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On Nonlinear Control of Induction Motors: Comparison of two Approaches

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***Abstract*— In this paper, a novel approach to nonlinear control of induction motors is developed. The proposed approach is used to design controllers for the rotor flux amplitude and the motor speed. The underlying design objective is to endow the closed loop system with high performance dynamics for high speed ranges while maximizing power efficiency and keeping the required stator voltage within the inverter ceiling limits. A comparative study between the performances of the proposed controller and field oriented control is carried out. The methods are compared in terms of their ability to handle loads on the motor shaft, their speed tracking capability and their sensitivity to operating condition variations. To estimate the rotor flux, an open loop observer is used.**

***Index Terms*— Nonlinear Control, induction motors, field orientation, input output linearization, observer.**

NOMENCLATURE

(d,q) : Rotating reference frame
i_{sd}, i_{sq} : d,q axis stator current components
V_{sd}, V_{sq} : d,q axis stator voltage components
i_{rd}, i_{rq} : d,q axis rotor current components
ϕ_{sd}, ϕ_{sq} : d,q axis stator flux linkage
ϕ_{rd}, ϕ_{rq} : d,q axis rotor flux linkage
$\Phi_r = (\phi_{rd}^2 + \phi_{rq}^2)$: Square rotor flux magnitude
Ω : Mechanical speed
ω : Shaft angular speed
ω_s : Stator electric angular pulsation
R_s, R_r : Stator,rotor resistances
L_s, L_r : Stator,rotor inductances
M_{sr} : Mutual inductance
J : Moment of inertia
C_r : Load torque
C_{em} : Electromagnetic torque

I. INTRODUCTION

Although all practical systems are nonlinear, most of the time linear control based on Jacobian linearization of nonlinear models is good enough [1]. However, when the system requirement is stringent and the nonlinearities are significant, linear control may not satisfy system specifications. In many cases, nonlinear control can utilize the structure of a specific system and improve the performance [2]. As pointed out in [10], electromechanical systems are good candidates for nonlinear control applications because the nonlinearities, being modeled on the basis of physical principles, are often significant and exactly known. Due to their reliability, ruggedness and relatively low cost, induction motors are widely used in industry. In contrast to DC motors, they can be used in aggressive environments since there is no problem with spark and corrosion [3]. However, from the control point of view, they represent a complex multivariable nonlinear problem and thus provide an interesting application area for nonlinear control theory. Induction motors constitute a class of highly coupled nonlinear multivariable systems with two control inputs (stator voltages) and two output variables (rotor speed, rotor flux magnitude) required to track desired reference signals [4]. An important practical task is to solve the induction motor control problem to achieve (simultaneously) high dynamic performance, high energy efficiency, robustness and simple implementation. One particular approach for the control of induction motors is the Field Oriented Control [5], [4]. Partial feedback linearization together with a proportional integral (PI) controller is used to regulate the motor states. In low speed ranges, this control strategy achieves the required control objective asymptotically. To improve the Field Oriented Control, full linearizing state feedback control based on differential geometric theory [2], has been proposed in [11], [9] for the electromagnetic torque control and in [10] for the adaptive speed control of a fifth order model of an induction motor.

In this paper, a comparative study between the classical Field Oriented Control [5], [4] and a newly proposed nonlinear controller has been carried out. The new controller is based on the theory of feedback linearization. The controller is used for the speed control of a fourth-order model of an induction motor. Since all the states are not available for direct measurement, an open loop flux observer is proposed for the estimation of the rotor flux. This paper is organized as follows. In section 2, the dynamic model of the induction motor is described. The Field Oriented Control is reviewed in section 3, and analysis of the necessity of a high performance control strategy for high speed ranges is carried on in the same section. In section 4, the nonlinear controller is designed for the speed and flux magnitude control of a fourth order model of an induction motor. The observer required for the rotor flux estimation is presented in section 5. Section 6 provides numerical simulation results, followed by the conclusion.

