Simulation Analysis of a Fuzzy Logic Based Rotor Resistance Estimator of an Indirect Vector Controlled Induction Motor Drive

Y. Miloud and A. Draou

***Abstract*—A simple method for estimating rotor resistance in an indirect vector-controlled induction motor drive is proposed in this paper. The proposed rotor resistance estimator which uses fuzzy logic principle is based on the improvement of control performance when mismatch between estimated rotor resistance and actual one occurs. Analysis, design, and digital simulations are carried out to demonstrate the effectiveness of the proposed estimator. Moreover, simplicity and direct approach for estimating the rotor resistance are other significant features for the described method.**

***Index Terms*—IFOC, rotor resistance estimator, field orientation.**

NOMENCLATURE

R_s, R_r	Stator and rotor resistances
L_s, L_r	Stator and rotor inductance
$\sigma = 1 - L_m^2/(L_s L_r)$	Leakage inductance
T_r	Rotor time constant
ψ_r	Rotor flux component
P	Number of pole pairs
T_e	Electromagnetic torque
θ_e	Stator electrical angle
ω_e	Electrical synchronous speed
ω_r	Electrical rotor speed
Ω_r, Ω_r^*	Mechanical rotor and reference speed
ω_{sl}	Slip speed
e	Speed error
v_{ds}, v_{qs}	d- and q- axis stator voltages
$i_{ds}, i_{qs}, i_{dr}, i_{qr}$	d- and q- axis stator and rotor currents in the stationary frame

I. INTRODUCTION

WITH THE RAPID the rapid development of microelectronics and power switches,most adjustable-speed drives are now realized with ac machines [1]. Vector control allows high-performance control of speed and torque to be achieved from an induction motor. In typical indirect vector controlled induction motor drive, the flux, torque and slip commands are calculated from the motor variables (voltage, current and speed) based on a model. The control performance is thus sensitive to the system parameters in particular the rotor resistance which changes significantly with temperature and often causes field orientation detuning and degrades system performance. Therefore, an important requirement to obtain good performance is to guarantee the coincidence between the parameters of model and those of the actual motor. Various methods are proposed to estimate the rotor resistance, as it is not possible to measure its value directly in a squirrel-cage motor. Therefore, an important requirement to obtain good performance is to guarantee the coincidence between the parameters of model and those of the actual motor. Various methods are proposed to estimate the rotor resistance, as it is not possible to measure its value directly in a squirrel-cage motor.

The online estimation method for static AC drives proposed by Garces in [2] uses a special adaptation function called the 'Modified Reactive Power Function' to avoid the effects of the stator resistance change. In [3], the rotor parameter estimation is proposed by estimating the rotor temperature. This is based on the fact that the heating influences the fundamental frequency component of terminal voltage for a given current input level. Lorenz in [4] developed an algorithm to correct the adverse effects of rotor resistance variations on torque and speed characteristics of motor. In [5], method used the thermal model of the induction motor to estimate at each operating point the values of stator and rotor resistances. In [6], rotor time constant measurement schemes for indirect vector control drives are proposed. The method proposed was conducted with the motor at standstill and used to adjust the slip gain for a self-tuning field-oriented controller. The measurements are made at specified periods. The induced stator voltage is measured at every zero crossing of the phase currents and the time constant is updated for every measurement.

In [7], the system is directly tuned on line for the rotor resistance variation for Direct Self-Control (DSC). DSC works for stator frequencies of up to a few Hz and can be adjusted to work at zero frequency stationary condition and locked rotor. In [8], rotor time constant measurement is proposed for an indirect vector controlled drive run by a voltage source inverter (VSI). The rotor time constant is measured from the induced stator voltage, which is measured directly using the special switching technique of the CRPWM (VSI). Similar to [6], the measurements are made at specific periods. In [9], a method for estimation of the rotor parameters in an IFOC system is proposed. In this case, the non-linear programming algorithm reconstructs

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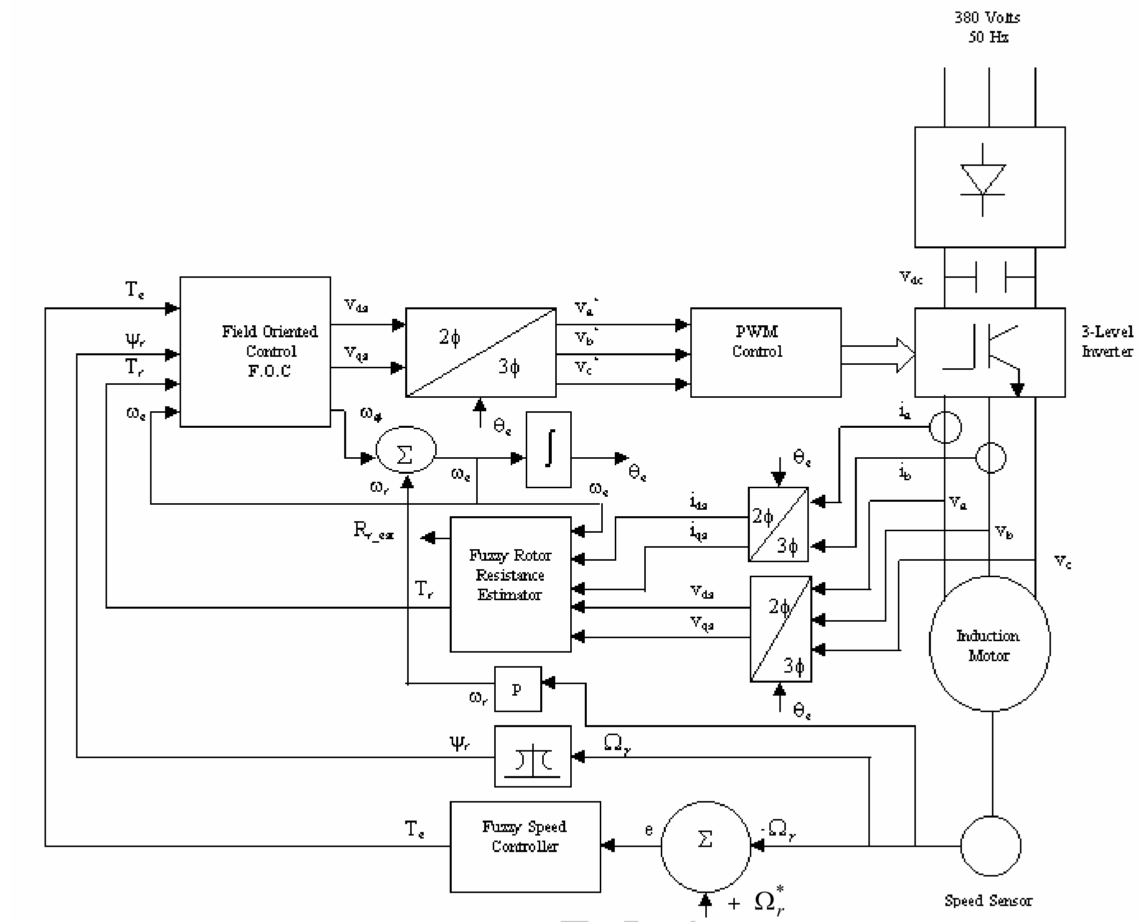


