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Torque Maximization and Sensorless Control of Induction Motor in a Flux Weakening region

A.BEN ALI(1), A.KHEDHER(2), M.F. MIMOUNI(3)et, R.DHIFAOUI(4)
(1)Institut Supérieur des Etudes Technologiques de Sousse, Cité Riadh,4000 Sousse.
(2)Institut Supérieur des Sciences Appliquées et de Technologie de Sousse, Rue 1er Juin 1955, 4003 Sousse
(3)Ecole Nationale d'Ingénieurs de Monastir, Route de Kairouan, 5000 Monastir.
(4)Institut National des Sciences Appliquées et de Technologie de Tunis, centre Urbain nord-1098 Tunis
Email : adnene792001@yahoo.fr

Abstract: *In this paper, one presents a transient state performance optimization of vector-controlled induction motor (IM) drives operating in the field-weakening region. The vector control scheme, based on the rotor flux field-orientation (RFOC), is considered to ensure maximum torque motor operation regime. Although in the high speed region, the measure of rotor speed and the sensitivity to parameters errors of the motor still remains a problem. In this context, one proposes an approach that ensures sensorless control and maximum torque operation maintaining robustness essentially to rotor resistance variation. Also, the speed sensorless control used for the motor drives able to obtain correct rotor speed through sliding mode rotor flux observer (SMRFO) under RFOC method. The estimation speed algorithm is sensitive to rotor resistance; witch is adapted on line using the Lyapunov stability theory. Simulation results demonstrate the performance of the proposed robust speed sensorless torque maximisation control scheme in response to high-speed demands and rotor resistance variations*

Keyword: *Induction machine, Torque Maximisation, Flux Weakening, Sensorless Control, Sliding Mode Observer.*

1. Introduction

In many applications of induction machine operating above nominal speeds such as spindle and traction drives, control law permits to generate in one hand constant torque at nominal speeds, and a decrease of torque and operation at constant power is realised at high speeds in other hand. In this regime, weakening of the motor flux is made in order to allow operation of variable-speed induction motor drives at high

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speeds, and to perform at transient-state maximum electromagnetic torque. This is desirable by taking into account the constraints linked to the limitations imposed by the maximum ratings currents and voltages magnitudes of the motor and those of inverter [1]. These constraints limit the reference flux and the q-axis current magnitudes for the field orientation scheme. Although, the reference flux in the field-weakening region are traditionally selected to vary with the speed. The slip frequency speed varies directly with speed in RFOC. If the maximum torque is required over a wide range of flux weakening, it is necessary to consider the nonlinearity induced by the magnetization curve of the machine [2].

Field-weakening operation involves the choice of the flux reference to obtain maximum torque and producing the necessary currents in order to converge to the flux and torque references. The conventional RFO technique has been shown to provide flux references that are near to nominal values, reducing the amount of current available to produce torque and, therefore, the maximum torque of the IM [3]. In the optimal current references for maximum torque were obtained by taking into account both current and voltage limits for the inverter and motor. In this case, only steady-state operation was considered and constant dc-bus is used. This paper considers the steady and transient state optimized operation of the induction motor drive below and above nominal speed under RFOC strategy.

Generally, DFOC is used for induction motor vector control technique, to ensure rapidly and accurately control the electromagnetic torque and rotor flux for a given speed. First, the rotor flux for DFOC is estimated via sliding mode rotor flux observer (SMRFO). However, the speed transducers affect generally the cost and reliability of DFOC based induction motor drives. Therefore, in high power, some class of vector controlled induction motor drives does not permit the use of speed transducers. Sensorless vector control techniques for induction motor drives are used in the literature [7] to provide a good estimation of the rotor speed. In sensorless vector control induction motor drives, developed in this paper is based on DFOC. The rotor speed is estimated from the slip angular frequency. However, the slip angular frequency is sensitive to the rotor resistance. This parameter is estimated on line via *Lyapunov* stability theory [8].

The paper is organized as follows. A mathematical model for the induction motor is given in section two. The mathematical model is established by using rotating reference frame equations, and then a steady state model is derived. In section three, the formulation of performance optimization problem is developed with the constraints and deals with torque maximization. The procedure used of the conception of SMRFO, the rotor speed estimation, and the adaptive rotor resistance is described in section four. Finally, simulations results are given in Section five.

