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Performance Analysis of Fuzzy based MRAC and Sliding Mode Controls of Vector Controlled Induction Motor Drive

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ABSTRACT: This paper presents the fuzzy based MRAC and sliding mode controls for Indirect vector controlled induction motor drive. In high performance AC drives the motor speed should closely match with the specified reference speed irrespective of the variations in the load, motor parameters and model uncertainties. Soft computing technique – Fuzzy logic is applied in this paper for the speed control of induction motor to achieve maximum torque with minimum loss. The fuzzy logic controller is implemented using the Field Oriented Control technique as it provides better control of motor torque with high dynamic performance. The proposed adaptive controller takes advantage of Model reference adaptive control and fuzzy logic control. An integrated mathematical model of the control scheme has been developed and simulated in MATLAB for Indirect vector control of an Induction motor. The simulated performances of the FL-MRAC slip gain tuner based IVCIM drive is compared to fuzzy based sliding mode controller. Simulation results conclude that the proposed Fuzzy based controller showed increased dynamic performance.

KEY WORDS: Induction motor, Field Oriented Control, Fuzzy logic Controller, Model reference adaptive control (MRAC) and Sliding Mode Control.

I. INTRODUCTION

AC Induction motors are being applied today to a wider range of applications requiring variable speed. field-oriented control technique has been widely used in industry for high-performance induction machine (IM) drive, where the knowledge of synchronous angular velocity is often necessary in the phase transformation for achieving the favorable decoupling control. However the speed control of the induction motors are not simple difficulties due to its complex and nonlinear mathematical model which involves parameters that vary with temperature, frequency and other operating conditions. The variations of parameters have significant effect on the accuracy of control speed and torque and other operating performance of the motor. It is therefore essential to optimize the motion control performance by designing intelligent adaptive controller based on fuzzy logic, neural network and expert systems; so that torque and flux have dynamic ideal response in high performance AC drives.

Advanced control based on artificial intelligence technique is called intelligent control. Intelligent control, act better than conventional adaptive controls. Fuzzy logic is a technique to embody human-like thinking into a control system. Fuzzy control has been primarily applied to the control of processes through fuzzy linguistic descriptions. The motor-control issues are traditionally handled by fixed gain proportional-integral (PI) and proportional-integral derivative (PID) controllers. However, the fixed-gain controllers are very sensitive to parameter variations, load disturbances, etc. Thus, the controller parameters have to be continually adapted or tuned. The problem can be solved by several adaptive control techniques such as model reference adaptive control (MRAC), sliding-mode control (SMC) variable structure control (VSC), and self-tuning PI controllers, etc. In this paper, fuzzy based sliding mode control and fuzzy based MRAC are proposed and the dynamic responses of vector controlled induction motor with the proposed controllers are compared. The basic concept of Mathematical Modeling of the induction motor is in Section 2. Field oriented control of induction is in section 3. The basic concept of sliding mode control and MRAC and a brief

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description of controllers are in Section 4. In section 4, the fuzzy based controller is proposed. Also the adaptive fuzzy based sliding mode control and fuzzy based MRAC are proposed. Simulation results are shown in Section 5. Finally, the paper is concluded in Section 6.

II. INDUCTION MOTOR MODELING

An Induction Motor of uniform air gap, with sinusoidal distribution of mmf is considered and the dynamic model [2] of the induction motor is derived by transforming the three phase quantitiesinto two phase direct and quadrature axes quantities. The equivalence between the three-phase and two-phase machine models [1] is derived from the concept of power invariance: the power must be equal in the three phase machine and its equivalent two-phase model. The d and q axes mmfs are found by resolving the mmfs of the three phases along the d and q axes. The mathematical model of the 3-phase IM could be represented by an equivalent 2-phase, where d^s, q^s, d^r *and* q^r correspond to the stator, rotor, direct and quadrature axes, respectively. The stator voltage equations formulated from stationary reference frame and the rotor voltage equations formulated to the rotating frame fixed to the rotor.The 3-phase stator and rotor voltage equations written in vector-matrix can be further transformed into 2-phase stator and rotor voltage equations using the well-known Park's transformation. The 3-phase stationary reference frame variables *as-bs-cs* are transformed into 2-phase stationary reference frame variables (d^s–q^s). Furthermore, these 2-phase variables are transformed into synchronously rotating reference frame variables (d^e–q^e) and vice-versa.

