Paper Review Report

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Abstract: A model reference adaptive control (MRAC)-based current control scheme of a PM synchronous motor with an improved servo performance is presented. Although the predictive current control is known to give ideal transient and steady-state responses among various PWM inverter-fed current control schemes for a PM synchronous motor, its steady-state response may be degraded under the motor parameter variations. To overcome such a limitation, the disturbances caused by the parameter variations will be estimated using an MRAC technique and compensated by a feedforward manner. Thus, the steady-state control performance can be effectively improved, while retaining its good dynamic performance. The proposed control scheme does not require the measurement of the phase voltage unlike the conventional disturbance estimation scheme using observer. This can be an effective way considering the phase voltage contains much harmonics as well as noise. The asymptotic stability of the overall system is proved and the adaptation laws are derived by the Lyapunov stability theory. The proposed scheme is implemented using DSP TMS320C31 and the effectiveness is verified through the comparative simulations and experiments.

This paper propose to control PM synchronous motor by a model reference adaptive control (MRAC)-based current control scheme for improve a servo drive performance. In this study, author begin form analysis problem of the current control schemes for a PWM inverter-fed PM synchronous motor drive. The author founded such problems in driving PM synchronous motor by a PWM inverter-fed, cannot be used in a high-performance drive due to some limitations such as a large current error, irregular PWM inverter operation and phase delay. Which this problems can be solve by the PI decoupling current control in the synchronous reference frame. However, this scheme still have problem about transient response which slow or may be decreased performance as parameter of PM synchronous motor variance such as the phase resistance, the synchronous inductance and the flux linkage. In other method, a predictive control scheme be proposed a dead-beat type controller, the switching of the space vector PWM inverter are determined by calculating the required reference voltage in a discrete domain that forces the motor currents to follow currents references. With this control scheme, system response have a fast transient, a
constant switching frequency and a lower current ripple. This scheme, however, requires the exactly knowledge of machine parameters and operating conditions which very difficult to know during the machine operate, especially back EMS which depending on the flux linkage ($\lambda_m$). This problem can be solve by delayed input voltage feedback scheme that independent of the back EMS but this scheme still require other motor parameter. Also, a multivariable state feedback control with an integrator and a generalized predictive control have been propose to control the induction motor drives. Although this scheme have a good performance and not considered the parameter but the controller design is quite complex. In addition, the author also mentioned the disturbance observer method for estimate disturbance. Which these methods are still required measured the phase voltages of the PWM inverter that much harmonics and noise caused by the switching.

Therefore, the author propose new method, MRAC technique used for estimated the disturbance voltages caused by the parameter variations that be abovementioned drawback of the predictive current control scheme. The author begin form designing the control inputs in the synchronous reference frame for decoupling compensation to prevent decreasing performance of a current control caused from the cross-coupling term ($L_{s0}\omega_r i_{ds}$ and $L_{s0}\omega_r i_{qs}$) and back EMF ($\lambda_m\omega_r$). Now PI control ($G_c(s)$) is employed for control currents Block diagram shown in fig 1.

![Figure 1 Block diagram for the synchronous PI plus decoupling current control and PMSM](image)

However, PI decoupling current regulator still use compensation form the back EMF and cross coupling terms. Under the parameter variations,
disturbances $f_{qs}$ and $f_{ds}$ in (1) (2) are directly influence on back EMF and cross coupling terms as a result to the current control performance.

$$f_{qs} = \Delta R_s i_{qs} + \Delta L_s i_{qs} + \Delta L_s \omega_r i_{ds} + \Delta \lambda_m \omega_r$$

(1)

$$f_{ds} = \Delta R_s i_{ds} + \Delta L_s i_{ds} - \Delta L_s \omega_r i_{qs}$$

(2)

In this article, the author propose estimating the disturbances caused by the parameter variations using the MRAC-base control scheme and by compensating using a feedforward control. The disturbances are estimated by MRAC and added into the voltage references for removed an influence of the parameter variations.

To estimate the parameters using the MRAC technique, an Adjustable model and a reference model shown in fig.2. This reference model represents an ideal dynamic behaviour of current when there are no parameter variations.

Figure 2 Overall block diagram for the proposed adaptive current control scheme

The error between two models will be used to update the estimated disturbance in an adjustable model to reduce the output error.

To derive the adaptation laws using the Lyapunov stability theory and if the adaptation rules for the disturbance estimation are chosen as
\[ \frac{d}{dt} \hat{f}_q = k_2 (-e^T PB_1) \]

and

\[ \frac{d}{dt} \hat{f}_d = k_2 (-e^T PB_2) \]

The derivative of Lyapunov function can be rewritten as

\[ \frac{d}{dt} V = -e^T Q e \leq 0 \]

This proves the asymptotic stability of the adjustable model into the reference model in. With using the proportional adaptation gains to improve the estimation performance during the transient periods, the disturbances can be estimated.

