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ROBUST ADAPTIVE CONTROL OF INDUCTION MOTORS WITH SPIRAL VECTOR

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Abstract: We have proposed a new parameter adjustment law that guarantees the stability of the error system that has a positive definite solution for a Riccati equation instead of the strictly positive realness condition, even if the plant has exogenous disturbances. However, the parameter adjustment law has a disadvantage that the transient response of the closed-loop system has large overshoot. In this paper, we try to improve the transient performance by introducing the imaginary axis shifting. Moreover, we apply an adaptive scheme to the position control of current-fed induction motors by the field-oriented control method. *Copyright © 2002 IFAC*

Keywords: Adaptive algorithm, Riccati equation, coprime factorization, field-oriented control, current-fed induction motor

1. INTRODUCTION

Strictly positive realness plays an important role in the model reference adaptive control theory(Narendra, 1989). If the error system in the model reference adaptive control system is strictly positive real, the gradient-type parameter adjustment law acts to stabilize the error system. However, the adaptive control synthesis is rendered difficult when the relative degree of the plant exceeds unity, because the error system is not strictly positive real. Recently, we proposed a new parameter adjustment law that guarantees the asymptotic stability of the error system that has a positive definite solution for a Riccati equation instead of the strictly positive realness condition(Matsuo, 1995). However, the parameter adjustment law has a disadvantage that the transient response of the closed-loop system has large overshoot.

In this paper, we try to improve the transient performance of closed-loop system. Yamamura proposed the spiral vector method to derive analytical solutions of electromagnetic transients of AC machines by introducing a new idea of damped inductance(Yamamura, 1992). We compare the spiral vector with the imaginary axis shifting in the control theory, and present an adaptive control system with a spiral vector to improve the transient response in the position control system of current-fed induction motors. An adaptive controller is designed based on the motor dynamics approximated by a second-order dynamical system since the current control loop time constant is small enough to be negligible in current-fed induction motor with the field-oriented control.

Our approach has the following features:

(1) The solution for the Riccati equation exists when the relative degree of the plant exceeds

unity. Thus, we can deal with plants with a relative degree >= 2 in much the way same as with a relative degree = 1.

(2) By increasing the exponential part of the spiral vector, we can reduce overshoots of the adaptive control systems.

(3) We can reduce the stability of the error system with bounded disturbances to the existence of a positive definite solution for a Riccati equation, and derive an L_2 performance condition.

2. RELATIONSHIP BETWEEN SPIRAL VECTOR AND IMAGINARY AXIS SHIFTING

The spiral vector method applied by Yamamura to analysis of AC motors provides a new analytical solution of both steady and transient states(Yamamura, 1992). The spiral vector is defined as the following exponential function:

$$i = Ae^{\delta t}, \quad \delta = -\lambda + j\omega \tag{1}$$

where A means the phaser.

We consider the following differential equation:

$$A(p)y(t) = B(p)u(t), \quad p = \frac{d}{dt} \tag{2}$$

where $y(t)$ is an output, $u(t)$ is an input,

$$\begin{aligned} A(p) &= a_0 + a_1 p + \cdots + a_n p^n \\ B(p) &= b_0 + b_1 p + \cdots + b_m p^m. \end{aligned}$$

If the input u is the following spiral vector:

$$u(t) = U_p e^{\delta t}, \quad U_p = |V|e^{j\phi},$$

the solution of (2) is given by

$$y(t) = \frac{B(\delta)}{A(\delta)} U_p e^{\delta t} + \sum_{i=1}^{n} A_i e^{\delta_i t} \tag{3}$$

where $\delta_1, \cdots, \delta_n$ are all the solutions of characteristic equation $A(p) = 0$, under the assumption that there are no multiple roots. $\frac{B(\delta)}{A(\delta)}$ is an extension of the frequency response.

