reduced model can be considered to be m.

After the reduced-model order is determined to be m, the reduced model can be expressed as

$$G^*(B) = \frac{b_0^* + b_1^* B + \cdots + b_{m-1}^* B^{m-1}}{1 + a_1^* B + a_2^* B^2 + \cdots + a_m^* B^m}. \tag{21}$$

If the numerical difference between $G(e^{-j\omega T})$ and $G^*(e^{-j\omega T})$ is $\epsilon(j\omega)$, then we have

$$G(e^{-j\omega T}) = G^*(e^{-j\omega T}) + \epsilon(j\omega). \tag{22}$$

To avoid the steady-state error, the steady-state value of the reduced model is first set equal to that of the original model:

$$G^*(B = 1) = \frac{b_0^* + b_1^* + \cdots + b_{m-1}^*}{1 + a_1^* + a_2^* + \cdots + a_m^*} = G(B = 1) = \frac{b_0 + b_1 + \cdots + b_{n-1}}{1 + a_1 + a_2 + \cdots + a_n}. \tag{23}$$

Since the parameters of the denominator of the reduced model are already known from the dispersion analysis, only the parameters of b_i^*, $i = 0, 1, \cdots, m - 1$ are needed to be found. These parameters can be found by using the least-squares fitting technique [11].

The proportional gain and integral gain of the controller listed in (11) are set to 8.0 and 4.5, respectively. A low-frequency white noise is superimposed to the speed command when the drive was operating at $\omega_{r0} = 1000$ r/min and $T_{LO} = 0.55$ kg-m. The waveforms of the recorded command signal and rotor speed signal are shown in Fig. 3. Following the procedure described above, the drive model estimated from the sampled command and rotor speed signal is

$$G^*(B) = \frac{0.2408B}{1 - 0.759B}. \tag{24}$$

This estimated discrete plant model will be used for the adaptive control system design.

III. A Discrete Output Adaptive Model-Following Controller

A discrete output adaptive model-following controller suitable for motor drive is presented in this section. Suppose that the plant to the controlled and the chosen reference model are expressed as

$$x_p(k+1) = A_p x_p(k) + B_p u_p(k) \tag{25}$$

$$y_p(k) = C_p x_p(k) \tag{26}$$

$$x_m(k+1) = A_m x_m(k) + B_m u_m(k) \tag{27}$$

$$y_m(k) = C_p x_m(k). \tag{28}$$

where $x_p \in R^n$, $x_m \in R^n$, $u_p \in R^p$, $u_m \in R^p$, $y_p \in R^q$, $y_m \in R^q$, and A_p, B_p, C_p, A_m, B_m are constant matrices of appropriate dimensions. The pairs (A_p, B_p) and (A_m, B_m) are stabilizable and A_m is a stable matrix. The objective is to find the control input $u_p(k)$ such that the plant states can track those of the reference model. Then the resulting y_p will follow y_m automatically. For easy implementation, the control input u_p is chosen to be

$$u_p(k) = u_{pl}(k) = K_x x_m(k) + K_e e_0(k) + K_u u_m(k) \tag{29}$$

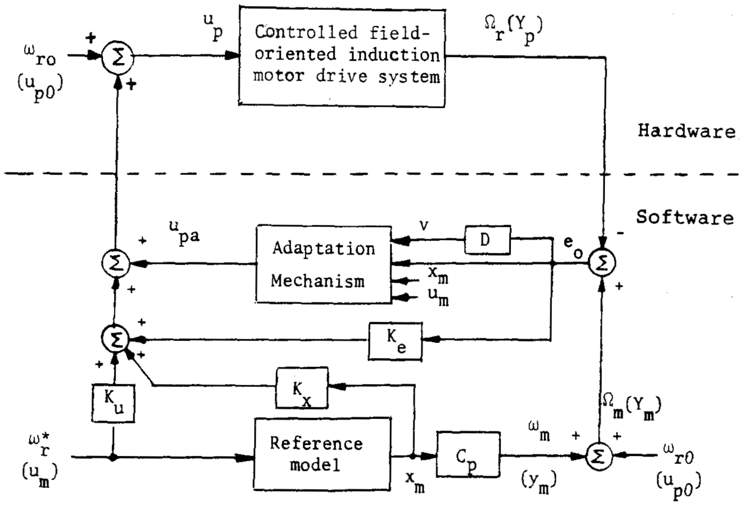

Fig. 4. The configuration of the proposed discrete model reference adaptive control system.

