ROBUST MRAS-BASED ALGORITHM FOR SIMULTANEOUS ESTIMATION OF ROTOR SPEED AND ROTOR RESISTANCE IN SENSORLESS INDIRECT VECTOR CONTROL OF INDUCTION MOTOR DRIVES

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***Abstract*— This paper presents a method for simultaneous rotor speed and rotor resistance estimation of an induction motor in an indirect rotor field oriented control system. This method is based on model reference adaptive system approach using only stator currents and voltages measurement. Finally the feasibility of the scheme is verified by simulation.**

***Keywords*— Induction motor, Speed sensorless control, Field-oriented control, rotor resistance estimation, MRAS estimator.**

I. INTRODUCTION

The induction motor has found use in a wide range of industrial application, due to its reliability, ruggedness, low maintenance requirement and relatively low cost [1].

On the other hand, ongoing research has concentrated on the elimination of the speed sensor at the machine shaft without deteriorating the dynamic performance of the drive control system. The advantages of speed sensorless induction motor drives are reduced hardware complexity and lower cost, reduces size of the drive machine, elimination of the sensor cable, better noise immunity, increased reliability and less maintenance requirements, in order to achieve good performance of sensorless speed different methods have been proposed, Model Reference Adaptive Systems (MRAS) schemes are the most common strategies employed due to their relative simplicity and low computational effort. Rotor flux MRAS, first proposed by Schauder, is the most popular MRAS strategy and a lot of effort has been focused on improving the performance of this scheme [2].

Indirect field oriented control IFOC on induction motor has been applied in many industrial applications due to high dynamics performance and no brushes and commutators as we have in separately excited dc motors [3], but the motor parameters used in the vector controller particularly to the rotor resistance, it changes widely with the rotor temperature, resulting in various harmful effects such as over (or under) excitation, the destruction of the decoupled condition of the flux and torque, etc. recently, attention has been given to the identification of the instantaneous value of the rotor resistance of sensorless induction motor drive.

To solve the above problems, many paper proposed the estimation methods. In [4] the rotor resistance identification based on the reactive power using fuzzy logic controller has been proposed. A model reference adaptive has been used for rotor resistance estimation in [5]. In [6] the rotor resistance estimation of sensorless induction motor drive using the Extended Kalman Filter. In [7] the rotor resistance estimated using artificial neural networks for vector controlled speed sensorless induction motor drive.

Out of these, MRAS is popular due to its simplicity, requirement of lesse computation time and good stability. On line adaptation of the rotor resistance can improve the performance of the MRAS sensorless drive. In this paper a simultaneous estimation of rotor speed and rotor resistance is presented based on MRAS scheme its performances are tested by simulation.

This paper is organized as follows. Section 2 shows the dynamic model of induction motor and principle of field-oriented controller; in section 3 the proposed system for speed estimation is presented. In section 4 MRAS rotor resistance estimator configuration is given. In section 5, the performances of the proposed sensorless control of induction motor with rotor resistance estimation are illustrated by simulation results. Finally section 6 draws the final conclusions.

II. DYNAMIC MODEL OF INDUCTION MOTOR

By referring to a rotating reference frame, denoted by the superscript (d,q), the dynamic model of a three–phase induction motor can be expressed as follows [8]:

$$
\begin{cases}
\frac{d}{dt} i_{ds} = -A_1 . i_{sd} + \omega_s i_{sq} + \frac{L_m}{\sigma . L_s . L_r . T_r} . \Phi_{rd} + A_2 \omega_r \Phi_{rq} + A_3 . V_{sd} \\
\frac{d}{dt} i_{qs} = -\omega_s i_{sd} - A_1 . i_{sq} - A_2 \omega_r \Phi_{rd} + \frac{L_m}{\sigma . L_r . L_s . T_r} . \Phi_{rq} + A_3 . V_{sq} \\
\frac{d}{dt} \Phi_{dr} = \frac{L_m}{T_r} i_{sd} - \frac{1}{T_r} \Phi_{rd} + (\omega_s - \omega_r) . \Phi_{rq} \\
\frac{d}{dt} \Phi_{qr} = \frac{L_m}{T_r} . i_{sq} - (\omega_s - \omega_r) . \Phi_{rd} - \frac{1}{T_r} \Phi_{rq} \\
\frac{d\omega_r}{dt} = \frac{p}{J} (T_{em} - T_l) - \frac{f}{J} \omega
\end{cases} \tag{1}
$$