The stator circuit equations can be modeled as follows:

$$v^s_{qs} = R_s i^s_{qs} + \frac{d}{dt}\psi^s_{qs} \qquad \text{-------- (2.1)}$$

$$v^s_{qs} = R_s i^s_{ds} + \frac{d}{dt}\psi^s_{qs} \qquad \text{-------- (2.2)}$$

Equations (2.1) and (2.2) are further converted into d^e– q^e frame. The flux linkage expressions in terms of the currents can be written as

$$\psi_{qs} = L_{ls} i_{qs} + L_m (i_{qs} + i_{qr}) \qquad \text{-------- (2.3)}$$

$$\psi_{qr} = L_{lr} i_{qr} + L_m (i_{qs} + i_{qr}) \qquad \text{-------- (2.4)}$$

$$\psi_{qm} = L_m (i_{qs} + i_{qr}) \qquad \text{-------- (2.5)}$$

$$\psi_{ds} = L_{ls} i_{ds} + L_m (i_{ds} + i_{dr}) \qquad \text{------- (2.6)}$$

$$\psi_{dr} = L_{lr} i_{dr} + L_m (i_{ds} + i_{dr}) \qquad \text{-------- (2.7)}$$

$$\psi_{dm} = L_m (i_{ds} + i_{dr}) \qquad \text{------- (2.8)}$$

Using above equations in voltage equations, the electrical transient model of the IM in terms of *v* and *i* is given in matrix form. The development of torque is also very important in the modeling of IMs. The speed ωr cannot be treated as a constant and is related to the torques as

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$$T_e = T_L + j\frac{d}{dt}\omega_m = T_L + \frac{2}{P}J\frac{d\omega_r}{dt} \quad \text{------- (2.9)}$$

where T_L is the load torque, J is the rotor inertia and ωm is the mechanical speed of the IM. Resolving the variables into *de–qe* components, we obtain

$$T_e = \frac{3}{2}\left(\frac{P}{2}\right)\left(\Psi_{dr} i_{qr} - \Psi_{qr} i_{dr}\right) \quad \text{------- (2.10)}$$

The dynamic machine model in stationary frame can be derived simply by substituting $\omega e = 0$. The corresponding stationary frame equations are given as follows:

$$v^s{}_{qs} R_s i^s{}_{qs} + \frac{d}{dt}\Psi^s{}_{qs} \quad \text{------- (2.11)}$$

$$v^s{}_{ds} R_s i^s{}_{ds} + \frac{d}{dt}\Psi^s{}_{ds} \quad \text{------- (2.12)}$$

$$0 = R_r i^s{}_{qr} + \frac{d}{dt}\Psi^s{}_{qr} - \omega_r \Psi^s{}_{dr} \quad \text{------- (2.13)}$$

$$0 = R_r i^s{}_{dr} + \frac{d}{dt}\Psi^s{}_{dr} - \omega_r \Psi^s{}_{qr} \quad \text{------- (2.14)}$$

The torque Equations can also be written with the corresponding variables in the stationary frame as follows:

$$T_e = \frac{3}{2}\left(\frac{P}{2}\right)\left(\Psi^s{}_{dr} i^s{}_{qr} - \Psi^s{}_{qr} i^s{}_{dr}\right) \quad \text{------- (2.15)}$$

The equations (1) and (15) form the mathematical model equations of a three phase induction motor.

III. VECTOR CONTROL OR FIELD ORIENTED CONTROL (FOC)

The Vector Control or Field Oriented Control is used to control Induction motor like a dc motor. Using vector control strategy, the torque and flux components can be controlled independently like dc motor. The basic principles of vector control can be explained with the help of dynamic model of induction motor where we need to convert 3Φ quantities into 2-axes system by 3Φ/2Φ transformation called d-q machine model.There are two methods of vector control, Direct Vector Control method & Indirect Vector Control (IFOC) method. In indirect vector control strategy rotor flux vector is estimated using the field oriented control equations requiring a rotor speed measurement. Due its implementation simplicity, Indirect Vector Control method is more popular than Direct Vector Control in industrial applications.

