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A Fuzzy Predictive Sliding Mode Control for High Performance Induction Motor Position Drives

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ABSTRACT

This paper presents a fuzzy predictive sliding mode control for high performance induction motor position drives. A new simplified inner-loop sliding-mode current control scheme based on a nonlinear mathematical model of an induction motor is introduced. Novel predictive fuzzy logic PI and PID controllers are used in speed and position loops, respectively. Sliding-mode current controllers and fuzzy predictive logic controllers are designed based on indirect vector control. The overall system performance is examined under different dynamic operating conditions. The performance of the drive system is robust and stable, and insensitive to parameters and operating condition variations even though non-exact system parameters are used in the implementation of the proposed controllers.

Keywords: Induction motor drive systems, fuzzy predictive control and sliding mode control.

1. Introduction

In recent years the control of high performance induction motor drives for general industrial applications and production automation has received widespread research interest. High performance servo position drives require high accuracy, fast response and robust dynamics. Attention has been given to the induction machine for reasons of cost, size, weight, reliability, ruggedness, simplicity, efficiency and ease of manufacture. The application of advanced control schemes for position-speed control of induction machines has been made possible by the increasing power and falling cost of microprocessors and digital signal processors(DSPs)[1].

It is well known that Field-Oriented Control(FOC) is an effective method for variable speed control of induction motor drives. However, difficulties arise due to modelling uncertainties which affect the linear controller such as parameter variations, magnetic saturation, load disturbances and un-modelled dynamics[2]. Thus, the controller parameters have to be continually adapted. To ensure good performance dynamics, various robust control strategies for induction motor(IM) drives have been reported in the literature. The problem can be solved by several adaptive control techniques, such as model reference adaptive control(MARC)[3], sliding mode control (SMC)[4], variable structure control(VSC)[5], and self-tuning PI controllers[6]. The design of all the above

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controllers depends on the exact system mathematical model. However, it is often difficult to develop an accurate system mathematical model due to unknown load variation, unknown and unavoidable parameter variations caused by saturation, temperature variations, and system disturbances. In order to overcome the above problems, the fuzzy-logic controller(FLC) has been introduced recently, and is being used for motor control purposes [8]-[11]. The mathematical tool for the FLC is the fuzzy set theory introduced by Zadeh[7]. As compared to the conventional PI, PID, and their adaptive versions, the FLC has the following advantages: 1) it does not need an exact system mathematical model; 2) it can handle nonlinear equations of arbitrary complexity; and 3) it is based on the general linguistic IF–THEN structure, which undergirds human logic. However, the application of FLC has faced some disadvantages during hardware and software implementation, due to its high computational burden[8].

Based on the inaccurate system mathematical model of induction motors due to unknown load variation, unavoidable parameter variations, saturation, temperature variations, and system disturbance may occur. This paper presents the hybrid controllers of SMC and Predictive Fuzzy PID. The Sliding Mode(SM) current control is designed in the phase coordinates with reference to the current being transformed from the field-oriented coordinates. The speed and position controllers are designed by using Predictive Fuzzy Logic Control(FLC). Their parameters are taken into account in the sliding mode current control loop.

2. IM Dynamic Model

The dynamic model of an IM can be represented by a set of highly nonlinear differential equations. Assuming linear magnetic circuits, identical mutual inductance, and neglecting the iron losses [2], the nonlinear dynamic model in synchronous reference frame is given by [2, 12]:

$$\frac{di_{ds}}{dt} = -\gamma\, i_{ds} + \eta\,\mu\,\lambda_{dr} + \omega_s\, i_{qs} + \omega_r\,\mu\,\lambda_{qr} + \frac{1}{\sigma L_s} V_{ds} \quad (1)$$

$$\frac{di_{qs}}{dt} = -\gamma\, i_{qs} - \omega_r\,\mu\,\lambda_{dr} - \omega_s\, i_{ds} + \eta\,\mu\,\lambda_{dr} + \frac{1}{\sigma L_s} V_{qs} \quad (2)$$

$$\frac{d\lambda_{dr}}{dt} = \eta\, L_m - \eta\,\lambda_{dr} + \omega_{sl}\,\lambda_{qr} \quad (3)$$