Fig. 1. Schematic diagram of the proposed control strategy with rotor resistance identification.

the non-measurement currents using reinitialized partial moments. The electrical parameters are then estimated from the stator current model in batches. In [10], a method for the identification of rotor time constant is proposed. The identification method is an optimization problem in which the objective function is the total square error between the motor and the commanded stator currents. Despite all these efforts, rotor resistance estimation remains a difficult problem.

This paper presents a method of estimation of the rotor resistance identification based on the reactive power using fuzzy logic controllers. One fuzzy controller is used to regulate the speed, and another is to correct detuning of field orientation. The estimator uses available system signal: stator phase currents and stator phase voltages. The control system has been designed and the overall system has been simulated using MATLAB/SIMULINK software. Furthermore, the machine and control equations are derived, effects of the rotor resistance variations in the fuzzy controllers are presented. Simulation results show the high performance of the method used by using a simple structure of F.O.C. block with d -q axis voltage as outputs.

II. Dynamic Model of the nduction Machine

The model of the squirrel-cage induction machine can be expressed in terms of d- and q-axes quantities resulting in the following equations

$$\dot{X} = AX + BU \tag{1}$$

where definitions are given in (2)-(5). The electromagnetic torque and the mechanical equations can be written as

$$T_e = \frac{3}{2} pL_m (i_{dr} i_{qs} - i_{qr} i_{ds}) \tag{2}$$

$$J\frac{d\Omega_r}{dt} + f\Omega_r = T_e - T_L \tag{3}$$

where J is the moment of inertia, f the viscous friction coefficient and T_L the load torque. A simulation model of the induction machine has been built using top and bottom equations.

$$A = \begin{bmatrix} -\frac{1}{\sigma T_s} & \omega_e + \frac{1-\sigma}{\sigma}\omega_r & \frac{L_m}{\sigma L_s T_r} & \frac{L_m}{\sigma L_s}\omega_r \\ -\omega_e - \frac{1-\sigma}{\sigma}\omega_r & -\frac{1}{\sigma T_s} & -\frac{L_m}{\sigma L_s}\omega_r & \frac{L_m}{\sigma L_s T_r} \\ \frac{L_m}{\sigma L_r T_s} & -\frac{L_m}{\sigma L_r}\omega_r & -\frac{1}{\sigma T_r} & \omega_e - \frac{1}{\sigma}\omega_r \\ \frac{L_m}{\sigma L_r}\omega_r & \frac{L_m}{\sigma L_r T_s} & -\omega_e + \frac{1}{\sigma}\omega_r & -\frac{1}{\sigma T_r} \end{bmatrix} \tag{4}$$

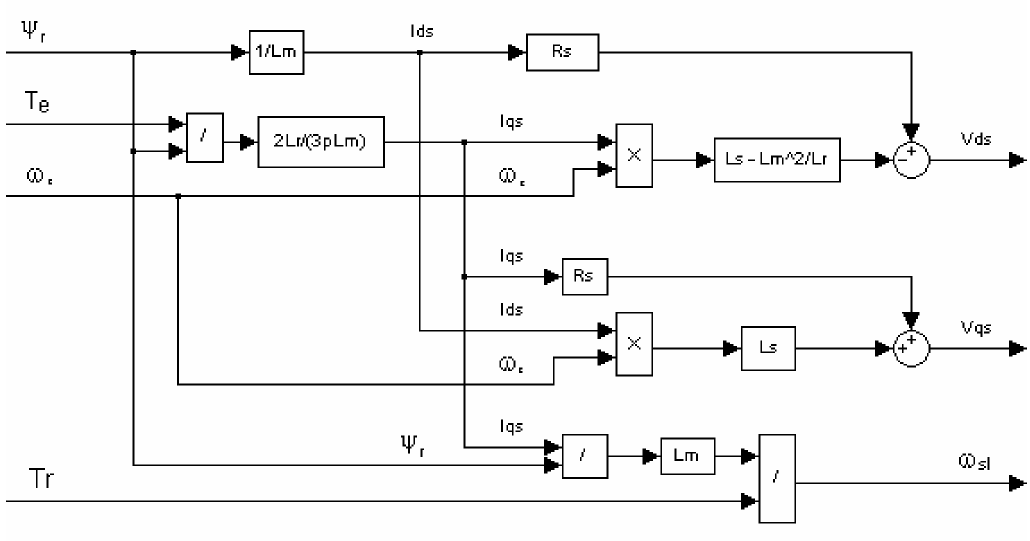


Fig. 2. Structure of F.O.C block.