3.1 INDIRECT FIELD ORIENTED CONTROL (IFOC)

In the Indirect Vector Control method, by using summation of rotor speed and slip frequency, the rotor flux angle is calculated. Hence the unit vectors are obtained indirectly. Then the d-q axis currents are obtained from the torque and flux producing components of stator current.

$$\theta_e = \int \omega_e dt = \int(\omega_r + \omega_{st})dt = \theta_r + \theta_{st} \quad \text{----- (3.1)}$$

The rotor circuit equations

$$\frac{d\psi_{dr}}{dt} + \frac{R_r}{L_r}\psi_{dr} - \frac{L_m}{L_r}R_r i_{ds} - \omega_{sl}\psi_{qr} = 0 \quad \text{----- (3.2)}$$

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$$\frac{d\psi_{qr}}{dt}+\frac{R_r}{L_r}\psi_{qr}-\frac{L_m}{L_r}R_r i_{qs}-\omega_{sl}\psi_{dr}=0 \quad \text{----- (3.3)}$$

For decoupling control $\Psi qr = 0$,So the total flux Ψr directs on the d^e axis. Now from equations 3.1 and 3.2, we get

$$\frac{L_r}{R_r}\frac{d\psi_{dr}}{dt}+\psi_r=\frac{L_m}{L_r}i_{ds} \quad \text{---- (3.4)}$$

As well , the slip frequency can be calculated as

$$\omega_{sl}=\frac{L_m}{\psi_r}\frac{R_r}{L_r}i_{qs}=\frac{R_r}{L_r}\frac{i_{qs}}{i_{ds}} \quad \text{---- (3.5)}$$

The slip gain is

$$K_s=\frac{\omega^{\bullet}_{sl}}{i^{\bullet}_{qs}}=\frac{L_m R_r}{L_r\psi_r} \quad \text{---- (3.6)}$$

It is found that the ideal decoupling can be achieved if the above slip angular command is used for making field orientation. The constant flux Ψr and $\Psi r^{I} = 0$can be substituted in equation 3.4 , so that rotor flux sets as

$$\psi_r=L_m i_{ds} \quad \text{---- (3.7)}$$

The electromechanical torque developed is given by

$$T_e=\frac{3}{2}\frac{P}{2}\frac{L_m}{L_r}\psi_r i_{qs} \quad \text{---- (3.8)}$$

IV. CONTROLLERS DESIGN

Speed controller is necessary to control the speed of the induction motor drive. Design of this speed controller greatly affects the performance of the electric drive. PI controllers are the most commonly used speed controllers before the introduction of fuzzy controller. Design and tuning of the fuzzy based controllers are defined in this section.

4.1 FUZZY LOGIC CONTROLLER

Fuzzy Logic implementation requires no exact knowledge of a model. The block diagram of a FLC is shown in Fig. 4.

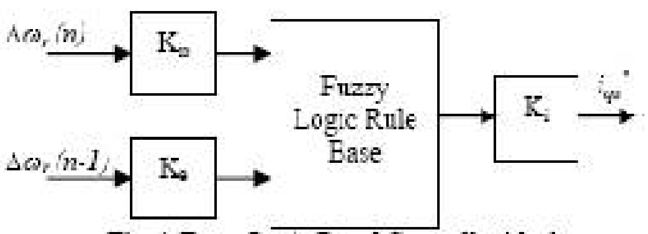

Fig. 4. Fuzzy Logic Based Controller block

It involves the use of the concept of fuzzy subset and rule based modeling. By permitting certain amount of imprecision, complex solutions are modeled with ease.

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4.2 PROPOSED ADAPTIVE FUZZY CONTROL SCHEMES

4.2.1 Fuzzy based SLIDINGMODE CONTROL BASIC CONCEPT

The basic principle of the sliding mode control consists in moving the state trajectory of the system toward a surface $S(X) = 0$ and maintaining it around this surface with the switching logic function Un . The basic sliding mode control law is expressed as.

$$Uc = Ueq + Un \qquad \text{----- (4.1)}$$

This expression uses two terms, Ueq and Un . Ueq : is determined off line with a model that represents the plant asaccurately as possible. It is used when the system state is in the sliding mode. The term Un : is a sign function defined as $Un= k$ sgn($S(X)$), where

$$sgn(S(X)) = \begin{cases} 1, & if \quad S(x) < 0 \\ -1, & if \quad S(x) > 0 \end{cases} \qquad \text{----- (4.2)}$$

This will guarantee that the state is attracted to the switching surface by satisfying the Lyapunov stability criteria .

$$S(x)\dot{S}(x) < 0 \qquad \text{----- (4.3)}$$

This strategy enforces the system trajectory to move toward and to stay on the sliding surface from any initial condition.Using a sign function often causes chattering in practice. One solution to reduce chattering is to introduce a boundary layer around the sliding surface [5], [6]. This is expressed by:

$$Un = \begin{cases} \frac{k}{\varepsilon} S(x), & if|S(x)| < \varepsilon \\ ksign(S(x)), & if|S(x)| > \varepsilon \end{cases} \qquad \text{------ (4.4)}$$

with k , a positive coefficient and ε, the thickness of the boundary layer. However, a small value of S might produce a boundary layer so thin that it can excite high frequency dynamics.

