An Efficient On-line Identification Approach to Rotor Resistance of Induction Motors Without Rotational Transducers

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Abstract

In this paper, we propose an effective on-line identification method for rotor resistance, which is useful in making speed control of induction motors without rotational transducers robust with respect to the variation in rotor resistance. Our identification method for rotor resistance is based on the linearly perturbed equations of the closed-loop system for sensorless speed control about the operating point. Our identification method for rotor resistance uses only the information of stator currents and voltages. It can provide fairly good identification accuracy regardless of load conditions. Some experimental results are presented to demonstrate the practical use of our identification method. For our experimental work, we have built a sensorless control system, in which all algorithms are implemented on a DSP. Our experimental results confirm that our on-line identification method allows for high precision speed control of commercially available induction motors without rotational transducers.

Nomenclature

V_{ds} (V_{qs})	d(q)-axis stator voltage
i_{ds} (i_{qs})	d(q)-axis stator current
ϕ_{dr} (ϕ_{qr})	d(q)-axis rotor flux
ϕ_{ds} (ϕ_{qs})	d(q)-axis stator flux
ϕ_s (ϕ_r)	magnitude of stator(rotor) flux vector ($\equiv \sqrt{\phi_{ds}^2+\phi_{qs}^2}$ ($\sqrt{\phi_{dr}^2+\phi_{qr}^2}$))
ω_r (ω_e)	angular velocity of motor speed(reference frame)
ω_s	angular velocity of rotor flux
ω_i	angular velocity of injected AC component
ω_r^* (ϕ_{dr}^*)	speed(rotor flux) command
T_e (T_L)	generated torque(load torque)
p	number of pole pairs
M	mutual inductance between stator and rotor
L_{sl} (L_{rl})	leakage inductance of stator(rotor)
L_s (L_r)	stator(rotor) self-inductance ($=M+L_{sl}$ ($=M+L_{rl}$))
R_s (R_r)	stator(rotor) resistance
$\hat{R}_r$	identified value of rotor resistance
$\hat{\phi}_{dr}$ ($\hat{\phi}_{qr}$)	estimated value of d(q)-axis rotor flux
$\hat{\phi}_r$	estimated value of rotor flux magnitude
$\bar{x}$	steady-state value of state x
Δx	perturbed component of state x
J_M (B_M)	inertia(viscous damping coefficient) of motor
σ	leakage coefficient ($\equiv 1-M^2/(L_sL_r)$)
K_T	torque constant ($\equiv 3pM/(2L_r)$)
sgn(x)	sign function
a_i, $i=0,\cdots,5$	induction motor parameters

$$a_0 \equiv 1/(\sigma L_s),\quad a_3 \equiv a_0M/L_r,\quad a_4 \equiv R_r/L_r,$$
$$a_2 \equiv a_3a_4,\quad a_5 \equiv Ma_4,\quad a_1 \equiv a_0R_s+Ma_2,$$
$$a_6 \equiv B_M/J_M,\quad a_7 \equiv 1/J_M.$$

$\hat{a}_i$, $i=1,2,4,5.$ estimated induction motor parameters

$$\hat{a}_4 \equiv \hat{R}_r/L_r,\quad \hat{a}_2 \equiv a_3\,\hat{a}_4,$$
$$\hat{a}_5 \equiv M\,\hat{a}_4,\quad \hat{a}_1 \equiv a_0R_s+M\,\hat{a}_2.$$

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I. Introduction

The recently proposed methods, for instance, in [1-3] for speed control of induction motors without rotational transducers can provide the desired high performance but request the accurate information of motor parameters. Especially, rotor resistance R_r needs to be known with fairly good accuracy. This is because rotor resistance varies greatly with machine temperature, while its large variation can degrade seriously the control performance of the recently proposed methods for speed control of induction motors without rotational transducers. Most of the previous results[4-6] on on-line identification of motor parameters need for the information of motor speed and hence cannot be used for induction motors without rotational transducers. On the other hand, only a few prior works[7-9] concern on-line parameter identification for induction motors without rotational transducers.

In [7], full order MRAS(Model Reference Adaptive System) is proposed for sensorless speed control of induction motors and is to identify both R_s and R_r as well as motor speed. However, a rigorous performance analysis is not given. In fact, our simulation study has shown that the proposed method may diverge when $\omega_r T_L < 0$, that is, the load torque helps the motor run.

A multi-stage recursive least square method[8] also was proposed for the identification of all motor parameters as well as motor speed. Unfortunately, the implementation of this method involves the second order time-derivatives of stator currents and the first order time-derivatives of stator voltages.

In [9], the ripples in stator currents and voltages caused by the PWM inverter are measured to identify R_r as well as motor speed. This method needs for high resolution sensors and A/D converters in order to detect the ripples within the desired accuracy and, furthermore, involves pure integration of some quantities, which may be very sensitive to DC offset of measurement data.

In this paper, we propose an effective method of identifying R_r on-line. Our identification method for R_r is based on the linearly perturbed equations of the closed-loop system about the operating point. Our on-line identification method for R_r can provide fairly good identification accuracy regardless of load conditions. Pure integrations of measured quantities are not required. Furthermore, high-order differentiations of stator currents and voltages are avoided by use of appropriate band pass filters.