2. Modelling of the induction machine

The transient state of the induction machine can be described in terms of space vectors by the following equations written in a rotating synchronous reference frame as:

$$\begin{cases} \bar{v}_s = R_s \bar{i}_s + \dfrac{d\bar{\varphi}_s}{dt} + j\omega_s \bar{\varphi}_s \\ 0 = R_r \bar{i}_r + \dfrac{d\bar{\varphi}_r}{dt} + j\omega_g \bar{\varphi}_r \end{cases} \tag{1}$$

$$\begin{cases} \bar{\varphi}_S = L_S \bar{i}_S + M \bar{i}_r \\ \bar{\varphi}_r = L_r \bar{i}_r + M \bar{i}_S \end{cases} \tag{2}$$

$$T = \frac{3}{2} p \frac{M}{L} \bar{\varphi} \wedge \bar{i} \tag{3}$$

where p is the pole pairs number, ω_S is the angular frequency of the rotor flux vector, ω_g is the sleep speed, and '$\wedge$' denotes the vectorial product.

By substituting the rotor current vector $\bar{i}_r$ with respect to $\bar{i}_S$ and the parameters of the machine, the rotor equation gives:

$$\frac{d\bar{\varphi}_r}{dt} + (-\frac{1}{T_r} + j\omega_g)\bar{\varphi}_r = \frac{M}{T_r}\bar{i}_s \tag{4}$$

In steady-state operation, (1) and (4) become:

$$\begin{cases} \bar{v}_s = R_s \bar{i}_s + j\omega_s \bar{\varphi}_s \\ (-\dfrac{1}{T_r} + j\omega_g)\bar{\varphi}_r = \dfrac{M}{T_r}\bar{i}_s \end{cases} \tag{5}$$

In the RFOC scheme, the rotor flux vector is aligned with the d-axis and it imposes the following condition:

$$\varphi_{rd} = \phi_r,\ \varphi_{rq} = 0 \tag{6}$$

By using condition described by (6), equation (5) gives:

$$\begin{cases} v_{sd} = R_s i_{sd} - \omega_s \sigma L_s i_{sq} \\ v_{sq} = R_s i_{sq} + \omega_s \sigma L_s i_{sd} \\ \phi_r = M i_{sd} \\ \omega_g = \dfrac{i_{sq}}{T_r i_{sd}} \end{cases} \tag{7}$$

The steady state torque equation can be written as:

$$T_{em} = \frac{3}{2} p \frac{M}{L_r} i_{sd} i_{sq} \tag{8}$$

These steady-state equations (7) and (8) will be utilized for the analysis of the maximum torque capability.

In the field-weakening region, the magnetizing inductance M varies easily, due to the variation of the rotor flux. To take into account the detuned effect of M, a magnetizing curve of the induction machine can be obtained from the no-load test, using the measurements of the stator currents and voltages. For the considered IM, the magnetizing inductance M variation with respect to the rotor magnetising current i_{mr} is shown in Fig.1.

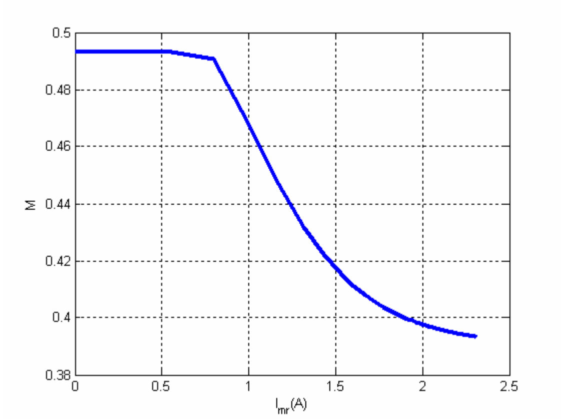

Fig.1 : Variation of the mutual inductance according to the rotor magnetizing current

The curve presented in fig.1 can be approximated by a simple model described the following equation [2]:

$$M = \begin{cases} 0.4928 \quad if \quad i_{mr} \leq 0,76 \\ 0,39 + 0.4933\, e^{-3(i_{mr}-0,76)} - 0.39\, e^{-3,65(i_{mr}-0,76)} \quad if \quad i_{mr} > 0,76 \end{cases} \tag{9}$$

According to [1], the expression of the rotor magnetizing current can be given by:

$$i_{mr} = i_{sd}\sqrt{1+\left(\frac{i_{sq}}{i_{sd}}\left(1-\frac{M}{L_r}\right)\right)^2} \tag{10}$$

Therefore, one introduces in the main algorithm to ensure the maximum torque capability, the model (9) of the mutual inductance curve.