If the input and output of (2) are written in the following spiral vector form:

$$u(t) = e^{-\delta t}\bar{u}(t), \quad y(t) = e^{-\delta t}\bar{y}(t). \tag{4}$$

The relationship between $\bar{y}$ and $\bar{u}$ is obtained as

$$A(p-\delta)\bar{y}(t) = B(p-\delta)\bar{u}(t). \tag{5}$$

The spiral vector can be considered as a kind of imaginary axis shifting.

3. CONTROLLER DESIGN WITH IMAGINARY AXIS SHIFTING

We consider the plant obtained by shifting the imaginary axis as a plant model to design a controller with a good tracking performance. Setting the original plant as (2), we have such a δ-shifted plant as

$$\begin{aligned} \bar{y}(t) &= \frac{B(p-\delta)}{A(p-\delta)}\bar{u}(t) = e^{\delta t}y(t) \\ &= e^{\delta t}\frac{B(p)}{A(p)}u(t) = e^{\delta t}\frac{B(p)}{A(p)}e^{-\delta t}\bar{u}(t). \end{aligned}$$

If the LTI controller,

$$\bar{u}(t) = C(p)\bar{y}(t) \tag{6}$$

stabilizes the δ-shifted plant, $\bar{y}(t)$ converges to 0 as $t \to \infty$. Thus, $y(t)$ converges to 0 faster than $e^{-\delta t}$ as $t \to \infty$. The representation of the LTI controller can be expressed as follows:

$$\begin{aligned} \bar{u}(t) &= C(p)\bar{y}(t), \quad e^{\delta t}u(t) = C(p)e^{\delta t}y(t) \\ u(t) &= e^{-\delta t}C(p)e^{\delta t}y(t), \quad u(t) = C(p+\delta)y(t). \end{aligned}$$

Therefore, the LTI controller for the plant (2) can be obtained by shifting inversely the imaginary axis.

In designing an adaptive controller, the imaginary axis in the transfer function of the error system is shifted to improve the tracking speed. If an adaptive controller for the δ-shifted plant is given by

$$\bar{u}(t) = f(p, t, \bar{y}(t), \bar{u}(t)),$$

the controller for the original plant becomes

$$u(t) = e^{-\delta t} f(p, t, e^{\delta t}y(t), e^{\delta t}u(t)). \tag{7}$$

4. PARAMETER ADJUSTMENT LAW BASED ON RICCATI EQUATION

In this section, we propose a modified parameter adjustment law that does not require the strictly positive realness.

The error system for the SISO plant in the presence of a bounded disturbance $w(t)$, is given by

$$\dot{e}(t) = Ae(t) + b_1(\delta_c(\hat{k}(t) - k)^T\zeta(t)) + b_2 w(t) \tag{8}$$

$$e_y(t) = c_1^T e(t) + d_1 w(t) \tag{9}$$

where $A \in R^{n\times n}, b \in R^n$ and $c \in R^n$ are known, A is asymptotically stable, $e \in R^n$ is an error state vector, $e_y \in R$ is an output error, k is an unknown parameter vector containing plant parameters, $\hat{k}$ is an estimate of k, $\zeta \in R^m$ is a piecewise continuous regressor vector, $w(t)$ is a disturbance with a bounded power norm, and δ_c is an unknown scalar parameter.

We suppose the following:

(1) δ_c is limited such that $0 < \delta_n < \delta_c < \delta_m$ in which the lower and upper bounds, δ_n and δ_m, are known.
(2) (A, b, c^T) is of minimal phase type.