$$\frac{d\lambda_{qr}}{dt} = \eta\, L_m\, i_{qs} - \eta\,\lambda_{qr} - \omega_{sl}\,\lambda_{dr} \quad (4)$$

$$T_e = \frac{3 P L_m}{4 L_r}(i_{qs}\,\lambda_{dr} - i_{ds}\,\lambda_{qr}) \quad (5)$$

$$T_e - T_L = \frac{2}{P}\left(J\frac{d\omega_r}{dt} + B\,\omega_r\right) \quad (6)$$

$$\dot{\theta}_m = \frac{\omega_r}{2P} \quad (7)$$

where,

$$\eta = \frac{R_r}{L_r},\ \sigma = 1 - \frac{L_m^2}{L_s L_r},\ \mu = \frac{L_m}{\sigma L_s L_r},$$

$$\gamma = \frac{1}{\sigma L_s}\left(R_s + \frac{L_m^2}{L_r^2} R_r\right), \text{ and } \omega_{sl} = \omega_s - \omega_r \quad (8)$$

v_{ds}, v_{qs} : Stator voltages in the *d-q* reference frame.
i_{ds}, i_{qs} : Stator currents in the *d-q* reference frame.
λ_{dr}, λ_{qr} : Rotor flux linkages in the d-q coordinates.
T_e : Electromagnetic torque of the IM.
T_L : Load or disturbance torque.
P : Number of poles of the IM.
ω_s : Stator angular frequency (rad/sec).
ω_r : Rotor electrical speed in angular frequency.
ω_{sl} : Slip angular frequency of the IM.
θ_m : Rotor mechanical angular position.
R_r, R_s :Rotor and stator resistance referred to the stator.
L_r, L_s : Rotor and stator inductance referred to the stator.
L_m : Mutual inductance referred to the stator.
J : Moment of inertia of the IM.
B : Motor frictions.

In vector control the rotor flux in the *q*-axis is set equal to zero (λ_{qr}=0 and $p\lambda_{qr}$ =0). Therefore, Eqs. (1-4) of the IM mathematical model becomes:

$$\frac{di_{ds}}{dt} = -\gamma\, i_{ds} + \eta\,\mu\,\lambda_{dr} + \omega_s\, i_{qs} + \frac{1}{\sigma L_s} V_{ds} \quad (9)$$

$$\frac{di_{qs}}{dt} = -\gamma\, i_{qs} - \omega_r\,\mu\,\lambda_{dr} - \omega_s\, i_{ds} + \frac{1}{\sigma L_s} V_{qs} \quad (10)$$

$$\frac{d\lambda_{dr}}{dt} = \eta L_m i_{ds} - \eta \lambda_{dr} \quad (11)$$

$$0 = \eta L_m i_{qs} - \omega_{sl} \lambda_{dr} \quad (12)$$

If the d-axis rotor flux λ_{dr} is kept constant, the generated torque will be linearly proportional to i_{qs}.

This yields $p\lambda_{dr}$ =0 and the last two Eqs. (11) and (12) give:

$$\omega_{sl} = \eta \frac{i_{qs}}{i_{ds}} \quad (13)$$

3. Sliding Mode Current Control

The IM drive system is a multi-input multi-output nonlinear system. The problem with the traditional PI linear control technique is that the controller design is dependent on the operating point, and the decoupling between d- and q-axis is sensitive to parameter errors. On the other hand, the dynamics of SMC are not dependent on the working point as long as it remains in sliding mode. Hence, compensation for nonlinearities as well as order reduction can be achieved. Decoupling can also be avoided due to the capability of model error compensation. SMC offers an alternative solution for the inner current control loop, as seen in the following section.

3.1 Current Control

Eq. (9) and (10) describe current components in d and q coordinates. We can rewrite them simply as:

$$\begin{aligned} \frac{di_{ds}}{dt} &= f_d + bu_d \\ \frac{di_{qs}}{dt} &= f_q + bu_q \end{aligned} \quad (14)$$

where,

$$\begin{aligned} f_d &= -\gamma i_{ds} + \eta \mu \lambda_{dr} + \omega_s i_{qs} \\ f_q &= -\gamma i_{qs} - \omega_r \mu \lambda_{dr} - \omega_s i_{ds} \\ b &= \frac{1}{\sigma L_s} \end{aligned} \quad (15)$$

