High Performance Control for SVPWM Inverter-Fed Induction Motor with Core Loss Consideration

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Abstract: The actual flux and torque of a field-orientated control (FOC) induction (IM) differ from the reference flux and torque due to the effects of core loss. Thus, it is desirable to design a controller to achieve desired speed and flux taking core loss into account. Moreover, it is also desirable to design an observer to estimate states of IM properly since the states of IM are increased with consideration of core loss. Therefore, in this paper speed and flux controller, and a full-order observer for space vector pulse width modulation (PWM) inverter-fed IM taking core loss into account are designed based on discrete-time proportional and integral (PI) controller. The effectiveness of proposed discrete-time PI controller and observer systems are verified by simulation. The simulation results confirmed that the proposed controller can properly track the reference speed and flux and the proposed observer can properly estimate the state quantities of IM taking core loss into account.

Keywords: Induction motor, Core loss, Speed and flux control, PI controller, Observer, pole placement technique, Space vector PWM inverter.

1 INTRODUCTION

The space vector modulation (SVM) technique of PWM inverter-fed IM has been applied to field-oriented control (FOC) (Dong-Choon 1994), (Lascu 2000) of IM because the switching loss and harmonic loss are minimized by this technique. The FOC was developed to obtain high performance by decoupling flux and torque generating stator current components by neglecting core loss (Uddin 2002). But the core loss exists to all ac electrical drives.

The core loss effects to ac electric machine have been investigated and an orthogonal model for all ac drive taking core loss into account has been proposed in (Boldea 1987). It has been clarified in (Boldea 1987), (Levi 1995), (Levi 1996) that the reference values of stator current components for decoupling flux and torque are differed from the actual stator current components values due to the effects of core loss of IM. The complexity is aroused to design FOC in terms of stator current components with compensation of core loss (Levi 1995), (Levi 1996). To overcome this complexity, PI controller based indirect FOC has been developed in terms of magnetizing current components where the magnetizing current components are decomposed to control flux and torque (Levi1995), (Levi 1996), (Mannan 2003). In (Levi1995), (Levi 1996), the conventional PI controller was applied to achieve desired speed where the gains of PI controller are chosen by trail and error method. To design discrete time PI controller, the controller gains are chosen by trail and error method for step change of speed and load torque to achieve desired speed and flux in (Mannan 2003). The steady state error and overshoot problems are aroused where PI controller gains selected by trail and error method. Moreover, the conventional PI controller is become unstable against the variation of load torque and parameters.

The proposed observers for IM drive such as (Verghese 1988), (Kubota 1993), (Maes 2000) were derived only to estimate rotor flux with neglected core loss. Due to the consideration of core loss, the states of IM are increased in the state space model of IM drive. It is desirable to design an observer to estimates states of IM taking core loss into account.

In this paper, discrete-time PI controller is proposed to achieve desired speed and rotor flux without overshoot and steady state error problems for indirect FOC of IM. The gains of PI rotor flux and speed controllers are chosen by placing pole into unit circle. Consequently, the designed controllers are become stable against the variations of load torque and parameters.

To perform the proposed speed and flux controller, a full order observer is also proposed. The gains of observer are chosen to obtain the poles of observer proportional to the poles of IM state space model (Kubota 1993), (Maes 2000). The simulation results are demonstrated to show the effectiveness of proposed controller and observer system for PWM inverter-fed IM where the PWM inverter are performed based on SVM technique. The desired speed and rotor d-axis flux can be achieved without overshoot and steady state error and against the variations of load torque and parameters. It is also comprehended from simulation results that the states can be estimated properly by using the proposed observer.