The adaptive gain error $\theta(t)$ is defined as

$$\theta(t) = \hat{k}(t) - k.$$

We propose the following parameter adjustment law:

$$\left.\begin{aligned} \dot{\hat{k}}(t) &= -\alpha\zeta(t)J(s)e_y(t), \\ J(s) &= F(s) + \frac{\gamma}{\alpha}, \end{aligned}\right\} \quad (10)$$

where

$$F(s) = \frac{1}{f}h(s)N_a(s)^{-1},$$
$$h(s) = \frac{1}{(\tau s+1)^{n_*+m}},$$
$$N_a(s) = c^T(sI-A)^{-1}b,$$

$m > 0$, n_* is the relative degree of $N_a(s)$ and f is any positive constant less than or equal to δ_n. The transfer function $h(s)$ is stable with the same relative degree as n_* and satisfies the following equation:

$$|1 - h(j\omega)| \le \epsilon_0 \text{ for all } \omega < \omega_0$$

The constant ϵ_0 decreases and the frequency ω_0 increases as the parameter $\tau > 0$ in $h(s)$ decreases. The proposed method can be considered as a frequency weighted gradient-type adaptive law. Noting that $N_a(s)$ is of minimal phase type and the transfer function form of $e_y(t)$ is given by

$$e_y(t) = N_a(s)\{\delta_c\theta^T\zeta(t)\},$$

we obtain

$$F(s)e_y(t) = \frac{\delta_c}{f}h(s)\{\theta^T\zeta(t)\}.$$

The following lemma assures that the parameter adjustment law (10) makes the error system (8) and (9) robustly stable.

Theorem 1. Consider the error equations (8) and (9) with the parameter adjustment law (10), where $w(t)$ is a disturbance with a bounded power norm satisfying the following inequality:

$$\lim_{T\to\infty}\frac{1}{T}\int_0^T w(t)^T w(t)dt \le M^2.$$

If the frequency of $\theta(t)^T\zeta(t)$ is lower than ω_0, and there exist positive definite matrices P, Q and positive numbers β, ϕ, ψ and ϵ_1 such that

$$\begin{aligned} &A^TP + PA + \frac{\delta_m}{\beta^2}(Pb_1 - c_1)(b_1^TP - c_1^T) \\ &+\frac{1}{\phi^2}Pb_2b_2^TP = -Q - 2\epsilon_1c_1c_1^T \quad (11) \\ &K_1 > 0, \quad (12) \end{aligned}$$

then $(e(t), \theta(t)^T\zeta(t))$ has a finite power norm. Moreover, the following inequality is satisfied:

$$\lim_{T\to\infty}\frac{1}{T}\int_0^T (c_1^Te(t))^2dt \le (\frac{\delta_m\psi^2d_1^2+\phi^2}{2\epsilon_1} + \frac{K_2^2}{4\epsilon_1K_3}\gamma_0^2)M^2$$

where

$$K_1 = \delta_m\left(\frac{\alpha}{\gamma}(1-\epsilon_0) - \frac{\beta^2}{2} - \frac{1}{2\psi^2}\right), K_2 = \frac{\alpha\delta_m}{\gamma}$$
$$K_3 = \delta_m\left(\frac{\alpha}{f\gamma} - \frac{\beta^2}{2} - \frac{1}{2\psi^2}\right)$$
$$\gamma_0 = ||F(s)c_1^T(sI-A)^{-1}b_2 + d_1||_\infty.$$

5. SYSTEM MODEL

5.1 *Motor model*

The induction motor can be described in a stator fixed (a, b) reference frame by the following equations (Bodson, 1994; Kim, 1997; Marino, 1998):

$$\dot{x} = A(\omega)x + Bu \quad (13)$$
$$\dot{\theta} = \omega \quad (14)$$
$$\dot{\omega} = -\frac{1}{J}\omega + \mu(\psi_{ra}i_{sb} - \psi_{rb}i_{sa}) - \frac{1}{J}\tau_L \quad (15)$$

$$x = \begin{bmatrix} i_{sa} \\ i_{sb} \\ \psi_{ra} \\ \psi_{rb} \end{bmatrix}, \quad u = \begin{bmatrix} u_{sa} \\ u_{sb} \end{bmatrix}$$

$$A(\omega) = \begin{bmatrix} -\gamma & 0 & \xi\eta & n_p\omega\xi \\ 0 & -\gamma & -n_p\omega\xi & \xi\eta \\ L_{sr}\eta & 0 & -\eta & -n_p\omega \\ 0 & L_{sr}\eta & n_p\omega & -\eta \end{bmatrix}$$

