Nonlinear Adaptive Speed and Torque Control of Induction Motors with Unknown Rotor Resistance

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***Abstract*—In this paper, we propose a nonlinear adaptive speed and torque controller of induction motors with unknown rotor resistance. All the system parameters except rotor resistance are assumed to be known, and only the stator currents and rotor speed are assumed to be available. The desired speed and torque should be a smooth bounded function. A complete proof of the global stability without singularity is given, and the output error will converge to zero asymptotically. Finally, the simulation and experimental results will be given to demonstrate the effectiveness of the proposed controller.**

***Index Terms*—Adaptive tracking control, induction motors, partial-state feedback, unknown rotor resistance.**

NOMENCLATURE

$V_a(V_b)$	a-axis (b-axis) stator voltage.
$I_a(I_b)$	a-axis (b-axis) stator current.
$\psi_a(\psi_b)$	a-axis (b-axis) rotor flux.
ω_r	Mechanical angular speed of rotor.
R_s	Stator resistance.
R_r	Equivalent rotor resistance.
L_s	Stator self-inductance.
L_r	Equivalent rotor self-inductance.
M	Mutual inductance.
p	Number of pole pairs.
J	Rotor inertia.
D	Damping coefficient.
T_L	Load torque.
k_T	Torque constant $[= (3pM/2L_r)]$.
L_o	$L_r^2/M\ [L_s - M^2/L_r]$.
β_1	$R_s L_r^2/M$.
β_2	pL_r.
β_3	L_r^2/M.

I. INTRODUCTION

DUE TO THE highly nonlinear dynamics, the control of an induction motor has become the benchmark of nonlinear control theory and has been extensively explored over the past decade. In the early years, much of the research was devoted to the control problems with full-state feedback, and either obtained the approximate stator–rotor flux decoupling or achieved exact input–output linearlization. So far, there have been several methods developed, such as the nonlinear geometric techniques [5], [8], or their adaptive version [9]. Although these control schemes could achieve satisfactory performance with mildly complex calculations, they have exhibited the problem with singularity and, in practice, the full-state feedback is undesirable since the rotor fluxes are very difficult to measure accurately.

In recent years, due to the great advances in nonlinear control theory, the observer-based controller has become one of the most commonly used schemes in induction motor control and an enormous amount of results on this have been published. Instead of measuring the rotor fluxes directly, the flux observers were constructed to furnish flux estimates needed in the controller. First of all, Kanellakopoulos *et al.* [4] proposed a nonlinear flux-observer-based speed control with known motor parameters. Thereafter, a variety of related research has also been presented [1], [10], [14].

All the control schemes mentioned above, whether full-state or partial-state feedback, are greatly dependent on the exact knowledge of the motor parameters. However, these are very sensitive to the variation of temperature and the environment of operation. Therefore, to attain the high control performance with sufficient robustness, the contemporary design of the induction motor controller should take the uncertainties of motor parameters into account. In [11], Marino *et al.* proposed a clever adaptive observer scheme which could compensate for the unknown but bounded rotor resistance. Later, Hu *et al.* [3] slightly modified the above observer structure and provided a rigorous proof for speed/position control with an uncertain mechanical subsystem. However, it was still subject to a singularity problem when the magnitude of the estimated rotor flux is zero.

For the torque control problem, many results have been presented and they have obtained satisfactory performance. In [2], Espinosa *et al.* proposed a globally stable output-feedback speed tracking controller for induction motors without any estimation of the rotor fluxes. However, this scheme needs to know exact values of the motor parameters.

In this paper, we will propose a new nonlinear adaptive speed and torque controller with unknown rotor resistance, which is free of the singularity problem. Only the stator currents and rotor speed are assumed available for the design. A complete Lyapunov-based proof is provided which shows that the errors of speed and torque tracking will asymptotically approach to zero.

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The simulation as well as the experimental results are also presented to demonstrate the performance and effectiveness of the hereby developed controller.

The organization of this paper is as follows. Section I briefly gives an introduction about induction motors and surveys some related research. In Section II, the mathematical model of an induction motor will be presented in detail. The state observers and controller design are developed in Section III, in which the asymptotical speed and torque tracking will be proved, and the simulation results of the control performance are given in Section IV. Finally, we will make some conclusions in Section V.

II. Preliminary and Problem Formulation

A. Mathematical Modeling

If the induction motor is assumed to be linear, i.e., it is never in the saturation region, and the waveform of air-gap magnetomotive force (MMF) is nearly sinusoidal, then the dynamic equations of the induction motor can be expressed as follows [6], [3]:

$$\begin{aligned} L_o\dot{I}_a &= -MR_rI_a - \beta_1 I_a + R_r\psi_a + \beta_2\omega_r\psi_b + \beta_3 V_a \\ L_o\dot{I}_b &= -MR_rI_b - \beta_1 I_b - \beta_2\omega_r\psi_a + R_r\psi_b + \beta_3 V_b \\ L_r\dot{\psi}_a &= -R_r\psi_a + MR_rI_a - \beta_2\omega_r\psi_b \\ L_r\dot{\psi}_b &= -R_r\psi_b + MR_rI_b + \beta_2\omega_r\psi_a \\ T_e &= k_T(\psi_a I_b - \psi_b I_a) \end{aligned} \tag{1}$$

where $L_o, \beta_1, \beta_2,$ and β_3 are constants defined in the Nomenclature. In the above equations, $(I_a, I_b, \psi_a, \psi_b)$ are referred to as the well-known stator fixed a–b model as presented in [3]. Moreover, to facilitate our subsequent controller design which intends to relax the need of the knowledge of the rotor resistance R_r, the terms associated with R_r are separated from the other terms in the system dynamic equations.

B. Problem Statement

Given the above equations, the control objective is now to design a nonlinear adaptive speed and torque servo controller of the induction motor with unknown rotor resistance which guarantees the asymptotical stability of the closed-loop system. In this paper, we will focus on the speed and torque tracking control, or, servo control with satisfactory performance, i.e., given the desired speed and torque trajectories, the controller should be able to control the induction motor to track those trajectories as closely as possible. Moreover, we only need to know the upper and lower bounds of the rotor resistance R_r instead of its exact knowledge.

III. Induction Motors Control

A. Problem Description

Before we proceed with the controller design and its analysis, some practical assumptions will be made as follows.

Assumptions:

A1) All the parameters of the motor except the rotor resistance R_r are known.

A2) The stator currents and rotor speed are measurable signals.

A3) The rotor resistance R_r is unknown but its upper and lower bounds are priorly known, i.e.,

$$\underline{R}_r < R_r < \overline{R}_r$$

for some known constants $\underline{R}_r$ and $\overline{R}_r$.

A4) The desired torque should be bounded continuously differentiable, i.e., $T_d \in BC^1$.

A5) The load torque T_L is a function which satisfies that, if T_e is bounded, then the rotor speed ω_r is also bounded.

Remark 1: In general, the load torque is a function of the rotor speed. In this paper, we will assume that the load torque can be expressed in the following form:

$$T_L = \mu_0 \cdot s(k, \omega_r) + \mu_1\omega_r + \mu_2\dot{\omega}_r$$

for some constants $k, \mu_0, \mu_1, \mu_2 > 0$, and the sigmoidal function $s(k, \omega_r) = (1 - e^{-k\cdot\omega_r}) \cdot (1 + e^{-k\cdot\omega_r})^{-1}$. This form includes the viscous friction, coulomb friction, and mechanical load. From the torque–load relation $T_e - T_L = J\dot{\omega}_r + D\omega_r$, where J is the rotor inertia, and D is the damping coefficient.

Due to the fact that the flux signals are not measurable (because lack of practical flux sensors), the nonlinear adaptive controller proposed here will have to involve state observation and, hence, the observer design.

