Indirect Field Oriented Control for High Performance Induction Motor Drives Using Space Vector Modulation with consideration of Core Loss

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Abstract- Indirect Field Oriented Control (IFOC) is known to produce high performance in induction Motor (IM) drives by decoupling rotor flux and torque producing current components of stator current. The decoupling control between the rotor flux and the torque is no longer achieved in terms of stator current components considering core loss into account. To overcome the perturbation effects of core loss, this paper introduces a decoupled rotor flux and torque control based on the magnetizing current components by incorporating PI control system and using space vector modulation (SVM) technique. It is able to perform high performance control in IM drives using SVM technique. In order to perform the control strategy, the rotor flux and magnetizing current estimator is proposed. The design of the proposed observer is based on Lyapunov's stability method whose estimation error converges to zero exponentially irrespective of the input. Simulation results are presented to show the validity of the proposed control as well as rotor flux and magnetizing current estimation for IFOC of IM drives. A good performance is obtained by means of PI controller and SVM technique, and the rotor flux and the magnetizing current are obtained properly by the proposed observer.

I. INTRODUCTION

The vector control of ac drives [1] has been widely used in high performance control system. Indirect field oriented control (IFOC) is one of the most effective vector control of induction motor due to the simplicity of designing and construction. In order to obtain the high performance of torque and speed of an IM drive, the rotor flux and torque generating current components of stator current must be decoupled suitably respective to the rotor flux vector like separately excited dc motor [2]. However, decoupled rotor flux and torque control in terms of stator current is not met in practice, where core loss exists [3]-[5]. The effects of core loss due to eddy current in IM drives perturb the performance of decoupled control system in terms of slip angular frequency.

To overcome the perturbation of core loss, this paper develops a decoupled rotor flux and torque control based on the magnetizing current components. However, this developed control system is valid only for steady state condition. By incorporating PI control [6]–[7] system and using space vector modulation technique [8]-[9], the decoupled control system is used to perform in both transient and steady state conditions.

In order to perform IFOC of IM drives using SVM shown in Fig.1, it is necessary to measure primary angular frequency, the stator voltages, the stator currents, the magnetizing currents, the rotor fluxes and the rotor speed. However, among those the rotor fluxes and magnetizing currents are difficult to measure. Therefore, in this paper, we proposed an observer system to estimate the rotor fluxes and magnetizing currents. The design of the observer is based on the solution of simultaneous Lyapunov equations [10]. It is proved that the estimation error converges to zero exponentially irrespective of the input, although the estimation error itself depends on the input.

The validity of IFOC using SVM and the proposed observer scheme are verified by simulation. Simulation results show that PI control using SVM technique is capable of realizing IFOC in spite of the degradation of core loss and the proposed estimation method is capable of estimating the rotor current and rotor flux.

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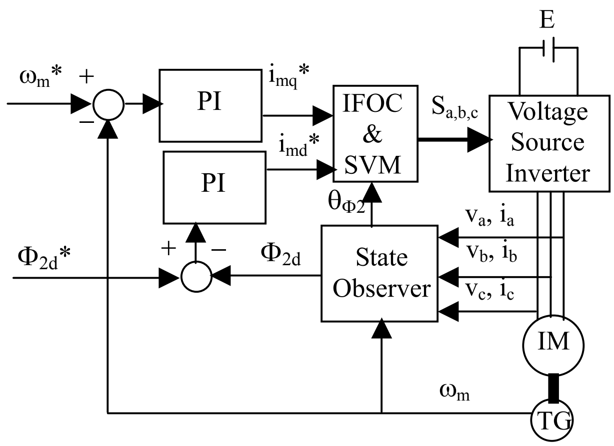

Fig. 1. The IFOC of IM drives using SVM.

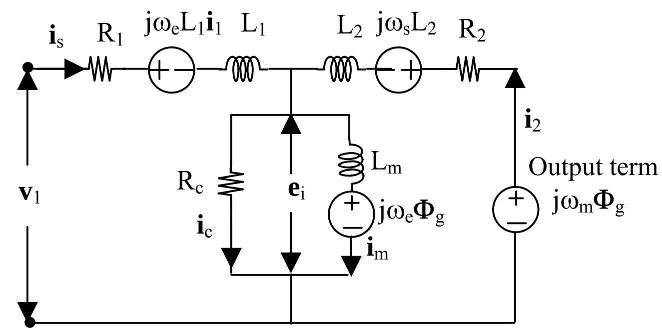

Fig. 2. Space vector equivalent circuit of an IM taking core loss into account.