$$B = \begin{bmatrix} \frac{1}{\sigma L_s} & 0 \\ 0 & \frac{1}{\sigma L_s} \\ 0 & 0 \\ 0 & 0 \end{bmatrix}$$

where

$$\sigma = 1 - \frac{L_{sr}^2}{L_s L_r}, \mu = \frac{n_p L_{sr}}{J L_r}$$
$$\eta = \frac{R_r}{L_r}, \xi = \frac{L_{sr}}{\sigma L_r L_s}$$
$$\gamma = \frac{L_{sr}^2 R_r + L_r^2 R_s}{\sigma L_r^2 L_s}.$$

In the above equations, we define the following notations:

i_{sa}, i_{sb}	stator currents
ψ_{ra}, ψ_{rb}	rotor fluxes
$L_s r$	mutual inductance
$L_s(, L_r)$	stator (, rotor) inductance
$R_s(, R_r)$	stator (, rotor) resistance
$u = \begin{bmatrix} u_{sa} \\ u_{sb} \end{bmatrix}$	stator voltages
θ	rotor position
ω	mechanical speed
τ	electromagnetic torque
τ_L	load torque
n_p	pole pair number
J	moment of inertia

5.2 *Feedback controller*

A current-fed induction motor with the field-oriented control (FOC) can be approximated by a DC servo-motor(Dote, 1998).

Denoting the modulus β and the angle ρ of the flux as

$$\beta = \sqrt{\psi_{ra}^2 + \psi_{rb}^2}$$
$$\rho = \arctan \frac{\psi_{rb}}{\psi_{ra}},$$

we have the following dynamic equations for the current-fed field-oriented controller (Kim, 1997):

$$\dot{\theta} = \omega \tag{16}$$
$$\dot{\omega} = -\frac{B}{J}\omega + \mu i_2^{new} - \frac{1}{J}\tau_L \tag{17}$$
$$\dot{\beta} = -\eta\beta + L_{sr}\eta i_1^{new} \tag{18}$$
$$\dot{\rho} = \frac{R_r \mu}{n_p \beta^2 J} i_2^{new}. \tag{19}$$

The input variables of the current-fed field-oriented controller are i_1^{new} and i_2^{new}, and are designed as the following scheme:

- The first input i_1^{new} is designed so that $\beta \to \beta_{dr}$ (β_{dr}:constant).
- The second input i_1^{new} is designed so that the rotor position and its speed converge to desired values.

The actual control signal of the current-fed FOC is given by

$$\begin{bmatrix} u_{sa} \\ u_{sb} \end{bmatrix} = \frac{1}{\epsilon}\{e^{\mathcal{T}\rho} \begin{bmatrix} 1 & 0 \\ 0 & \frac{1}{\beta} \end{bmatrix} \begin{bmatrix} i_1^{new} \\ i_2^{new} \end{bmatrix} - \begin{bmatrix} i_{sa} \\ i_{sb} \end{bmatrix}\}$$

where

$$e^{\mathcal{T}\rho} = \begin{bmatrix} \cos\rho & -\sin\rho \\ \sin\rho & \cos\rho \end{bmatrix}.$$

We employ the indirect FOC since there is no need to estimate the rotor flux.

- **Indirect FOC :** The rotor fluxes are replaced to their estimates.

$$\begin{bmatrix} u_{sa} \\ u_{sb} \end{bmatrix} = \frac{1}{\epsilon}\{e^{\mathcal{T}\hat{\rho}_d} \begin{bmatrix} 1 & 0 \\ 0 & \frac{1}{\beta_d} \end{bmatrix} \begin{bmatrix} i_1^{new} \\ i_2^{new} \end{bmatrix} - \begin{bmatrix} i_{sa} \\ i_{sb} \end{bmatrix}\}$$

β_d and $\hat{\rho}_d$ are the estimates of rotor flux given by

$$\beta_d = \beta_{dr} \tag{20}$$
$$\dot{\hat{\rho}}_d = \frac{\hat{R}_r \mu}{n_p \beta_{dr}^2 J} i_2^{new} \tag{21}$$

where $\hat{R}_r$ is the estimate of the rotor resistance.