In matrix form:

$$\dot{I}_{dqs} = f_{dq}(I_{dqs}, \lambda_{dr}, \omega_r, \omega_s) + bU_{dq} \quad (16)$$

where,

$$I_{dqs} = \begin{bmatrix} i_{ds} \\ i_{qs} \end{bmatrix},\ U_{dq} = \begin{bmatrix} u_d \\ u_q \end{bmatrix},\ f_{dq} = \begin{bmatrix} f_d \\ f_q \end{bmatrix} \quad (17)$$

To design the SMC for the current loop, the following tasks must be performed. First, design the switching functions for the current control, which is:

$$\begin{aligned} s_d &= i_{ds}^* - i_{ds} \\ s_q &= i_{qs}^* - i_{qs} \end{aligned} \quad (18)$$

The second task is to determine the conditions under which sliding mode can be enforced. In our system there is no control gain to be adjusted. The right domain is to be found in the space where the trajectory converges to the sliding manifold. So, let

$$S_{dq} = \begin{bmatrix} s_d \\ s_q \end{bmatrix} \quad (19)$$

Taking the time derivative of S_{dq} then

$$\dot{S}_{dq} = \dot{I}_{dq}^* - f_{dq}(I_{dq}, U, \omega) + bU_{dq} \quad (20)$$

Design a Lyapunov function candidate

$$V = \frac{1}{2} S_{dq}^T S_{dq} \quad (21)$$

Differentiate and substitute in (20) so,

$$\dot{V} = (S^*)^T F^* + (S^*)^T D^T DU_{gate} \quad (22)$$

where,

$$F^* = \begin{bmatrix} F_1^* \\ F_2^* \\ F_3^* \end{bmatrix} = D^T F_{dq} \quad (23)$$

$$D^T D = \left(\frac{2}{3}b\right)^2 \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ -\frac{1}{2} & 1 & -\frac{1}{2} \\ -\frac{1}{2} & -\frac{1}{2} & 1 \end{bmatrix} \quad (24)$$

$$U_{gate} = \begin{bmatrix} u_1 \\ u_2 \\ u_3 \end{bmatrix},\quad U_{gate} = -sign(S^*) \quad (25)$$

where,

$$S^* = \begin{bmatrix} s_1^* \\ s_2^* \\ s_3^* \end{bmatrix} = \frac{3}{2b} \begin{bmatrix} \cos\rho_a & -\sin\rho_a \\ \cos\rho_b & -\sin\rho_b \\ \cos\rho_c & -\sin\rho_c \end{bmatrix} \begin{bmatrix} s_d \\ s_q \end{bmatrix} \quad (26)$$

$$\rho_a = \rho = \tan^{-1}(\frac{\lambda_\beta}{\lambda_\alpha}),\ \rho_b = \rho - \frac{2\pi}{3},\ \rho_c = \rho + \frac{2\pi}{3} \quad (27)$$

and,

λ_β, λ_α : rotor flux linkage at α-β co-ordinate.

S^* is a vector of the transformed switching functions. The controls u_1, u_2 and u_3 take values from the discrete set {-1,1}. The transformed vector S^* should be designed such that under controls of (25), s_d and s_q vanish in finite time. Vector S^* is selected as

$$S^* = \frac{3}{2b^2} D^T S_{dq} \quad (28)$$

The condition which verify the convergence of the system to the manifold ($\dot{V}<0$) is

$$(\frac{4}{9})b^2 \geq \max(|F_1^*|,|F_2^*|,|F_3^*|) \quad (29)$$

The inequality (28) defines a subspace in the system space, in which the state trajectories converge to the sliding manifold S_{dq}=0 in finite time. This shows that the attraction domain of the sliding mode manifold is bounded in the state space.

The final task is to map the resulting controls u_1, u_2 and u_3 into the switching patterns which are applied to the power inverter.