2 IM DYAMIC MODEL INCLUDING CORE LOSS

Fig.1 shows the *d* and *q*- axes equivalent circuits of an induction motor including stator eddy current core loss. The mathematical model of an induction motor according to Fig.1 can be written as the follows (Levi 1995):

$$v_{1d} = R_1 i_{1d} + d\Phi_{1d}/dt - \omega_e \Phi_{1q}\,;\; v_{1q} = R_1 i_{1q} + d\Phi_{1q}/dt + \omega_e \Phi_{1d} \tag{2.1}$$

$$0 = R_2 i_{2d} + d\Phi_{2d}/dt - \omega_s \Phi_{2q}\,;\; 0 = R_2 i_{2q} + d\Phi_{2q}/dt + \omega_s \Phi_{2d} \tag{2.2}$$

$$R_c i_{cd} = d\Phi_{md}/dt - \omega_e \Phi_{mq}\,;\; R_c i_{cq} = d\Phi_{mq}/dt + \omega_e \Phi_{md} \tag{2.3}$$

$$\Phi_{1d} = L_1 i_{1d} + \Phi_{md}\,;\; \Phi_{1q} = L_1 i_{1q} + \Phi_{mq} \tag{2.4}$$

$$\Phi_{2d} = L_2 i_{2d} + \Phi_{md}\,;\; \Phi_{2q} = L_2 i_{2q} + \Phi_{mq} \tag{2.5}$$

$$\Phi_{md} = L_m i_{md}\,;\; \Phi_{mq} = L_m i_{mq} \tag{2.6}$$

$$i_{cd} + i_{md} = i_{1d} + i_{2d}\,;\; i_{cq} + i_{mq} = i_{1q} + i_{2q} \tag{2.7}$$

$$T_e = p(L_m/L_2)(i_{mq}\Phi_{2d} - i_{md}\Phi_{2q}) \tag{2.8}$$

$$d\omega_m/dt = -(D/J)\,\omega_m + (p/J)(T_e - T_L) \tag{2.9}$$

Symbols v, i, Φ, ω_e, ω_s, ω_m, p, T_e, T_L, D and J indicate voltage, current, flux linkage, primary angular frequency, slip frequency, rotor speed, number of pole pair, developed torque, load torque, damping factor and total inertia respectively. L_1, L_2 and L_m denote stator leakage, rotor leakage and mutual inductances. R_1, R_2 and R_c denote stator, rotor and core loss resistances. Subscripts 1, 2, *m* and *c* are used to represent stator, rotor, magnetizing branch and core loss quantities respectively. Indices *d* and *q* stand for *d*-axis and *q*-axis components.

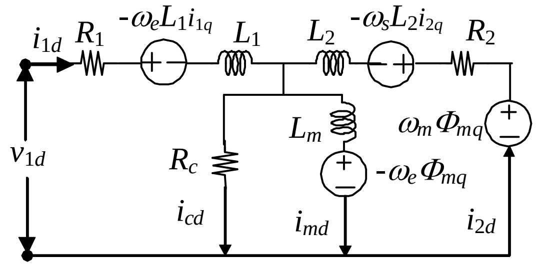
The values of the parameters and rating used in the simulation are given in **Table 1** (Paraskevopoulos 1996). The core loss resistance is constant since in this work only stator eddy current core loss only consider.

3 PI CONTROLLER DESIGN

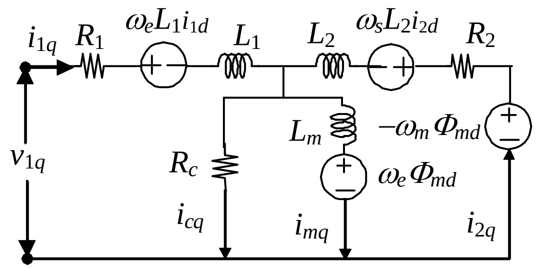
The constraints of indirect FOC can be defined as follows:

$$\Phi_{2d} = \text{Constant}, \Phi_{2q} = 0 \tag{3.1}$$

The rotor flux, torque and slip angular frequency at steady state condition with the constraint (3.1) of indirect FOC can be written from (2.1)-(2.8) as follows:



(a) *d*-axis



(b) *q*-axis

Fig. 1. Synchronous frame *d-q* axis model of the induction motor taking core loss into account.

