Reactive power model used for adaptive and indirect field-orientated induction motor control

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Abstract

This paper studies the use of reactive power for the indirect-field-orientated control (IFOC) of induction motors and a method to achieve a simple and robust drive control in spite of rotor resistance variations. Firstly, the reactive power model is presented showing how the rotor-flux vector velocity can be calculated without using the rotor resistance. Secondly, a Model Reference Adaptive Controller (MRAC) is reviewed to perform IFOC while estimating the rotor resistance on-line. The convergence of the rotor resistance estimator is presented in detail. Finally, an algorithm to achieve indirect field orientation is proposed based on the reactive power "consumed" by the induction motor. It is shown how two different algorithms must be used during braking and motoring for correct IFOC convergence. Experimental results for the MRAC and the proposed controller are presented and compared with results obtained using both a tuned and de-tuned IFOC.

1. Introduction

The *classic*, field-orientated controller is an established method of providing induction motor drives with a torque response similar to that offered by d.c. machines (see [1]). Essentially, it positions the stator current with respect to the rotating flux vector at each sample instant such that the resulting flux vector magnitude remains constant and the electromagnetic torque *tracks* the desired reference input. The controller must know the spatial position of the flux vector, a variable not normally available for measurement. Estimation of the flux vector using a model of the machine rotor circuit remains the only feasible solution. Estimators based on such a model suffer from the following deficiencies:

- The rotor circuit model depends on the equivalent resistance of the rotor coils. This resistance is known to vary with temperature and results in *de-tuned* operation of the controller.
- The rotor circuit model includes the mechanical velocity which must be measured to a high precision. Transducers, such as optical encoders, capable of adequate measurement, are expensive and may require frequent maintenance.

To reduce the *de-tuned* effect of rotor resistance mismatch, adaptive schemes have been proposed to adjust the rotor circuit model on-line (see for example [2] or [3]). An early example of such a controller was proposed by Garces in 1980 in which the reactive power is used for a second estimation of the rotor flux vector velocity (or synchronous frequency, ω_S) which is used to correct the rotor circuit model (see [4]). The technique has received much attention although its convergence properties are deduced from intuitive reasoning, lacking analytic confirmation.

Eliminating the need for a mechanical velocity measurement has introduced the new class of controllers known as *sensorless* field-orientated controllers, examples of which have shown responses equivalent to a tuned field-orientated controller (see [5]). However, often they too are sensitive to resistive machine parameters.

Here, the reactive-power model is presented first as the *reference model* for a model reference adaptive controller (MRAC) in a manner similar to that of Garces. It is used later as the basis of a *sensorless* torque controller from which a dynamic response similar to that of a tuned field-orientated controller is achieved although not requiring either resistive parameters or mechanical velocity measurements. Experimental results are included to demonstrate the operation of the controllers.

2. Machine reactive power model.

The induction motor stator equations may be written as (see [1]),

$$L_\sigma\left(\frac{di^*_{Sd}}{dt}-\omega_c i^*_{Sq}\right)= -R_S i^*_{Sd}+v_{Sd}-\frac{L_M}{L_R}\left(\frac{d\psi_{Rd}}{dt}-\omega_c\psi_{Rq}\right) \quad (1a)$$

$$L_\sigma\left(\frac{di^*_{Sq}}{dt}+\omega_c i^*_{Sd}\right)= -R_S i^*_{Sq}+v_{Sq}-\frac{L_M}{L_R}\left(\frac{d\psi_{Rq}}{dt}+\omega_c\psi_{Rd}\right) \quad (1b)$$

where $\vec{v}_{Sdq}=\begin{pmatrix}v_{Sd} & v_{Sq}\end{pmatrix}^T$ is the stator voltage space vector, $\vec{\psi}_{Rdq}=\begin{pmatrix}\psi_{Rd} & \psi_{Rq}\end{pmatrix}^T$ is the rotor flux space vector and ω_c is the angular velocity of the rotating coordinate system. $\vec{i}^*_{Sdq}=\begin{pmatrix}i^*_{Sd} & i^*_{Sq}\end{pmatrix}^T$ is the vector of *impressed* stator current references when using current control as depicted in Fig. 1.

