Proceedings of the American Control Conference
San Diego, California • June 1999

Nonlinear Sensorless Indirect Adaptive Speed Control of Induction Motor with Unknown Rotor Resistance and Load

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Abstract

In this paper, a nonlinear indirect adaptive sensorless speed controller for induction motors is proposed. In the controller, only the stator currents are assumed to be measurable. Flux observers and rotor speed estimator are designed to relax the need of flux and speed measurement. Besides, two estimators are also designed to overcome drifting problem of the rotor resistance and unknown load torque.

Nomenclature

{ V_a, V_b } are stator voltages; { I_a, I_b } are stator currents; { ψ_a, ψ_b } are rotor fluxes; ω_r is the mechanical angular speed of the rotor; R_s is the stator resistance; R_r is the rotor resistance; L_s is the stator self-inductance; L_r is the rotor self-inductance; M is the mutual inductance; p is the number of pole-pairs; J is the rotor inertia; D is the damping coefficient; T_L is the load torque; k_T is the torque constant $(=\frac{3pM}{2L_r})$; L_o equals to $\frac{L_r^2}{M}(L_s - \frac{M^2}{L_r})$; β_1 equals to $\frac{R_s L_r^2}{M}$; β_2 equals to pL_r; β_3 equals to $\frac{L_r^2}{M}$

1 Introduction

The induction motor is a coupled system with highly nonlinear dynamics. In the early years, all system states are assumed to be measurable and parameters are assumed to be known. Under these assumptions, techniques such as classical field-orientation [1] and input-output linearization [2] are utilized to design the controller. Then flux observers are then designed to relax the need of flux measurement [3]. These flux observers are designed under the assumption that the rotor resistance is known. Following these researches, further efforts were then to design controllers and flux observers which are adaptive with respect to both system parameters and/or the load [7] [9].

All the schemes above require speed measurement. However, the speed measuring device is rather costly relative to the price of an induction motor in general. Besides, the measured signals are usually noisy and difficult to deal with. Therefore, controllers that does not require speed measurement are obviously preferable for practical implementation. Many research results on sensorless vector control have been proposed [6], of which analyze are mainly based on the steady state behavior and only rough proves are supplied. On the other hand, in [8], many researches on sensorless control were discussed and compared, including vector control and other modern control theory, such as robust and MRAS (Model Reference Adaptive System). Here, we follow this trend to design a full nonlinear adaptive sensorless speed controller for induction motors based on flux observer.

In this paper, an introduction of induction motors and related researches are discussed in section 1. In section 2, the mathematical model will be presented . The main part of this paper is section 3 in which observers and controller are designed and proved in detail. Experimental results are presented in section 4. Finally, we will make some conclusions in section 5.

2 Mathmetical Model

In this section, we introduce the induction motor model. If the induction motor never goes into the saturation region, and the air-gap MMF is sinusoidal, then it can be characterized by the following dynamic equations:

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

where L_o, β_1, β_2, β_3 are constants defined in the nomenclature. The mathematical model listed above is referred

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to the well-known stator fixed reference frame.Also, the dynamics of the mechanical part can be derived.

$$J\dot{\omega}_r + D\omega_r + T_L = T_e \quad (2)$$

where $J > 0$ is the rotor inertia, $D > 0$ is the damping coefficient and T_L is the load torque.

3 Induction Motor Control

Before the thorough investigation on the observers and controllers, several assumptions will be presented below to make the problem more precise.

Assumptions :

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

(A2) The stator currents are measurable.

(A3) The load torque T_L is an unknown constant.

(A4) The desired rotor speed should be a bounded smooth function with known first and second order time derivatives.

Control Objective :

Given the desired rotor speed trajectory $\omega_{rd}(t)$, our control goal is to design control laws such that the rotor speed ω_r and flux Ψ can track the desired trajectory asymptotically in finite time with all internal signals being bounded, subject to assumptions (A1)~(A4).