$$B = \frac{1}{\sigma L_s}\begin{bmatrix} 1 & 0 \\ 0 & 1 \\ -\frac{L_m}{L_r} & 0 \\ 0 & -\frac{L_m}{L_r} \end{bmatrix},\ X = \begin{bmatrix} i_{ds} \\ i_{qs} \\ i_{dr} \\ i_{qr} \end{bmatrix},\ U = \begin{bmatrix} V_{ds} \\ V_{qs} \end{bmatrix} \tag{5}$$

$$\sigma = 1 - \frac{L_m^2}{L_s L_r},\ T_r = \frac{L_r}{R_r},\ T_s = \frac{L_s}{R_s} \tag{6}$$

III. Description of the Approach Used for the Rotor Resistance Identification

The system, which is presented in Fig. 1, is an indirect field oriented control (IFOC)-based induction motor drive. It consists mainly of a squirrel- cage induction motor, a voltage-regulated pulse width modulated inverter, fuzzy speed controller and fuzzy rotor resistance estimator. The induction motor is a three phase, Y connected, four pole, 1.5 kW, 1420 rpm, 220/380 V, 50 Hz, and 6.4/3.7 A.

Under field orientation condition, the d-q equations of the motor in the synchronous reference frame are

$$R_r i_{qr} + \omega_{sl}\psi_{dr} = 0 \tag{7}$$

$$R_r i_{dr} + \frac{d}{dt}\psi_{dr} = 0 \tag{8}$$

$$L_m i_{qs} + L_r i_{qr} = 0 \tag{9}$$

$$L_m i_{ds} + L_r i_{dr} = \psi_{dr} \tag{10}$$

$$v_{ds} = R_s i_{ds} - \sigma L_s \omega_e i_{qs} + \sigma L_s \frac{di_{ds}}{dt} + \frac{L_m}{L_r}\frac{d\psi_{dr}}{dt} - \frac{L_m}{L_r}\omega_e \psi_{qr} \tag{11}$$

$$v_{qs} = R_s i_{qs} + \sigma L_s \omega_e i_{ds} + \sigma L_r \frac{di_{qs}}{dt} + \frac{L_m}{L_r}\frac{d\psi_{qr}}{dt} + \frac{L_m}{L_r}\omega_e \psi_{dr} \tag{12}$$

where R_s, R_r, L_r, L_s, and L_m are motor parameters, $i_{dr}, i_{qr}, i_{ds}, i_{qs}, \psi_{dr}$, and ψ_{ds} are motor currents and fluxes, and ω_{sl} is slip frequency. The equations describing the motor operation in decoupling mode are deduced from (2) and (7)-(10)

$$\omega_{sl} = \frac{L_m}{\psi_r}\left(\frac{R_r}{L_r}\right) i_{qs} \tag{13}$$

$$T_e = \frac{3}{2} p \frac{L_m}{L_r}\psi_r i_{qs} \tag{14}$$

$$\left(\frac{L_r}{R_r}\right)\frac{d\psi_r}{dt} + \psi_r = L_m i_{ds} \tag{15}$$

and in the steady state mode and from (15),

$$\psi_r = L_m i_{ds} \tag{16}$$

then (11) and (12) become

$$v_{ds} = R_s i_{ds} - \sigma L_s \omega_e i_{qs} \tag{17}$$

$$v_{qs} = R_s i_{qs} + L_s \omega_e i_{ds}\ . \tag{18}$$

From (13), (14), (16), (17), and (18), the structure of F.O.C. block can be obtained as shown in Fig. 2. The outputs of the block of F.O.C. are d-q axis voltages and slip frequency (ω_{sl}).

Because of the variation of R_r and L_r, the desired field orientation condition can not always be maintained and the drive performance can be significantly affected. For the normal operation of the drive and without considering the effects derived from the saturation (L_r constant), this rotor resistance can change up to 200% over operation.

In order to adapt this parameter without unwanted measurements inside of the machine, a function was sought which could give the necessary information of the state of the flux (magnitude and position). This function was found in a modified expression of the reactive power. It can be computed by using the stator currents and voltages, and it does not depend on the stator resistance R_s. For the deduction of this function, both stator voltage (11) and (12) is used. However, the reactive power of the machine in the steady state is given by

