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Minimal Order Bilinear Observer for High Performance Control of Induction Motor Taking Core Loss into Account†

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This paper presents a minimal order rotor current and rotor flux observer for an indirect field oriented induction motor drive with consideration of core loss due to eddy current from the viewpoint of nonlinear observer using bilinear model. The state equations for an induction motor were derived which behaved as a bilinear system in terms of the product of input and state variables. The design of the proposed observer is based on Lyapunov's stability method whose estimation error converges to zero exponentially irrespective of the inputs. We have also proposed a control system by using multi-input and multi-output optimal regulator theory. The estimated rotor current and rotor flux are fed back to the multi-input and multi-output optimal regulator. Simulation results are presented to show the validity of the proposed controller as well as rotor current and rotor flux estimation of induction motor drive.

Key Words: Minimal order bilinear observer, optimal regulator control, induction motor, core loss

1. Introduction

Due to the inherent limitations, DC motors have been replaced in the industrial application by the induction motor (IM) drive in terms of the vector control technique, also known as field oriented control (FOC) for high performance. However, the general FOC theory was derived under the assumption that the core loss may be neglected. This idealization is not met in practice where core loss exists.

The effects of core loss in the orthogonal-axis model of electrical machines have been investigated[1]. Also, a model with a steady state equivalent resistance for representing the core loss has been suggested in [1]. The proposed equivalent resistance model in [1] has been applied in [2, 5] for the detail analysis of different types of FOC of IM under the steady state condition.

The authors of [2, 3] have investigated the core loss effects in the FOC of IM and showed that the core loss deteriorates the performance of FOC because the output torque and flux differ from the reference torque and flux. The proper decoupling control of rotor flux and torque by setting the stator d-q axis current as a reference for the indirect FOC (IFOC) neglecting core loss has been developed in many research works such as [4]. The decoupling control with consideration of core loss for IFOC[3] in steady state condition and direct FOC[3] have been developed for precise operation in the industrial applications which are very complex in terms of the mathematical model and controller design. The IFOC in [2] has been accomplished using conventional PI controller. However, the conventional PI controller can be designed simply but it is not stable under the variation of parameters and load torque disturbance. Also, slip calculation of IFOC depends on rotor resistance, which changes due to the rising temperature.

The control of IM drive using multi-input and multi-output (MIMO) optimal regulator has been proposed in [4, 5]. It is clarified in [4, 5] that the MIMO optimal regulator is stable against the variation of parameter and load torque disturbance. Therefore, MIMO optimal regulator is better than conventional PI controller. In [5], we designed MIMO optimal regulator to improve efficiency of IM drive taking core loss into account without any observer for estimation of IM drive.

In the IFOC, the reference magnetizing current and slip angular frequency are calculated by the rotor flux where core loss is taken into account[2]. Consequently, the estimation of states of IM has become a very important issue in control applications of IM drives. The estimation techniques for the IM drives are very fascinating and challenging subjects of R & D, and recently, they have received wide attention in the research works[6),7)]. Most of those

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observers are designed only to estimate rotor flux with neglected core loss in stationary reference frame in a linear form. The rotor current and rotor flux interfere with each other due to the effects of core loss. Therefore, it is important to estimate rotor current and rotor flux as a compensation for the degradation of core loss.

In this paper, we proposed a control system by using MIMO optimal regulator theory from the standpoint of modern control to overcome the complexity of conventional decoupling control in which the simple state equations are used to realize the FOC. The FOC with consideration of core loss can be easily done by the combination of stator voltage and primary angular frequency as inputs in the controller[7]. The slip frequency is found out directly from the proposed controller to achieve the constraints of IFOC and the proposed controller is stable for the variation of load disturbance torque and parameters unlike conventional vector control.

To implement MIMO optimal regulator, the state equations are derived as internal states of the stator current, rotor current, rotor flux and rotor speed where the stator current and rotor speed are measurable only. Since the state equations of an IM with consideration of core loss in a synchronously rotating reference frame have behaved as bilinear, we proposed a minimal order observer for bilinear system of IM to estimate the rotor current and rotor flux. The state computation time of the proposed observer is less than that of the full-order observer. The estimation error using the proposed observer method converges to zero irrespective of inputs. As a result, the proposed observer is suitable for working with the MIMO optimal regulator.

Simulation results demonstrate that MIMO optimal regulator control is capable of realizing FOC in spite of the degradation of core loss and the proposed estimation method is capable of estimating the rotor current and rotor flux while the rotor speed is either constant or time varying. Using Matlab/Simulink[11], simulations were carried out. The simulation results show the good response of proposed controller and observer system.

2. Model and Equations for the Induction Motor with Consideration of Core Loss

A. Physical Model

In the d-q axis equivalent circuit for FOC of IM considering core loss due to eddy current[2),5)], a resistance R_m is connected in parallel across the internal induced voltage branch. The d-q axis equivalent circuit of an IM in a synchronously rotating reference frame is shown in **Fig. 1**.

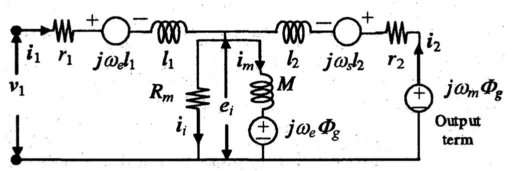

Fig. 1 Space vector equivalent circuit of an IM taking core loss into account

Fig. 1 represents the combination of a well-known field oriented control d-q axis equivalent circuit with neglected core loss and a conventional approach to core loss modeling in a steady state phasor equivalent circuit. The resistance of this equivalent circuit depends on the frequency and flux level, and the ohmic value of R_m can be determined experimentally, which is investigated in detail in [1, 2].

B. Voltage and Flux Linkage Equations

The voltages and fluxes linkage equations with consideration of core loss due to the eddy current of an IM in a d-q axis synchronously rotating reference frame (called arbitrary reference frame) can be written as:

$$\boldsymbol{v}_1 = r_1 \boldsymbol{i}_1 + l_1(d\boldsymbol{i}_1/dt) + j\omega_e l_1 \boldsymbol{i}_1 + \boldsymbol{e}_i \quad (2.1)$$

$$0 = r_2 \boldsymbol{i}_2 + (d\boldsymbol{\Phi}_2/dt) + j(\omega_e - \omega_m)\boldsymbol{\Phi}_2 \quad (2.2)$$

$$\boldsymbol{e}_i = R_m \boldsymbol{i}_i = (d\boldsymbol{\Phi}_g/dt) + j\omega_e \boldsymbol{\Phi}_g \quad (2.3)$$

$$\boldsymbol{\Phi}_1 = l_1 \boldsymbol{i}_1 + \boldsymbol{\Phi}_g \quad (2.4)$$

$$\boldsymbol{\Phi}_2 = l_2 \boldsymbol{i}_2 + \boldsymbol{\Phi}_g \quad (2.5)$$

$$\boldsymbol{\Phi}_g = M \boldsymbol{i}_m$$

$$\boldsymbol{i}_i + \boldsymbol{i}_m = \boldsymbol{i}_1 + \boldsymbol{i}_2 \quad (2.6)$$

The torque developed in an IM is obtained from the power balance at the rotor terminals. This torque is given by

$$T_e = pIm(\boldsymbol{i}_2^* \boldsymbol{\Phi}_2)$$

$$(d\omega_m/dt) = -(D/J)\omega_m + (p/J)(T_e - T_L) \quad (2.7)$$

Symbols r_1, r_2, R_m indicate stator, rotor and core loss resistances. L_s, L_r indicate stator and rotor self-indutances. l_1, l_2 indicate stator and rotor leakage inductances. M, p indicate mutual inductance and number of pole pair. ω_e, ω_m, ω_s indicate primary, rotor and slip speeds. $\boldsymbol{v}_1$, $\boldsymbol{e}_i$ indicate stator and internal induced voltage vectors. $\boldsymbol{i}_1$, $\boldsymbol{i}_2$, $\boldsymbol{i}_m$, $\boldsymbol{i}_i$ indicate stator, rotor, magnetizing and core loss current vectors. $\boldsymbol{\Phi}_1$, $\boldsymbol{\Phi}_2$, $\boldsymbol{\Phi}_g$ indicate stator, rotor and air gap flux vectors. T_e, T_L indicate electromagnetic and load torques. J, D indicate total inertia and damping factor. Δ, k indicate the first difference value and sampling time. "T", "s" indicate the transposition of matrix and steady state value.