3.2 Rotor Flux Estimation

From the motor model in α-β coordinates, the rotor flux equations are:

$$\frac{d\lambda_\alpha}{dt} = -\eta\,\lambda_\alpha - \omega_s\,\lambda_\beta + \eta\,L_m\,i_\alpha \quad (30)$$

$$\frac{d\lambda_\beta}{dt} = -\eta\,\lambda_\beta + \omega_s\,\lambda_\alpha + \eta\,L_m\,i_\beta \quad (31)$$

The original equations for the rotor flux can be used directly as the observer equations. This observer is stable and the rate of convergence depends on the motor time constant. The flux observer equations are given by:

$$\frac{d\hat{\lambda}_\alpha}{dt} = -\eta\,\hat{\lambda}_\alpha - \omega_s\,\hat{\lambda}_\beta + \eta\,L_m\,i_\alpha \quad (32)$$

$$\frac{d\hat{\lambda}_\beta}{dt} = -\eta\,\hat{\lambda}_\beta + \omega_s\,\hat{\lambda}_\alpha + \eta\,L_m\,i_\beta \quad (33)$$

where $\hat{\lambda}_\alpha$, $\hat{\lambda}_\beta$ are the estimated rotor flux components. Defining the estimation error as:

$$\bar{\lambda}_\alpha = \hat{\lambda}_\alpha - \lambda_\alpha \quad (34)$$

$$\bar{\lambda}_\beta = \hat{\lambda}_\beta - \lambda_\beta \quad (35)$$

and, the error dynamics as:

$$\frac{d\bar{\lambda}_\alpha}{dt} = -\eta\,\bar{\lambda}_\alpha - \omega_s\,\bar{\lambda}_\beta \quad (36)$$

$$\frac{d\bar{\lambda}_\beta}{dt} = -\eta\,\bar{\lambda}_\beta + \omega_s\,\bar{\lambda}_\alpha \quad (37)$$

To test the stability of the observer, design a Lyapunov function candidate as:

$$V = 0.5(\bar{\lambda}_\alpha^2 + \bar{\lambda}_\beta^2) > 0 \quad (38)$$

Differentiate and substitute in (36) and (37) then,

$$\begin{aligned} \dot{V} &= -\eta(\bar{\lambda}_\alpha^2 + \bar{\lambda}_\beta^2) \\ &= -2\eta V < 0 \end{aligned} \quad (39)$$

Inequality (39) proves that; the difference between the real and the estimated rotor flux tends to zero asymptotically and the rate of convergence depends on the rotor time constant (L_r/R_r). The rotor flux angle $\hat{\rho}$ can be calculated as:

$$\hat{\rho} = \tan^{-1}(\frac{\hat{\lambda}_\beta}{\hat{\lambda}_\alpha}) \quad (40)$$

4. Fuzzy Predictive Logic Control

Fuzzy control is widely used for controlling nonlinear systems whose plant models are unknown or vague. On the other hand, the increasing demands on quality and economic performance of control systems has gradually led to predictive controls gaining more attention[13, 14]. Combining Fuzzy control with predictive control may avoid the drawbacks of model-based generalized predictive control(GPC). Recently, Fuzzy Predictive

Control has received more more attention[15].

Fuzzy Predictive PI and PID controllers are used in speed and position loops respectively. Fig. 2 shows the details of a predictive PID Fuzzy controller.

4.1 Speed Control

The proposed Fuzzy Predictive PI controller is designed by considering the speed error ($e_\omega(k)$) and rate of change of the speed error ($\Delta e_\omega(k)$) as the input linguistic variables and the quadrate stator reference current ($i_q^*(k)$) as the output linguistic variable. The block diagram structure of the proposed PI-fuzzy speed controller is shown in Fig. 2. The equation, which describes the input-output relationship, is given below:

$$u_\omega(k) = i_{qs}^{e*}(k) = \int \Delta i_{qs}^{e*}(k) = f(e_\omega(k), \Delta e_\omega(k)) \qquad (41)$$

Where, $e_\omega(k) = \omega_r^*(k) - \omega_r(k)$ is the speed error and $\Delta e_\omega(k) = e_\omega(k) - e_\omega(k-1)$ is the change in the speed error.