Table 1 Ratings and Parameters of IM

2.24 kW, 14.96 Nm, 230 V rms (line), 4 pole, 1430 rpm
R_1 =0.55 Ω, R_2 =0.75 Ω, R_c =320 Ω, L_1 = 5.0 mH, L_2 = 5.0 mH, L_m = 63.0 mH

$$\Phi_{2d} = L_m i_{md}\,; \quad T_e = p(L_m/L_2) i_{mq} \Phi_{2d}\,; \quad \omega_s = (R_2 L_m/L_2)(i_{mq}/\Phi_{2d}) \tag{3.2}$$

It is clear from (3.2) that the rotor flux and torque can be generated in terms of *d*-axis and *q*-axis magnetizing current components respectively. The *d*-axis flux can be kept constant by regulating slip angular frequency.

According to (2.2), (2.9) and (3.1) – (3.2) the dynamics of speed and rotor flux can be expressed as follows:

$$d\omega_m / dt = -(D/J)\,\omega_m + (p/J)\left[(pL_m/L_2) i_{mq} \Phi_{2d} - T_L\right] \tag{3.3}$$

$$d\Phi_{2d} / dt = a_{r32} i_{md} + a_{r33} \Phi_{2d} \tag{3.4}$$

The discrete-time PI controller can be defined by the following expressions (3.3) and (3.4) to calculate the first difference of desired magnetizing current components.

$$\Delta i_{md}^{\ *}(k) = K_{I\Phi} e_{\Phi}(k) - K_{P\Phi} \Delta\Phi_{2d}(k); \quad \Delta i_{mq}^{\ *}(k) = K_{I\omega} e_{\omega}(k) - K_{P\omega} \Delta\omega_m(k) \tag{3.5}$$

where $e_{\Phi}(k) = \Phi_{2d}^{*}(k) - \Phi_{2d}(k)$, $e_{\omega}(k) = \omega_m^{*}(k) - \omega_m(k)$, $\Delta\Phi_{2d}(k) = \Phi_{2d}(k) - \Phi_{2d}(k-1)$, $\Delta\omega_m(k) = \omega_m(k) - \omega_m(k-1)$, $\Delta i_{md}^{*}(k) = i_{md}^{*}(k) - i_{md}^{*}(k-1)$, $\Delta i_{mq}^{*}(k) = i_{mq}^{*}(k) - i_{mq}^{*}(k-1)$.

k is sampling instant. $K_{P\omega}$ and $K_{I\omega}$ indicate the proportional and integral constant respectively for speed controller loop. Similarly, $K_{P\Phi}$ and $K_{I\Phi}$ indicate the proportional and integral constant respectively for flux controller loop. Superscript * indicates the desired or reference value.

Since the first difference $\Delta\Phi_{2d}(k)$ and $\Delta\omega_m(k)$ are used to design PI controller instead of $\Delta e_{\Phi}(k)$ and $\Delta e_{\omega}(k)$, the overshoot problem is not aroused. By using the relation (3.3) and (3.4) and choosing PI controller gains the poles of the dynamics (3.5) can be placed into unit circle for discrete time system. Hence, the PI controller gains are chosen as follows:

$$K_{P\omega} = (2\alpha_{\omega} - k_2)/k_1;\ K_{I\omega} = \alpha_{\omega}^2/k_1;\quad K_{P\Phi} = (2\alpha_{\Phi} + a_{r33})/a_{r32},\ K_{I\Phi} = \alpha_{\Phi}^2/a_{r32} \quad (3.6)$$

where $k_1 = p^2 L_m \Phi_{2d}^R / L_2 J$, $k_2 = D/J$ and Φ_{2d}^R is rated value of rotor d-axis flux. α_{ω} and α_{Φ} are the negative poles for speed and flux dynamics.