Where

$$
A_1 = \left(\frac{R_s}{\sigma . L_s} + \frac{1-\sigma}{\sigma . T_r} \right); \qquad A_2 = \frac{L_m}{\sigma . L_s . L_r}
$$

$$
A_3 = \frac{1}{\sigma . L_s}; \; \sigma = 1 - \frac{L_m^2}{L_s L_r};
$$

$$
\omega_g = \omega_s - \omega_r
$$

$$
T_{em} = \frac{3}{2} P \frac{L_m}{L_r} (\Phi_{rd} . i_{sq} - \Phi_{rq} . i_{sd})
$$

ω_s and ω_r are the electrical synchronous stator and rotor speed; σ is the linkage coefficient, and T_r is the rotor time constants.

III. Rotor Flux Orientation Strategy

There are tow categories of vector control strategy. We are interested in this study to the so-called IFOC. As shows in Eq (1) that the expression of the electromagnetic torque in the dynamic regime presents a coupling between stator current and rotor flux, [9].

The main objective of the vector control of induction motors is, as in DC machines, to independently control the torque and the flux; this is done by using a *d-q* rotating reference frame synchronously with the rotor flux space vector, [6]-[10]. The d-axis is then aligned with the rotor flux space vector (Blaschke, 1972). Under this condition we get:

$$
\varphi_{rd} = \varphi_r \text{ and } \varphi_{rq} = 0
$$

The torque equation becomes analogous to the DC machine and can be described as follows:

$$
T_e = \frac{3}{2} p \frac{L_m}{L_r} (\varphi_r i_{sq}) \tag{2}
$$

It is right to adjust the flux while acting on the stator current component i_{sd} and to adjust the torque while acting on the i_{sq} component.

Using the Eq (1) we get:

$$
_r \; i_{sd} = p \frac{(1 + T_r s)}{L_m} \varphi_r^{*} \tag{3}
$$

$$
i_{sq} = \frac{T_r}{L_m} \omega_{gl}^{*} \varphi_r^{*} \tag{4}
$$

We replace i_{sq} by its expression to obtain T_e as function of the reference slip speed ω_{gl}^{*}

$$
T_e = \frac{3}{2} p \frac{\varphi_r^{*2}}{R_r} \omega_{gl}^{*} \tag{5}
$$

The stator voltage commands are:

$$
\begin{cases}
v_{sd} = R_s i_{sd} - \sigma L_s \omega_s i_{sq} + \sigma L_s \frac{di_{sd}}{dt} + \frac{L_m}{L_r} \frac{d\varphi_r}{dt} \\
\quad = v_{sd1} - \omega_s \sigma . L_S . i_{Sq} \\
v_{sq} = R_s i_{sq} + \sigma L_s \omega_s i_{sd} + \sigma L_s \frac{di_{sq}}{dt} + \frac{L_m}{L_r} \omega_s \varphi_r \\
\quad = v_{sq1} - \omega_s \sigma . L_S . i_{Sd} - \frac{L_m}{L_r} . \omega \varphi_r
\end{cases} \tag{6}
$$

The voltages v_{sd} and v_{sq} should act on the current i_{sd} and i_{sq} separately and consequently the flux and the torque. The two-phase stators current are controlled by two PI controllers taking as input the reference values i^*_{sd} , i^*_{sq} and the measured values. Thus, the common thought is to realize the decoupling by adding the compensation terms ($\tilde{e}_{sd}$ and $\tilde{e}_{sq}$), [11].

The block decoupling is described by the followingg equations:

$$
\begin{aligned}
\tilde{e}_{sd} &= \omega_s \sigma . L_S . i_{Sq} \\
\tilde{e}_{sq} &= -\omega_s \sigma . L_S . i_{Sd} - \frac{L_m}{L_r} . \omega \varphi_r
\end{aligned} \tag{7}
$$

It is necessary to determine the amplitude and the position of rotor flux. In the case of an indirect field oriented control, the module is obtained by a block of field weakening given by the following non linear relation:

$$¤_r^* = \begin{cases} ¤_{rn} & if \ |\mu_r| \le \mu_{rn} \\ ¤_{rn} \dfrac{\mu_{rn}}{|\mu_r|} & if \ |\mu_r| > \mu_{rn} \end{cases} \quad (8)$$

