Adaptive Input-Output Linearization Technique of Interior Permanent Magnet Synchronous Motor with Specified Output Dynamic Performance

Kyeong-Hwa Kim, In-Cheol Baik, Gun-Woo Moon, Dae-Sik Lee, and Myung-Joong Youn

Abstract

An adaptive input-output linearization technique of an interior permanent magnet synchronous motor with a specified output dynamic performance is proposed. The adaptive parameter estimation is achieved by a model reference adaptive technique where the stator resistance and the magnitude of flux linkage can be estimated with the current dynamic model and state observer. Using these estimated parameters, the linearizing control inputs are calculated. With these control inputs, the input-output linearization is performed and the load torque is estimated. The adaptation laws are derived by the Popov's hyperstability theory and the positivity concept. The robustness and the output dynamic performance of the proposed control scheme are verified through the computer simulations.

I. Introduction

Permanent magnet (PM) synchronous motors have been gradually replacing DC motors in a wide range of drive applications such as machine tools and industrial robots. The advantage of using a PM synchronous motor is that many drawbacks caused by the brushes and commutators of a DC motor can be eliminated. Furthermore, the PM synchronous motor has the high power density, large torque to inertia ratio, and high efficiency as compared with a DC motor having the same output rating[1]-[2]. An interior PM (IPM) synchronous motor is a special type of a PM synchronous motor. In an IPM synchronous motor, the magnets are buried inside the rotor so that the mechanically robust construction is obtained which can be used for high speed applications. However, since the q-axis stator inductance is larger than the d-axis stator inductance in the IPM synchronous motor, an additional nonlinear salient torque is produced in the developed torque. Therefore, to be met with the specified output dynamic performance under the existence of nonlinearity such as the salient torque and inherent coupling, a careful controller design must be required.

In the most servo controller design, the electrical dynamics are often neglected because the electrical subsystem dynamics are inherently faster than the associated mechanical subsystem dynamics. An example of this control simplification is the common assumption that the current is a control input for the PM synchronous motor drive with the high gain current feedback. This assumption yields an acceptable performance for many low performance drive applications due to the large time scale seperation between the electrical and mechanical dynamics[3]. But, robotics, machine tools, and direct drive motors are characterized by a smaller degree of dynamic time scale separation[3]. For these high performance drive applications, a conventional controller (inner-loop current control plus outer-loop speed control) often fails to perform satisfactorily. In recent years, state feedback linearizations and input-output decoupling techniques have been applied to the control of the nonlinear plants such as the robot manipulators, induction motors, and PM synchronous motors in order to specify the output dynamic performance based on the linear design techniques [4]-[9]. The basic idea is to first transform a nonlinear

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system into a linear one by a nonlinear state feedback, and then use the well-known linear design techniques to complete the controller design[4]. In these schemes, the nonlinear terms can be effectively cancelled out, and the output error dynamics can be specified to the linear-based model. These techniques, however, require the full knowledge of system parameters and load conditions with the sufficient accuracy. When the controller parameters are mismatched with the real system parameters, a satisfactory output performance cannot be obtained. In order to guarantee the robust response under the parameter variation, the controller parameters must be adaptively changed with the variations of the plant parameters. In reference [8], an input-output linearization technique has been applied to a PM synchronous motor using the speed and d-axis current as the outputs. In this scheme, an integral control has been introduced to improve the robustness with respect to an inaccurate speed measurement. However, other motor parameter variations have not been considered. In addition, the integral control can not give a good improvement of the transient response under the parameter mismatch. In reference [9], an adaptive input-output linearizing controller of an induction motor has been designed using the speed and squared flux as the outputs, and the load torque and rotor resistance have been estimated by choosing an appropriate Lyapunov function. In reference [10], an adaptive load torque observer against the variation of the flux linkage of a PM synchronous motor has been designed by the gradient method. However, this scheme has been designed based on the conventional inner-loop current control and outer-loop position control. In addition, the gradient method is based on the local parametric optimization theory, and requires two supplementary assumptions, that is, the initial values of the estimated parameters must be in the neighborhood of the true parameters and the speed of adaptation must be low[11].

In this paper, an adaptive input-output linearization technique of an IPM synchronous motor with a specified output dynamic performance is proposed. The adaptive parameter estimation is achieved by a model reference adaptive system (MRAS) technique and the adaptive laws are derived by the hyperstability theory and positivity concept. The flux linkage and stator resistance are estimated by an MRAS technique with the current dynamic model and state observer. Using these estimated parameters, the linearizing control inputs of the IPM synchronous motor are calculated. The resultant system becomes a linear and decoupled one except for the unknown load torque disturbance term. To get a perfect input-output linearized system, the load torque disturbance will be estimated. As a result, a perfect input-output linearized motor model which has no nonlinearity and coupling can be obtained. A linear-based specified output response will be obtained as soon as the estimated parameters converge to the true parameter values.

