A ROBUST CONTROL OF PM SYNCHRONOUS MOTOR USING ACCELERATING TORQUE FEEDBACK

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ABSTRACT - A robust control technique is presented for a high performance control of a permanent-magnet(PM) synchronous motor. In order to deal with the internal and external disturbances of a PM synchronous motor drive system, a new feedback control structure is proposed. Since the dynamic behavior of the PM synchronous motor drive system is mainly concerned with the difference between the electro-magnetically developed torque and the load torque which generally referred to as an accelerating torque, the estimation and control techniques of this torque are introduced. The simulations and experiments are carried out for the DSP-based PM synchronous motor drive system and the results well demonstrate the effectiveness of the proposed control technique.

NOMENCLATURE

v_{ds}, v_{qs}	:	d and q axis voltages, respectively
i_{ds}, i_{qs}	:	d and q axis currents, respectively
$\lambda_{ds}, \lambda_{qs}$	:	d and q axis stator fluxes, respectively
L_d, L_q	:	d and q axis inductances, respectively
M_{dq}, M_{qd}	:	mutual inductance between d and q axis
M_{df}, M_{qf}	:	mutual inductance between d/q axis and permanent magnet
τ_e	:	electro-magnetically developed torque
τ_L	:	load torque
τ_a	:	accelerating torque
θ_r	:	electrical angular position of rotor
ω_r	:	electrical angular velocity of rotor
P	:	number of poles
J	:	inertia moment of motor and load
B	:	viscous damping coefficient
'^'	:	estimated value
'*'	:	reference value
'n'	:	nominal value

1. INTRODUCTION

Permanent magnet (PM) synchronous motors are generally used in high performance drive applications such as industrial robots, machine tools, and aerospace actuators because of their high power density, high torque to inertia ratio, low maintenance, and ability to operate in explosive environment. Moreover, using the current regulated PWM (CRPWM) technique associated with the concept of the field-orientated control, the PM synchronous motor can be controlled to have the speed-torque characteristics similar to that of the DC motor [5], [8], [10].

However, there exist some problems in the control of the PM synchronous motor. First, there exists the load torque and inertia variation. Most of previous approaches dealing the control of the PM synchronous motor mainly consider the rejection of the effects of the load torque and inertia variation regarding as the disturbances which affect on the speed control performance and shows robustness against mechanical parameter variation [9], [10]. But, they can not consider next problem.

The next problem in the control of the PM synchronous motor is from the assumption that the linkage flux of the motor is constant in space. It is, however, impossible in practical, i.e. the linkage flux of the motor contains some undesirable harmonic components which cause torque ripple even with sinusoidal feed currents, and varies nonlinearly with the temperature rise. To solve the harmonics problem. Some approaches are presented. One is the harmonic cancellation technique by using the combination of the back EMF harmonics and feed current waveform [5],[6]. But, this approach is only implemented off-line manner, so can not consider various operating condition. The other is the dynamic feedback control of the instantaneous torque, which is the electro-magnetically developed torque in the motor and generally referred to as an instantaneous torque control. This technique employs a least square method or a model reference adaptive system (MRAS) technique to estimate the instantaneous torque [7], [8].

But, the dynamic behavior of the PM synchronous motor depends on not the developed torque, τ_e but the difference between the developed torque and the load torque, $\tau_a = \tau_e - \tau_l$, which is referred to as accelerating torque. So, to deal with these problems, an accelerating torque feedback control scheme is proposed in this paper. If the accelerating torque is perfectly controlled in the inner control loop, the speed dynamics of the motor is independent on the load torque and parameter variation

This paper describes the estimation and control of the accelerating torque using adaptive and variable structure control techniques. By using an adaptive torque observer, accelerating torque is estimated and then this estimated torque is fed to the torque controller. The proposed control scheme is applied to the PM synchronous motor drive system and implemented in a digital manner using DSP TMS320C30. The simulations and experiments are carried out for this system to show the effectiveness of the proposed scheme.

