Integral Plus Proportional Controller Based Speed Control of Direct Torque Controlled IPMSM

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Abstract: Recently, interior permanent magnet synchronous motors (IPMSMs) have gained an increasing popularity in a variety of industrial applications. As the technology gets improved, studies on IPMSM such as direct torque control (DTC) method have been improved as well. The main idea in DTC is to use the motor flux and torque as basic control variables. To control speed of an IPMSM incorporated of DTC, in this paper, the authors design and simulate a controller, which is called integral plus proportional (IP) controller, to control the speed of IPMSM incorporated DTC. In order to obtain the stable performance of speed of IPMSM, the gains of designed IP controller are chosen by choosing the proper value of poles. Moreover, the chosen gains of IP controller confirm that the steady state error and the overshoot problems can be minimized and the controller becomes robust against the disturbance of load torque. The effectiveness of our designed IP controller to control speed of IPMSM incorporated with DTC method is verified by Matlab/Simulink software. It is seen from simulation works that the performance of IP controller is better as compared with the conventional proportional integral (PI) controller.

Keywords: Direct Torque Control, Speed Controller, PI Controller, IP Controller, Interior Permanent Magnet Synchronous Motor.

1. INTRODUCTION

The interior permanent magnet synchronous motor (IPMSM) in high performance industrial applications is becoming popular as compared to other type of ac motor due to some of its advantageous features including high torque to current ratio as well as high power to weight ratio, high efficiency, low noise, wider speed range, compact construction and robust operation (Jahns T. M. et al 1986). The main concerns for high performance variable speed drive (HPVSD) are fast and precise speed response, quick recovery of speed from any disturbances and parameter insensitivity, robustness in variable speed domain and maintenance free operations.

Unfortunately, the operation of the IPMSM is strongly affected by the motor magnetic saliency, saturation and armature reaction effects (Rahman M. A. et al 1996). Thus, the precise torque and speed control of IPMSM becomes a complex issue due to nonlinear coupling among its winding currents and the rotor speed as well as the nonlinearity present in the electromagnetic developed torque due to magnetic saturation of the rotor core particularly at high speeds (Morimoto S. et al 1994).

Generally, two of the control techniques of HPVSD, which are called field-oriented control (FOC) and direct torque control (DTC), respectively (Fodorean D. et al 2007), are widely used in the industrial environment. The operating principles of these control strategies are different but their objectives are the same. The aim of both controllers is to control effectively the motor torque and flux in order to force the motor to accurately track the command trajectory regardless of the machine and load parameter variation or any extraneous disturbances. Both control strategies have been successfully implemented in industrial products (Farid N. et al 2007).

The fast torque and flux control with proper estimation of torque and flux can be achieved using the DTC. Moreover, DTC is robust with respect to motor parameters and perturbations. The DTC based controllers are very little sensible to the parameters detuning in comparison with FOC, and operation with variable switching frequency and large torque ripple (due to the hysteresis comparators) are the disadvantages of these controllers (Rodriguez J. et al 2004).

It is important to design an appropriate speed controller to achieve high performance in industrial applications of IPMSM incorporated with DTC. Conventionally, the control issues are handled by conventional proportional– integral (PI) (Gumaste A. V. et al 1981), fuzzy logic control (Uddin M. N. et al 2000) and various adaptive controllers such as model reference adaptive controller, sliding-mode controller (Lin F. J. et al 2004), MIMO optimal regulator (Murata T. et al 2012), etc. All of these control techniques however increase the complexity of the HPVSD systems due to the controller design complexity. Moreover, the design of intelligent control (such as fuzzy logic, artificial neural network) in HPVSD is time consuming task since its performance depends on the expert knowledge and its need to tune a number of parameters. Fixed gain PI controllers for speed control have been used in industry for a long time because of simplicity, satisfactory steady state performance and easier real-time implementation. However, conventional controllers such as PI, PID are not suitable for HPVSD because of their sensitivity to plant parameter variations, load disturbance and any other kinds of disturbances like temperature change, command speed change, etc. (Uddin M. N. et al 2002). Moreover, the overshoot and steady state error problem cannot be minimized using one fixed set of gains.