When the response of the current controller is *significantly* faster than any other machine variable, it may be assumed that $\vec{i}_{Sdq} \approx \vec{i}^*_{Sdq}$ where $\vec{i}_{Sdq} = \begin{pmatrix} i_{Sd} & i_{Sq} \end{pmatrix}^T$ are the *actual* machine stator currents. Note that, i^*_{Sd} is often referred to as the flux current and i^*_{Sq} as the torque current since, during steady-state field orientation $|\vec{\psi}_{Rab}| = L_M i^*_{Sd}$ and the electromagnetic torque $t_e = P(L_M/L_R)|\vec{\psi}_{Rab}| i^*_{Sq}$ with the torque controlled using i^*_{Sq}. P is the number of pole pairs.

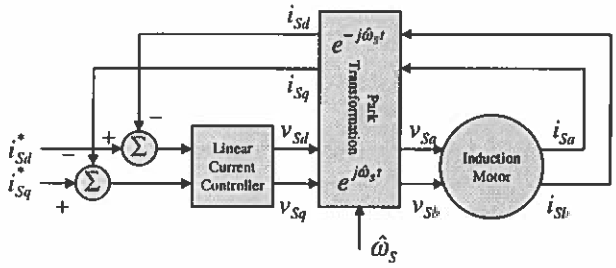

Figure 1. Pseudo current controller.

During *field orientation*, the coordinate system is aligned with the d-component rotor flux and hence $\psi_{Rd} = |\vec{\psi}_{Rdq}|$ and $\psi_{Rq} = 0$ with the coordinate system rotating at the *synchronous* frequency, ω_S . During steady-state stator current, equations (1) may be combined to eliminate the rotor resistance yielding,

$$\hat{\omega}^q_S = \frac{q}{L_\sigma |\vec{i}^*_{Sdq}|^2 + \frac{L_M}{L_R} i^*_{Sd} |\vec{\psi}_{Rdq}|^*} \tag{2}$$

where $q = v_{Sq} i^*_{Sd} - v_{Sd} i^*_{Sq}$ is the reactive power "consumed" by the machine ($*$ means impressed or reference value). Clearly, (2) is independent of the rotor resistance and may be used to estimate the synchronous frequency under the conditions of steady-state field orientation. This property is used as the basis of a MRAC system in which the rotor resistance is updated online. Also, the rotor velocity is not present in (2), precisely the property exploited later in the form of a *sensorless* torque controller, the subject of Section 4.

3. Model reference adaptive induction motor controller (MRAC)

An adaptive Indirect Field Orientated Controller (IFOC) is depicted in Figure 2. The IFOC synchronous frequency estimate, $\hat{\omega}^{IFOC}_S$, is determined using the *classic* slip frequency relationship,

$$\hat{\omega}^{IFOC}_S = \omega_R + \hat{\omega}^{IFOC}_{slip} \tag{3}$$

where the estimated slip frequency, $\hat{\omega}^{IFOC}_{slip}$, is determined from,

$$\omega^{IFOC}_{slip} = \hat{R}_R(t) \frac{L_M}{L_R} \frac{i^*_{Sq}}{|\vec{\psi}_{Rdq}|^*} \tag{4}$$

The rotor resistance estimate, $\hat{R}_R(t)$, used in (4) is defined as *slowly* time varying for the adaption process. An error, ε, is formed by comparing the estimate from (4) with the estimate

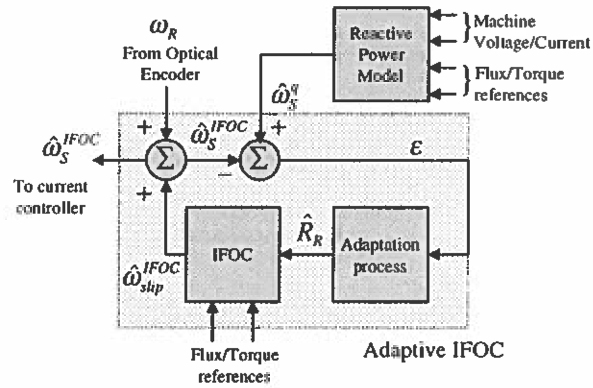

Figure 2. Schematic diagram of the MRAC. The machine and current controller is shown in Figure 1.

using (2), which is assumed to produce the correct slip frequency with the actual rotor resistance, R_R. The *open-loop* error equation may be written as,

$$\frac{d\varepsilon}{dt} = -\frac{1}{L_R} \frac{i^*_{Sq}}{i^*_{Sd}} \frac{d\hat{R}_R(t)}{dt} \quad \text{where} \quad \varepsilon = \hat{\omega}^q_S - \hat{\omega}^{IFOC}_S \tag{5}$$

