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Two simple and novel SISO controllers for induction motors based on adaptive passivity

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Abstract

The design of two single-input single-output (SISO) controllers for induction motors based on adaptive passivity is presented in this paper. The two controllers work together with a field orientation block. Because of the adaptive nature of the proposed controllers, the knowledge of the set motor–load parameters is not needed and robustness under variations of such parameters is guaranteed. Simple proportional controllers for the torque, rotor flux and stator current control loops are used, due to the control simplification given by the use of feedback passive equivalence. A new principle called the "Torque–Flux Control Principle" is also stated in this article, which considerably simplifies the controller design, diminishing the control efforts and avoiding also rotor flux estimation.

Keywords: Control of induction motors; Adaptive passivity; Adaptive control; Electric drive control; Nonlinear control of single-input single-output systems

1. Introduction

The design of suitable control algorithms for induction motors has been widely investigated during the last decade and at the beginning of this. Although the field orientation principle has been imposed as an obligatory part of the control schemes that guarantees high performance, the speed, torque and rotor flux controllers continue evolving. Since the beginning of the vector control of AC drives, seen as a viable replacement of the traditional DC drives, several techniques of linear control have been used, such as P, PI and PID regulators, and exact feedback linearization [1,2]. Due to their characteristics these techniques do not guarantee a suitable machine operation for all operation ranges and they do not consider the variations of the set motor–load parameters. In this sense, there are some applications where it is required to operate at several speeds between zero and the nominal, the motor can warm up diminishing the stator and rotor resistances, or perhaps the inertia of the load can vary (e.g. a reel being filled in the textile industry). Several nonlinear control techniques have been proposed in order to overcome the above problems.

Sliding mode techniques [3–5] guarantee robustness under variations of some parameters such as load torque and rotor resistance. Nevertheless this control technique presents the chattering effect and acoustic noise as main disadvantages. Artificial intelligence techniques using fuzzy logic, neuronal networks or their combinations, which can even include optimization algorithms based on genetic algorithms, have also been proposed by some authors [6,7]. The fuzzy logic can guarantee robustness under parameter variations and load disturbances, but it uses membership functions which require a large practical experience for their choice. On the other hand, the neuronal networks guarantee good results but they require a training process that can be off-line, on-line or a combination of these. Control techniques based on passivity have also been suggested [5,8] but they are not robust under variations of the set motor–load parameters.

All the previous techniques are based on rather complex control schemes that might involve a great deal of work in: the choice of the right parameters and functions (in the case of fuzzy control), the off-line training or the on-line parameter estimation at every instant of time (in the case of ANN), the

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state variable estimation (rotor flux estimation) and the design of the control strategy based on the estimations. For induction motor control it becomes necessary to use robust adaptive techniques, to avoid the parameter and rotor flux estimations to end up with simple controllers considering the model nonlinearities. Recently, in [9,10,14,15] two novel adaptive feedback passive equivalence techniques were proposed for SISO and MIMO systems. They consider the nonlinear model characteristics and they are adaptive in nature, guaranteeing robustness under all model parameter variations.

Based on the control techniques presented in [9,10] for SISO systems, two control strategies for induction motors are proposed in this paper, one with fixed adaptive gains and the other with time-varying adaptive gains. These strategies are suitably simplified by using the new principle called the "Torque–Flux Control Principle" which is also proposed in the paper. This principle simplifies the design of the torque and flux controller that work together with a field orientation block. By applying this principle to the design of the two controller given by [9,10], two control schemes without requiring parameter and variable estimations and using simple proportional gains for the speed, rotor flux and stator currents control loops are obtained. A MIMO version of the techniques proposed in this paper was developed in [16], having a larger number of adjustable parameters. The results proposed here are much simpler since both controllers present only two adjustable parameters using simple adaptive laws, guaranteeing results similar to those presented in [16] from the robustness viewpoint, under a wide range of motor–load parameter variations as well as under a wide range of proportional gain variations.

The Controller with Fixed Adaptive Gains (CFAG) and the Controller with Time-Varying Adaptive Gains (CTVAG) are preliminary results of a research project aimed to develop a beta product that will impact the induction motor control industry. The first strategy (CFAG) is simpler but a better transient behavior is expected when using the second one (CTVAG). Further experimental studies will determine which technique is better, depending perhaps on technical requirements, such as precision, transient response, operation range or on load characteristics. If the alpha product justifies marketing this approach, then movement onto beta site development should proceed.