C. State Equations of the Physical Model

The state equations with neglecting core loss were derived[4] in which the state variables are considered as stator current, rotor flux and rotor speed. Only by these states the physically insight of FOC of IM cannot be realized where core loss exists. For this purpose, we considered the state variables are rotor speed (ω_m), stator current ($\boldsymbol{i}_1$), rotor current ($\boldsymbol{i}_2$) and rotor flux ($\boldsymbol{\Phi}_2$).

From equations (2.1)～(2.7), we can derive the state equations as

$$(d\omega_m/dt)=-(D/J)\omega_m+(p/J)T_e-(p/J)T_L \quad (2.8)$$

$$(d\boldsymbol{i}_1/dt)=a_{r11}\boldsymbol{i}_1+ja_{i11}\boldsymbol{i}_1+a_{r12}\boldsymbol{i}_2+a_{r13}\boldsymbol{\Phi}_2-j\omega_s\boldsymbol{i}_1+b_1\boldsymbol{v}_1 \quad (2.9)$$

$$(d\boldsymbol{i}_2/dt)=a_{r21}\boldsymbol{i}_1+a_{r22}\boldsymbol{i}_2+ja_{i22}\boldsymbol{i}_2+a_{r23}\boldsymbol{\Phi}_2+ja_{i23}\boldsymbol{\Phi}_2-j\omega_s\boldsymbol{i}_2 \quad (2.10)$$

$$(d\boldsymbol{\Phi}_2/dt)=a_{r32}\boldsymbol{i}_2-j\omega_s\boldsymbol{\Phi}_2 \quad (2.11)$$

where, $a_{r11}=-(r_1+R_m)/l_1$, $a_{i11}=-\omega_m$,

$a_{r12}=-R_mL_r/Ml_1$, $a_{r13}=R_m/Ml_1$, $a_{r21}=-R_m/l_2$,

$a_{r22}=-R/l_2$, $a_{i22}=-\omega_m$, $a_{r23}=R_m/Ml_2$,

$a_{i23}=\omega_m/l_2$, $a_{r32}=-r_2$, $b_1=1/l_1$, $R=r_2+(R_mL_r/M)$.

Equation (2.10) includes state equation to eliminate the degradation of dynamic response for the undesired effects of core loss.

3. Bilinear Model to Estimate Rotor Current and Rotor Flux

A. Bilinear Model

It is necessary to know the values of rotor current and rotor flux to overcome the effects of core loss. But it is very difficult to measure the rotor current and rotor flux of IM practically. Therefore, the rotor current and rotor flux must be calculated or estimated from the motor parameters and output information such as speed and stator current.

In this paper, an observer method for bilinear system[8),9)] is applied to estimate the rotor current and rotor flux. Equations (2.9)～(2.11) can be written as equation (3.1)

$$\begin{bmatrix}\dot{x}_1\\ \dot{x}_2\end{bmatrix}=\begin{bmatrix}\boldsymbol{A}_{11} & \boldsymbol{A}_{12}\\ \boldsymbol{A}_{21} & \boldsymbol{A}_{22}\end{bmatrix}\begin{bmatrix}\boldsymbol{x}_1\\ \boldsymbol{x}_2\end{bmatrix}+\omega_s\begin{bmatrix}\boldsymbol{A}_{11}^i & \boldsymbol{0}_{2\times4}\\ \boldsymbol{0}_{4\times2} & \boldsymbol{A}_{22}^i\end{bmatrix}\begin{bmatrix}\boldsymbol{x}_1\\ \boldsymbol{x}_2\end{bmatrix}+\begin{bmatrix}\boldsymbol{B}_1\\ \boldsymbol{0}_{4\times2}\end{bmatrix}\boldsymbol{u}$$

$$\dot{\boldsymbol{x}}=\boldsymbol{A}\boldsymbol{x}+\boldsymbol{A}^i\omega_s\boldsymbol{x}+\boldsymbol{B}\boldsymbol{u} \quad (3.1)$$

The stator current, which is measurable, is taken as the output and expressed as

$$\boldsymbol{y}=\boldsymbol{C}\boldsymbol{x} \quad (3.2)$$

where

$\boldsymbol{x}=[\boldsymbol{x}_1, \boldsymbol{x}_2]^T$, $\boldsymbol{x}_1=[i_{1d}, i_{1q}]^T$, $\boldsymbol{x}_2=[i_{2d}, i_{2q}, \Phi_{2d}, \Phi_{2q}]^T$,

$\boldsymbol{u}=[v_{1d}, v_{1q}]^T$, $\boldsymbol{y}=[i_{1d}, i_{1q}]^T$, $\boldsymbol{A}_{11}=\boldsymbol{a}_{11}$, $\boldsymbol{A}_{12}=[\boldsymbol{a}_{12}, \boldsymbol{a}_{13}]$,

$\boldsymbol{A}_{21}=[\boldsymbol{a}_{21}, \boldsymbol{0}_{2\times2}]^T$, $\boldsymbol{A}_{22}=\begin{bmatrix}\boldsymbol{a}_{22} & \boldsymbol{a}_{23}\\ \boldsymbol{a}_{32} & \boldsymbol{0}_{2\times2}\end{bmatrix}$, $\boldsymbol{A}_{11}^i=\boldsymbol{a}_{11}^i$,

$\boldsymbol{A}_{22}^i=\begin{bmatrix}\boldsymbol{a}_{22}^i & \boldsymbol{0}_{2\times2}\\ \boldsymbol{0}_{2\times2} & \boldsymbol{a}_{33}^i\end{bmatrix}$, $\boldsymbol{I}=\begin{bmatrix}1 & 0\\ 0 & 1\end{bmatrix}$, $\boldsymbol{J}=\begin{bmatrix}0 & -1\\ 1 & 0\end{bmatrix}$, $\boldsymbol{C}=[\boldsymbol{I}\ \boldsymbol{0}_{2\times4}]$,

$\boldsymbol{a}_{11}=a_{r11}\boldsymbol{I}+a_{i11}\boldsymbol{J}$, $\boldsymbol{a}_{12}=a_{r12}\boldsymbol{I}$, $\boldsymbol{a}_{13}=a_{r13}\boldsymbol{I}$, $\boldsymbol{a}_{21}=a_{r21}\boldsymbol{I}$,

$\boldsymbol{a}_{22}=a_{r22}\boldsymbol{I}+a_{i22}\boldsymbol{J}$, $\boldsymbol{a}_{23}=a_{r23}\boldsymbol{I}+a_{i23}\boldsymbol{J}$, $\boldsymbol{a}_{32}=a_{r32}\boldsymbol{I}$,

$\boldsymbol{a}_{11}^i=\boldsymbol{a}_{22}^i=\boldsymbol{a}_{33}^i=-\boldsymbol{J}$, $\boldsymbol{B}_1=b_1\boldsymbol{I}$.