II. NONLINEAR INDUCTION MOTOR MODEL

The dynamic equations of an induction motor in the synchronously rotating reference frame are formulated as:

$$\begin{aligned} V_{sd} &= R_s i_{sd} + \tfrac{d\phi_{sd}}{dt} - \omega_s \phi_{sq} \\ V_{sq} &= R_s i_{sq} + \tfrac{d\phi_{sq}}{dt} + \omega_s \phi_{sd} \\ 0 &= R_r i_{rd} + \tfrac{d\phi_{rd}}{dt} - (\omega_s - \omega)\phi_{rq} \\ 0 &= R_r i_{rq} + \tfrac{d\phi_{rq}}{dt} - (\omega_s + \omega)\phi_{rd} \end{aligned} \quad (1)$$

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The above model is based on the two-phase equivalent machine representation, assuming equal mutual inductances and linear magnetic circuits, and neglecting iron (core) losses.The stator and rotor winding flux linkages are expressed as:

$$\begin{array}{l} \phi_{sd} = L_s i_{sd} + M_{sr} i_{rd} \\ \phi_{sq} = L_s i_{sq} + M_{sr} i_{rq} \\ \phi_{rd} = M_{sr} i_{sd} + L_r i_{rd} \\ \phi_{rq} = M_{sr} i_{sq} + L_r i_{rq} \end{array} \quad (2)$$

Where: $\phi_{sd}, \phi_{sq}, \phi_{rd}$ and ϕ_{rq} are respectively the stator and rotor fluxes projections on the (d, q) axis reference frame. L_s,L_r are the stator and rotor self-inductances and M_{sr} is the mutual inductance.

The electromagnetic torque developed by the motor is expressed in terms of rotor fluxes and stator currents as:

$$C_{em} = np\frac{M_{sr}}{L_r}(i_{sd}\phi_{rd} - i_{sd}\phi_{rq}) \quad (3)$$

While the load torque acts as a disturbance via the mechanical relation:

$$J\frac{d\Omega}{dt} = C_{em} - C_r \quad (4)$$

By adding the rotor dynamics (1) and (2) to the electromagnetic dynamics (4), and considering a reference frame for the Park transformation, rotating at the same angle as the magnetizing current θ_s such as $\frac{d\theta_s}{dt} = \omega_s$, the overall dynamics of an induction motor can be written in a state space representation. By considering as state variables the stator currents (i_{sd}, i_{sq}), the rotor fluxes (ϕ_{rd}, ϕ_{rq}) and the rotor angular speed ω and two control variables, the stator voltages (V_{sd}, V_{sq}) yielding the following fifth-order model:

$$\dot{x} = f(x) + gu \quad (5)$$

with: $x = \begin{bmatrix} \omega & \phi_{rd} & \phi_{rq} & i_{sd} & i_{sq} \end{bmatrix}^T$, the state vector

$u = \begin{bmatrix} V_{sd} & V_{sq} \end{bmatrix}^T$, the control vector

$$f(x) = \begin{bmatrix} \mu(\phi_{rd}i_{sq} - \phi_{rq}i_{sd}) - \frac{npC_r}{J} \\ -\frac{1}{\tau_r}\phi_{rd} + (\omega_s - \omega)\phi_{rq} + \frac{M_{sr}}{\tau_r}i_{sd} \\ -\frac{1}{\tau_r}\phi_{rq} - (\omega_s - \omega)\phi_{rd} + \frac{M_{sr}}{\tau_r}i_{sq} \\ \frac{\beta}{\tau_r}\phi_{rd} + \beta\omega\phi_{rq} - \frac{1}{\tau_1}i_{sd} + \omega_s i_{sq} \\ \frac{\beta}{\tau_r}\phi_{rq} - \beta\omega\phi_{rd} - \frac{1}{\tau_1}i_{sq} - \omega_s i_{sd} \end{bmatrix}$$

and

$$g = \begin{bmatrix} g_a & g_b \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & \frac{1}{L_1} & 0 \\ 0 & 0 & 0 & 0 & \frac{1}{L_1} \end{bmatrix}^T$$

The parameters of the considered motor and time constants are given by these expressions: $\mu = np^2\frac{M_{sr}}{JL_r}$;$\tau_r = \frac{L_r}{R_r}$; $\beta = \frac{M_{sr}}{L_r L_1}$; $L_1 = L_s - \frac{M_{sr}^2}{L_r}$; $R_1 = R_s + R_r(\frac{M_{sr}}{L_r}^2)$; $\tau_1 = \frac{L_1}{R_1}$.