Since the proposed control scheme uses only the control inputs in the predictive current control for the input signal of the reference and adjustable models to estimate the disturbances. It does not require the measurement of the phase voltages often contain much harmonics as well as noise due to the switching.

For proving propose scheme, the author demonstrate simulation and experimental. The overall system consists of a proposed current controller, a PWM inverter, a PM synchronous motor and adaptive algorithm for the disturbance estimation. Form the simulation result for the synchronous PI decoupling current control under the parameter variations shown that the current has relatively a large overshoot and slow transient response. For simulation result of the predictive current control under the same parameter variations shown that it have a faster transient response than PI decoupling current control but still have steady state error. The simulation result for the proposed MRAC-based adaptive current control scheme under the same parameter variations, it is clearly as compared with the other method shows a smaller overshoot and a faster transient response and improved steady-state response similar in case of the predictive current control without parameter variations.

Form experimental result, the author show that proposed control scheme is these errors are effectively removed in current responses. The proposed control scheme have a good estimating performance of the disturbances similar to the simulation result.

Thus, proposed MRAC-based adaptive current control scheme can be effectively improved the steady state control performance, while retaining its good dynamic performance. The proposed control scheme does not require the measurement of the phase voltages unlike the conventional disturbance
estimation scheme using observer. A systematic design approach for a robust current control can be accomplished.

However, performance overall of the servo drive not yet proved in this study because the current control, it be only inner loop of servo drive control. Therefore, proving performance of PM synchronous motor driving, it should have outer loop or speed loop control to shown effectiveness of the servo drive.

Other than, selecting the gains for the PI control, the author don’t shown that use what method to select the gains, hence are not sure that selected gains be most suitable for the PI control.

In the future this study can be extended the study to have several, such as the using optimize method to chosen the adaptive gains of the adaptation rules. Moreover, using this proposed scheme for speed loop control of the servo drive for estimating load torque or external disturbance, or take this proposed scheme to control current in the other type of motor such as DC motor or Induction Motor etc.
Model reference adaptive control-based adaptive current control scheme of a PM synchronous motor with an improved servo performance

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Abstract: A model reference adaptive control (MRAC)-based current control scheme of a PM synchronous motor with an improved servo performance is presented. Although the predictive current control is known to give ideal transient and steady-state responses among various PWM inverter-fed current control schemes for a PM synchronous motor, its steady-state response may be degraded under the motor parameter variations. To overcome such a limitation, the disturbances caused by the parameter variations will be estimated using an MRAC technique and compensated by a feedforward manner. Thus, the steady-state control performance can be effectively improved, while retaining its good dynamic performance. The proposed control scheme does not require the measurement of the phase voltage unlike the conventional disturbance estimation scheme using observer. This can be an effective way considering the phase voltage contains much harmonics as well as noise. The asymptotic stability of the overall system is proved and the adaptation laws are derived by the Lyapunov stability theory. The proposed scheme is implemented using DSP TMS320C31 and the effectiveness is verified through the comparative simulations and experiments.

1 Introduction

Recently, the field-oriented control and high-quality current control schemes have been an essential part of AC motor drive systems to implement a high-performance servo system [1, 2]. The current control schemes for a PWM inverter-fed PM synchronous motor drive can be classified as hysteresis control, ramp comparison control, synchronous frame proportional integral (PI) control and predictive control [3–6]. The basic requirements for the current controller are fast dynamic response during the transient state, zero steady-state error, stable PWM inverter operation and robustness against the variations of machine parameters. Based on this, the hysteresis and ramp comparison controllers cannot be used in a high-performance drive due to some limitations such as a large current error, irregular PWM inverter operation and phase delay [4]. To overcome such problems, the PI decoupling current control in the synchronous reference frame has been proposed. In this control scheme, the controlled variables are DC quantities. By employing the PI control and decoupling inputs for the back EMF and cross-coupling terms, this control gives ideal steady-state control characteristics having zero steady-state error irrespective of operating conditions. However, the transient response is generally slow or may be degraded due to the inexact decoupling input under the parameter variations. On the other hand, in a predictive control scheme, the switching instants of the inverter switches are determined by calculating the required reference voltage in a discrete domain that forces the motor currents to follow their references, yielding a dead-beat type controller. With the space vector PWM technique, this control scheme is known to provide the advantages such as a fast transient response, a constant switching frequency and a lower current
ripple [4, 5]. This scheme, however, requires the full knowledge of machine parameters and operating conditions with the sufficient accuracy and cannot give a satisfactory response under the parameter mismatch. Generally, it is very difficult to exactly obtain the information on the machine parameters since they may vary during operation due to the changes in the temperature, current level and operating frequency as well as machine saturation. In particular, the inaccuracy of the back EMF influences primarily on the control performance.