3.1 Observer Design and Analysis

The system mode is expressed as section 2. We set $R_r = R_{rn} + \theta_1$ and $\omega_r = \omega_{rd} + \theta_2$ where θ_1 stands for (unknown) difference between the actual resistanve value and its nominal value, and θ_2 denotes the speed tracking error. For easy reference, we define the notations for the observed values and the observation errors as $\tilde{I}_a = I_a - \hat{I}_a, \tilde{I}_b = I_b - \hat{I}_b, \tilde{\psi}_a = \psi_a - \hat{\psi}_a, \tilde{\psi}_b = \psi_b - \hat{\psi}_b, \tilde{R}_r = R_r - \hat{R}_r, \tilde{\omega}_r = \omega_r - \hat{\omega}_r$ where the symbol ^ denotes that it is an observed value and the symbol ~ denotes an observation error. According to the structure of the dynamics in (1), the observers are proposed as in [4] :

$$\begin{aligned} L_o\dot{\hat{I}}_a &= k_0\tilde{I}_a - (M\hat{R}_r + \beta_1)I_a + \hat{R}_r\hat{\psi}_a \\ &\quad +\beta_2\hat{\omega}_r\hat{\psi}_b + \beta_3 V_a + v_1 + v_5 \\ L_o\dot{\hat{I}}_b &= k_0\tilde{I}_b - (M\hat{R}_r + \beta_1)I_b - \beta_2\hat{\omega}_r\hat{\psi}_a \\ &\quad +\hat{R}_r\hat{\psi}_b + \beta_3 V_b + v_2 + v_6 \\ L_r\dot{\hat{\psi}}_a &= -\hat{R}_r\hat{\psi}_a + M\hat{R}_r I_a - \beta_2\hat{\omega}_r\hat{\psi}_b + v_3 \\ L_r\dot{\hat{\psi}}_b &= -\hat{R}_r\hat{\psi}_b + M\hat{R}_r I_b + \beta_2\hat{\omega}_r\hat{\psi}_a + v_4 \end{aligned} \quad (3)$$

where $\hat{R}_r = R_{rn} + \hat{\theta}_1, \hat{\omega}_r = \omega_{rd} + \hat{\theta}_2$ and the constant $k_0 > 0$ is a control gain. In order to utilize this property to cancel the unmeasurable terms, we design two auxiliary observation errors [4] :

$$\begin{aligned} \tilde{Z}_a &= L_o\tilde{I}_a + L_r\tilde{\psi}_a, \\ \tilde{Z}_b &= L_o\tilde{I}_b + L_r\tilde{\psi}_b \end{aligned} \quad (4)$$

Another interesting characteristic is studied here. Although the auxiliary observation errors, $\tilde{Z}_a$ and $\tilde{Z}_b$, are not measurable (because they are composed of $\tilde{I}$ and $\tilde{\psi}$, and $\tilde{\psi}$ is not measurable), their first-order time derivatives are measurable. Two additional error signals η_a and η_b are then defined as follows :

$$\begin{aligned} \eta_a &= \tilde{Z}_a - \zeta_a, \\ \eta_b &= \tilde{Z}_b - \zeta_b \end{aligned} \quad (5)$$

where ζ_a and ζ_b are auxiliary control signals. Again, η_a and η_b are unmeasurable errors but with measurable first-order time derivatives. Motivated by how the coupling terms can be canceled, we design the observer inputs v_1 , v_2 , v_3 and v_4 to be

$$\begin{aligned} v_1 &= -\frac{\hat{R}_r}{L_r}(L_o\tilde{I}_a - \zeta_a) - \frac{\beta_2\hat{\omega}_r}{L_r}(L_o\tilde{I}_b - \zeta_b) \\ v_2 &= -\frac{\hat{R}_r}{L_r}(L_o\tilde{I}_b - \zeta_b) + \frac{\beta_2\hat{\omega}_r}{L_r}(L_o\tilde{I}_a - \zeta_a) \\ v_3 &= -k_0\tilde{I}_a + \frac{L_o}{L_r}(\hat{R}_r\tilde{I}_a + \beta_2\hat{\omega}_r\tilde{I}_b) \\ v_4 &= -k_0\tilde{I}_b + \frac{L_o}{L_r}(\hat{R}_r\tilde{I}_b - \beta_2\hat{\omega}_r\tilde{I}_a) \end{aligned} \quad (6)$$