II. IFOC OF IM DRIVES

In the d-q axis equivalent circuit of field oriented control induction motor considering core loss due to eddy current, a resistance R_c is connected in parallel across the internal induced voltage branch. The d-q axis equivalent circuit of induction motor in a synchronously rotating reference frame is shown in Fig. 2.

From Fig.2, the rotor flux, torque, and slip angular velocity in steady state with the constraints of IFOC can be written as the following:

$$\Phi_{2d} = L_m i_{md} \tag{2.1}$$

$$T_e = p(L_m/L_2) i_{mq} \Phi_{2d} \tag{2.2}$$

$$R_c = (\omega_e^2/P_i)(L_m^2 i_{mq}^2 + \Phi_{2d}^2) \tag{2.3}$$

$$\omega_s = (R_2 L_m/L_2)(i_{mq}/\Phi_{2d}) \tag{2.4}$$

From (2.3), it is clear that the core loss resistance depends on the frequency and flux level where P_i is the measured core loss power in different operating conditions. According to (2.1), (2.2) and (2.4), decoupling of torque and flux exists in terms of magnetizing currents in steady state condition.

III. PI CONTROLLER DESIGN

The control of torque and flux can be established in steady state condition by using (2.1)-(2.4). PI controllers are dynamic controllers, very useful in practice and superior to static controllers, since they offer great simplicity and flexibility in satisfying the closed-loop system's specifications. Fig. 1 shows a block diagram illustrating the proposed IFOC scheme based on PI controller.

The torque control can be achieved from speed error, since the developed electromagnetic torque affects the speed dynamic. Applying PI controller into the error signal between reference and measured or calculated speed and flux, it is possible to find out the desired magnetizing component i_{md}* and i_{mq}*. Therefore, the equations for PI controller can be written as follows:

$$i_{md}^* = \{K_{P\Phi} + (K_{I\Phi}/s)\}(\Phi_{2d}^* - \Phi_{2d}) \tag{3.1}$$

$$i_{mq}^* = \{K_{P\omega} + (K_{I\omega}/s)\}(\omega_m^* - \omega_m) \tag{3.2}$$

where "*" denotes reference or desired value.

The required inputs, stator voltages and primary angular frequency can be calculated from the outputs i_{md}* and i_{mq}* of PI controller by using (3.3)-(3.6) and (2.4).

$$i_{1d}^* = i_{md}^* - (L_m/R_c)\omega_e i_{mq}^* \tag{3.3}$$

$$i_{1q}^* = (L_r/L_2) i_{mq}^* + (L_m/R_c)\omega_e i_{md}^* \tag{3.4}$$

$$v_{1d}^* = (R_1 + R_c) i_{1d}^* - R_c i_{md}^* - \omega_e L_1 i_{1q}^* \tag{3.5}$$

$$v_{1q}^* = (R_1 + R_c) i_{1q}^* - (R_c L_r/L_2) i_{mq}^* + \omega_e L_1 i_{1d}^* \tag{3.6}$$

For speed control, it is needed to modify $K_{P\omega}$ and $K_{I\omega}$ constants.

IV. States Estimation

An IM drive can be described by following state equations in a synchronously rotating reference frame, which are derived from Fig. 2.

$$\begin{aligned} \dot{\mathbf{x}} &= \mathbf{A}\mathbf{x} + \omega_e \mathbf{A}^i \mathbf{x} + \mathbf{B}\mathbf{v}_1 \\ \mathbf{i}_1 &= \mathbf{C}\mathbf{x} \end{aligned} \tag{4.1}$$

where $\mathbf{x} = \begin{bmatrix} \mathbf{i}_1 & \mathbf{i}_m & \mathbf{\Phi}_2 \end{bmatrix}^T$, $\mathbf{i}_1 = \begin{bmatrix} i_{1d}, i_{1q} \end{bmatrix}^T$,