To overcome such problems, the current control scheme independent of the back EMF variation has been introduced through the estimation of the back EMF using the feedback of the delayed input voltages and currents in a discrete domain. This scheme still, however, requires other motor parameters such as the phase resistance and synchronous inductance as well as the measurement of phase voltages, and also unstable regions may exist under a certain variation of the synchronous inductance [7]. Also, a multivariable state feedback control with an integrator and a generalised predictive control [6] have been reported. Although a good performance can be obtained, the controller design is quite complex and the parameter variations are not considered. Recently, to improve the control performance against the disturbances caused by the motor parameter variations, the current control with a feedforward compensation using observer or time-delayed estimator has been proposed. In [8], though the disturbance is not a state variable, it has been estimated using a disturbance observer under the assumption that its variation is not so fast during each sampling interval. To construct this observer, however, the phase voltages of the PWM inverter need to be measured. In general, it is not easy to measure the phase voltage exactly since it often contains much harmonics and noise caused by the switching. Also, in spite of its verified performance, the stability has not been proved in [9]. Since the disturbances are estimated through a model-based calculation and low-pass filter to reduce the high-frequency noise in stator currents, the estimating performance is severely dependent on the filter bandwidth. In addition, the delayed time step 

In this article, a model reference adaptive control (MRAC)-based adaptive current control scheme of a PM synchronous motor is presented to improve the performance of a servo drive. Although the predictive control provides a fast transient response and a stable performance independent of the operating condition, its steady-state response may be degraded under the motor parameter variations such as the flux linkage, synchronous inductance and phase resistance. This is more serious at high-speed operations since the disturbance is proportional to the product of the operating speed and these parameters. To overcome this drawback, the disturbance voltages caused by the parameter variations will be estimated by an MRAC technique and used for the compensation of the reference voltages by a feedforward manner. The asymptotic stability of the overall system is proved and the adaptation laws are derived by the Lyapunov stability theory. Thus, the steady-state control performance can be significantly improved, while retaining the good transient characteristics of the predictive control. Also, the proposed control scheme does not require the measurement of the phase voltages unlike the conventional observer-based disturbance estimation scheme. Generally, to estimate the parameters using the MRAC technique, an adjustable model and a reference model are required, and the same control inputs applied to each one have to be determined. In the proposed scheme, the MRAC-based parameter estimation is carried out not by using the phase voltages of inverter but by using the transformed control inputs as the input on each model. This can be an effective way considering that the measurement of the phase voltages is not simple because they often contain much harmonics as well as noise. The whole control processing is implemented by the software of DSP TMS320C31 for a PM synchronous motor driven by a three-phase voltage-fed PWM inverter [10, 11] and the effectiveness of the proposed control scheme is verified through the simulations and experiments.

2 Modelling of a PM synchronous motor

A PM synchronous motor consists of permanent magnets mounted on the rotor surface and three-phase stator windings that are sinusoidally distributed and displaced by 120°. The stator voltage equations of a PM synchronous motor in the synchronously rotating reference frame are described as follows [1]

\[
v_p = R_i i_p + L_s \omega_s i_p + \lambda_m \omega_s (1)
\]

\[
v_d = R_i i_d + L_s \omega_s i_d - L_s \omega_s i_p (2)
\]

where \(R_i\) is the phase resistance, \(L_s\) the synchronous inductance, \(\omega_s\) the electrical rotor angular velocity and \(\lambda_m\) the flux linkage established by the permanent magnet. Using the nominal parameters, (1) and (2) can be rewritten as follows

\[
v_p = R_n i_p + L_n i_p + L_n \omega_s i_p + \lambda_m \omega_s + f_{ip} (3)
\]

\[
v_d = R_n i_d + L_n i_d - L_n \omega_s i_p + f_{id} (4)
\]

where \(f_{ip}\) and \(f_{id}\) represent the disturbances caused by the parameter variations and can be expressed as

\[
f_{ip} = \Delta R_s i_p + \Delta L_s i_p + \Delta \lambda_m \omega_s (5)
\]

\[
f_{id} = \Delta R_s i_d + \Delta L_s i_d - \Delta \lambda_m i_p (6)
\]

where \(\Delta R_s = R_n - R_n\) and \(\Delta L_s = L_n - L_n\) and subscript ‘o’ denotes the nominal value. Using \(i_p\) and
\(i_d\) as the state variables in (3) and (4), the state equation of a PM synchronous motor can be expressed as follows