The slip frequency can be calculated from the values of the stator current quadrate and the rotor flux oriented reference frame as follow:

$$\mu_g = \mu_s - \mu_r = \frac{L_m}{T_r} \cdot \frac{i_{Sq}}{¤_{rd}} = \frac{1}{T_r} \frac{i_{sq}}{i_{sd}} \quad (9)$$

The rotor flux position is given by:

$$—_S = \int \mu_S .dt = \int \left(p.\Omega + \frac{L_m .i_{Sq}}{T_R .¤_r^\bullet} \right) .dt \quad (10)$$

3.1 Rotor Speed Regulation

The use of a classical PI controller makes appear in the closed loop transfer function a zero, which can influence the transient of the speed. Therefore, it is more convenient to use the so-called IP controller which has some advantages as a tiny overshoot in its step tracking response, good regulation characteristics compared to the proportional plus integral (PI) controller and a zero steady-state error

$$\frac{\mu_r(s)}{\mu_r^*(s)} = \frac{k_i .k_p .k_t .p}{J.s^2 + (B + k_p .k_t .p).s + k_i .k_p .k_t .p} \quad (11)$$

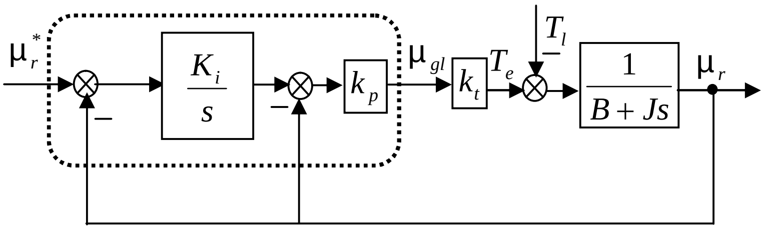

Fig.1. Bloc diagram of IP speed controller

The gains of IP controller, K_p and K_i, are determined using a design method to obtain a trajectory of speed with the desired parameters (¶ and μ_n). The gains parameters values of the IP speed controller are easily obtained as:

$$\begin{cases} K_{p\mu} = \dfrac{(2.¶.\mu_n .J - B) R_r}{P.¤_r^2} \\ K_{i\mu} = \dfrac{J.\mu_n^2}{K_{p\mu} .p^2 .¤_r^2} \end{cases} \quad (12)$$

According to the above analysis, the indirect field oriented control (IFOC), of induction motor with current- regulated with PWM inverter control system can reasonably be presented by the block diagram shown in the Fig. 6.

The two PI current controllers (fig. 6) act to produce the decoupled voltages v_{sd1} and v_{sq1}

The reference voltages v_{sd}^* and v_{sq}^* determined by (6) ensure decoupled two-axes control of the induction motor drive.

IV. Proposed System for Speed Estimation

The model reference adaptive system is one of many promising techniques employed in adaptive control. Among various types of adaptive system configuration, MRAS is important since it leads to relatively easy-to-implement systems with high speed of adaptation for a wide range of applications.
The complete IM drive along the proposed MRAS based speed estimator is shown in figure 2 the parameters of the motor are available in Appendix. A constant reference flux of 0.695Wb is assumed, the speed estimation algorithm is enclosed by dotted line and consists of three main blocs. “Reference

Model”, “Adjustable Model”, and “Adaptation Mechanism”.

The reference model, usually expressed by the voltage model, represents the stator equation. It generates the reference value of the rotor flux components in the stationary reference frame from the monitored stator voltage and current components. The reference rotor flux components obtained from the reference model are given by [2, 12]:

$$p \cdot^{s}_{r\sim} = \frac{L_r}{L_m}(v_{s\sim} - R_s i_{s\sim} - \pm L_s p i_{s\sim})$$
$$p \cdot^{s}_{r} = \frac{L_r}{L_m}(v_{s} - R_s i_{s} - \pm L_s p i_{s}) \qquad (13)$$