Experimental results show that our identification algorithm can identify R_r within the accuracy of $\pm 10\%$, and allows for high precision speed control of commercially available induction motors without rotational transducers.

II. Main Results

The dynamic model of induction motors can be described in the d-q synchronous reference coordinate frame as follows:

$$\dot{i}_{ds} = -a_1 i_{ds} + \omega_e i_{qs} + a_2\phi_{dr} + pa_3\omega_r\phi_{qr} + a_0 V_{ds}, \quad \text{(1a)}$$

$$\dot{i}_{qs} = -a_1 i_{qs} - \omega_e i_{ds} + a_2\phi_{qr} - pa_3\omega_r\phi_{dr} + a_0 V_{qs}, \quad \text{(1b)}$$

$$\dot{\phi}_{dr} = -a_4\phi_{dr} + (\omega_e - p\omega_r)\phi_{qr} + a_5 i_{ds}, \quad \text{(1c)}$$

$$\dot{\phi}_{qr} = -a_4\phi_{qr} - (\omega_e - p\omega_r)\phi_{dr} + a_5 i_{qs}, \quad \text{(1d)}$$

$$\dot{\omega}_r = -a_6\omega_r + a_7(T_e - T_L), \quad \text{(1e)}$$

where the generated torque T_e is given by

$$T_e \equiv K_T(-\phi_{qr} i_{ds} + \phi_{dr} i_{qs}), \quad \text{(2)}$$

The definitions of the variables and the constants that appear in the above equations are given in Nomenclature.

The magnitudes of L_{sl} and L_{rl} are generally a few percent of that of M and their variations are relatively small compared with those of the other parameters. Thus, L_{sl} and L_{rl} may be regarded as constant in the operating region of interest[4]. On the other hand, M depends mainly on stator flux but is not affected seriously by machine temperature and motor speed. For this reason, the variation of M is compensated on-line by using the data of M vs. ϕ_s, which are obtained via appropriate off-line test and are stored in microprocessor memory.

The stator resistance R_s and rotor resistance R_r are known to depend on machine temperature[4]. Thus we need to identify on-line both stator and rotor resistances in order to have recently proposed sensorless speed control method provide the desired high performance. However, in this paper, we focus only on the on-line identification of rotor resistance because of limited space. The on-line identification of stator resistance will be presented in a separate paper. Therefore we assume that M, L_{sl}, L_{rl}, and R_s are constant and are known a priori.

The proposed identification algorithm for R_r id derived under the following assumptions.

(A.1) The motor speed is constant.

(A.2) The angular velocity and the magnitude of rotor flux vector are constant.

(A.3) The rotor resistance as well as load torque do not vary during the identification process.

Note that the above assumptions are practically reasonable and can be well accepted.

Before presenting our identification method for R_r, we need some preliminary results. Our closed-loop system for sensorless speed control consists of an induction motor without any rotational transducer, our feedback linearizing controller proposed in [5]:

$$\omega_e \equiv p\,\hat{\omega}_r + (\hat{a}_5\, i_{qs}) / \hat{\phi}_{dr}, \quad \text{(3a)}$$

$$V_{ds} \equiv -(\omega_e i_{qs}) / a_0 + v_1, \quad \text{(3b)}$$

$$V_{qs} \equiv p\,\hat\omega_r(i_{ds}+a_3\,\hat\phi_{dr})/a_0+v_2/\,\hat\phi_{dr}, \tag{3c}$$

$$v_1 \equiv K_{pd}(\tilde u_1-i_{ds})+K_{id}\int_0^t(\tilde u_1-i_{ds})dt, \tag{3d}$$

$$v_2 \equiv K_{pq}(\tilde u_2-\widehat{\phi_{dr}}i_{qs})+K_{iq}\int_0^t(\tilde u_2-\widehat{\phi_{dr}}i_{qs})dt, \tag{3e}$$

$$\tilde u_1 \equiv -K_{p\phi}\,\hat\phi_{dr}+K_{i\phi}\int_0^t(\phi_{dr}^*-\hat\phi_{dr})dt, \tag{3f}$$

$$\tilde u_2 \equiv -K_{p\omega}\,\hat\omega_r+K_{i\omega}\int_0^t(\omega_r^*-\hat\omega_r)dt, \tag{3g}$$

and our rotor flux and speed estimator proposed in [3]:

$$\begin{aligned}\dot{\hat\phi}_r = &-\left\{\hat a_4-\hat a_2\,\hat a_5\frac{(\hat\alpha_d i_{ds}+\hat\alpha_q i_{qs})}{|\hat\alpha_d|^2+|\hat\alpha_q|^2}\right\}\hat\phi_r\\ &+\hat a_5\,\mathrm{sgn}(\omega_r)(\hat\alpha_d i_{qs}-\hat\alpha_q i_{ds})/(|\hat\alpha_d|^2+|\hat\alpha_q|^2)\\ &\times\sqrt{|\hat\alpha_d|^2+|\hat\alpha_q|^2-|\hat\phi_r|^2},\end{aligned} \tag{4a}$$

$$\hat\omega_r \equiv \mathrm{sgn}(\omega_r)\sqrt{|\hat\alpha_d|^2+|\hat\alpha_q|^2-|\hat\phi_r|^2}/pa_3\,\hat\phi_r, \tag{4b}$$