\[
\begin{align*}
\dot{x} &= Ax + Bv_i + d - Bf_i, \\
A &= \begin{pmatrix}
-R_{so}/L_{so} & -L_{so}/\omega_t \\
L_{so}/\omega_t & -R_{so}/L_{so}
\end{pmatrix}, \\
B &= \begin{pmatrix}
1/L_{so} & 0 \\
0 & 1/L_{so}
\end{pmatrix}, \\
d &= \begin{pmatrix}
-L_{so}/\omega_t \omega_t \\
0
\end{pmatrix}
\end{align*}
\]  

where \(x = [i_q, i_d]^T, v_i = [v_{qs}, v_{ds}]^T, f_i = [f_{qs}, f_{ds}]^T\).

### 3 Current control algorithm

In the state equation of a PM synchronous motor in (7), the cross-coupling term and back EMF are represented as functions of the operating speed and current. To prevent a current control performance from degrading caused by these terms, the conventional current controller in the synchronous reference frame often uses the additional control inputs for decoupling compensation as follows

\[
\begin{align*}
\tilde{v}_q &= u_{qs} + L_{so}\omega_t i_q + \lambda_{m}\omega_t \\
\tilde{v}_d &= u_{ds} - L_{so}\omega_t i_d
\end{align*}
\]

where \(\tilde{v}_q\) and \(\tilde{v}_d\) are the reference voltages applied to a PM synchronous motor through the PWM inverter, respectively, and \(u_{qs}\) and \(u_{ds}\) are the control inputs using current errors that are added to the decoupling control. Fig. 1 shows the block diagram for the current controller in (8) and (9) and PM synchronous motor in the synchronous reference frame. In this figure, the left shaded area represents the current controller including the decoupling control. The computed reference voltages are applied to a PM synchronous motor using the PWM technique. For the simplification of analysis, it is reasonable to assume that the reference voltages are applied to the terminals of the motor without any loss and deformation in the PWM inverter, i.e., \(\tilde{v}_q = v_{qs}\) and \(\tilde{v}_d = v_{ds}\). Based on this, the model can be transformed into a new one in terms of new control inputs [12]. Fig. 2 shows a simplified block diagram used for the controller design. When the controller parameters in (8) and (9) are mismatched with the real motor parameters, the resultant model has the nonlinear disturbances in its input–output relation caused by the parameter variations as shown in Fig. 2. Since such disturbances directly influence on the control performance, their effects must be quickly removed.

In the conventional synchronous PI decoupling current control, the PI control is employed for the transfer function \(G_i(s)\) in Fig. 2, where \(u_{qs}\) and \(u_{ds}\) are determined using current errors as follows

\[
\begin{align*}
u_{qi} &= \left(K_P + \frac{K_i}{s}\right)(i_{q*} - i_q) \\
u_{di} &= \left(K_P + \frac{K_i}{s}\right)(i_{d*} - i_d)
\end{align*}
\]

where \(i_{q*}\) and \(i_{d*}\) are the commands for the \(q\)-axis and \(d\)-axis currents, respectively, \(K_P\) and \(K_I\) are the proportional and integral gains of the PI control, respectively, \(s\) is a Laplace operator and the symbol ‘*’ denotes the reference quantities. With the parameter matching between the controller and motor, that is, \(f_{qs} = f_{ds} = 0\), a closed-loop transfer function for the \(q\)-axis current can be obtained from Fig. 2 as follows

\[
T(s) = \frac{i_{q*}(s)}{i_{q*}(s)} = \frac{K_P f_{qs} + K_I}{L_{so}w^2 + (R_{so} + K_P)w + K_I}
\]

The closed-loop transfer function for the \(d\)-axis current can be determined in a similar way and a bandwidth of the current controller can be selected using (12).

In the predictive current control, the voltage references in (8) and (9) are calculated to force current error at next step to zero. In this control, \(u_{qs}\) and \(u_{ds}\) in (8) and (9) are obtained using the current reference and current at present step in a discrete domain as follows

\[
\begin{align*}
u_{qs} &= \frac{L_{so}}{T}(i_{q*}(k + 1) - i_q(k)) \\
u_{ds} &= \frac{L_{so}}{T}(i_{d*}(k + 1) - i_d(k))
\end{align*}
\]

where \(T\) is a sampling period. Using the reference voltages
in (8) and (9) and the control inputs in (13) and (14), a dead-beat type current controller having a quite fast transient response can be designed [5, 13].

4 MRAC-based adaptive current control

Generally, the controlled variables are DC quantities in the synchronous frame PI decoupling current regulator. Through the exact compensation for the back EMF and cross-coupling terms, this control may give ideal steady-state control characteristics irrespective of operating conditions. However, the transient response is generally slow or may be degraded due to the inexact cancellation input under the parameter variations. On the other hand, although a fast transient current response can be obtained in the predictive control scheme, this control gives a steady-state current error under the parameter mismatch between the motor and controller.