From the above representation, we can see that the dynamic model of an induction motor is a strongly coupled nonlinear multivariable system. The control problem is to choose V_{sd}, V_{sq} in such a way as to force the motor electrical angular speed ω and the rotor flux magnitude $\phi_r = \sqrt{\phi_{rd}^2 + \phi_{rq}^2}$ to track given reference values denoted by ω_{ref} and ϕ_{ref}, respectively. Note that the choice of a reference frame rotating at the same angle as the magnetizing current $\dot{\theta}_s = \omega_s$ is more suitable for this control problems since in this frame the steady state signals are constant.

III. FIELD ORIENTED CONTROL

One particular approach for the control of induction motors is the Field Oriented Control (FOC) introduced by Balaschke in [5]. This control strategy is based on the orientation of the flux vector along the d axis [4], [6] which can be expressed by considering :

$$\phi_r^{d,q} = (\phi_r, 0)^T \quad (6)$$

Using (6), we eliminate all the terms with quadratic rotor flux and reduce the third equation in (5) to this expression of the synchronous angular speed:

$$\omega_s = \omega + \frac{M_{sr}i_{sq}}{\tau_r\phi_r} \quad (7)$$

The rotor flux orientation leads to the following expression of the electromagnetic torque, which is equivalent to a separately excited DC motors:

$$C_{em} = np\frac{M_{sr}}{L_r}(i_{sd}\phi_{rd}) \quad (8)$$

With the above assumptions (6) and (7), the fifth order model (5) is reduced to the fourth-order:

$$\begin{cases} \frac{d\omega}{dt} = \mu(\phi_r i_{sq}) - \frac{npC_r}{J} \\ \frac{d\phi_r}{dt} = -\frac{1}{\tau_r}\phi_r + \frac{M_{sr}}{\tau_r}i_{sd} \\ \frac{di_{sd}}{dt} = \frac{\beta}{\tau_r}\phi_r - \frac{1}{\tau_1}i_{sd} + \omega_s i_{sq} + \frac{1}{L_1}V_{sd} \\ \frac{di_{sq}}{dt} = -\beta\omega\phi_r - \frac{1}{\tau_1}i_{sq} - \omega_s i_{sd} + \frac{1}{L_1}V_{sq} \end{cases} \quad (9)$$

Note that the flux is now just proportional to the stator current and the speed is proportional to the product of the flux and the quadratic component of the stator current.

If we choose a nonlinear state feedback control (V_{sd}, V_{sq}) such as:

$$\begin{bmatrix} V_{sd} \\ V_{sq} \end{bmatrix} = L_1 \begin{bmatrix} \frac{-\beta}{\tau_r}\phi_r - \omega i_{sq} - \frac{M_{sr}}{\tau_r}(\frac{i_{sq}^2}{\phi_r}) + V_d \\ \beta\omega\phi_r + \omega i_{sd} + \frac{M_{sr}}{\tau_r}(\frac{i_{sd}i_{sq}}{\phi_r}) + V_q \end{bmatrix} \quad (10)$$

the system (9) presents a simple structure and the dynamics of the flux, now linear, can be expressed as:

$$\begin{cases} \frac{d\phi_r}{dt} = -\frac{1}{\tau_r}\phi_r + \frac{M_{sr}}{\tau_r}i_{sd} \\ \frac{di_{sd}}{dt} = -\frac{1}{\tau_1}i_{sd} + V_d \end{cases} \quad (11)$$

The control signal V_d can now be designed as a PI loop of the form:

$$V_d = -K_{d1}(\phi_r - \phi_{ref}) - K_{d2}\int_0^t(\phi_r(\tau) - \phi_{ref}(\tau)d\tau \quad (12)$$

When the flux reaches the desired value ϕ_{ref}, the dynamic of the speed is linear and is described by:

$$\left\{ \begin{array}{l} \frac{d\omega}{dt} = \mu(\phi_r i_{sq}) - \frac{npC_r}{J} \\ \frac{di_{sq}}{dt} = -\frac{1}{\tau_1} i_{sq} + V_q \end{array} \right. \tag{13}$$

and the speed can be controlled by V_q using a PI control loop:

$$V_q = -K_{q1}(\omega - \omega_{ref}) - K_{q2}\int_0^t (\omega(\tau) - \omega_{ref}(\tau) d\tau \tag{14}$$

Field Oriented Control makes it possible to control the induction motor in a manner similar to a separately excited DC motor; this allows the induction motor to be used in applications requiring high dynamic performances where traditionally only DC drives could be applied [4]. One drawback of this method is that it assumes the magnitude of the rotor flux to be regulated to a constant value. In reality, to prevent saturation of the stator voltages in high speed ranges, the field needs to be weakened. This is shown in the following section.