B. State Observer Design

From the dynamic equations in (1), it is clear that many terms on the right-hand side (RHS) of the expressions involving $L_o\dot{I}_a$, $L_o\dot{I}_b$ and those involving $L_r\dot{\psi}_a$, $L_r\dot{\psi}_b$ are the same. Therefore, motivated by the observer scheme in [3], in addition to the flux observers, we construct the current observers which takes the unknown rotor resistance into account. Define the observation errors as follows:

$$\begin{aligned} \tilde{I}_a &= I_a - \hat{I}_a \\ \tilde{I}_b &= I_b - \hat{I}_b \\ \tilde{\psi}_a &= \psi_a - \hat{\psi}_a, \\ \tilde{\psi}_b &= \psi_b - \hat{\psi}_b \\ \tilde{R}_r &= R_r - \hat{R}_r \end{aligned}$$

and the following auxiliary weighted observation quantities:

$$\tilde{Z}_a = L_o\tilde{I}_a + L_r\tilde{\psi}_a \quad \tilde{Z}_b = L_o\tilde{I}_b + L_r\tilde{\psi}_b. \tag{2}$$

Let the state observers be designed as follows:

$$\begin{aligned} L_o\dot{\hat{I}}_a &= k_0\tilde{I}_a - M\hat{R}_rI_a - \beta_1 I_a + \hat{R}_r\hat{\psi}_a + \beta_2\omega_r\hat{\psi}_b + \beta_3 V_a \\ &\quad + u_{o1} + u_{c1} \\ L_o\dot{\hat{I}}_b &= k_0\tilde{I}_b - M\hat{R}_rI_b - \beta_1 I_b - \beta_2\omega_r\hat{\psi}_a + \hat{R}_r\hat{\psi}_b + \beta_3 V_b \\ &\quad + u_{o2} + u_{c2} \\ L_r\dot{\hat{\psi}}_a &= -\hat{R}_r\hat{\psi}_a + M\hat{R}_rI_a - \beta_2\omega_r\hat{\psi}_b + u_{o3} + u_{c3} \\ L_r\dot{\hat{\psi}}_b &= -\hat{R}_r\hat{\psi}_b + M\hat{R}_rI_b + \beta_2\omega_r\hat{\psi}_a + u_{o4} + u_{c4} \end{aligned} \tag{3}$$

where $k_0 > 0$ is a design parameter which is quite important for subsequent stability analysis, and $u_{oi}, u_{ci}, 1 \le i, j \le 4$, are some auxiliary signals to be designed subsequently.

Combining (1) and (3), we can obtain the following sets of dynamic equations of observation errors:

$$\begin{aligned} L_o\dot{\tilde{I}}_a &= -k_0\tilde{I}_a - M\tilde{R}_rI_a + R_r\tilde{\psi}_a + \tilde{R}_r\hat{\psi}_a + \beta_2\omega_r\tilde{\psi}_b \\ &\quad - u_{o1} - u_{c1} \\ L_o\dot{\tilde{I}}_b &= -k_0\tilde{I}_b - M\tilde{R}_rI_b + R_r\tilde{\psi}_b + \tilde{R}_r\hat{\psi}_b - \beta_2\omega_r\tilde{\psi}_a \\ &\quad - u_{o2} - u_{c2} \\ L_r\dot{\tilde{\psi}}_a &= -R_r\tilde{\psi}_a - \tilde{R}_r\hat{\psi}_a + M\tilde{R}_rI_a - \beta_2\omega_r\tilde{\psi}_b - u_{o3} - u_{c3} \\ L_r\dot{\tilde{\psi}}_b &= -R_r\tilde{\psi}_b - \tilde{R}_r\hat{\psi}_b + M\tilde{R}_rI_b + \beta_2\omega_r\tilde{\psi}_a - u_{o4} - u_{c4} \end{aligned} \tag{4}$$

whereby $\dot{\tilde{Z}}_a$ and $\dot{\tilde{Z}}_b$ can be straightforwardly derived as

$$\begin{aligned} \dot{\tilde{Z}}_a &= -k_0\tilde{I}_a - u_{o1} - u_{c1} - u_{o3} - u_{c3} \\ \dot{\tilde{Z}}_b &= -k_0\tilde{I}_b - u_{o2} - u_{c2} - u_{o4} - u_{c4} \end{aligned} \tag{5}$$

which clearly are computable since $\tilde{I}_a = I_a - \hat{I}_a$ are both available signals and u_{oi}, u_{ci}, and $1 \le i, j \le 4$ are all design signals. Comparing (4) and (5), it can be easily seen that the RHS of (5) does not depend on any unknown parameters or signals, unlike the situation for (4). This motivates us to build an observer for $\tilde{Z}_a, \tilde{Z}_b$, which leads to additional observation errors η_a, η_a defined as follows:

$$\eta_a = \tilde{Z}_a - \zeta_a + \epsilon_a \quad \eta_b = \tilde{Z}_b - \zeta_b + \epsilon_b$$

where ϵ_a, ϵ_b indicate the errors caused by the initial conditions, i.e.,

$$\begin{aligned} \epsilon_a &= \eta_a(0) + \zeta_a(0) - \tilde{Z}_a(0) \\ \epsilon_b &= \eta_b(0) + \zeta_b(0) - \tilde{Z}_b(0). \end{aligned}$$

Because $\tilde{Z}_a(0), \tilde{Z}_b(0)$ are not priorly known, ϵ_a and ϵ_b can be regarded as some unknown constants.

Using the related terms presented above, we now design the observer inputs u_{o1}, u_{o2} to achieve the necessary stability property as follows:

$$\begin{aligned} u_{o1} &= -\frac{\beta_2}{L_r}\,\omega_rL_o\tilde{I}_b - \frac{\hat{R}_r}{L_r}\,(L_o\tilde{I}_a - \zeta_a) \\ u_{o2} &= \frac{\beta_2}{L_r}\,\omega_rL_o\tilde{I}_a - \frac{\hat{R}_r}{L_r}\,(L_o\tilde{I}_b - \zeta_b). \end{aligned} \tag{6}$$

Remark 2: Note that the terms introduced in the design of u_{o1} and u_{o2} are to cancel the unmeasurable terms in the observer error dynamics so that a well-known Lyapunov stability analysis can be made.

Likewise, the input signals u_{o3} and u_{o4} are designed as follows:

$$\begin{aligned} u_{o3} &= -k_0\tilde{I}_a - \frac{\beta_2}{L_r}\,\omega_r\tilde{I}_b - u_{o1} - u_{c1} \\ u_{o4} &= -k_0\tilde{I}_b + \frac{\beta_2}{L_r}\,\omega_r\tilde{I}_a - u_{o2} - u_{c2}. \end{aligned} \tag{7}$$

Given the input design of $u_{oi}, i = 1, 2, 3, 4$, for the observer, we can simplify the dynamic equations (5) as

$$\dot{\tilde{Z}}_a = \frac{\beta_2}{L_r}\,\omega_r\tilde{I}_b - u_{c3}$$

$$\dot{\tilde{Z}}_b = -\frac{\beta_2}{L_r}\,\omega_r\tilde{I}_a - u_{c4}.$$

Now, construct a Lyapunov function candidate as follows:

$$\begin{aligned} V_0 &= \frac{1}{2}\,L_o\tilde{I}_a^2 + \frac{1}{2}\,L_o\tilde{I}_b^2 + \frac{1}{2}\,\tilde{Z}_a^2 + \frac{1}{2}\,\tilde{Z}_b^2 + \frac{1}{2}\,R_r(\eta_a - \epsilon_a)^2 \\ &\quad + \frac{1}{2}\,R_r(\eta_b - \epsilon_b)^2 + \frac{1}{2}\,\Gamma_r^{-1}\tilde{R}_r^2, \qquad \Gamma_r^{-1} > 0 \end{aligned}$$

and substitute u_{o1}, u_{o2}, u_{o3}, and u_{o4} into its time derivative $\dot{V}_0$, then, we can obtain the following result:

$$\begin{aligned} \dot{V}_0 &= -k_0\tilde{I}_a^2 - k_0\tilde{I}_b^2 - u_{c1}\tilde{I}_a - u_{c2}\tilde{I}_b - u_{c3}\tilde{Z}_a - u_{c4}\tilde{Z}_b \\ &\quad + \tilde{R}_r(\Omega_0 - \Gamma_r^{-1}\dot{\hat{R}}_r) + R_r(\eta_a - \epsilon_a)\left(\dot{\eta}_a + \frac{1}{L_r}\,\tilde{I}_a\right) \\ &\quad + R_r(\eta_b - \epsilon_b)\left(\dot{\eta}_b + \frac{1}{L_r}\,\tilde{I}_b\right) \end{aligned}$$

where

$$\begin{aligned} \Omega_0 = \ & \tilde{I}_a\left(-MI_a + \hat{\psi}_a - \frac{1}{L_r}(L_o\tilde{I}_a - \zeta_a)\right) \\ & + \tilde{I}_b\left(-MI_b + \hat{\psi}_b - \frac{1}{L_r}\,(L_o\tilde{I}_b - \zeta_b)\right) \end{aligned}$$

and $u_{c1}, u_{c2}, u_{c3}, u_{c4}, \dot{\eta}_a, \dot{\eta}_b$ (in fact, $\dot{\zeta}_a\ \dot{\zeta}_b$) are some auxiliary inputs yet to be specified.