2. Motor model and passive decomposition scheme

2.1. Motor model

An induction motor model obtained from the generalized electrical machine equations will be used for the controller design. Magnetic field linearly distributed through the air gap is assumed and the resulting model is applicable to a p-poles machine. Iron losses, saturation and hysteresis phenomena are neglected. The reference axes are fixed in arbitrary x–y coordinates that rotate at a generic speed ω_g. A generic coordinate system is adopted to allow appreciation of the advantages of the controller if it is expressed on diverse axes. The resulting equations are [11]

$$\begin{aligned}
\dot{i}_{sx} &= -\left(\frac{R_s}{\sigma L_s} + \frac{L_m^2 R_r}{\sigma L_s L_r^2}\right) i_{sx} + \omega_g i_{sy} \\
&\quad + \frac{L_m}{\sigma L_s L_r^2} R_r \psi_{rx} + \frac{L_m}{\sigma L_s L_r} \omega_r \psi_{ry} + \frac{1}{\sigma L_s} u_{sx} \\
\dot{i}_{sy} &= -\omega_g i_{sx} - \left(\frac{R_s}{\sigma L_s} + \frac{L_m^2 R_r}{\sigma L_s L_r^2}\right) i_{sy} \\
&\quad - \frac{L_m}{\sigma L_s L_r} \omega_r \psi_{rx} + \frac{L_m R_r}{\sigma L_s L_r^2} \psi_{ry} + \frac{1}{\sigma L_s} u_{sy} \\
\dot{\psi}_{rx} &= R_r \frac{L_m}{L_r} i_{sx} - \frac{R_r}{L_r} \psi_{rx} + (\omega_g - \omega_r) \psi_{ry} \\
\dot{\psi}_{ry} &= R_r \frac{L_m}{L_r} i_{sy} - (\omega_g - \omega_r) \psi_{rx} - \frac{R_r}{L_r} \psi_{ry} \\
\dot{\omega}_r &= (T_{em} - T_c) \frac{1}{J} - \frac{B_p}{J} \omega_r
\end{aligned} \tag{1}$$

where i_{sx}, i_{sy} are the stator currents, ψ_{rx}, ψ_{ry} are the rotor fluxes, ω_r is the rotor speed and u_{sx}, u_{sy} are the stator voltages, considered as control inputs. L_m, L_s, L_r are the mutual, stator and rotor inductances respectively, R_s, R_r are the stator and rotor resistances, J is the rotor inertia, T_{em} is the electromagnetic torque produced by the motor, T_c is the load torque and B_p is the mechanical viscous damping coefficient. We also define

$\sigma = 1 - \frac{L_m^2}{L_s L_r}$, leakage or coupling factor.

$R_s' = R_s + \frac{L_m^2}{L_r^2} R_r$, stator transient resistance.

σL_s, stator transient inductance.

The electromagnetic torque is given by

$$\begin{aligned}
T_{em} &= \frac{3}{2}\frac{p}{2}\frac{L_m}{L_r}(\bar{\psi}_{rg} \times \bar{i}_{sg}) = \frac{3}{2}\frac{p}{2}\frac{L_m}{L_r} |\bar{\psi}_{rg}| \, |\bar{i}_{sg}| \sin\alpha \\
&= \frac{3}{2}\frac{p}{2}\frac{L_m}{L_r}(\psi_{rx} i_{sy} - \psi_{ry} i_{sx}).
\end{aligned} \tag{2}$$

The representation of the vectors, together with the nomenclature used in the study, is shown in Fig. 1.

2.2. Passive decomposition

A passive decomposition is used in [5,8], valid also for the generalized machine. It considers the induction motor as the interconnection of the electrical and the mechanical systems by means of a negative feedback, as seen in Fig. 2. Starting from the previous statement the induction motor control can be carried out by controlling the electrical system, considering the mechanical system as a passive disturbance of $(T_{em} - T_c)$ on the variable ω_r.

Starting from the previous idea the following assumption is stated.

Assumption 1. There exists an input vector voltage $u_s^* = [u_{sx}^* \;\; u_{sy}^*]^T$ which guarantees that the desired currents $i_s^* =$

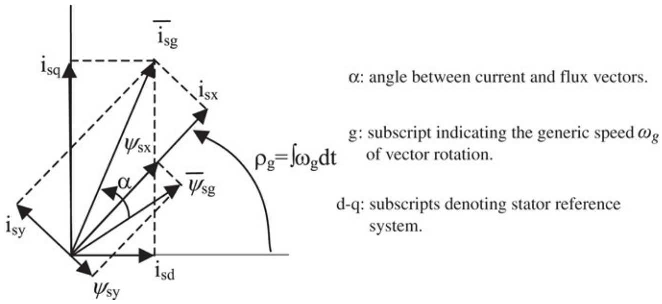

Fig. 1. Angle between stator current and rotor flux vectors.

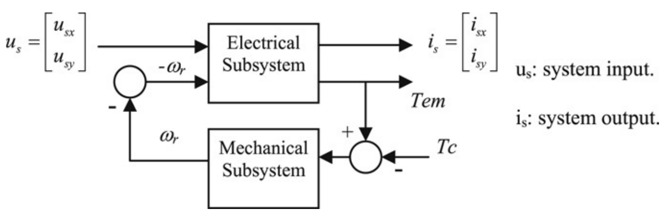

Fig. 2. Electrical system to be controlled with a passive disturbance.

$[i_{sx}^* \ i_{sy}^*]^{\mathrm{T}}$ and fluxes $\psi_r^* = [\psi_{rx}^* \ \psi_{ry}^*]^{\mathrm{T}}$ can be established in the induction motor.

