DECOUPLING CONTROL OF AN INDUCTION MOTOR WITH RECURSIVE ADAPTATION OF ROTOR RESISTANCE

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ABSTRACT - We propose a nonlinear feedback controller that can control the induction motors with high dynamic performance by means of decoupling of motor speed and rotor flux. The nonlinear feedback controller needs the information on some motor parameters. Among them, rotor resistance varies greatly with machine temperature. A new recursive adaptation algorithm for rotor resistance which can be applied to our nonlinear feedback controller is also presented in this paper. The recursive adaptation algorithm makes the estimated value of rotor resistance track its real value. Some simulation results show that the adaptation algorithm for rotor resistance is robust against the variation of stator resistance and mutual inductance. In addition, it is computationally simple and has small estimation errors. To demonstrate the practical significance of our results, we present some experimental results.

NOMENCLATURE

$v_{ds}(v_{qs})$	d(q)-axis stator voltage
$i_{ds}(i_{qs})$	d(q)-axis stator current
$\phi_{dr}(\phi_{qr})$	d(q)-axis rotor flux
ω_r	motor speed
ω_{sl}	slip speed
ω_s	angular speed of rotor flux
$R_s(R_r)$	stator (rotor) resistance
$L_s(L_r)$	stator (rotor) self-inductance
$L_{so}(L_{ro})$	stator (rotor) leakage inductance
M	mutual inductance between stator and rotor
J	moment of inertia
B	damping coefficient
K_T	torque constant $(=3pM/2L_r)$
p	number of pole pairs
σ	leakage coefficient $(=1-M^2/L_s L_r)$
a_0	$1/\sigma L_s$
a_1	$a_0(R_s+M^2R_r/L_r^2)$
a_2	$a_0 M R_r/L_r^2$
a_3	$a_0 M/L_r$
a_4	R_r/L_r
a_5	MR_r/L_r
a_6	B/J
a_7	1/J
T_L	external load torque
z^s	steady state value of the variable z
$\hat{a}_i$	estimated value of a_i

I. INTRODUCTION

Along with the rapid growth in microelectronics and power electronics technologies, various advanced control methods have been successfully implemented in real time and shown to be useful in controlling induction motors with high dynamic performance. The controllers in [1] - [6] can control induction motors to behave like dc motors. In particular, the controllers proposed recently in [4] - [6] can control induction motors with motor speed and rotor flux dynamically decoupled.

In this paper, we propose a nonlinear feedback controller that can control the induction motors with high dynamic performance by means of decoupling of motor speed and rotor flux. For power efficiency control or field weakening, the rotor flux may be reduced at light loads. Since the motor speed is dynamically decoupled from the rotor flux, this can be done successfully without any degradation of motor speed responses. Our nonlinear feedback controller needs the information on some motor parameters. Among them, rotor resistance varies greatly with machine temperature. In certain cases, it can increase by 100% over its ambient or nominal value. The values of rotor resistance at different machine temperatures can not be obtained by an off-line test and must be estimated instantly during the operation.

Some efficient estimation algorithms for rotor resistance can be found in [2], [6]-[13]. The estimation methods in [7] and [9] need the information on stator voltages. Because PWM inverters are commonly used as the voltage sources for the induction motors, such voltage sensing can deliver some error in the estimation of rotor resistance. In [2], the stator voltages are calculated from the stator voltage commands with dead time compensation instead of using voltage sensors. In [12], a fuzzy logic technique is used in order to estimate rotor resistance.

We present a new recursive adaptation algorithm for rotor resistance, which seems to have some advantages over the previous methods. For instance, our algorithm does not request voltage sensors. It does not depend on stator resistance and stator inductance. Some simulation results show that it is relatively robust against the variation of mutual inductance. In addition, it is computationally simple and has small estimation errors. The results in this paper improve our results in [11] proposed previously for high dynamic performance. To demonstrate the practical significance of our results, we present some experimental results as well as mathematical analyses.