In this article, the steady-state performance of the predictive control scheme will be improved by estimating the disturbances caused by the parameter variations using the MRAC scheme and by compensating using a feedforward control, while retaining its good transient response. When the parameters of the PM synchronous motor are exactly known, \( f_{qs} \) and \( f_{ds} \) become zeros. In this case, the disturbances in Fig. 2 can be removed by the decoupling control in (8) and (9), and thus the transfer function as (12) can be obtained through the PI control, from which a desired current controller bandwidth can be assigned. Also, in the predictive current control, a dead-beat type controller having a fast transient and zero steady-state error can be designed using (8), (9), (13) and (14). Under the parameter variations, however, \( f_{qs} \) and \( f_{ds} \) in Fig. 2 are not zeros any longer and act as disturbances in its input–output relation. Since such disturbances directly influence on the current control performance, their effects must be quickly removed. To cope with such an influence, the voltage references in the proposed MRAC-based control scheme are determined using estimates for the disturbances as follows

\[
v_{qi}^* = u_{qi} + L_{so} \omega_{i} i_{di} + \lambda_{so} \omega_{i} + \hat{f}_{qi} \quad (15)
\]

\[
v_{di}^* = u_{di} - L_{so} \omega_{i} i_{qi} + \hat{f}_{di} \quad (16)
\]

where the symbol ‘\(^*\)’ denotes the estimated quantities and the control inputs \( u_{qi} \) and \( u_{di} \) are obtained based on the predictive control scheme in (13) and (14). When the reference voltages in (15) and (16) are applied to (7), the state equation model of the PM synchronous motor can be transformed as

\[
\dot{x} = A_1 x + B_1 u - B_2 \Delta f_{qs} - B_2 \Delta f_{ds} \quad (17)
\]

where \( u = [u_{qi} \quad u_{di}]^T \), \( \Delta f_{qs} = f_{qs} - \hat{f}_{qs} \), \( \Delta f_{ds} = f_{ds} - \hat{f}_{ds} \)

\[
B_1 = \left( \begin{array}{cc} \frac{1}{L_{so}} & 0 \\ 0 & \frac{1}{L_{so}} \end{array} \right)^T, \quad B_2 = \left( \begin{array}{c} \frac{R_{so}}{L_{so}} \\ 0 \end{array} \right), \quad A_1 = \left( \begin{array}{cc} \frac{-R_{so}}{L_{so}} & 0 \\ 0 & \frac{-R_{so}}{L_{so}} \end{array} \right)
\]

To estimate the parameters using the MRAC technique, an adjustable model and a reference model should be required [12–14]. An adjustable model is obtained from (17). A reference model is obtained from (17) with the parameter matching condition, namely, \( \Delta f_{qs} = 0 \) and \( \Delta f_{ds} = 0 \) as follows

\[
\dot{x}_M = A_1 x_M + B u \quad (18)
\]

where \( x_M = [x_{1M} \quad x_{2M}]^T \). This reference model represents an ideal dynamic behaviour of current when there are no parameter variations. An error between the adjustable model and reference model is defined as

\[
e = x - x_M \quad (19)
\]

The error between two models will be used to update the estimated disturbance in an adjustable model to reduce the output error. By subtracting (18) from (17), the error dynamics can be obtained as

\[
\dot{e} = A_1 e - B_1 \Delta f_{qs} - B_2 \Delta f_{ds} \quad (20)
\]

The design objective of the adaptive control using the error dynamics in (20) becomes as [14]

1. \( \lim_{t \to \infty} e(t) = 0 \) for any initial conditions \( e(0), \Delta f_{qs}(0) \) and \( \Delta f_{ds}(0) \)

2. Find the adaptation rules that lead to \( \lim_{t \to \infty} \hat{f}_{qs} = f_{qs} \) and \( \lim_{t \to \infty} \hat{f}_{ds} = f_{ds} \).