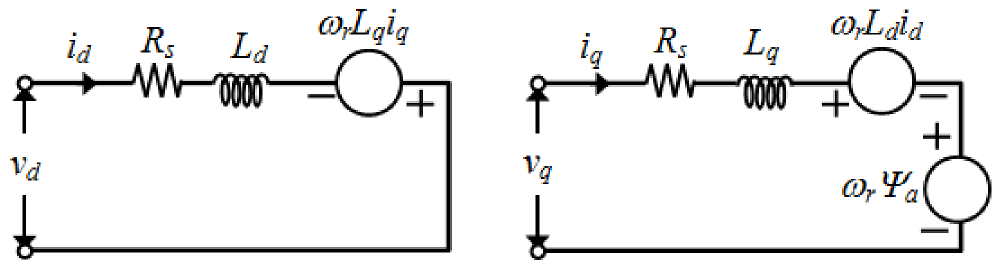
To overcome the problems of conventional PI controller, in this paper we design an IP controller (Liaw C. M. et al 1993) to control the speed of IPMSM incorporated with DTC. The design and implementation of IP controller is very simple like of those of PI controller. The designed IP controller is able to overcome the steady state error and overshoot problems and to reject the disturbance. The poles are selected to locate the eigenvalues of the transfer functions in the left half of complex s-plane.

The effectiveness of our designed IP controller to control speed of IPMSM incorporated with DTC method is verified by simulation results, which were carried out by Matlab/Simulink software. It is seen from simulation works that the performance of IP controller is better as compared with the conventional proportional integral (PI) controller.

2. DYAMIC MODEL IPMSM

Fig. 1 shows the *d*- axis and *q*- axis equivalent circuit of an IPMSM. In the *d*-*q* coordinates, which rotate synchronously with an electrical angular velocity ω_r, the voltage equation, speed equation and torque equation of IPMSM can be described as follows (Zhong L. et al 1997):

$$\left.\begin{aligned} v_d &= R_s i_d + p\lambda_d - \omega_r \lambda_q \\ v_q &= R_s i_q + p\lambda_q + \omega_r \lambda_d \end{aligned}\right\} \tag{1}$$



(a) *d*-axis (b) *q*-axis

Fig. 1 IPMSM equivalent circuit in a synchronously rotating reference frame.

$$\left.\begin{aligned}\lambda_d &= L_d i_d + \Psi_a \\ \lambda_q &= L_q i_q\end{aligned}\right\} \tag{2}$$

$$T_e = P_n[\lambda_d i_q - \lambda_q i_d] \tag{3}$$

$$J_m p\omega_r = -B_m \omega_r + P_n(T_e - T_L) \tag{4}$$

where, v_d and v_q are the *d*-axis and *q*-axis components of terminal voltage, i_d and i_q are the *d*-axis and *q*-axis components of armature current, λ_d and λ_q are the *d*-axis and *q*-axis components of flux, L_d and L_q are the *d*-axis and *q*-axis components of armature self-inductance, R_s is the armature resistance, $p = d/dt$, P_n is the number of pole pairs, T_e is the electromagnetic torque, T_L is the load torque, B_m is friction coefficient, J_m is the inertia and Ψ_a is the rotor magnetic flux linkage.

The stator flux linkage and the torque can be estimated in the stationary (α-β) frame. The equations are as follows (Peter Vas. 1990):

$$\left.\begin{aligned}\lambda_\alpha &= \int (v_\alpha - R_s i_\alpha)dt \\ \lambda_\beta &= \int (v_\beta - R_s i_\beta)dt\end{aligned}\right\} \tag{5}$$

$$T_e = P_n[\lambda_\alpha i_\beta - \lambda_\beta i_\alpha] \tag{6}$$

The absolute value and the phase angle of stator flux are given by the following equation.

$$|\lambda_s| = \sqrt{\lambda_\alpha^2 + \lambda_\beta^2} \tag{7}$$

$$\theta_s = \tan^{-1}(\lambda_\beta / \lambda_\alpha) \tag{8}$$

The estimated flux and torque by means of using Eqs. (5) to (7) are used to calculate the difference between the desired quantities and actual quantities to apply in DTC.