A. Requirement of Field Weakening for High Performance Control Strategy

Equations (9) show that the speed is controlled by the quadrature component of the stator current i_{sq}. In order to increase the speed, V_{sq} must be chosen such that $\frac{di_{sq}}{dt} \succ 0$, that is:

$$-\beta\omega\phi_r - \frac{1}{\tau_1} i_{sq} - \omega_s i_{sd} + \frac{1}{L_1} V_{sq} \succ 0 \tag{15}$$

replacing β and τ_1 by their values we have:

$$V_{sq} \succ L_1(-\omega + \frac{M_{sr}}{\tau_r}\frac{i_{sq}}{\phi_r}) + \frac{M_{sr}}{L_r}\omega\phi_r + R_1 i_{sq} \tag{16}$$

As the quantity L_1 is quite small, the dominant term on the right hand side of this inequality is $(\frac{M_{sr}}{\tau_r}\frac{i_{sq}}{\phi_r})$, which means high speeds require rather large input voltages. In practice, the voltages must be kept within the inverter ceiling limits [4]; so the flux ϕ_r is decreased from the nominal as the speed ω increases above rated speed. This method of reducing the flux at high speeds is called "flux weakening" [8]. That is, the flux is required to reach the nominal value ϕ_n for $\omega \prec \omega_n$, ω_n denotes the nominal speed, and the rotor flux amplitude has to be weakened according to the rule $\phi_{ref} = |\phi_n|\frac{\omega_n}{\omega}$ for $\omega > \omega_n$ [4], [8].

Operating in the flux weakening regime will maximize power efficiency so that only the minimum stator input power needed to operate at the desired speed is used. That is, even when the motor is operating below the nominal speed, flux may be varied in order to maximize power efficiency [4], [7].

A disadvantage of the Field Oriented Controller is that the method assumes the magnitude of the rotor flux to be regulated to a constant value. Therefore, the dynamics of the speed and flux may interfere in the high speed ranges. Eliminating this coupling and achieving high performance dynamics for all speed ranges can be realized by considering an input-output linearization technique [2].

IV. DESIGN OF A NONLINEAR INPUT-OUTPUT LINEARIZING CONTROLLER FOR INDUCTION MOTORS

A. Input-Output Linearization Technique

The input-output control problem is to find a state feedback such that the transformed system is input-output decoupled that is, one input influence one output only [2]. The technique requires measurements of the state vector x in order to transform a multi-input nonlinear control system:

$$\left\{ \begin{array}{l} \dot{x} = f(x) + \sum_{i\in m} g_i(x)u_i \\ y = h(x) \end{array} \right. (z \in \Re^n, v \in \Re^m) \tag{17}$$

into a linear and controllable one

$$\dot{z} = Az + Bv, \quad (z \in \Re^n, v \in \Re^m) \tag{18}$$

by means of nonlinear state feedback:$u = \alpha(x) + \beta(x)v$, with $\beta(x)$ a nonsingular $(m \times m)$ matrix and nonlinear state space change of coordinates $z = h(x)$. Linear techniques can then be applied in the design of the control v [2].

The outputs to be controlled are :

$$\left[\begin{array}{ccc} h_1(x) & .. & h_m(x) \end{array} \right]^T \tag{19}$$

The manipulated quantities are differentiated with respect to time until the input appears and the derivatives of the state variables are eliminated using the state space model of the system. This can be done introducing the directional or Lie derivative of a state function $h(x) : \Re^n \longrightarrow \Re$ along a vector field $f(x) = (f_1(x), ..., f_n(x)$

$$L_f h(x) = \sum_{i=1}^{n} f_i(x)\frac{\partial h}{\partial x_i}(x) \tag{20}$$

Iteratively, $L_f^i h(x) = L_f L_f^{i-1}(h(x))$. That is, once the system is linearized, one can use a linear controller in the design of the control signal input v for the system in the new reference.

B. Application to the Speed Control of Induction Motors

The controller design is based on the fourth order dynamic model obtained from the (d,q) axis model of the motor under the field oriented assumptions so that either speed or flux magnitude control objective can be fulfilled. The underlying design concept is to endow the closed loop system with high performance dynamics for high speed ranges while maximizing power efficiency and keeping the required stator voltage within the inverter ceiling limits.