3. Control strategy used for maximum torque capability in the flux weakening region

The maximum value of equation (8) means the maximum torque. Therefore, for the maximum torque operation the value of (8) should be maximized under the voltage and current ratings of the PWM inverter. The maximum voltage that the inverter can apply to the machine is limited by available dc-link voltage and (PWM) strategy. In this paper, a PWM strategy based on voltage space vector modulation is used, and then is limited to $V_{dc}/\sqrt{3}$ [1]. The maximum machine current is limited by the inverter current rating and the machine thermal rating. Therefore the machine voltage and current have the following limits:

$$|V_s|^2 = V_{sd}^2 + V_{sq}^2 \le V_{max}^2 \qquad (11)$$

$$|I_s|^2 = i_{sd}^2 + i_{sq}^2 \le I_{max}^2 \qquad (12)$$

In this part, one supposes that the magnetic circuit of the asynchronous machine is unsaturated what permits to have a constant mutual inductance.

3.1. Maximum Torque Algorithm

Generally, for yielding the maximum torque under the limit conditions of voltage and current, one distinguishes three operative zones [1, 4, 5].

A first zone concerns the regime of the motor with needs a maximal current to produce a maximum torque. The speed being less than the nominal speed, the magnitude of electromotive voltage is always less than the peak value V_{max}, by consequence the current is the only constraint to respect.

$$i_{sd}^{*\,2} + i_{sq}^{*\,2} = I_{max}^2 \qquad (13)$$

By using the steady state relation of the sleep speed ω_g and equation (13), the electromagnetic torque is expressed as:

$$T_{em} = \frac{3}{2} p \frac{M^2}{L_r} I_{max}^2 \frac{\omega_g T_r}{1 + \omega_g^2 T_r^2} \qquad (14)$$

One shows that the maximum torque is gotten for an optimal slip speed $\omega_g = \frac{1}{T_r}$. In this case, the stator current components are given by [11,12]:

$$\begin{cases} i_{ds}^* = \dfrac{I_{max}}{\sqrt{2}} \\ i_{qs}^* = \dfrac{I_{max}}{\sqrt{2}} \end{cases} \qquad (15)$$

The second zone is linked to the increase of the speed; it generates an increase in the electromotive voltage of the machine. In this case, the magnitude of required flux is reduced in an optimal manner to respect the two constraints.

Under the two constraints (11) and (12) and by using equation (7), the maximum value of (8) consists in solving of the following system:

$$\begin{cases} \left(R_s^2+\omega_s^2L_s^2\right)i_{sd}^2+\left(R_s^2+\omega_s^2\sigma^2L_s^2\right)i_{sq}^2+2\omega_s\dfrac{M^2}{L_r}R_si_{sd}i_{sq}=V_{\max}^2 \\ i_{sd}^2+i_{sq}^2=I_{\max}^2 \end{cases} \quad (16)$$

From the expression (16), we deduce $i_{sq}=\sqrt{I_{\max}^2-i_{sd}^2}$, and by replacing the component i_{sq} in the first equation of (16), one can obtain:

$$\left(R_s^2+\omega_s^2L_s^2\right)i_{ds}^2+\left(R_s^2+\omega_s^2\sigma^2L_s^2\right)\left(I_{max}^2-i_{ds}^2\right)+2\omega_s\frac{M^2}{L_r}R_si_{ds}\sqrt{I_{max}^2-I_{ds}^2}=V_{\max}^2 \quad (17)$$

The resolution of the expression (17) makes itself numerically via the Matlab environment.

The third zone permits to the motor to operating beyond the nominal frequency, the electromotive voltage becomes much larger so that the magnitude of the stator current of the machine is lower than the nominal value, and the attainable maximum output torque is determined only by the voltage-limit constraint.