II. DECOUPLING CONTROL OF MOTOR SPEED AND ROTOR FLUX

In this section, we describe our approach to control of induction motors whose dynamic equations are described, in the d-q synchronously rotating frame[14], as

$$
\begin{aligned}
\dot{i}_{ds} &= -a_1 i_{ds} + \omega_s i_{qs} + a_2\phi_{dr} + pa_3\omega_r\phi_{qr} + a_0 v_{ds}\\
\dot{i}_{qs} &= -a_1 i_{qs} - \omega_s i_{ds} + a_2\phi_{qr} - pa_3\omega_r\phi_{dr} + a_0 v_{qs}\\
\dot{\phi}_{dr} &= -a_4\phi_{dr} + (\omega_s - p\omega_r)\phi_{qr} + a_5 i_{ds}\\
\dot{\phi}_{qr} &= -a_4\phi_{qr} - (\omega_s - p\omega_r)\phi_{dr} + a_5 i_{qs}\\
\dot{\omega}_r &= -a_6\omega_r + a_7 K_T(\phi_{dr} i_{qs} - \phi_{qr} i_{ds}) - a_7 T_L
\end{aligned}
\tag{1}
$$

Here, the constants K_T and a_i, i = 0, . . . , 7 are the parameters of the induction motor. See the Nomenclature for the symbols and notations that appear frequently in our development. The angular speed of rotor flux ω_s in eq. (1) is chosen as

$$\omega_s = p\omega_r + \hat{a}_5 i_{qs} / \hat{\phi}_{dr} \tag{2}$$

from which it follows that the slip speed ω_{sl} becomes $\hat{a}_5 i_{qs}/\hat{\phi}_{dr}$. If $\hat{a}_5 (\equiv \hat{M}\hat{R}_r/\hat{L}_r)$ is equal to a_5 and $\hat{\phi}_{dr}$ is equal to ϕ_{dr}, then the q-axis rotor flux ϕ_{qr} will approach to zero. As the result, the dynamic equations in (1) and (2) can be approximated to

$$
\begin{aligned}
\dot{i}_{ds} &= -a_1 i_{ds} + \omega_s i_{qs} + a_2\phi_{dr} + a_0 v_{ds}\\
\dot{i}_{qs} &= -a_1 i_{qs} - \omega_s i_{ds} - pa_3\omega_r\phi_{dr} + a_0 v_{qs}\\
\dot{\phi}_{dr} &= -a_4\phi_{dr} + a_5 i_{ds}\\
\dot{\omega}_r &= -a_6\omega_r + a_7 K_T \phi_{dr} i_{qs} - a_7 T_l
\end{aligned}
\tag{3}
$$

To obtain the information on rotor flux, we adopt the following well-known rotor flux simulator.

$$\dot{\hat{\phi}}_{dr} = -\hat{a}_4\hat{\phi}_{dr} + \hat{a}_5 i_{ds} \tag{4}$$

If the output to be controlled is chosen as [6]

$$y \equiv \begin{bmatrix} y_1 \\ y_2 \end{bmatrix} = \begin{bmatrix} \phi_{dr}^2 \\ \omega_r \end{bmatrix} \tag{5}$$

then one can find a nonlinear feedback controller that decouples the reduced system consisting of (3) and (5).

$$\mathrm{u} \equiv \begin{bmatrix} v_{ds} \\ v_{qs} \end{bmatrix} = \begin{bmatrix} -(\omega_s i_{qs} + \hat{a}_5 i_{ds}^2 / \hat{\phi}_{dr})/\hat{a}_0 \\ p\omega_r (i_{ds} + \hat{a}_3\hat{\phi}_{dr})/\hat{a}_0 \end{bmatrix} + \bar{u}/\hat{\phi}_{dr} \tag{6}$$

where $\bar{u} \equiv [\bar{u}_1 \;\; \bar{u}_2]^T$ is the new input and $\hat{a}_0$, $\hat{a}_3$, $\hat{a}_5$ represent the estimated values of a_0, a_3, a_5, respectively.