Now we are ready to perform the Lyapunov stability analysis to examine the stability condition of these observers. Define a Lyapunov function candidate as :

$$V_o = \frac{1}{2}\tilde{I}_a^2 + \frac{1}{2}\tilde{I}_b^2 + \frac{1}{2}k_\eta\eta_a^2 + \frac{1}{2}k_\eta\eta_b^2 + \frac{1}{2}k_\omega\tilde{\omega}_r^2 + \frac{1}{2}k_R\tilde{R}_r^2$$

where control gains k_η, k_ω and k_R should satisfy

$$k_\eta,\ k_\omega,\ k_R > 0$$

Here, to complete the final analysis an additional assumption is required.

(A5) The speed tracking error and the change of the rotor resistanceresistance from itd nominal value, i.e., θ_2 and θ_1, varies slowly so that its first order time derivative can be negligible.

Again, if $\dot{\hat{\eta}}_a$, $\dot{\hat{\eta}}_b$, $\dot{\hat{R}}_r$ and $\dot{\hat{\omega}}_r$ are designed properly which then leads to the design of those signal as

$$\begin{aligned} \dot{\hat{\eta}}_a &= -\frac{1}{k_\eta}\left(\frac{\hat{R}_r}{L_rL_o}\tilde{I}_a - \frac{\beta_2\hat{\omega}_r}{L_rL_o}\tilde{I}_b\right) \\ \dot{\hat{\eta}}_b &= -\frac{1}{k_\eta}\left(\frac{\hat{R}_r}{L_rL_o}\tilde{I}_b + \frac{\beta_2\hat{\omega}_r}{L_rL_o}\tilde{I}_a\right) \\ \dot{\hat{R}}_r &= -\frac{1}{k_R}\left(\frac{\hat{\Omega}_{11}}{L_o}\tilde{I}_a + \frac{\hat{\Omega}_{12}}{L_o}\tilde{I}_b\right) \\ \dot{\hat{\omega}}_r &= -\frac{1}{k_\omega}\left(\frac{\hat{\Omega}_{21}}{L_o}\tilde{I}_a + \frac{\hat{\Omega}_{22}}{L_o}\tilde{I}_b\right) \end{aligned} \quad (7)$$

where $\hat{\eta}_a$ and $\hat{\eta}_b$ are estimated values of η_a and η_b, respectively, satisfying $\dot{\hat{\eta}}_a = \dot{\eta}_a, \hat{\eta}_a(0) = 0; \dot{\hat{\eta}}_b = \dot{\eta}_b, \hat{\eta}_b(0) = 0$. By substituting the rotor resistance estimator and the rotor speed observer, we can re-assess $\dot{V}$ as :

$$\dot{V}_o \le -k_0\tilde{I}_a^2 - \tilde{I}_a v_5 + \frac{1}{L_rL_o}|\tilde{I}_a||\tilde{R}_r\tilde{\eta}_a|$$

$$
\begin{aligned}
&+\frac{\beta_2}{L_r L_o}\,|\tilde I_a||\tilde\omega_r\tilde\eta_b| \\
&-k_0\tilde I_b^2-\tilde I_b v_6+\frac{1}{L_r L_o}\,|\tilde I_b||\tilde R_r\tilde\eta_b| \\
&+\frac{\beta_2}{L_r L_o}\,|\tilde I_b||\tilde\omega_r\tilde\eta_a|
\end{aligned}
$$

Clearly, the upper bound on $\dot V$ on the right hand side (RHS) of equation (8) still does not have definite negative sign. To cope with that, we incorporate the variable structure design (VSD) into our controller. We evaluate the four functions $F_i(t)$, $i=1,2,3,4$ which satisfy

$$
\begin{aligned}
F_1(t) &\geq \delta_a\,|\tilde R_r|, \quad F_2(t)\geq \delta_b\,|\tilde\omega_r|, \\
F_3(t) &\geq \delta_b\,|\tilde R_r|, \quad F_4(t)\geq \delta_a\,|\tilde\omega_r|
\end{aligned}
$$