$\mathbf{i}_m = \begin{bmatrix} i_{md}, i_{mq} \end{bmatrix}^T$, $\mathbf{\Phi}_2 = \begin{bmatrix} \Phi_{2d}, \Phi_{2q} \end{bmatrix}^T$,

$$\mathbf{A} = \begin{bmatrix} \mathbf{A}_{11} & \mathbf{A}_{12} & \mathbf{A}_{13} \\ \mathbf{A}_{21} & \mathbf{A}_{22} & \mathbf{A}_{23} \\ 0 & \mathbf{A}_{32} & \mathbf{A}_{33} \end{bmatrix} \quad \mathbf{A}^i = \begin{bmatrix} \mathbf{A}^i{}_{11} & 0 & 0 \\ 0 & \mathbf{A}^i{}_{22} & 0 \\ 0 & 0 & \mathbf{A}^i{}_{33} \end{bmatrix} \quad \mathbf{B} = \begin{bmatrix} \mathbf{B}_1 \\ 0 \\ 0 \end{bmatrix}$$

$\mathbf{A}_{11} = -\{(R_1 + R_c)/L_1\}\mathbf{I} = a_{r11}\mathbf{I}$, $\mathbf{A}_{12} = (R_c L_r / L_1 L_2)\mathbf{I} = a_{r12}\mathbf{I}$,

$\mathbf{A}_{13} = -(R_c / L_1 L_2)\mathbf{I} = a_{r13}\mathbf{I}$, $\mathbf{A}_{21} = (R_c / L_m)\mathbf{I} = a_{r21}\mathbf{I}$,

$\mathbf{A}_{22} = -(R_c L_r / L_m L_2)\mathbf{I} = a_{r22}\mathbf{I}$, $\mathbf{A}_{23} = (R_c / L_m L_2)\mathbf{I} = a_{r23}\mathbf{I}$,

$\mathbf{A}_{32} = (R_2 L_m / L_2)\mathbf{I} = a_{r32}\mathbf{I}$, $\mathbf{A}^i{}_{11} = \mathbf{A}^i{}_{22} = \mathbf{A}^i{}_{33} = -\mathbf{J}$,

$\mathbf{A}_{33} = -(R_2 / L_2)\mathbf{I} + \omega_m \mathbf{J} = a_{r33}\mathbf{I} + a_{i33}\mathbf{J}$, $\mathbf{B}_1 = (1/L_1)\mathbf{I} = b_1\mathbf{I}$,

$L_r = L_2 + L_m$, $\mathbf{C} = \begin{bmatrix} \mathbf{I} & \mathbf{0} & \mathbf{0} \end{bmatrix}$, $\mathbf{I} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$, $\mathbf{J} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$

The full order state observer can be written by the following equation.

$$\dot{\hat{\mathbf{x}}} = \mathbf{A}\hat{\mathbf{x}} + \omega_e \mathbf{A}^i \hat{\mathbf{x}} + \mathbf{B}\mathbf{v}_1 + \mathbf{G}(\hat{\mathbf{i}}_1 - \mathbf{i}_1) \tag{4.2}$$

where ^ indicates estimated value and **G** is observer gain.

The poles of the observer are made proportional to those of the IM. Therefore, the values of gain matrix can be calculated from (4.1) and (4.2) as follows:

$$\mathbf{G} = \begin{bmatrix} g_1\mathbf{I} + g_2\mathbf{J} & g_3\mathbf{I} + g_4\mathbf{J} & g_5\mathbf{I} + g_6\mathbf{J} \end{bmatrix}^T$$

$g_1 = (k-1)(a_{r11} + a_{r22} + a_{r33})$ $g_2 = (k-1)(a_{i33} - 3\omega_e)$

$$g_3 = (k-1)(1/a_{r12})[a_{r22}^2 + a_{r33}a_{r22} + a_{r33}^2 + a_{r12}a_{r21} - (a_{r33}^2 / a_{i33})(\omega_e - a_{i33})]$$

$$g_4 = (k-1)(1/a_{r12})[\omega_e a_{r22} + 2\omega_e a_{r33} - (\omega_e^2 / a_{i33})(2a_{r22} + 3a_{r33})]$$

$$g_5 = (k-1)\{a_{r33}^2/(a_{r13}a_{i33})\}(\omega_e - a_{i33})$$

$$g_6 = (k-1)(1/a_{r13})[a_{i33}a_{r33} - \omega_e(2a_{r22} + 4a_{r33}) + (\omega_e^2 / a_{i33})(2a_{r22} + 3a_{r33})]$$

where k (k>0) is the proportional constant.