C. Nonlinear Adaptive Controller Design

Define the state variables in a more compact form as follows:

$$u = \begin{bmatrix} u_1 \\ u_2 \end{bmatrix} = \begin{bmatrix} V_a \\ V_b \end{bmatrix}$$

$$x = \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{bmatrix} = \begin{bmatrix} I_a \\ I_b \\ \psi_a \\ \psi_b \end{bmatrix}$$

$$e = \begin{bmatrix} e_1 \\ e_2 \\ e_3 \\ e_4 \end{bmatrix} = \begin{bmatrix} I_a - I_{ad} \\ I_b - I_{bd} \\ \psi_a - \psi_{ad} \\ \psi_b - \psi_{bd} \end{bmatrix}$$

then, the state equations can be written concisely in a matrix form as follows:

$$\dot{x} = (A_0 + A_1)x + Bu$$

where A_0, A_1, and B are defined at the bottom of the next page.

Remark 3: From the foregoing compact redefinition, we can observe that, A_0 is a negative-definite matrix, so that $x^TA_0x < 0$.

1) Torque Control:

Lemma 1: If the desired torque T_d can be expressed in the following form:

$$T_d = k_T(\psi_{ad}I_{bd} - \psi_{bd}I_{ad})$$

then the bound on the difference between the electrical torque and the desired torque can be expressed as follows:

$$|T_e - T_d| \le \frac{k_T}{2}\,(\|e\|^2 + 2\|e\|\,\|x_d\|) \tag{8}$$

where

$$x_d = \begin{bmatrix} x_{1d} \\ x_{2d} \\ x_{3d} \\ x_{4d} \end{bmatrix} = \begin{bmatrix} I_{ad} \\ I_{bd} \\ \psi_{ad} \\ \psi_{bd} \end{bmatrix}.$$

Proof: Please refer to [14].

With this lemma, it is quite obvious that the following corollary will naturally hold.

Corollary 1: If the state error $e \to 0$ as $t \to \infty$, then $T_e \to T_d$ as $t \to \infty$. □

As a result, we can realize that the objective of torque tracking will be achieved once the state tracking is achieved. This, thus, allows us to concentrate on the dynamics of state tracking error as derived in the following:

$$\begin{aligned} \dot{e} &= (A_0 + A_1)e - \dot{x}_d + (A_0 + A_1)x_d + Bu \\ &= A_0 e + C \end{aligned} \tag{9}$$

where

$$C = A_1 e - \dot{x}_d + (A_0 + A_1)x_d + Bu = [c_1\ c_2\ c_3\ c_4]^T$$

with

$$\begin{aligned} c_1 = {} & \frac{R_r}{L_o}\,\psi_a + \frac{\beta_2}{L_o}\,\omega_r\psi_b - \dot{I}_{ad} - \left(\frac{M}{L_o}\,R_r + \frac{\beta_1}{L_o}\right) I_{ad} \\ & + \frac{\beta_3}{L_o}\,u_1 \\ c_2 = {} & \frac{R_r}{L_o}\,\psi_b - \frac{\beta_2}{L_o}\,\omega_r\psi_a - \dot{I}_{bd} - \left(\frac{M}{L_o}\,R_r + \frac{\beta_1}{L_o}\right) I_{bd} \\ & + \frac{\beta_3}{L_o}\,u_2 \\ c_3 = {} & -\dot{\psi}_{ad} + \frac{M}{L_r} R_r I_a - \frac{R_r}{L_r}\,\psi_{ad} - \frac{\beta_2}{L_r}\,\omega_r\psi_b \\ c_4 = {} & -\dot{\psi}_{bd} + \frac{M}{L_r}\,R_r I_b + \frac{\beta_2}{L_r}\,\omega_r\psi_a - \frac{R_r}{L_r}\,\psi_{bd}. \end{aligned}$$

In the above equations, the input voltages u_1 and u_2 can be used to completely eliminate the terms of c_1 and c_2. However, there are not enough inputs which can be used to cancel c_3 and c_4. Therefore, we will regard the desired currents and fluxes as the pseudo inputs and design them properly so as to make the error system asymptotically stable.

Now, choosing the desired fluxes and currents as follows:

$$\begin{aligned} \dot{\psi}_{ad} &= -\left(\frac{\beta_2}{L_r}\,\omega_r + \frac{MT_d}{\gamma^2 L_r k_T}\,\hat{R}_r\right)\psi_{bd} \\ \dot{\psi}_{bd} &= \left(\frac{\beta_2}{L_r}\,\omega_r + \frac{MT_d}{\gamma^2 L_r k_T}\,\hat{R}_r\right)\psi_{ad} \\ I_{ad} &= -\frac{T_d}{\gamma^2 k_T}\,\psi_{bd} + \frac{1}{M}\,\psi_{ad} \\ I_{bd} &= \frac{T_d}{\gamma^2 k_T}\,\psi_{ad} + \frac{1}{M}\,\psi_{bd} \end{aligned} \tag{10}$$

for some constant $\gamma > 0$, we can directly get the following result:

$$c_3 = -\frac{MT_d}{\gamma^2 L_r k_T}\,\tilde{R}_r\psi_{bd} + \frac{M}{L_r}\,R_r e_1 - \frac{\beta_2}{L_r}\,\omega_r e_4$$

$$A_0 = \begin{bmatrix} -\left(\frac{M}{L_o}\,R_r + \frac{\beta_1}{L_o}\right) & 0 & 0 & 0 \\ 0 & -\left(\frac{M}{L_o}\,R_r + \frac{\beta_1}{L_o}\right) & 0 & 0 \\ 0 & 0 & -\frac{1}{L_r}\,R_r & 0 \\ 0 & 0 & 0 & -\frac{1}{L_r}\,R_r \end{bmatrix}$$

$$A_1 = \begin{bmatrix} 0 & 0 & \frac{1}{L_o}\,R_r & \frac{\beta_2}{L_o}\,\omega_r \\ 0 & 0 & -\frac{\beta_2}{L_o}\,\omega_r & \frac{1}{L_o}\,R_r \\ \frac{M}{L_r}\,R_r & 0 & 0 & -\frac{\beta_2}{L_r}\,\omega_r \\ 0 & \frac{M}{L_r}\,R_r & \frac{\beta_2}{L_r}\,\omega_r & 0 \end{bmatrix}$$

$$B = \begin{bmatrix} \frac{\beta_3}{L_o} & 0 \\ 0 & \frac{\beta_3}{L_o} \\ 0 & 0 \\ 0 & 0 \end{bmatrix}$$

$$c_4 = \frac{MT_d}{\gamma^2 L_r k_T}\,\tilde{R}_r\psi_{ad} + \frac{M}{L_r}\,R_r e_2 + \frac{\beta_2}{L_r}\,\omega_r e_3.$$

It can, in fact, be verified that the choice (10) will satisfy the condition in *Lemma 1*, namely, $T_d = k_T(\psi_{ad}I_{bd} - \psi_{bd}I_{ad})$. Besides, the desired fluxes and currents are also subject to the following condition.

Corollary 2: If T_d is a bounded function, then $(I_{ad}, I_{bd}, \psi_{ad}, \psi_{bd})$, namely, the desired fluxes and currents, are all bounded functions. □

Proof: Please refer to [14].