$\boldsymbol{0}_{n\times m}$ indicates a zero matrix with dimension $n\times m$.

Equation (3.1) shows that the state equations of an IM become a bilinear system in terms of the product of the input slip angular frequency and internal state variables.

B. Rotor Current and Rotor Flux Estimation Method

From equation (3.1), the minimal order observer equation with an error correction term is derived as follows:

$$\dot{\hat{\boldsymbol{x}}}_2=\boldsymbol{A}_{21}\boldsymbol{x}_1+\boldsymbol{A}_{22}\hat{\boldsymbol{x}}_2+\omega_s\boldsymbol{A}_{22}^i\hat{\boldsymbol{x}}_2+\boldsymbol{G}[\dot{\boldsymbol{x}}_1-\{(\boldsymbol{A}_{11}\boldsymbol{x}_1+\boldsymbol{A}_{12}\hat{\boldsymbol{x}}_2)+\omega_s\boldsymbol{A}_{11}^i\boldsymbol{x}_1+\boldsymbol{B}_1\boldsymbol{u}\}] \quad (3.3)$$

where $\hat{\boldsymbol{x}}_2$ is the estimated rotor current and rotor flux and $\boldsymbol{G}$ is gain matrix.

It is desirable to eliminate the differential term of the $\boldsymbol{x}_1$ in (3.3). Thus, substituting a new variable as

$$\boldsymbol{z}=\hat{\boldsymbol{x}}_2-\boldsymbol{G}\boldsymbol{x}_1 \quad (3.4)$$

into (3.3), $\dot{\boldsymbol{x}}_1$ disappears as in (3.5).

$$\dot{\boldsymbol{z}}=[\boldsymbol{D}+\omega_s\boldsymbol{D}^i]\boldsymbol{z}+[\boldsymbol{E}+\omega_s\boldsymbol{E}^i]\boldsymbol{y}+\boldsymbol{L}\boldsymbol{u} \quad (3.5)$$

We can define a new variable in terms of $\boldsymbol{z}$ and $\boldsymbol{y}$ as follows:

$$\boldsymbol{w}=\boldsymbol{F}\boldsymbol{z}+\boldsymbol{H}\boldsymbol{y} \quad (3.6)$$

where $\boldsymbol{D}=\boldsymbol{A}_{22}-\boldsymbol{G}\boldsymbol{A}_{12}$, $\boldsymbol{D}^i=\boldsymbol{A}_{22}^i$, $\boldsymbol{L}=-\boldsymbol{G}\boldsymbol{B}_1$,

$\boldsymbol{H}=[\boldsymbol{I}_2|\boldsymbol{G}]^T$, $\boldsymbol{E}=\boldsymbol{A}_{21}-\boldsymbol{G}\boldsymbol{A}_{11}+\boldsymbol{D}\boldsymbol{G}$,

$\boldsymbol{E}^i=-\boldsymbol{G}\boldsymbol{A}_{11}^i+\boldsymbol{D}^i\boldsymbol{G}$, $\boldsymbol{F}=[\boldsymbol{0}_{2\times4}|\boldsymbol{I}_4]^T$.

From equation (3.3) and (3.5), the error dynamics is governed by the following equation:

$$\boldsymbol{e}=\boldsymbol{z}-\boldsymbol{U}\boldsymbol{x} \quad (3.7)$$

$$\dot{\boldsymbol{e}}=[\boldsymbol{D}+\omega_s\boldsymbol{D}^i]\boldsymbol{e}+[\boldsymbol{D}\boldsymbol{U}+\boldsymbol{E}\boldsymbol{C}-\boldsymbol{U}\boldsymbol{A}]\boldsymbol{x}+\omega_s[\boldsymbol{D}^i\boldsymbol{U}+\boldsymbol{E}^i\boldsymbol{C}-\boldsymbol{U}\boldsymbol{A}^i]\boldsymbol{x}+[\boldsymbol{L}-\boldsymbol{U}\boldsymbol{B}]\boldsymbol{u} \quad (3.8)$$

For a given slip frequency ω_s or rotor angular frequency ω_m, equation (3.7) is a linear system. Therefore, the error dynamics (3.8) is enough for estimating the rotor current and rotor flux at the constant ω_s or ω_m by satisfying the following conditions:

(i) $\boldsymbol{D}\boldsymbol{U}+\boldsymbol{E}\boldsymbol{C}-\boldsymbol{U}\boldsymbol{A}=0$ (3.9.1)

(ii) $\boldsymbol{D}^i\boldsymbol{U}+\boldsymbol{E}^i\boldsymbol{C}-\boldsymbol{U}\boldsymbol{A}^i=0$ (3.9.2)

(iii) $\boldsymbol{L}-\boldsymbol{U}\boldsymbol{B}=0$ (3.9.3)

(iv) $\boldsymbol{F}\boldsymbol{U}+\boldsymbol{H}\boldsymbol{C}=\boldsymbol{K}=\boldsymbol{I}_6$ (3.9.4)

In order to satisfy the above equations (3.9.1)～(3.9.4), it is required to properly choose arbitrary matrices $\boldsymbol{G}$ and $\boldsymbol{U}$. Hence, we have chosen these matrices as:

$$\boldsymbol{G}=[g_1\boldsymbol{I}+g_2\boldsymbol{J}, g_3\boldsymbol{I}+g_4\boldsymbol{J}]^T \text{ and } \boldsymbol{U}=[-\boldsymbol{G}|\boldsymbol{I}_4]$$

The estimation error converges to zero not depending on $\boldsymbol{x}$ and input voltages v_{1d} and v_{1q} because the chosen matrices fulfill the condition (3.9.1)～(3.9.4). Consequently, the error (3.8) can be written as follows:

$$\dot{\boldsymbol{e}}=[\boldsymbol{D}+\omega_s\boldsymbol{D}^i]\boldsymbol{e} \quad (3.10)$$

For choosing the gain matrix $\boldsymbol{G}$ within stable condition, the Lyapunov's function can be taken as follows:

$$V(\boldsymbol{e})=\boldsymbol{e}^T\boldsymbol{e} \quad (3.11)$$

According to (3.10) and (3.11), the derivative of Lyapunov's function can be written as follows:

$$\dot{V}(\boldsymbol{e})=\boldsymbol{e}^T\{[\boldsymbol{D}+\boldsymbol{D}^T]+\omega_s[\boldsymbol{D}^i+\boldsymbol{D}^{iT}]\}\boldsymbol{e} \quad (3.12)$$

For the stability of equation (3.12) and convergence of error to zero, it is required to satisfy the following conditions:

(v) $\boldsymbol{D}^i+\boldsymbol{D}^{iT}=0$ (3.13.1)

(vi) $\boldsymbol{D}+\boldsymbol{D}^T=-\boldsymbol{N}$ (3.13.2)

(vii) All eigenvalues of $\boldsymbol{D}+\omega_s\boldsymbol{D}^i$ have negative real parts.