Then we can define the variable errors as the difference between the desired variable and the real one. For example, $e_{i_{sx}} = i_{sx}^* - i_{sx}$, where i_{sx}^* is the reference current component and i_{sx} the current component obtained through measurements. Then we can write the electrical system in terms of the variable errors (deviation variables) as follows:

$$\begin{aligned}\dot{y} &= \begin{bmatrix} \dot{e}_{i_{sx}} \\ \dot{e}_{i_{sy}} \end{bmatrix} \\ &= \begin{bmatrix} -\dfrac{R_s'}{\sigma L_s} e_{i_{sx}} + \omega_g e_{i_{sy}} + \dfrac{L_m R_r}{\sigma L_s L_r^2} e_{\psi_{rx}} + \dfrac{L_m}{\sigma L_s L_r} \omega_r e_{\psi_{ry}} \\ -\omega_g e_{i_{sx}} - \dfrac{R_s'}{\sigma L_s} e_{i_{sy}} - \dfrac{L_m}{\sigma L_s L_r} \omega_r e_{\psi_{rx}} + \dfrac{L_m R_r}{\sigma L_s L_r^2} e_{\psi_{ry}} \end{bmatrix} \\ &\quad + \begin{bmatrix} \dfrac{1}{\sigma L_s} & 0 \\ 0 & \dfrac{1}{\sigma L_s} \end{bmatrix} u \\ \dot{z} &= \begin{bmatrix} \dot{e}_{\psi_{rx}} \\ \dot{e}_{\psi_{ry}} \end{bmatrix} = \begin{bmatrix} R_r \dfrac{L_m}{L_r} e_{i_{sx}} - \dfrac{R_r}{L_r} e_{\psi_{rx}} + (\omega_g - \omega_r) e_{\psi_{ry}} \\ R_r \dfrac{L_m}{L_r} e_{i_{sy}} - (\omega_g - \omega_r) e_{\psi_{rx}} - \dfrac{R_r}{L_r} e_{\psi_{ry}} \end{bmatrix}\end{aligned} \tag{3}$$

where $y = i_s^* - i_s = [(i_{sx}^* - i_{sx}) \ (i_{sy}^* - i_{sy})]^{\mathrm{T}} = [e_{i_{sx}} \ e_{i_{sy}}]^{\mathrm{T}} = [y_1 \ y_2]^{\mathrm{T}} \in \Re^2$, $z = \psi_r^* - \psi_r = [(\psi_{rx}^* - \psi_{rx}) \ (\psi_{ry}^* - \psi_{ry})]^{\mathrm{T}} = [e_{\psi_{rx}} \ e_{\psi_{ry}}]^{\mathrm{T}} = [z_1 \ z_2]^{\mathrm{T}} \in \Re^2$ and

$$\begin{aligned} u &= u_s^* - u_s = [(u_{sx}^* - u_{sx}) \ (u_{sy}^* - u_{sy})]^{\mathrm{T}} \\ &= [e_{u_{sx}} \ e_{u_{sy}}]^{\mathrm{T}} = [u_1 \ u_2]^{\mathrm{T}} \in \Re^2. \end{aligned}$$

In order to apply the techniques proposed in [9,10] it is necessary to express the electrical system as two SISO systems. Then the MIMO system (3) will be divided into two SISO subsystems defined as follows:

$$\left.\begin{aligned} \dot{y}_1 &= A_1'(y_i, z) + B_1'(y_i, z) u_1 \\ \dot{z} &= f_0'(z) + P_1'(y_i, z) y_1 \end{aligned}\right\} \quad \text{for } i = 1, 2 \tag{4}$$

$$\left.\begin{aligned} \dot{y}_2 &= A_2'(y_i, z) + B_2'(y_i, z) u_2 \\ \dot{z} &= f_0'(z) + P_2'(y_i, z) y_2 \end{aligned}\right\} \quad \text{for } i = 1, 2 \tag{5}$$

with y_1, y_2 scalar outputs, u_1, u_2 scalar inputs, A_1', A_2', B_1', B_2', scalar functions, $f_0' \in \Re^2$, P_1', $P_2' \in \Re^2$ and $z \in \Re^2$. For our particular case we will have:

Subsystem 1:

$$\begin{aligned}\dot{y}_1 &= \dot{e}_{i_{sx}} \\ &= \left[-\frac{R_s'}{\sigma L_s} e_{i_{sx}} + \omega_g e_{i_{sy}} + \frac{L_m R_r}{\sigma L_s L_r^2} e_{\psi_{rx}} + \frac{L_m}{\sigma L_s L_r} \omega_r e_{\psi_{ry}} \right] \\ &\quad + \left[\frac{1}{\sigma L_s} \right] e_{u_{sx}} \\ \dot{z} &= \begin{bmatrix} \dot{e}_{\psi_{rx}} \\ \dot{e}_{\psi_{ry}} \end{bmatrix} = \begin{bmatrix} -\dfrac{R_r}{L_r} e_{\psi_{rx}} - (\omega_g - \omega_r) e_{\psi_{ry}} \\ (\omega_g - \omega_r) e_{\psi_{rx}} - \dfrac{R_r}{L_r} e_{\psi_{ry}} \end{bmatrix} \\ &\quad + \begin{bmatrix} \dfrac{L_m}{T_r} \\ \dfrac{L_m}{T_r} \dfrac{\dot{e}_{i_{sy}}}{\dot{e}_{i_{sx}}} \end{bmatrix} e_{i_{sx}}. \end{aligned} \tag{6}$$