Taking k = 1.0, the error between estimated and actual values can be written as follows:

$$\dot{\mathbf{e}} = [\mathbf{A} + \omega_e \mathbf{A}^i]\mathbf{e} \tag{4.3}$$

Now let us consider the Lyapunov function as

$$\mathbf{V} = \mathbf{e}^T \mathbf{P} \mathbf{e} \tag{4.4}$$

The time derivative of Lyapunov function by using (4.3) and (4.4) can be written as follows:

$$\dot{\mathbf{V}} = \mathbf{e}^T \{\mathbf{P}[\mathbf{A} + \omega_e \mathbf{A}^i] + [\mathbf{A} + \omega_e \mathbf{A}^i]^T \mathbf{P}\}\mathbf{e} \tag{4.5}$$

All eigenvalues of $\mathbf{A} + \omega_e \mathbf{A}^i$ have negative real parts according to the inherent characteristic of IM drive, which can be confirmed by Fig. 3. Hence, the observer system is stable at constant speed. The ratings and parameters of IM are given in TABLE I as shown in next page.

By taking $\mathbf{P} = \mathbf{I}_6$, it can be written as following:

$$\mathbf{P}\mathbf{A}^i + \mathbf{A}^{iT}\mathbf{P} = 0 \quad \text{and} \quad \mathbf{P}\mathbf{A} + \mathbf{A}^T\mathbf{P} = -\mathbf{Q}$$

where **Q** is a positive symmetric matrix.

TABLE I

Ratings and Parameters of Induction Motor

1.1 Kw, 200/√3 V/phase, 6 Poles,50 Hz
R_1 = 0.2842 Ω, R_2 = 0.2878 Ω, R_c=404.66 Ω, slip=0.03, L_1 = 28.3 mH, L_2 = 28.8 mH, L_m = 26.8 mH, J = 0.0179 Kg-m^2, D = 0.0

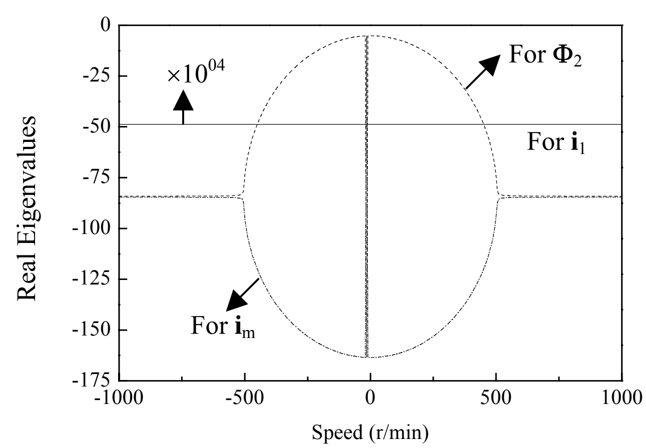

Fig. 3. Real eigenvalues vs speed at full load torque.

Hence, equation (4.5) can be written as follows:

$$\dot{\mathbf{V}} = -\mathbf{e}^T \mathbf{Q} \mathbf{e} \quad (4.6)$$

Since **P** and **Q** are positive definite, the error becomes

$$\lim_{t \to \infty} \mathbf{e}(t) = 0 \quad (4.7)$$

by the well-known stability theorem of Lyapunov. Therefore, the observer system is stable in the case of variable speed.