II. Input-Output Linearization with Load Torque Estimation

1. Modeling of IPM synchronous motor

The IPM synchronous motor considered in this paper consists of a permanent magnet rotor and three phase stator windings which is sinusoidally distributed and displaced by 120°. Fig. 1 shows the typical IPM synchronous motor configuration. With the geometry as in Fig. 1, the magnet thicknesses appear as large series air gaps in the d-axis magnetic flux paths since the incremental permeability of ceramic and rare-earth magnet materials is nearly that of free space. Since the q-axis magnetic flux can pass through the steel pole pieces without crossing the magnet air gaps, the stator phase inductance is noticeably higher with the q-axis rotor orientation. The voltage equations of an IPM synchronous motor in the synchronous reference frame are described as follows:

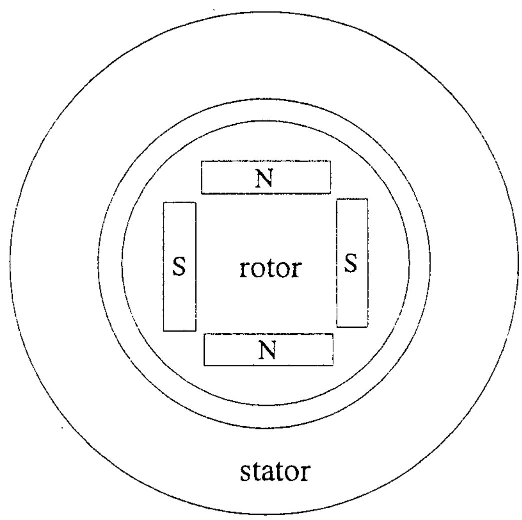


Fig. 1. Typical IPM synchronous motor configuration.

$$v_{qs} = R_s i_{qs} + L_q \dot{i}_{qs} + L_d \omega_r I_{ds} + \lambda_m \omega_r \quad (1)$$
$$v_{ds} = R_s i_{ds} + L_d \dot{i}_{ds} + L_d \dot{i}_{ds} - L_q \omega_r i_{qs}$$

where, R_s is the stator resistance, L_q and L_d are the q-axis and d-axis stator inductances, respectively, ω_r is the electrical rotor angular velocity, and λ_m is the flux linkage established by the permanent magnet. The torque developed by the machine is expressed as

$$T_e = \frac{3}{2} P\{\lambda_m i_{qs} + (L_d - L_q) i_{qs} i_{ds}\} \quad (2)$$

so that the speed dynamics becomes

$$\dot{\omega}_r = \frac{3}{2}\frac{P^2}{J}\{\lambda_m i_{qs} + (L_d - L_q) i_{qs} i_{ds}\} - \frac{P}{J} T_L \quad (3)$$

where J is the moment of inertia of the rotor and load, P is the number of pole pairs, and T_L is the load torque. Using ω_r, i_{qs}, and i_{ds} as the state variables, the state equation of an IPM synchronous motor can be expressed as follows:

$$\dot{x} = f(x) + g_1 v_{qs} + g_2 v_{ds} \tag{4}$$

where $x = [\omega_r \ i_{qs} \ i_{ds}]^T$, $g_1 = \left(0 \ \frac{1}{L_q} \ 0\right)^T$, $g_2 = \left(0 \ 0 \ \frac{1}{L_d}\right)^T$

$$f(x) = \begin{pmatrix} \frac{3}{2}\frac{P^2}{J}\{(L_d - L_q) i_{qs} i_{ds} + \lambda_m i_{qs}\} - \frac{P}{J} T_L \\ -\frac{R_s}{L_q} i_{qs} - \frac{L_d}{L-q}\omega_r i_{ds} - \frac{\lambda_m}{L_q}\omega_r \\ -\frac{R_s}{L_d} i_{ds} + \frac{L_q}{L_d}\omega_r i_{qs} \end{pmatrix}.$$

Clearly, the nonlinearity of an IPM synchronous motor comes from the inherent cross-coupling and the salient component of the developed torque.

2. *Input-output linearization*

To linearize the nonlinear model (4), the controlled variable is differentiated with respect to time until the input appears. This can be easily done by introducing the Lie derivative of a state function $h(x): R^n \to R$ along a vector field $f(x) = (f_1, \cdots, f_n)$ as follows:

$$L_f h = \nabla h \cdot f = \sum_{i=1}^{n} \frac{\partial h}{\partial x_i} f_i(x). \tag{5}$$

Iteratively, it is defined that $L_f^i h = L_f(L_f^{(i-1)} h)$. When all the nominal motor parameters are to be known and the unknown load torque disturbance exists, (4) can be expressed using the estimated load torque $\hat{T}_L$ as follows:

$$\dot{x} = f(x) + g_i v_{qs} + g_2 v_{ds} + d \triangle T_L \tag{6}$$

where $\triangle T_L = T_L - \hat{T}_L$, $d = \left(-\frac{P}{J} \ 0 \ 0\right)^T$

$$f(x) = \begin{pmatrix} \frac{3}{2}\frac{P^2}{J}\{(L_d - L_q) i_{qs} i_{ds} + \lambda_m i_{qs}\} - \frac{P}{J} \hat{T}_L \\ -\frac{R_s}{L_q} i_{qs} - \frac{L_d}{L_q}\omega_r i_{ds} - \frac{\lambda_m}{L_q}\omega_r \\ -\frac{R_s}{L_d} i_{ds} + \frac{L_q}{L_d}\omega_r i_{qs} \end{pmatrix}.$$