The parameter update-law is chosen as,

$$\frac{d\hat{R}_R(t)}{dt} = \gamma \, i^*_{Sq} \, \varepsilon \tag{6}$$

where γ is a constant that affects the convergence rate of the controller. The closed-loop error dynamic is given by,

$$\frac{d\varepsilon}{dt} = -\frac{\gamma}{L_R} \frac{(i^*_{Sq})^2}{i^*_{Sd}} \varepsilon \tag{7}$$

During non-zero torque demands, the stability of (7) is assured since $i^*_{Sd} > 0$ for a non-zero flux magnitude. Adaptation is not possible during zero torque demand since $i^*_{Sq} = 0$. The resulting IFOC is not dependent on the rotor resistance. Clearly, the adaption rate is dependent on the torque current demand, i^*_{Sq}, which may reduce to low values around the operating point in applications that have low load torque, for example a simple inertia. This represents a fundamental limitation of all IFOC adaptive controllers since zero torque demand results in almost-zero slip-frequency and hence low rotor circuit currents. From the error dynamics, (7), it can be seen that the adaptation may start with a zero initial rotor resistance estimate. However performance is improved if an initial value close to the *cold* value determined during controller commissioning is used.

4. Direct synchronous frequency calculation controller (DωS)

Simulation studies based on the MRAC have shown that the reactive power model may be used as a method of achieving field-orientated control directly without mechanical velocity measurement. The controller shown here is similar in structure to a previously proposed reactive-power estimator (see [6]). The DωS controller uses the estimated synchronous frequency, $\hat{\omega}^q_S$, provided by the reactive-power model for the Park transformation (see Figure 3). As can be seen, the need for a mechanical velocity measurement is removed and, additionally, the reactive power model does not contain resistive parameters. A fundamental problem with the reactive power model, as exposed by Dalton [6], is that it ceases to converge during machine braking thus limiting its applications considerably. Here, a modification is proposed in which the transition into braking mode is detected and the model is *switched* into a second form capable of maintaining field orientation. Correct field orientation implies $\psi_{Rq} = 0$. Assuming the machine is *close* to correct field orientation, the q-component rotor flux may be used as an error which must be designed to converge to zero using the synchronous frequency estimate, $\hat{\omega}_S$, as a control input. *Near* field orientation implies $|\vec{\psi}_{Rdq}| \approx |\vec{\psi}_{Rdq}|^*$ and $d|\vec{\psi}_{Rdq}|/dt \approx 0$. If the stator currents are assumed to be in steady-state, the stator equations, (1) may be combined to yield an Ordinary Differential Equation (ODE) in the form,

$$\frac{d\psi_{Rq}}{dt} = -\hat{\omega}_S \frac{i^*_{Sq}}{i^*_{Sd}} \psi_{Rq} + F \tag{8}$$

where F is given by

$$F = -\frac{L_\sigma |\vec{i}^*_{Sdq}|^2 + \frac{L_M}{L_R} i^*_{Sd} |\vec{\psi}_{Rdq}|^*}{\frac{L_M}{L_R} i^*_{Sd}} \hat{\omega}_S + \frac{L_R}{L_M} \frac{q}{i^*_{Sd}} \tag{9}$$

$\hat{\omega}_S$ is chosen such that $F = 0$ in (9) yielding,

$$\hat{\omega}_S \bigg|_{F=0} = \frac{q}{L_\sigma |\vec{i}^*_{Sdq}|^2 + \frac{L_M}{L_R} i^*_{Sd} |\vec{\psi}_{Rdq}|^*} \tag{10}$$

which is the reactive power model given by equation (2). The closed-loop dynamics are then governed by the equation,

$$\frac{d\psi_{Rq}}{dt} = -\hat{\omega}_S \frac{i_{Sq^*}}{i^*_{Sd}} \psi_{Rq} \tag{11}$$

Equation (11) shows that $\psi_{Rq} \to 0$ providing the product $\hat{\omega}_S \, i_{Sq} > 0$, corresponding to the machine acting in motoring mode. In situations when the machine enters into braking, for example during velocity reversal of inertial loads, the convergence is not stable and field orientation is lost.