$$Q_r = v_{ds} i_{qs} - v_{qs} i_{ds}\ . \tag{19}$$

By substituting (11) and (12), in (19), the reactive power becomes

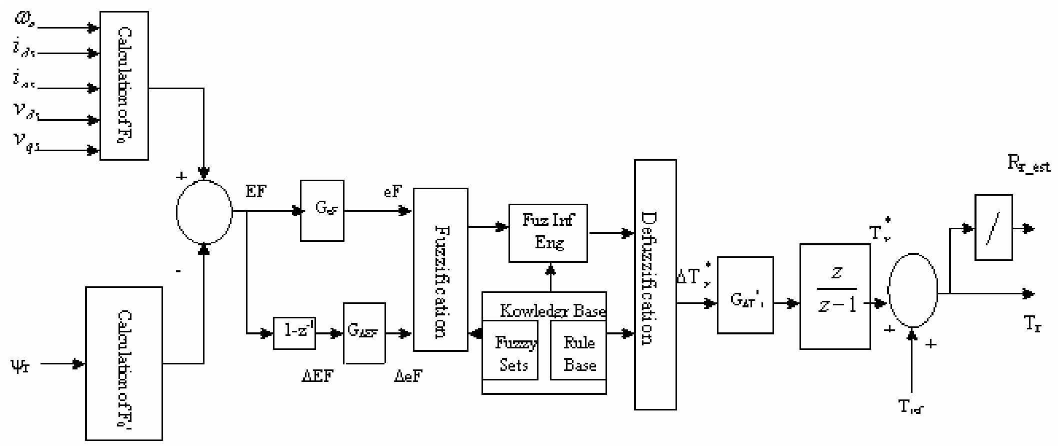
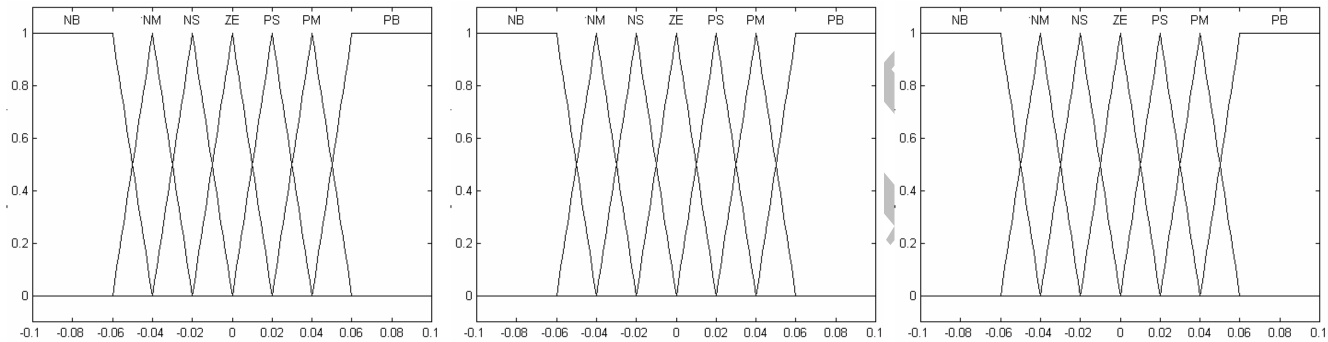


Fig. 3. Fuzzy controller block diagram for rotor resistance estimation.



(a) (b) (c)

Fig. 4. Input and output membership functions, (a) eF , (b) ΔeF , and (c) ΔT_r .

$$
\begin{aligned}
Q_r = {} & \sigma L_s \frac{di_{ds}}{dt} i_{qs} - \sigma L_s \frac{di_{qs}}{dt} i_{ds} - \sigma L_s \omega_s (i^2{}_{ds} + i^2{}_{qs}) \\
& - \frac{L_m}{L_r} \frac{d\psi_{qr}}{dt} i_{ds} + \frac{L_m}{L_r} \frac{d\psi_{dr}}{dt} i_{qs} \\
& - \frac{L_m}{L_r} \omega_e (\psi_{qr} i_{qs} + \psi_{dr} i_{ds})
\end{aligned} \tag{20}
$$

By equalizing (19) and (20), a characteristic function can be defined as

$$
\begin{aligned}
F = {} & \frac{1}{\omega\,\omega_e}\left[\left(v_{ds} - \sigma L_s \frac{di_{ds}}{dt}\right) i_{qs} - \left(v_{qs} - \sigma L_s \frac{di_{qs}}{dt}\right) i_{ds}\right] \\
& + \sigma L_s \left(i_{ds}^2 + i_{qs}^2\right)
\end{aligned} \tag{21}
$$

or as a function of the terms containing the rotor flux

$$
F = \frac{L_m}{L_r} \frac{d\psi_{dr}}{dt} i_{qs} - \frac{L_m}{L_r} \frac{d\psi_{qr}}{dt} i_{ds} - \frac{L_m}{L_r} \omega_e (\psi_{dr} i_{ds} + \psi_{qr} i_{qs}) \tag{22}
$$

where ω_e is the electrical synchronous speed. By introducing field orientation conditions $(\psi_{qr} = 0, \psi_{dr} = \psi_r)$, F is written as:

$$
F = \frac{L_m}{L_r}\left(\frac{d\psi_r}{dt} i_{qs} - \omega_e \psi_r i_{ds}\right). \tag{23}
$$

In the steady state $\left(\frac{d\psi_r}{dt} = 0\right)$, this equation becomes

$$
F_0 = -\frac{L_m}{L_r} \omega_e \psi_r i_{ds} = -\frac{1}{L_r} \omega_e (\psi_r)^2 . \tag{24}
$$

The error function ($EF = F - F_0$) as will be shown later by simulation, reflects the rotor resistance variation [2] and can be used as a correction function for the adaptation of the rotor time constant $T_r = L_r / R_r$ in the fuzzy controller shown in Fig. 3.

IV. ROTOR RESISTANCE ESTIMATOR USING FUZZY LOGIC

Fig. 3 shows the configuration of the proposed fuzzy logic rotor resistance estimation. The functions F and F_0 are first calculated respectively from the measured variables $i_{ds}, i_{qs}, v_{ds}, v_{qs}, \omega_e$ and the reference value Ψ^* .