Now, we will show that the input-output dynamic characteristics of the system given by (3) - (6) can be linear. Let the new state variable z be defined as

$$
\begin{aligned}
z &= [z_1^T \;\; z_2^T]^T = [z_{11} \;\; z_{12} \;\; z_{21} \;\; z_{22}]^T\\
&= [\phi_{dr} i_{ds} \;\; \phi_{dr}^2 \;\; \phi_{dr} i_{qs} \;\; \omega_r]^T
\end{aligned}
\tag{7}
$$

If $\hat{a}_0 = a_0$, $\hat{a}_3 = a_3$, $\hat{a}_5 = a_5$, and $\hat{\phi}_{dr} = \phi_{dr}$, then the system given by (3) - (6) can be represented in the new state space as

$$
\begin{aligned}
\dot{z} &= \begin{bmatrix} \dot{z}_1 \\ \dot{z}_2 \end{bmatrix} = \begin{bmatrix} A_1 z_1 + b\bar{u}_1 \\ A_2 z_2 + b\bar{u}_2 + LT_L \end{bmatrix}\\
y_i &= cz_i, \;\; i = 1, 2.
\end{aligned}
\tag{8}
$$

where

$$
A_1 = \begin{bmatrix} -a_1 - a_4 & a_2 \\ 2a_5 & -2a_4 \end{bmatrix},
$$

$$
A_2 = \begin{bmatrix} -a_1 - a_4 & 0 \\ a_7 K_l & -a_6 \end{bmatrix},
$$

$$b = [a_0 \;\; o]^T, \;\; c = [0 \;\; 1], \;\; L = [0 \;\; -a_7]^T \tag{9}$$

The block diagram representation of the decoupled linear system (8) is shown in Fig. 1.

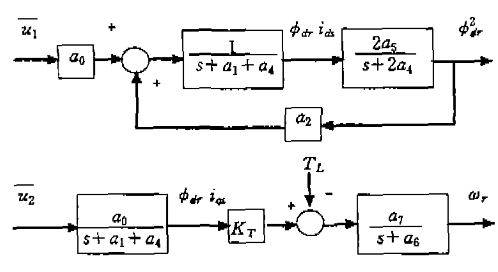


Fig 1. Block diagram of the decoupled linear system

Note that the responses of motor speed and rotor flux are dynamically decoupled and linear. Also note that the input-output dynamic characteristics of the system given by (3) - (6) are identical with those of the decoupled linear system (8). So, even if the rotor flux is reduced for power efficiency control at light loads or for field weakening at high speeds, the dynamic performance of the motor speed responses will not be deteriorated because of the complete decoupling of motor speed and rotor flux. For successful set-point tracking of motor speed and rotor flux, the new inputs $\bar{u}_1$ and $\bar{u}_2$ are chosen as follows[15]:

$$\bar{u}_1 = k_{p1}(\tilde{u}_1 - \hat{\phi}_{dr} i_{ds}) + k_{i1}\int_0^t (\tilde{u}_1 - \hat{\phi}_{dr} i_{ds})dt$$

$$\bar{u}_2 = k_{p2}(\tilde{u}_2 - \hat{\phi}_{dr} i_{qs}) + k_{i2}\int_0^t (\tilde{u}_2 - \hat{\phi}_{dr} i_{qs})dt \qquad (10)$$

where

$$\tilde{u}_1 = -k_{p3}\hat{\phi}_{dr}^2 + k_{i3}\int_0^t (\phi^* - \hat{\phi}_{dr}^2)dt$$

$$\tilde{u}_2 = -k_{p4}\omega_r + k_{i4}\int_0^t (\omega_r^* - \omega_r)dt \qquad (11)$$

Here, constants k_{ij}, k_{pj}, j=1, 2, 3, 4 are controller gains and ω_r^*, ϕ^* represent the command inputs for ω_r, $\hat{\phi}_{dr}^2$, respectively.

III. RECURSIVE ADAPTATION OF ROTOR RESISTANCE

As mentioned earlier, if the estimated values $\hat{a}_0$, $\hat{a}_3$, $\hat{a}_4$, and $\hat{a}_5$ which are used in the control inputs (2), (6), and the rotor flux simulator (4) are equal to their real values a_0, a_3, a_4, and a_5, respectively, the nonlinear dynamic equations of induction motors (1) with the control inputs (2), (6), and the rotor flux simulator (4) can be transformed into the decoupled linear system (8). By the way, the motor parameters a_0, a_3, a_4, and a_5 are composed entirely of M, L_s, L_r, and R_r. Accordingly, the accurate estimation of the motor parameters M, L_s, L_r, and R_r is of crucial importance to high dynamic performance. The motor inductances M, L_s(=M + L_{so}), and L_r (=M + L_{ro}) depend on rotor flux level and do not vary with machine temperature and load conditions. So, they should be exactly known at different rotor flux levels. In practice, they can be obtained by an off-line test. In what follows, we assume that $\hat{M} = M$, $\hat{L}_s = L_s$, and $\hat{L}_r = L_r$. This leads us to assume that $\hat{a}_0 = a_0$ and $\hat{a}_3 = a_3$.