In order to avoid any zero dynamics and to get a total input-output linearization, ω_r and i_{ds} are chosen as the outputs[8]. The objectives of the control are to maintain the speed and d-axis current to their reference values with the specified behavior on the linear-based model. Define the state variables as follows:

$$\begin{aligned} y_1 &= h_1(x) = \omega_r \\ y_2 &= L_f h_1(x) = \frac{3}{2}\frac{P^2}{J}\{(L_d - L_q) i_{qs} i_{ds} + \lambda_m i_{qs}\} - \frac{P}{J}\hat{T}_L \\ y_3 &= h_2(x) = i_{ds} \end{aligned} \tag{7}$$

which is the speed, acceleration determined from the estimated load torque, and d-axis current. The dynamic equation of new state variables is described as follows:

$$\begin{aligned} \dot{y}_1 &= y_2 + L_d h_1 \cdot \triangle T_L \\ \dot{y}_2 &= L_f^2 h_1 + L_{g1} L_f h_1 \cdot v_{qs} + L_{g2} L_f h_1 \cdot v_{ds} + \frac{d}{dt}\hat{T}_L \cdot L_d h_1 \\ \dot{y}_3 &= L_f h_2 + L_{g2} h_2 \cdot v_{ds} \end{aligned} \tag{8}$$

where $L_d h_1 = -\frac{P}{J}$

$$L_{g1} L_f h_1 = \frac{3}{2}\frac{P^2}{JL_1}\{(L_d - L_q) i_{ds} + \lambda_m\}$$

$$L_{g2} L_f h_1 = \frac{3}{2}\frac{P^2}{JL_d}(L_d - L_q) i_{qs}$$

$$L_{g2} h_2 = \frac{1}{L_d}$$

$$L_f h_2 = -\frac{R_s}{L_d} i_{ds} + \frac{L_q}{L_d}\omega_r i_{qs}$$

$$\begin{aligned} L_f^2 h_1 &= \frac{3}{2} P^2 \frac{\lambda_m}{J}\left\{-\frac{R_s}{L_q} i_{qs} - \frac{L_d}{L_q}\omega_r i_{ds} - \frac{\lambda_m}{L_q}\omega_r\right\} \\ &+ \frac{3}{2} P^2 \frac{L_d - L_q}{J}\left\{-\frac{R_s}{L_q} i_{qs} i_{ds} - \frac{L_d}{L_q}\omega_r i_{ds}^2 - \frac{\lambda_m}{L_q}\omega_r i_{ds} - \frac{R_s}{L_d} i_{qs} i_{ds} + \frac{L_q}{L_d}\omega_r i_{qs}^2\right\} \end{aligned}$$

To estimate the nonlinear states in (8), the required control inputs which are applied to the inputs of motor v_{qs} and v_{ds} are calculated as follows:

$$\begin{pmatrix} v_{qs} \\ v_{ds} \end{pmatrix} = D(x)^{-1}\begin{pmatrix} -L_f^2 h_1 - \frac{d}{dt}\hat{T}_L \cdot L_d h_1 + u_1 \\ -L_f h_2 + u_2 \end{pmatrix} \tag{9}$$

where u_1 and u_2 are the equivalent control inputs, and the decoupling matrix is defined as

$$D(x) = \begin{pmatrix} L_{g1} L_f h_1 & L_{g2} L_f h_1 \\ 0 & L_{g2} h_2 \end{pmatrix} \tag{10}$$

If the d-axis current reaches the demagnetization current, the decoupling matrix $D(x)$ becomes singular[8]. The singular point of $D(x)$ can be obtained using $\det D(x)=0$ as follows:

$$(L_d - L_q) i_{ds,dem} + \lambda_m = 0 \quad \text{or} \quad i_{ds,dem} = \frac{\lambda_m}{(L_d - L_q)}. \tag{11}$$

Since the demagnetization current is often high value, it does not occur in practice. With the assumption that the demagnetization current does not flow in the motor terminal, the linearizing control inputs (9) can be always calculated. When the linearizing control inputs (9) are applied, the nonlinear motor model (8) becomes

$$\begin{aligned} \dot{y}_1 &= y_2 + L_d h_1 \cdot \triangle T_L \\ \dot{y}_2 &= u_1 \\ \dot{y}_3 &= u_2 \end{aligned} \tag{12}$$

which does not include the cross-coupled nonlinearity and nonlinear salient torque component. Only the existing nonlinearity comes from an uncertain load torque disturbance. Fig. 2 shows the calculation of the linearizing control inputs and the resultant linearized output dynamics.