Now, we are ready to introduce our control law as well as to analyze the resulting system stability. Select the Lyapunov function candidate as follows:

$$V_1 = \frac{1}{2}\,e^T G e, \qquad G = \mathrm{diag}[g_1,\, g_1,\, g_2,\, g_2]$$

for some constants $g_1,\, g_2 > 0$, then its first-order time derivative can be found as follows:

$$\begin{aligned}
\dot{V}_1 &= e^T\dot{e}\\
&= e^T G A_0 e + e^T G C\\
&= e^T G A_0 e + g_1 e_1\\
&\quad\cdot\left(\frac{R_r}{L_o}\,\psi_a + \frac{\beta_2}{L_o}\,\omega_r\psi_b - \dot{I}_{ad} - \left(\frac{M}{L_o}\,R_r + \frac{\beta_1}{L_o}\right)I_{ad} + \frac{\beta_3}{L_o}\,u_1\right)\\
&\quad + g_1 e_2\left(\frac{R_r}{L_o}\,\psi_b - \frac{\beta_2}{L_o}\,\omega_r\psi_a - \dot{I}_{bd} - \left(\frac{M}{L_o}R_r + \frac{\beta_1}{L_o}\right)I_{bd}\right.\\
&\qquad\qquad \left. + \frac{\beta_3}{L_o}\,u_2\right)\\
&\quad + g_2 e_3\left(-a\tilde{R}_r\psi_{bd} + \frac{M}{L_r}\,R_r e_1 - \frac{\beta_2}{L_r}\,\omega_r e_4\right)\\
&\quad + g_2 e_4\left(a\tilde{R}_r\psi_{ad} + \frac{M}{L_r}\,R_r e_2 + \frac{\beta_2}{L_r}\,\omega_r e_3\right)
\end{aligned}\tag{11}$$

where $a = MT_d/\gamma^2 L_r k_T$. From the definitions of I_{ad} and I_{bd}, it is clear that $\dot{I}_{ad}$ and $\dot{I}_{bd}$ are both measurable since they can be expressed as follows:

$$\begin{aligned}
\dot{I}_{ad} &= -\frac{\dot{T}_d}{\gamma^2 k_T}\,\psi_{bd} - \frac{T_d}{\gamma^2 k_T}\,\dot{\psi}_{bd} + \frac{1}{M}\,\dot{\psi}_{ad}\\
&= -\frac{T_d}{\gamma^2 k_T}\left(\frac{\beta_2}{L_r}\,\omega_r + a\hat{R}_r\right)\psi_{ad}\\
&\quad - \left(\frac{\dot{T}_d}{\gamma^2 k_T} + \frac{\beta_2}{ML_r}\,\omega_r + \frac{T_d}{\gamma^2 L_r k_T}\,\hat{R}_r\right)\psi_{bd}\\
\dot{I}_{bd} &= \frac{\dot{T}_d}{\gamma^2 k_T}\,\psi_{ad} + \frac{T_d}{\gamma^2 k_T}\,\dot{\psi}_{ad} + \frac{1}{M}\,\dot{\psi}_{bd}\\
&= -\frac{T_d}{\gamma^2 k_T}\left(\frac{\beta_2}{L_r}\,\omega_r + a\hat{R}_r\right)\psi_{bd}\\
&\quad + \left(\frac{\dot{T}_d}{\gamma^2 k_T} + \frac{\beta_2}{ML_r}\,\omega_r + \frac{T_d}{\gamma^2 L_r k_T}\,\hat{R}_r\right)\psi_{ad}.
\end{aligned}$$

Let the control inputs u_1 and u_2 be designed as follows:

$$\begin{aligned}
u_1 &= \frac{L_o}{\beta_3}\left(-\frac{1}{L_o}\,\hat{R}_r\hat{\psi}_a - \frac{\beta_2}{L_o}\,\omega_r\hat{\psi}_b + \dot{I}_{ad}\right.\\
&\qquad\left. + \frac{M}{L_o}\,\hat{R}_r I_{ad} + \frac{\beta_1}{L_o}\,I_{ad} - k_1 e_1 + v_1\right)\\
u_2 &= \frac{L_o}{\beta_3}\left(-\frac{1}{L_o}\,\hat{R}_r\hat{\psi}_b + \frac{\beta_2}{L_o}\,\omega_r\hat{\psi}_a + \dot{I}_{bd} + \frac{M}{L_o}\,\hat{R}_r I_{bd}\right.\\
&\qquad\left. + \frac{\beta_1}{L_o}\,I_{bd} - k_1 e_2 + v_2\right)
\end{aligned}\tag{12}$$

for some constant $k_1 > 0$, where v_1 and v_2 are another auxiliary control inputs to be defined later, then it is clear that such control is indeed realizable.

Therefore, by substituting u_1 and u_2 into (11) and using the inequality of $XY \le (1/2)(X^2 + Y^2)$, we can attempt to derive the upper bound on $\dot{V}_1$ as follows:

$$\begin{aligned}
\dot{V}_1 \le &-\left(g_1\left(\frac{M}{L_o}\,R_r + \frac{\beta_1}{L_o} + k_1\right) - \frac{1}{2}\,g_2\,\frac{M^2}{L_r^2}\,\overline{R}_r^2\right)(e_1^2 + e_2^2)\\
&- g_2\left(\frac{R_r}{L_r} - 1\right)(e_3^2 + e_4^2) + g_1\,\frac{\beta_2}{L_o}\,\omega_r e_1\tilde{\psi}_b\\
&- g_1\,\frac{\beta_2}{L_o}\,\omega_r e_2\tilde{\psi}_a + g_1 e_1\left(\frac{R_r}{L_o}\,\tilde{\psi}_a + v_1\right)\\
&+ g_1 e_2\left(\frac{R_r}{L_o}\,\tilde{\psi}_b + v_2\right) + \frac{1}{2}\,g_2 a^2\tilde{R}_r^2\psi_{bd}^2 + \frac{1}{2}\,g_2 a^2\tilde{R}_r^2\psi_{ad}^2\\
&+ g_1 e_1\left(\frac{1}{L_o}\,\hat{\psi}_a - \frac{M}{L_o}\,I_{ad}\right)\tilde{R}_r\\
&+ g_1 e_2\left(\frac{1}{L_o}\,\hat{\psi}_b - \frac{M}{L_o}\,I_{bd}\right)\tilde{R}_r
\end{aligned}\tag{13}$$

where one has to beware that the terms of $bR_r\tilde{\psi}_a$ and $bR_r\tilde{\psi}_b$ are not measurable. Following a similar way of designing u_{o1} and u_{o2}, we let the auxiliary control inputs v_1 and v_2 be designed mainly to cancel those terms with either positive or uncertain sign.

Suppose v_1 and v_2 are designed as follows:

$$\begin{aligned}
v_1 &= \frac{1}{L_o L_r}\hat{R}_r(L_o\tilde{I}_a - \zeta_a)\\
v_2 &= \frac{1}{L_o L_r}\,\hat{R}_r(L_o\tilde{I}_b - \zeta_b)
\end{aligned}\tag{14}$$

which again is a realizable design. After somewhat ingenious manipulation, we have

$$\begin{aligned}
v_1 + \frac{1}{L_o}\,R_r\tilde{\psi}_a &= -\frac{1}{L_o L_r}\,\tilde{R}_r(L_o\tilde{I}_a - \zeta_a) + \frac{1}{L_o L_r}\,R_r(\eta_a - \epsilon_a)\\
v_2 + \frac{1}{L_o}\,R_r\tilde{\psi}_b &= -\frac{1}{L_o L_r}\,\tilde{R}_r(L_o\tilde{I}_b - \zeta_b) + \frac{1}{L_o L_r}\,R_r(\eta_b - \epsilon_b)
\end{aligned}$$

which along with the relation $\psi_{ad}^2 + \psi_{bd}^2 = \gamma^2$, derived from the proof of *Lemma 2*, can allow us to rederive the upper bound on $\dot{V}_1$ as

$$\begin{aligned}
\dot{V}_1 \le &-\left[g_1\left(\frac{M}{L_o}\,R_r + \frac{\beta_1}{L_o} + k_1\right) - \frac{1}{2}\,g_2\,\frac{M^2}{L_r^2}\,\overline{R}_r^2\right](e_1^2 + e_2^2)\\
&- g_2\left(\frac{R_r}{L_r} - 1\right)(e_3^2 + e_4^2) + g_1\,\frac{\beta_2}{L_o}\,\omega_r e_1\tilde{\psi}_b\\
&- g_1\,\frac{\beta_2}{L_o}\,\omega_r e_2\tilde{\psi}_a + \frac{g_1}{L_o L_r}\,e_1 R_r(\eta_a - \epsilon_a)\\
&+ \frac{g_1}{L_o L_r}\,e_2 R_r(\eta_b - \epsilon_b) + \tilde{R}_r\Omega_1 + \frac{1}{2}\,g_2\gamma^2 a^2\tilde{R}_r^2
\end{aligned}$$

where

$$\Omega_1 = g_1 e_1 \left(\frac{1}{L_o}\,\hat{\psi}_a - \frac{M}{L_o}\,I_{ad} - \frac{1}{L_o L_r}\,(L_o \tilde{I}_a - \zeta_a) \right) + g_1 e_2 \left(\frac{1}{L_o}\,\hat{\psi}_b - \frac{M}{L_o}\,I_{bd} - \frac{1}{L_o L_r}\,(L_o \tilde{I}_b - \zeta_b) \right).$$