To derive the adaptation laws using the Lyapunov stability theory, the Lyapunov function be selected as

\[
V = e^T P e + \frac{1}{k_2} (\Delta f_{qs})^2 + \frac{1}{k_2} (\Delta f_{ds})^2 \quad (21)
\]

where \( P \) is a symmetric positive definite matrix and \( k_2 \) is a positive constant. The derivative of (21) can be obtained as

\[
\frac{d}{dt} V = e^T P \dot{e} + e^T \dot{P} e + \frac{2}{k_2} (\Delta f_{qs}) \frac{d}{dt} (\Delta f_{qs}) + \frac{2}{k_2} (\Delta f_{ds}) \frac{d}{dt} (\Delta f_{ds}) \quad (22)
\]

If \( \hat{f}_{qs} = 0 \) and \( \hat{f}_{ds} = 0 \), that is, \( d/dt(\Delta f_{qs}) = -d/dt f_{qs} \) and...
\[
\frac{d}{dt}(\Delta \theta) = -\frac{d}{dt} \hat{f}_d, \quad (22)
\]
can be expressed as
\[
\frac{d}{dt} V = -e^T Q e - 2\Delta \theta \left( e^T P B + \frac{1}{k_2} \frac{d}{dt} \hat{f}_d \right)
\]
where \( Q \) is a symmetric positive definite matrix expressed as
\[
A_1^T P + PA_1 = -Q \quad (24)
\]
If the adaptation rules for the disturbance estimation are chosen as
\[
\frac{d}{dt} \hat{f}_q = k_2 \cdot (-e^T P B_1) \quad (25)
\]
\[
\frac{d}{dt} \hat{f}_d = k_1 \cdot (-e^T P B_2) \quad (26)
\]
Equation (23) can be rewritten as
\[
\frac{d}{dt} V = -e^T Q e \leq 0 \quad (27)
\]
This proves the asymptotic stability of the adjustable model in (17) into the reference model in (18) [12]. Using (25) and (26) with the proportional adaptation gains to improve the estimation performance during the transient periods, the disturbances can be estimated as [14]
\[
\hat{f}_q = \left( k_1 + \frac{k_2}{s} \right) \cdot (-e^T P B_1) \quad (28)
\]
\[
\hat{f}_d = \left( k_1 + \frac{k_2}{s} \right) \cdot (-e^T P B_2) \quad (29)
\]
where \( k_1 \) and \( k_2 \) are the proportional and integral adaptation gains for the disturbance estimation, respectively. The estimated disturbances in (28) and (29) are employed to obtain the reference voltages of the proposed control scheme in (15) and (16) with (13) and (14).

5 Simulation and experimental results

The overall block diagram for the proposed MRAC-based adaptive current control scheme is shown in Fig. 3. The overall system consists of a proposed current controller, a PWM inverter, a PM synchronous motor and adaptive algorithm for the disturbance estimation. In this figure, the largest shaded area in the right side represents the adjustable model resulting from the parameter mismatch between the controller and motor, which is composed of a current controller, a PWM inverter and a motor. For the current control algorithm, the predictive control scheme with a feedforward disturbance compensation is employed. By using this technique, the reference voltages are composed of the predictive control part and disturbance compensation part. During these operations, the adaptive algorithms are used to estimate the disturbances caused by

![Figure 3 Overall block diagram for the proposed adaptive current control scheme](image-url)
the parameter variations in real time, and the estimated values are used to compute the reference voltages in (15) and (16). Fig. 4 shows the simplified block diagram of the proposed adaptive current control scheme when (15) and (16) are applied to the PM synchronous motor. Unlike Fig. 2, the disturbances appear as $D_{qfs}$ and $D_{fds}$ in system model that approach to zeros as the estimated values in (28) and (29) converge to real ones. Since the proposed control scheme uses only the control inputs in (13) and (14) for the input signal of the reference and adjustable models to estimate the disturbances, it does not require the measurement of the phase voltages unlike the conventional observer-based estimation scheme requiring the phase voltage as input [9].

Generally, this can be an effective way considering it is not simple to measure phase voltages exactly, since they often contain much harmonics as well as noise due to the switching. The calculated reference voltages $v^{*}_{qs}$ and $v^{*}_{ds}$ are applied to a PM synchronous motor through the space vector PWM technique [15]. Based on this technique, the conduction periods of the inverter switches are determined to obtain the average voltage equal to the reference. Also, during the large transient periods such as a starting and a sudden load change where the average voltages equal to $v^{*}_{qs}$ and $v^{*}_{ds}$ cannot be produced, the conduction durations of the inverter switches are simply adjusted as the method in [16] to approximate the average voltage to the reference.

The configuration of the experimental system is shown in Fig. 5. The whole control algorithms are implemented using DSP TMS320C31 [10, 11] and the sampling period is set to 128 μs both in the simulations and experiments, which yields a switching frequency of 7.8 kHz. The PM synchronous motor is driven by a three-phase PWM inverter employing the intelligent power module (IPM). The rotor speed and absolute rotor position are detected through a 12 bit/rev resolver-to-digital converter (RDC) using a brushless resolver. The phase currents are detected by the Hall-effect devices and are converted through two 12-bit A/D converters, where the resolution of current is $8/2^{12}$ A. The detected phase currents are transformed todqvalues in a DSP using the rotor position. Fig. 6 shows the experimental test setup consisting of PM synchronous motor, an inertial load and PM DC generator. Fig. 7 shows the photograph of experimental system. The nominal parameters of a PM synchronous motor used for the simulations and experiments are listed in Table 1.