The error EF and its time variation ΔEF are then calculated as

$$
EF(k) = F(k) - F_0(k). \tag{25}
$$

$$
\Delta EF(k) = EF(k) - EF(k-1). \tag{26}
$$

These variables are used as inputs for the FLC. The internal structure of the fuzzy logic rotor resistance estimation is chosen similar to that of a fuzzy logic controller, which consists of fuzzification, inference engine and defuzzification. The $eF(k)$ and $\Delta eF(k)$ fuzzification stage input signals are derived from the actual $EF(k)$ and $\Delta EF(k)$ signals by dividing with the respective gain factors G_{eF} and $G_{\Delta eF}$. For the successful design of FLC's proper selection of these gains are crucial jobs, which in many cases are done through trial and error to achieve the best possible control performance. Then the crisp variables are converted into fuzzy variable eF and ΔeF using triangular membership functions, Fig. 4. These input membership functions are used to transfer crisp inputs into fuzzy sets. These variables are used as inputs for the FLC. The internal structure of the fuzzy logic rotor resistance estimation is chosen similar to that of a fuzzy logic controller, which consists of fuzzification, inference engine

TABLE I
RULE BASE FOR ROTOR RESISTANCE ESTIMATION

$eF(k)$ / $\Delta eF(k)$	NB	NM	NS	ZE	PS	PM	PB
NB	NB	NM	NM	NS	NS	NS	ZE
NM	NM	NM	NS	NS	NS	ZE	PS
NS	NM	NM	NS	NS	ZE	PS	PM
ZE	NB	NM	NS	ZE	PS	PM	PM
PS	NS	NS	ZE	PS	PS	PM	PM
PM	NS	ZE	PS	PS	PS	PM	PM
PB	ZE	PS	PS	PM	PM	PB	PB

TABLE II
RULE BASE FOR SPEED CONTROL

e_1 / e_2	NB	NM	NS	ZE	PS	PM	PB
NB	NB	NB	NB	NM	NS	NS	ZE
NM	NB	NM	NM	NM	NS	ZE	PS
NS	NB	NM	NS	NS	ZE	PS	PM
ZE	NB	NM	NS	ZE	PS	PM	PB
PS	NM	NS	ZE	PS	PS	PM	PB
PM	NS	ZE	PS	PM	PM	PM	PB
PB	ZE	PS	PS	PM	PB	PB	PB

and defuzzification. The $eF(k)$ and $\Delta eF(k)$ fuzzification stage input signals are derived from the actual $EF(k)$ and $\Delta EF(k)$ signals by dividing with the respective gain factors G_{eF} and $G_{\Delta eF}$. For the successful design of FLC's proper selection of these gains are crucial jobs, which in many cases are done through trial and error to achieve the best possible control performance. Then the crisp variables are converted into fuzzy variable eF and ΔeF using triangular membership functions, Fig. 4. These input membership functions are used to transfer crisp inputs into fuzzy sets.

The expert's experience is incorporated into a knowledge base with 49 rules (7x7). This experience is synthesized by the choice of the input-output (I/O) membership functions and the rule base. Then, in the second stage of the FLC, the inference engine, based on the input fuzzy variables eF and ΔeF, uses appropriate IF-THEN rules in the knowledge base to imply the final output fuzzy sets as shown in the Table I, where NB, NM, NS, ZE, PS, PM, PB correspond to Negative Big, Negative Medium, Negative Small, Zero, Positive Small, Positive Medium, Positive Big respectively.

In the defuzzification stage, the implied fuzzy set is transformed to a crisp output by the center of gravity [11] defuzzification technique as given by the formula (27) where z_i is the numerical output at the ith number of rules and $\mu(z_i)$ corresponds to the value of fuzzy membership function at the ith number of rules. The summation is from one to n, where n is the number of rules that apply for the given fuzzy inputs.

$$z = \frac{\sum_{i=1}^{n} z_i \cdot \mu(z_i)}{\sum_{i=1}^{n} \mu(z_i)} . \tag{27}$$

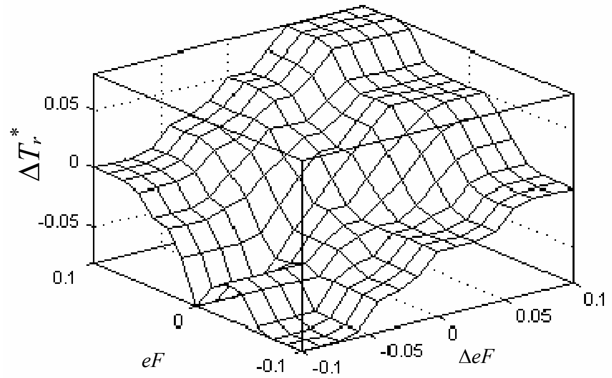


Fig. 5. Crisp input/output map, fuzzy controller nonlinear input/output map for rotor resistance estimation.

The crisp output ΔT_r^* is multiplied by the gain factor $G_{\Delta T_r^*}$ and then integrated to give

$$T_r^*(k) = T_r^*(k-1) + G_{\Delta T_r^*} \Delta T_r^*(k) . \tag{28}$$

This value added to the reference rotor time constant (T_{ref}) gives the estimated time constant (T_r) which is used as an input to the F.O.C block of Fig. 1 to ensure the correct field orientation operation of the drive. Without considering the effects derived from the saturation (L_r constant), $R_{r-\mathrm{est}}$ is obtained from the estimated rotor time constant Fig. 3. Therefore this rotor resistance estimation value used in the control model must match its real value in order to maintain a high performance of the induction motor drive as will be shown later.