where $e_0(k) \triangleq y_m(k) - y_p(k)$ is the error between the system output and the model output. Define the error vector

$$e(k) \triangleq x_m(k) - x_p(k). \tag{30}$$

Then from (25)–(29), one can obtain the following equation:

$$\begin{aligned} e(k+1) &= (A_p - B_p K_e C_p) e(k) \\ &\quad + (A_m - A_p - B_p K_x) x_m(k) \\ &\quad + (B_m - B_p K_u) u_m(k). \end{aligned} \tag{31}$$

Equation (31) shows that if A_m, B_m, K_x, K_e, and K_u are chosen to let $(A_p - B_p K_e C_p)$ be a Hurwitz matrix, and

$$K_x = B_p^+ (A_m - A_p) \tag{32}$$

$$K_u = B_p^+ B_m \tag{33}$$

where B_p^+ is the left Penrose pseudoinverse of B_p, $B_p^+ = (B_p^T B_p)^{-1} B_p^T$ [9], then the error system of (31) will be asymptotically stable and the output of the controlled plant will follow that of the reference model.

The linear model-following control system (LMFC) proposed above can lead to perfect model-following characteristics only when the plant is invariant. Thus, in addition to the regular input u_{pl} of (29), an adaptation signal u_{pa} is added to reduce the effect of motor parameter changes,

$$\begin{aligned} u_{pa}(k) &= \Delta K_x(k, v(k)) x_m(k) + \Delta K_e(k, v(k)) e_0(k) \\ &\quad + \Delta K_u(k, v(k)) u_m(k) \end{aligned} \tag{34}$$

$$v(k) \triangleq D e_0(k) = D C_p e(k) \tag{35}$$

where $\Delta K_x(k)$, $\Delta K_e(k)$, and $\Delta K_u(k)$ are adaptive gain matrices, and D is a $(p \times q)$ gain matrix for e_0. Fig. 4 shows the block diagram of the proposed adaptive controller. By adding u_{pa} to (29), one can obtain an equivalent feedback system described by the following equations:

$$e(k+1) = (A_p - B_p K_e C_p) e(k) + B_p w_1(k) \tag{36}$$

$$v(k) = D e_0(k) = D C_p e(k) \triangleq D_c e(k) \tag{37}$$

$$\begin{aligned} w_1(k) &= [B_p^+ (A_m - A_p) - K_x - \Delta K_x(k)] x_m(k) \\ &\quad - \Delta K_e(k) e_0(k) \\ &\quad + [B_p^+ B_m - K_u - \Delta K_u(k)] u_m(k). \end{aligned} \tag{38}$$

According to Popov's theorem [9], for the above system to be asymptotically hyperstable, it is necessary and sufficient that

1) The nonlinear time-varying part must satisfy the following inequality:

$$\eta(k_0, k_1) = \sum_{k=k_0}^{k_1} v^T(k) w_1(k) \geq -\gamma_0^2 \tag{39}$$

for all $k_1 \geq k_0$ where γ_0^2 is a real finite positive constant.

2) The transfer matrix of the equivalent linear part

$$G(z) = D_c (zI - A_p + B_p K_e C_p)^{-1} B_p \tag{40}$$

must be strictly positive real.