V. SVM Technique in PWM Inverter

SVM technique is more preferable scheme for PWM voltage source inverter since it gives a large linear control range, less harmonic distortion and fast transient response. The SVM units in Fig. 1 produces inverter control signal. It receives the reference voltages v^*_{1d} (3.5) and v^*_{1q} (3.6) in a synchronously rotating reference frame. These voltages are converted into a stator reference frame such as $v^*_{1\alpha}$ and $v^*_{1\beta}$. The SVM principle is based on the switching between two adjacent active vectors and a zero vectors during one switching period. The six active-states are forming a regular hexagon and dividing it into six equal sectors denoted as **I**, **II**, **III**, **IV**, **V**, **VI** shown in Fig. 4. The reference voltage vector v_{1s}^* defined by its magnitude and V_{1s} (5.1) and angle θ (5.2) in a stator reference frame can be produced by adding two adjacent active vectors $\mathbf{U}_a$ (U_a, θ_a) and $\mathbf{U}_b$ (U_b, θ_b) and, if necessary, a zero vector **U**0(000) or **U**7(111).

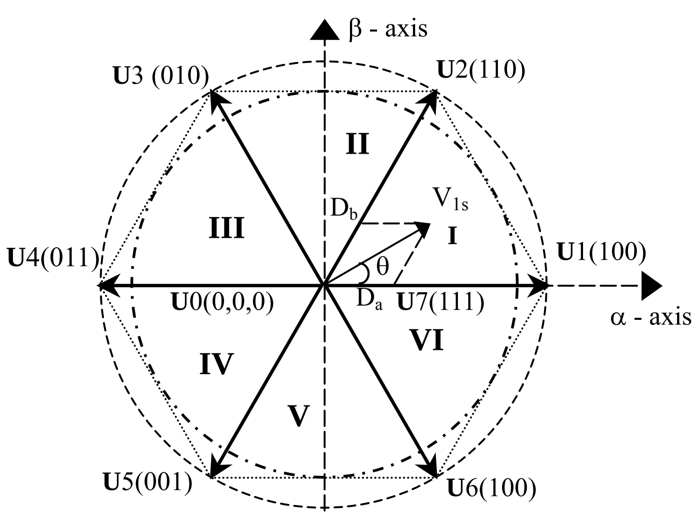

Fig. 4. Representation of the inverters states in the stationary reference frame.

$$V_{1s} = \sqrt{v^{*2}_{1\alpha} + v^{*2}_{1\beta}} \quad (5.1)$$

$$\theta = \tan^{-1}(v^*_{1\alpha} / v^*_{1\beta}) \quad (5.2)$$

The duty cycles Da and D_b for each active vector are the solutions of the complex equation

$$\begin{aligned} V_{1s}(\cos\theta + j\sin\theta) &= D_a U_a(\cos\theta_a + j\sin\theta_a) \\ &\quad + D_b U_b(\cos\theta_b + j\sin\theta_b) \end{aligned} \quad (5.3)$$

$$D_a = \frac{\sqrt{3}V_{1s}}{E}\sin\theta \quad (5.4)$$

$$D_b = \frac{3V_{1s}}{2E}\cos\theta - \frac{1}{2}Da \quad (5.5)$$

where E is the dc link voltage.

The duty cycle for the zero vector is the remaining time inside the switching period Ts

$$D_0 = 1 - D_a - D_b \quad (5.6)$$

The vector sequence and timing during one switching period is given in TABLE II. Fig. 5 shows the command signals for the inverter when vector **U**1(100) and **U**2(110) and zero vectors **U**0(000) and **U**7(111) are applied. The request of minimal number of commutations per cycle is met if the $\mathbf{U}_a$ and $\mathbf{U}_b$ vectors are selected regarding to TABLE III.

TABLE II
Time Duration for Selected Vectors

U0	$\mathbf{U}_a$	$\mathbf{U}_b$	**U**7	$\mathbf{U}_b$	$\mathbf{U}_a$	**U**0
$D_0/4$	$D_a/2$	$D_b/2$	$D_0/2$	$D_b/2$	$D_a/2$	$D_0/4$

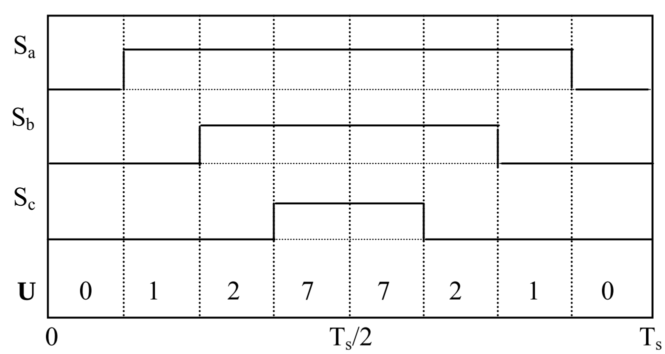

Fig. 5. The SVM voltage vector timing.