The Adjustable model, usually represented by the current model, describes the rotor equation where the rotor flux components are expressed in terms of stator current components and the rotor speed. The rotor flux components obtained from the adaptive model are given by [2]:

$$p \hat{\cdot}^{r}_{r\sim} = \frac{L_m}{T_r} i_{s\sim} - \frac{1}{T_r}\hat{\cdot}_{r\sim} - \hat{\mu}_r \hat{\cdot}_{r}$$
$$p \hat{\cdot}^{r}_{r} = \frac{L_m}{T_r} i_{s} - \frac{1}{T_r}\hat{\cdot}_{r} - \hat{\mu}_r \hat{\cdot}_{r\sim} \qquad (14)$$

Finally, the adaptation scheme generates the value of the estimated speed to be used in such a way as to minimize the error between the reference and estimated fluxes. In view of the overall gradual stability of the system, we make use of the Popov hyper stability theorem. The universal error is [2, 13]:

$$\pounds = \cdot_{r} \hat{\cdot}^{r}_{r\sim} - \cdot^{s}_{r\sim} \hat{\cdot}^{r}_{r}$$
$$\hat{\mu}_r = (K_p + \frac{K_i}{p})\pounds \qquad (15)$$

Where $i_{s\sim}$ and $v_{s\sim}$ are measured values and “^” signifies the estimated value.

IV. MRAS Based Rotor Resistance Estimation for IFOC IM Drive

There are many methods for estimation of the rotor resistance. One group of on-line rotor resistance adaptation methods is based on the principles of MRAS. This is the approach with relatively simple implementation requirements.

In any model reference adaptive system-based rotor resistance estimation, one quantity is formed in two different ways. One of them is independent of rotor resistance and other is dependent on this parameter. The computed two quantities are used to formulate the error signal. The error between the states of the two models is used to drive a suitable adaptation mechanism, which in most cases is a PI controller that generates the estimate resistance.

$$\pounds_{\sim} = \text{¤}_{r\sim} - \hat{\text{¤}}_{r\sim}$$
$$\pounds = \text{¤}_{r} - \hat{\text{¤}}_{r} \qquad (16)$$

$$\hat{R}_r = (K_p + \frac{K_i}{p})\pounds$$

Figure 2 shows the bloc diagram of the estimation technique with adaptive reference model.

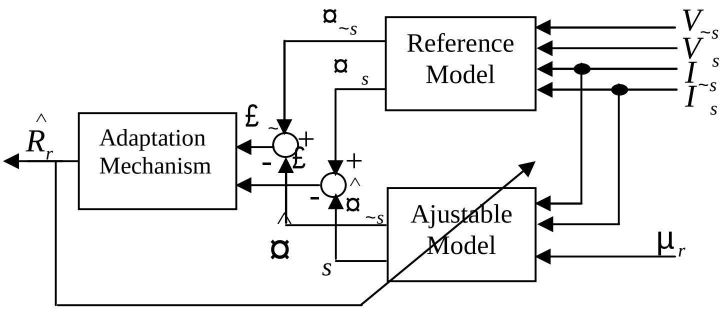

Fig.2 MRAS adaptive Scheme to estimate rotor resistance of induction motor

Both parameters R_r and $\hat{R}_r$ vary with time and each may be seen as an input to the adjustable model. To investigate the dynamic response of the MRAS rotor resistor estimation, it is necessary to linearize the stator and rotor equations for small deviation around a working point. So, the deviations of the error ε are given by:

$$\Delta\pounds = \frac{\text{¤}_{\sim r0}\Delta\text{¤}_{\sim r0} + \text{¤}_{r0}\Delta\text{¤}_{r0}}{\sqrt{\text{¤}^2_{\sim r0} + \text{¤}^2_{r0}}} - \frac{\hat{\text{¤}}_{\sim r0}\Delta\hat{\text{¤}}_{\sim r0} + \hat{\text{¤}}_{r0}\Delta\hat{\text{¤}}_{r0}}{\sqrt{\hat{\text{¤}}^2_{\sim r0} + \hat{\text{¤}}^2_{r0}}} \qquad (17)$$

And the transfer function relating $\Delta £$ to $\Delta\hat{R}_r$ is:

$$\frac{\Delta £}{\Delta \hat{R}_r}\bigg|_{\Delta\hat{R}_r \to \infty} = \frac{(p+¢_0)(|¤_0|^2 - L_s(¤_{\sim r0} i_{\sim r0} + ¤_{\ r0} i_{\ r0}))}{((p+¢_0)^2+µ^2)|¤_0|} \tag{18}$$

Where at steady state we have $¤_0^2 = ¤_{\sim r0}^2 + ¤_{\ r0}^2$ and $R_r = \hat{R}_r$

The closed loop diagram of the dynamic response of MRAS rotor resistance estimation can be drawn as in Fig. 3.