The input/output mapping of the FLC rotor resistance estimation is shown in Fig. 5, which is a continuous highly non-linear function. Detailed discussion about FLC construction is referred in [12] The same type of membership functions used in fuzzy logic rotor resistance estimation are applied in fuzzy sets for speed fuzzy logic controller. The inputs are e_1 and e_2 as defined in (29) and (30), where G_1 and G_2 are adjustable input gains.

$$e_1 = G_1\left(\Omega_r^*(k) - \Omega_r(k)\right) \tag{29}$$

$$e_2 = G_2\left(e_1(k) - e_1(k-1)\right) \tag{30}$$

A knowledge base of 7×7 rules, as shown in Table II, is applied to tune T_e to reduce the speed error to zero. The final output of speed fuzzy logic controller is expressed in

$$T_e(k) = T_e(k-1) + G_{T_e^*} . T_e^*(k) . \tag{31}$$

These fuzzy rules can be understood easily and can be explained intuitively. For example, IF error of speed is negative big ($e_1 = \mathrm{NB}$) and change of error is negative small ($e_2 = \mathrm{NS}$) then it is quite natural that the fuzzified torque command should be negative big ($\Delta T_e^* = \mathrm{NB}$). The other rules can be understood in a similar way [12].

V. SIMULATION RESULTS

The configuration of the overall control system is shown in Fig. 1. It is essential that the simulation model is designed to approach as close to reality as possible. Therefore, for the simulation of the whole drive system according to Fig. 1, a mathematical model has been developed based on the induction motor equations and the equations for estimating the rotor resistance, which have been derived in Section III. In addition, a mathematical

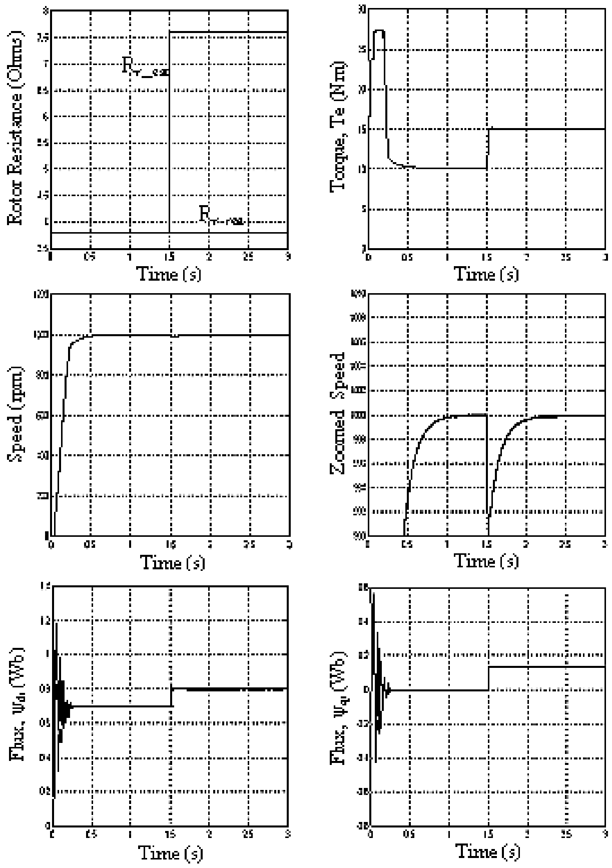


Fig. 6. Simulation waveforms illustrating the effects of the rotor resistance variation on the IM drive performance.

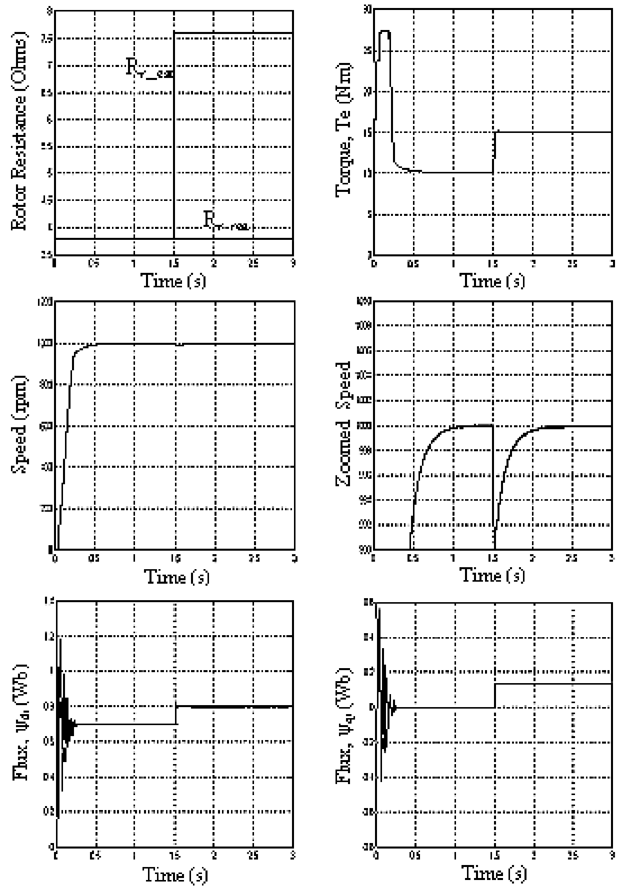


Fig. 7. Effect of the rotor resistance variation with fuzzy estimator for uncompensated ramp change of R_r .

model for all the remaining drive system units was necessary to complete the simulation model.

In order to analyze the drive system performance for their flux and torque responses, with rotor resistance variation, the above-presented system has been simulated using MATLAB/SIMULINK software. A squirrel cage induction motor with a rated power of 1.5 kW has been used. The specifications and parameters of the induction motor are listed in Appendix constant reference flux of 0.695 Wb is assumed and the speed was held constant at 1000 rpm. The rotor resistance was stepped or ramped from 100% to 200% of its rated value, thereby simulating a change in rotor resistance due to a temperature change. The system was first started up to 1000 rpm with a full load of 10 N.m. at 1.5 s, the rotor resistance was stepped from 100% to 200% of its rated value. The responses of the uncompensated step case are shown in Fig. 6.