It is seen that the matrix $\boldsymbol{D}^i$ is complex matrix without any real value and $\boldsymbol{D}^{iT}$ becomes complex conjugate of $\boldsymbol{D}^i$. As a result, the condition (v) is satisfied and the convergence of error does not depend on input ω_s. From the previous values and condition (3.13.2), the matrix $\boldsymbol{N}$ can be written as the following:

$$\boldsymbol{N}=\begin{bmatrix} n_{r11}\boldsymbol{I} & n_{r12}-n_{i12}\boldsymbol{J} \\ n_{r12}\boldsymbol{I}+n_{i12}\boldsymbol{J} & n_{r22}\boldsymbol{I} \end{bmatrix}$$

where

$n_{r11}=2(a_{r12}g_1-a_{r22})$, $n_{i12}=a_{i23}-a_{r13}g_2+a_{r12}g_4$,

$n_{r12}=-a_{r23}+a_{r13}g_1-a_{r32}+a_{r12}g_3$, $n_{r22}=2a_{r13}g_3$.

If we choose $n_{r12}=0$ and $n_{i12}=0$, the matrix $\boldsymbol{N}$ becomes a diagonal matrix. Therefore, g_4, g_2 and g_1 can be chosen as follows:

$$g_1=(a_{r32}+a_{r23}-a_{r12}g_3)/a_{r13}$$

$$g_4=0,\ g_2=a_{i23}/a_{r13} \quad (3.14)$$

To become the matrix $\boldsymbol{N}$ as positive definite matrix, all diagonal elements should be positive. Since a_{r13} is positive, the value of g_3 must be positive for the positive value of n_{r22}. For the positive value of n_{r11} the value of g_1 should satisfy the condition $g_1<(a_{r22}/a_{r12})$. It has been tested that the calculated value of g_1 from (3.14) is less than (a_{r22}/a_{r12}) if g_3 is chosen less than 0.0001. We chose the value of $g_4=0$, $g_3=0.00001$ for simulation work. The value of g_2 is changed with changing of speed and g_1 is changed with changing of a_{r32}, a_{r23} and a_{r12}. From above discussion, it is cleared that the matrix $\boldsymbol{N}$ becomes a positive definite matrix. Also, the condition (vii) is as satisfied shown in **Table 1** for different speeds at full load torque and equation (3.12) becomes

$$\dot{V}(\boldsymbol{e})=-\boldsymbol{e}^T\boldsymbol{N}\boldsymbol{e} \quad (3.15)$$

and by the well-known stability theorem of Lyapunov

$$\mathrm{Lim}_{t\to\infty}\boldsymbol{e}(t)=0 \quad (3.16)$$

As the matrix $\boldsymbol{N}$ becomes positive definite, the proposed observer system is stable and the error converges to zero irrespective of inputs.

Combining equations (3.2), (3.6) and (3.7), we obtained

$$\boldsymbol{w}=[\boldsymbol{FU}+\boldsymbol{HC}]\boldsymbol{x}+\boldsymbol{Fe} \quad (3.17)$$

Since condition (iv) is satisfied, then by (3.16)

$$\mathrm{Lim}_{t\to\infty}\|\boldsymbol{w}-\boldsymbol{Kx}\|=0 \quad (3.18)$$

Therefore, the dynamic system (3.5) is a stable $\boldsymbol{Kx}$ observer for (3.1).

Table 1 Real roots and value of g_4 in different speeds

Speed [r/min]	g_4	Real roots for equation (3.10)	
		$a_{i2d}=a_{i2q}\times10^{03}$	$a_{\Phi2d}=a_{\Phi2q}\times10^{03}$
100.0	0.0016	−0.1438	−1.0052
500.0	0.0078	−0.1432	−1.0058
800.0	0.0125	−0.1431	−1.0059
900.0	0.0141	−0.1431	−1.0059

4. Influence of Core Loss into Field Oriented Control

The ideal FOC of IM is illustrated in Fig. 2, where it is shown the rotor flux and torque decoupling control in terms of the stator d-axis and q-axis current. The position of rotor flux is calculated using rotor speed and slip frequency values. Therefore, the manipulation of IFOC is implemented by setting the q-axis component of rotor flux and d-axis component of rotor current zero.

With the constraints for IFOC of IM, we can derive the following equations from equations (2.1)～(2.6) at steady state condition:

$$\Phi_{2d}=Mi_{1d}-(Ml_2/R_m)\omega_e i_{2q} \quad (4.1)$$

$$T_e=p\frac{M}{L_r}[i_{1q}\Phi_{2d}+\frac{M}{R_m}(l_2\omega_e^2 i_{2q}\Phi_{2d}-\omega_e i_{1d}\Phi_{2d})] \quad (4.2)$$

$$\omega_s=(r_2M/L_r)[(i_{1q}/\Phi_{2d})-(\omega_e/R_m)] \quad (4.3)$$

$$R_m=\omega_e^2(l_2^2i_{2q}^2+\Phi_{2d}^2)/P_i \quad (4.4)$$

where P_i is experimentally measured core loss.

From equation (4.4), it can easily be understood that the core loss resistance is a function of frequency and flux level, which is explained in detail in [2]. The frequency

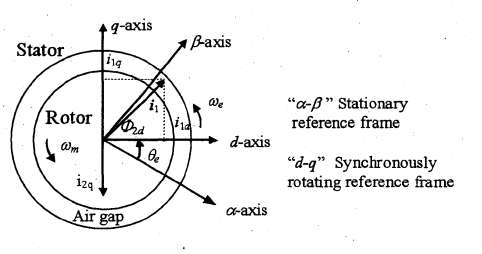

Fig. 2 Ideal field oriented control representation

and flux level characteristics of the core loss resistance can be derived at steady state condition using equation (4.4). The core loss resistance values have been determined experimentally whose procedure was explained in [2].

The first terms of equations (4.1)～(4.3) are used for the case of ideal condition of FOC and others terms are for the influence of core loss.

It is clear from the aforementioned equations that the degradation is occurred where the reference value of rotor flux and torque are calculated only by using stator current. Consequently, the decoupling control becomes very complex in the mathematical model and controller design.

To overcome the complexity of FOC in terms of stator current components, an IFOC in [2] has been proposed in terms of magnetizing current components. The rotor flux and torque linear equations in terms of magnetizing currents are given[2] as the following:

$$\Phi_{2d}=Mi_{md} \quad (4.5)$$

$$T_e=p(M/l_2)i_{mq}\Phi_{2d} \quad (4.6)$$

$$\omega_s=(r_2M/l_2)(i_{mq}/\Phi_{2d}) \quad (4.7)$$

$$i_{1d}=i_{md}-(M/R_m)\omega_e i_{mq} \quad (4.8)$$

$$i_{1q}=(L_r/l_2)i_{mq}+(M/R_m)\omega_e i_{md} \quad (4.9)$$

$$v_{1d}=(r_1+R_m)i_{1d}-R_m i_{md}-\omega_e l_1 i_{1q} \quad (4.10)$$

$$v_{1q}=(r_1+R_m)i_{1q}-(R_mL_r/l_2)i_{mq}+\omega_e l_1 i_{1d} \quad (4.11)$$

The IFOC, according to (4.5)～(4.7), is valid for steady state and it is sensitive to parameter variations. Using the equations (4.5)～(4.6), the stator voltage, stator current and rotor current can be carried out as (4.8)～(4.11).

For the compensation of undesired effects of core loss, we proposed the MIMO optimal regulator from the standpoint of modern control for the voltage source inverter-fed induction motor drive. The position of rotor flux can be adjusted by the control inputs such as slip frequency, stator d-q axis voltages.

5. MIMO Optimal Regulator Control as Compensation for Core Loss

The state equations (2.8)～(2.11) show the nonlinear property of an IM drive. The optimal regulator theory is based on the linear system in an operating point. After linearization of state equations, it can be written in a compact form as following.