Accordingly, we design the four control signals v_5 and v_6 as :

$$
\begin{aligned}
v_5 &= \frac{1}{L_r}\mathrm{sgn}(\tilde I_a)\cdot\Big(F_1(t)+\beta_2F_2(t)\Big) \\
v_6 &= \frac{1}{L_r}\mathrm{sgn}(\tilde I_b)\cdot\Big(F_3(t)+\beta_2F_4(t)\Big)\,, \qquad (8)
\end{aligned}
$$

where $\mathrm{sgn}(\cdot)$ is the sign function defined as

$$
\mathrm{sgn}(x)=\begin{cases}1 & \text{if } x\geq 0\\ 0 & \text{if } x<0\end{cases}\qquad \forall\, x\in\Re\,,
$$

As a result, $\dot V_o$ becomes

$$
\dot V_o \leq -\frac{k_o}{L_o}(\tilde I_a^2+\tilde I_b^2)
$$

It follows that $\tilde I_a$, $\tilde I_b$, $\tilde R_r$, $\tilde\omega_r$, η_a and η_b are all bounded. We now impose that $\hat\theta_1(t)+R_{rn}\geq\frac{R_{rn}}{2}>0$, $\forall t\geq 0$. Denoting by $\hat\theta_{1p}(t)$ the modification of $\hat\theta_1(t)$ given by the projection algorithm, $\hat\theta_{1p}(t)$ is chosen such that

$$
(\theta_1-\hat\theta_1(t))^2\geq(\theta_1-\hat\theta_{1p}(t))^2
$$

which guarantees that the value of V_o does not increase when $\hat\theta_1$ is replaced by $\hat\theta_{1p}$ given by the projection algorithm. Since $R_r=R_{rn}+\theta_1$ is positive, the estimate $\hat\theta_1(t)$ is given as

$$
\hat\theta_1(t)=\hat\theta_{1p}(t)=\frac{-R_{rn}}{2}+\epsilon, if\ \hat\theta_1(t)\leq -R_{rn}
$$

with $\epsilon>0$ a positive constant such that $\epsilon\leq 2(R_{rn}+\theta_1)=2R_r$. Since (I_a,I_b) are bounded in finite time ,i.e.(I_a, $I_b\in L_{\infty e}$), by assumption, (ψ_a,ψ_b,ω_r) are also bounded. Since $(\psi_a,\psi_b,I_a,I_b,\omega_r)$ are bounded, if $\hat R_r=R_{rn}+\hat\theta_1\geq\frac{R_{rn}}{2}>0$, it follows that $\hat\psi_a$ and $\hat\psi_b$ are also bounded according to (3). So, $(\hat\psi_a,\hat\psi_b)$ are bounded, and therefore $\tilde Z_a$ and $\tilde Z_b$ are bounded. This implies that ζ_a and ζ_b are bounded and that v_1, v_2, v_3 and v_4 are bounded as well. So, $\dot{\tilde I}_a$ and $\dot{\tilde I}_b$ are bounded. On the other hand, $\dot{\tilde I}_a$ and $\dot{\tilde I}_b$ are L_2 signals. By Barbalat's lemma, it follows that $\lim_{t\to\infty}\tilde I_a(t)=0, \lim_{t\to\infty}\tilde I_b(t)=0$. We can rewrite error system in matrix form as

$$
\begin{aligned}
\dot{\tilde I} &= A\tilde I+W^T(t)X+B(t), &(9)\\
\dot X &= -\Lambda W(t)\tilde I+\Lambda C\tilde I &(10)
\end{aligned}
$$

in which $\tilde I=[\tilde I_a,\tilde I_b]^T$, $X=[\tilde R_r,\tilde\omega_r,\eta_a,\eta_b]$, $\Lambda=diag[\frac{1}{k_\eta},\frac{1}{k_\eta},\frac{1}{k_R},\frac{1}{k_\omega}]$ and