3. DIRECT TORQUE CONTROL STRATEGY OF IPMSM

DTC was introduced in Japan by Takahashi and Nagochi (Takahashi I. et al 1986) and also in Germany by Depenbrock (Baader, U. et al 1992; Depenbrock M. 1988). These innovations depart from the vector control in which the coordinate transformation is crucially needed. The DTC control method relies on a bang-bang control instead of a decoupling control which is the characteristic of vector control. Their technique (bang-bang control) works very well with the on-off operation of inverter semiconductor power devices.

The basic concept behind the DTC of AC drive is to control the torque and flux linkage directly and independently by the use of six or eight voltage space vectors found in lookup tables. The possible eight voltage space vectors used in DTC are shown in **Fig. 2** (Peter Vas. 1990). The voltage vector plane is divided into six sectors so that each voltage vector divides

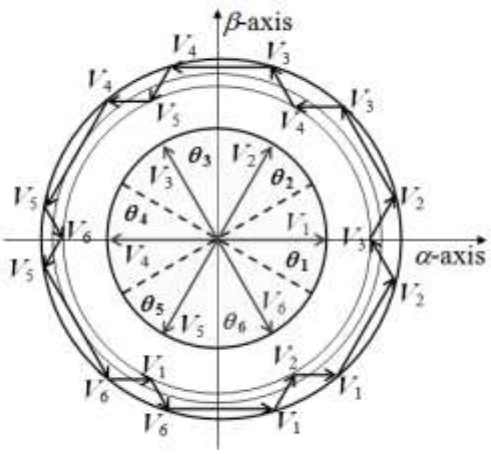


Fig. 2 Voltage Vectors for direct torque control.

Table1
Switching Table Presented by Takahashi and Noguchi

θ, C_T, C_F		θ_6	θ_6	θ_6	θ_6	θ_6	θ_6
	C_T =1	$\boldsymbol{V}_2$	$\boldsymbol{V}_3$	$\boldsymbol{V}_4$	$\boldsymbol{V}_5$	$\boldsymbol{V}_6$	$\boldsymbol{V}_1$
C_F =1	C_T =0	$\boldsymbol{V}_7$	$\boldsymbol{V}_0$	$\boldsymbol{V}_7$	$\boldsymbol{V}_0$	$\boldsymbol{V}_7$	$\boldsymbol{V}_0$
	C_T =-1	$\boldsymbol{V}_6$	$\boldsymbol{V}_1$	$\boldsymbol{V}_2$	$\boldsymbol{V}_3$	$\boldsymbol{V}_4$	$\boldsymbol{V}_5$
	C_T =1	$\boldsymbol{V}_3$	$\boldsymbol{V}_4$	$\boldsymbol{V}_5$	$\boldsymbol{V}_6$	$\boldsymbol{V}_1$	$\boldsymbol{V}_2$
C_F =0	C_T =0	$\boldsymbol{V}_0$	$\boldsymbol{V}_7$	$\boldsymbol{V}_0$	$\boldsymbol{V}_7$	$\boldsymbol{V}_0$	$\boldsymbol{V}_7$
	C_T =-1	$\boldsymbol{V}_5$	$\boldsymbol{V}_6$	$\boldsymbol{V}_1$	$\boldsymbol{V}_2$	$\boldsymbol{V}_3$	$\boldsymbol{V}_4$

each region into two equal parts. In each sector, four of the six non-zero voltage vectors may be used. Also zero vectors are allowed.

The appropriate stator voltage vector out of eight possible inverter states (according to the difference between the reference and actual torque and flux linkage that means according to the outputs of torque hysteresis and flux hysteresis comparators) is chosen so that the stator flux linkage vector rotates along the stator reference frame (α-β frame) trajectory and produces the desired torque.

The output of the torque hysteresis comparator is denoted as C_T, the output of the flux hysteresis comparator as C_F and the flux linkage sector is denoted as θ. The torque hysteresis comparator is a three valued comparator. C_T = **-** 1 means that the actual value of the torque is above the reference and out of the hysteresis limit, and C_T = 1 means that the actual value is below the reference and out of the hysteresis limit. C_T = **-** 1 means that the difference between actual and reference values of the torque and inside the hysteresis limit. The flux hysteresis comparator is a two valued comparator. C_F = 0 means that the actual value of the flux linkage is above the reference and out of the hysteresis limit and C_F = 1 means that the actual value of the flux linkage is below the reference and out of the hysteresis limit (Rahman M. F. et al 2003). All the possibilities can be tabulated into a switching table as shown in **Table 1** (Takahashi I. et al 1986).