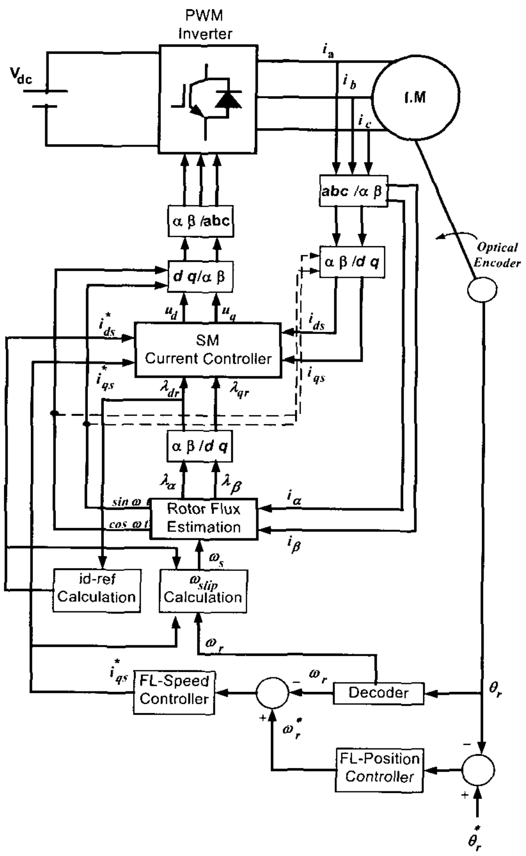

Fig. 1 The proposed predictive fuzzy PDI control system

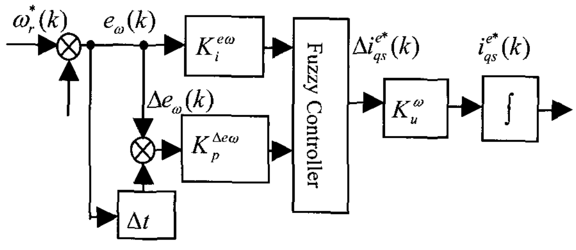

Fig. 2 PI-Fuzzy speed controller

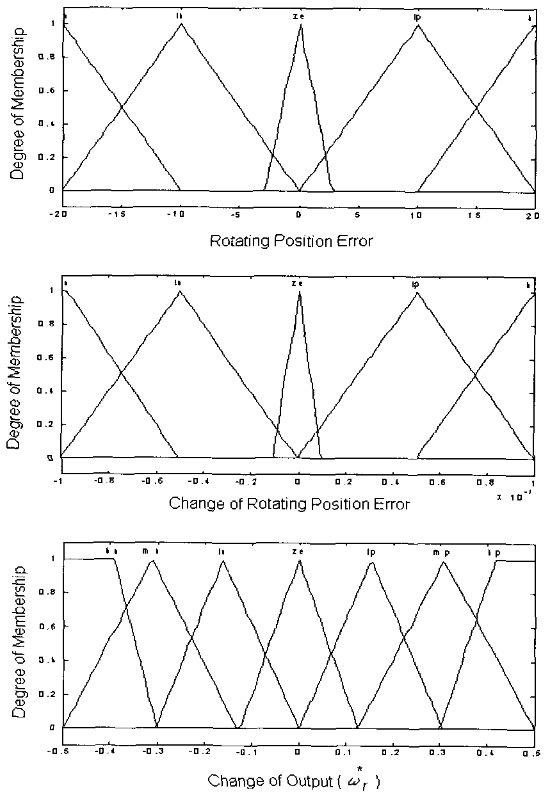

Fig. 3 Membership functions for input and output linguistic variables of the speed control loop

Table 1 Fuzzy-Rule based matrix of speed loop

	Speed Error ($\Delta\omega_r$)							
		hn	mn	ln	ze	lp	mp	hp
Change Of Speed Error (Δe_s)	hp	ze	lp	mp	hp	hp	hp	hp
	mp	ln	ze	lp	mp	hp	hp	hp
	lp	mn	ln	ze	lp	mp	hp	hp
	ze	hn	mn	ln	ze	lp	mp	hp
	ln	hn	hn	mn	ln	ze	lp	mp
	mn	hn	hn	hn	mn	ln	ze	lp
	hn	hn	hn	hn	hn	mn	ln	ze

The scaling factors $K_p^{\Delta e\omega}$, $K_i^{e\omega}$ and K_u^{ω} are selected precisely to tune the output of the Fuzzy controller for the desired speed response.

The membership functions used for $e_\omega(k)$, $\Delta e_\omega(k)$ and $i_q^*(k)$ are triangular for all the Fuzzy sets except the fuzzy sets hn (high negative) and hp (high positive) which are trapezoidal. The membership function employed for the input and the output fuzzy sets are given in Fig. 3. The controller rules are summarized in Table 1. Where, hn (high negative), mn (medium negative), ln (low negative), ze (Zero), lp (low positive), mp (medium positive) and hp (high positive) are linguistic variables. The max-min inference method was used and the defuzzification procedure was based on the centre of area method. Fig. 4 shows the output surface for the speed control.