In addition to fulfilling those control objectives, our control design aims to reduce the complexity of the control scheme, saving thereby the computation time of the control algorithm, which is an improvement over previous work found in the technical literature [9], [10]. The outputs to be controlled are the speed ω and the square of the rotor flux magnitude $\Phi_r = \phi_r^2$. The output vector is:

$$\left[\begin{array}{cc} h_1(x) & h_2(x) \end{array} \right]^T = \left[\begin{array}{cc} \omega & \phi_r^2) \end{array} \right]^T \tag{21}$$

Define the change of coordinates:

$$\begin{cases} z_1 = h_1(x) = \omega \\ z_2 = L_f h_1(x) = \mu(\phi_r i_{sq}) - \frac{npC_r}{J} \\ z_3 = h_2(x) = \phi_r^2 \\ z_4 = L_f h_2(x) = -\frac{2}{\tau_r}\phi_r^2 + \frac{2M_{sr}}{\tau_r}\phi_r i_{sd} \end{cases} \quad (22)$$

Thus the derivatives of the outputs are given in the new coordinate system by:

$$\begin{cases} \dot{z}_1 = h_1(x) = z_2 \\ \dot{z}_2 = \ddot{h}_1(x) = L_f^2 h_1(x) + Lg_a L_f h_1(x) V_{sd} \\ \qquad + Lg_b L_f h_2(x) V_{sq} \\ \dot{z}_3 = \dot{h}_2(x) = z_4 \\ \dot{z}_4 = \ddot{h}_2(x) = L_f^2 h_2(x) + Lg_a L_f h_2(x) V_{sd} \\ \qquad + Lg_b L_f h_2(x) V_{sq} \end{cases} \quad (23)$$

This system can be written as:

$$\begin{bmatrix} \ddot{z}_1 \\ \ddot{z}_3 \end{bmatrix} = \begin{bmatrix} L_f^2 h_1(x) \\ L_f^2 h_2(x) \end{bmatrix} + \triangle(x) \begin{bmatrix} V_{sd} \\ V_{sq} \end{bmatrix} \quad (24)$$

with:

$$L_f^2 h_1(x) = -\mu(\tfrac{1}{\tau_r} + \tfrac{1}{\tau_1})(i_{sd} + i_{sq})\phi_r + \tfrac{\mu}{\tau_r} M_{sr} i_{sd} i_{sq} \\ - \mu\beta\omega\phi_r^2$$

$$L_f^2 h_2(x) = \tfrac{2}{\tau_r^2}(2 + \beta M_{sr})\phi_r^2 - (\tfrac{6M_{sr}}{\tau_r^2} + \tfrac{2M_{sr}}{\tau_r\tau_1}) i_{sd}\phi_r \\ + 2\tfrac{M_{sr}}{\tau_r}\omega_s i_{sq}\phi_r + 2\tfrac{M_{sr}^2}{\tau_r^2} i_{sd}^2$$

$$Lg_a L_f h_1(x) = 0 \quad Lg_b L_f h_1(x) = \tfrac{\mu}{L_1}\phi_r$$

$$Lg_a L_f h_2(x) = 0 \quad Lg_b L_f h_2(x) = \tfrac{2M_{sr}}{L_r\tau_r}\phi_r$$

The decoupling matrix $\Delta(x)$ is defined as:

$$\triangle(x) = \begin{bmatrix} L_{ga} L_f h_1(x) & L_{gb} L_f h_1(x) \\ L_{ga} L_f h_2(x) & L_{gb} L_f h_2(x) \end{bmatrix} \\ = \begin{bmatrix} 0 & \frac{\mu}{L_1}\phi_r \\ \frac{2M_{sr}}{\tau_r L_1}\phi_r & 0 \end{bmatrix}$$

and:

$$det(\Delta(x)) = -\frac{2\mu M_{sr}}{\tau_r L_1^2}\phi_r^2 \quad (25)$$

The decoupling matrix $\Delta(x)$ is singular if and only if ϕ_r^2 is zero which only occurs at the start up of the motor. That is, to fulfill this condition one can use in a practical setting, an open loop controller at the start up of the motor, and then switch to the nonlinear controller as soon as the flux goes up to zero.