The poles are placed into unit circle for discrete time PI controller. The desired primary frequency can be obtained as follow:

$$\omega_e^*(k) = \omega_m(k) + (R_2/L_2)[i_{mq}^*(k)/i_{md}^*(k)] \quad (3.7)$$

Using (3.1)-(3.5) and (2.1)-(2.8), the desired stator current components can be given by the following:

$$i_{1d}^*(k) = i_{md}^*(k) - (L_m/R_c)\omega_e^*(k) i_{mq}^*(k) \quad (3.8)$$

$$i_{1q}^*(k) = (L_r/L_2) i_{mq}^*(k) + (L_m/R_c)\omega_e^*(k) i_{md}^*(k) \quad (3.9)$$

The desired stator voltage components can be obtained by:

$$v_{1d}^*(k+1) = v_{1d}^*(k) + \Delta v_{1d}^*(k);\quad v_{1q}^*(k+1) = v_{1q}^*(k) + \Delta v_{1q}^*(k) \quad (3.10)$$

The first difference values of stator voltages can be given by:

$$\Delta v_{1d}^*(k) = K_{Ii} e_{id}(k) - K_{Pi}\Delta i_{1d}(k);\quad \Delta v_{1q}^*(k) = K_{Ii} e_{iq}(k) - K_{Pi}\Delta i_{1q}(k) \quad (3.11)$$

where, $e_{id}(k) = i_{1d}^*(k) - i_{1d}(k)$, $e_{iq}(k) = i_{1q}^*(k) - i_{1q}(k)$, $\Delta i_{1d}(k) = i_{1d}(k) - i_{1d}(k-1)$, $\Delta i_{1q}(k) = i_{1q}(k) - i_{1q}(k-1)$. K_{Pi} and K_{Ii} indicate the proportional and integral constant respectively for current controller loop.

To achieve the desired speed and rotor d-axis flux the desired magnetizing components can be obtained using PI controller (3.5). Also, if any difference is occurred between the desired and actual value of stator current components the compensating stator voltage components can be obtained using PI controller (3.11). **Fig. 2** shows the block diagram of discrete-time PI speed controller to obtained desired q-axis magnetizing current components. The others PI controller structures are same as Fig. 2 where the gains, input and output are

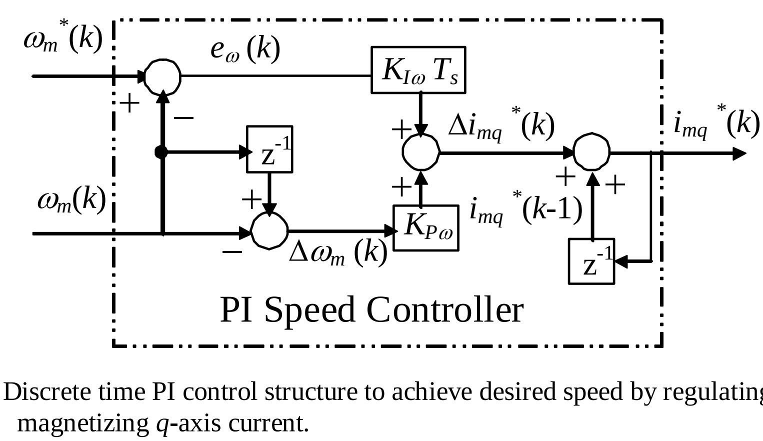


Fig. 2. Discrete time PI control structure to achieve desired speed by regulating magnetizing q-axis current.

different.