In this zone, the speed of the machine is very important; one obeys to the constraint of stator voltage. Therefore, the direct current command is limited current constraint. The current command is usually limited by the torque-limit constraint (12) because the rotor flux is reduced in the field weakening region. By taking into account the equations (8) and (9), the torque is written as follows:

$$T_{em}=\frac{3}{2}p\frac{M^2}{Lr}\frac{V_{\max}^2}{R_s^2}\frac{\omega_gT_r}{\left(1-\sigma T_s\omega_gTr\omega_s\right)^2+\left(T_r\omega_g+T_s\omega_s\right)^2} \quad (18)$$

The optimal sleep speed ω_g^* is given by:

$$\omega_g^*=\frac{1}{T_r}\sqrt{\frac{1+\left(T_s\omega_s\right)^2}{1+\left(\sigma T_s\omega_s\right)^2}} \quad (19)$$

The correspondent maximum torque is given by:

$$T_{e\max}=\frac{3}{2}p\frac{M^2\omega_g^*}{R_r}i_{ds}^{*\,2} \quad (20)$$

The correspondent's components commands of stator currents are given by:

$$\begin{cases} i_{ds}^*=\sqrt{\dfrac{2}{3}\dfrac{R_rT_{\max}}{p\omega_g^*}} \\ i_{qs}^*=\dfrac{T_r}{M}\omega_g^*i_{ds}^* \end{cases} \quad (21)$$

3.2. Control Algorithm

The induction motor maximum torque control amounts generating in a first phase the reference torque according to the reference speed and the following expression [9,10]:

$$T_{ref} = \frac{J}{P}\frac{\omega_{ref} - \omega_r}{T} + f\,\omega_r \qquad (22)$$

and to calculate the motor maximum torque using the suggested method.

In one second phase one proceeds oneself by comparison, because two cases can to present:

- If $T_{ref} \geq T_{\max}$ one calculation the torque by using the maximization torque proposed algorithm.
- If not the currents are determined by fixing the direct current $i_{sd}^* = \frac{I_{\max}}{\sqrt{2}}$ and while choosing the quadratic current such as: $i_{sq}^* = \dfrac{T_{ref}}{\frac{3P}{2}\frac{M^2}{L_r} i_{sd}^*}$

4. Rotor flux Observer based on the sliding mode approach associated to the rotor resistance estimation

In the RFOC technique, the rotor resistance varies essentially to temperature, possessed a direct impact on the estimation of the rotor flux level and sleep frequency variables [6]. So, in order to overcome the control sensitive to rotor resistance variation, we subject in this paper the use of a SMRFO, associated to an adapted control law to estimate the rotor resistance.

4.1. Formulation of the Proposed Sliding Mode Rotor Flux Observer

The design procedure of the sliding mode observer consists of performing the following two steps. First, design the manifold (the intersection of the sliding mode surface SMS) S, such that the estimation error trajectories restricted to S have the desired stable dynamics. Second, the observer gain is determined to drive the estimation error trajectories to S and maintain it on the set, once intercepted, for all subsequent time

Denoting $\hat{x}_1, \hat{x}_2, \hat{x}_3$ *and* $\hat{x}_4$ as estimates of state variables $i_{s\alpha}, i_{s\beta}, \varphi_{r\alpha}, \varphi_{r\beta}$ respectively, the sliding mode observer of electrical subsystem is designed as follows:

$$
\begin{cases}
\dfrac{d\hat{x}_1}{dt} = -\gamma \hat{x}_1 + \dfrac{k}{Tr}\hat{x}_3 + pk\,\hat{x}_4\,\hat{x}_5 + \dfrac{1}{\sigma Ls}Vsd + A1Is \\
\dfrac{d\hat{x}_2}{dt} = -\gamma \hat{x}_2 + \dfrac{k}{Tr}\hat{x}_4 + -pk\,\hat{x}_3\,\hat{x}_5 + \dfrac{1}{\sigma Ls}Vsq + A2Is \\
\dfrac{d\hat{x}_3}{dt} = \dfrac{M}{Tr}\hat{x}_1 - \dfrac{1}{Tr}\hat{x}_3 - p\,\hat{x}_4\,\hat{x}_5 + A3Is \\
\dfrac{d\hat{x}_4}{dt} = \dfrac{M}{Tr}\hat{x}_2 - \dfrac{1}{Tr}\hat{x}_4 + p\,\hat{x}_3\,\hat{x}_5 + A4Is
\end{cases}
\tag{23}
$$