$$\hat\phi_{dr} \equiv |\hat\phi_r|^2\frac{\hat a_2\,\hat\alpha_d-a_3\,\hat\alpha_q p\,\hat\omega_r}{|\hat\alpha_d|^2+|\hat\alpha_q|^2}, \tag{4c}$$

$$\hat\phi_{qr} \equiv |\hat\phi_r|^2\frac{\hat a_2\,\hat\alpha_q+a_3\,\hat\alpha_d p\,\hat\omega_r}{|\hat\alpha_d|^2+|\hat\alpha_q|^2}, \tag{4d}$$

where the quantities $\hat\alpha_d$ and $\hat\alpha_q$ are defined by

$$\begin{aligned}\hat\alpha_d &\equiv \beta_d+M\,\hat a_2 i_{ds}, \qquad \hat\alpha_q \equiv \beta_q+M\,\hat a_2 i_{qs},\\ \beta_d &\equiv \dot i_{ds}-\omega_e i_{qs}+a_0R_s i_{ds}-a_0V_{ds},\\ \beta_q &\equiv \dot i_{qs}+\omega_e i_{ds}+a_0R_s i_{qs}-a_0V_{qs}.\end{aligned} \tag{5}$$

Here, (4) is simply the d-q synchronous reference frame representation of flux and speed estimator proposed in [3]. Note that, $|\hat\alpha_d|^2+|\hat\alpha_q|^2$, $\hat\alpha_d i_{ds}+\hat\alpha_q i_{qs}$, and $\hat\alpha_d i_{qs}-\hat\alpha_q i_{ds}$ are invariant quantities under d-q coordinate transformation. Thus, (4) has the same performance described in [3].

In (3) and (4), $\hat R_r$ represents the value of rotor resistance identified by our identification algorithm for R_r. And, it is clear that if $\hat R_r=R_r$ then our feedback linearizing controller and our rotor flux and speed estimator will play the desired roles described in [3, 5]. That is, the feedback linearizing controller in (3) will make the dynamic characteristics between the input $(\omega_r^*,\ \phi_{dr}^*)$ and the output $(\omega_r,\ \phi_{dr})$ decoupled and linear, while the rotor flux and speed estimator in (4) will assure that $\hat\omega_r=\omega_r$, $\hat\phi_{dr}=\phi_{dr}$, and $\hat\phi_{qr}=\phi_{qr}$.

First, we introduce some identities which will be used in deriving our identification algorithm for R_r. First, we can easily obtain the following identities using (4b) - (4d).

$$\begin{aligned}\hat\alpha_d &= \hat a_2\,\hat\phi_{dr}+pa_3\,\hat\omega_r\,\hat\phi_{qr},\\ \hat\alpha_q &= \hat a_2\,\hat\phi_{qr}-pa_3\,\hat\omega_r\,\hat\phi_{dr}.\end{aligned} \tag{6}$$

Next, we define the quantities α_d, α_q by

$$\alpha_d \equiv \beta_d+Ma_2 i_{ds},\quad \alpha_q \equiv \beta_q+Ma_2 i_{qs}. \tag{7}$$

Then we can obtain from (1a) and (1b) the following identities similar to those in (6).

$$\alpha_d = a_2\phi_{dr}+pa_3\omega_r\phi_{qr},\quad \alpha_q = a_2\phi_{qr}-pa_3\omega_r\phi_{dr}. \tag{8}$$

Next, we attempt to find an expression of the rotor flux at steady-state. To do so, we multiply both sides of (1c) and (1d) by ϕ_{dr} and ϕ_{qr}, respectively, and add the resulting two equations. Then, we obtain the following equation for $\Phi_r \equiv |\phi_{dr}|^2+|\phi_{qr}|^2$.

$$\dot\Phi_r = -2a_4\Phi_r+2a_5(\phi_{dr}i_{ds}+\phi_{qr}i_{qs}). \tag{9}$$

By (9) along with **(A.2)**, we see that the following algebraic equation holds at steady-state.

$$|\bar\phi_{dr}|^2+|\bar\phi_{qr}|^2-M(\bar\phi_{dr}\,\bar i_{ds}+\bar\phi_{qr}\,\bar i_{qs}) = 0. \tag{10a}$$

On the other hand, it follows from (7), (8), and (10a) that

$$\bar\beta_d\,\bar\phi_{dr}+\bar\beta_q\,\bar\phi_{qr} = 0. \tag{10b}$$

Solving (10) for $\bar\phi_{dr}$ and $\bar\phi_{qr}$, we finally have the following expression of the rotor flux at steady-state.

$$\begin{aligned}\bar\phi_{dr} &= M\frac{\bar\beta_q\,\bar i_{ds}-\bar\beta_d\,\bar i_{qs}}{|\bar\beta_d|^2+|\bar\beta_q|^2}\,\bar\beta_q,\\ \bar\phi_{qr} &= -M\frac{\bar\beta_q\,\bar i_{ds}-\bar\beta_d\,\bar i_{qs}}{|\bar\beta_d|^2+|\bar\beta_q|^2}\,\bar\beta_d.\end{aligned} \tag{11}$$