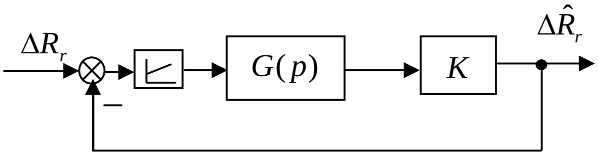

Fig. 3. Closed loop diagram of the dynamic response of MRAS

Where:

$$K = \frac{|¤_0|^2 - L_s(¤_{\sim r0} i_{\sim r0} + ¤_{\ r0} i_{\ r0})}{L_r|¤_0|} \tag{19}$$

And

$$G(p) = \frac{p + \dfrac{1}{T_r}}{(p + \dfrac{1}{T_r})^2 + µ_r^2} \tag{20}$$

The transfer function $G(p)$ allows two complex poles:

$$p_1 = -\frac{1}{T_r} + jµ_r \text{ and } p_2 = -\frac{1}{T_r} - jµ_r$$

The $\dfrac{1}{T_r}$ is always positive, the poles *p1* and *p2* have negative real parts. So *G(p)* stability is confirmed. The PI regulator is justified by the fact that the estimator has to perform with no error at steady state and to converge in a reasonable bandwidth compared to the dynamics speed response.

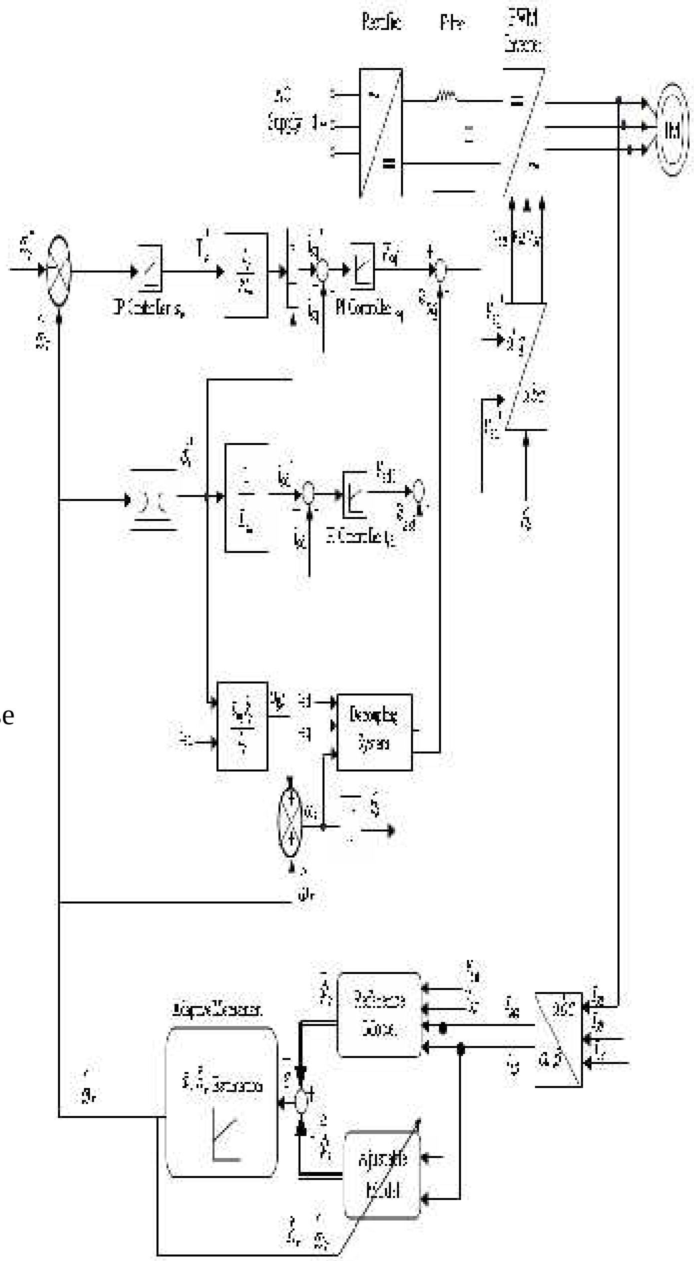

Figure 4. Block Diagram of Simultaneous Estimation of Rotor Speed and Rotor Resistance

6. Simulation Results and Discussion:

In order to verify the effectiveness and feasibility of the proposed method, a digital computer simulation model has been developed in MATLAB/SIMULINK and the results are presented below.