On the other hand, rotor resistance varies greatly with machine temperature but is not influenced by rotor flux level. In certain cases, rotor resistance can increase by 100% over its ambient or nominal value. The values of rotor resistance at different machine temperatures can not be obtained by an off-line test and must be estimated instantly during the operation.

Now, we will consider the situation that the rotor resistance is estimated accurately (i.e. $\hat{R}_r = R_r$). Then, $\hat{\phi}_{dr}$ will be equal to ϕ_{dr}. From the first and second equations of (1) with the control input (6), we obtain

$$\dot{i}_{ds} = -a_1 i_{ds} + a_2\hat{\phi}_{dr} - \hat{a}_5 i_{ds}^2 / \hat{\phi}_{dr} + \hat{a}_0\bar{u}_1 / \hat{\phi}_{dr}$$
$$\dot{i}_{qs} = -a_1 i_{qs} - \hat{a}_5 i_{ds} i_{qs} / \hat{\phi}_{dr} + \hat{a}_0\bar{u}_2 / \hat{\phi}_{dr} \qquad (12)$$

Because $\dot{i}_{ds}$ and $\dot{i}_{qs}$ are zero in the steady-state, this leads to (13).

$$i_{ds}^s \dot{i}_{qs}^s - \dot{i}_{ds}^s i_{qs}^s = -a_2\hat{\phi}_{dr}^s i_{qs}^s + \frac{\hat{a}_0}{\hat{\phi}_{dr}^s}(\bar{u}_2^s i_{ds}^s - \bar{u}_1^s i_{qs}^s)$$
$$= 0 \qquad (13)$$

where the superscript s denotes the steady-state value. In the steady-state, the eq. (4) becomes

$$\hat{\phi}_{dr}^s = \hat{M} i_{ds}^s \qquad (14)$$

Because it is assumed that $\hat{L}_s = L_s$, $\hat{M} = M$, and $\hat{L}_r = L_r$, the motor parameter a_2 can be written as

$$a_2 = a_0 M R_r / L_r^2 = \hat{a}_0 \hat{M} R_r / \hat{L}_r^2 \qquad (15)$$

From (13) - (15), we have a_2

$$R_r = \frac{\hat{L}_r^2}{\hat{M}^2\hat{\phi}_{dr}^s}\left(\frac{\bar{u}_2^s}{i_{qs}^s} - \frac{\bar{u}_1^s}{i_{ds}^s}\right) \qquad (16)$$

By the way, the eq. (16) is obtained on the assumption that $\hat{R}_r$ is equal to R_r. So, we can have

$$\hat{R}_r = R_r = \frac{\hat{L}_r^2}{\hat{M}^2\hat{\phi}_{dr}^s}\left(\frac{\bar{u}_2^s}{i_{qs}^s} - \frac{\bar{u}_1^s}{i_{ds}^s}\right) \tag{17}$$

The eq. (17) is effective only in case that $\hat{R}_r = R_r$. However, it can be utilized to construct a recursive adaptation algorithm for R_r. Assume that the n-th estimated rotor resistance $\hat{R}_r$ (n) is applied to (2), (4), and (6) and then, in the steady state, the (n+1)th estimated rotor resistance $\hat{R}_r$ (n+1) is calculated using (17). Then, we find that $\hat{R}_r$ (n+1) is closer to R_r than $\hat{R}_r$ (n), which will be shown in later simulations and experiments. Now, we describe our recursive adaptation algorithm for R_r

Step 1) Calculate $\hat{R}_r$ using (17) and update $\hat{a}_4$ and $\hat{a}_5$ in (2), (4), and (6).

Step 2) Wait until all the state variables reach the steady state.