So far, we have presented the design of control inputs, u_i and v_i, $i = 1, 2$, as well as the previous observer design, namely, u_{oi}, $i = 1, 2, 3, 4$. In order to complete the whole adaptive control design and to analyze the overall system stability, we choose a natural Lyapunov function candidate as follows:

$$V = V_0 + V_1.$$

To evaluate its time derivative, obviously we only need to sum up the time derivatives of V_1 and V_2 as follows:

$$\begin{aligned}
\dot{V} &= \dot{V}_0 + \dot{V}_1 \\
&\le -k_0 \tilde{I}_a^2 - k_0 \tilde{I}_b^2 - u_{c1}\tilde{I}_a - u_{c2}\tilde{I}_b - u_{c3}\tilde{Z}_a - u_{c4}\tilde{Z}_b \\
&\quad - \left(g_1 \left(\frac{M}{L_o} R_r + \frac{\beta_1}{L_o} + k_1 \right) - \frac{1}{2}\, g_2\, \frac{M^2}{L_r^2}\, \overline{R}_r^2 \right) (e_1^2 + e_2^2) \\
&\quad - g_2 \left(\frac{R_r}{L_r} - 1 \right) (e_3^2 + e_4^2) + g_1\, \frac{\beta_2}{L_o}\, \omega_r e_1 \tilde{\psi}_b - g_1\, \frac{\beta_2}{L_o} \\
&\quad \cdot \omega_r e_2 \tilde{\psi}_a + \tilde{R}_r \left(\Omega_0 + \Omega_1 + \frac{1}{2} g_2 \gamma^2 a^2 \tilde{R}_r - \Gamma_r^{-1} \dot{\hat{R}}_r \right) \\
&\quad + R_r(\eta_a - \epsilon_a) \left(\dot{\eta}_a + \frac{1}{L_r}\,\tilde{I}_a + \frac{g_1}{L_o L_r}\, e_1 \right) \\
&\quad + R_r(\eta_b - \epsilon_b) \left(\dot{\eta}_b + \frac{1}{L_r}\,\tilde{I}_b + \frac{g_1}{L_o L_r}\, e_2 \right)
\end{aligned}$$

which with the following design of auxiliary inputs:

$$\begin{aligned}
u_{c3} &= -g_1\, \frac{\beta_2}{L_r L_o}\, \omega_r e_2 \\
u_{c4} &= g_1\, \frac{\beta_2}{L_r L_o}\, \omega_r e_1 \\
u_{c1} &= -L_o u_{c3} \\
u_{c2} &= -L_o u_{c4} \\
\dot{\eta}_a &= -\frac{1}{L_r} \left(\tilde{I}_a + \frac{g_1}{L_o}\, e_1 \right) \\
\dot{\eta}_b &= -\frac{1}{L_r} \left(\tilde{I}_b + \frac{g_1}{L_o}\, e_2 \right) \\
\dot{\zeta}_a &= \frac{1}{L_r} \left(\tilde{I}_a + \frac{g_1}{L_o}\, e_1 + \beta_2 \omega_r \tilde{I}_b \right) - u_{c3} \\
\dot{\zeta}_b &= \frac{1}{L_r} \left(\tilde{I}_b + \frac{g_1}{L_o}\, e_2 - \beta_2 \omega_r \tilde{I}_a \right) - u_{c4}
\end{aligned} \tag{15}$$

apparently leads to a much more concise form for the estimate of $\dot{V}$

$$\begin{aligned}
\dot{V} &\le -k_0 \tilde{I}_a^2 - k_0 \tilde{I}_b^2 \\
&\quad - \left(g_1 \left(\frac{M}{L_o} R_r + \frac{\beta_1}{L_o} + k_1 \right) - \frac{1}{2}\, g_2\, \frac{M^2}{L_r^2}\, \overline{R}_r^2 \right) \\
&\quad \cdot (e_1^2 + e_2^2) - g_2 \left(\frac{R_r}{L_r} - 1 \right) (e_3^2 + e_4^2) \\
&\quad + \tilde{R}_r \left(\Omega_r + \frac{1}{2}\, g_2 \gamma^2 a^2 \tilde{R}_r - \Gamma_r^{-1} \dot{\hat{R}}_r \right)
\end{aligned} \tag{16}$$

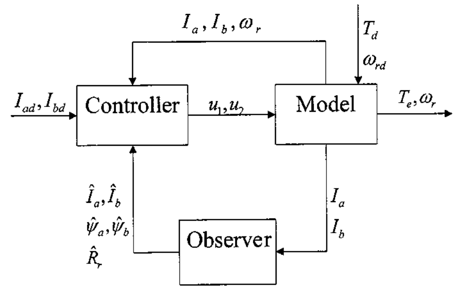

Fig. 1. Control block diagram.

where $\Omega_r = \Omega_0 + \Omega_1$, and for some constants k_0, $g_1 > 0$. By Lyapunov stability theory, all the signals involved in the system will be bounded provided the RHS of (16) can be established to be nonpositive. Clearly, it is not the case now simply because of the last term on the RHS of (16).

Remark 4: In general, the value of R_r is several Ω and L_r is hundreds of millihenrys and, therefore, we can directly make an assumption that $R_r/L_r - 1 \gg 0$

Since $\tilde{R}_r$ is not measurable so that the last term on the RHS of (16) cannot be easily canceled, we can design a projection law to eliminate it as follows to serve this purpose:

$$\dot{\hat{R}}_r = \begin{cases} \Gamma_r \left(\Omega_r + \frac{1}{2}\, g_2 \gamma^2 a^2 (\underline{R}_r - \hat{R}_r) \right), & \text{if } \hat{R}_r > \overline{R}_r + \delta_1 \\ \delta_2, & \text{if } \hat{R}_r = \overline{R}_r + \delta_1 \end{cases}$$

where the initial condition of $\hat{R}_r$ should satisfy

$$\hat{R}_r(0) > \overline{R}_r + \delta_1$$

for some constants δ_1, $\delta_2 > 0$, so that (16) can be further simplified as

$$\dot{V} \le -k_0 \tilde{I}_a^2 - k_0 \tilde{I}_b^2 - \left[g_1 \left(\frac{M}{L_o} R_r + \frac{\beta_1}{L_o} + k_1 \right) - \frac{1}{2}\, g_2\, \frac{M^2}{L_r^2}\, \overline{R}_r^2 \right] \cdot (e_1^2 + e_2^2) - g_2 \left(\frac{R_r}{L_r} - 1 \right) (e_3^2 + e_4^2). \tag{17}$$

By assigning g_1, g_2 properly, $V \ge 0$ and $\dot{V} \le 0$, it can be shown that $\tilde{I}_a$, $\tilde{I}_b$, $\tilde{Z}_a$, $\tilde{Z}_b$ and η_a, η_b, $\tilde{R}_r$, e_1, e_2, e_3, e_4 are all bounded. From *Corollary 2*, I_{ad}, I_{bd}, ψ_{ad}, ψ_{bd} are bounded and, hence, I_a, I_b, ψ_a, ψ_b are also bounded. By definitions of $\tilde{Z}_a$, $\tilde{Z}_b$, it is obvious that $\tilde{\psi}_a$ and $\tilde{\psi}_b$ are also bounded. Since the true states and the estimated errors are all bounded, then the state estimations $\hat{I}_a$, $\hat{I}_b$, $\hat{\psi}_a$, $\hat{\psi}_b$ are also bounded. Therefore, all the internal signals and, hence, the signal T_e as well as ω_r are bounded and, especially by assumption A5), $\dot{\tilde{I}}_a$, $\dot{\tilde{I}}_b$ from (9) and $\dot{e}_1$, $\dot{e}_2$, $\dot{e}_3$, $\dot{e}_4$ from (9) are all bounded as well.