The control commands i_1^{new} and i_2^{new} are selected as follows:

· i_1^{new} is given by the following equation so as to trace the desired value β_{dr}:

$$i_1^{new} = \frac{\beta_{dr}}{L_{sr}}$$

· i_2^{new} is generated by the adaptive controller so as to satisfy the desired position and speed.

6. ADAPTIVE CONTROLLER DESGIN OF INDUCTION MOTOR BASED ON DC MOTOR MODEL

6.1 *Control method*

The controller design procedure consists of the following steps:

1. Derivation of the approximate model of the current-fed induction motor with FOC.
2. Derivation of the input-output relation based on the coprime factorization.
3. Design of the adaptive controller with the spiral vector.

6.2 *Design model of induction motor*

When the high-gain current feedback of a servo-motor is carried out, the mechanical equation of a motor with model uncertainties is given by the following equation (Dote, 1998):

$$(\hat{J}+\Delta J)\frac{d^2\theta}{dt^2}+(\hat{D}+\Delta D)\frac{d\theta}{dt} = (\hat{K}_T+\Delta K_T)i - T_L \quad (22)$$

θ	rotor angle (output)
$\omega = \frac{d\theta}{dt}$	rotor speed
i	current in the motor (input)
T_L	load
$\hat{J}$	nominal value of inertia
$\hat{D}$	nominal value of friction coefficient
$\hat{K}_T$	nominal value of torque constant
ΔJ	variation of inertia
ΔD	variation of friction coefficient
ΔK_T	variation of torque constant

It is assumed that the available signal is just a rotor angle θ. The transfer function from the current to the rotor angle is

$$\theta(t) = \frac{\hat{K}_T}{p(\hat{J}p+\hat{D})}\{i(t) - \frac{1}{\hat{K}_T}T_e(t)t\} = G(p)\{i(t) - \frac{1}{\hat{K}_T}T_e(t)\}$$

where $T_e(t)$ is the following equivalent torque including the parameter variations:

$$T_e(t) = \Delta J\frac{d^2\theta(t)}{dt^2} + \Delta D\frac{d\theta(t)}{dt} - \Delta K_T i(t) + T_L(t)$$

The reference model is

$$(p^2+m_1p+m_2)\theta_m(t) = m_2 i_m(t)$$

where $r_m(t)$ is a reference input, $y_m(t)$ is a reference output, and $p^2+m_1p+m_2$ is a stable polynomial.

The spiral vector forms of the signals of the motor are defined as

$$i(t) = e^{-\delta t}\bar{i}(t), \quad \theta(t) = e^{-\delta t}\bar{\theta}(t)$$
$$T_e(t) = e^{-\delta t}\bar{T}_e(t), \quad \theta_m(t) = e^{-\delta t}\bar{\theta}_m(t)$$
$$\bar{r}_m(t) = e^{-\delta t}\bar{i}_m(t).$$

The input-output relation between the spiral vector signals $\bar{\theta}$ and $\bar{i}$, is given by

$$\bar{\theta}(t) = e^{\delta t}G(p)e^{-\delta t}(\bar{i}(t) - \bar{T}'_e(t)) = G(p-\delta)(\bar{i}(t) - \bar{T}'_e(t))$$

where

$$G(p) = \frac{\hat{K}_T}{p(\hat{J}p+\hat{D})}, \quad G(p-\delta) = \frac{\bar{\beta}_1}{p^2+\bar{\delta}_1p+\bar{\delta}_2}$$
$$\bar{T}'_e(t) = \frac{1}{\hat{K}_T}T_e(t).$$

6.3 *Input-output relation*

In this section, we design an adaptive controller of the servomotor using the coprime factorization(Matsuo, 1995).