Step 3) Go to step 1.

Our new recursive adaptation algorithm for rotor resistance has some advantages over the previous methods. As can be seen from (17), it does not need the information on stator voltage, which is required in the prior works [7] and [9]. Therefore, its estimation accuracy is not affected by the use of a PWM inverter as a voltage source. Furthermore, it is independent of stator resistance and stator inductance. However, the performances of our algorithm in (17) depend on the accuracy of $\hat{M}$ and $\hat{L}_r$. In addition, it requires that the stator voltages v_{ds} and v_{qs} in (1) should be equal to their commands v_{ds} and v_{qs} in (6), respectively. Some dead-time compensation techniques can solve this problem [2].

IV. SIMULATION AND EXPERIMENTAL RESULTS

The performances of our control scheme developed in the preceding sections were studied through some simulations and experiments. For simulation and experimental work, we have chosen a two pole squirrel cage induction motor whose motor data are listed in Table 1.

First, we present the simulation results, which show good performance of our controller with R_r adaptation. For this aim, the value of $\hat{R}_r$ was initially assumed to be 1.425 Ω. This corresponds to a 25% estimation error in R_r. In this simulation, we assumed that $\hat{R}_s = R_s$, $\hat{M} = M$, and the induction motor was driven with rated rotor flux, rated load torque, and motor speed of 30r/min. Our decoupling controller was executed every 0.1msec. At t = 1sec, our recursive adaptation algorithm in (17) was started. It was calculated and updated every 0.5sec. The simulation result for this case is shown in Fig. 2. We can see that $\hat{R}_r$ approaches to R_r step by step. Observe that the motor speed response is disturbed by the abrupt change in $\hat{R}_r$. Thus, in the presence of large estimation error in R_r, our controller fails to decouple the motor speed and the rotor flux.

Table 1. Data of the induction motor used for simulations and experiments

Nameplate Data	Nominal	Parameters
220 V 50 Hz	R_s	1.09 Ω
3 phase	R_r	1.14 Ω
Y connected	L_s	100 mH
2 poles	L_r	100 mH
Rated power 600W	L_{ro} (L_{so})	7.7 mH
Rated speed 3000r/min	M	92.3 mH
Rated rotor flux 0.3Wb	J	3.2×10^{-4} kgm^2
Rated current 4.2A(r m s)	B	4.2×10^{-4} kgm^2/s

Second, the recursive adaptation algorithm in (17) was updated every 0.5msec and the rate of change of $\hat{R}_r$ was restricted within 0.2 Ω/sec. As can be seen in Fig. 3, the ripple in the motor speed response was decreased. We can see that the slow change in $\hat{R}_r$ minimize the influence on the motor speed response and it took a shorter time for R_r adaptation in the case of Fig. 3 than in the case of Fig.2.

Third, we investigate the robustness of our recursive adaptation algorithm with respect to R_s and M. The value of $\hat{R}_s$ was assumed to be 1.635Ω and to be unchanged throughout this simulation. This corresponds to a 50% estimation error in R_s. The simulation result for this case is shown in Fig. 4 (a). This simulation result shows that our recursive adaptation algorithm is perfectly robust against the variation of R_s. This is because the recursive adaptation algorithm in (17) does not need the information on R_s. Now, we will consider the situation that $\hat{M} = 1.2M$. We can see from Fig. 4 (b) that the 20% estimation error in $\hat{M}$ caused about 1.44% increase in the R_r estimation error. As for this simulation result, we can say that our recursive adaptation algorithm is relatively robust against the uncertainty in M. This is because the rotor leakage inductance L_{ro} is negligibly small compared with $\hat{L}_r$ (= $\hat{M} + L_{ro}$) and $\hat{L}_r^2$ is divided by $\hat{M}^2$ in eq. (17).

Fourth, our control algorithm was implemented on a DSP chip TMS320C25. The motor speed was detected by an optical encoder whose resolution was 4000 pulses/rev. The decoupling control algorithm was executed every 0.1msec and the recursive adaptation algorithm in (17) was updated every 0.5msec. The experimental result for this case is shown in Fig. 5. We can see that the ripple in the motor speed response is negligible. As can be seen from Fig. 3 and Fig. 5, the experimental results agree well with the simulation results. Slight differences between the simulation and experimental results are unavoidable due to imperfect hardware implementation, some uncertainties in motor data, and etc.