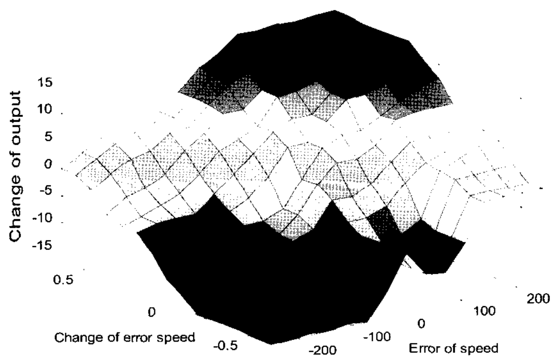

Fig. 4 The output surface for the speed control

4.2 Position Control

The position control loop needs to be tuned precisely. So, the required controller should have fast response, small overshoot, small settling time and approximately zero steady state error. The Predictive Fuzzy PD+I controller is chosen as it has all these characteristics. In speed control loops not all of these features are required for controlling the loop, therefore, a Predictive Fuzzy PI is used. The proposed Predictive Fuzzy PD+I controller is designed by considering the rotating position error ($e_\theta(k)$) and the rate of change of the rotating position error ($\Delta e_\theta(k)$) as the input linguistic variables, and the speed reference *($\omega_r^*(k)$)* as the output linguistic variable.

The block diagram of the position controller is presented in Fig. 5.

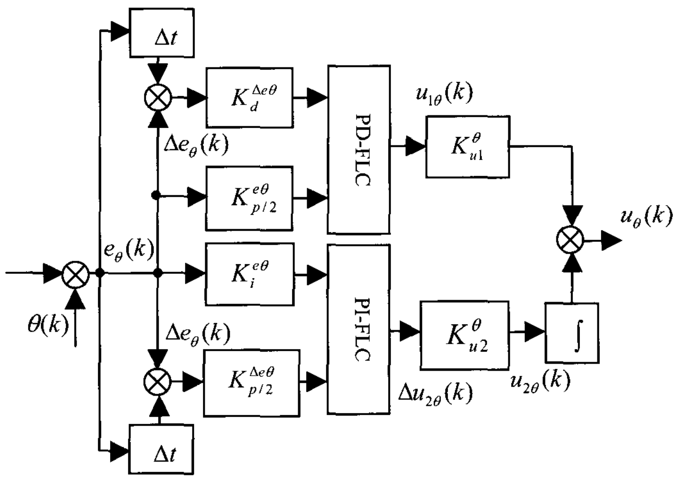

Fig. 5 PID-fuzzy position controller

The input-output relationship of the PID fuzzy position controller is described by:

$$u_\theta(k) = \omega_r^*(k) = u_{1\theta}(k) + \int u_{2\theta}(k) = f(e_\theta(k), \Delta e_\theta(k)) \quad (42)$$

Where, $e_\theta(k) = \theta^*(k) - \theta(k)$ is the position error and

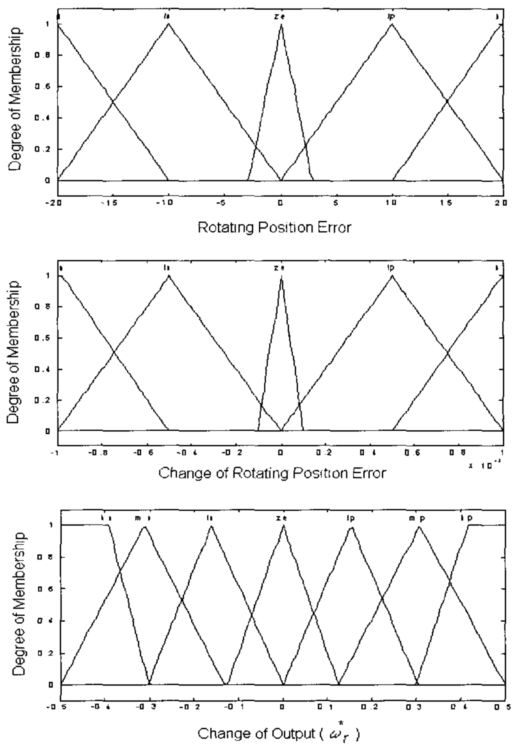

Fig. 6 Membership functions for input and output linguistic variables of the rotating position control loop

$\Delta e_\theta(k) = e_\theta(k) - e_\theta(k-1)$ is the change in the position error.