4 DESIGN OF FULL-OREDR OBSERVER

Using the measurable quantities of IM the observer can be designed. For constructing the observer system, using (2.1)-(2-9) the state equation of IM can be written as follows:

$$d\boldsymbol{x}/dt = \boldsymbol{A}\boldsymbol{x} + \omega_e \boldsymbol{A}^i \boldsymbol{x} + \boldsymbol{B}\boldsymbol{v}_1 \tag{4.1}$$

where, $\boldsymbol{A} = \begin{bmatrix} a_{r11}\boldsymbol{I} & a_{r12}\boldsymbol{I} & a_{r13}\boldsymbol{I} \\ a_{r21}\boldsymbol{I} & a_{r22}\boldsymbol{I} & a_{r23}\boldsymbol{I} \\ 0 & a_{r32}\boldsymbol{I} & a_{r33}\boldsymbol{I} + a_{i33}\boldsymbol{J} \end{bmatrix}$, $\boldsymbol{A}^i = -\begin{bmatrix} \boldsymbol{J} & 0 & 0 \\ 0 & \boldsymbol{J} & 0 \\ 0 & 0 & \boldsymbol{J} \end{bmatrix}$, $\boldsymbol{J} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$,

$\boldsymbol{I} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$

$\boldsymbol{x} = \left[i_{1d}, i_{1q}, i_{md}, i_{mq}, \Phi_{2d}, \Phi_{2q}\right]^T$, $\boldsymbol{v}_1 = \left[v_{1d}, v_{1q}\right]^T$, $\boldsymbol{B} = \left[b_1\boldsymbol{I}, 0, 0\right]^T$, $a_{r11} = -(R_1 + R_c)/L_1$,

$a_{r12} = R_c L_r / L_1 L_2$, $b_1 = 1/L_1$, $a_{r13} = -R_c / L_1 L_2$, $a_{r21} = R_c / L_m$, $a_{r22} = -R_c L_r / L_m L_2$,

$a_{r23} = R_c / L_m L_2$, $a_{r32} = R_2 L_m / L_2$, $a_{r33} = -R_2 / L_2$, $a_{i33} = \omega_m$, $L_r = L_2 + L_m$.

The full order state observer can be written by the following expression (4.2).

$$d\hat{\boldsymbol{x}}/dt = \boldsymbol{A}\hat{\boldsymbol{x}} + \omega_e \boldsymbol{A}^i \hat{\boldsymbol{x}} + \boldsymbol{B}\boldsymbol{v}_1 + \boldsymbol{G}(\hat{\boldsymbol{i}}_1 - \boldsymbol{i}_1) \tag{4.2}$$

where, $\boldsymbol{i}_1 = \left[i_{1d}, i_{1q}\right]^T$, $\hat{\boldsymbol{i}}_1 = \left[\hat{i}_{1d}, \hat{i}_{1q}\right]^T$, superscripts "^" and "$T$" indicate estimated value and transpose matrix respectively. $\boldsymbol{G}$ is observer gain.

The error between estimated (4.2) and actual values (4.1) can be written as follows:

$$d\boldsymbol{e}/dt = [\hat{\boldsymbol{A}} + \omega_e \boldsymbol{A}^i]\boldsymbol{e} \tag{4.3}$$

where, $\hat{\boldsymbol{A}} = \boldsymbol{A} + \left[\boldsymbol{G} \mid 0_{6\times4}\right]$, $\boldsymbol{e} = [e_{i1d}, e_{i1q}, e_{imd}, e_{imq}, e_{\Phi 2d}, e_{\Phi 2q}]^T$, $e_{i1d} = \hat{i}_{1d} - i_{1d}$,

$e_{i1q} = \hat{i}_{1q} - i_{1q}$, $e_{imd} = \hat{i}_{md} - i_{md}$, $e_{imq} = \hat{i}_{mq} - i_{mq}$, $e_{\Phi 2d} = \hat{\Phi}_{2d} - \Phi_{2d}$, $e_{\Phi 2q} = \hat{\Phi}_{2q} - \Phi_{2q}$.

The poles of the observer are placed such that the observer error decreases faster than the IM transients (Kubota 1993). The dynamics of IM state equation model (4.1) is stable because the real parts of poles of IM dynamics are negative. If the poles of the IM are given by P_{IM} then the observer poles P_O are selected as

$$P_O = cP_{IM} \tag{4.4}$$

where, c is proportional constant and c>1.0.