In order to let the inequality (39) be satisfied, the adaptive gain matrices $\Delta K_x(k, v(k))$, $\Delta K_e(k, v(k))$, and $\Delta K_u(k, v(k))$ in (34) with PI adaptation are set as

$$\Delta K_x(k) = \Delta K_x^I(k) + \Delta K_x^P(k)$$

$$\Delta K_x^I(k) = \Delta K_x^I(k-1) + L_1 v(k) [Q_1 x_m(k-1)]^T$$

$$\Delta K_x^P(k) = L_2 v(k) [Q_2 x_m(k-1)]^T$$

$$\Delta K_e(k) = \Delta K_e^I(k) + \Delta K_e^P(k)$$

$$\Delta K_e^I(k) = \Delta K_e^I(k-1) + M_1 v(k) [R_1 e_0(k-1)]^T$$

$$\Delta K_e^p(k) = M_2 v(k)[R_2 e_0(k-1)]^T$$
$$\Delta K_u(k) = \Delta K_u^I(k) + \Delta K_u^p(k)$$
$$\Delta K_u^I(k) = \Delta K_u^I(k-1) + N_1 v(k)[S_1 u_m(k-1)]^T$$
$$\Delta K_u^p(k) = N_2 v(k)[S_2 u_m(k-1)]^T \quad (41)$$

where $L_i, M_i, N_i \in R^{p\times p}$, $Q_i \in R^{n\times n}$, $R_i \in R^{q\times q}$, $S_i \in R^{p\times p}$, $i = 1, 2$; $Q_i, R_i, S_i, i = 1, 2$, and L_1, M_1, N_1 are positive definite matrices, and L_2, M_2, N_2 are positive or nonnegative definite matrices. The proof of the satisfaction of inequality (39) is straightforward and is neglected here.

The derivation of the control algorithm just listed does not consider the effect of time delay in the adaptation mechanism. However, the presence of this inherent time delay may introduce a supplementary condition that must be considered in the adaptation algorithm [9]. The controlled plant at time instant k with control input using the adaptation signal at instant $k - 1$ can be described by

$$x_p^0(k+1) = A_p x_p(k) + B_p u_p^0(k) \quad (42)$$

where

- $u_p^0(k)$ = the control input with adaptation signal synthesized at time instant $k-1$.
- $u_p(k)$ = the control input with adaptation signal synthesized at time instant k.
- $x_p^0(k)$ = priori state variables of the plant using $u_p^0(k)$
- $x_p(k)$ = posteriori state variables of the plant using $u_p(k)$.

Accordingly, the control input vector, the state error vector, and the compensated output error signal vector before adaptation can be written as

$$u_p^0(k) = u_{pl}(k) + u_{pa}^0(k) = u_{pl}(k) + \Delta K_x^I(k-1)x_m(k) + \Delta K_e^I(k-1)e_0(k) + \Delta K_u^I(k-1)u_m(k) \quad (43)$$

$$e^0(k) = x_m(k) - x_p^0(k) \quad (44)$$

$$v^0(k) = D_c e^0(k). \quad (45)$$

In practical implementation, the adaptation algorithms of (34)–(41) must be expressed in terms of $v^0(k)$ instead of $v(k)$. The relation between $v^0(k)$ and $v(k)$ can be found as follows. Using (44), one can find the explicit expression of $e^0(k+1)$ as

$$\begin{aligned} e^0(k+1) &= x_m(k+1) - x_p^0(k+1) \\ &= A_m x_m(k) + B_m u_m(k) - A_p x_p(k) \\ &\quad - B_p(K_x x_m(k) + K_e e_0(k) + K_u u_m(k)) \\ &\quad - B_p[\Delta K_x^I(k-1)x_m(k) \\ &\quad + \Delta K_e^I(k-1)e_0(k) \\ &\quad + \Delta K_u^I(k-1)u_m(k)]. \end{aligned} \quad (46)$$

Then, using (45), and the fact that the incremental changes of $v(k) - v^0(k)$ between adjacent sample instants are almost identical, one can obtain

$$\begin{aligned} v(k) = \{I &+ D_c B_p [L_1 x_m(k-1)^T Q_1^T x_m(k-1) \\ &+ L_2 x_m(k-1)^T Q_2^T x_m(k-1) \\ &+ M_1 e_0(k-1)^T R_1^T e_0(k-1) \\ &+ M_2 e_0(k-1)^T R_2^T e_0(k-1) \\ &+ N_1 u_m(k-1)^T S_1^T u_m(k-1) \\ &+ N_2 u_m(k-1)^T S_2^T u_m(k-1)]\}^{-1} v^0(k). \end{aligned} \quad (47)$$