TABLE III
Vector Selections in Different Sections

Section	θ	$\mathbf{U}_a$	$\mathbf{U}_b$
I	0 to π/3	**U**1	**U**2
II	π/3 to 2π/3	**U**3	**U**2
III	2π/3 to π	**U**3	**U**4
IV	π to 4π/3	**U**5	**U**4
V	4π/3 to 5π/3	**U**5	**U**6
VI	5π/3 to 2π	**U**1	**U**6

VI. Simulation Results

In order to verify the performance of the proposed IFOC of IM drive and observer system, simulations were carried out where (3.1) and (3.2) are changed into discrete-time form. The ratings and parameters of the induction motor model are listed in TABLE I. The values of sampling period 75 μsec and k=1.0 are used for simulation.

According to Fig. 1, the transient responses of the proposed IFOC using SVM technique, where the estimated value from observer is fed back are shown in Fig. 6. The dc link voltage is taken 200 Volts. Fig. 6 shows the transient response for step change of speed from 500 r/min to 900 r/min at 50% load torque. The speed controller's parameters used for simulation are the following.

(i) The PI controller for the flux in Fig.1 uses the values $K_{P\Phi}$= 0.5 and $K_{I\Phi}$= 0.05 and

(ii) The PI controller for the speed in Fig.1 uses the values $K_{P\omega}$= 0.02 and $K_{I\omega}$= 10^{-6}.

From Fig. 6 (a), it is clear that the actual speed follows the desired speed by using the proposed controller and observer system incorporating SVM technique. It is observed from Fig. 6 that the torque, d-q axis currents are constant at steady state condition with changes of speed. This situation can be considered as an indirect field oriented control where rotor flux and speed PI controllers act as a compensator for transient response.

Fig. 7 shows the transient response for change of load torque from 50% to 100% at 900 r/min. The torque controller's parameters used for simulation are the following.

(i) The PI controller for the flux in Fig.1 uses the values $K_{P\Phi}$= 0.5 and $K_{I\Phi}$= 0.05 and

(ii) The PI controller for the torque in Fig.1 uses the values $K_{P\omega}$= 0.03 and $K_{I\omega}$= 0.0002.

Fig. 7(a) shows that the actual torque follows the desired torque. From Fig. 7 (b), it is seen that the q-axis magnetizing current changes for changing load torque. Hence, the torque can be controlled by the q-axis component of magnetizing current.

Figs. 6 and 7 show the comparing result of observer based and full state measurement based controller. The initial estimation values are taken zero. It is clear from the simulation results that the estimation values of states converge to the actual values of states within a short time. Therefore, the proposed controller can be estimate accurately