![Figure 4](image1)

**Figure 4** Simplified block diagram using the proposed adaptive current control scheme

![Figure 5](image2)

**Figure 5** Configuration of the experimental system

Although the synchronous PI decoupling current control provides an ideal steady-state response without parameter variations since $f_{qs}$ and $f_{ds}$ in Fig. 2 become zeros, the current has relatively a large overshoot and slow transient under the parameter variations due to the mismatch of the cancellation inputs. Fig. 8 shows the simulation result for the synchronous PI decoupling current control under the parameter variations ($\Delta L_m = -0.5L_m, \Delta L_s = 1.0L_s$, and $\Delta R_s = 1.0R_s$). The $q$-axis and $d$-axis current references are given as 2 A and zero, respectively, and the motor is operated at a constant speed of 1200 rpm. The gains for the PI control are selected as $K_P = 17$ and $K_I = 20000$ so that the bandwidth of the current controller becomes 2000 rad/s. From the top, this figure shows the $q$-axis current, $d$-axis current, $a$-phase current at each sampling instant and real $a$-phase current responses. It can be clearly shown in this figure that the current has relatively a large

![Figure 6](image3)

**Figure 6** Experimental test setup

![Figure 7](image4)

**Figure 7** Photograph of experimental system
overshoot and slow transient response of about 4.5 ms. This degradation becomes more serious at high speed since the disturbance is proportional to operating speed as in (5) and (6).

Fig. 9 shows the simulation result for the predictive current control without parameter variations. The \( q \)-axis and \( d \)-axis current references are the same as Fig. 8. As is shown in Fig. 9, this control gives quite a fast transient response and the current response reaches its steady-state within 1 ms. However, this control requires the exact motor parameters and gives a severe steady-state error under the parameter mismatch.

Fig. 10 shows the simulation result for the predictive current control under the same parameter variations as Fig. 8. The steady-state errors are observed in the \( q \)-axis current, \( d \)-axis current and \( a \)-phase current responses caused by the inexact cancellation inputs in (8), (9), (13) and (14) as a result of the parameter variations. Even though an additional integrator can be used in the controller to improve this steady-state error, introducing an integral action may cause a slow transient.

Fig. 11 shows the simulation result for the proposed MRAC-based adaptive current control scheme under the same parameter variations as Figs. 8 and 10. Similarly, the \( q \)-axis and \( d \)-axis current references are given as 2 A and zero, respectively. The gains of the adaptation algorithm for the disturbance estimation are chosen as follows: \( k_1 = 120 \), \( k_2 = 50 \, 000 \) and \( Q = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix} \). From the top, this figure shows the \( q \)-axis current, \( d \)-axis current, \( a \)-phase current at each sampling instant, real \( a \)-phase current responses and the estimating performance of the disturbances, where the real disturbances are computed using (5) and (6) for the performance comparison with the estimates. It is noted that both estimated values well converge to their real values. This effectively suppresses the influence of the disturbance on the current responses. It is clearly shown that the current control performance can be significantly improved through the effective feedforward compensation of the disturbances estimated by the proposed MRAC scheme.

### Table 1 Specifications of a PM synchronous motor

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>rated power</td>
<td>400 W</td>
</tr>
<tr>
<td>rated torque</td>
<td>1.274 Nm</td>
</tr>
<tr>
<td>magnetic flux</td>
<td>0.16 Wb</td>
</tr>
<tr>
<td>synchronous inductance</td>
<td>5 mH</td>
</tr>
<tr>
<td>rated speed</td>
<td>3000 rpm</td>
</tr>
<tr>
<td>number of poles</td>
<td>4</td>
</tr>
<tr>
<td>phase resistance</td>
<td>3 Ω</td>
</tr>
<tr>
<td>moment of inertia</td>
<td>( 1.54 \times 10^{-4} ) Nms(^2)</td>
</tr>
</tbody>
</table>

![Figure 8](image_url) **Figure 8** Synchronous PI decoupling current control under the parameter variations

![Figure 9](image_url) **Figure 9** Predictive current control without parameter variations
where the current response reaches its steady-state within 1.5 ms and has a zero steady-state error. As compared with the method in Fig. 8, Fig. 11 shows a smaller overshoot and a faster transient response. Also, as compared with Fig. 10, Fig. 11 shows an improved steady-state response similar to Fig. 9 in case of the predictive current control without parameter variations. Thus, it is well verified that the influence of the disturbances caused by the parameter variations can be effectively suppressed.