In Figure 5, we see that the estimated rotor speed converge to the actual value, with almost zero static error. The actual and estimated quadratic rotor flux ψqr are stabilises to almost zero.

Also in this figure the actual and estimated ψ_{dr} are stabilises to its rated value 0.695 wb. The obtained result demonstrates that the proposed system gives a good estimate of speed based on MRAS scheme.

In Fig. 6, the figure shows the simulation results of actual and estimated speed for step changing of reference from 100 rad/sec to -100 rad/sec, and the nether one shows the speed error in the corresponding process.

The rotor resistance was ramped to 100% of its rates value at 3s. The responses of the uncompensated ramp case are shown in Figure 7.

We notice from this figure that there is the drop in speed which equals approximately 2.75 (rad/sec), the quadratic rotor flux is no longer zero. 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 Figure.8.

In Figure 9, the detuned problem is removed completely and ψ_{qr} stabilises to almost zero and ψ_{dr} to its rated value 0.695 wb. The fall in speed is negligible and excellent tracking of rotor resistance is obtained.

From the simulation results, the simultaneous speed and rotor resistance estimation has proves the effectiveness of the proposed method based on MRAS scheme.

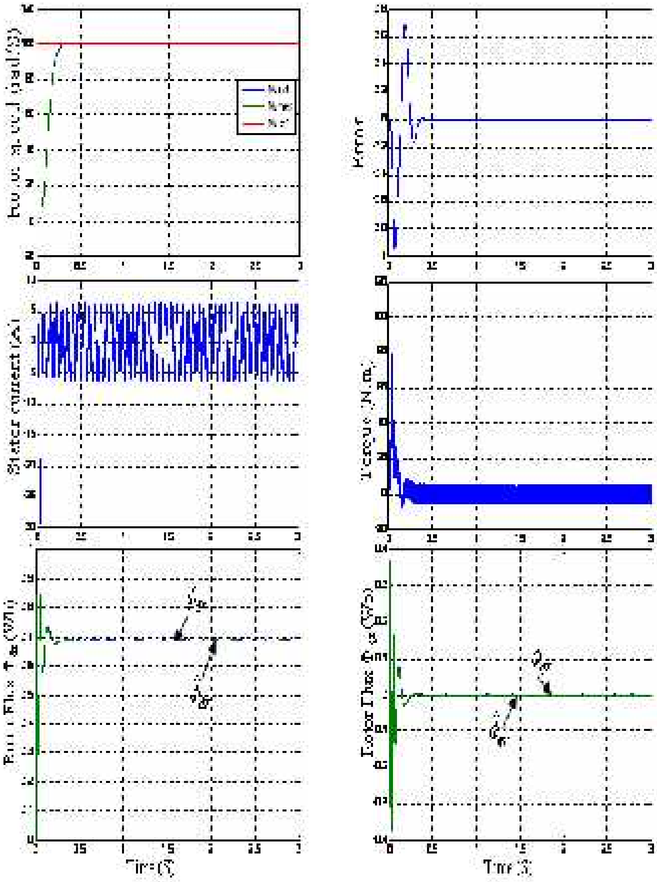

Fig.5: Dynamic performance of the control system

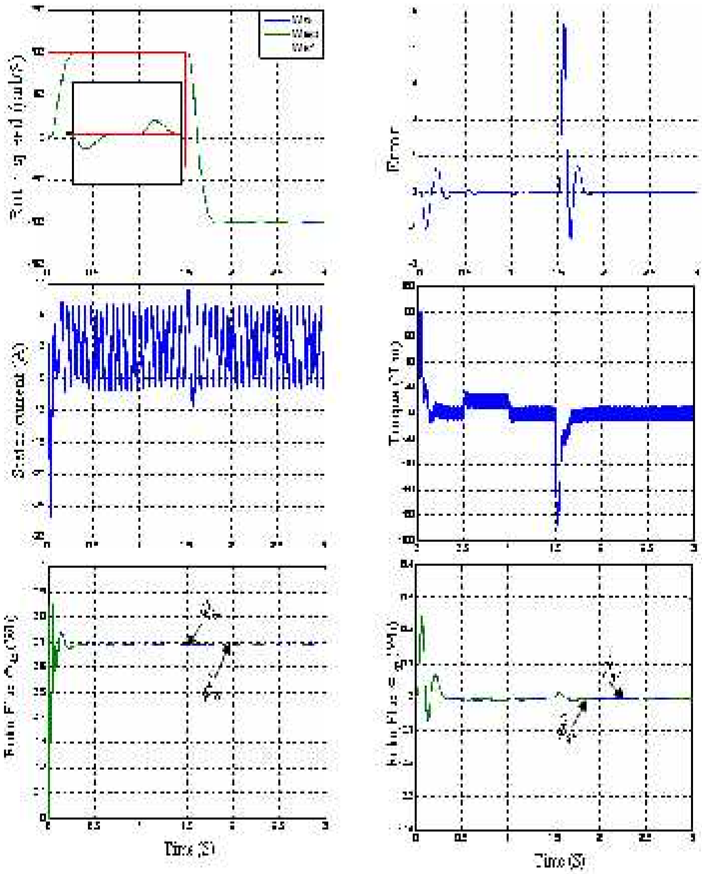

Fig.6. Performance of the proposed sensorless vector controlled induction motor drive with a speed reverse and under load change.

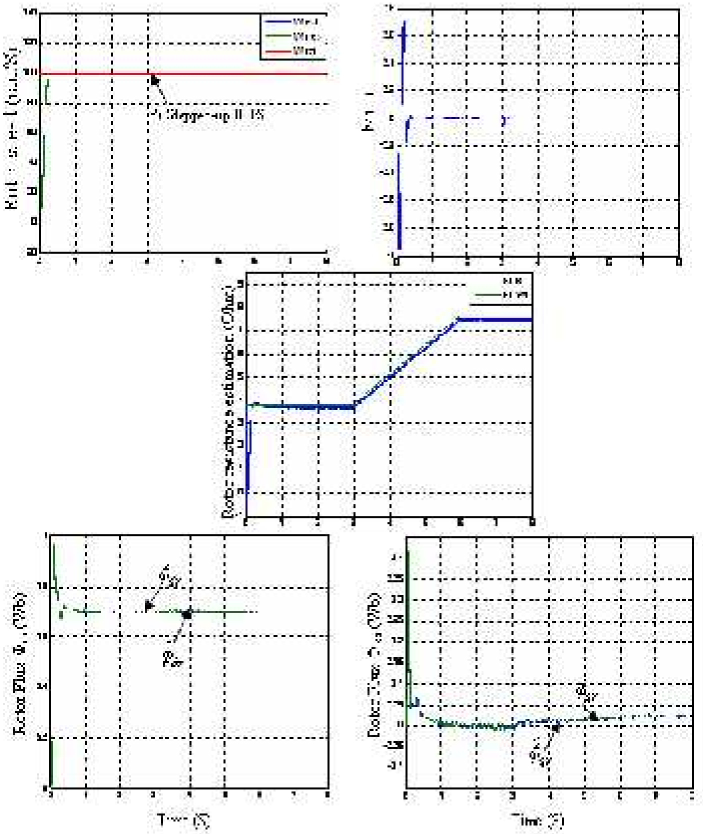

Fig.8: Performance of the MRAS based rotor resistance estimator with Kp = 0.5 and Ki = 4