As a consequence, all the input design is, indeed, a feasible design and by Barbalat's Lemma, it can be shown that

$$\tilde{I}_a, \tilde{I}_b, e_1, e_2, e_3, e_4 \to 0, \qquad \text{as } t \to \infty \tag{18}$$

and, according to the results of *Lemma 1*,

$$T_e \to T_d \qquad t \to \infty. \tag{19}$$

This shows that the objective of torque tracking is, thus, achieved. The following is a summarization of this result.

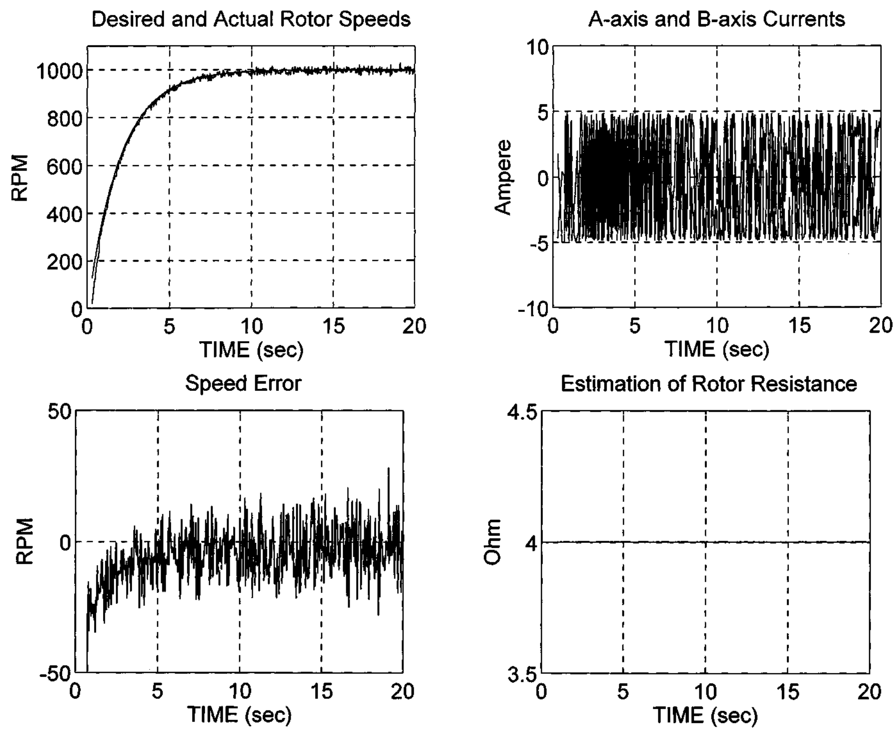

Fig. 2. $\omega_{rd} = 1000(1 - \exp^{-0.5t})$ r/min.

Theorem 1: Consider an induction motor whose dynamics are governed by (1) under the assumptions A1)–A5). Then, the output electrical torque T_e will approach the desired torque T_d asymptotically by the following controller:

$$\begin{aligned} u_1 &= \frac{L_o}{\beta_3}\left(-\frac{1}{L_o}\hat{R}_r\hat{\psi}_a - \frac{\beta_2}{L_o}\omega_r\hat{\psi}_b + \dot{I}_{ad} + \frac{M}{L_o}\hat{R}_r I_{ad}\right.\\ &\qquad \left. + \frac{\beta_1}{L_o} I_{ad} - k_1 e_1 + v_1\right)\\ u_2 &= \frac{L_o}{\beta_3}\left(-\frac{1}{L_o}\hat{R}_r\hat{\psi}_b + \frac{\beta_2}{L_o}\omega_r\hat{\psi}_a + \dot{I}_{bd} + \frac{M}{L_o}\hat{R}_r I_{bd}\right.\\ &\qquad \left. + \frac{\beta_1}{L_o} I_{bd} - k_1 e_2 + v_2\right)\end{aligned}$$

for some constant $k_1 > 0$, where subject to (4)

$$\begin{aligned} v_1 &= \frac{1}{L_o L_r}\hat{R}_r(L_o\tilde{I}_a - \zeta_a)\\ v_2 &= \frac{1}{L_o L_r}\hat{R}_r(L_o\tilde{I}_b - \zeta_b)\end{aligned}$$

and the auxiliary control inputs u_{ci}, u_{oi}, i =1, 2, 3, 4, and ζ_a, ζ_b, η_a, and eta_b are designed according to (6), (7), and (15) with the following parameter adaptation law:

$$\dot{\hat{R}}_r = \begin{cases} \Gamma_r\left(\Omega_r + \frac{1}{2}g_2\gamma^2 a^2(\underline{R}_r - \hat{R}_r)\right), & \text{if } \hat{R}_r > \overline{R}_r + \delta_1\\ \delta_2, & \text{if } \hat{R}_r = \overline{R}_r + \delta_1\end{cases}$$

subject to

$$\hat{R}_r(0) > \overline{R}_r + \delta_1$$

for some constants δ_1, δ_2, g_2, $\gamma > 0$.

2) Speed Control: For the rotating machines, such as induction motors, the rotor speed ω_r is the function of electrical torque and load torque, which can be expressed as follows:

$$J\dot{\omega}_r + D\omega_r = T_e - T_L \tag{20}$$

where $J > 0$ is the rotor inertia and $D > 0$ is the damping coefficient. Therefore, in order to achieve speed control of the induction motor, we should first take the effect of T_L into account. In this paper, the load torque T_L is as specified in *Remark 1*, and the electrical torque T_e can be re-expressed as

$$T_e = (J + \mu_2)\dot{\omega}_r + (D + \mu_1)\omega_r + \mu_0 \cdot s(k, \omega_r). \tag{21}$$

Define the desired torque T_d in the following form:

$$T_d = (J + \mu_2)\dot{\omega}_{rd} + (D + \mu_1)\omega_{rd} + \mu_0 \cdot s(k, \omega_{rd}) \tag{22}$$

where ω_{rd} is the desired speed trajectory. Subtracting (21) from (22) with T_L as specified in *Remark 1*, we can obtain

$$(J+\mu_2)\dot{e}_5+(D+\mu_1)e_5+\mu_0[s(k, \omega_r) - s(k, \omega_{rd})] = T_e - T_d \tag{23}$$

where $e_5 = \omega_r - \omega_{rd}$. If ω_{rd} is an arbitrary bounded second-order continuously differentiable function, then T_d from (22) obviously satisfies the assumption A4). In the following, we will present a theorem which describes conditions under which ω_r will approach ω_{rd} asymptotically.

A6) The desired rotor speed should be a bounded, second-order continuously differentiable function, i.e., $\omega_{rd} \in BC^2$, such that T_d satisfies A4).

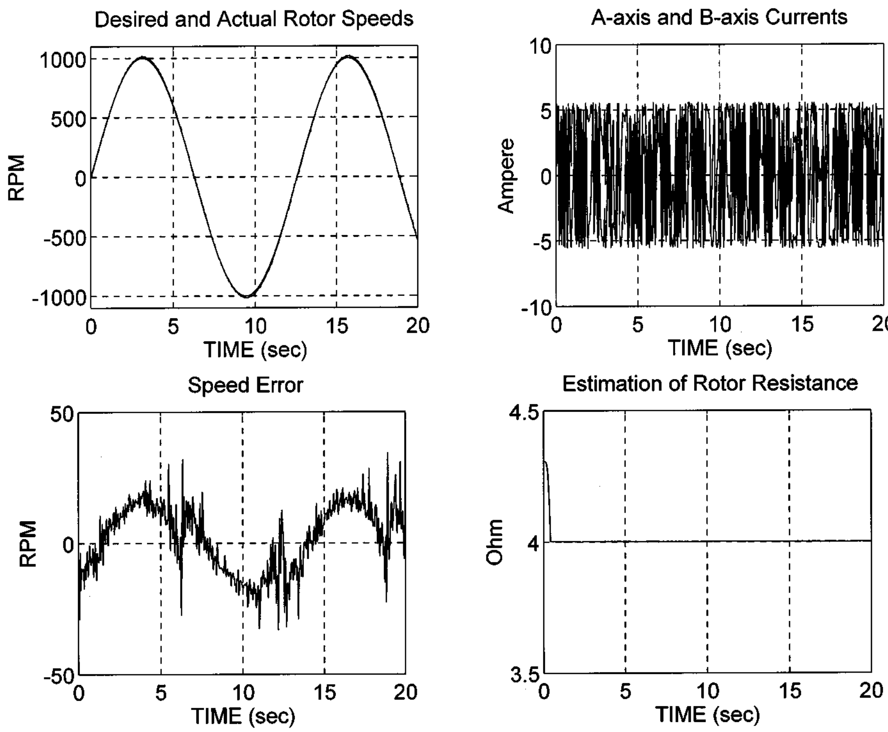

Fig. 3. $\omega_{rd} = 1000\ \sin(0.5t)$ r/min.