Subsystem 2:

$$\begin{aligned}\dot{y}_2 &= \dot{e}_{i_{sy}} \\ &= \left[-\omega_g e_{i_{sx}} - \frac{R_s'}{\sigma L_s} e_{i_{sy}} - \frac{L_m}{\sigma L_s L_r} \omega_r e_{\psi_{rx}} + \frac{L_m R_r}{\sigma L_s L_r^2} e_{\psi_{ry}} \right] \\ &\quad + \left[\frac{1}{\sigma L_s} \right] e_{u_{sy}} \\ \dot{z} &= \begin{bmatrix} \dot{e}_{\psi_{rx}} \\ \dot{e}_{\psi_{ry}} \end{bmatrix} = \begin{bmatrix} -\dfrac{R_r}{L_r} e_{\psi_{rx}} - (\omega_g - \omega_r) e_{\psi_{ry}} \\ (\omega_g - \omega_r) e_{\psi_{rx}} - \dfrac{R_r}{L_r} e_{\psi_{ry}} \end{bmatrix} \\ &\quad + \begin{bmatrix} \dfrac{L_m}{T_r} \dfrac{\dot{e}_{i_{sx}}}{\dot{e}_{i_{sy}}} \\ \dfrac{L_m}{T_r} \end{bmatrix} e_{i_{sy}} \end{aligned} \tag{7}$$

with $T_r = L_r / R_r$.

3. Controller design through passive equivalence

3.1. Model adjustments

In order to apply the theorems of [9,10], both subsystems must be parameterized in the following normal form [13] with linear explicit parametric dependence:

$$\left.\begin{aligned}\dot{y}_i &= \Lambda_{ai}^{\mathrm{T}} A_i(y_i, z) + \Lambda_{bi} B_i(y_i, z)\mu_i \\ \dot{z} &= \Lambda_0 f_0(z) + \Lambda_{pi} P_i(y_i, z) y_i\end{aligned}\right\} \quad i = 1, 2 \tag{8}$$

with $z \in \Re^2$, $y_i \in \Re$, $u_i \in \Re$, $A_i(y_i, z) \in \Re^4$, $B_i(y_i, z) \in \Re$, $f_0 \in \Re^2$, $P_i(y_i, z) \in \Re^2$. The parameters $\Lambda_{ai} \in \Re^4$, $\Lambda_{bi} \in \Re$, $\Lambda_0 \in \Re^{2\times 2}$, $\Lambda_{pi} \in \Re^{2\times 2}$. The term function $\dot{z} = f_0(z)$, in both subsystems, is the so called zero dynamics [17,18]. Now it is necessary to check whether both subsystems are locally weakly minimum phase by finding a positive definite differentiable function $W_0(z)$ satisfying $\left[\frac{\partial W_0(z)}{\partial z}\right]^{\mathrm{T}} \Lambda_0 f_0(z) \le 0, \forall \Lambda_0$ [13]. In both cases we define the function $W_0(z) = \frac{1}{2}(e_{\psi_{rx}}^2 + e_{\psi_{ry}}^2)$; then $\frac{\partial W_0(z)}{\partial z} = [e_{\psi_{rx}} e_{\psi_{ry}}]^{\mathrm{T}}$. Since $\Lambda_0 f_0(z) = \begin{bmatrix} -\frac{R_r}{L_r} e_{\psi_{rx}} - (\omega_g - \omega_r) e_{\psi_{ry}} \\ (\omega_g - \omega_r) e_{\psi_{rx}} - \frac{R_r}{L_r} e_{\psi_{ry}} \end{bmatrix}$ it can be checked that

$$\begin{aligned}
&\left[\frac{\partial W_0(z)}{\partial z}\right]^{\mathrm{T}} \Lambda_0 f_0(z) \\
&= \begin{bmatrix} e_{\psi_{rx}} & e_{\psi_{ry}} \end{bmatrix} \begin{bmatrix} -\frac{R_r}{L_r} e_{\psi_{rx}} - (\omega_g - \omega_r) e_{\psi_{ry}} \\ (\omega_g - \omega_r) e_{\psi_{rx}} - \frac{R_r}{L_r} e_{\psi_{ry}} \end{bmatrix} \\
&= -\frac{R_r}{L_r} e_{\psi_{rx}}^2 - (\omega_g - \omega_r) e_{\psi_{rx}} e_{\psi_{ry}} \\
&\quad + (\omega_g - \omega_r) e_{\psi_{rx}} e_{\psi_{ry}} - \frac{R_r}{L_r} e_{\psi_{ry}}^2 \\
&= -\frac{R_r}{L_r}\left(e_{\psi_{rx}}^2 + e_{\psi_{ry}}^2\right) \le 0
\end{aligned} \tag{9}$$

leading us to conclude that both subsystems are locally weakly minimum phase. Now we proceed to express them in the normal form defined by (8).