2. MODELING OF PMSM

In general, the PM synchronous motor has three phase stator windings and a permanent magnet rotor. Assuming the PM rotor is a fictitious winding with a constant current source i_f, the voltage equation and the developed torque equation of a PM synchronous motor in the synchronous rotating reference frame can be expressed as follows [1], [2], [7],[8]:

$$v_{ds} = r_s i_{ds} + L_{d0}\frac{di_{ds}}{dt} - L_{q0} i_{qs}\omega_r + \psi_{qm}\omega_r \qquad (1)$$

$$v_{qs} = r_s i_{qs} + L_{q0}\frac{di_{qs}}{dt} + L_{d0} i_{ds}\omega_r + \psi_{dm}\omega_r \,. \qquad (2)$$

$$\tau_e = \frac{3}{2}\frac{P}{2}[\psi_{dm} i_{qs} + \psi_{qm} i_{ds} + (L_{d0} - L_{q0}) i_{qs} i_{ds}] \qquad (3)$$

where

$$\psi_{dm} = \sum_{n=0}^{\infty}\left(6nM_{qf6n} + M_{df6n}\right)\cos 6n\theta_r \cdot i_f$$

$$\psi_{qm} = -\sum_{n=0}^{\infty}\left(6nM_{df6n} + M_{qf6n}\right)\sin 6n\theta_r \cdot i_f \,.$$

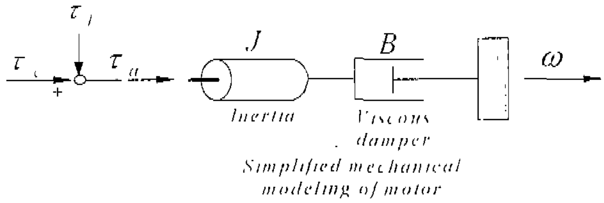
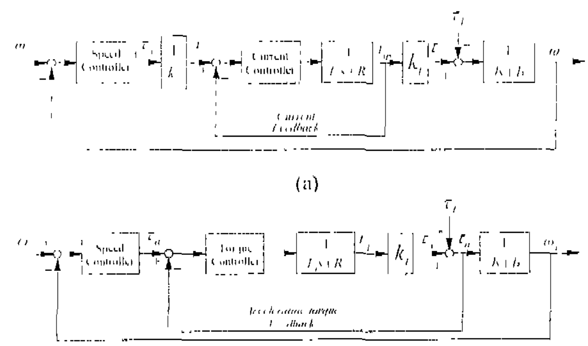


Fig. 1 Schematic diagram of the mechanical system of the motor


(a)

(b)

Fig. 2 Comparison of speed control scheme of the PMSM
(a) Conventional scheme employing CRPWM
(b) Proposed scheme employing accelerating torque feedback

By using the concept of the field orientation, it can be assumed that the d axis current i_d is controlled to be zero, in which the developed torque of the motor is maximized for a given stator current. Under this assumption, the torque and current relation can simply be described as

$$\tau_e = k_t i_{qs} = \frac{3}{2}\frac{P}{2}\psi_{dm} i_{qs} \,. \qquad (4)$$

The motor model is completed by the inclusion of the mechanical equation of the motor and load and it can be represented as

$$\tau_e = J\left(\frac{2}{P}\right)\frac{d\omega_r}{dt} + B\left(\frac{2}{P}\right)\omega_r + \tau_l \,. \qquad (5)$$

3. ROBUST CONTROL USING ACCELERATING TORQUE FEEDBACK

The basic concept of the proposed scheme is derived from a schematic diagram of the mechanical system of the motor as shown in Fig. 1. It can be observed in this figure and equation (5) that the dynamic behavior of the motor depends on not the developed torque τ_e, but the difference between the developed torque and load torque $\tau_a = \tau_e - \tau_l$, i.e. accelerating torque. However, most existing approaches mainly consider the control of the developed torque of the motor and thus the load torque is regarded as disturbance which largely affects the speed control performance. Fig. 2(a) shows the conventional speed control scheme of the PM synchronous motor employing a CRPWM drive. If the torque constant, k_t, is constant and exactly known, the developed torque can be controlled by the inner loop current controller.

However, the torque constant is not constant due to deviations from a sinusoidal flux density distribution around the airgap. Also, the flux linkage varies nonlinearly the temperature rise. In order to solve these problems, several techniques using the preknowledge of the motor parameters have been proposed [5]-[8]. One of the most popular approaches is the dynamic feedback of the developed torque employing an on-line flux estimator [7],[8]. However, to obtain high quality control performance, the effect of the load torque should be rejected by the speed controller in the outer control loop [9], [10].