A second synchronous frequency estimate is proposed which satisfies the desired closed-loop dynamics during braking. The ODE this time is given by,

$$\frac{d\psi_{Rq}}{dt} = \hat{\omega}_S \frac{i^*_{Sq}}{i^*_{Sd}} \psi_{Rq} + F' \tag{12}$$

Again, choosing the synchronous frequency estimate such that $F' = 0$ yields stable field orientation when the product $\hat{\omega}_S \, i^*_{Sq} < 0$ corresponding to the machine in a braking mode. F' may be determined by substituting ψ_{Rq} and $d\psi_{Rq}/dt$ from the stator equations (1). The corresponding F' is given as,

$$F' = -2R_S \frac{L_R}{L_M} i^*_{Sq} - \frac{\frac{L_M}{L_R} i^*_{Sd} |\vec{\psi}_{Rdq}|^* + L_\sigma \left((i^*_{Sd})^2 - (i^*_{Sq})^2 \right)}{\frac{L_M}{L_R} i^*_{Sd}} \hat{\omega}_S + \frac{L_R}{L_M} \frac{q'}{i^*_{Sd}} \tag{13}$$

where $q' = v_{Sd} i_{Sq} + i_{Sd} v_{Sq}$ and $\hat{\omega}_S$ is given by,

$$\hat{\omega}_S \bigg|_{F'=0} = \frac{q' - 2R_S i^*_{Sq} \, i^*_{Sd}}{L_\sigma \left((i^*_{Sd})^2 - (i^*_{Sq})^2 \right) + \frac{L_M}{L_R} i^*_{Sd} |\vec{\psi}_{Rdq}|^*} \tag{14}$$

Application of the controllers requires the detection of motor or braking modes. Here, the stability criterion is assessed and used to determine which synchronous frequency estimate must be used. The detection may be summarized using,

$$\hat{\omega}_S = \begin{cases} \hat{\omega}_S \, i^*_{Sq} > 0 & \text{motor mode, eqn. (2)} \\ \hat{\omega}_S \, i^*_{Sq} < 0 & \text{brake mode, eqn. (14)} \end{cases} \tag{15}$$

A schematic of the controller is shown in Figure 3.

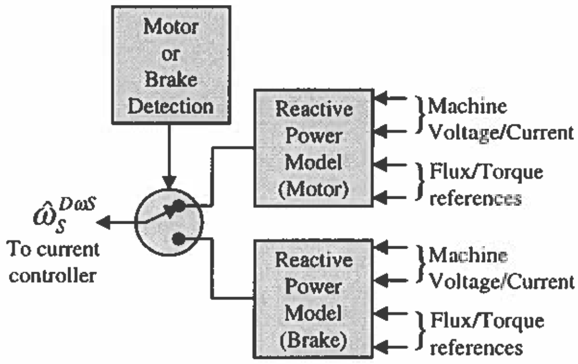

Figure 3. Schematic diagram of the DωS controller. The machine and current controller are not shown for clarity

5. Experimental results

The controllers described above have been implemented and tested on a 0.3kW, squirrel-cage laboratory machine in the LAE[1] laboratory at E.T.S. de Ingeniería-ICAI, Madrid. The experiments have been carried out using Matlab®, Simulink® and the Real-Time Workshop. The Simulink® models have been compiled and executed in both a Texas TMS320C30 and a Pentium 200MHz platforms. It has been shown how the Pentium outperforms the DSP not only in execution speed but also in the total acquisition cost of the system. The experimental setup is illustrated in Figure 4. The voltage source inverter (VSI) shown in Figure 4 controls the motor currents with simple PI controllers with a settling time of around 12ms. Other machine parameters are given in Appendix A. The experimental measurements are presented in a form which allows a comparison between a conventional IFOC, the MRAC and the new DωS controller.

The velocity responses of a conventional tuned IFOC, the MRAC and the DωS controllers are shown in Figure 5. The experiment is performed during low speed operation with low machine torque. The MRAC's rotor resistance estimate, $\hat{R}_R$, converges to the actual value, R_R in finite time. The rotor velocity overshoot of the MRAC shows the effect of a poor rotor resistance estimate during start-up. At