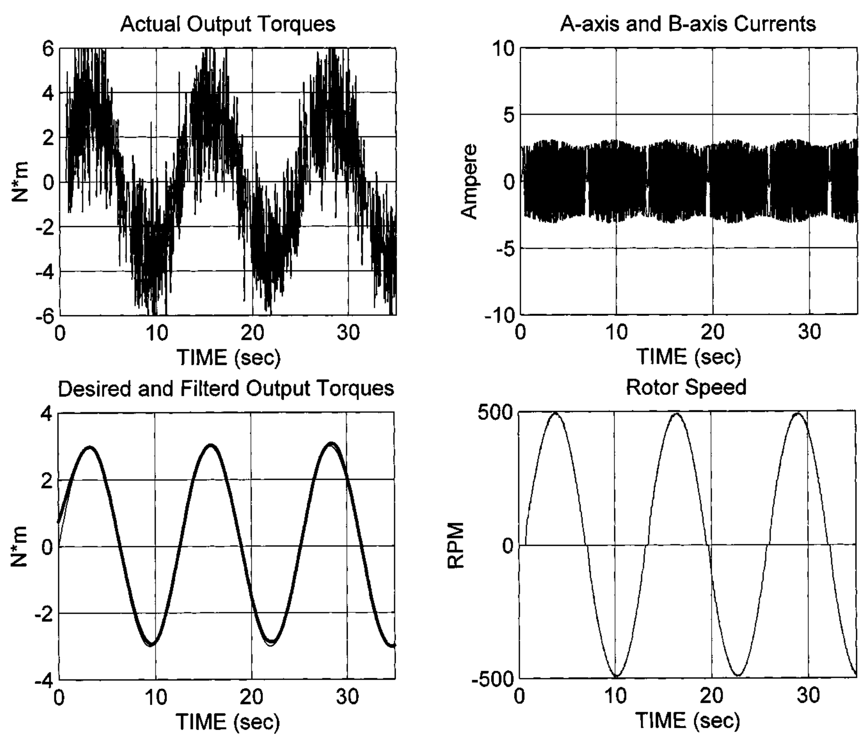

Fig. 4. $T_d = 3\ \sin(0.5t)$ N·m.

Before the theorem is presented, a useful working lemma will be introduced first, which helps to establish the fast speed convergence.

Lemma 2: Consider the following linear time-varying equation:

$$\dot{x}(t) + p(t) \cdot x(t) = q(t) \tag{24}$$

where $p(t) \geq p_0 > 0$, and $q(t)$ is a bounded function, which approaches 0 asymptotically. Then, $x(t)$ approaches 0 asymptotically.

Proof: Please refer to [15].

Theorem 2: Consider an induction motor whose dynamics are governed by (1) under the assumptions A1)–A3) and A5). If

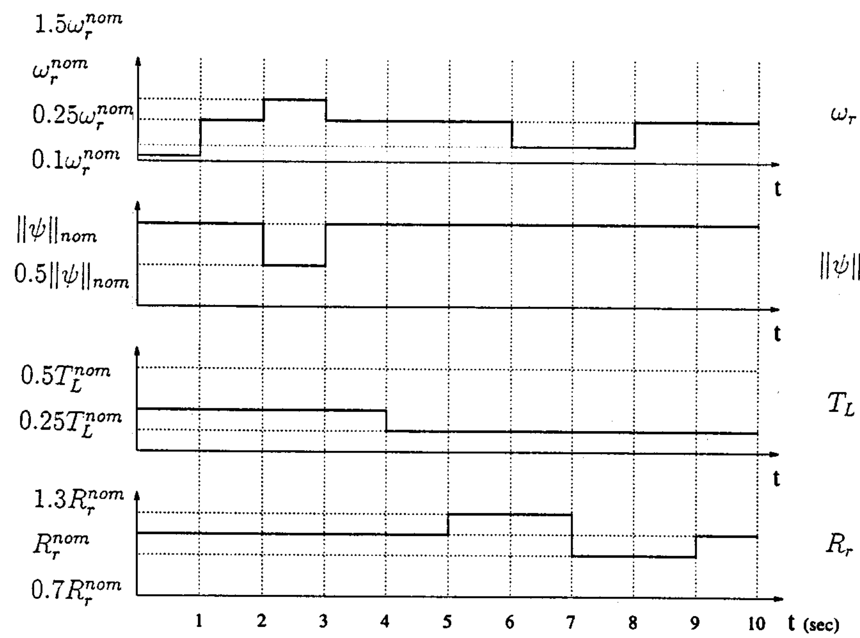

Fig. 5. Benchmark specification for simulation.

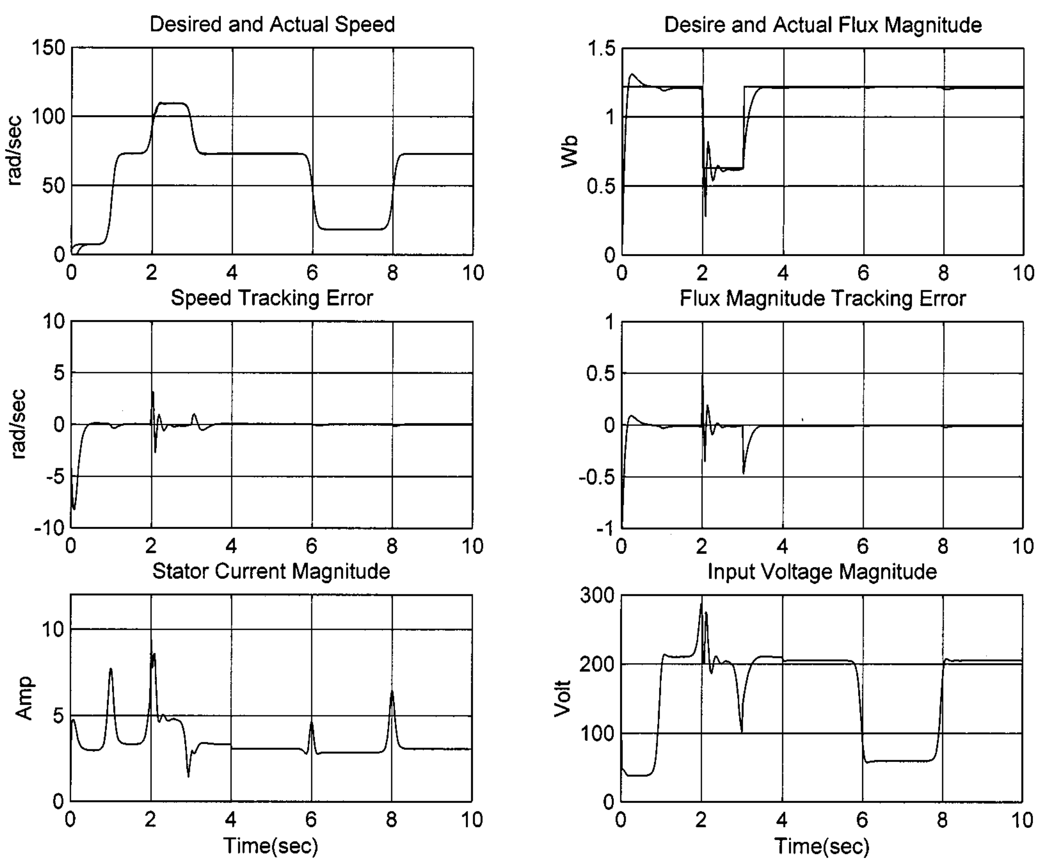

Fig. 6. Simulation of the benchmark test.

the desired rotor speed is a bounded, second-order continuously differentiable function and T_d is defined in the following form:

$$T_d = (J + \mu_2)\dot{\omega}_{rd} + (D + \mu_1)\omega_{rd} + \mu_0 \cdot s(k, \omega_{rd}).$$

Then, the rotor speed ω_r will approach ω_{rd} asymptotically by the controller the same as that of *Theorem 1*.