The vector sliding surface (S), the vector sign (Is) and the gain matrices for the observer are given by:

$$
s = \begin{pmatrix} S_1 \\ S_2 \end{pmatrix} = D^{-1} * \begin{pmatrix} i_{sd} - \hat{i}_{sd} \\ i_{sq} - \hat{i}_{sq} \end{pmatrix},\ Is = \begin{pmatrix} sign(S_1) \\ sign(S_2) \end{pmatrix},\ A_{1,2} = \begin{pmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{pmatrix} = D * \begin{pmatrix} \delta_1 & 0 \\ 0 & \delta_2 \end{pmatrix}
$$

$$
A_{3,4} = \begin{pmatrix} A_{31} & A_{32} \\ A_{41} & A_{42} \end{pmatrix} = \begin{pmatrix} (q_1 - \frac{1}{\hat{T}_r}) * \delta_1 & \hat{x}_5 * \delta_2 \\ \hat{x}_5 * \delta_1 & (q_2 - \frac{1}{\hat{T}_r}) * \delta_2 \end{pmatrix},\ D = \begin{pmatrix} \frac{k}{\hat{T}_r} & k * \hat{x}_5 \\ -k * \hat{x}_5 & \frac{k}{\hat{T}_r} \end{pmatrix}
$$

where q_1,q_2, δ_1 and δ_2 are positive constants; and T_{re} is the estimate of rotor time constant.

The sliding mode observer gains are chosen in such away that the *Lyapunov* stability conditions are satisfied [6] .This yields:

$$
\begin{cases}
\delta_1 \geq \left| \left(\left(\dfrac{kM}{Lr}(Rr - \hat{Rr}) \left(\dfrac{x_3}{M} - x_1 \right) + kx_5 e_4 + \dfrac{\hat{Rr}}{Lr} k e_3 \right) \right| \\
\delta_2 \geq \left| \left(\left(\dfrac{kM}{Lr}(Rr - \hat{Rr}) \left(\dfrac{x_4}{M} - x_2 \right) + kx_5 e_3 + \dfrac{\hat{Rr}}{Lr} k e_4 \right) \right|
\end{cases}
\tag{24}
$$

where e_3, e_4 are flux errors in the $(\alpha,\ \beta)$ frame.

It is well known that sliding mode techniques generate undesirable chattering. This problem can be remedied by replacing the switching function by a smooth continuous function [6, 7]. Then, we propose replace the **Sign** function by the following function:

$$Sat(S_i)=\begin{cases} 1 & if \quad S_i > \lambda \\ -1 & if \quad S_i < -\lambda \\ \frac{S_i}{\lambda} & if \quad |S_i| < \lambda \end{cases} \tag{25}$$

Where λ is a positive small constant, $i=1,2$.

4.2. Estimate speed and electromagnetic torque

After having estimated the rotor flux, the estimate of the electric speed machine is determine via the basic control law formalism: $\omega_r = \omega_s - \omega_g$. The synchronous speed is deduced from the calculation of the derivative rotor flux position and the sleep speed [8]:

$$\hat{\omega}_r = \frac{\phi_{r\alpha}\frac{d\phi_{r\beta}}{dt} - \phi_{r\beta}\frac{d\phi_{r\alpha}}{dt}}{\phi_{r\alpha}^2 + \phi_{r\beta}^2} - \frac{2}{3p}\frac{R_r \hat{T}_{em}}{\phi_{r\alpha}^2 + \phi_{r\beta}^2} \tag{26}$$

The electromagnetic torque is estimated via the observation of the rotor flux and the measurements of the stator currents as follows [7,8]:

$$\hat{T}_{em} = \frac{3}{2} p \frac{M}{L_r}(\hat{\phi}_{r\alpha} i_{s\beta} - \hat{\phi}_{r\beta} i_{s\alpha}) \tag{27}$$

4.3. Estimation of the Rotor Resistance Using the Direct Lyapunov Approach.

In order to reach an asymptotic convergence of the estimated current value to the actual current value, and to take into account the perturbation of the rotor resistance, we considered the *Lyapunov* function given by [7]:

$$V = \frac{1}{2}S^T S + \frac{(R_r - \hat{R}r)^2}{2q_3} \tag{28}$$

where q_3 is a positive constant.