$$
\begin{aligned}
A &= \begin{bmatrix} -\frac{k_o}{L_o} & 0\\ 0 & -\frac{k_o}{L_o}\end{bmatrix}, \\
W^T(t) &= \begin{bmatrix} \frac{\Omega_{11}}{L_o} & \frac{\Omega_{21}}{L_o} & \frac{\hat R_r}{L_rL_o} & \frac{\beta_2\hat\omega_r}{L_rL_o}\\ \frac{\Omega_{12}}{L_o} & \frac{\Omega_{22}}{L_o} & -\frac{\beta_2\hat\omega_r}{L_rL_o} & \frac{\hat R_r}{L_rL_o}\end{bmatrix}, \\
B(t) &= \begin{bmatrix} v_5\\ v_6\end{bmatrix}, \\
C^T &= \begin{bmatrix} \frac{\tilde\eta_a}{L_rL_o} & \frac{\tilde\eta_b}{L_rL_o} & 0 & 0\\ \frac{\tilde\eta_b}{L_rL_o} & \frac{\tilde\eta_a}{L_rL_o} & 0 & 0\end{bmatrix} \qquad (11)
\end{aligned}
$$

where

$$
\begin{aligned}
\Omega_{11} &= \hat\psi_a-MI_a-\frac{L_o}{L_r}\tilde I_a+\frac{1}{L_r}\zeta_a+\frac{1}{L_r}\eta_a\\
\Omega_{12} &= \hat\psi_b-MI_b-\frac{L_o}{L_r}\tilde I_b+\frac{1}{L_r}\zeta_b+\frac{1}{L_r}\eta_b\\
\Omega_{21} &= \beta_2\hat\psi_b-\frac{\beta_2L_o}{L_r}\tilde I_b+\frac{\beta_2}{L_r}\zeta_b+\frac{\beta_2}{L_r}\eta_b\\
\Omega_{22} &= -\beta_2\hat\psi_a+\frac{\beta_2L_o}{L_r}\tilde I_a-\frac{\beta_2}{L_r}\zeta_a-\frac{\beta_2}{L_r}\eta_a
\end{aligned}
\qquad (12)
$$

Lemma 1.

If, in addition, there exist two positive constants T and ϵ such that

$$
\int_t^{t+T} W(\tau)W^T(\tau)\geq\epsilon I>0, \forall t\geq 0 \qquad (13)
$$

with $W(t)$ defined as in (11), then according to Appendix A applied to

$$
\begin{aligned}
\dot{\tilde I} &= A(t)\tilde I+W^T(t)X,\\
\dot X &= -W(t)\tilde I \qquad (14)
\end{aligned}
$$

□

We can get $\tilde I$ asymptoically tend to zero and X is bounded by matrix B according to Lemma 1. While $\tilde I$ is small, we can give small control signals (v_5,v_6). Finally, if $\tilde I$ tend to zero, we give v_5, v_6 zero. Then, the equilibrium point ($\tilde I=0$, $X=0$) is asymptotically stable, namely $\tilde R_r$, $\tilde\omega_r$, η_a, η_b, $\tilde I_a$ and $\tilde I_b$ tend asymptotically to zero, which implies that I_a and I_b are bounded and $\hat\psi_a$ and $\hat\psi_b$ are also bounded, that $\tilde\psi_a$ and $\tilde\psi_b$ also tend asymptotically to zero. in conclusion, equation (3) with (v_1, v_2, v_3, v_4) are given in (6), and $\hat\theta_1$, $\hat\theta_2$, η_a and η_b updated according to (7) constitute an adaptive observer provided that (13) is satisfied.

3.2 Controller Design and Analysis

In this section, we propose a state feedback control for induction motor system which is adaptive with respect to the unknown constant load torque T_L, assuming all states (I_a,I_b,ψ_a,ψ_b) are measurable and the rotor resistance (R_r) is a known constant. The controller objective is to guarantee asymptotical zero convergence of rotor speed tracking error, rotor flux tracking error, and load torque estimation error under any intial condition. The dynamics model of the induction motor is expressed as (1). And, we define the speed tacking error, flux tacking error and load torque estimation error as follows : $e_\omega = \omega_r - \omega_{rd}, e_\Psi = \psi_a^2 + \psi_b^2 - \Psi_d^2, e_T = T_L - \hat{T}_L$, where ω_{rd} and Ψ_d are the reference signals and $\hat{T}_L$ is a time-varying estimated of T_L.