4.2.1.1 IM SLIDING MODE CONTROL

The ‘d’ axis, has the stator current component (Ids) loop and the ‘q’ axis allows the control stator current component (Iqs), whereas the external loop provide the regulation of the speed.

A. Speed SMC

Under field oriented assumptions, the electromagnetic torque can be expressed as:

$$Te = \frac{P\, lm}{lm+ir}(Iqs)\phi^{*}_{r} = k_T\ I\, qs \qquad \text{------- (4.5)}$$

Basically, the control law for Tm_ is divided into two parts: equivalent control Um9 which defines the control action when the system is on the sliding mode and switching part U: which ensures the existence condition of the sliding mode. If the friction k_f is neglected expressions for Ueq and Us can be written as:

$$\begin{cases} Ueq = ke(t) \\ Us = -\beta sign(s(t)) \end{cases} \qquad \text{------- (4.6)}$$

To guarantee the existence of the switching surface consider a Lyapunov function [6, 9]:

$$V(t) = \frac{1}{2} S^2(t) \qquad \text{------- (4.7)}$$

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Based on Lyapunov theory, if the function VG_t_ is negative definite, this will ensure that the system trajectory will be driven and attracted toward the sliding surface *s(t)* and once reached, it will remain sliding on it until the origin is reached asymptotically.

$$S(t)\dot{S} = S(t)\{-\beta sign(s(t)) - d(t)\} \leq 0 \text{ ------ (4.8)}$$

To ensure that above function will be always negative definite, the value of the hitting control gain β should be designed as the upper bound of the lumped uncertainties d(t), i.e.

$$\beta \geq |d(t)| \text{ -------- (4.9)}$$

Therefore the speed control law defined previously will guarantee the existence of the switching surface s(t) and when the error function e(t) reaches the sliding surface, the system dynamics will be governed by equation which is always stable . Moreover, the control system will be nsensitive to the uncertainties Oa, Ob and the load disturbance Tw. The use of the sign function in the sliding mode control will cause high frequency chattering due to the discontinuous control action which represents a severe problem when the system state is close to the sliding surface. To overcome this problem an approach which combines FL with SM is used. The saturation function is replaced by a fuzzy inference system to smooth the control action. The membership functions for the input and output of the FL controller are obtained by trial error to ensure optimal performance.

B. Current SMCs

$$S(Iqs) = (Iqs^* - Iqs)$$

$$S(Ids) = (Ids^* - Ids)$$

The control law development for each variable in sliding mode theory is deduced from the reaching condition () and is indicated below

The current regulators laws in the ‘d’ axis and ‘*q*’ axis can be written as

Current Sliding Mode Control Law of Iqs

$$S(Iqs).\dot{S}(Iqs) < 0 \Rightarrow Vqs^S = Vqs\_eq + Vqs\_n \text{ ------- (4.10)}$$

$$Vqs_{eq} = RsIqs + IsIqs + \omega_s^*(IsIds + \phi^*_r) \text{ ------- (4.11)}$$

$$Vqs\_n = \begin{cases} \frac{kq}{\varepsilon q} S(Iqs), & if \ |S(Iqs)| < \varepsilon q \\ kqsign(S(Iqs)) & if \ |S(Iqs)| > \varepsilon q \end{cases} \text{ ------ (4.12)}$$

Current Sliding Mode Control Law of Ids

$$S(Ids).\dot{S}(Ids) < 0 \Rightarrow Vds^S = Vds\_eq + Vds\_n \text{ ------- (4.13)}$$

$$Vds_{eq} = RsIds + IsIds + \omega_s^*(IsIqs + T_r\,\phi^*_r\,\omega_{sl}^*) \text{ ------- (4.14)}$$

$$Vds\_n = \begin{cases} \frac{kd}{\varepsilon d} S(Ids), & if \ |S(Ids)| < \varepsilon d \\ kqsign(S(Ids)) & if \ |S(Ids)| > \varepsilon d \end{cases} \text{ ------ (4.15)}$$