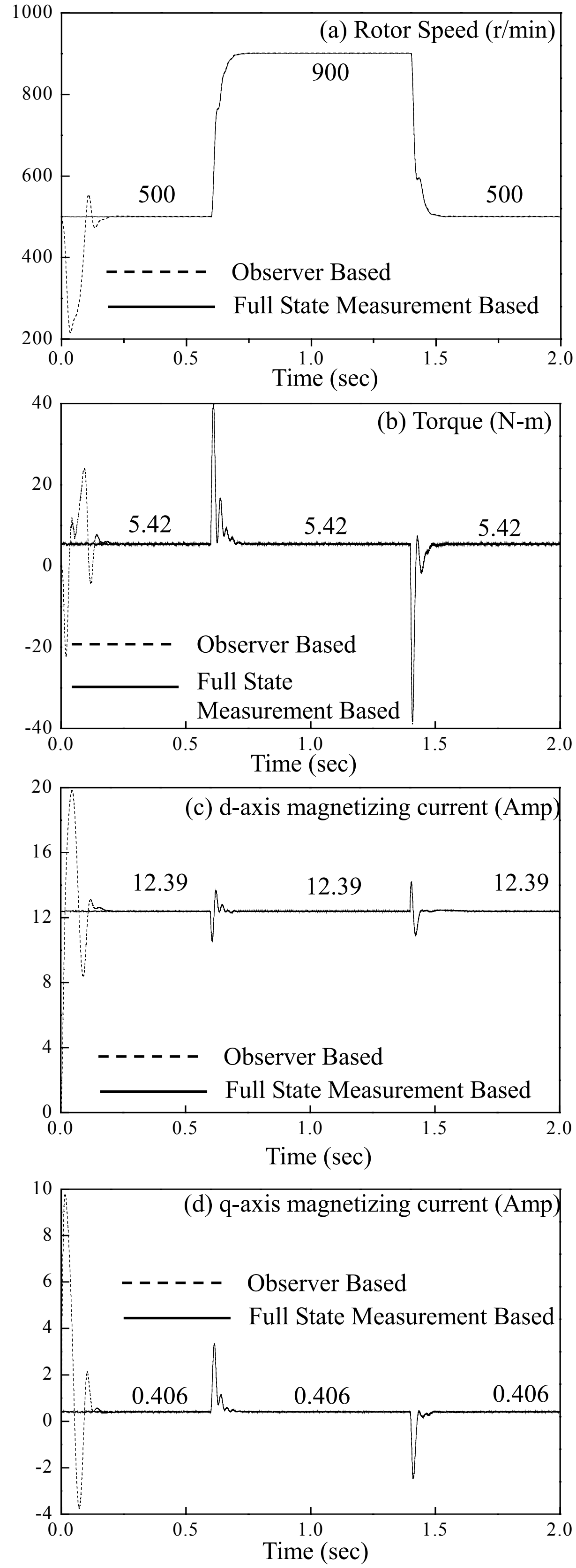

Fig. 6. Transient response for step change of speed from 500[r/min] to 900[r/min].at 50% load torque.

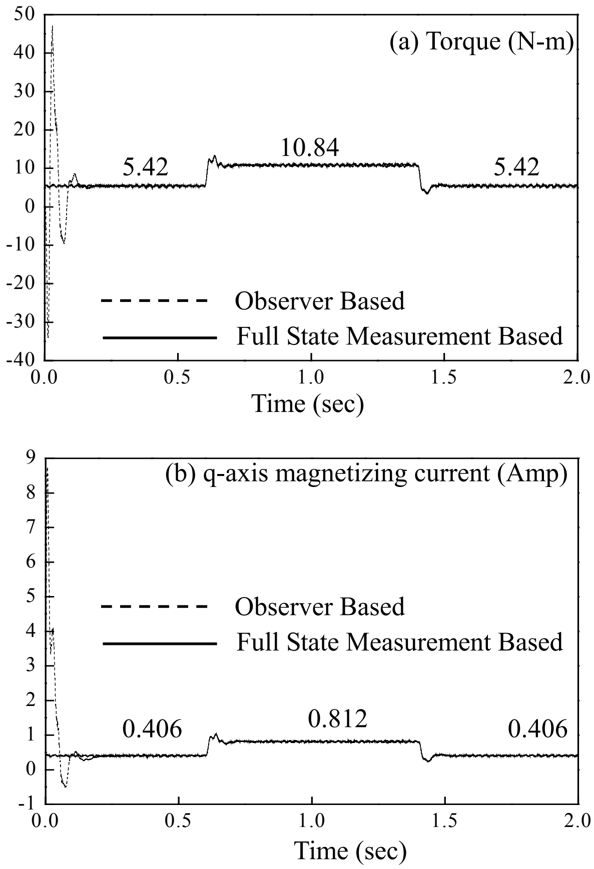

Fig. 7. Transient response for change of load torque from 50 [%] to 100 [%] at 900 r/min.

of state quantities of field oriented control induction motor drive. Also, it is clear from simulation results that the control of indirect field oriented control of induction motor drive can be achieved by using SVM technique of PWM inverter.

VI. Conclusion

This paper has presented the IFOC of IM drives using SVM technique with consideration of core loss. An observer for estimating the rotor flux and the magnetizing current in a synchronously rotating reference frame based on the Lyapunov stability theory has been presented. The verification of the proposed control is performed by simulation results. The dynamical requirement of IFOC is achieved by applying PI controller in transient and steady state conditions. The simulation results confirmed us that the proposed observer and control method using SVM technique can be used to indirect field-oriented induction motor control.

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