With the tracking errors and load torque estimation error defined above, their dynamics can be derived from (1) as :

$$\begin{aligned} J\dot{e}_\omega &= k_T(\psi_a I_b - \psi_b I_a) - D\omega_r - T_L - J\dot{\omega}_{rd}, \\ L_r\dot{e}_\Psi &= -2R_r\Psi^2 - 2L_r\dot{\Psi}_d^2 \\ &\quad +2MR_r(\psi_a I_a + \psi_b I_b) \end{aligned} \tag{15}$$

By designing the input signal I_a and I_b, we first consider a Lyapunov function candidate defined as :

$$V_1 = \frac{1}{2}(r_1 J e_\omega^2 + r_2 L_r e_\Psi^2 + r_3 e_T^2)$$

where gains r_1, r_2 and r_3 are positive. The choice of the desired currents I_{ad} and I_{bd} are equivalent to the following :

$$\begin{aligned} I_{ad} &= I_a = \frac{1}{\Psi^2}(\Omega_1\psi_a - \Omega_2\psi_b) \\ I_{bd} &= I_b = \frac{1}{\Psi^2}(\Omega_2\psi_a + \Omega_1\psi_b) \end{aligned} \tag{16}$$

where

$$\begin{aligned} \Omega_1 &= \frac{2R_r\Psi^2 + 2L_r\dot{\Psi}_d^2 - K_\Psi e_\Psi}{2MR_r} \\ \Omega_2 &= \frac{D\omega_{rd} + J\dot{\omega}_{rd} - (K_\omega - D)e_\omega + \hat{T}_L}{k_T} \end{aligned} \tag{17}$$

Define the current errors, namely, the differences between the actual currents and their desired values, as$e_a = I_a - I_{ad}, e_b = I_b - I_{bd}$. Then, the error dynamics involving $(e_\omega, e_\Psi, e_a, e_b)$ can be summerized as follows :

$$\begin{aligned} J\dot{e}_\omega &= k_T(\psi_a e_b - \psi_b e_a) - K_\omega e_\omega - e_T, \\ L_r\dot{e}_\Psi &= 2MR_r(\psi_a e_a + \psi_b e_b) - K_\Psi e_\Psi \\ L_o\dot{e}_a &= -MR_r I_a - \beta_1 I_a + R_r\psi_a \\ &\quad +\beta_2\omega_r\psi_b + \beta_3 V_a - L_o\dot{I}_{ad} \\ L_o\dot{e}_b &= -MR_r I_b - \beta_1 I_b - \beta_2\omega_r\psi_a \\ &\quad +R_r\psi_b + \beta_3 V_b - L_o\dot{I}_{bd} \end{aligned} \tag{18}$$

The derivatives of the reference currents are found as :

$$\begin{aligned} \dot{I}_{ad} &= \Omega_3 - \frac{L_o\psi_b(K_\omega - D)}{\Psi^2 k_T J}e_T \\ \dot{I}_{bd} &= \Omega_4 + \frac{L_o\psi_a(K_\omega - D)}{\Psi^2 k_T J}e_T \end{aligned} \tag{19}$$

where Ω_3 and Ω_4 are known function. Finally, we chose the control input terms V_a and V_b as

$$\begin{aligned} V_a &= \frac{1}{\beta_3}(MR_r I_a + \beta_1 I_a - R_r\psi_a \\ &\quad -\beta_2\omega_r\psi_b + \Omega_3 - K_a e_a + u_a) \\ V_b &= \frac{1}{\beta_3}(MR_r I_b + \beta_1 I_b - R_r\psi_b \\ &\quad +\beta_2\omega_r\psi_a + \Omega_4 - K_b e_b + u_b) \end{aligned} \tag{20}$$

where $K_a, K_b \geq 0$, and u_a and u_b are to be designed later. Given such design arrangement the error dynamics of I_a and I_b become

$$\begin{aligned} L_o\dot{e}_a &= -K_a e_a + u_a + \frac{L_o\psi_b(K_\omega - D)}{\Psi^2 k_T J}e_T \\ L_o\dot{e}_b &= -K_b e_b + u_b - \frac{L_o\psi_a(K_\omega - D)}{\Psi^2 k_T J}e_T \end{aligned} \tag{21}$$