4. DESIGN OF IP SPEED CONTROLLER OF IPMSM

The speed controller is design to obtain the desired torque from the difference between the reference and actual values of speed. Here, we design an IP controller to obtain the desired torque. Assuming that the change of load torque is a step function (*i.e.* the change of load torque is zero), the open loop transfer function, which is obtained from Eqs. (4) of speed controller in terms of electromagnetic torque can be expressed by the following equation:

$$G_P(s) = \frac{\omega_r(s)}{T_e(s)} = \frac{A}{s+B} \tag{9}$$

where, $A = P_n / J_m$, $B = B_m / J_m$.

The transfer function of IP controllers is written as follows (Liaw C. M. et al 1993):

$$G_{IP}(s) = \frac{K_I}{s}[\omega_r^*(s) - \omega_r(s)] - K_P\omega_r(s) \tag{10}$$

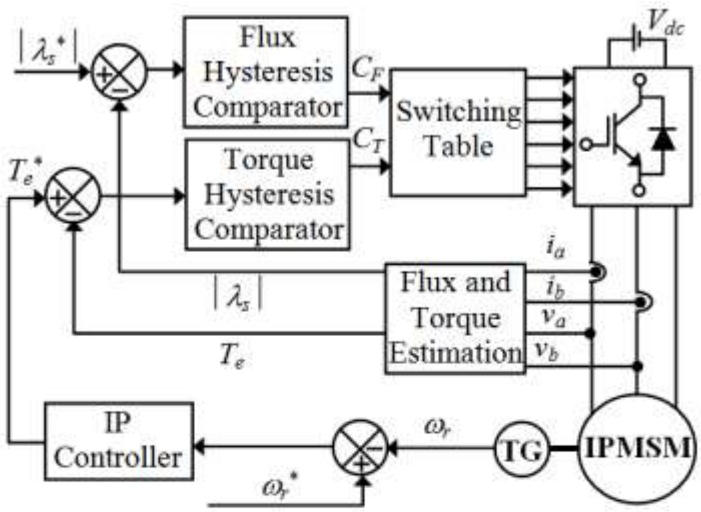


Fig. 3 Block diagram of overall system.

where, $\omega_r^*(s)$ indicates the reference speed.

Therefore, the transfer function of closed loop system based on IP controller can be given by the following equation:

$$G_{CS}(s) = \frac{\omega_r(s)}{\omega_r^*(s)} = \frac{BK_I}{s^2 + (A + BK_P)s + BK_I} \tag{11}$$

The transfer function (11) can be expressed as follows:

$$G_{CS}(s) = \frac{h_1}{s+\alpha_1} + \frac{h_2}{s+\alpha_2} \tag{12}$$

where, α_1 and α_2 represent the eigenvalues or poles of closed loop dynamics Eq. (12) and

$$\left.\begin{array}{ll} h_1 + h_2 = 0; & \alpha_1 + \alpha_2 = A + BK_P \\ \alpha_1\alpha_2 = BK_I; & \alpha_2 h_1 + \alpha_1 h_2 = BK_I \end{array}\right\} \tag{13}$$

According to the theory which is explained in (Liaw C. M. et al 1993), the solution of α_1, α_2 , K_P and K_I can be given by the following equation:

$$\left.\begin{array}{l} \alpha_1 = \dfrac{2.9697}{t_r} \\ \alpha_2 = 2\alpha_1 \end{array}\right\} \tag{14}$$

$$\left.\begin{array}{l} K_P = \dfrac{3\alpha_1 - A}{B} \\ K_I = \dfrac{2\alpha_1^2}{B} \end{array}\right\} \tag{15}$$

where, t_r is the rise-time.

Choosing the poles and controller parameters according to (14) and (15), the tracking steady-state error is zero. There has also no overshoot problem and the proposed IP controller is stable under the variations of load torque because the poles are negative real values.