For the Subsystem 1 we have

$$\begin{aligned}
\Lambda_{a1} &= \begin{bmatrix} -\frac{R_s'}{\sigma L_s} & 1 & \frac{L_m R_r}{\sigma L_s L_r^2} & \frac{L_m}{\sigma L_s L_r} \end{bmatrix}^{\mathrm{T}}, \\
A_1(y_i, z) &= \begin{bmatrix} e_{i_{sx}} \\ \omega_g e_{i_{sy}} \\ e_{\psi_{rx}} \\ \omega_r e_{\psi_{ry}} \end{bmatrix}, \\
\Lambda_{b1} &= \frac{1}{\sigma L_s}, \qquad B_1(y_i, z) = 1, \\
\Lambda_{p1} &= \begin{bmatrix} \frac{L_m}{T_r} & 0 \\ 0 & \frac{L_m}{T_r} \end{bmatrix}, \qquad y_1 = e_{i_{sx}}, \\
P_1(y_i, z) &= \begin{bmatrix} P_{11} \\ P_{21} \end{bmatrix} = \begin{bmatrix} 1 \\ \frac{\dot{e}_{i_{sy}}}{\dot{e}_{i_{sx}}} \end{bmatrix}, \qquad u_1 = e_{u_{sx}}.
\end{aligned} \tag{10}$$

For the Subsystem 2 we can write

$$\begin{aligned}
\Lambda_{a2} &= \begin{bmatrix} -\frac{R_s'}{\sigma L_s} & -1 & \frac{L_m R_r}{\sigma L_s L_r^2} & -\frac{L_m}{\sigma L_s L_r} \end{bmatrix}^{\mathrm{T}}, \\
A_2(y_i, z) &= \begin{bmatrix} e_{i_{sy}} \\ \omega_g e_{i_{sx}} \\ e_{\psi_{ry}} \\ \omega_r e_{\psi_{rx}} \end{bmatrix}, \\
\Lambda_{b2} &= \frac{1}{\sigma L_s}, \qquad B_2(y_i, z) = 1, \\
\Lambda_{p2} &= \begin{bmatrix} \frac{L_m}{T_r} & 0 \\ 0 & \frac{L_m}{T_r} \end{bmatrix}, \qquad y_2 = e_{i_{sy}}, \\
P_2(y_i, z) &= \begin{bmatrix} P_{12} \\ P_{22} \end{bmatrix} = \begin{bmatrix} \frac{\dot{e}_{i_{sx}}}{\dot{e}_{i_{sy}}} \\ 1 \end{bmatrix}, \qquad u_2 = e_{u_{sy}}.
\end{aligned} \tag{11}$$

For both subsystems (10) and (11) we have

$$\Lambda_0 f_0(z) = \begin{bmatrix} -\frac{R_r}{L_r} e_{\psi_{rx}} - (\omega_g - \omega_r) e_{\psi_{ry}} \\ (\omega_g - \omega_r) e_{\psi_{rx}} - \frac{R_r}{L_r} e_{\psi_{ry}} \end{bmatrix}.$$

3.2. *Proposed controller with fixed adaptive gains (CFAG)*

From [9] the subsystems (10) and (11) must be locally weakly minimum phase and matrices $B_i(y_i, z)$ for $i = 1, 2$ should be invertible. The first condition was already demonstrated in (9). From Eqs. (10) and (11) it can be readily checked that matrices $B_i(y_i, z) = 1$ for $i = 1, 2$, so they are in fact invertible.

Then, according to [9] there exists an adaptive controller of the form

$$\begin{aligned}
u_i(y_i, z, \theta_{hi}) &= \frac{1}{B_i(y_i, z)} \\
&\times \Big[\theta_{1i}^{\mathrm{T}}(t) A_i(y_i, z) - \theta_{2i}(t) P_{1i}(y_i, z) \frac{\partial W_0(z)}{\partial z_1} \\
&- \theta_{3i}(t) P_{2i}(y_i, z) \frac{\partial W_0(z)}{\partial z_2} + \theta_{4i}(t)\mu_i\Big] \quad \text{for } i = 1, 2
\end{aligned} \tag{12}$$

with $z \in \Re^2$, $y_i \in \Re$, $u_i \in \Re$, $\mu_i \in \Re$, $A_i(y_i, z) \in \Re$, $B_i(y_i, z) \in \Re$, $f_0 \in \Re^2$, $P_{1i}(y_i, z) \in \Re^2$, $P_{2i}(y_i, z) \in \Re^2$ and $\theta_{hi}(t)$ for $i = 1, 2$ and $h = 1, \ldots, 4$ adjustable parameters with $\theta_{1i}(t) \in \Re^4$, and $\theta_{2i}(t), \theta_{3i}(t), \theta_{4i}(t) \in \Re$ updated with the adaptive laws