If the decoupling matrix is not singular, the nonlinear state feedback control is given by:

$$\begin{bmatrix} V_{sd} \\ V_{sq} \end{bmatrix} = \triangle(x)^{-1} \begin{bmatrix} -L_f^2 h_1(x) + V_1 \\ -L_f^2 h_2(x) + V_2 \end{bmatrix} \quad (26)$$

This controller linearizes and decouples the system, resulting in:

$$\begin{cases} \ddot{h}_1 = V_1 \\ \ddot{h}_2 = V_2 \end{cases} \quad (27)$$

The closed loop system (27) is input-output decoupled and linear. To ensure perfect tracking of speed and flux references, V_1 and V_2 are chosen as follows:

$$\begin{cases} V_1 = -k_{a1}(\omega - \omega_{ref}) - k_{a2}\dot{\omega} - \int_0^t \omega \\ V_2 = -k_{b1}(\Phi_r - \Phi_{ref}) - k_{b2}\dot{\Phi}_r \end{cases} \quad (28)$$

Where k_{a1}, k_{a2} and k_{b1}, k_{b2} are positive non-zero constants to be determined in order to make the closed loop system (27) stable and to have fast response in variable tracking. The proposed control strategy is illustrated by the control block diagram reported in Fig.1. The flux loop and speed loop regulation are given by Fig.2 and Fig.3 respectively.

V. INDUCTION MOTOR ROTOR FLUX ESTIMATION

In practice, only the stator currents and rotor speed or position are available for measurement; therefore we propose to estimate the rotor flux using an open loop observer [12]. A number of flux observers have been developed and documented in the literature [12], [13]. In this work, we adopt the following observer to estimate the rotor flux. The flux estimator is governed by the state space model:

$$\begin{cases} \frac{d\widehat{\phi}_{rd}}{dt} = -\frac{1}{\tau_r}\widehat{\phi}_r + \frac{M_{sr}}{\tau_r} i_{sd} \\ \frac{d\widehat{\theta}_s}{dt} = \omega + \frac{M_{sr} i_{sq}}{\tau_r \widehat{\phi}_r} \end{cases} \quad (29)$$

where $\widehat{\phi}_r$ is the estimated rotor flux, $\widehat{\theta}_s$ is the estimated flux angular position and i_{sd} and i_{sq} are respectively, the direct and quadratic components of the stator current, available from measurement.

The flux estimation error is:

$$e_r = \begin{bmatrix} \widehat{\phi}_r - \phi_r \end{bmatrix} \quad (30)$$

and its dynamic is governed by the following equation:

$$\frac{de_r}{dt} = -\frac{1}{\tau_r} e_r. \quad (31)$$

Equation (31) shows that the rotor flux magnitude converges exponentially to its actual value. Hence, the convergence of the error dynamics of this observer is limited by the rotor time constant τ_r. The algorithm proposed above, applied to the nonlinear control of an induction motor, has been implemented in the MATLAB environment. The rotor flux has been estimated with the flux observer (29).

VI. SIMULATION RESULTS

To validate the performances of the proposed controller, we provide a series of simulations and a comparative study between the performances of the proposed control strategy and those of the classical Field Oriented Control. The simulations are conducted by a 1.5 kW, two pole-pair cage rotor induction machine. The motor parameters and specifications are listed in the appendix. The simulation test involves the following operating sequences: the motor is required to reach the reference value $\omega_{ref} = 200\ rad/s$

in the interval of time $[0-3s]$ and $\omega_{ref} = 400\ rad/s$ for $t > 3s$. The controller gains have been chosen as follows: $k_{a1} = 2.10^3; k_{a2} = 200; k_{b1} = 10^3; k_{b2} = 100$ in order to obtain fast and precise response in speed and flux tracking.

In the first simulation, load torque disturbance and field weakening function are omitted. That is, the rotor flux is required to track a constant reference value of 1Wb for all speed ranges and the motor shaft is subject to the nominal load torque.The same simulation conditions are applied for both methods. The time histories of speed and flux magnitude tracking behavior are reported on figure 4, for the nonlinear controller (NLC) and on figure 5 for the Field oriented Control. As the figures show, it is observed that the speed and flux tracks the reference values adequately well, for both methods. That is, without load torque perturbation and with considering a constant value for the reference flux, the two methods demonstrate nearly the same dynamic behavior.