The analysis is also based on the Lyapunov stability theory. First, we define a Lyapunov function candidate for the controller as :

$$V = \frac{1}{2}(r_1 J e_\omega^2 + r_2 L_r e_\Psi^2 + r_3 e_T^2 + r_4 L_o(e_a^2 + e_b^2) + \epsilon e_T e_\omega)$$

for some positive constants r_1, r_2, r_3 and r_4 and for some sufficiently small $\epsilon(> 0)$. After a careful choice, we following payload estimation algorithn as follow:

$$\begin{aligned} \dot{\hat{T}}_L &= \frac{1}{r_3}(-r_1 e_\omega + r_4 e_a\frac{L_o\psi_b(K_\omega - D)}{\Psi^2 k_T J} \\ &\quad -r_4 e_b\frac{L_o\psi_a(K_\omega - D)}{\Psi^2 k_T J}) \\ &\quad +\frac{\epsilon}{J}(k_T(\psi_a e_b - \psi_b e_a) - K_\omega e_\omega)) \end{aligned} \tag{22}$$

If u_a and u_b are chosen as

$$\begin{aligned} u_a &= \frac{1}{r_4}(r_1 e_\omega k_T\psi_b - 2r_2 e_\Psi MR_r\psi_a \\ &\quad +r_4\frac{\epsilon e_\omega}{r_3}\frac{L_o\psi_b(K_\omega - D)}{\Psi^2 k_T J}) - \frac{\epsilon e_\omega}{r_3 J}\epsilon k_T\psi_b \\ u_b &= \frac{1}{r_4}(-r_1 e_\omega k_T\psi_a - 2r_2 e_\Psi MR_r\psi_b \\ &\quad -r_4\frac{\epsilon e_\omega}{r_3}\frac{L_o\psi_a(K_\omega - D)}{\Psi^2 k_T J}) + \frac{\epsilon e_\omega}{r_3 J}\epsilon k_T\psi_a \end{aligned} \tag{23}$$

There exist $K_v > 0$ then

$$\dot{V} \leq -K_v \parallel [e_\omega, e_\Psi, e_a, e_b, e_T]^T \parallel^2$$

Consequently, we know the signals (e_ω, e_Ψ, e_a, e_b, e_T) in the closed-loop system is bounded from Lyapunov stability theory. Since (I_a, I_b, ψ_a, ψ_b, ω_r) are bounded and the initial flux value Ψ is not equal to zero, it follows that the time derivative of the tracking errors , namely, ($\dot{e}_\omega$, $\dot{e}_\Psi$, $\dot{e}_a$, $\dot{e}_b$), the control inputs (V_a, V_b, u_a, u_b) and the estimated signal ($\dot{\hat{T}}_L$) are also bounded. On the other hand, e_ω, e_Ψ, e_a, e_b and e_T are L_2 signals. So, they are asymptotically stable by Barbalat's lemma.

4 Experimental Result

Experiments are done with a three horse power induction motor which is manufactured by TECO Co. Ltd. Taiwan. In order to check the performance, we have done two experiments with exponential and sinusoidal speed command as in figure 1 and figure 2. Obviously, experiments show that the sensorless controller is indeed effective to drive the motor to track a given smooth speed command.