Table 2 Ratings and Parameters of IPMSM.

Stator resistance, R_s	5.8 Ω
d-axis self inductance, L_d	44.8 mH
q-axis self inductance, L_q	102.7 mH
Rotor flux constant, λ_s	0.533 WB
Moment of Inertia, J_m	0.00529 kg-m^2
Damping coefficient, B_m	0.00006 kg-m^2/s
Number of pole pairs, P_n	2
Mechanical torque, T_m	10 Nm

5. SIMULATION RESULTS

In order to verify the performance of proposed IP controller based speed control of IPMSM incorporated with DTC, computer simulations were performed using Matlab/Simulink. In the simulation studies different operating conditions are observed for DTC and the speed controller strategies. The simulation has done like the overall block diagram is shown in **Fig.3**. The rating and the parameters values of used IPMSM in the simulation are given in **Table 2**. The sampling time and dc voltage of inverter for this simulation is considered 50 μsec, and 350 V, respectively.

At first the results are observed for the DTC to control the torque and flux. **Fig. 4** shows the transient responses of torque and flux for the step change of reference torque while the reference value of magnitude of stator flux is kept constant at 0.6 Wb. The reference torque is stepped-up from its 50% rated value to 100% of its rated value at t=1.0 sec and again stepped-down from its 100% rated value to 50%of its rated value at t=2.0 sec. It is observed from Fig. 4(a) the actual torque follows the reference torque quickly using the DTC strategy by using the switching Table 1. The actual magnitude of flux can also follow the reference of magnitude of stator flux as shown if Fig. 4(b).

The results obtained by conventional DTC for the step change of magnitude of stator flux are represented in **Fig. 5**. The reference torque is stepped-up from 0.3 Wb to 0.6 Wb at t=1.0 sec and again stepped-down from its 0.6 Wb to 0.3 Wb at t=2.0 sec. It is observed from Fig. 5(b) the actual magnitude of flux follows the reference of magnitude of flux quickly. The actual torque can also follow the reference of torque as shown in Fig. 5(a). From Figs. 4 and 5 it is clear that the torque and magnitude of stator flux can be control with the DTC strategy.

In order to control of speed of IPMSM with DTC, in this paper we design an IP controller. From the IP controller, the reference torque, which is used a reference torque of DTC, is obtained. Moreover, the simulation has done using the PI controller for different gains of its to clarify the problems of a conventional PI controller. **Fig. 6** shows the transient response of IPMSM due to the step change of speed and load torque based on the PI controller. In this work, three different cases of gains of PI controller are considered. The gains, which are used this simulation, are as follows:

For Case 1: $K_{P\omega}$=0.09; $K_{I\omega}$=1.0e-06

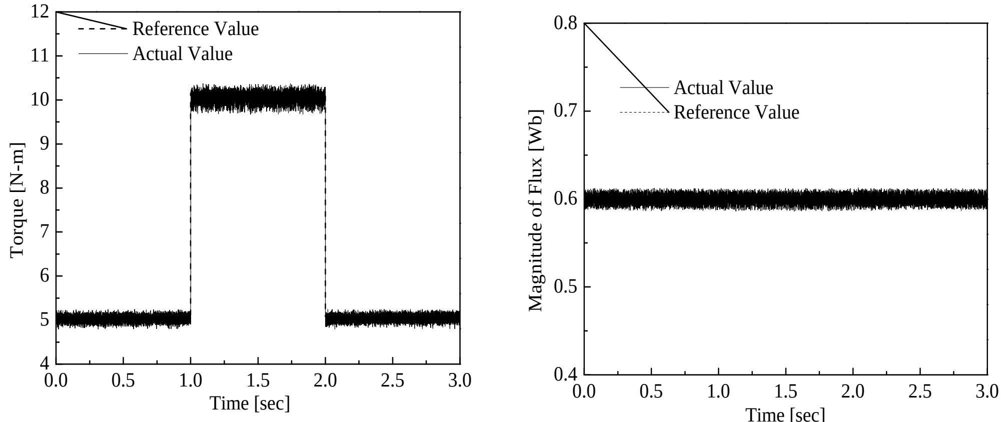
For Case 2: $K_{P\omega}$=0.1; $K_{I\omega}$=5.0e-02

For case 3: $K_{P\omega}$=0.12; $K_{I\omega}$=1.17

It is observed from Fig. 6 that the steady-state error problem, the overshoot problem, and the stable performance under the variation of load disturbance cannot be eliminated by using a one fixed set of gains of a PI controller.