The coprime factorization of $G(p-\delta)$ is given by

$$G(p-\delta)) = \frac{\bar{\beta}_1}{p^2+\bar{\delta}_1p+\bar{\delta}_2} = \frac{N(p)}{M(p)}$$

where

$$N(p) = \frac{\bar{\beta}_1}{p^2+d_1p+d_0}, \quad M(p) = \frac{p^2+\bar{\delta}_1p+\bar{\delta}_2}{p^2+d_1p+d_0}.$$

$N(p)$ and $M(p)$ satisfy the following Bezout identity:

$$\tilde{Y}(p)M(p) + \tilde{X}(p)N(p) = 1 \quad (23)$$

From the above equation, the input-output relation is obtained such that

$$\bar{\theta}(t) = N(p)\{\bar{i}(t) + \tilde{X}(p)\bar{\theta}(t) - (1-\tilde{Y}(p))\bar{i}(t) - \tilde{Y}(p)\bar{T}'_e(t)\}$$

Using the spiral vector signals $\bar{\theta}_m$ and $\bar{i}_m$ the reference model is rewritten as

$$\bar{\theta}_m(t) = \frac{m_2}{p^2+\bar{m}_1p+\bar{m}_2}\bar{i}_m(t) = \frac{\bar{\beta}_1}{p^2+d_1p+d_0}\{\frac{1}{\bar{\beta}_1}\frac{m_2(p^2+d_1p+d_0)}{p^2+\bar{m}_1p+\bar{m}_2}\bar{i}_m(t)\} = N(p)\{\frac{1}{\bar{\beta}_1}\zeta_0(t)\}$$
$$\zeta_0 = \frac{m_2(p^2+d_1p+d_0)}{p^2+\bar{m}_1p+\bar{m}_2}\bar{i}_m(t).$$

6.4 *Controller without disturbance estimator*

Neglecting the equivalent disturbance term in designing a controller, we apply the parameter adjustment law based on a Riccati equation to the error equation. After re-shifting the imaginary axis, the actual input of the original plant satisfies

$$i_2^{new}(t) = e^{-\delta t}(\hat{k}(t)^T\xi(t))$$
$$\xi(t) = \begin{bmatrix} \frac{m_2(p^2+d_1p+d_0)}{p^2+\bar{m}_1p+\bar{m}_2}e^{\delta t}r_m(t) \\ \frac{p}{p^2+d_1p+d_0}e^{\delta t}\theta(t) \\ \frac{1}{p^2+d_1p+d_0}e^{\delta t}\theta(t) \\ \frac{p}{p^2+d_1p+d_0}e^{\delta t}i(t) \\ \frac{1}{p^2+d_1p+d_0}e^{\delta t}i(t) \end{bmatrix}.$$

The parameter adjustment law is given by

$$\frac{d}{dt}\hat{k}(t) = \{-\alpha\xi(p)\frac{p^2 + d_1 p + d_0}{(\tau p + 1)^2}\bar{e}(t) - \gamma\xi(t)\bar{e}(t)\}.$$

6.5 *Controller with disturbance estimator*

If the variations of the motor parameters $\Delta J, \Delta D$ and ΔK_T are almost constant, it is not necessary to add the adaptive estimator of these parameters to the adaptive controller. Since the adaptive controller can estimate the plant parameters including the variations, we add the following adaptive estimator of the load T_L to the controller. The input law is given by

$$i_2^{new}(t) = e^{-\delta t}\left[\,\hat{k}^T(t)\ \hat{l}(t)\,\right]\begin{bmatrix}\xi(t)\\ e^{\delta t}\end{bmatrix}.$$

To avoid the instability of the internal signal in the adaptive loops, δ is switched as follows:

$$\delta = \begin{cases}\delta_{sp} & (t \le 1)\\ 0 & (t > 1)\end{cases} \tag{24}$$