To verify the system stability condition, the gains kd, kq, and εd, εq should be taken positive by selecting the appropriate values. This sliding mode functions introduce some undesirable chattering. Hence, we will substitute it by the fuzzy logic function. In order to reduce the chattering, two current FSMCs are added to FSMC of the speed outer

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loop. These controllers are used under the same rules IF…THEN, max-min inference mechanism and center of gravity defuzzyfier. The FSMCs are chosen as follows

$$\begin{cases} Vds^f = Vds\_eq + Vds\_f \\ Vqs^f = Vqs\_eq + Vqs\_f \end{cases} \quad \text{------- (4.16)}$$

Vds_f and Vqs_f are calculated with the fuzzy sliding rules described up.

4.2.2 FUZZY LOGIC BASED MODEL REFERENCE ADAPTIVE CONTROL

4.2.2.1 Fuzzy Logic Based Model Reference Adaptive Control with Slip Gain Tuner for IVCIM Drive:

The MRAC method based on reactive power and stator d-axis voltage are combined together with a weighting factor which is generated by a fuzzy controller. The weighting factor ensures the dominant use of reactive power method in low speed high torque region whereas the d – axis voltage method is dominant in high speed low torque region (Gilberto C.D. Sousa et al.1993). A second fuzzy controller tunes the slip gain based on combined detuning error and its slope so as to ensure fast convergence at any operating point on torque-speed plane. The rule base matrix for the fuzzy logic controller generating detuning factor (Kf) is given in Table 2: It clearly shows that if speed is low (L) and torque is high (H) then weighting factor is high(H).

The reference model output is compared with that of adaptive model and the resulting error generates the estimated slip gain through a fuzzy controller. The objective is to provide an adaptive feedback control for fast convergence at any operating point, irrespective of the strength of error signal E and its derivative signal.
From the d^e-q^e model of IM, the stator equations are

$$Vqs = Rsiqs + \frac{d}{dt}(\psi qs) + \omega_e\, \psi ds \quad \text{------ (4.1.1)}$$
$$Vds = Rsids + \frac{d}{dt}(\psi ds) - \omega_e\, \psi qs \quad \text{------ (4.1.2)}$$

At steady state condition under vector control,

$$\frac{d}{dt}(\psi qs) = 0 \quad \text{------ (4.1.3)}$$
$$\frac{d}{dt}(\psi ds) = 0 \quad \text{------ (4.1.4)}$$

$$\Psi ds = Ls\ ids \quad \text{------ (4.1.5)}$$
$$\Psi ds = Ls\ ids\text{-}\frac{Lm}{Lr}\, iqs\ Lm = \left(Ls - \frac{Lm^2}{Lr}\right) iqs \quad \text{------ (4.1.6)}$$

$$Vqs = \omega_e\, Ls\ ids \quad \text{------ (4.1.7)}$$
$$Vds = -\ \omega_e \left(Ls - \frac{Lm^2}{Lr}\right) iqs \quad \text{------ (4.1.8)}$$

$$Q* = Vqs\ ids - Vds\ iqs \quad \text{(reference)} \quad \text{------ (4.1.9)}$$
$$Q = Vqs^s\ ids^s - Vds^s\ iqs^s \quad \text{(actual)} \quad \text{------ (4.1.10)}$$

$$Vds = Vqs^s\ sin\theta_e - Vds^s\ cos\theta_e \quad \text{(actual)} \quad \text{------ (4.1.11)}$$

$$Vds^* = Rs\ ids^* - \widehat{\omega_e}\ L_\sigma\ iqs^* \quad \text{(reference)} \quad \text{------ (4.1.12)}$$

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where cosθe and sinθe are the unit vector components. The loop errors are divided by the respective scaling factor to derive the per unit variable ΔQ and the ΔVds for manipulation by fuzzy controller. Fuzzy controller generates the corrective incremental slip gain ΔKs based on the combined detuning error E and its derivative CE as shown in figures 4.1 and 4.2. Membership function for output variable is shown in figure 4.3.