The pole placement can be achieved by defining the observer gain matrix $\boldsymbol{G}$ in a special form

$$\boldsymbol{G} = \left[g_1\boldsymbol{I} + g_2\boldsymbol{J} \quad g_3\boldsymbol{I} + g_4\boldsymbol{J} \quad g_5\boldsymbol{I} + g_6\boldsymbol{J}\right]^T \tag{4.5}$$

The expressions for g_1, g_2, g_3, g_4, g_5 and g_6 can be derived using (4.1)-(4.4). The derived expression for g_1 and g_2 can be given by

$$g_1 = (c-1)(a_{r11} + a_{r22} + a_{r33}); \quad g_2 = (c-1)(a_{i33} - 3\omega_e) \tag{4.6}$$

After calculating the value of g_1 and g_2 the other values of gain matrix $\boldsymbol{G}$ can be derived by

$$\begin{bmatrix} g_1 \\ g_2 \\ g_3 \\ g_4 \end{bmatrix} = \begin{bmatrix} a_{r12} & 0 & a_{r13} & 0 \\ 0 & a_{r12} & 0 & a_{r13} \\ m_{31} & m_{32} & m_{33} & m_{34} \\ m_{41} & m_{42} & m_{43} & m_{44} \end{bmatrix}^{-1} \begin{bmatrix} C_1 \\ C_2 \\ C_3 \\ C_4 \end{bmatrix} \tag{4.7}$$

where, $m_{31}=a_{r12}a_{r33}-a_{r13}a_{r32}$, $m_{32}=\omega_e a_{r12}-a_{r12}a_{i33}$, $m_{33}=a_{r13}a_{r22}-a_{r12}a_{r23}$, $m_{34}=a_{r13}\omega_e$, $m_{41}=a_{r12}a_{i33}-a_{r12}\omega_e$, $m_{42}=a_{r12}a_{r33}-a_{r13}a_{r32}$, $m_{43}=-a_{r13}\omega_e$, $m_{44}=a_{r13}a_{r22}-a_{r12}a_{r23}$,

$$C_1=-(k^2-1)(a_{r22}a_{r33}+a_{r11}a_{r22}+a_{r11}a_{r33}+2\omega_e a_{i33}-3\omega_e^2-a_{r23}a_{r32}-a_{r12}a_{r21})$$
$$+(a_{r22}+a_{r33})g_1+(2\omega_e-a_{i33})g_2$$

$$C_2=-(k^2-1)[a_{r22}a_{i33}+a_{r11}a_{i33}-(a_{r33}+a_{r11}+a_{r22}+a_{r11}+a_{r33}+a_{r22})\omega_e]$$
$$+(a_{i33}-2\omega_e)g_1+(a_{r33}+a_{r22})g_2$$

$$C_3=(k^3-1)[(a_{r11}+a_{r22}+a_{r33})\omega_e^2-a_{r11}a_{r22}a_{r33}-(a_{r11}a_{i33}+a_{r22}a_{i33})\omega_e+a_{r11}a_{r23}a_{r32}$$
$$+a_{r12}a_{r33}a_{r21}-a_{r13}a_{r32}a_{r21}]-(\omega_e^2-a_{i33}\omega_e-a_{r22}a_{r33}+a_{r23}a_{r32})g_1(a_{i33}a_{r22}-a_{r33}\omega_e-a_{r22}\omega_e)g_2$$

$$C_4=(k^3-1)[(a_{r11}a_{r33}+a_{r11}a_{r22}+a_{r22}a_{r33}-a_{r23}a_{r32}-a_{r12}a_{r21})\omega_e-a_{r11}a_{r22}a_{i33}+a_{i33}\omega_e^2$$
$$-\omega_e^3+a_{r12}a_{r21}a_{i33}]-(a_{r33}\omega_e+a_{r22}\omega_e-a_{r22}a_{i33})g_1-(a_{r23}a_{r32}+\omega_e^2-a_{r22}a_{r33}-a_{i33}\omega_e)g_2$$

Due to the poles of observer dynamics are proportional to the poles of IM dynamics, the error (4.3) convergence rates faster than the dynamics of the IM. The proposed such type of observer is working well from very low to high speed (Kubota 1993), (Maes 2000). It has been proven using Lyapunov direct method in (Kubota 1993), (Maes 2000) that the dynamics of observer system is stable if the real part of poles of the observer system is negative.