To deal with this problem, an accelerating torque feedback control is proposed. Fig. 2(b) shows the concept of the proposed control. If the accelerating torque is perfectly controlled in the inner control loop, the speed dynamics of the motor is independent on the load torque. Therefore, the speed controller in the outer control loop can be simply design without considering the effect of the load torque.

Unfortunately, the major difficulty of this scheme is how to obtain the information on the accelerating torque which is not accessible. Since the measuring system of this torque is very expensive and bulky, it can not be employed in industrial servo systems.

Therefore, an estimating technique is introduced in this paper. The linkage flux of a motor is estimated by the MRAS technique and the developed torque is calculated by using the estimated linkage flux and measured currents. The load torque which includes the effect of the inertia variation is estimated by the adaptive observer and the accelerating torque can be calculated from the estimated developed torque and load torque. Then, the accelerating torque is controlled by the torque controller using variable structure control (VSC) with an integral action, so called IVSC.

4. DESIGN OF PROPOSED TORQUE CONTROLLER

The block diagram for the proposed torque control scheme is shown in Fig. 3. The overall system consists of a developed torque estimator, a load torque observer, and a torque controller using IVSC.

Design of developed torque estimator using MRAS technique

The mathematical model of the PM synchronous motor given in (1) and (2) can be represented in the state space form as follows:

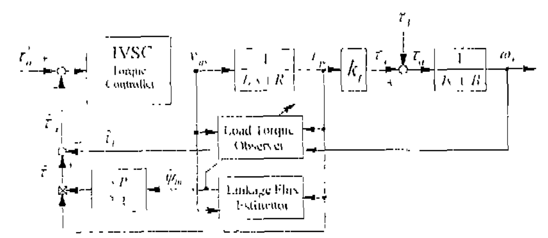


Fig. 3 Proposed torque control scheme employing torque estimator and controller

$$\dot{x} = Ax + Bu + D\psi \tag{6}$$

where

$$x = [i_{ds} \;\; i_{qs}]^T, \quad u = [v_{ds} \;\; v_{qs}]^T, \quad \psi = [\psi_{qm} \;\; \psi_{dm}]^T$$

$$A = \begin{bmatrix} -\frac{r_s}{L_{d0}} & \frac{L_{q0}}{L_{d0}}\omega_r \\ -\frac{L_{d0}}{L_{q0}}\omega_r & -\frac{r_s}{L_{q0}} \end{bmatrix}, \qquad B = \begin{bmatrix} \frac{1}{L_{d0}} & 0 \\ 0 & \frac{1}{L_{q0}} \end{bmatrix},$$

$$D = \begin{bmatrix} -\frac{\omega_r}{L_{d0}} & 0 \\ 0 & \frac{\omega_r}{L_{q0}} \end{bmatrix}.$$

For this model, the adjustable system to estimate the flux ψ can be chosen as follows:

$$\dot{\hat{x}} = A\hat{x} + Bu + D\hat{\psi} - Fe \tag{7}$$

where $e = x - \hat{x}$ and F is the gain matrix to determine the convergence rate and to decouple the time varying terms. The adaptation rule to estimate the linkage flux ψ can be given as

$$\dot{\hat{\psi}} = \gamma D^T G e \tag{8}$$

where γ is an adaptation gain, and G is a solution of the Lyapunov equation [8].

Using the estimated linkage flux, the developed torque of the PM synchronous motor can be estimated as follows:

$$\hat{\tau}_e = \frac{3}{2}\frac{P}{2}\left(\hat{\psi}_{dm} i_{qs} + \hat{\psi}_{qm} i_{ds}\right). \tag{9}$$

Design of load torque observer

By using the concept of a field orientation, it can be assumed that the d axis current i_d is controlled to be zero, in which the developed torque of the motor is maximized for a given stator current. Under this assumption, the torque and current relation can simply be described as

$$\tau_e = k_t i_{qs} \tag{4}$$

where

$$k_t = \frac{3}{2}\frac{P}{2}\psi_{dm}.$$

And another assumption of the load torque observer is that the load torque is constant as $\dot{\tau}_l = 0$. Under the above assumptions, the observer to estimate the load torque is given as follows :

$$\begin{pmatrix}\dot{\hat{\omega}}_r \\ \dot{\hat{\tau}}_l\end{pmatrix} = \begin{pmatrix}-\frac{B}{J} & -\frac{P}{2}\frac{1}{J} \\ 0 & 0\end{pmatrix}\begin{pmatrix}\hat{\omega}_r \\ \hat{\tau}_l\end{pmatrix} + \begin{pmatrix}\frac{P}{2}\frac{\hat{k}_t}{J} \\ 0\end{pmatrix} i_{qs} - \begin{pmatrix}l_1 \\ l_2\end{pmatrix}(\omega_r - \hat{\omega}_r) \qquad (10)$$

where $\hat{k}_t = \frac{3}{2}\frac{P}{2}\hat{\psi}_{dm}$.