Physically, $v^0(k)$ is the error between reference model output and plant output, which is measured at instant k when the LMFC is actuated but the adaptation signal is not applied yet. However, the data sampling process must be carried twice in each time period, which is quite unreasonable as far as minimized computation time is concerned. Thus, in the implementation, $v^0(k)$ is approximated by

$$v^0(k) = De_0^0(k) \cong De_0(k-1). \quad (48)$$

The design procedure of the proposed adaptive controller can be summarized as follows:

1) Select a proper reference model such that the desired performance can be achieved.
2) Choose K_e vector such that $(A_p - B_p K_e C_p)$ is a Hurwitz matrix.
3) Calculate K_x and K_u vectors using (32) and (33).
4) Parameters $D, L_i, M_i, N_i, Q_i, R_i, S_i, i = 1, 2$ of the adaptation mechanism are chosen to satisfy the conditions of (39) and (40).

IV. Design of the Proposed Adaptive Field-Oriented Induction Motor Drive

The dynamic model of the field-oriented induction motor drive obtained by the stochastic dynamic modeling technique described in the previous section is repeated as follows:

$$G_p(B) = \frac{0.2408B}{1 - 0.759B}. \quad (49)$$

If the reference model is chosen as

$$G_m(B) = \frac{0.4B}{1 - 0.6B} \quad (50)$$

then the parameters of the proposed adaptive controller are found following the procedure described in the previous sections as

$$\begin{aligned} &K_e = 1.0 \quad K_x = -0.66 \quad K_u = 1.66 \\ &D = 2.0 \quad L_i = M_i = N_i = Q_i = R_i = S_i = 1.0 \\ &\qquad i = 1, 2. \end{aligned} \quad (51)$$

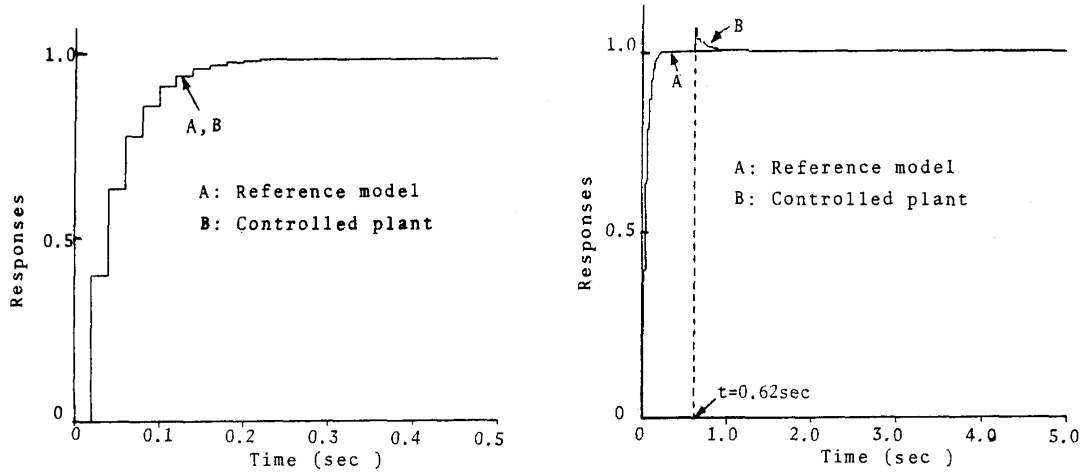

Fig. 5. The simulated output responses of the reference model and the controlled drive system.