Controlled object: $$\dot{\boldsymbol{x}}_c=\boldsymbol{A}_c\boldsymbol{x}_c+\boldsymbol{B}_c\boldsymbol{u}_c+\boldsymbol{E}_c\boldsymbol{d}_c \quad (5.1)$$

Output: $$\boldsymbol{y}_c=\boldsymbol{C}_c\boldsymbol{x}_c \quad (5.2)$$

where $\boldsymbol{x}_e=[x_{c1}, x_{c2}, x_{c3}, x_{c4},]^T$, $x_{c1}=\omega_m$, $\boldsymbol{x}_{c2}=[i_{1d}, i_{1q}]^T$, $\boldsymbol{x}_{C3}=[i_{2d}, i_{2q}]^T$, $\boldsymbol{x}_{c4}=[\Phi_{2d}, \Phi_{2q}]^T$, $\boldsymbol{u}_c=[u_{c1}, \boldsymbol{u}_{c2}]^T$, $u_{c1}=\omega_s$, $\boldsymbol{u}_{c2}=[v_{1d}, v_{1q}]^T$, $\boldsymbol{d}_c=\boldsymbol{T}_L$, $\boldsymbol{y}_c=[\boldsymbol{y}_{c1}, \boldsymbol{y}_{c2}]^T$, $\boldsymbol{y}_{c1}=\omega_m$,

$$\boldsymbol{y}_{c2}=[\Phi_{2d}, \phi_{2q}]^T,\ \boldsymbol{C}_c=\begin{bmatrix}1 & \boldsymbol{0}_{1\times2} & \boldsymbol{0}_{1\times2} & \boldsymbol{0}_{1\times2}\\ 0 & \boldsymbol{0}_{2\times2} & \boldsymbol{0}_{2\times2} & \boldsymbol{I}_{2\times2}\end{bmatrix},\ \boldsymbol{E}_c=\begin{bmatrix}e_{c11}\\ \boldsymbol{0}_{6\times1}\end{bmatrix}$$

$$\boldsymbol{A}_c=\begin{bmatrix}a_{c11} & \boldsymbol{0}_{1\times2} & \boldsymbol{a}_{c13} & \boldsymbol{a}_{c14}\\ \boldsymbol{a}_{c21} & \boldsymbol{a}_{c22} & \boldsymbol{a}_{c23} & \boldsymbol{a}_{c24}\\ \boldsymbol{a}_{c31} & \boldsymbol{a}_{c32} & \boldsymbol{a}_{c33} & \boldsymbol{a}_{c34}\\ \boldsymbol{0}_{2\times1} & \boldsymbol{0}_{2\times2} & \boldsymbol{a}_{c43} & \boldsymbol{a}_{c44}\end{bmatrix},\ \boldsymbol{B}_c=\begin{bmatrix}0 & \boldsymbol{0}_{1\times2}\\ \boldsymbol{b}_{c21} & \boldsymbol{b}_{c22}\\ \boldsymbol{b}_{c31} & \boldsymbol{0}_{1\times2}\\ \boldsymbol{0}_{1\times2} & \boldsymbol{b}_{c42}\end{bmatrix},$$

$a_{c11}=-D/J$, $\boldsymbol{a}_{c13}=(p^2/J)[\Phi^s_{2q}, -\Phi^s_{2d}]$,
$\boldsymbol{a}_{c14}=(p^2/J)[-i^s_{2q}, i^s_{2d}]$, $\boldsymbol{a}_{c21}=[i^s_{1q}, -i^s_{1d}]^T$,
$\boldsymbol{a}_{c22}=-a_{r11}\boldsymbol{I}-(\omega^s_m i_{1d}+\omega^s_s)\boldsymbol{J}$, $\boldsymbol{a}_{c23}=a_{r12}\boldsymbol{I}$, $\boldsymbol{a}_{c24}=a_{r13}\boldsymbol{I}$,
$\boldsymbol{a}_{c31}=[\{i^s_{2q}-(\Phi^s_{2q}/l_2)\}, \{i^s_{2d}+(\Phi^s_{2d}/l_2)\}]^T$,
$\boldsymbol{a}_{c32}=a_{r21}\boldsymbol{I}$, $\boldsymbol{a}_{c33}=a_{r22}\boldsymbol{I}-(\omega^s_m+\omega^s_s)\boldsymbol{J}$,
$\boldsymbol{a}_{c34}=a_{r23}\boldsymbol{I}+(\omega^s_m/l_2)\boldsymbol{J}$, $\boldsymbol{a}_{c43}=a_{r32}\boldsymbol{I}$, $\boldsymbol{a}_{c44}=-\omega^s_s\boldsymbol{J}$,
$\boldsymbol{b}_{c21}=[i^s_{1q}, i^s_{1d}]^T$, $\boldsymbol{b}_{c22}=(1/l_1)\boldsymbol{I}$, $\boldsymbol{b}_{c31}=[i^s_{2q}, -i^s_{2d}]^T$,
$\boldsymbol{b}_{c41}=[\Phi^s_{2q}, -\Phi^s_{2d}]^T$, $e_{c11}=-p/J$.

The state space model of the IM system transformed into the synchronous reference frame becomes a three-input and three-output system as (5.1) and (5.2). Equations (5.3) and (5.4) are obtained by transforming (5.1) and (5.2) into the discrete time form[4),5)], where the sampling period is τ.

$$\boldsymbol{x}_c(k+1)=\boldsymbol{A}_c(k)\boldsymbol{x}_c(k)+\boldsymbol{B}_c(k)\boldsymbol{u}_c(k)+\boldsymbol{E}_c(k)\boldsymbol{d}_c(k) \quad (5.3)$$

$$\boldsymbol{y}_c(k)=\boldsymbol{C}_c\boldsymbol{x}_c(k) \quad (5.4)$$

Matrices of equation (5.3) can be defined as follows:

$$\boldsymbol{A}_c(k)=e^{A_c\tau},\ \boldsymbol{B}_c(k)=e^{(A_c\tau/2)}\boldsymbol{B}_c,\ \boldsymbol{E}_c(k)=e^{(A_c\tau/2)}\boldsymbol{E}_c$$

The error term $\boldsymbol{e}_o(k)$ between the reference value $\boldsymbol{R}(k)$ and the output $\boldsymbol{y}(k)$ is introduced as a new state.

$$\boldsymbol{e}_o(k)=\boldsymbol{R}(k)-\boldsymbol{y}_c(k) \quad (5.5)$$

$$\boldsymbol{e}_o(k)=[e_{\omega m}, e_{\Phi 2d}, e_{\Phi 2q}]^T$$

where

$$\boldsymbol{R}(k)=[\omega^R_m, \Phi^R_{2d}, \Phi^R_{2q}]^T \quad (5.6)$$

Here, $\boldsymbol{R}(k)$ is reference signal for the controlled variable.