[1] Laboratorio de Accionamientos Eléctricos

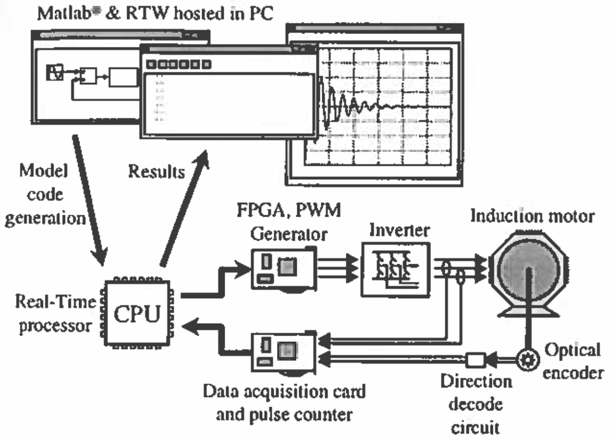

Figure 4. Experimental setup showing the Real-Time Workshop and the associated hardware interfaces.

the point where the machine velocity is reversed this overshoot no longer appears and the response becomes similar to the tuned IFOC. The DωS controller response is not dependent on the rotor resistance and matches the tuned IFOC both during start-up and reversal. The machine stator cur-

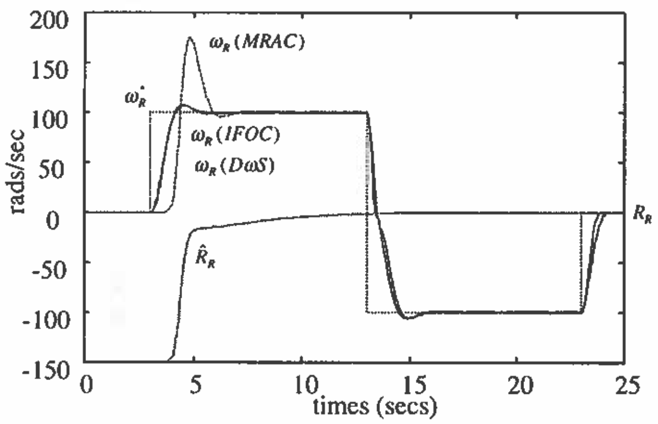

Figure 5. Experimental results showing the rotor velocity responses for a tuned IFOC, the MRAC and the DωS controllers. The rotor resistance estimate, $\hat{R}_R$, is shown converging to the actual machine value, R_R, in the case of the MRAC.

rents are shown in Figure 6. The MRAC quickly converges to the tuned IFOC corresponding to the tuning phase depicted in Figure 5. A de-tuned IFOC is also shown in Figure 6 in which the controller is operating with +40% error in the rotor resistance estimate.

6. Conclusions

Field-orientated control of the induction motor has been predominantly implemented using the rotor circuit model for correct stator current positioning. Here, MRAC applied to the rotor resistance estimation for an IFOC has been reviewed. In addition, an alternative field orientated controller with independence to the machine rotor resistance and rotor velocity has been presented. The reactive power is used on both occasions due to it's independence of machine resistive

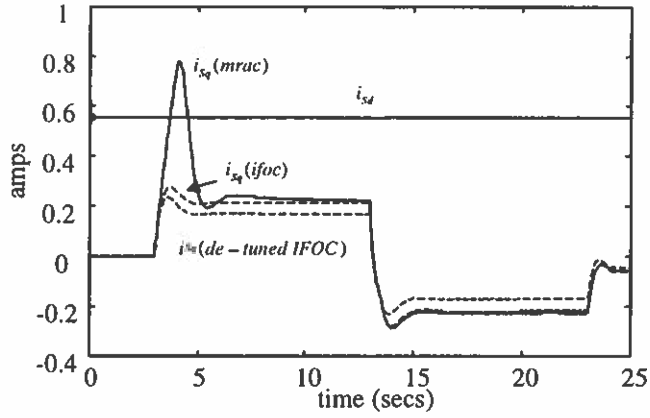

Figure 6. Machine stator currents for a tuned IFOC, a de-tuned IFOC and the MRAC.

parameters and mechanical velocity. Experimental results have demonstrated their performance in comparison with a tuned, classic IFOC.

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A. Nomenclature

A hat is used to indicate vector quantities, e.g. $\hat{x}$.
An asterisk is used to denote reference values, e.g. x^*

R_S	Stator resistance	48	Ohms
L_S	Stator inductance	1.1338	Henries
R_R	Rotor resistance	24.6	Ohms
L_R	Rotor inductance	1.1338	Henries
L_M	Magnetizing inductance	1.0282	Henries
P	Machine pole-pairs	2	
L_σ	Stator leakage inductance	$L_S\left(1-\frac{L_M^2}{L_S L_R}\right)$	