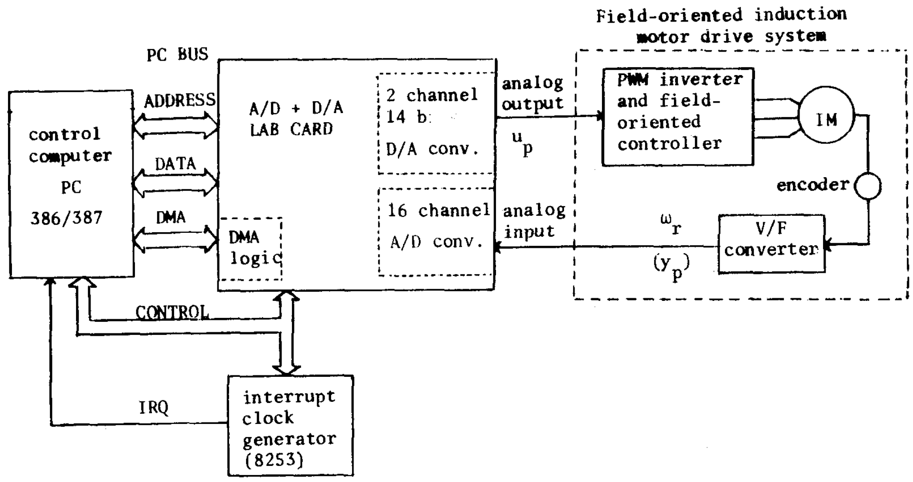

Fig. 6. The schematic diagram of the computer closed-loop control system for implementing the proposed adaptive controller.

Before implementing the proposed adaptive controller, the computer simulation is made to test the effectiveness of the controller. The simulation results shown in Fig. 5(a) indicate that the perfect model following control can be achieved. Now suppose that the plant is changed to

$$G_{p1}(B) = \frac{0.3B}{1 - 0.759B} \tag{52}$$

at $t = 0.62$ s. The simulation results shown in Fig. 5(b) indicate that rather good performance under parameter changes can still be obtained by the proposed adaptive controller.

V. Implementation of the Proposed Controller and Experimented Results

Having tested the performance of the proposed controller by simulation, the hardware implementation of the drive system and the software realization of the proposed adaptive controller are carried out. Fig. 6 shows the schematic diagram of the computer control system for implementing the proposed adaptive controller. The sampling interval is controlled by an interrupt facility using the 8253 timer circuit. A laboratory card, which has 2-channel D/A converters and 16-channel A/D converters, is used as the interfaces between the controlled motor drive and the computer.

Based on the parameters of the proposed adaptive controller found in the previous section, Fig. 7 gives the dynamic rotor speed response to a step command applied when the motor was operated at (ω_{r0} = 1000 r/min, T_{LO} = 0.55 kg-m) and (ω_{r0} = 500 r/min, T_{LO} = 0.25 kg-m), respectively. The rotor speed responses of the proposed drive system without adaptive control are also shown in Fig. 7 for comparison. The results show that better following characteristics using the proposed adaptive control