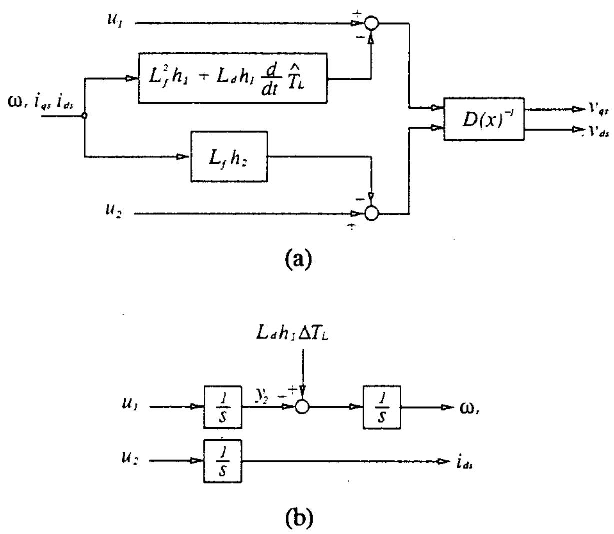


Fig. 2. Calculation of the linearizing control inputs and block diagram of the resultant linearized system. (a) Calculation of the linearizing control inputs (b) Block diagram of the resultant linearized system

3. *Load torque estimation*

In the previous section, an input-output linearized and decoupled model of an IPM synchronous motor was obtained. Using the resultant model, output responses can be specified. However, an uncertain load torque disturbance term still exists in (12). To get a specified speed response under the existence of an uncertain load torque disturbance, the load torque disturbance will be estimated. Choose the equivalent control inputs as follows:

$$u_1 = -k_{\omega 1}(y_1 - \omega_r^*) - k_{\omega 2}(y_2 - \dot{\omega}_r^*) + \ddot{\omega}_r^*$$
$$u_2 = -k_{id}(y_3 - i_{ds}^*) + \dot{i}_{ds}^* \quad (13)$$

where ω_r^* and i_{ds}^* are the speed and d-axis current references. These equivalent control inputs give the second order speed error dynamics and the first order d-axis current error dynamics. The desired poles can be easily chosen by means of the gains $k_{\omega 1}$, $k_{\omega 2}$, and k_{id}. To estimate the load torque disturbance with the MRAS technique, a reference model is chosen as follows:

$$\dot{y}_{1M} = y_{2M}$$
$$\dot{y}_{2M} = -k_{\omega 1}(y_{1M} - \omega_r^*) - k_{\omega 2}(y_{2M} - \dot{\omega}_r^*) + \ddot{\omega}_r^* \quad (14)$$
$$\dot{y}_{3M} = -k_{id}(y_{3M} - i_{ds}^*) + \dot{i}_{ds}^*$$

where y_{1M}, y_{2M}, and y_{3M} denote the states of the reference model. By subtracting (14) from (12), the error dynamic equation can be obtained as follows:

$$\dot{\tilde{y}} = A_M \tilde{y} - W_1 \quad (15)$$

where $\tilde{y} = y - y_M$, $W_1 = -B \cdot \triangle T_L$

$$A_M = \begin{pmatrix} 0 & 1 & 0 \\ -k_{\omega 1} & -k_{\omega 2} & 0 \\ 0 & 0 & -k_{id} \end{pmatrix}, \quad B = (L_d h_1 \ 0 \ 0)^T.$$

Define the adaptation mechanism as follows:

$$v = D\tilde{y} \quad (16)$$

$$\hat{T}_L(v,t) = \int_0^t \Psi_1(v,t)d\tau + \Psi_2(v,t) + \hat{T}_L(0) \quad (17)$$

where D is a linear compensator, V is the output of a linear compensator which processes the state error, and Ψ_1 and Ψ_2 are the nonlinear adaptation mechanisms for the estimation of the load torque. The design problem to obtain the asymptotic adaptation is as follows:

1. Determine D, Ψ_1, and Ψ_2 such that $\lim_{t\to\infty} \tilde{y}(t) = 0$ for any initial conditions $y_M(0)$, $y(0)$, and $T_l - \hat{T}_L(0)$.
2. Find the supplementary conditions which lead to $\lim_{t\to\infty} \hat{T}_L(v,t) = T_L$.

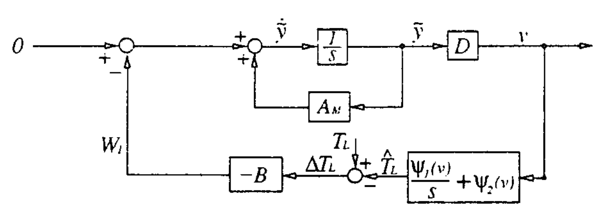


Fig. 3. Structure of MRAS for the load torque estimation.

Based on (15), (16), and (17), an MRAS system for the load torque estimation can be constructed as shown in Fig. 3, which consists of a linear time invariant forward block and a nonlinear feedback block. This system is hyperstable if the forward transfer function matrix is strictly positive real and the input-output inner product of the nonlinear feedback block satisfies the Popov's integral inequality as follows[11]:

$$\int_0^{t_1} v^T w_1 dt = \int_0^{t_1} -v^T B \cdot \triangle T_L dt \geq -\gamma_o^2 \quad \textbf{for all} \ t_1 \geq 0 \quad (18)$$

where γ_0^2 is a finite positive constant. It is shown that for a given matrix A_M, the strictly positive real transfer function matrix

$$H_1(s) = D(sI - A_M)^{-1} \quad (19)$$

can be obtained by choosing a linear compensator $D = P$ in which P is a symmetric positive definite matrix and the solution of the following Lyapunov equation:

$$A_M^T P + P A_M = -Q \tag{20}$$

where Q is a symmetric positive definite matrix[11]. If and only if the reference model is asymptotically stable, which is generally the case, (20) always has a positive definite matrix solution P. In order to derive the adaptation mechanism for the load torque disturbance, (18) can be expressed using (17) as follows:

$$\int_o^{t_1} -v^T B \cdot \left(T_L - \int_0^t \Psi_1(v,t) d\tau - \Psi_2(v,t) - \hat{T}_L(0) \right) dt \geq -\gamma_0^2. \tag{21}$$