Now, we derive some important steady-state characteristics of our closed-loop system given by (1) - (5). To do so, we first write the differential equation in (4a) in the following form using (4b) - (4d).

$$\dot{\hat\phi}_r = -\hat a_4\,\hat\phi_r+\hat a_5\,(\hat\phi_{dr}i_{ds}+\hat\phi_{qr}i_{qs})/\,\hat\phi_r. \tag{12a}$$

On the other hand, by the definitions of $\hat\alpha_d$ and $\hat\alpha_q$ in (5), we have the following equality.

$$\beta_d\,\hat\phi_{dr}+\beta_q\,\hat\phi_{qr}+M\,\hat a_2(\hat\phi_{dr}i_{ds}+\hat\phi_{qr}i_{qs}) = \hat a_2\,\hat\Phi_r. \tag{12b}$$

Now, from (12), we can see that the steady-state values of $\hat\phi_{dr}$ and $\hat\phi_{qr}$ can be written as

$$\begin{aligned}\bar{\hat\phi}_{dr} &= M\frac{\bar\beta_q\,\bar i_{ds}-\bar\beta_d\,\bar i_{qs}}{|\bar\beta_d|^2+|\bar\beta_q|^2}\,\bar\beta_q,\\ \bar{\hat\phi}_{qr} &= -M\frac{\bar\beta_q\,\bar i_{ds}-\bar\beta_d\,\bar i_{qs}}{|\bar\beta_d|^2+|\bar\beta_q|^2}\,\bar\beta_d,\end{aligned} \tag{13}$$

and, in turn that

$$\bar{\hat\phi}_{dr} = \bar\phi_{dr},\qquad \bar{\hat\phi}_{qr} = \bar\phi_{qr}. \tag{14}$$

Furthermore, the following identity can be derived from **(A.2)**, (3a), (6)-(8) and (14).

$$\bar{\hat\phi}_{qr} = 0. \tag{15}$$

Note that, as is shown below, the estimation error in rotor resistance directly affects the accuracy of the set-point tracking for motor speed. That is,

$$p\,\bar{\omega}_r - p\,\overline{\hat{\omega}}_r = -(a_5 - \hat{a}_5)\,\bar{i}_{qs} / \overline{\hat{\phi}}_{dr}. \tag{16}$$

The above identity can be derived from the following relationships at steady-state among motor speed and the angular velocities of the reference frame and the rotor flux.

$$\begin{aligned} \bar{\omega}_s &= \bar{\omega}_e, \\ \bar{\omega}_s &= p\,\bar{\omega}_r + (a_5\,\bar{i}_{qs} / \overline{\hat{\phi}}_{dr}). \end{aligned} \tag{17}$$

This shows that the regulation error of motor speed is proportional to the estimation error of rotor resistance.

Now, we are ready to describe our identification method for R_r. When our closed-loop system with both the motor speed command ω_r^* and the rotor flux command ϕ_{dr}^* kept constant reaches its steady-state, we add a small AC component, say,

$$\Delta\phi^*_{dr} = A\sin\omega_i t, \tag{18}$$

to the flux command. Then, Wait until the sensorless speed control system reaches a kind of sinusoidal steady-state. Then the state variable x of the system can be approximated as the sum of the DC component $\bar{x}$ and the AC component Δx of frequency ω_i, that is, $x = \bar{x} + \Delta x$. Recall that the feedback linearizing controller in (3) proposed in [5] allows for decoupled control of rotor flux and motor speed. Hence, this small AC component will not cause serious disturbance in motor speed.

Through the linear perturbation of (1c) and (1d) about the operating point, we have

$$\Delta\ddot{\phi}_{dr} = -a_4\Delta\phi_{dr} + (\bar{\omega}_e - p\,\bar{\omega}_r)\Delta\phi_{qr} + a_5\Delta i_{ds}, \tag{19a}$$

$$\begin{aligned} \Delta\ddot{\phi}_{qr} = &-a_4\Delta\phi_{qr} - (\bar{\omega}_e - p\,\bar{\omega}_r)\Delta\phi_{dr} \\ &-(\Delta\omega_e - p\Delta\omega_r)\,\overline{\hat{\phi}}_{dr} + a_5\Delta i_{qs}. \end{aligned} \tag{19b}$$

Similarly, we can also obtain

$$\begin{aligned} \Delta\omega_e = &\; p\Delta\hat{\omega}_r + (\hat{a}_5\Delta i_{qs} / \overline{\hat{\phi}}_{dr}) \\ &-(\hat{a}_5\,\bar{i}_{qs}\Delta\hat{\phi}_{dr} / |\overline{\hat{\phi}}_{dr}|^2), \end{aligned} \tag{19c}$$

$$\begin{aligned} 0 = &\; (\hat{a}_2\Delta\hat{\phi}_{dr} - a_2\Delta\phi_{dr}) \\ &+ pa_3(\overline{\hat{\omega}}_r\Delta\hat{\phi}_{qr} - \bar{\omega}_r\Delta\phi_{qr}) - M(\hat{a}_2 - a_2)\Delta i_{ds}, \end{aligned} \tag{19d}$$

and

$$\begin{aligned} p\Delta\hat{\omega}_r = &\; p\Delta\omega_r + (1/\overline{\hat{\phi}}_{dr})\{(\hat{a}_4\Delta\hat{\phi}_{qr} - a_4\Delta\phi_{qr}) \\ &+ p(\bar{\omega}_r\Delta\phi_{dr} - \overline{\hat{\omega}}_r\Delta\hat{\phi}_{dr}) + M(a_4 - \hat{a}_4)\Delta i_{qs}\}. \end{aligned} \tag{19e}$$

Note that (A.1), (A.3), (14), and (15) have been used in the derivation of (19).