The state feedback control is basically a proportional control, the steady state error may exist due to the model uncertainty. To remove this error, the deviation term $\Delta\boldsymbol{x}(k+1)$ between the state $\boldsymbol{x}(k+1)$ and $\boldsymbol{x}(k)$ is introduced as a new state.

$$\Delta\boldsymbol{x}(k+1)=\boldsymbol{x}(k+1)-\boldsymbol{x}(k) \quad (5.7)$$

We assumed that all states of equation (5.1) are controllable and observable and that the equation shows the actual controlled object. An augmented system called the error system can be written as

$$\begin{bmatrix}\boldsymbol{e}_o(k+1)\\ \Delta\boldsymbol{x}(k+1)\end{bmatrix}=\begin{bmatrix}\boldsymbol{I} & -\boldsymbol{CA}\\ 0 & \boldsymbol{A}\end{bmatrix}\begin{bmatrix}\boldsymbol{e}_o(k)\\ \Delta\boldsymbol{x}(k)\end{bmatrix}+\begin{bmatrix}-\boldsymbol{CB}\\ \boldsymbol{B}\end{bmatrix}\begin{bmatrix}\Delta u_{c1}(k)\\ \Delta\boldsymbol{u}_{c2}(k)\end{bmatrix} \quad (5.8)$$

Let us assume that the reference signal for speed control is a step function, therefore the state equation (5.8) of the error system can be written by the new variable as:

$$\boldsymbol{X}(k+1)=\boldsymbol{\Psi}\boldsymbol{X}(k)+\boldsymbol{G}\Delta\boldsymbol{u}(k) \quad (5.9)$$

where $X(k)=[e_o(k) \quad \Delta x(k)]^T$ (5.10)

Ψ : 10×10 matrix, G : 3×10 matrix.

It is clear that the sensitivity to the parameter variation can be attained as long as the error system (5.9) exists. The system (5.9) is controlled to be stable in spite of the parameter variations of the control object (5.1).

The performance index or cost function at each sampling time $k(k=1, 2, \cdots)$, $J(k)$ is defined as

$$J(k)=\sum_{k=1}^{\infty}[X(k)^T QX(k)+u(k)^T Ru(k)] \tag{5.11}$$

where Q is 10×10 and R is 3×3 matrices.

The matrix R is assumed to be symmetrical and positive definite, whereas the matrix Q is assumed to be symmetrical and positive semi-definite. The matrices Q and R were chosen as following:

$$Q=\text{diag}[Q_{11}, Q_{22}, Q_{33}, Q_{44}, 0, 0, 0, 0, 0, 0]$$

$$R=\text{diag}[R_{11}, R_{22}, R_{33}]$$

By using optimal control theory [12], the control input can be derived easily as

$$\Delta u(k)=F_B X(k) \tag{5.12}$$

$$F_B=-[R+G^T PG]^{-1}G^T P\Psi \tag{5.13}$$

where F_B is 3×10 matrix.

To evaluate the feedback gain matrix, it is necessary to solve the Riccati equation in steady state. Then the matrix P becomes

$$P=Q+\Psi^T P\Psi-\Psi^T PG[R+G^T PG]^{-1}G^T P\Psi \tag{5.14}$$

According to the solution of (5.12) and (5.13), the poles of augmented system (5.9) lie into the unit circle[12]. Consequently, the design controller is stable against the variation of parameters and load torque disturbance of IM drive. To realize the control system by the simple block diagram, it is needed the solution of equation (5.12) as

$$u(k)=F_e\sum_{i=0}^{k}e_0(i)+F_x x(k)-F_x x(0)+u(0) \tag{5.15}$$

where F_e is 3×3 and F_x is 3×7 matrices which are carried out from the matrix F_B.

Since the feedback gain of this method is derived from the linearization at operating point, it should be followed a look up table in terms of required implementation specific operating points such as speed and torque, provided that the appropriate feedback gain matrix values have been determined by simulation.

From the above equation (5.15), we can define that first term represents integral action, second term represents proportional action and other terms represent compensating action. **Fig. 3** shows the block diagram for the overall system of proposed controller and observer system of IM drive.

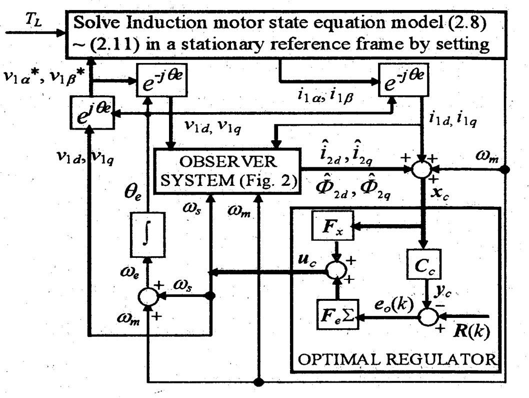

Fig. 3 Overall system block diagram

The block involving exponential $e^{-j\theta e}$ in Fig. 3 represents a coordinate transformation from a stationary to a synchronously rotating coordinate system given by

$$i_{1d}=i_{1\alpha}\cos\theta_e+i_{1\beta}\sin\theta_e \tag{5.16}$$

$$i_{1q}=-i_{1\alpha}\sin\theta_e+i_{1\beta}\cos\theta_e \tag{5.17}$$

The inverse $e^{j\theta e}$ quantity represents the transformation from synchronous rotating to stationary coordinate system given by

$$i_{1\alpha}=i_{1d}\cos\theta_e-i_{1q}\sin\theta_e \tag{5.18}$$

$$i_{1\beta}=i_{1d}\sin\theta_e+i_{1q}\cos\theta_e \tag{5.19}$$

The superscript * in Fig. 3 denotes the desired quantities. The voltage and flux can be transformed from stationary frame to rotating frame or rotating frame to stationary frame according to the same equations (5.16)～(5.19).

6. Controller with Observer

The controller using MIMO optimal regulator was designed in section 5 to minimize the performance index (5.11) and the poles are placed into the unit circle of augmented system (5.9). Since the poles are located into unit circle, the controller is stable under the variation of parameters and load torque disturbance. The stability of observer system was established in section 3. Therefore, the controller and the observer system are stable individually.

This estimated rotor current and rotor flux are fed back to the controller. The augmented system (5.9) of controller can be written as

$$\hat{X}(k+1)=\Psi\hat{X}(k)+G\Delta u(k) \tag{6.1}$$

Equation (6.1) differs from equation (5.9) by the fact that we use estimated values of the state variables. Now, the whole stability depends on the augmented system (6.1).

To show the stability of system (6.1) by Lyapunov's stability theorem, we chose the Lypunov function as follows:

$$V[\hat{X}(k)]=\hat{X}^T(k)P_w\hat{X}(k) \tag{6.2}$$

where $\boldsymbol{P}_w$ is a positive definite matrix.

The equations (6.1) can be deduced as follows:

$$\hat{X}(k+1)=\boldsymbol{\Psi}_c\hat{X}(k) \quad (6.3)$$

where $\boldsymbol{\Psi}_c=\boldsymbol{\Psi}+\boldsymbol{G}\hat{\boldsymbol{F}}$ is closed loop matrix and $\hat{\boldsymbol{F}}$ is feedback gain of system (6.1).

According to Lyapunov's stability theorem[10], the following condition should be satisfied for stability of system (6.1) in spite of the variation of parameters and disturbance.

$$\Delta V[\hat{X}(k)]<0 \quad (6.4)$$

From (6.2) and (6.3), the first difference of Lyapunov function can be chosen as

$$\Delta V[\hat{X}(k)]=\hat{X}^T(k)[\boldsymbol{\Psi}_c^T\boldsymbol{P}_w\boldsymbol{\Psi}_c-\boldsymbol{P}_w]\hat{X}(k) \quad (6.5)$$

From (6.4) and (6.5), it can be written as follows:

$$\boldsymbol{\Psi}_c^T\boldsymbol{P}_w\boldsymbol{\Psi}_c-\boldsymbol{P}_w=-\boldsymbol{Q}_w \quad (6.6)$$

where $\boldsymbol{Q}_w$ is a positive definite matrix.