By solving the inequality (21), the load torque disturbance can be estimated as follows:

$$\hat{T}_L(v,t) = \left(k_{PT} + \frac{k_{IT}}{s} \right) \cdot (v^T B) = \left(k_{PT} + \frac{k_{IT}}{s} \right) \cdot (v_1 \cdot L_d h_1) \tag{22}$$

where $v = [v_1\ v_2\ v_3]^T$, k_{PT} and k_{IT} are the PI gains for the load torque estimation, respectively. With this estimated load torque, (12) becomes a linear state equation with no disturbance term. When the estimated load torque converges to the real value, the equivalent control inputs (13) applied to (12) make the output error dynamics as follows:

$$\dot{e}_\omega + k_{\omega 2} \dot{e}_\omega k_{w1} e_w = 0 \tag{23}$$

$$\dot{e}_{id} + k_{id} e_{id} = 0$$

where $e_\omega = \omega_4 - \omega_r^*$ and $e_{id} = i_{ds} - i_{ds}^*$.

III. Estimation of Flux Linkage and Stator Resistance

When the linearizing control inputs (9) are calculated to get a linearized model (12) in section II, all the nominal motor parameters were assumed to be known. However, there exists a certain variation of the parameters such as the flux linkage and stator resistance. Under the mismatch of these parameters, a linearized and decoupled model cannot be obtained. To overcome this problem, the flux linkage and stator resistance will be estimated with the current dynamic model and state observer. The dynamic model of the q-axis and d-axis currents is described as follows:

$$\dot{i}_s = A i_s + B v_s + f \tag{24}$$

where $i_s = [i_{qs}\ i_{ds}]^T$, $v_s = [v_{qs}\ v_{ds}]^T$

$$A = \begin{pmatrix} -\frac{R_s}{L_q} & -\frac{L_d}{L_q}\omega_r \\ \frac{L_q}{L_d}\omega_r & -\frac{R_s}{L_d} \end{pmatrix},\quad B = \begin{pmatrix} \frac{1}{L_q} & 0 \\ 0 & \frac{1}{L_d} \end{pmatrix},\quad f = \begin{pmatrix} -\frac{\lambda_m}{L_q}\omega_r \\ 0 \end{pmatrix}.$$

The full state observer can be expressed as follows:

$$\dot{\hat{i}}_s = \hat{A}\,\hat{i}_s + B v_s + \hat{f} + G(\hat{i}_s - i_s) \tag{25}$$

where '∧' denotes the estimated quantities and G is a gain matrix. From (24) and (25), the state error equation of the state observer can be obtained as follows:

$$\dot{e} = (A + G)e - W_2 \tag{26}$$

where $e = i_s\ \hat{i}_s$ and $W_2 = -\triangle A\ \hat{i}_s - \triangle f$. The error matrices $\triangle A$ and $\triangle f$ are caused by the variations of the flux linkage and stator resistance, respectively, and can be written as

$$\triangle A = A - \hat{A} = \begin{pmatrix} -\frac{\triangle R_s}{L_q} & 0 \\ 0 & -\frac{\triangle R_s}{L_d} \end{pmatrix} \tag{27}$$

$$\triangle f = f - \hat{f} = \begin{pmatrix} -\frac{\omega_r}{L_q} \\ 0 \end{pmatrix} \triangle\lambda_m \tag{28}$$

where $\triangle R_s = R_s - \hat{R}_s$ and $\triangle\lambda_m = \lambda_m - \hat{\lambda}_m$. Using the similar procedures in section II, an MRAS system for the estimation of the flux linkage and stator resistance is constructed as shown in Fig. 4. By appropriately choosing the gain matrix G, the forward transfer function matrix

$$H_2(s) = (sI - (A+G))^{-1} \tag{29}$$

can be strictly positive real without any linear compensator D in the forward path of Fig. 4, and the Popovs integral inequality $\int_0^{t_1} e^T W_2 dt \geq -\gamma_o^2$ can be used to derive the adaptation mechanisms for the flux linkage and stator resistance. By solving this type of inequality, the flux linkage and stator resistance can be estimated as follows:

$$\hat{\lambda}_m = -\left(k_{P\lambda} + \frac{k_{I\lambda}}{S} \right) \cdot e_{qs}\omega_r \tag{30}$$

$$\hat{R}_s = -\left(k_{PR} + \frac{k_{IR}}{s} \right) \cdot \left(\frac{1}{L_q} e_{qs}\ \hat{i}_{qs} + \frac{1}{L_d} e_{ds}\ \hat{i}_{ds} \right) \tag{31}$$

where $k_{P\lambda}$ and $k_{I\lambda}$ are the PI gains for the flux linkage estimation, respectively, and k_{PR} and k_{IR} are the PI gains for the stator resistance estimation, respectively. Using these estimated parameters (22), (30), and (31), the linearizing control inputs (9) are calculated. In order to assign two poles of the state observer at the specified locations on the complex plane and to be satisfied with the strictly positive real condition without any linear compensator, the gain matrix can be selected as follows:

$$G = \begin{pmatrix} -g_1 & -g_2 \\ g_3 & -g_4 \end{pmatrix} \tag{32}$$

where, $g_1 = k\frac{\hat{R}_s}{L_d} - \frac{\hat{R}_s}{L_q}$, $g_2 = \left(k - \frac{L_d}{L_q}\right)\omega_r$,

$g_3 = \left(k - \frac{L_q}{L_d}\right)\omega_r$, and $g_4 = (k-1)\frac{\hat{R}_s}{L_d}$.