$$\left.\begin{aligned}
\dot{\theta}_{1i}(t) &= -\mathrm{sign}(\Lambda_{bi}) A_i(y_i, z) y_i \\
\dot{\theta}_{2i}(t) &= -\mathrm{sign}(\Lambda_{bi}) y_i P_{1i}(y_i, z) \left(\frac{\partial W_0(z)}{\partial z_1}\right) \\
\dot{\theta}_{3i}(t) &= -\mathrm{sign}(\Lambda_{bi}) y_i P_{2i}(y_i, z) \left(\frac{\partial W_0(z)}{\partial z_2}\right) \\
\dot{\theta}_{4i}(t) &= -\mathrm{sign}(\Lambda_{bi}) y_i \mu_i
\end{aligned}\right\} \quad i = 1, 2 \tag{13}$$

that applied separately to each subsystem (8) make them locally feedback equivalent to a C^2-passive subsystem from the new input μ_i to the output y_i [9]. The parameters $\Lambda_{ai} \in \Re$, $\Lambda_{bi} \in \Re$, $\Lambda_0 \in \Re$, $\Lambda_{pi} \in \Re^{2\times 2}$ for $i = 1, 2$ represent constant but unknown parameters from a bounded compact set Ω. μ_1, μ_2 signals will be chosen as in (27).

3.3. Proposed controller with time-varying adaptive gains (CTVAG)

Another adaptive controller with time-varying gains [10] can also be proposed for this case. This controller denoted as **CTVAG** has the following form:

$$u_i(y_i, z, \theta_{hi}) = \frac{1}{B_i(y_i, z)}\left[\theta_{1i}^{\mathrm{T}}(t)A_i(y_i, z) - \theta_{2i}(t)P_{1i}(y_i, z)\frac{\partial W_0(z)}{\partial z_1} - \theta_{3i}(t)P_{2i}(y_i, z)\frac{\partial W_0(z)}{\partial z_2} + \theta_{4i}(t)\mu_i\right] \quad \text{for } i = 1, 2 \tag{14}$$

with $z \in \Re^2, y_i \in \Re, u_i \in \Re, \mu_i \in \Re$, $A_i(y_i, z) \in \Re$, $B_i(y_i, z) \in \Re$, $f_0 \in \Re^2, P_{1i}(y_i, z) \in \Re^2$, $P_{2i}(y_i, z) \in \Re^2$, for $i = 1, 2$ and $\theta_{hi}(t)$ for $i = 1, 2$ and $h = 1, \ldots, 4$, are adjustable parameters with $\theta_{1i}(t) \in \Re^4$ and $\theta_{2i}(t), \theta_{3i}(t), \theta_{4i}(t) \in \Re$, updated with the adaptive laws

$$\left.\begin{aligned}
\dot\theta_{1i}(t) &= -\mathrm{sign}(\Lambda_{bi})\frac{\gamma_{1i}^{-1}(t)}{\sqrt{1+\frac{1}{\gamma_i(t)^{\mathrm{T}}\gamma_i(t)}}} \times A_i(y_i, z)y_i\\
\dot\theta_{2i}(t) &= -\mathrm{sign}(\Lambda_{bi})\frac{\gamma_{2i}^{-1}(t)}{\sqrt{1+\frac{1}{\gamma_i(t)^{\mathrm{T}}\gamma_i(t)}}} \times P_{1i}(y_i, z)\left(\frac{\partial W_0(z)}{\partial z_1}\right)y_i\\
\dot\theta_{3i}(t) &= -\mathrm{sign}(\Lambda_{bi})\frac{\gamma_{3i}^{-1}(t)}{\sqrt{1+\frac{1}{\gamma_i(t)^{\mathrm{T}}\gamma_i(t)}}} \times P_{2i}(y_i, z)\left(\frac{\partial W_0(z)}{\partial z_2}\right)y_i\\
\dot\theta_{4i}(t) &= -\mathrm{sign}(\Lambda_{bi})\frac{\gamma_{4i}^{-1}(t)}{\sqrt{1+\frac{1}{\gamma_i(t)^{\mathrm{T}}\gamma_i(t)}}}\mu_i y_i
\end{aligned}\right\} \quad i = 1, 2 \tag{15}$$

where $\gamma_{1i}(t) \in \Re^{4\times 4}$ and $\gamma_{2i}(t), \gamma_{3i}(t), \gamma_{4i}(t) \in \Re$ are time-varying adaptive gains defined by