A numerical procedure has been developed to calculate the *Lyapunov* function which converges to zero. Thus satisfy the condition of the *Lyapunov* stability. Under such condition, the control law of the rotor resistance is given by [6,8]:

$$\frac{d\hat{Rr}}{dt} = q_3 (A_{1,2} I_S)^T C \qquad (29)$$

where $C = \frac{\varphi_{re}}{Lr} - \frac{M}{Lr} i_s$ (30)

5. Simulation results

The proposed vector control system is applied to a 1.5 kW IM that its parameter are given in the appendix (A). The stator current is limited to 1.5 of rated current. Fig. 2 shows a block diagram of the proposed scheme applied to an induction machine drive system. The whole system consists of speed regulator, current regulator, SMRFO, and a sensorless speed algorithm associated to adaptive rotor resistance estimation. Extensive simulations results are presented to evaluate the motor drive performance.

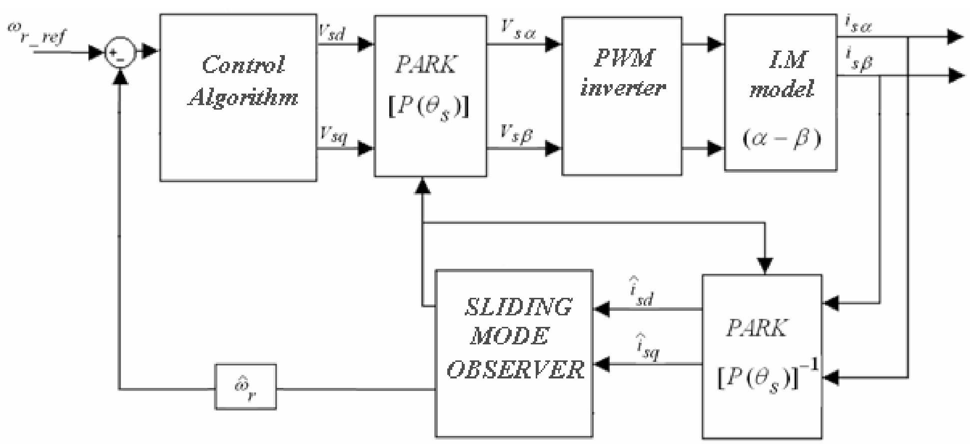

Fig.2 : Proposed Control Structure

By figure 3 one illustrates the machine response for a speed varying from 0 to 400 rd/s while considering the mutual inductance variation expressed by equation 8.

Knowing the speed reference is a ramp, the current remains practically constant for speeds weak (lower at the speed limits), in the same way for the electromagnetic torque and flux. Once that speed exceeds a certain speed, known as speed limit, the system enters in the weakening region explained by a flux, current and electromagnetic torque reduction.