Finally, we show that our controller with adaptation of rotor resistance can guarantee almost exact decoupling of motor speed and rotor flux. Fig. 6 shows the experimental results for the case of step change in motor speed from 1500r/min to 3000r/min after our R_r adaptation algorithm was executed. The rotor flux was changed from 0.5p.u. (= 0.0225Wb^2) to 1.0p.u. (= 0.09Wb^2) at t = 0.4sec and from 1.0p.u. to 0.5p.u. at t = 1.4sec. However, the motor speed response is not affected by the abrupt change in the rotor flux because of the complete decoupling of motor speed and rotor flux.

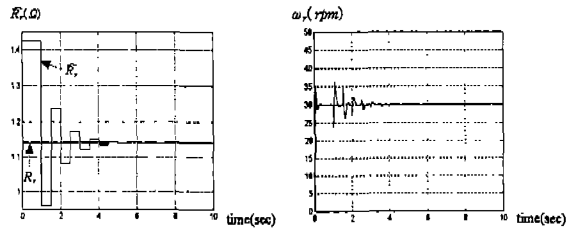


Fig. 2. Simulation of R_r adaptation and motor speed ω_r.

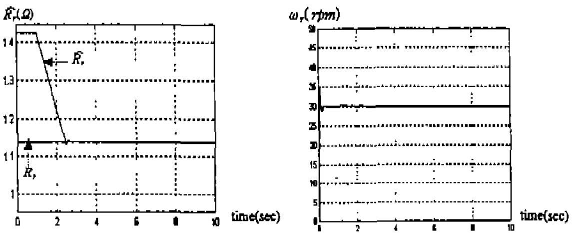
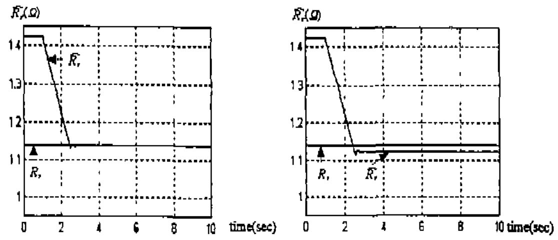


Fig. 3. Simulation of R_r adaptaion with updating period of 0.5msec.



(a) (b)

Fig. 4. Simulation of R_r adaptaion (a) when $\hat{R}_s = 1.5R_s$ and (b) when $\hat{M} = 1.2M$.

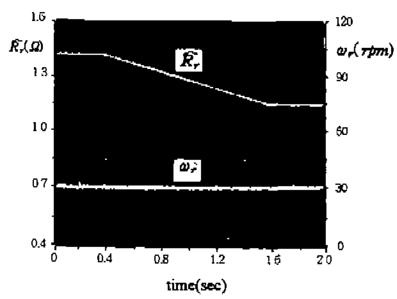


Fig.5.Experimental results of R_r adapation and motor speed ω_r.

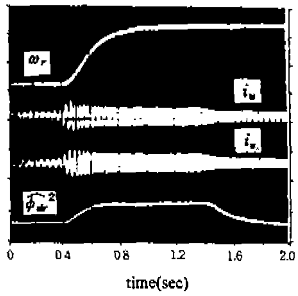


Fig.6. Experimental results of motor speed of motor speed (700r/min/div), u(v) phase stator currents (10A/div), and rotor flux (0.05 Wb^2/div).

V. CONCLUSION

In this paper, a decoupling controller with adaptation of rotor resistance has been proposed. Some simulation results showed that the recursive adaptation algorithm for rotor resistance proposed in this paper was robust against the variation of stator resistance and mutual inductance. In addition, our adaptation algorithm is computationally simple and has small estimation errors. However, the adaptation algorithm was designed to be applied to our decoupling controller. So, further research will focus on studies related to its application to the general vector control scheme. The mathematical proof for the convergence of the proposed adaptation algorithm is still underway.

ACKNOWLEDGEMENT

This work was sponsored by the 1997 research support program of R & D Management Center for Energy and Resources.

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