With this type of observer gain G, the stable error dynamics which means the negative diagonal term of the matrix$(A+G)$ guarantees the satisfaction of strictly positive real condition. When the estimated parameters converge to their real values, the closed loop observer error dynamics can be represented as follows:

$$\dot{e} = (A+G)e = \begin{pmatrix} -k\frac{R_s}{L_d} & -k\omega_r \\ k\omega_r & -k\frac{R_s}{L_d} \end{pmatrix} \cdot e. \quad (33)$$

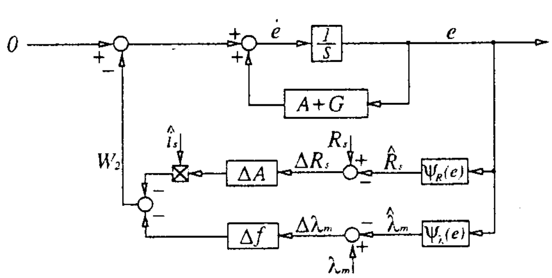


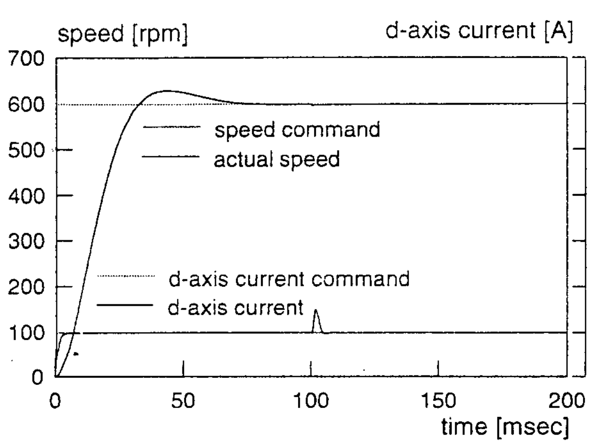
Fig. 4. Structure of MRAS for the flux linkage and stator resistance estimation.

IV. Simulation Results

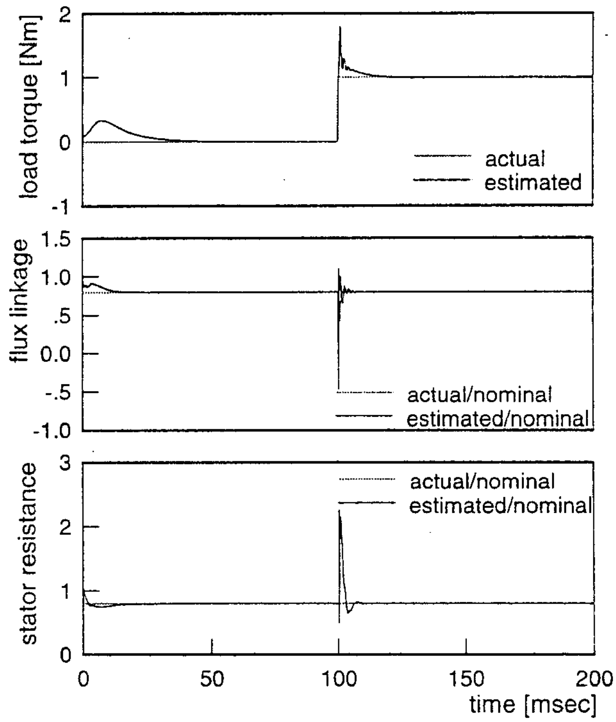
The nominal parameters of an IPM synchronous motor used for the simulations are as follows:

$R_{s,nom} = 1.07[\Omega]$, $L_{q,nom} = 4.6[mH]$, $L_{d,nom} = 2.3[mH]$

$\lambda_{m,nom} = 0.2[Wb]$, and $J_{nom} = 0.001[Nmsec^2]$.