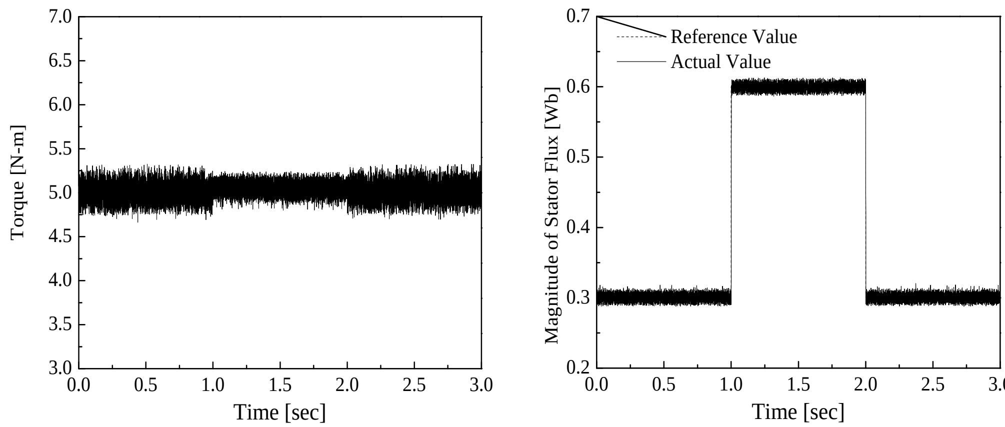
The effectiveness of the designed of IP controller is also verified by simulation results. **Fig. 7** and **8** shows the responses of speed, flux, torque and stator current for the step change of speed and the load torque. The reference speed stepped-up from 500 rpm to 1000rpm at t=1.0 sec and again stepped-down from 1000 rpm to 500 rpm at t=3.0 sec. At t = 2.0 sec the load torque is changed from 5 N-m to 10 N-m. To calculate the gains of IP controller, the rise-time is chosen as t_r = 0.075 sec.

The response of speed is obtained without steady-state error and overshoot problems as shown in Fig. 7 (a). The IP controller is also performed as stable controller under the variations of load torque which is confirmed from Fig. 7(a). The actual flux can follow the reference flux even the change of speed and load torque. The compensation of change of load torque is possible by electromagnetic torque is shown in Fig. 8(a). Fig. 8(a) and (b) shows the α-axis and β-axis stator current of IPMSM. It is seen from the current response, the magnitude of current will not be changed due to change of speed but the frequency is changed to reach to the desired value. The magnitude of stator current is increased as increases the load torque. From the observation of simulation results it is cleared the performance of IP controller is better as compared with those of PI controller.



(a) Torque (b) Magnitude of Stator Flux.

Fig. 4 Transient responses of DTC for step change of reference torque.



(a) Torque (b) Magnitude of Stator Flux

Fig. 5 Transient responses of DTC for step change of reference of magnitude of stator flux.