Now, it is clear that an equation which does not involve inaccessible quantities $\Delta\phi_{dr}$, $\Delta\phi_{qr}$, and $\Delta\omega_r$ can be derived from the above five equations. In fact, we can show through some tedious calculations that the following identity holds at sinusoidal steady-state.

$$R_r X + L_r Y = 0, \tag{20}$$

Here, X and Y are given by

$$\begin{aligned} X \equiv &\; M\Delta\ddot{i}_{ds} - M\frac{\hat{R}_r}{L_r}\Delta\dot{i}_{ds} \\ &+ M\bar{\omega}_e p\,\overline{\hat{\omega}}_r\Delta i_{ds} + \left(\frac{\hat{R}_r}{L_r} + \bar{\omega}_e M\frac{i_{qs}}{\hat{\phi}_{dr}}\right)\Delta\dot{\hat{\phi}}_{dr} \\ &- \bar{\omega}_e p\,\overline{\hat{\omega}}_r\Delta\hat{\phi}_{dr} + \left(p\,\overline{\hat{\omega}}_r - \frac{\hat{R}_r}{L_r}M\frac{i_{qs}}{\hat{\phi}_{dr}}\right)\Delta\dot{\hat{\phi}}_{qr} \\ &+ \left(\frac{\hat{R}_r}{L_r} + p\,\overline{\hat{\omega}}_r M\frac{i_{qs}}{\hat{\phi}_{dr}}\right)\Delta\hat{\phi}_{qr}, \end{aligned} \tag{21a}$$

$$\begin{aligned} Y \equiv &-M\frac{\hat{R}_r}{L_r}\Delta\ddot{i}_{ds} + \frac{\hat{R}_r}{L_r}\Delta\ddot{\hat{\phi}}_{dr} \\ &-|\bar{\omega}_e|^2\Delta\dot{\hat{\phi}}_{dr} + p\,\overline{\hat{\omega}}_r\Delta\ddot{\hat{\phi}}_{qr} + \frac{\hat{R}_r}{L_r}\bar{\omega}_e\Delta\dot{\hat{\phi}}_{qr}. \end{aligned} \tag{21b}$$

Note that the RHS of (21) involve only the quantities that can be measured and/or estimated. Therefore, we can identify the true value of rotor resistance by applying least square algorithm to the above equation (20).

As can be seen from (21), X and Y involve high order time-derivatives of measured and/or estimated quantities, which can be computed only with large errors. However, (20) implies that at sinusoidal steady-state,

$$R_r H(X) + L_r H(Y) = 0, \tag{22}$$

for any low pass filter H. In other words, the following identity holds at sinusoidal steady-state.

$$R_r \tilde{X} + L_r \tilde{Y} = 0, \tag{23}$$

Table 1. Data of the induction motor used for Experiment.

Nameplate Data	Nominal Values of Motor Parameters	
220V 60Hz	p	2
3phase Y-connected	Rs	3.46 Ω
Rated Power 750W	Rr	1.9 Ω
Rated Flux 0.4Wb	Ls	148.5 mH
Rated Torque 4.3Nm	Lr	148.5 mH
Rated Current 3.6A	M	140.3 mH

for any low pass filter H, where

$$\begin{aligned} \tilde{X} \equiv &\; MH(\Delta\ddot{i}_{ds}) - M\frac{\hat{R}_r}{L_r}H(\Delta\dot{i}_{ds}) \\ &+ \left(\frac{\hat{R}_r}{L_r} + \bar{\omega}_e M\frac{i_{qs}}{\hat{\phi}_{dr}}\right)H(\Delta\dot{\hat{\phi}}_{dr}) \end{aligned} \tag{24a}$$

$$
\begin{aligned}
&- \overline{\omega}_e p \overline{\widehat{\omega}}_r H(\varDelta \widehat{\phi}_{dr}) + M \overline{\omega}_e p \overline{\widehat{\omega}}_r H(\varDelta i_{ds}) \\
&+ (p \overline{\widehat{\omega}}_r - \frac{\widehat{R}_r}{L_r} M \frac{i_{qs}}{\widehat{\phi}_{dr}}) H(\varDelta \dot{\widehat{\phi}}_{qr}) \\
&+ (\frac{\widehat{R}_r}{L_r} + p \overline{\widehat{\omega}}_r M \frac{i_{qs}}{\widehat{\phi}_{dr}}) H(\varDelta \widehat{\phi}_{qr}), \\
\widehat{Y} \equiv &- M \frac{\widehat{R}_r}{L_r} H(\varDelta \ddot{i}_{ds}) \\
&+ \frac{\widehat{R}_r}{L_r} H(\varDelta \ddot{\widehat{\phi}}_{dr}) - |\overline{\omega}_e|^2 H(\varDelta \dot{\widehat{\phi}}_{dr}) \\
&+ p \overline{\widehat{\omega}}_r H(\varDelta \ddot{\widehat{\phi}}_{qr}) + \frac{\widehat{R}_r}{L_r} \overline{\omega}_e H(\varDelta \dot{\widehat{\phi}}_{qr}).
\end{aligned} \tag{24b}
$$