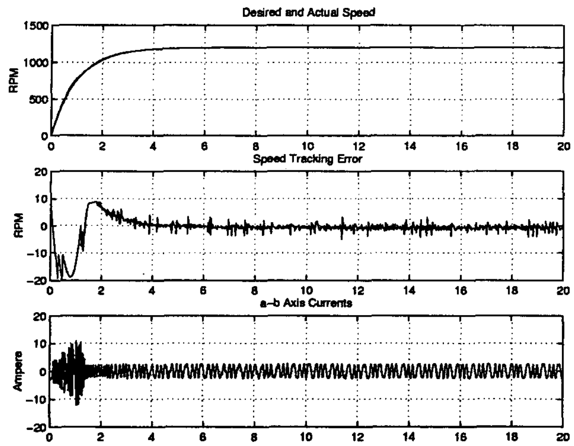

Figure 1: $\omega_{rd} = 1200(1 - e^{-t})$ RPM with no load

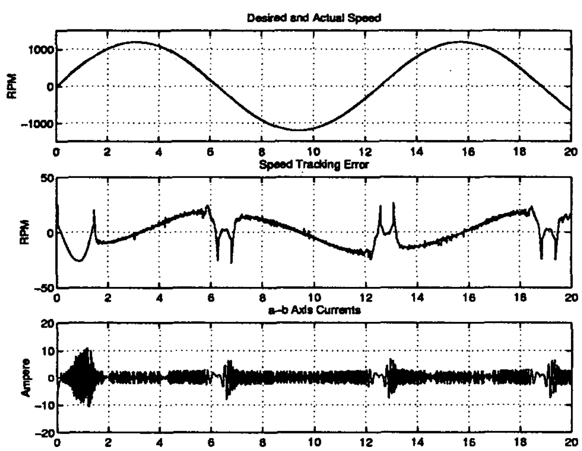

Figure 2: $\omega_{rd} = 1200 \cdot \sin(0.5t)$ RPM with no load

5 Conclusion

In this paper, we have presented a partial-state feedback adaptive sensorless speed and flux tracking controller for induction motors with fifth-order nonlinear dynamic model which is actuacted by a voltage source. The main contribution of the controller is that asymptotic tracking of rotor speed and rotor fluxes are achieved without measurement of both the rotor fluxes and the rotor speed. Moreover, the variations of the rotor resistance and load torque are also taken into account.

References

[1] W. Leonhard, "Microcomputer Control of High Dynamic Performance Ac-Drives-a Survey", *Automatica*, Vol. 22, pp. 1-19, 1986.

[2] R. Marino, S. Peresada and P. Valigi, "Adaptive Input-Output Linearizing Control of Induction Motors", *IEEE Trans. on Automatic Control*, Vol. 38, pp. 208-221, 1993.

[3] G. C. Verghese and S. R. Sanders, "Observers for Flux Estimation in Induction Machines", *IEEE Trans. on Industrial Electronics*, Vol. 35, pp. 85-94, 1988.

[4] J. Hu and D. M. Dawson, "Adaptive Control of Induction Motor Systems Despite Rotor Resistance Uncertainty", *Proceedings of the American Control Conference*, Jun. 1996, pp. 1397-1402

[5] A. M. Lee and L. C. Fu, "Nonlinear Adaptive Speed and Torque Control of Induction Motors with Unknown Rotor Resistance", Master Thesis National Taiwan University Taiwan R.O.C., 1996.

[6] H. Kubota and K. Matsuse, "Speed Sensorless Field-Oriented Control of Induction Motor with Rotor Resistance Adaptation", *IEEE Trans. on Industrial Application*, Vol. 30, No. 5, Sep./Oct. 1994, pp. 1219-1224.

[7] Jung-Hua Yang, Wen-Hai Yu, and Li-Chen Fu, "Nonlinear Observer-Based Adaptive Tracking Control for Induction Motors with Unknown Load", *IEEE Trans on Industrial Electronics*, Vol. 42, No. 6, Dec. 1995, pp. 579-586.

[8] C. Has, A. Bettini, L. Feraris, G. Griva and F. Profumo, "Comparison of Different Schemes without Shaft Sensors for Field Oriented Control Drives", *Proceedings of IEEE IECON'94*, pp. 1579-1588, 1994.

[9] R. Marino, S. Peresada and P. Tomei, "Adaptive Observer-Based Control of Induction Motors with Unknown Rotor Resistance", *IEEE International Journal of Adaptive Control and Signal Processing,* Vol. 10, 1996, pp.345-363.