(a)



(b)

Fig. 5. Output responses and parameter estimations when $\lambda_m = 0.8\lambda_{m,nom}$ and $R_x = 0.8R_{s,nom}$ at t=0 and $T_L = 1[Nm]$ at t=100[msec]
(a)output responses (speed and d-axis current)
(b) parameter estimations

Fig. 5 shows the output responses and the parameter estimations for the proposed control scheme when R_s and λ_m are varied to 80 [%] of their nominal values at t=0, respectively, and the step load torque of 1 [Nm] is applied at t=100 [msec]. This step load torque value corresponds to half the rated load. The speed and d-axis current commands are given as 600 [rpm] and 1 [A], respectively, and the sampling period is 0.1 [msec]. The selected gains of the reference model for the specified output performance are set as follows: $k_{\omega 1} = 10000$, $k_{\omega 2} = 140$, and $k_{id} = 1000$. Fig. 5(a) shows the speed and d-axis current responses. As can be shown in this figure, the specified performance of the speed and d-axis current response can be obtained in the presense of the parameter variation and the step load torque disturbance. Fig. 5(b) shows the estimations of T_L, λ_m, and R_s. The adaptive PI gains for the parameter estimation are selected as follows: $k_{PT} = 3\times 10^{-5}$, $k_{IT} = 4\times 10^{-4}$, $k_{PR} = 0$, $k_{IR} = 6.9$, $k_{P\lambda} = 0$, $k_{I\lambda} = 1000/|\omega_r|$, and $Q = I_3$. It is shown that all the estimated parameters converge to their true values within finite steps. Fig. 6 shows the reference model y_M and adjustable model for the speed and d-axis current. Fig. 7 shows the real motor currents and the estimated states of current observer. Fig. 8 shows the output responses and the parameter estimations when the moment of

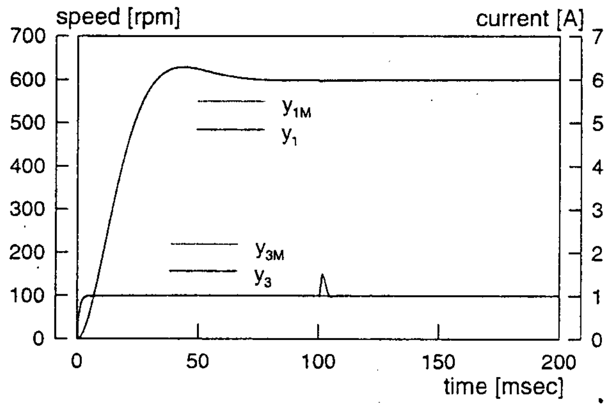


Fig. 6. Reference and adjustable models for the speed and d-axis current.

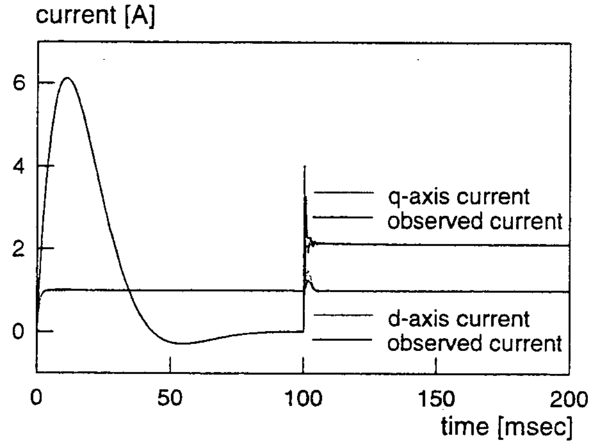
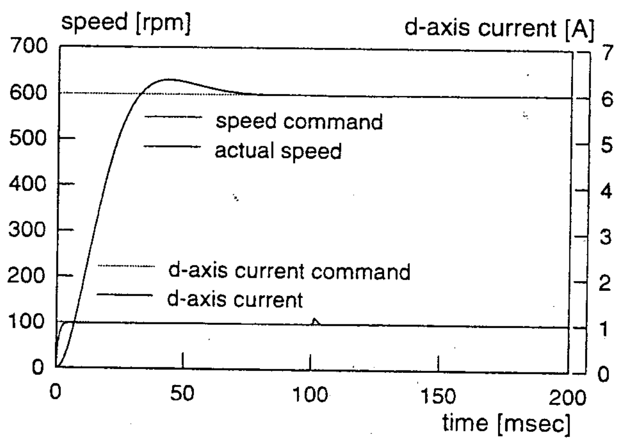
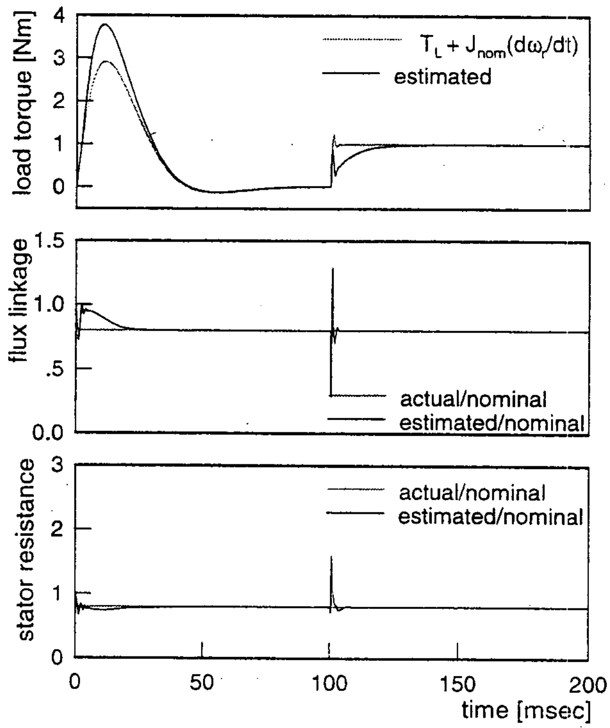


Fig. 7. Real and observed currents for the q- and d-axes.