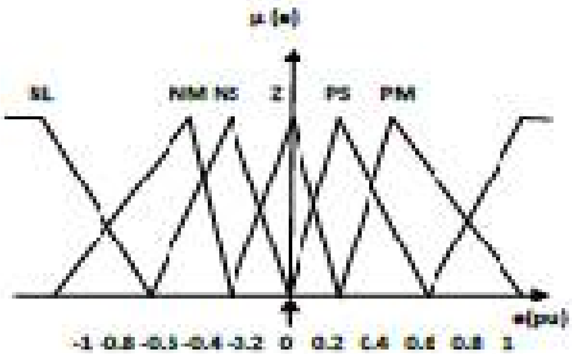

Fig 4.1 Membership function for error

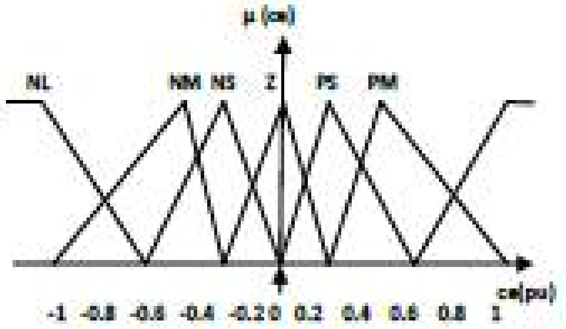

Fig 4.2 Membership function for change in error

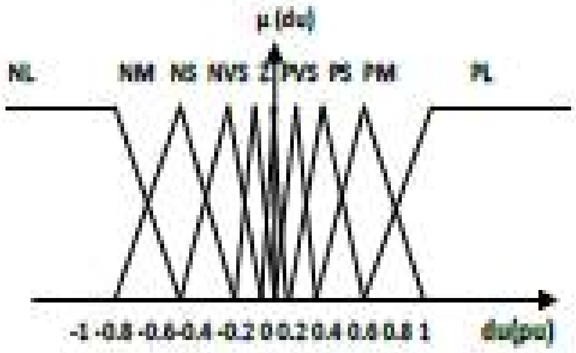

Fig 4.3 Membership function for output

CE/E	NL	NM	NS	Z	PS	PM	PL
PL	Z	PS	PM	PL	PL	PL	PL
PM	NS	Z	PS	PM	PL	PL	PL
PS	NM	NS	Z	PS	PS	PL	PL
Z	NL	NM	NS	Z	PM	PM	PL
NS	NL	NL	NM	NS	Z	PS	PM
NM	NL	NL	NL	NM	NS	Z	PS
NL	NL	NL	NL	NL	NM	NS	Z

Table 1. Rule base matrix for fuzzy controller

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$I_{qs}\backslash\omega e$	H	L
H	M	H
L	L	M

Table 2. Rule base matrix for weighting factor (k_f)

V. SIMULATION RESULTS

The performance of indirect vector control induction motor drive has been simulated in MATLAB environment using simulink.

5.1 Simulation Model of Indirect Vector Control

The model for indirect vector control induction motor drive is shown in the Figures below. The induction motor output results with fuzzy based MRAC controller and Fuzzy based sliding mode controller are obtained using simulation and are analyzed in Table.3. The results are shown in below figures5.2 and 5.4

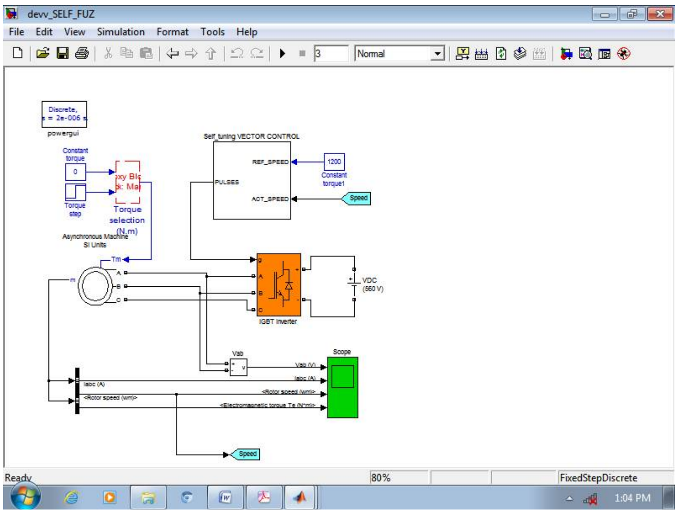

Fig 5.1. Simulation Model for Indirect Vector Control with Fuzzy based MRAC controller