The scaling factors $K_{p/2}^{\Delta e\theta}$, $K_{p/2}^{e\theta}$, $K_i^{e\theta}$, $K_d^{\Delta e\theta}$ and output gains, K_{u1}^{θ} and K_{u2}^{θ} are selected precisely for the desired output.

The membership function employed for the input and the output fuzzy sets are given in Fig. 6.

The controller rules are summarized in Table 2. Fig. 7 shows the output surface for the speed rotating position control.

Table 2 Fuzzy-Rule based matrix of position loop

	Position Error ($\Delta\theta_r$)					
Change of Position Error (Δe_c)		hn	ln	ze	lp	hp
	hn	hn	hn	mn	ln	ze
	ln	hn	ln	ln	ze	lp
	ze	mn	ze	ze	lp	mp
	lp	ln	ze	lp	lp	mp
	hp	ze	lp	mp	hp	hp

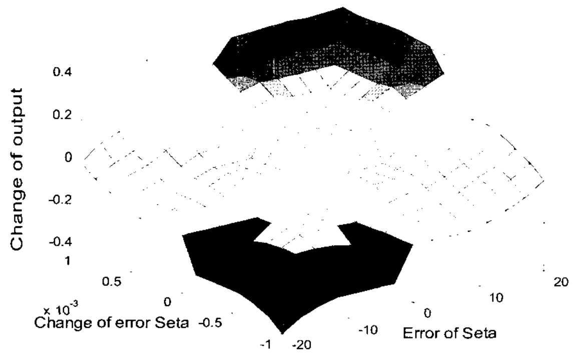

Fig. 7 The output surface for the position control

5. Results

The proposed Fuzzy predictive SMC for high performance IM position drives is simulated. The motor specifications and parameters are given in table 3. Several tests were performed to evaluate the performance of the proposed system. Fig. 8 shows the capability of the controller to resist the effects of load disturbance, which is an important issue for electrical drives. Fig. 8a presents the overall responses of the electromechanical torque and the current component (i_q). The system started at no-load then at t = 2 sec it is suddenly subjected to full load torque. At t = 3 sec the load torque is decreased to 75% of the full load and at t = 4 sec it is increased again to the full load torque. Fig. 8(b-g) shows the responses of the torque, current, speed and position during the sudden step in the load torque from no-load to 100% load then to 75% load and again to 100% load. This reveals that the performance of the overall system is robust in the face of load disturbances.

Fig. 9 presents the overall performance of the system during starting and position step change. At t = 3 sec the system is subjected to a position step change from 12.65 rad to 9.42 rad. Fig. 9(a,b) shows the responses of the position, speed, torque and current during this step change.

This figure shows that the designed system can handle the sudden decrease in the position command quickly without overshoot, under-shoot or steady-state error. This is in contrast to the linear-controller-based drive system which suffers from steady-state error and does not respond quickly to changes. Thus, we can see that the proposed Fuzzy predictive SMC-based drive is superior to other conventional linear-controller-based drive systems.

Table 3 The IM specifications and parameters

Rated Power	4 KW	R_s	1.95 Ω
Rated Current	9.6 A	R_r	1.58 Ω
Rated Speed	2840 rpm	J	0.0127 N.m.s^2/rad
Rated Voltage	400 V	B	0.00195 N.m.s/rad
Rated Torque	12.8 N.m	L_s	0.1554 H
No. of pole	2	L_r	0.1568 H
		L_m	0.1503 H

6. Conclusions

Based on the nonlinear model of induction motors a new simplified inner-loop sliding-mode current control scheme for IM drive has been presented. Due to the invariant property of the sliding-mode control, the current inner-loop possesses very good dynamics. Novel Fuzzy Predictive PI and PID controllers are used in speed and position loops, respectively. Since exact system parameters are not required in the implementation of the proposed controllers, the performance of the drive system is robust and stable, and insensitive to parameter and operating condition variations. Based on the experiments that have been presented which take into account different operating conditions, one can conclude that the proposed Fuzzy predictive SMC offers superior performance and stability, compared to linear-controller-based drive systems, even in the presence of position variation and external load disturbances.