In the second test, both the field weakening function and load torque disturbance are considered. That is, the flux is required to reach the nominal value $\phi_{ref} = 1wb$ when $\omega \prec \omega_{ref}$ and the rotor flux amplitude is then weakened according to the rule $\phi_{ref} = |\phi_n|\frac{\omega_n}{\omega}$ for $\omega \succ \omega_n$ so as not to saturate voltages for high speed ranges [4], where ω_n and ϕ_n denote the nominal speed and flux respectively.

To investigate the disturbance rejection of the controlled system and the decoupling between the speed and the flux magnitude, the motor shaft is subject to a step load torque of $5Nm$ for $1 < t < 2s$ and $4 < t < 6s$, respectively. Note that the load torque is applied in two regions, when the speed is lower than the nominal value, $\omega_n = 314\ rad/s$, and then again when the speed exceeds the nominal value. The time histories of speed and flux magnitude tracking behavior are shown in figure 7 for the nonlinear controller. Note that speed and flux reach the desired reference values, and there is no effect of the load torque variation on the flux for all speed ranges; that is, speed and flux are decoupled even when the flux had to be lowered. Thus, the input output control loop achieves good tracking performance and disturbance rejection in all speed ranges. Moreover, power efficiency is improved by adjusting flux levels, without affecting speed regulation. The plots of speed and flux magnitude tracking behavior are shown in figure 9 for the Field Oriented Control. The simulations show that the controller performs satisfactorily. However, a coupling between the speed and the flux appears in the flux weakening region. That is, operating in flux-weakening regime excites the coupling between flux and speed in the classical Field Oriented Control, causing undesired speed fluctuation. The time behavior of the (d, q) stator current components is shown in figure 6 for the first test and in figures 8 and 10 for the second test. Those figures show that currents are within acceptable limits. Notice the peak in stator current corresponding to the reference velocity changes. This peak current is required to accelerate the rotor to the desired speed.

As the figures show, the two methods demonstrate nearly the same dynamic behavior; however, when flux has to be lowered, there is coupling between flux and speed for the Field Oriented controller. On the contrary, the nonlinear controller achieves exact decoupling between outputs, which leads to better performance in all speed ranges.

VII. CONCLUSION AND FUTURE WORK

In this paper, two control techniques have been compared for induction motors: classical Field Oriented control, and input-output linearizing control, proposed by the current authors. From the comparative study, one can conclude that the two methods demonstrate nearly the same dynamic behavior. However, the input-output linearizing controller shows better performance than the Field Oriented controller in speed tracking at high speed ranges. Currently, the input-output linearizing controller gains have been determined by trial and error. Determining these gains with regard to closed-loop stability as well as performance objectives will be considered in future work.

Appendix
Induction Motor Data

Rated power	P	1.5Kw
Rated voltage		380/220 v
Rated speed		1480 rpm
Number of pole pairs	np	2
Rotor inductance	L_r	0.1568 H
Rotor resistance	R_r	1Ω
Stator inductance	L_s	0.1554 H
Stator resistance	R_s	1.2 Ω
Mutual inductance	M_{sr}	0.15H
Rotor inertia	J	0.013 Kgm^2
Rated Load torque	C_r	5Nm

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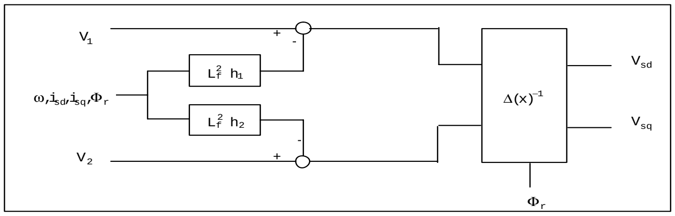

Fig. 1. Control Block Diagram

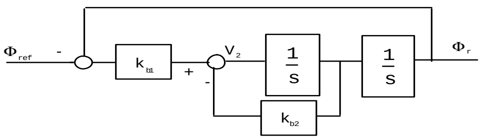

Fig. 2. Flux Loop Regulation

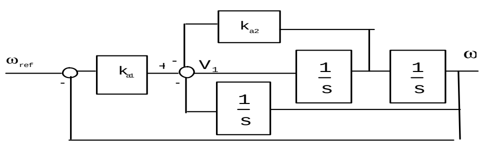

Fig. 3. Speed Loop Regulation

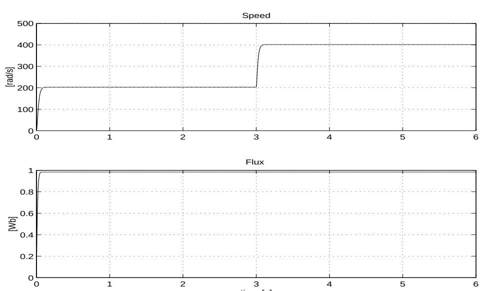

Fig. 4. Dynamics of the Speed and the Flux magnitude (NLC)