From the above results, the accelerating torque can be estimated as follow:

$$\hat{\tau}_a = \hat{\tau}_e - \hat{\tau}_l .$$

and this estimated accelerating torque is feedback to torque controller.

Design of torque controller using VSC with integral action

In the proposed control scheme, the control variables of interest are the d axis current i_{ds} and the accelerating torque τ_a. The sliding surface of the proposed control scheme can be chosen as [3]

$$s_i = X_i + c_i \int_0^t X_i(\varsigma)d\varsigma = 0 \qquad (11)$$

$$s_\tau = X_\tau + c_\tau \int_0^t X_\tau(\varsigma)d\varsigma = 0 \qquad (12)$$

where $X_i \equiv i_{ds} - i_{ds}^*$, $X_\tau \equiv \tau_a - \tau_a^*$ and c_i, c_τ are the coefficients of the sliding surface. The control input of the torque controller consists of the equivalent control input v^{eq} and the switching control input $\Delta v_{i\tau}$, and can be given as follows:

$$v_{i\tau} = v_{i\tau}^{eq} + \Delta v_{i\tau}$$

where

$$v_{i\tau} = \begin{pmatrix}v_{ds} \\ v_{qs}\end{pmatrix},\ v_{i\tau}^{eq} = \begin{pmatrix}v_i^{eq} \\ v_\tau^{eq}\end{pmatrix},\ \text{and}\ \Delta v_{i\tau} = \begin{pmatrix}\Delta v_i \\ \Delta v_\tau\end{pmatrix}.$$

By using the condition of the equivalent control given as $\dot{s}_{i\tau} = 0$, the equivalent control $v_{i\tau}^{eq}$ can be derived as

$$v_i^{eq} = r_s^o i_{ds} + e_d^o - L_{d0}^o c_i X_i + L_{d0}^o \dot{i}_d^* \qquad (13)$$

$$v_\tau^{eq} = r_s^o \frac{\tau_e}{k_t} + e_q^o - L_{q0}^o c_\tau X_\tau + L_{q0}^o \frac{\dot{\tau}_e^*}{k_t} . \qquad (14)$$

The switching control input $\Delta v_{i\tau}$ can be given as

$$\Delta v_i = K_{i1} X_i + K_{i2} \qquad (15)$$

$$\Delta v_\tau = K_{\tau 1} X_\tau + K_{\tau 2} . \qquad (16)$$

The gains of the control inputs K_{i1}, K_{i2}, $K_{\tau 1}$, and $K_{\tau 2}$ can be determined by using the well known sliding mode existence condition given as[3],[8]

$$s_i \dot{s}_i < 0 \qquad (17)$$

$$s_\tau \dot{s}_\tau < 0 . \qquad (18)$$

From the above inequalities, the gains of the control inputs are determined as follows:

$$K_{i1} = \begin{cases}\alpha_{i1} > \max(\Delta r_s - \Delta L_d c_i) & \text{for } s_i X_i < 0 \\ \beta_{i1} < \min(\Delta r_s - \Delta L_d c_i) & \text{for } s_i X_i > 0\end{cases} \qquad (19)$$

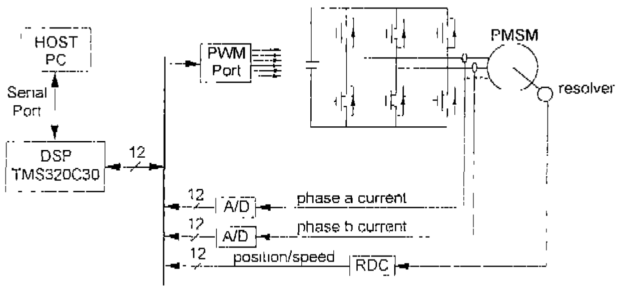


Fig. 4 Configuration of experimental system

$$K_{i2} = \begin{cases}\alpha_{i2} > \max(\Delta r_s i_d^* + \Delta L_d \dot{i}_d^* + \Delta e_d) & \text{for } s_i < 0 \\ \beta_{i2} < \min(\Delta r_s i_d^* + \Delta L_d \dot{i}_d^* + \Delta e_d) & \text{for } s_i > 0\end{cases} \qquad (20)$$