$$\left.\begin{aligned}
\dot\gamma_{1i}(t) &= -\left[\gamma_{1i}A_i(y_i, z)A_i^{\mathrm{T}}(y_i, z)\gamma_{1i}\right]\\
\dot\gamma_{2i}(t) &= \left[\gamma_{2i}(t)P_{1i}(y_i, z)\left(\frac{\partial W_0(z)}{\partial z_1}\right)\right]^2\\
\dot\gamma_{3i}(t) &= \left[\gamma_{3i}(t)P_{2i}(y_i, z)\left(\frac{\partial W_0(z)}{\partial z_2}\right)\right]^2\\
\dot\gamma_{4i}(t) &= [\gamma_{4i}(t)\mu_i]^2
\end{aligned}\right\} \quad i = 1, 2 \tag{16}$$

with $\gamma_i(t) = \left[\mathrm{Trace}\{\gamma_{1i}(t)\} \quad \gamma_{2i}(t) \quad \gamma_{3i}(t) \quad \gamma_{4i}(t)\right] \in \Re^4$. This controller can be applied to each separate subsystem (8) and according to [10] this will convert each of them into C^2-passive equivalent systems. The parameters $\Lambda_{ai} \in \Re$, $\Lambda_{bi} \in \Re$, $\Lambda_{0i} \in \Re$, $\Lambda_{pi} \in \Re^{2\times 2}$ represent constant but unknown parameters from a bounded compact set Ω. μ_1, μ_2 signals will be chosen as in (27).

By applying the **CFAG** given by Eqs. (12) and (13) and **CTVAG** given by Eqs. (14)–(16), we will obtain two control schemes that do not need the knowledge of any motor–load parameters. However it is necessary to know the error of the rotor flux components (z_1 and z_2), i.e. rotor flux estimation is needed. We will show in Section 4 that this will be not necessary.

4. Principle of Torque–Flux Control

4.1. Basis and statement

From now on we will work in a control scheme consisting of two kinds of blocks: the adaptive passive controllers, to control the stator current components and rotor flow (Fig. 3(a) and (b)), and the field orientation block (Fig. 3(c)). As can be seen in Fig. 3(a), the block will have an external closed-loop speed control feeding the internal closed-loop i_{sy} current control.

In Fig. 3(b) it can be seen there is a block with an external open-loop ψ_{rx} control feeding the internal closed-loop i_{sx} current control. The second block is $e^{j\rho_g}$, represented in Fig. 3(c), which transforms from a stationary to a rotating coordinate system. This is a very important control block employed to guarantee an adequate field orientation such that the maximum output torque is obtained.

Once this point is made clear and since the controller works together with the field orientation block the following assumption can be made.

Assumption 2. It is assumed that the coordinate transformation block is equal to $e^{j\rho_g}$, guaranteeing an adequate field orientation. Then we can write

$$\dot\rho_g = \omega_g = \omega_r + \frac{L_m}{T_r}\frac{i_{sy}}{\psi_{rx}}. \tag{17}$$

The previous assumption implies several consequences that are mentioned next.

Consequences:

1. From the fourth part of Eq. (1) we can write

$$\begin{aligned}
&\dot\psi_{ry} = R_r\frac{L_m}{L_r}i_{sy} - (\omega_g - \omega_r)\psi_{rx} - \frac{R_r}{L_r}\psi_{ry},\\
&\text{with } \omega_g = \omega_r + \frac{L_m}{T_r}\frac{i_{sy}}{\psi_{rx}}\\
&\dot\psi_{ry} = R_r\frac{L_m}{L_r}i_{sy} - \left[\left(\omega_r + \frac{L_m}{T_r}\frac{i_{sy}}{\psi_{rx}}\right) - \omega_r\right]\psi_{rx} - \frac{R_r}{L_r}\psi_{ry}\\
&\dot\psi_{ry} + \frac{R_r}{L_r}\psi_{ry} = 0 \Rightarrow \lim_{t\to\infty}\psi_{ry} = 0.
\end{aligned} \tag{18}$$

In practice the quadrature flux component will be practically zero at $t = 5T_r$, because ψ_{ry} is given by a first-order differential equation.

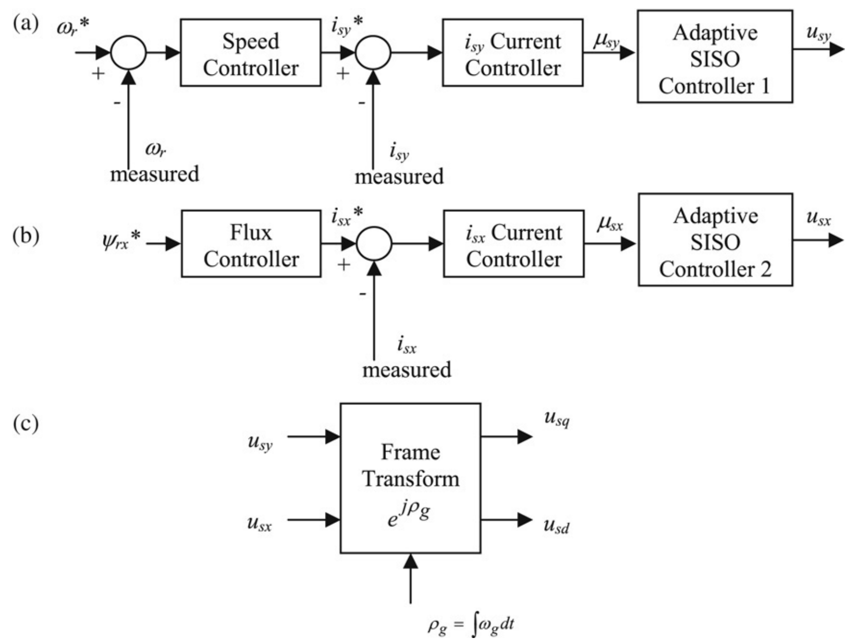

Fig. 3. Control blocks involved in the control scheme.