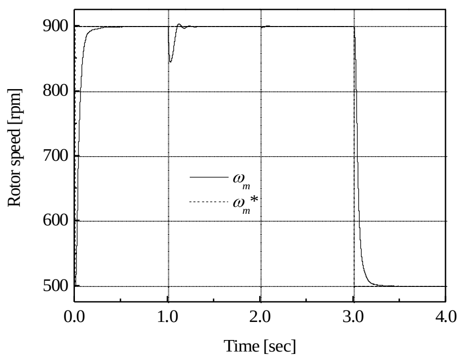
5 SIMULATION RESULTS

In order to verify the effectiveness of proposed PI controller and observer system incorporating SVM technique of PWM inverter which follow the desired speed and rotor *d*-axis flux, the simulations were carried out. The gains of the PI controllers chosen are: (i) PI speed controller: $K_{P\omega}$ =5.64×10^{-02}, $K_{I\omega}$ =1.26×10^{-04}, (ii) PI flux controller: $K_{P\Phi}$ =10.58, $K_{I\Phi}$ =2.64×10^{-02}, and (iii) PI *d*- and *q*-axis current loop: K_{Pi} =5.0, K_{Ii} =0.125.

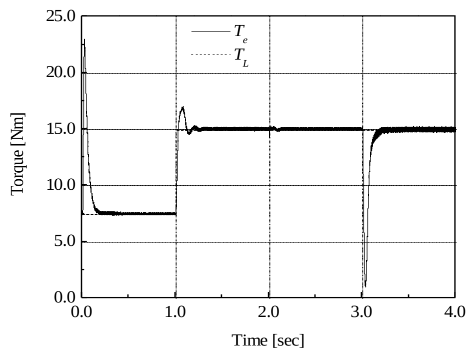
The sampling period, inverter DC link voltage and observer proportional constant are chosen as T_s=100 μsec, *E*=400 volt and *c*=1.00001 respectively. The proportional constant should be selected as near as possible to 1.0 because for the large value of *c* the gain matrix values are very large and the observer system is going to the unstable region. After selecting the all constants of controller and observer system, the proposed controller and observer system of IM drive taking core loss into account of **Fig. 3** was verified using simulation which was performed by FORTRAN. To solve the dynamic state equations of IM drives fourth-order *Runge-Kutta* method was used. In this simulation studies, the transient behavior of indirect field-oriented IM drive is evaluated for different operating conditions.

Fig. 4 shows the transient response for indirect FOC of IM using the proposed PI controller and observer system. It is seen from Figs. 4 (a) and (c) that the actual speed and rotor *d*-axis flux follow the step change of desired values without any steady state error. Figs. 4 (c) and (d) confirmed that the proposed PI controller satisfied the constraint (3.1) of FOC for the trajectory of speed and constant rotor *d*-axis flux. The proposed PI controller is also stable for the sudden change of load torque, which is clarified in Fig. 4. Fig. 4(b) shows that the electromagnetic torque can be followed the load torque. From Fig. 4, it has also been seen that the actual rotor speed and *d*-axis flux can follow the reference rotor speed and *d*-axis flux under the variation of rotor resistance. In (Uddin 2002), it has been proven that the steady state error has been occurred by using conventional PI controller. From Fig. 4, it can be state that the proposed discrete-time PI controller is better than the conventional PI controller.