It is observed in this figure that when the estimated rotor resistance has changed from its rated value, a detuning problem has occurred and the command torque (T_e) is no longer 10 N.m but rather has increased to 15 N.m to compensate the drop in speed which equals approximately 9.5 rpm. But in the actual operating conditions, the rate of change of temperature is very slow and so the resistance variation. Fig. 7 covers this situation, where at 1.5 s a ramp change of rotor resistance for an uncompensated case is applied linearly from 100% of its rated value to 200% till 4.5 s, after that, this value is maintained constant for 2.5 s. We notice from this figure, that the torque command deviates from the motor torque and therefore the quadratic rotor fluxes and the error $F - F_0$ are no longer zero. The direct rotor flux has increased from its rated value giving a detuning problem. However, the performance of the control system is affected when the rotor resistance value used in the control algorithm does not match properly the real value.

Therefore in order to maintain a high performance of the induction motor drive, it is required that the rotor resistance value used in the control model should be updated regularly to track its real value. In this case, the field orientation condition can be maintained which is illustrated in Fig. 8 by applying a ramp change of rotor resistance for a compensated case. In this figure, the detuned problem is removed completely and ψ_{qr} stabilizes to almost zero and ψ_{dr} to its rated value 0.695 wb. The drop in speed is negligible and excellent tracking of rotor resistance is obtained. The error $F - F_0$ is equal to zero and command torque matches perfectly the motor torque at steady state.

The initial oscillations immediately after zero instant time in rotor d-axis flux can be minimized by increasing response time in speed to 2 s rather than 0.5 s.

However, by comparing the results for uncompensated and compensated cases of a ramp change variation of rotor resistance, one can say that the association of rotor resistance estimator using fuzzy logic controller provides excellent dynamic performance to an induction motor drive.

Fig. 9 demonstrates the high performance of the fuzzy logic based rotor resistance estimator when the rated value (R_r) has changed initially to ±20% and at the same time a ramp change of rotor resistance was applied at 1.5 s. In both cases, the rotor resistance tracking is excellent and

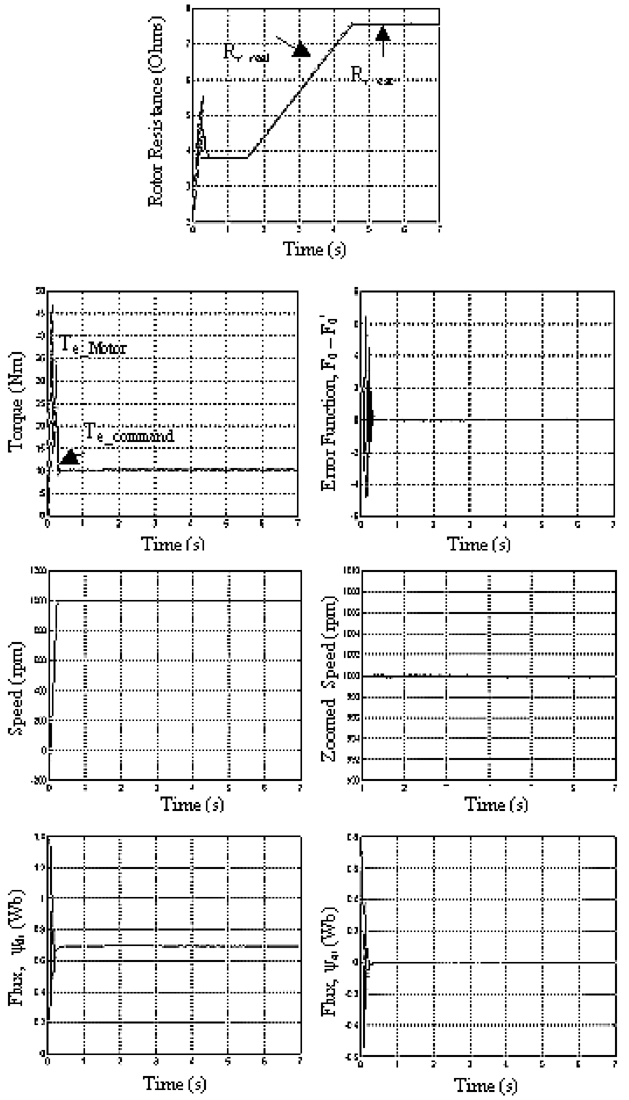
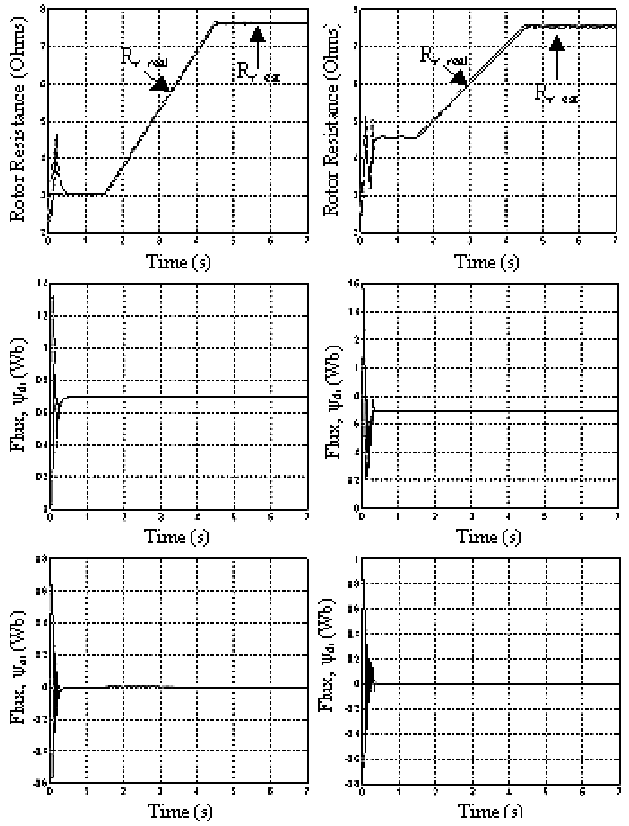


Fig. 8. Effects of rotor resistance variation with Fuzzy estimator for compensated ramp change of R_r .

the field orientation condition is still maintained. However, insensitivity to the drive parameter variations and working conditions can thus be obtained. It is clear that the proposed scheme achieves good performance as it achieves compensation of the rotor resistance changes.