Proof: Please refer to [15].

Finally, the control block diagram is shown in Fig. 1.

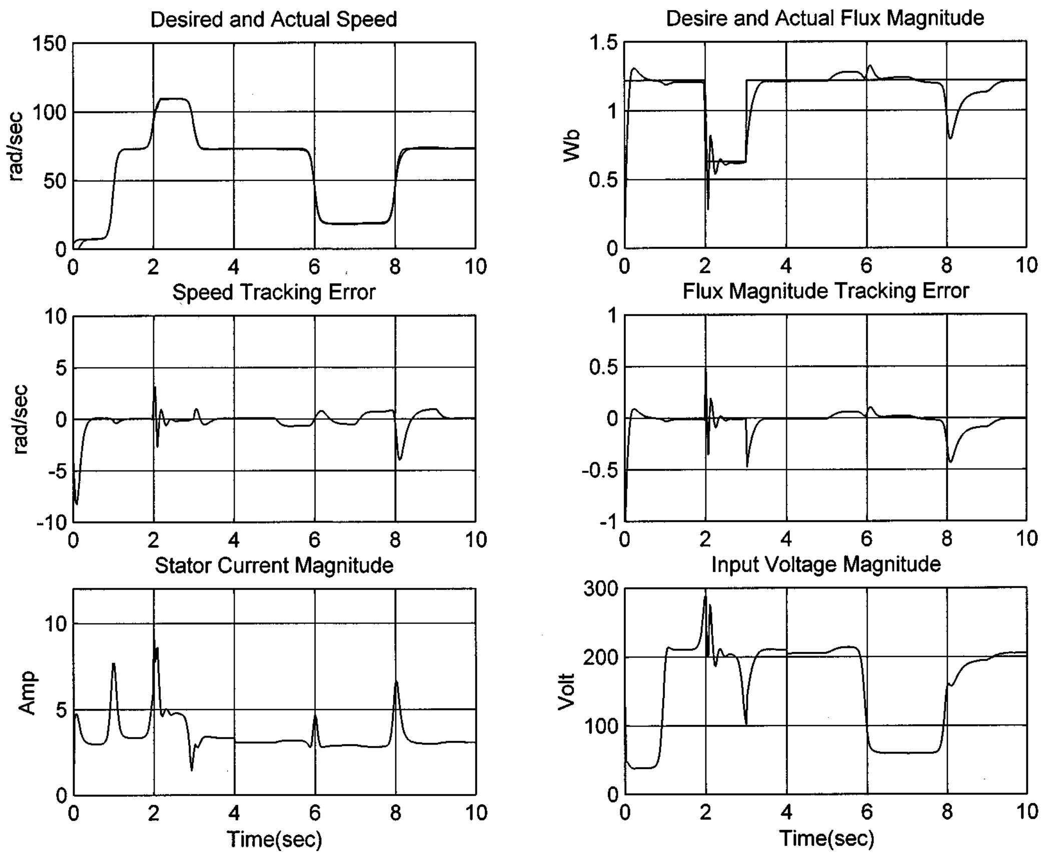

Fig. 7. Simulation of the benchmark test under the rotor resistance variation.

IV. Results and Discussions

A. Experiment Apparatus

The experiment consists of a Pentium PC, one power bipolar junction transistor (BJT) inverter with sinusoidal pulsewidth modulation (SPWM) which switches in 2 KHz, one 16-b D/A board, and a decoder board. A six-pole 1-hp three-phase induction motor with a squirrel-cage rotor is used and is equipped with a 1024-pulse/rev encoder. The motor is manufactured by ELMA Motors Company with delta-connected stator, and the specifications are as follows:

rated power—0.75 KW;
rated current—4 A;
rated voltage—220 V;
rated frequency—60 Hz;
rated speed—1120 r/min;
poles—six;
$R_s = 3.745\ \Omega$;
$R_r = 3.583\ \Omega$;
$L_s = 163.3$ mH;
$L_r = 163.3$ mH;
$M = 154.67$ mH;
$J = 0.05$ N·m^2.

To demonstrate the performance of our controller persuasively, we will make a variety of experiments on the induction motor. The test cases include speed control for desired (asymtotically) constant speed and desired sinusoidal speed, and torque control for desired constant torque. All the results are illustrated in Figs. 2–4. In Fig. 2, the desired rotor speed is specifically given as $\omega_{rd} = 1000(1 - \exp^{-0.5t})$ r/min and, in Fig. 3, the desired rotor speed is $\omega_{rd} = 1200 \sin(0.5t)$ r/min with the maximum tracking error ± 20 r/min at the positive and negative peak speeds. Both figures apparently demonstrate the satisfactory speed tracking performance. For torque control, the successful torque tracking is illustrated in Fig. 4.

B. Simulations with Benchmark Specification

We also perform computer simulations for the benchmark example. The rotor speed is required to change between these values: ω_{nom}, $0.1\omega_{\text{nom}}$, $0.25\omega_{\text{nom}}$, $1.5\omega_{\text{nom}}$ during $t = [0, 10]$ s, where the $\omega_{\text{nom}} = 700$ r/min. The rotor flux is required to change from $\|\psi\|_{\text{nom}}$ to $0.5\|\psi\|_{\text{nom}}$ when the rotor speed is $1.5\omega_{\text{nom}}$, where $\|\psi\|_{\text{nom}} = 1.22$ Wb. The load torque is $0.5{T_L}^{\text{nom}}$ at $t = 0$ s and changes to $0.25{T_L}^{\text{nom}}$ at $t = 4$ s where ${T_L}^{\text{nom}} = 7$ N·m, and the rotor resistance variation is $\pm 30\%$. The benchmark specification is shown in Fig. 5.

The benchmark assumptions are as follows.

- B1) Measurable signals are the stator currents (I_a, I_b) and rotor speed ω.
- B2) Load torque T_L is constant, although unknown.
- B3) All parameters are known, but the rotor resistance R_r will change.
- B4) Stator and rotor currents and stator voltages are constrained by 12 A and 300 V.
- B5) The system begins with all zero initial conditions.

The simulation results for the speed tracking are shown in Figs. 6 and 7. In order to comply with the theory developed in this paper, especially the assumptions on the smoothness of the desired speeds, we adopt a smooth function to approximate the step function so that the desired speed is continuously differentiable. One can find that the speed tracking is insensitive to the load torque variation. The speed tracking error will converge no matter what the desired values of the rotor flux norm and rotor resistance are. When the rotor resistance variation is considered, the error convergence for speed tracking will be slightly slower, which is shown in Fig. 7.

V. CONCLUSIONS

From the simulation and experiment results, we can make the following observations.

- The desired speed and torque trajectories can be assigned arbitrarily as long as they satisfy assumptions A6) and A4).
- In addition to the speed and torque control, we are able to control the flux magnitude $\sqrt{\psi_{ds}^2 + \psi_{qs}^2}$ as well.
- From the experiment results, we can see that the good tracking performance of the proposed servo controller will not be affected by the existence of the load torque.

From (8), the error between T_e and T_d is dependent on the desired currents and the desired fluxes, which is directly related to T_d and the desired flux magnitude γ by (10). Therefore, the convergence rate of the tracking may be different for different values of T_d and γ. It is noteworthy that the estimation of rotor resistance $\hat{R}_r$ will not approach the true value of the rotor resistance R_r.

In this paper, we have successfully proposed a nonlinear adaptive speed and torque servo controller, in which the rotor resistance is not required to be known exactly. We also give a rigorous proof of global stability without singularity and guarantee that the tracking error will converge to zero asymptotically.

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