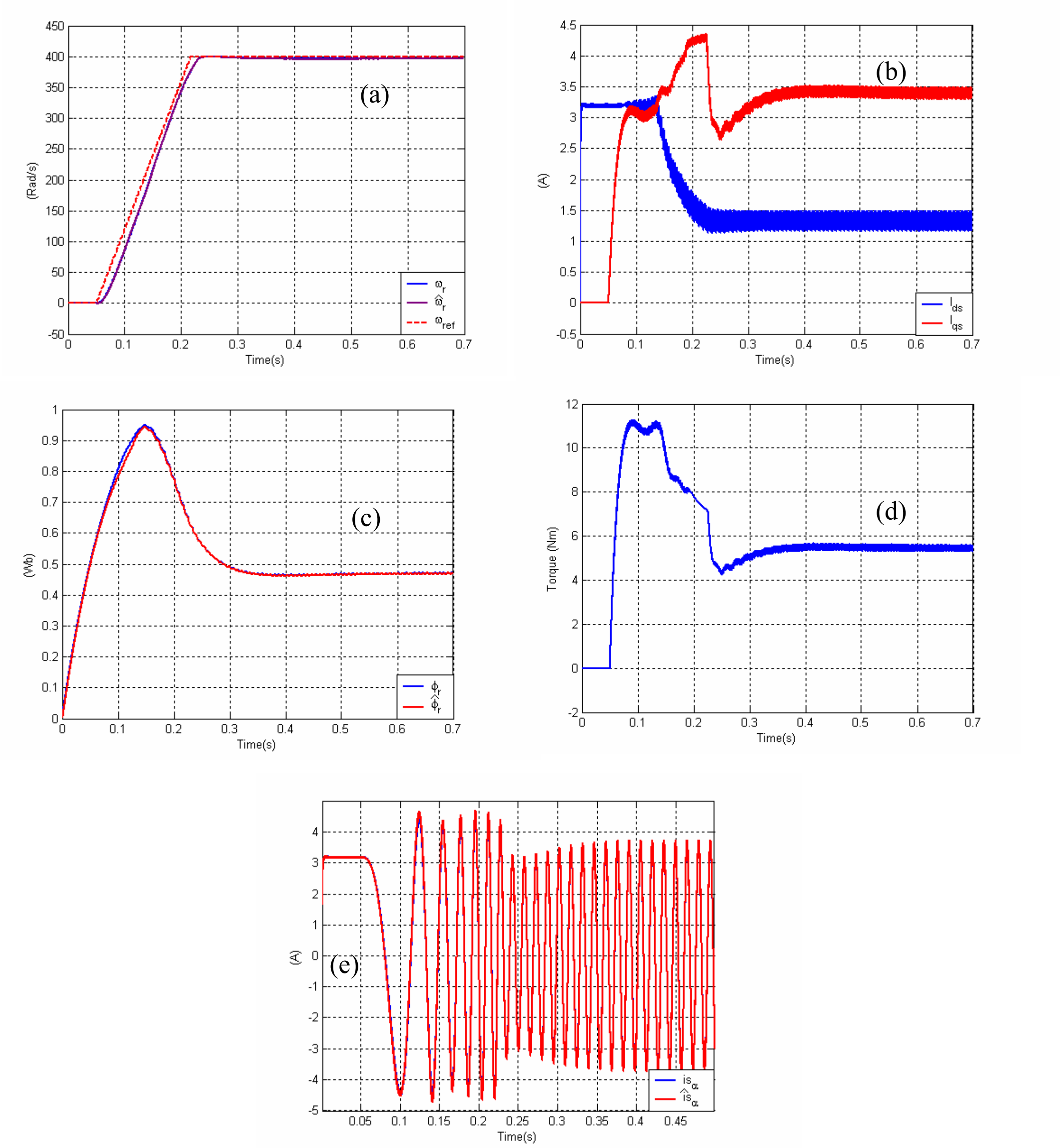

Fig. 3: Response of the machine without parameter variations

Fig.4 shows, the results of the transient response of rotor flux, d and q "axis currents", and torque as the speed changes from 0 rd/s to 400 rd/s in the case of using rated voltage constraint.

To investigate the drive performance and to characterize the steady-state regime with respect to rotor resistance variation, several tests have been carried out. Initially, the rotor resistance is choose equal to the rated value, after a time equal to 0.3 sec, we introduce an increase step equal to 50% of the nominal value. At a time equal to 0.4 sec, the adaptive control law for rotor resistance estimation is used.

Fig. 4(d) shows the transition into field weakening operation can be obtained without reduction in the output torque. The estimated rotor speed converges to real speed practically with zero steady state error.

By taking into account, the magnetising inductance model, the q“-axis current” magnitude is decreased as shown in Fig. 4(b). As a result, the output torque is reduced.