If we chose $\boldsymbol{P}_w=\boldsymbol{P}$ and $\hat{\boldsymbol{F}}=\boldsymbol{F}$, then the Riccati equation (5.14) can be written as follows:

$$\boldsymbol{\Psi}_c^T\boldsymbol{P}\boldsymbol{\Psi}_c-\boldsymbol{P}=-\boldsymbol{F}^T\boldsymbol{R}\boldsymbol{F}-\boldsymbol{Q} \quad (6.7)$$

Therefore, the expression for $\boldsymbol{Q}_w$ can be written as follows:

$$\boldsymbol{Q}_w=\boldsymbol{F}^T\boldsymbol{R}\boldsymbol{F}+\boldsymbol{Q} \quad (6.8)$$

Since $\boldsymbol{Q}$ is positive semi-definite and $\boldsymbol{R}$ is positive definite, according to expression (6.8) $\boldsymbol{Q}_w$ becomes positive definite matrix. Therefore, it can be concluded that the controller system with the proposed observer is stable.

7. Simulation Results

In order to verify the performance of the proposed MIMO optimal regulator and observer for bilinear system of IM drive as shown in Fig. 3, simulations were carried out using Matlab/Simulink[11]. The ratings and parameters of the IM model are listed in **Table 2**.

The value of sampling period is chosen 75 μsec. We consider the core loss power P_i is equal to 27 watts at rated condition. Using this value, the core loss resistance is calculated according to equation (4.4). The core loss resistance was considered constant for simulation study. The values of g_1, g_2, g_3 and g_4 used are given in section 3.

To verify the validity of proposed minimal order observer, the initial estimation values of rotor current and rotor flux components are chosen zero. **Fig. 4** shows the response of proposed observer to estimate rotor current and rotor flux components. The actual values given in Fig. 4 at $t=0$ sec are calculated using (2.1)～(2.6) and (4.1)～(4.3) at steady state for 800 r/min and full load torque. From Fig. 4, it is clear that the estimation values of rotor current and rotor flux components reach the actual values from zero values.

At time $t=0.2$, the actual values of rotor q-axis current and rotor d-axis flux are changed. In this case it is assumed that the IM runs into field-weakening region by increasing slip frequency 4.713 rad/sec to 9.0 rad/sec. By changing the slip frequency with 800 r/min, full load torque, zero rotor d-axis current and zero rotor q-axis flux, the actual values of rotor q-axis flux and rotor d-axis flux are changed and the others values are calculated using (2.1)～(2.6) and (4.1)～(4.3) at steady state. From Fig. 4, it is seen that the estimation values of rotor current and rotor flux components reached the actual values after changing the rotor q-axis current and rotor d-axis flux at $t=0.2$ sec. These situations clarified that the proposed observer can estimate accurately rotor current and rotor flux components of IM drive taking core loss into account.

The most important control object of FOC of IM drive is the constant linkage rotor flux that means Φ_{2d}=constant with $\Phi_{2q}=0$. The core loss resistance R_m is calculated using (4.4) by taking the core loss power 27 watts. To verify the performance of proposed controller with observer system, the simulation work has been done where three cases are considered. The cases are as follows:

(i) Speed is changed from 800 r/min to 900 r/min at 0.1 sec with 50% load torque. Since the core loss resistance changes with frequency, the core is also changed from 268.61 Ω to 338.57 Ω.

(ii) The load torque is changed from 50% to 100% at 0.6 sec with 900 r/min. And the actual core loss resistance is changed from 338.57 Ω to 349.76 Ω.

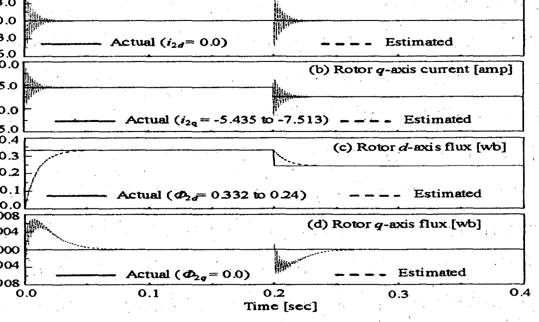

Fig. 4 Rotor current and flux components estimate using the proposed minimal order observer system

Table 2 Ratings and Parameters of Induction Motor

1.1 kW, 200/$\sqrt{3}$ V/phase, 6 Poles, 50 Hz
$r_1=0.2842\ \Omega$, $r_2=0.2878\ \Omega$, $R_m=404.397\ \Omega$, $L_s=28.3$ mH, slip=0.03, $L_r=28.8$ mH, $M=26.8$ mH, $J=0.0179$ kg-m², $D=0$

But it is assumed that the changing value of core loss resistance is unknown. Therefore, we used the previous value of core loss resistance i.e. 338.57 Ω in observer and controller.

(iii) Stator resistance r_1 and rotor resistance r_2 are increased by 30% of each rated value at 1.1 sec and it is assumed that the changing values of resistances are unknown. The rated values of resistances are used to calculate control input and estimate the state quantities. The actual value indicates that these quantities are real value of simulated IM drive and used value indicates that these quantities are used in controller and observer system for estimation.

In this simulation work, the values of g_1 and g_2 are calculated using (3.14) where $g_3=0.00001$ and $g_4=0.0$. The steady state values of rotor current and rotor flux components at 800 r/min and 50% load torque are considered as the initial estimation values and the estimation values of rotor current and rotor flux components are fed back to calculate the state of augmented system (6.1).

Fig. 5 shows the response for step change of desired rotor speed from 800 r/min to 900 r/min with 50% load at $t=0.1$ sec where $R_{11}=150.00$, $R_{22}=10.0$, $R_{33}=300.0$, $Q_{11}=0.05$, $Q_{22}=10^5$, $Q_{33}=5\times10^5$ and $Q_{44}=2.0\times10^6$ are used for the proposed control system. The values of weighting factor matrices $\boldsymbol{R}$ and $\boldsymbol{Q}$ are determined by trial and error method to obtain good performance. These figures also illustrate the comparison between the proposed control system and IFOC in terms of magnetizing current according to equations (4.5)$\sim$(4.7).

The actual speed follows the reference speed without any overshoot as shown in Fig. 5a after changing the step change of reference speed from 800 r/min to 900 r/min with 50% load torque at $t=0.1$ sec. For changing the reference speed, the actual rotor fluxes and the generating torque are suddenly changed in transient condition but after some time the actual rotor fluxes return to the reference values shown in Figs. 5b and 5c and the generating torque returns to the load torque value shown in Fig. 5d. At $t=0.6$ sec the disturbance load torque is changed from 50% to 100% with 900 r/min. The generating torque follows the disturbance load torque shown in Fig. 5d.

The generating torque can compensate the effects of disturbance load torque. In this situation the actual speed and the actual rotor fluxes are changed at transient condition and return to the reference values after some time as shown in Fig. 5a$\sim$5c. As the actual rotor fluxes return to the reference values, the controller satisfied the requirement of FOC at steady state condition. From Fig. 5, it can be concluded that the speed control with the proposed observer system for FOC of IM drive is achieved under the variation of load torque. It is clear from Fig. 5 that the transient responses of proposed control system are better than those of the conventional IFOC.