$$K_{\tau 1} = \begin{cases}\alpha_{\tau 1} > \max\left[(\Delta r_s - \Delta L_q c_\tau)\frac{1}{k_t}\right] & \text{for } s_\tau X_\tau < 0 \\ \beta_{\tau 1} < \min\left[(\Delta r_s - \Delta L_q c_\tau)\frac{1}{k_t}\right] & \text{for } s_\tau X_\tau > 0\end{cases} \qquad (21)$$

$$K_{\tau 2} = \begin{cases}\alpha_{\tau 2} > \max\left[(\Delta r_s \tau_e^* + \Delta L_q \dot{\tau}_e^*)\frac{1}{k_t} + \Delta e_q\right] & \text{for } s_\tau < 0 \\ \beta_{\tau 2} < \min\left[(\Delta r_s \tau_e^* + \Delta L_q \dot{\tau}_e^*)\frac{1}{k_t} + \Delta e_q\right] & \text{for } s_\tau > 0\end{cases} \qquad (22)$$

where the symbol ' Δ ' denotes the difference between real value and nominal value of motor parameter.

For the practical implementation, the switching input $\Delta v_{i\tau}$ is approximated to the continuous input by using the saturation function.

5. SIMULATIONS AND EXPERIMENTS

Configuration of overall system

The configuration of the experimental system is shown in Fig. 4. The processor is DSP TMS320C30 with a clock frequency of 33MHz. The whole control algorithm including the torque estimator, load torque observer, and torque controller is implemented by the assembly language program of TMS320C30. The sampling frequency is set to 7.8 kHz. The PM synchronous motor is driven by a three-phase PWM inverter employing the intelligent power module (IPM) with switching frequency 7.8 kHz. The specification of the motor is listed in the Table 1.

Table 1 Specifications of experimental PMSM

Rated power	400 W	Rated speed	3000 rpm
Rated torque	1 274 Nm	Torque constant	0.501 Nm/A
Stator resistance	3.0 Ω	Stator inductance	20 mH
No. of poles	4	Moment of inertia	1.54×10^{-4} Nms[2]

Determination of estimator and controller gains

In order to show the usefulness of the proposed scheme, the simulations and experiments are carried out for the PM synchronous motor drive system described in the previous section.

The gains of the torque estimator and controller are determined using the actual parameters of the experimental PM synchronous motor. The developed torque estimator gain matrix F given in (4) is chosen as

$$\begin{bmatrix} 400 & -\omega_r \\ \omega_r & 400 \end{bmatrix}$$

which cancels the time varying terms, so that the poles of $\bar{A} = A + F$ are determined as $p_i = 500$ and $p_\tau = 500$. The solution of the Lyapunov equation G is obtained as

$$\begin{bmatrix} 0.001 & 0 \\ 0 & 0.001 \end{bmatrix}$$

for $Q = I_2$, where I_2 denotes the identity matrix of 2×2 dimension. The adaptation gain γ is chosen as 40. And the gains of the load torque observer are chosen as

$$\begin{pmatrix} l_1 \\ l_2 \end{pmatrix} = \begin{pmatrix} -793 \\ 16.425 \end{pmatrix}$$

so that observer poles are placed at $s = -400 \pm j300$.

The coefficients of the sliding surface are chosen as c_i = 100 and c_τ = 100, respectively, so that the time constants of X_i and X_τ are both determined as 10 msec. The gains of control inputs are chosen from (19) to (22) as $\alpha_{i1} = 3$, $\beta_{i1} = -3$, $\alpha_{i2} = 10$, $\beta_{i2} = -10$, $\alpha_{\tau 1} = 70$, $\beta_{\tau 1} = -70$, $\alpha_{\tau 2} = 80$, and $\beta_{\tau 2} = -80$, respectively.