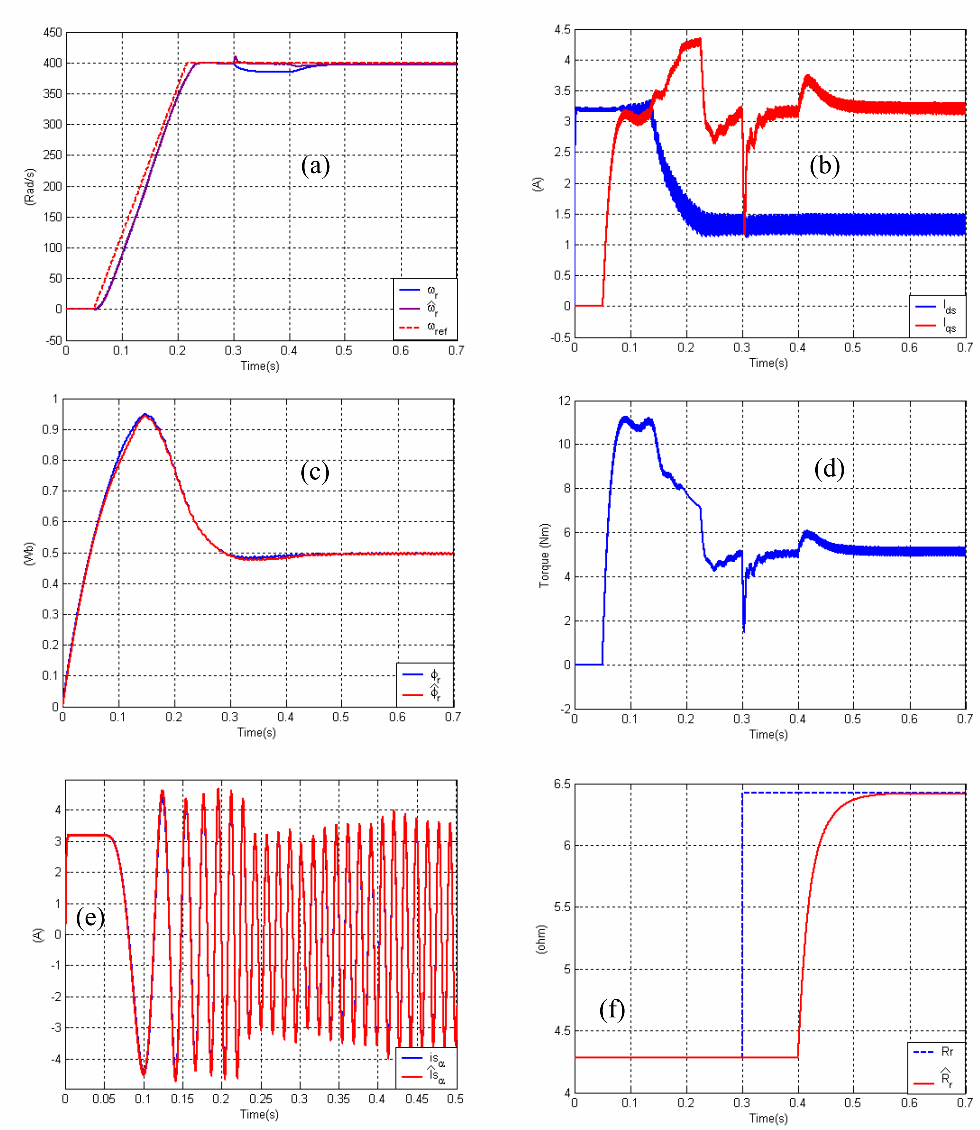

Fig. 4: Response of the machine with parameter variations

6. Conclusion

The performances sensorless RFOC scheme under field weakening operation has been studied in this paper. The proposed sensorless control approach for IM drives is based on the FOC technique. Considering the constraints of voltage and current, a field-weakening scheme giving maximum torque of the IM was presented. For achieving a robust drive system to parameter variations, the DFOC combining width an adaptive control law for rotor resistance based on *Lyapunov* stability theory, is used in the field-weakening region. Simulation results confirm the validity of the proposed control algorithm and verify that the adopted scheme provides an improved maximum torque in transient state regime.

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Appendix A: Parameters of the motor

Rated power: 1.5 Kw,
Rated stator voltage: 231/400V, 50Hz,
Rated stator current: 3.2/5.5A,
R_{r0}=4.282 Ω, R_s=5.717Ω,
L_0=0.0223H,L_{s0}=L_{r0}=0.4466H,M_0=0.4246H,
σ =0.09734, J=0.0049 kgm^2, f=0.029.