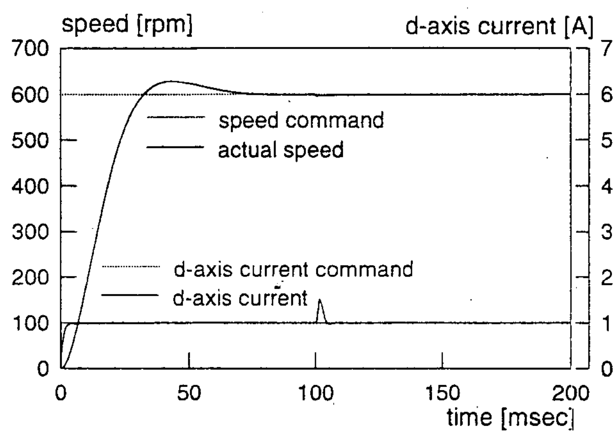
(a)



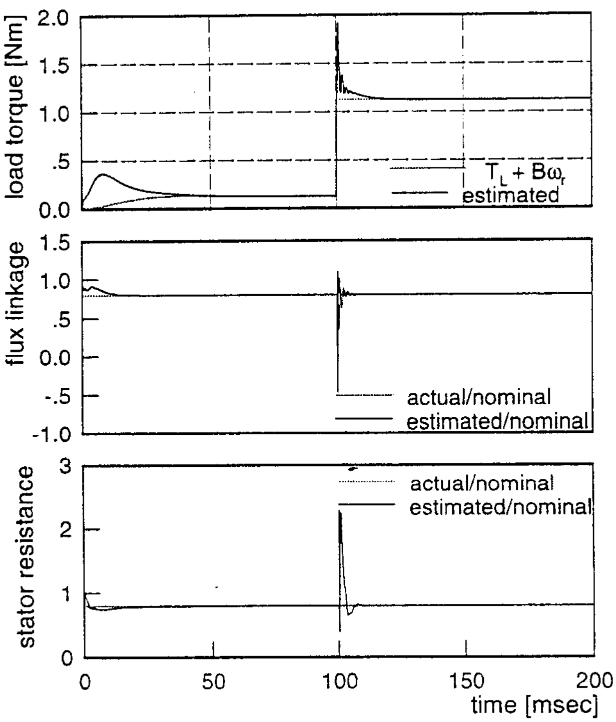
(b)

Fig. 8. Output responses and parameter estimations when $\lambda_m = 0.8\lambda_{m,nom}$ and $R_s = 0.8R_{s,nom}$ at t=0 and $T_L = 1[Nm]$ at t=100[msec] with J=2J_{nom} at t=0 (a) Output responses (speed and d-axis current) (b) Parameter estimations

inertia is varied to double the nominal value under the same condition of Fig. 5. In this case, the estimated load torque converges to $T_L + (J - J_{nom})\frac{d\omega_r}{dt}$. The estimation error in the load torque is observed at t=15 [msec] because $\hat{\lambda}_m$ and $\hat{R}_s$ do not yet converge to their true values. Fig. 9



(a)



(b)

Fig. 9. Output responses and parameter estimations when $\lambda_m = 0.8\lambda_{m,nom}$ and $R_s = 0.8R_{s,nom}$ at t=0 and $T_L = 1[Nm]$ at t=100[msec] with B=0.002[NMSec] at t=0 (a) Output responses (speed and d-axis current) (b) Parameter estimations

shows the output responses and the parameter estimations when the unmodeled viscous friction term B=0.002 [Nmsec] is added under the same condition of Fig. 5. It can be shown that the estimated load torque converges to $T_L + B\omega_r$. This comes from the fact that this load torque estimator has the characteristics of the disturbance observer. As can be seen in Fig. 9(a), a high performance specified output response can be obtained in spite of unmodeled disturbance term.

V. Conclusions

In this paper, an adaptive input-output linearization technique of an IPM synchronous motor with a specified output dynamic performance has been proposed and shown its robust performance against the motor parameter variations, load torque disturbances, and unmodeled disturbance term. The adaptive parameter estimation is achieved by an MRAS technique where the stator resistance and the magnitude of flux linkage can be estimated with the current dynamic model and state observer. Using these estimated parameters, an input-output feedback linearization technique is performed, and the load torque is estimated. As a result, a perfectly linearized and decoupled motor model and a specified output dynamic performance can be obtained.

Acknowledgement

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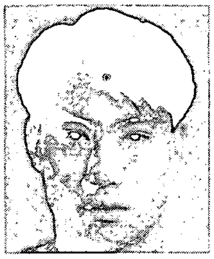

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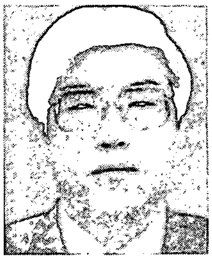

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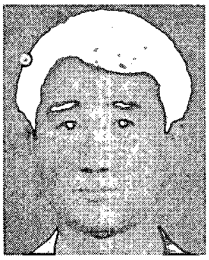

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