Here, $H(z)$ is the output of the low pass filter H to the input z. Therefore, we can avoid the undesirable differentiations of stator currents and other quantities by choosing a filter H satisfying

$$\text{degree of denominator} \geq 2 + \text{degree of numerator} \tag{25}$$

For an instance, the filtered quantities in computing $\widehat{X}$ and $\widehat{Y}$ in (24) can be obtained from the measured or estimated quantity through a filter of the following form.

$$H(s) \equiv H_1(s)\ H_2(s), \tag{26}$$

where

$$
\begin{aligned}
H_1(s) &\equiv |\omega_L|^2 / \{s^2 + (\omega_L/Q)\ s + |\omega_L|^2\}, \\
H_2(s) &\equiv (\omega_i/Q)\ s / \{s^2 + (\omega_i/Q)\ s + |\omega_i|^2\}.
\end{aligned}
$$

Hence, $H(s)$ is the cascade realization of a second order band pass filter $H_2(s)$ and a low pass filter $H_1(s)$.

Recall that only the AC components with frequency ω_i are of interest. For this reason, the center frequency of the band pass filter $H_2(s)$ is chosen to be ω_i. On the other hand, the second order low pass filter $H_1(s)$ is used to eliminate the measurement noise and to meet the requirement in (25), that is, to avoid the undesirable differentiations.

III. Experimental Results

In this section, we present some experimental results in order to demonstrate the practical use of our on-line identification method for rotor resistance. To this aim, we have built an all-digital drive system for speed control of an induction motor without any rotational transducer. Our overall controller consisted of (i) the feedback linearizing controller in (3) for decoupled control of rotor flux and motor speed, (ii) the flux and speed estimator in (4) for induction motors without rotational transducers, and (iii) our on-line identification algorithms for R_r in (23). All algorithms for control and identification were implemented digitally on a DSP chip(TMS320C32). The block diagram representation of the DSP based drive system built for our experiment is shown in Fig 1. In our drive system, PWM signals were generated by the well-known space vector modulation technique and dead-time compensation technique. And, stator currents and voltages were processed by 12-bit A/D converters.

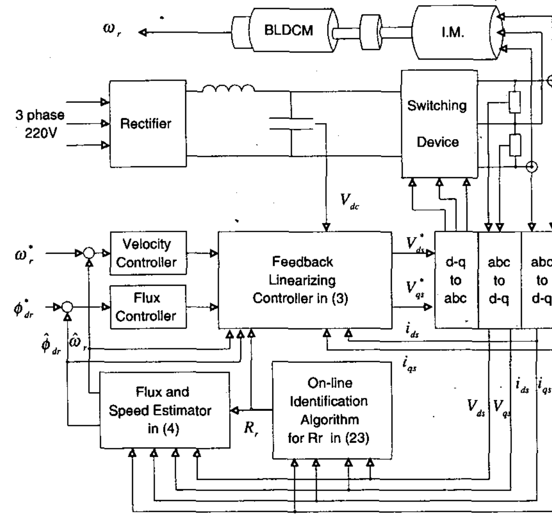

Fig. 1. Configuration of the DSP based drive system built for Experiment.

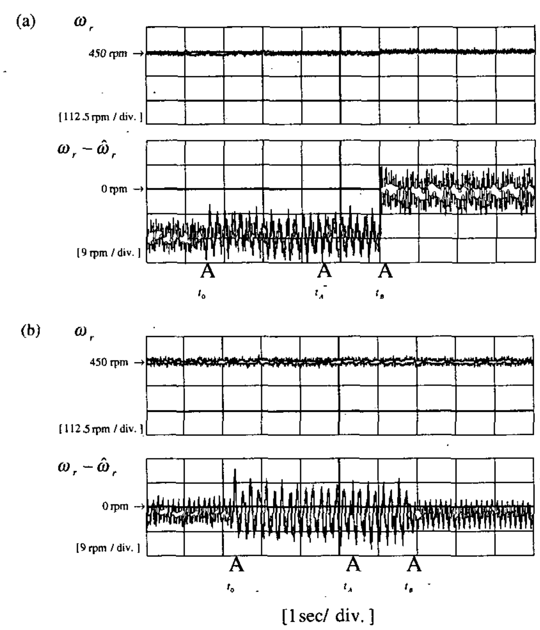

Fig. 2. Rr identification under (a) a half load condition, (b) no load condition.

The commercially available induction motor which was used for our experimental work was a four pole squirrel cage induction motor manufactured by Hyosung Industry Co., Korea. The motor data are listed in Table 1. For load test, a

BLDCM was coupled mechanically to the induction motor and the motor speed was monitored by the optical encoder(8192 pulse/rev) mounted on the BLDCM.