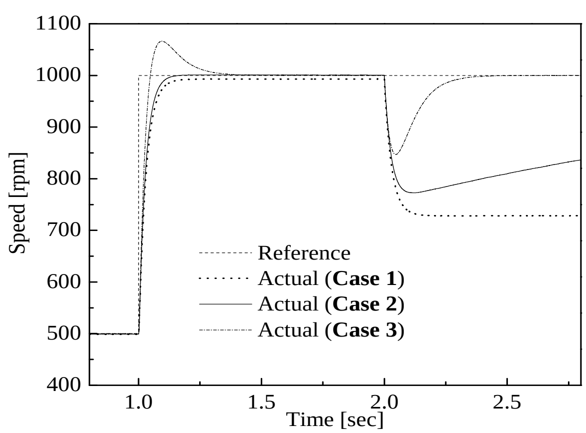
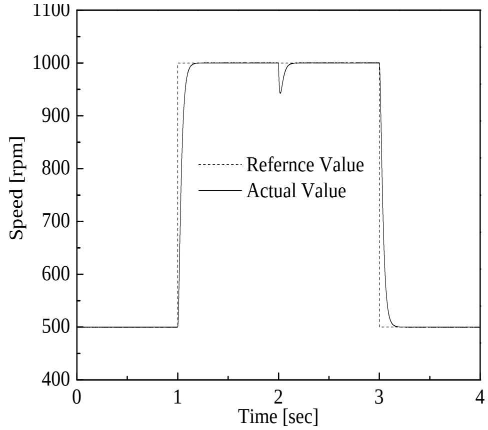
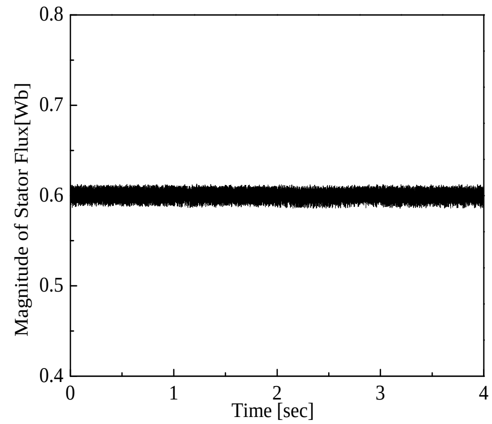


Fig. 6 Transient responses of speed based of PI controller for three different values of its gains.

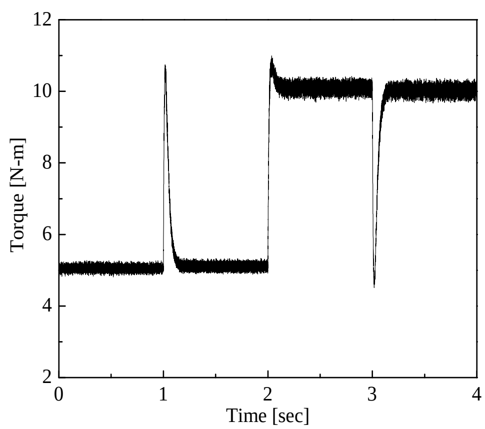


(a) Speed

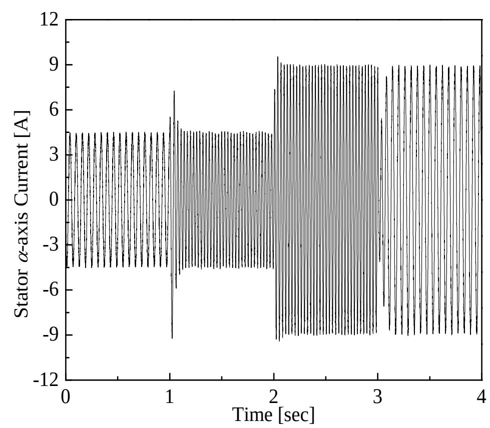


(b) Magnitude of Stator Flux.

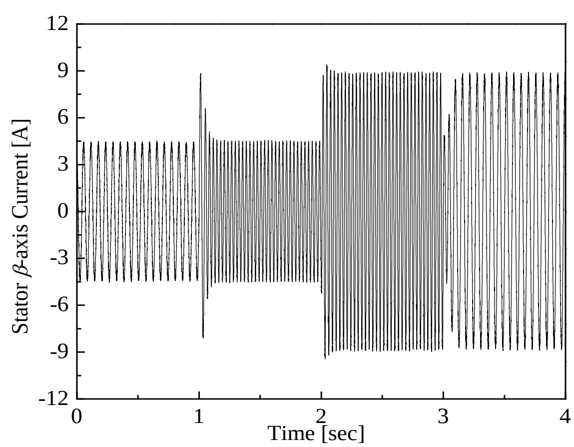
Fig. 7 Transient responses of speed and flux for step change of speed based of IP controller.



(a) Torque



(b) Stator α-axis Current



(c)a Stator β-axis Current

Fig. 8 Transient responses of torque and stator current for step change of speed based of IP controller.

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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