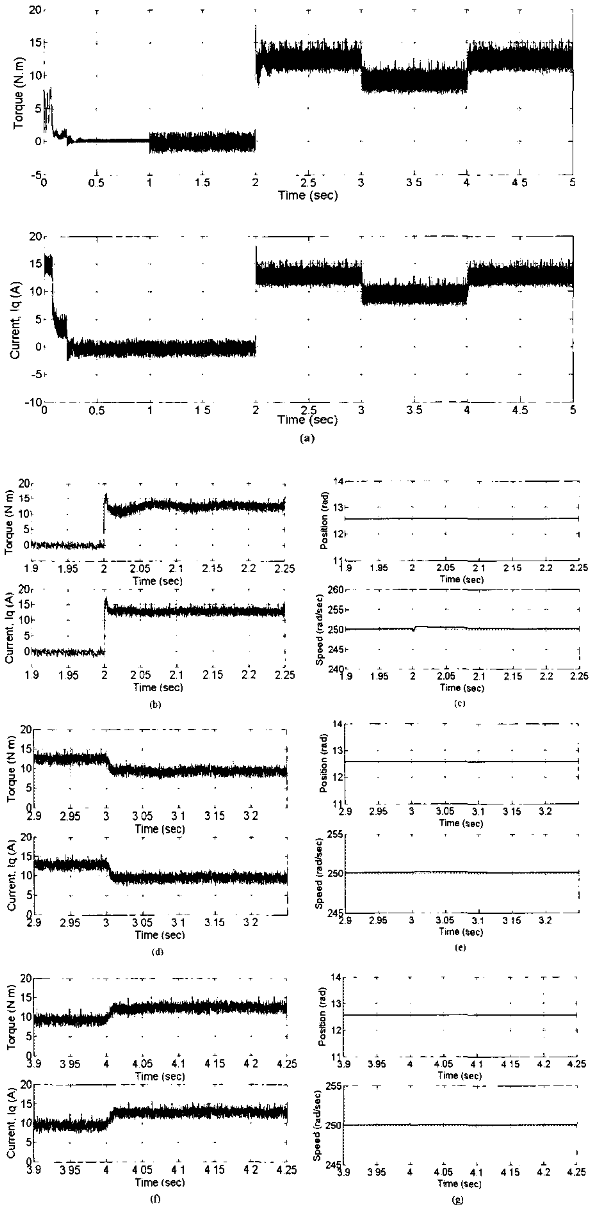

Fig. 8 Torque, current (iq), speed and position during sudden changes in load torque, (a) the overall waveforms of torque and current (iq) during sudden changes, (b, c) from 0% to 100% load, (d, e) from 100% to 75% load, and (f, g) from 75% to 100% load

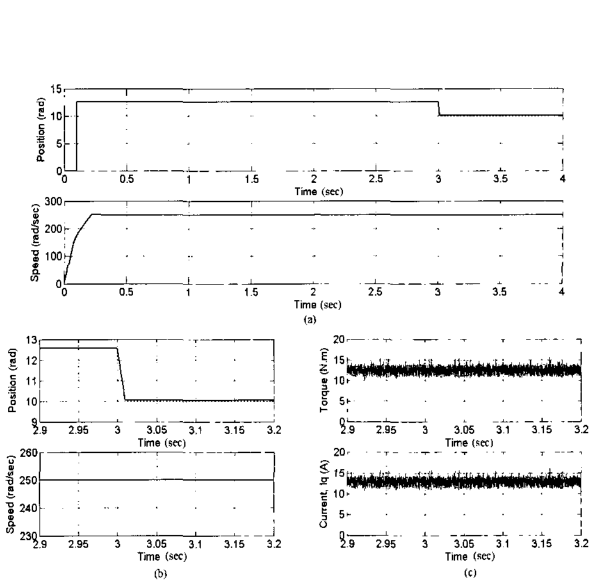

Fig. 9 Position, speed, torque and current during sudden changes in the position, (a) overall position and speed during starting and position sudden change, (b,c) Position step change from 12.65 rad to 9.42 rad

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