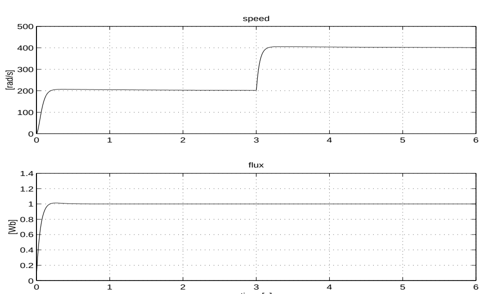

Fig. 5. Dynamics of the Speed and the Flux magnitude (FOC)

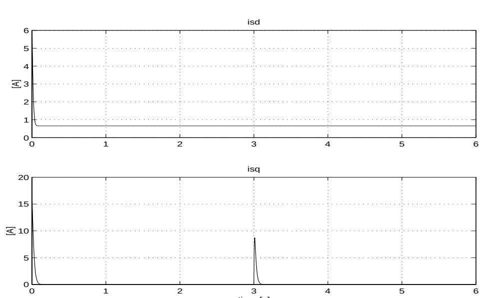

Fig. 6. Direct and Quadratic Components of the Stator Current i_{sd}, i_{sq}

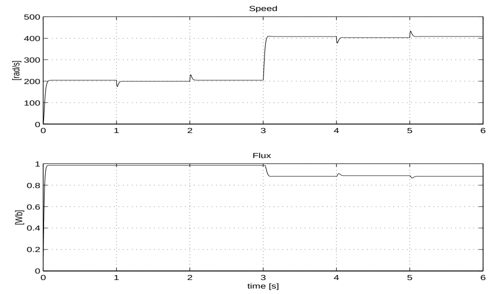

Fig. 7. Dynamics of the Speed and the Flux magnitude in the presence of sudden changes in the load torque (NLC)

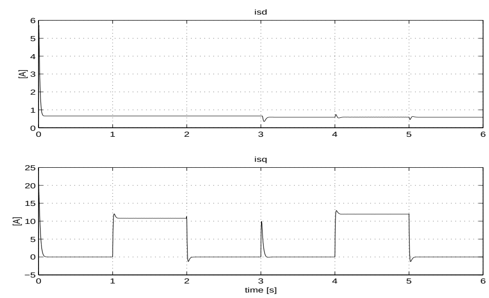

Fig. 8. Direct and Quadratic Components of the Stator Current i_{sd}, i_{sq} in the presence of sudden changes in the load torque (NLC)

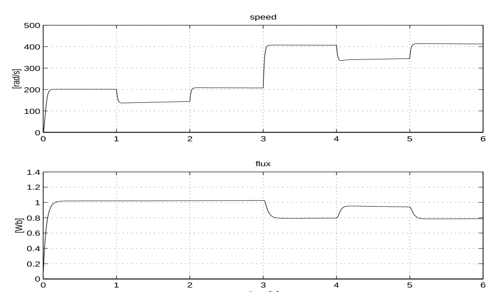

Fig. 9. Direct and Quadratic Components of the Stator Current i_{sd}, i_{sq} in the presence of sudden changes in the load torque (FOC)

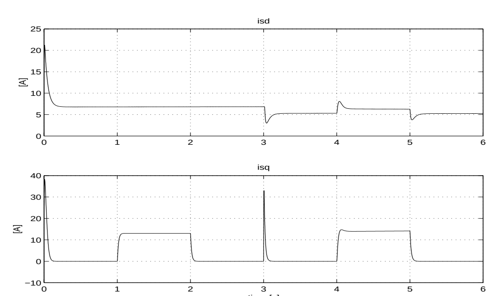

Fig. 10. Direct and Quadratic Components of the Stator Current i_{sd}, i_{sq} in the presence of sudden changes in the load torque (FOC)