The experimental results shown in Fig 2 demonstrate the performance of our identification algorithm for rotor resistance. Initially, the induction motor was operated with $\omega_r^* = 450$rpm (0.25pu) and $\phi_{dr}^* = 0.4$Wb (1.0pu). And, we chose the quality factor of $H_2(s)$ and the cut-off frequency of $H_1(s)$ as follows.

$$Q = 3, \quad \omega_L = 10\omega_i.$$

The value of stator resistance used in our overall controller was taken by 3.57 Ω, the DC test value of stator resistance. However, the value of rotor resistance used in our overall controller was set to 1.33 Ω. Note that this corresponds to 70% of the nominal value 1.9 Ω of rotor resistance. A small AC component $\Delta\phi_{dr}^*$ in (18}) with $A = 0.02$Wb (0.05pu) and $\omega_i = 10\pi$ rad/sec was injected at $t = t_0$, as indicated in Fig 2. Then, our identification algorithm for R_r was initiated at $t = t_A$. At $t = t_B$, the injected AC component $\Delta\phi_{dr}^*$ was removed and the initial value of $\hat{R}_r$ was updated by the value of rotor resistance identified by our identification algorithm for R_r. Particularly, in case of the experimental result in Fig 2.(a), the load torque $T_L = 2.15$Nm (0.5pu) was applied during the experiment, while, in case of the experimental results in Fig 2.(b), no load torque was applied. The identified value of rotor resistance was 1.83 Ω in case of a half load condition, while it was 1.98 Ω in case of no load condition.

We can see from Fig 2.(a) that the speed control accuracy of our sensorless speed control system has been improved significantly by using the identified value of R_r. As can be seen from Fig 2.(b), however, the identified value of R_r cannot improve the speed control accuracy of our sensorless speed control system when it is operated in no load condition. This is mainly because, in case of small i_{qs}, the motor speed error does not depend significantly on the identification error of R_r, as can be seen from (16). However, even in no load condition, our identification method for rotor resistance can identify R_r with fairly good accuracy. And, our extensive experimental study have shown that our identification algorithm for R_r can identify the rotor resistance within $\pm$ 10% accuracy. It is also worthy to note from Fig 2 that, even in presence of 30% identification error in rotor resistance, a 5% oscillation in rotor flux command does not cause any serious performance degradation in speed control.

In Fig 3, we present the waveforms of $\tilde{X}$ and $\tilde{Y}$ in (24). As can be seen from Fig 3, the waveforms of $\tilde{X}$ and $\tilde{Y}$ are quite smooth although X and Y involve high order derivatives which can be obtained only with unacceptably large errors.

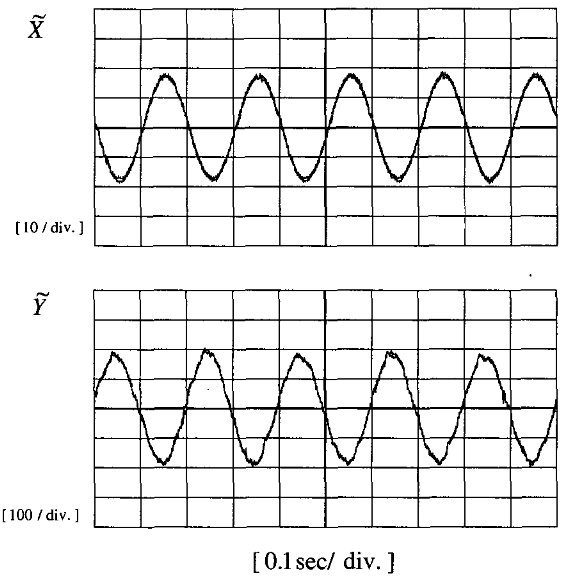

Fig. 3. Typical waveforms of $\tilde{X}$ and $\tilde{Y}$.

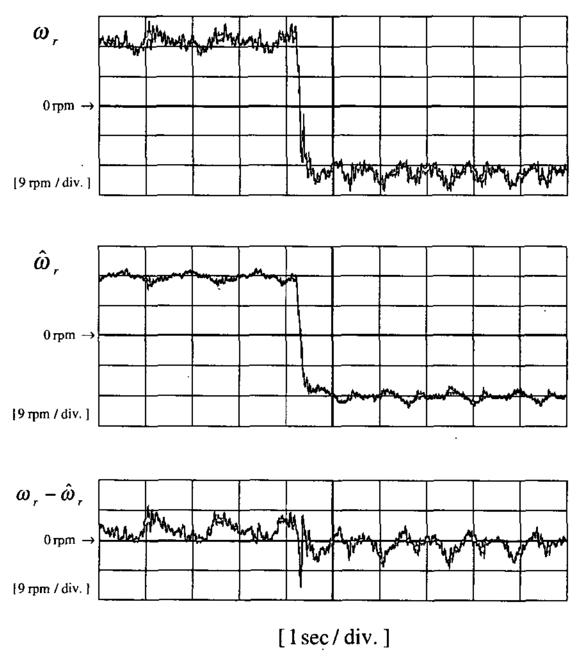

Fig. 4. Forward-reverse operation at very low speed under no load condition(18rpm).