7. SIMULATION RESULTS

The performance of the adaptive control scheme for rotor position was investigated through simulation with MATLAB/SIMULINK. We used the motor parameters in Table 1 as in Bodson *et al*(1994). The reference model was $\frac{100}{(p+10)^2}$. Figure

Table 1. Motor parameters

R_r	3.9	Ω	R_s	1.7	Ω
L_r	0.014	H	L_s	0.014	H
L_{sr}	0.011	H	n_p	3	
J	0.00011	kgm^2	B	0.00014	Nm/rad/s

1 shows the closed-loop responses of the rotor position when the estimate of the rotor resistace $\hat{R}_r$ is $\frac{1}{3}R_r$ and the adaptive controller without the disturbance estimator is implemented with the spiral vector factor $\delta = 0.01, 6$ and 9. The simulation results showed that the adaptive controller did not need the disturbance estimator and the switching of the spiral vector factor was available to reduce the overshoot maginitude of the closed-loop system in the transient period.

8. CONCLUSIONS

In this paper, we considered the relationship between the spiral vector method in the motor transient analysis and the imaginary axis shifting in the control theory. Comparing the proposed adaptive controller with other nonlinear control design methods of induction motors, our adaptive mechanism has a simple structure, because it

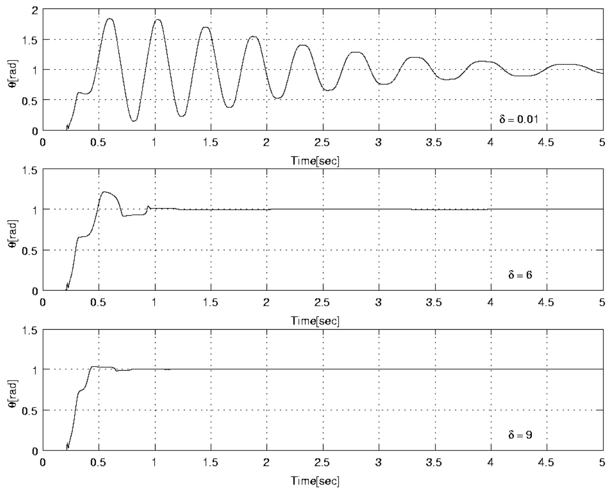

Fig. 1. Closed-loop position responses when $\delta_{sp} = 0.01(top), 6(middle), 9(bottom)$.

can be designed based on modified gradient-type adaptive algorithm and the plant model for the controller design is a 2nd order linear system.

9. REFERENCES

Bodson, M. , J. Chiasson and R. Novotnak (1994). High-Performance Induction Motor Control via Input-Output Linearization. *IEEE Control System Magazine, August*, 25-33.

Dote,Y. and R .G Hoft (1998). *Intelligent Control, Power Electronic Systems*, Oxford University Press.

Ioannou,P.A. and J. Sun (1996). *Robust Adaptive Control.* Prentice-Hall.

Kim,K.-C., R. Ortega, A. Charara and J-P. Vilain (1997). Theoretical and Experimental Comparison of Two Nonlinear Controllers for Current-Fed Induction Motors. *IEEE Transaction on Control Systems Technology*, **5-3**, 338-348.

Marino,R., S.Peresada and P.Tomei (1998). Adaptive Output Feedback Control of Current-Fed Induction Motors with Uncertain Rotor Resistance and Load Torque. *Automatica*, **34-5**, 617-624.

Matsuo,T. , K. Tsunetsugu and K. Nakano (1995). A Design of Robust Adaptive Control System with a Relaxed Strictly-Positive-Realness Condition. *Proc. of the 34th Conf. on Decision and Control*, 2322–2327.

Narendra,K. S. and A. M. Annaswamy (1989). *Stable Adaptive Systems*, Prentice Hall.

Yamamura,S. (1992). *Spiral Vector Theory of AC Circuit and Machines*, Oxford university Press.