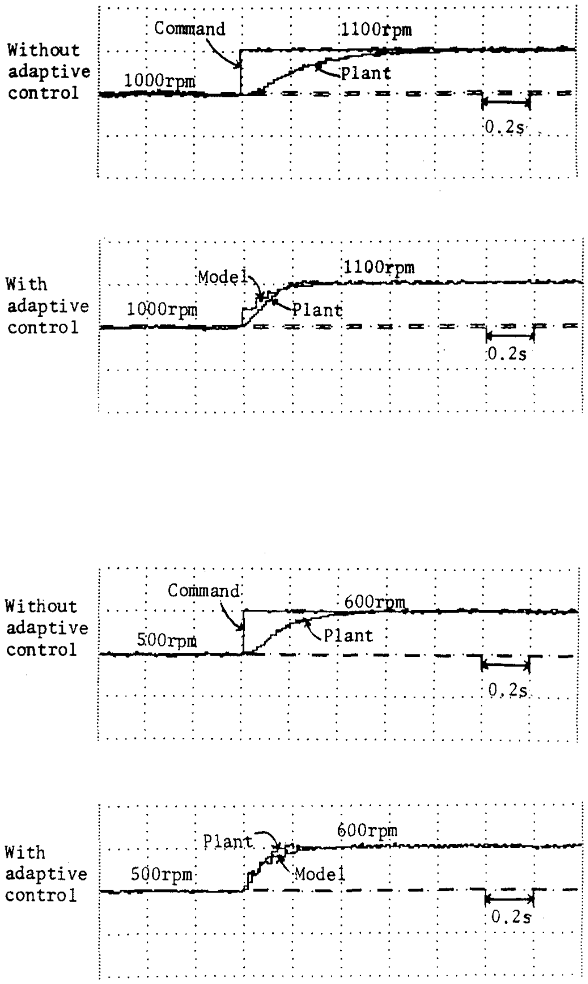

Fig. 7. The measured step rotor speed response of the reference model and the controlled drive system at different operating conditions.

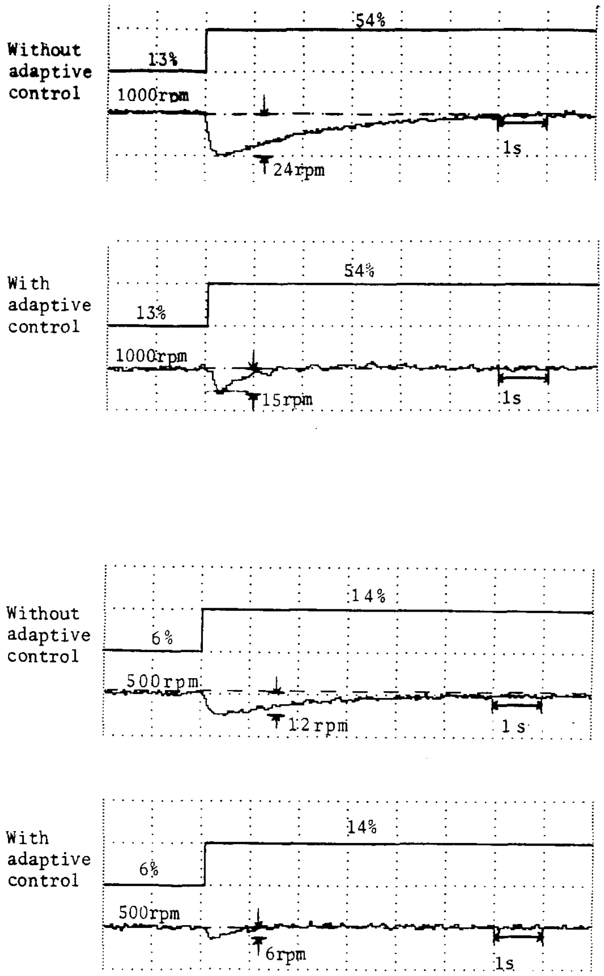

Fig. 8. The measured rotor speed responses of the controlled drive system due to a step-load resistance change of $R_L = 160\text{–}45\ \Omega$ occurring at different operating conditions.

are achieved. To test the speed regulation characteristics, a permanent dc generator with switched resistors is used as the load of the drive system. Fig. 8 shows the dynamic rotor speed responses without and with adaptive control due to a step-load resistance change from 160 to 45Ω, the corresponding load torque changes are estimated to be from 13 to 54% rated load torque (4.2 kg-m) and from 6 to 14% rated load torque for 1000 and 500 r/min operating speeds, respectively. One can observe from Fig. 8 that the smaller speed dip and faster restoration are also obtained using this proposed adaptive controller. Furthermore, the performances of the the proposed controller are rather insensitive to the operating condition changes.

VI. Conclusions

The design and implementation of a discrete adaptive speed controller for field-oriented induction motor drives have been presented. The dynamic model of the field-oriented induction motor drive system is found using a stochastic dynamic modeling technique. The main features of the proposed controller are that not all system states are required and an explicit parameter identification is not needed. After the effectiveness of the proposed controller by computer simulations is tested, the adaptive controller is successfully implemented. The experimental results show that good performances in both the rotor speed following and regulation characteristics are obtained, and the performances are rather insensitive to operating condition changes.

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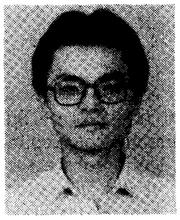

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