Finally, we present the experimental results on the dynamic performance of our closed-loop system for sensorless speed control, when our overall controller used the identified values of stator and rotor resistance. Fig 4 presents the experimental results for the motor operation with the rated value of rotor flux under no load condition. Note that we have made step changes in speed command. Nonetheless, our closed-loop system allows for fast forward-reverse operations and can keep the steady-state speed within $\pm$9rpm ($\pm$0.5%). In particular, Fig 4 shows that our closed-loop system can provide good speed control performance even at very low speed 18rpm (0.01pu). On the other hand, the experimental result in Fig 5 shows the performance of our closed-loop system under some load condition. By using the BLDCM which was coupled mechanically to the induction motor, a full load torque was applied at $t = t_A$ and removed at $t = t_B$, as indicated in Fig 5. In this while, we made a step change in speed command. We can see from Fig 5 that our closed-loop system can provide good speed control accuracy even under a full load condition and, furthermore, can reject the disturbance torque promptly.

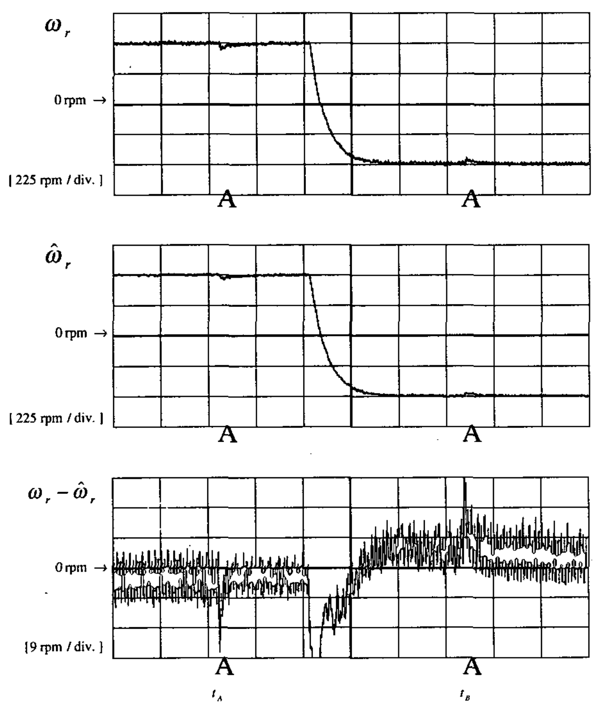

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Fig. 5. Forward-reverse operation under a full load condition (450rpm).

IV. Conclusion

In this paper, we have proposed an efficient on-line identification method for rotor resistance of induction motors without rotational transducers. We have demonstrated through various experiments that it can provide fairly good identification accuracy regardless of load conditions and, in turn, allows for high precision speed control of commercially available induction motors without rotational transducers.

References

[1] T.Ohtani, N.Takada, and K.Tanaka, "Vector Control of Induction Motor without Shaft Encoder," *IEEE Trans. Ind. Applicat.*, vol.28, no.1, pp.157-164, 1992.

[2] C.Shauder, "Adaptive Speed Identification for Vector Control of Induction Motors without Rotational Transducers," *IEEE Trans. Ind. Applicat.*, vol.28, no.5, pp.1054-1061, 1992.

[3] H.S.Yoo and I.J.Ha, "A Polar Coordinate-Oriented Method of Identifying Rotor Flux and Speed of Induction Motors without Rotational Transducers," *IEEE Trans. Control Syst. Technol.*, vol.4, no.3, pp.230-243, 1996.

[4] N.R.Klaes, "Parameter Estimation of an induction Machine with Regard to Dependencies on Saturation," *IEEE Trans. Ind. Applicat.*, vol.29, no.6, pp.1135-1140, 1993.

[5] D.I.Kim, I.J.Ha, and M.S.Ko, "Control of Induction Motors via Feedback Linearization with Input-output Decoupling," *Int. J. Contr.*, vol.51, pp.863-883, 1990.

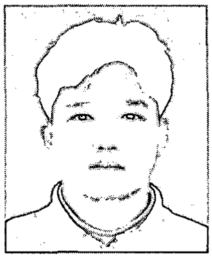
[6] G.S.Kim, I.J.Ha, and M.S.Ko, "Control of Induction Motors for Both High Dynamic Performance and High Power Efficiency," *IEEE Trans. Ind. Electron.*, vol.39, no.4, pp.323-333, 1992.

[7] H.Kubota, and K.Matsuse, "Speed Sensorless Field Oriented Control of Induction Motor with Rotor Resistance Adaptation," *IEEE Trans. Ind. Applicat.*, vol.30, no.5, pp.1219-1224,1994.

[8] M.Velez-Reyes, K.Minami, and G.C.Verghese, "Recursive Speed and Parameter Estimation for Induction Machines," *Conf. Rec. of 1989 IEEE IAS Annual Meeting*, pp.607-610.

[9] T.Kanmachi and I.Takahashi, "Sensor-less Speed Control of An Induction Motor with No Influence of Secondary Resistance Variation," *Conf. Rec. of 1993 IEEE IAS Annual Meeting*, pp.408-413.

[10] T. Soderstrom and P. Stoica, *System Identification*, Prentice Hall, 1989



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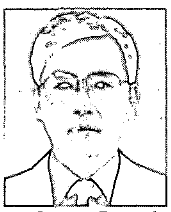

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