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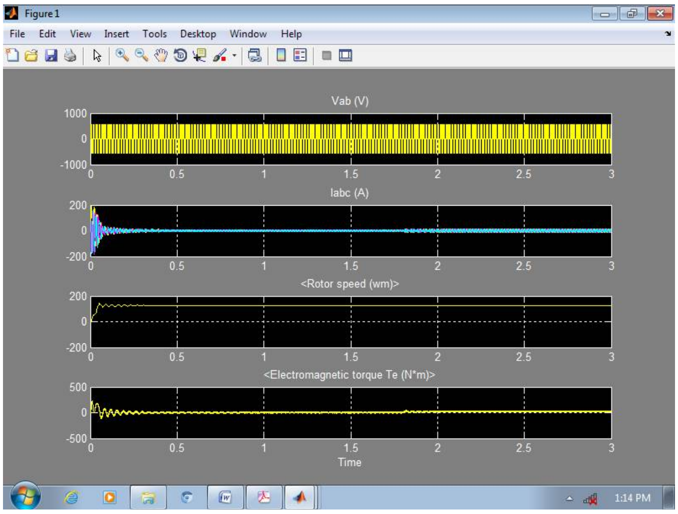

Fig 5.2 Simulation results with Fuzzy based MRAC controller

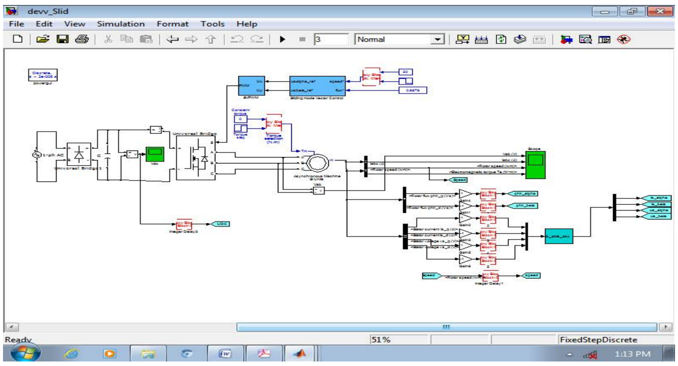

Fig 5. 3. Simulation Model for Indirect Vector Control with Fuzzy based SMC controller

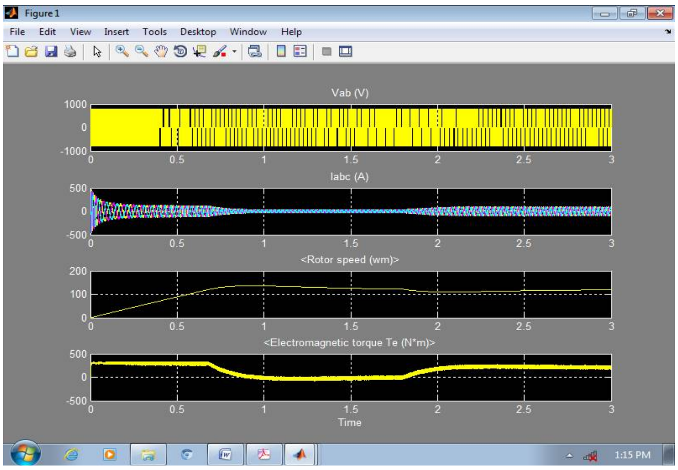

Fig 5.4 Simulation results with Fuzzy based SMC controller

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Controller	Rise Time (sec)	Settling Time (sec)	Peak overshoot (%)
Fuzzy based SMC	0.07	0.17	4.3
Fuzzy based MRAC	0.05	0.15	3.8

Table.3 Summary of Results

VI. CONCLUSION

In this paper, the fuzzy based MRAC and fuzzy based sliding mode control of vector controlled induction motor drive are proposed and the performances are analyzed. From simulation results it was shown that the proposed the fuzzy based MRAC Controller is robust to external variations and has given satisfactory performances in speed response with no overshoot, rapid time response error and a good tracking reference speed. The decoupling between the stator flux and the torque (speed) is maintained with regard to the application of external load disturbance. The fuzzy based MRAC has shown superior performance than that of fuzzy based sliding mode control. The results obtained from simulation shows that the fuzzy based MRAC Controller has increased dynamic response and superior performance.

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