To verify the robustness of proposed controller, stator resistance and rotor resistance were increased 30% of each rated value at 1.1 sec. Fig. 5 shows the responses of desired speed and rotor fluxes for changing the resistances. When the resistances are changed, the actual speed and the

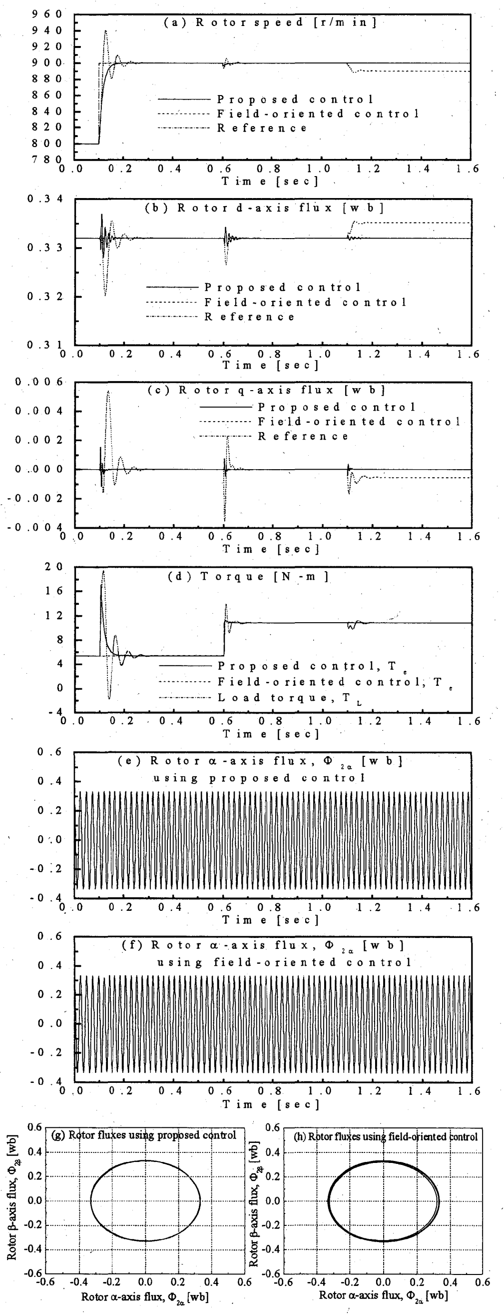

Fig. 5 Transient responses for step change of speed and disturbance load torque with the variation of resistance

actual rotor fluxes are changed at transient condition. The change of actual speed and rotor fluxes are almost negligible. Since the conventional FOC is sensitive to the parameter variations, the actual speed and rotor fluxes cannot follow the reference speed and rotor fluxes after changing the resistances. From Fig. 5, it can also be concluded that the speed control with the proposed observer system for FOC of IM drive is stable under the variation of parameters.

Fig. 5e and 5f show the rotor α-axis flux in a stationary reference frame using the proposed control and field-oriented control respectively. Fig. 5g and 5h demonstrate the resultant trajectory of rotor fluxes in a stationary frame using the proposed control and field-oriented control, respectively. As demonstrated by the simulation results shown in these figures, the flux responses using the proposed controller are better than those responses using the field-oriented control. Using the proposed control system, the rotor flux can be followed constant rotor flux magnitude as shown in Fig. 5g. Moreover, the flux phase error is almost negligible since the resultant magnitude of rotor flux is constant using the proposed control. In the case field-oriented control, the rotor flux cannot always be followed constant flux magnitude as shown in Fig. 5h. So, some flux phase error is aroused in the conventional field-oriented control.

Fig. 6a～6d shows the estimation errors for whole simulation range from 0.0 sec to 1.6 sec for both the proposed control and field-oriented control systems. These figures demonstrated the estimation error in a d-q axis synchronously rotating reference frame. It is seen from these figures that the estimation errors are zero in steady state as shown in these figures from 0.0 sec to 1.1 sec where the parameters are known. And some error is occurred due to the change of parameters as shown in these figures from 1.1 sec to 1.6 sec. Fig. 6e represents the actual values and estimation values of rotor flux components in case of the proposed control system in a stationary frame from 1.4 sec to 1.48 sec. Fig. 6f represents the actual values and estimation values of rotor flux components in case of conventional field-oriented control system in a stationary frame from 1.4 sec to 1.48 sec. From these figures, it is clear that there have some error of rotor flux components due to the change of stator, rotor and core loss resistances. Fig. 6 ensures that the proposed observer can estimate rotor current and flux components accurately where the values of parameters are known and some negligible error is aroused where the values of changing parameters are unknown.

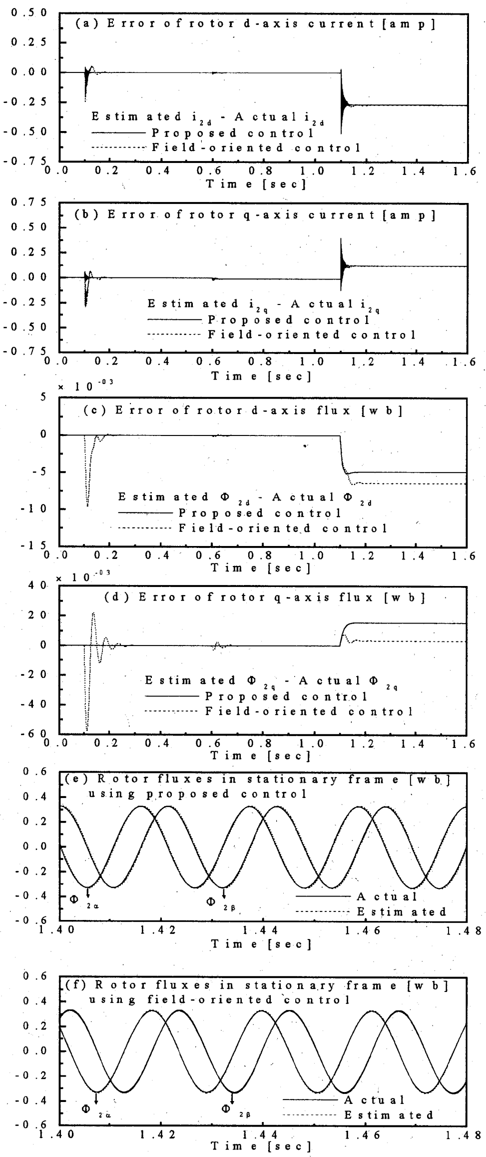

Fig. 6 Estimation error for changing speed, load torque and resistances

These situations can be considered as the high performance control for FOC of IM since the torque is unchanged (i.e. T_e is constant which is shown in Fig. 5d) when the speed is changed and the speed is unchanged (i.e. ω_m is constant which is shown in Fig. 5a) when the load torque is changed. Also, the proposed controller is stable with the proposed observer system against the variation of parameters. In both cases, the rotor d-axis flux is constant and the rotor q-axis flux follows zero, which are the main constraints of FOC. It is seen that the high performance control for FOC of IM can be realized by three-inputs in order to take core loss into account.

8. Conclusion

A high performance control for FOC of IM with consideration of core loss is proposed which is based on the

MIMO optimal regulator control theory. To eliminate the transient and the steady state error due to model uncertainty, the augmented system is incorporated. As the poles of designed controller are located in unit circle, the controller is stable for variation of disturbance load torque and parameters. The performances of proposed controller are compared with those of the conventional indirect FOC. The simulation results confirm that the proposed control system is better than the conventional FOC.

The rotor current and rotor flux is estimated with the minimal order observer for bilinear system. The proposed bilinear observer is stable which is confirmed using Lyapunov's stability theorem. The estimation performance of proposed bilinear observer is excellent. At last we can conclude that, by virtue of both the MIMO optimal regulator control and minimal order observer for bilinear system, the high performance for FOC of IM is very satisfactory not only at steady state condition but also in the transient condition.

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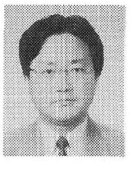

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