Simulations and experiment results

To show the effectiveness of the proposed torque control scheme, the performance is compared with that of the conventional control scheme. The conventional control scheme consists with the PI synchronous current controller with back-EMF compensation and PI controller for outer loop controller, i.e. speed controller. The proposed control scheme is also using PI speed controller which has gains same as these of PI controller of the conventional scheme. Two control schemes are designed to have same performance at same condition. Fig. 5 shows the speed responses of two control schemes for various load conditions. As the load torque is increased, it is shown that the dynamic performance is degraded in the conventional control scheme. However, in the proposed scheme, the nearly same results can be obtained. Fig. 6 shows the responses of the proposed accelerating torque estimator at $\tau_l = 0.34[Nm]$. Fig. 6(a) show the estimated developed torque. Fig. 6(b) and (c) show the estimated load torque and accelerating torque, respectively. Fig. 7 shows the speed responses under the moment of inertia variation. In the conventional scheme, the speed dynamic is degraded as the moment of inertia is increased. But, in the proposed control scheme, the moment of inertia

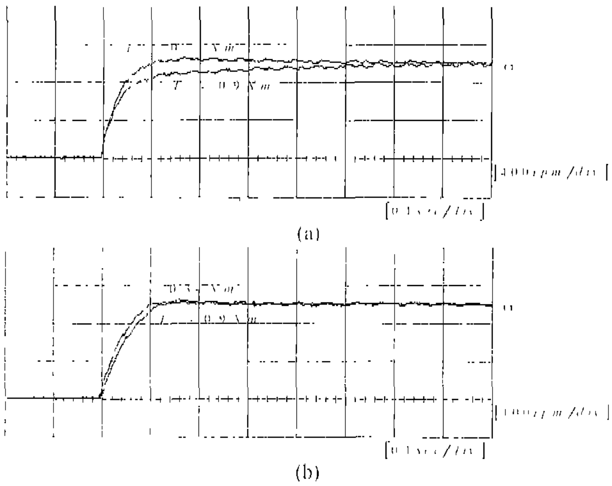


Fig. 5 Comparison of speed responses under load torque variation
(a) With PI synchronous current controller
(b) With proposed torque controller

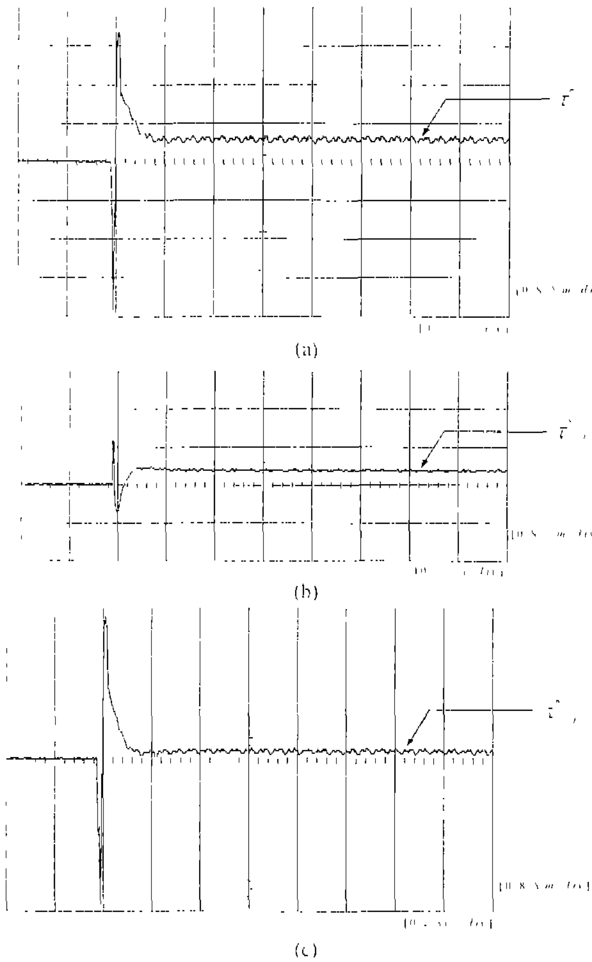


Fig. 6 Responses of proposed accelerating torque estimator
(a) Estimated developed torque
(b) Estimated load torque
(c) Estimated accelerating torque