Fig. 10 confirms the robustness of the proposed estimated method at low speed. At starting, few oscillations appear till 1s, and then, in steady state the oscillations disappear completely and the speed stabilizes to its reference value. The tracking of rotor resistance was achieved.

Finally, a comparison between fuzzy estimator for compensated ramp change of R_r and PI rotor resistance variation Figs. 11 and 12., was carried out. The PI controller presents an overshoot at transient time of 86 rpm (8.6%) whereas the proposed controller has no overshoot. When a ramp change of rotor resistance was applied at 1.5 s till 4.5 s, and after that this value was kept constant to 7 s, the fuzzy controller reacts perfectly without almost any change in speed but the PI controller has first a drop of speed of 5 rpm at 1.5 s, and then, an overspeed speed of 2 rpm at 4.5 s. One can see clearly the difference between these two controllers.



(a) (b)

Fig. 9. Performance of the fuzzy logic based rotor resistance estimator when its rated value (R_r) has changed initially to (a) -20% and (b) +20%.

VI. Conclusion

A simple fuzzy logic based rotor resistance estimator of an indirect vector-controlled induction motor drive has been presented. The estimator uses available system variable signals as stator currents and stator voltages.

Digital simulation results show that this estimator can minimize the detuning effects and enhance dramatically the performance of an indirect field oriented induction motor drive. An excellent tracking performance was obtained and hence the drive operation was optimized. At low speed, the proposed estimated method showed its robustness by tracking the real rotor resistance value and the speed at steady state stabilized at its reference value.

Appendix

Induction Motor Parameters

1.5 kW, 1420 rpm	$R_s = 4.85\Omega$	$L_s = 274$ mH
220/380 V, 6.4/3.7 A	$R_r = 3.805\Omega$	$L_r = 274$ mH
3 Phases, 50 Hz, 4 poles	$L_m = 258$ mH	
$J = 0.031$ kg.m^2	$B = 0.00114$ kg.m^2/s	

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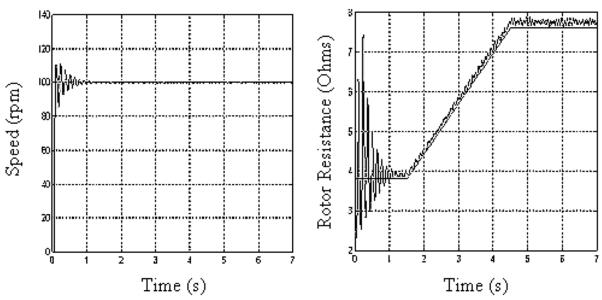

Fig. 10. Proposed estimated method at low speed.

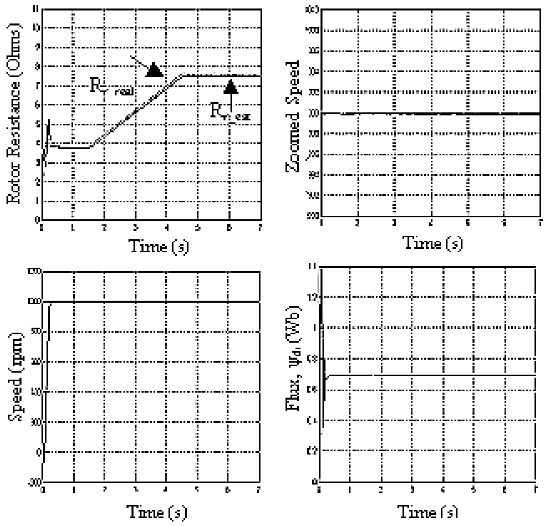

Fig. 11. Effect of rotor resistance variation with Fuzzy estimator for compensated ramp change of R_r .

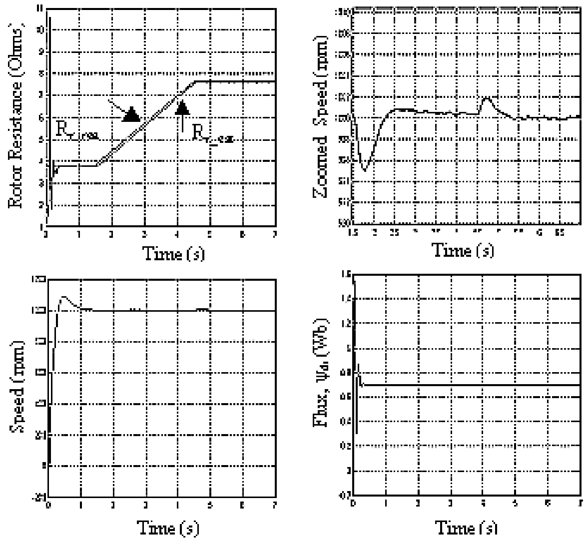

Fig. 12. Effect of rotor resistance variation with PI estimator for compensated ramp change of R_r .

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