Fig. 12 shows the performance comparison between the predictive current control and proposed MRAC-based current control scheme under the variations of the phase resistance and synchronous inductance. In the predictive control, the steady-state error and phase delay are clearly observed. On the other hand, in the proposed scheme, it is shown that these errors are effectively removed in current responses. Fig. 13 shows the estimating performance of the disturbances for the proposed control scheme in Fig. 12. Fig. 14 shows the performance comparison under the further variation of the synchronous inductance, and Fig. 15 shows the estimating performance of the disturbances for the proposed control scheme in this case. From these results, although \(f_{qs}\) and \(f_{ds}\) in (5) and (6) are not constant under the parameter mismatches in the phase resistance and synchronous inductance, the assumption of \(\hat{f}_{qs} = 0\) and \(\hat{f}_{ds} = 0\) can be reasonable during each fast sampling interval, since a sufficiently fast sampling period as compared with the current dynamics is often employed in digital controller to control the motor current.

Concerning to the parameter convergence, \(f_{qs}\) and \(f_{ds}\) are regarded as total disturbances as in (17) and they are related with the \(q\)-axis and \(d\)-axis current dynamics.
respectively. If the current errors in (20) are driven to zero asymptotically \((e \to 0)\), it implies \(\Delta f_{p_i} \to 0\) and \(\Delta f_{d_i} \to 0\) as follows:

\[
-B_1 \cdot \Delta f_{p_i} - B_2 \cdot \Delta f_{d_i} = \begin{pmatrix} -1 \\ -1 \end{pmatrix} f_{p_i} + \begin{pmatrix} 1 \\ 1 \end{pmatrix} f_{d_i}
\]

which results in the parameter convergence into real ones without requiring a complex current reference for the persistent excitation. This can be observed in Figs. 13 and 15 where the simple current reference does not provide the parameter convergence error.

Fig. 16 shows the performance comparison under the same parameter variations as Fig. 11 when a compensation of the inverter nonlinearity such as the dead-time does not exist. As expected, the effects of dead time as well as parameter variations influence current responses at the same time. Generally, the dead time acts as a dominant disturbance at low speed, on the other hand, the parameter variation acts as a dominant disturbance voltage at high speed. Fig. 17 shows the estimating performance of disturbances for the proposed scheme in Fig. 16. In this case, the estimated disturbances converge to \(f_{p_i} + V_{d,\text{dead}}\) and \(f_{d_i} + V_{d,\text{dead}}\), instead of \(f_{p_i}\) and \(f_{d_i}\), where \(V_{d,\text{dead}}\) and \(V_{d,\text{dead}}\) denote the output voltage errors in the synchronous frame caused by the dead time. This is due to the fact that the proposed adaptation scheme has the nature of disturbance estimator [17].

Fig. 18 shows the experimental result for the predictive current control without the parameter variations. The \(q\)-axis and \(d\)-axis current references are set to 2 A and zero, respectively, and all the experiments are carried out under the same conditions as the simulations. It is clearly observed to have a good control performance similar to the simulation result in Fig. 9.
Fig. 19 shows the experimental result for the predictive current control under $\Delta \lambda_m = -0.5\lambda_m$. In all current responses, the steady-state errors are observed as in Fig. 10 as a result of the parameter variations, which are expected to be proportional to the rotor speed.

Fig. 20 shows the experimental result for the proposed MRAC-based adaptive current control scheme under the parameter variations. Since the estimated disturbances are effectively compensated through a feedforward control by using the proposed control scheme, it is confirmed that the current control performance can be significantly improved, which is well coincident with the simulation result in Fig. 11.

6 Conclusions

To improve a performance of the servo drive such as the transient and steady-state responses, an MRAC-based adaptive current control scheme of a PM synchronous motor has been presented. Although the predictive current control is known to give ideal transient and steady-state responses, its steady-state response may be degraded in the presence of the motor parameter variations. This is more serious at high-speed operations since the disturbance is proportional to operating speed. To overcome this problem, the disturbances caused by the parameter variations are estimated by using the MRAC technique and are employed for the compensation by a feedforward manner. Thus, the steady-state control performance can be effectively improved, while retaining its good dynamic performance. The asymptotic stability of the proposed control scheme is proved and the adaptation laws are derived by the Lyapunov stability theory. The proposed control scheme does not require the measurement of the phase voltages unlike the conventional disturbance estimation scheme using observer. This can be an effective way considering that the phase voltages contain much harmonics as well as noise by the inverter switching.
By using the proposed scheme, a systematic design approach for a robust current control can be accomplished. The overall control system is implemented using DSP TMS320C31 and its feasibility and effectiveness are verified through the comparative simulations and experiments.

7 References