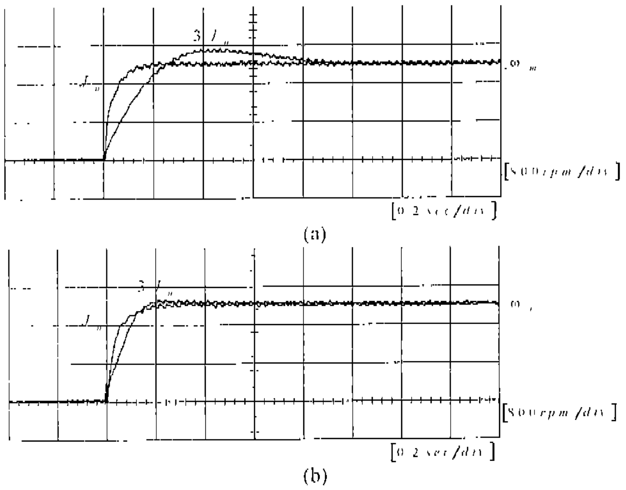


Fig. 7 Comparison of speed responses under inertia variation
(a) With PI synchronous current controller
(b) With proposed torque controller

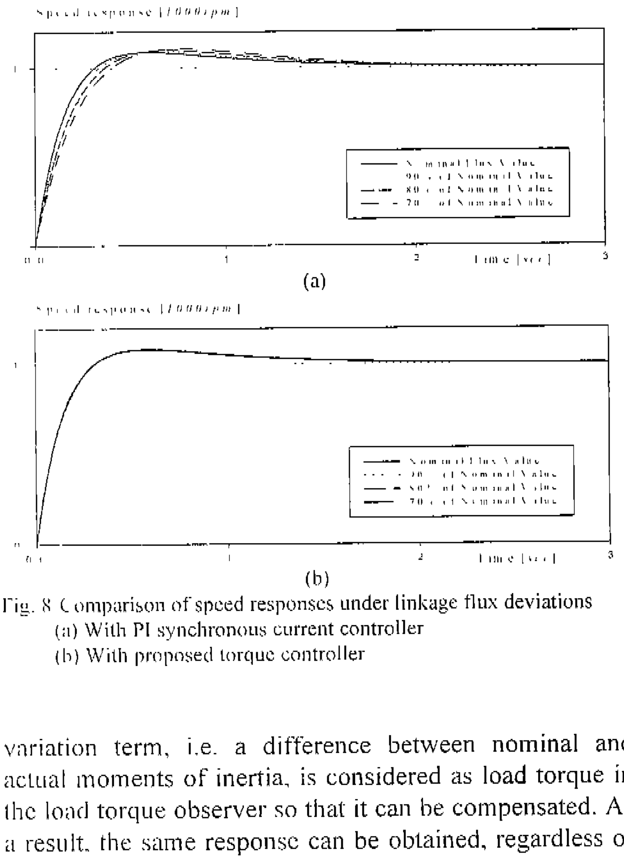


Fig. 8 Comparison of speed responses under linkage flux deviations
(a) With PI synchronous current controller
(b) With proposed torque controller

variation term, i.e. a difference between nominal and actual moments of inertia, is considered as load torque in the load torque observer so that it can be compensated. As a result, the same response can be obtained, regardless of the variation of the moment of inertia.

Fig. 8 shows the speed responses under the linkage flux deviation. This simulation is carried out under assuming that the linkage flux is varied to 90%, 80%, and 70% of its nominal value. In the response of the conventional scheme, it is observed that the overshoot is larger and the settling time is longer as the linkage flux deviation is larger, while it is regarded that the responses of the proposed scheme are independent of the linkage flux deviation. If the linkage flux is varied to smaller than its nominal value, the torque constant is also varied to smaller. In the conventional scheme, it makes the current command of the current controller smaller and the developed torque smaller than that of nominal value case. So the slower response is obtained. However, in the proposed scheme, the linkage flux is estimated before the developed torque is calculated. So nearly exact developed torque can be obtained.

6. CONCLUSIONS

An accelerating torque feedback control scheme is proposed for the robust control of the PM synchronous motor. The accelerating torque is calculated by the developed torque and load torque which are estimated by using the developed torque estimator using MRAS technique and the adaptive load torque observer, respectively. And this torque is controlled by using IVSC technique. The computer simulations and experiments are carried out for the DSP-based high torque PM synchronous motor control system and the results well demonstrate that the proposed scheme provides a robust control performance against the load torque and inertia variations. It is expected that the proposed scheme is applied to the high performance applications of the PM synchronous motor.

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