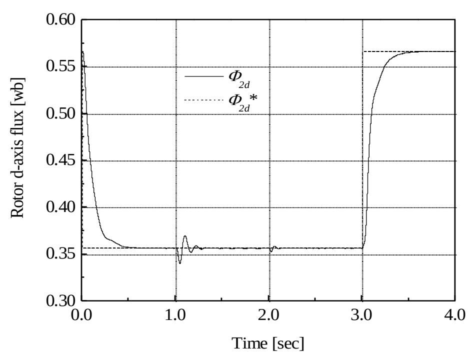
Fig. 5 shows the estimation errors between actual value and estimated value of IM state quantities. The estimation errors of stator current, magnetizing current and rotor flux



(a) Rotor speed

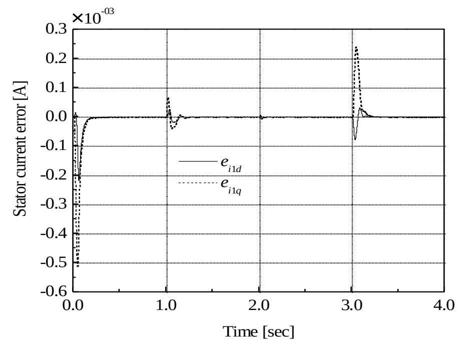


(b) Torque

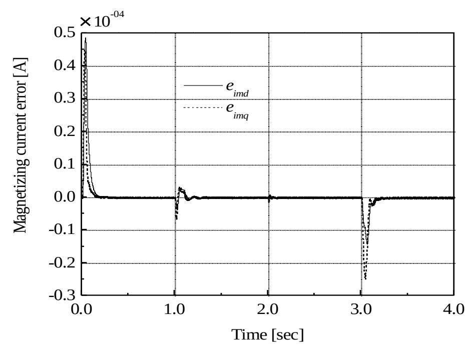


(c) Rotor *d*-axis flux

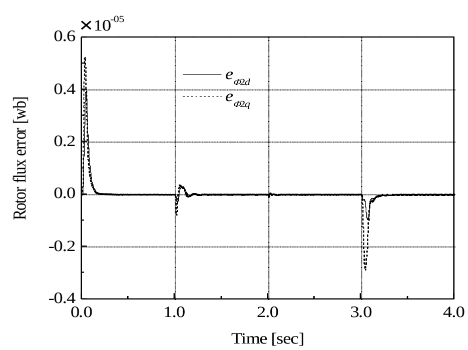
Fig. 4. Transient responses for indirect FOC of IM using the proposed PI controller and observer system.



(a) Stator current error



(b) Magnetizing current error



(c) Rotor flux error

Fig. 5. Estimation error between estimated value and actual value.

components are very small. From Fig. 5, it is clear that the errors are almost negligible. Therefore, the states of IM can be estimated properly by using the proposed observer.

As demonstrated above regarding simulation results, when the designed controller and observer are applied to SVM technique of PWM inverter-fed IM drive taking core loss into account, the trajectory of desired speed and rotor flux is obtained accurately. The proposed discrete-time PI controller is stable for uncertainty of parameters unlike the conventional PI controller. The simulation results show that the transient stability of the indirect IM drive is effectively achieved by using the proposed PI controller and observer system.

6 CONCLUSION

This paper has discussed to design discrete time PI controller and observer for indirect field oriented IM drive taking core loss into account. The discrete-time PI controller has been designed to achieve the desired speed and rotor flux. The proposed discrete-time PI controller is stable under the variation of load torque and parameters of IM drive. As the real parts of poles of proposed observer are negative, the proposed observer is also stable. Therefore, the proposed controller and observer system for SVM technique of PWM inverter-fed IM drive taking core loss into account enable to provide high performance. The simulation results show that the desired speed and rotor *d*-axis flux of an IM drive is effectively achieved by using the proposed controller and observer system incorporating SVM technique of PWM inverter. Finally, it can be concluded that the proposed discrete-time PI controller with observer system has good performance for controlling space vector PWM inverter–fed indirect field-oriented IM drive.

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