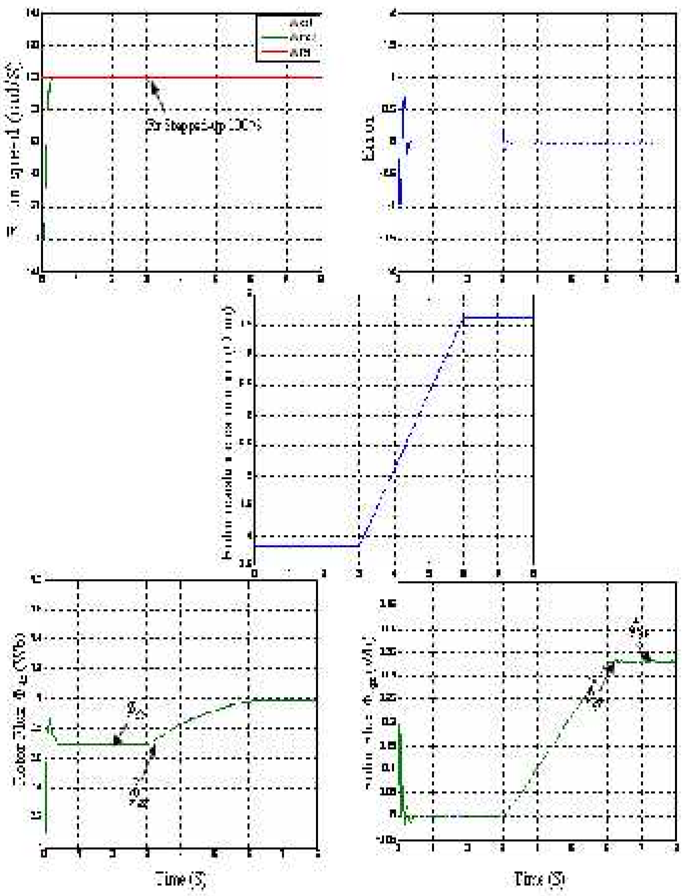

Fig.7. Waveforms illustrating the effects of ramp rotor resistance variation in sensorless IFOC Induction motor drive

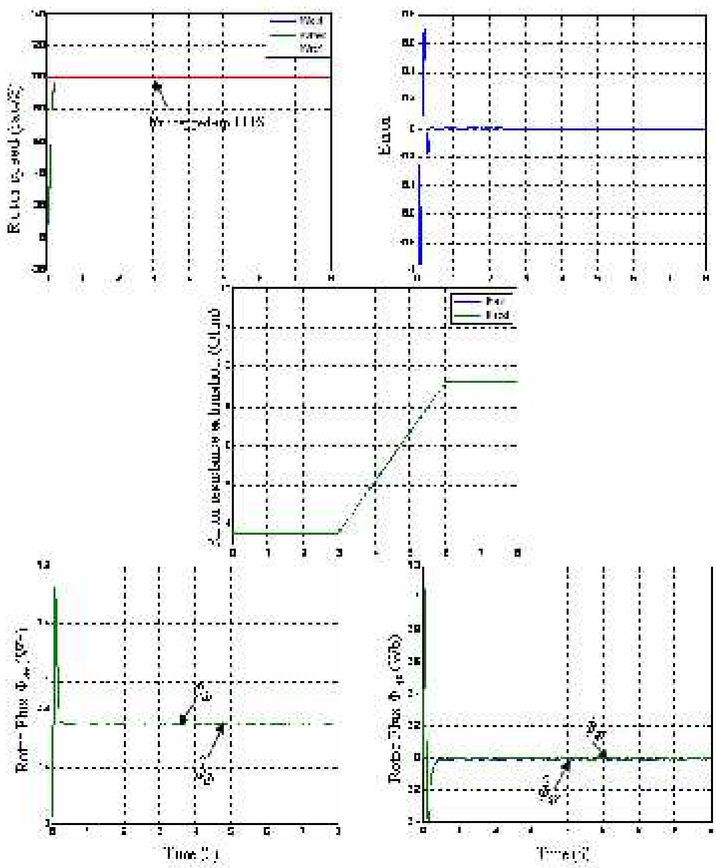

Fig.8. Performance of the MRAS based rotor resistance estimator with Kp = 0,21 and Ki = 3

VI. Conclusion

This paper addresses the problems of rotor resistance and speed estimation in sensorless IFOC induction motor drive based MRAS scheme.
The validity of the proposed method has been verified by simulation. So the influence of the rotor resistance variations on the field oriented control of the induction motor can be removed

Appendix

Induction Motor Parameters

50 Hz, 1.5 Kw , 1420 rpm, 380 V, 3.7A
$R_r = 3.805\Omega$, $R_s = 4.85\Omega$, $L_s = 274$ mH,
$L_s = 274$ mH
$J = 0.031$ kg.m^2, F=0.00114kg.m^2/s

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