2. From Eq. (1) and considering the previous consequence given by (18) the following model for the electrical system is obtained:

$$
\begin{aligned}
\dot{i}_{sx} &= -\left(\frac{R_s}{\sigma L_s} + \frac{L_m^2 R_r}{\sigma L_s L_r^2}\right) i_{sx} + \omega_g i_{sy} + \frac{L_m}{\sigma L_s L_r^2} R_r \psi_{rx} \\
&\quad + \frac{L_m}{\sigma L_s L_r} \omega_r \psi_{ry} + \frac{1}{\sigma L_s} u_{sx} \\
\dot{i}_{sy} &= -\omega_g i_{sx} - \left(\frac{R_s}{\sigma L_s} + \frac{L_m^2 R_r}{\sigma L_s L_r^2}\right) i_{sy} - \frac{L_m}{\sigma L_s L_r} \omega_r \psi_{rx} \\
&\quad + \frac{L_m R_r}{\sigma L_s L_r^2} \psi_{ry} + \frac{1}{\sigma L_s} u_{sy} \\
\dot{\psi}_{rx} &= R_r \frac{L_m}{L_r} i_{sx} - \frac{R_r}{L_r} \psi_{rx}.
\end{aligned}
\tag{19}
$$

3. From third part of equation (19) it can be concluded that ψ_{rx} is a function of de i_{sx}. In order to emulate an independent field DC motor, the objective will be to maintain the rotor flux constant.

4. From Eq. (2) we have $T_{em} = \frac{3}{2}\frac{p}{2}\frac{L_m}{L_r}(\psi_{rx} i_{sy} - \psi_{ry} i_{sx})$ and due to Consequence 1 given by Eq. (18), the electromagnetic torque can be expressed as $T_{em} = \frac{3}{2}\frac{p}{2}\frac{L_m}{L_r}\psi_{rx} i_{sy}$ where it can be seen that T_{em} is a function of i_{sy}.

5. The control of the rotor flux $\bar{\psi}_r = \psi_{rx}$ is carried out through the stator current direct component; it is desirable to keep it fixed. The control of torque T_{em} is carried out through the stator current quadrature component. All the above is obtained by using the coordinate transformation block and the corresponding field orientation.

Based on above assumption and the five consequences derived from it, the following control principle is established.

The Principle of Torque–Flux Control

When designing controllers for an alternating current motor based on a model of the Generalized Electrical Machine, working in a control scheme with the coordinate transformation block $e^{j\rho_g}$ *(the Field Oriented Scheme) to transform from a stationary to a rotating coordinate system, the control of the torque and flux can be done by controlling the stator currents. So, it is useless to make efforts to control the rotor flux or rotor current components. The controller still guarantees a suitable torque and flux control and it is possible to discard all the terms related to the rotor current or rotor flux components.*

By applying this principle not only is the controller design considerably simplified but also the rotor current or rotor flux estimations are avoided. It is important to emphasize that sometimes, even if the field orientation effect is used, the consequences mentioned above are not fulfilled due to a bad orientation, e.g. due to a bad estimation of parameter L_m/T_r. Nevertheless, experimental results show that the validity of the Principle still holds in those cases. The controller will take care of an effective control of the components of the stator current whether the field orientation guarantees or not that the rotor flux component errors tend to zero.

4.2. Application to the controller CFAG

Considering the "Principle of Torque–Flux Control", we will use the theorem stated in [9]. For the **CFAG** we will choose adaptive laws such that the storage functions V_{1i} for $i = 1, 2$ satisfy the following condition:

$$
\left.
\begin{aligned}
\dot{V}_{1i} &= \left(\frac{\partial W_o(z)}{\partial z}\right) \Lambda_o f_o(z) + \left\{ \sum_{j=1}^{n-1} \left[\Phi_{\theta_{(j+1)i}} P_{ji}(y_i, z) \right.\right. \\
&\qquad \left.\left. \times \left(\frac{\partial W_o(z)}{\partial z_j}\right) y_i \Lambda_{bi}^* + \Phi_{\theta_{(j+1)i}} \dot{\Phi}_{\theta(j+1)i} \left|\Lambda_{bi}^*\right| \right] \right\} \\
&\quad + \left\{ \Phi_{\theta_{1i}} A_i^{\mathrm{T}}(y_i, z) y_i \Lambda_{bi}^* + \Phi_{\theta_{1i}} \dot{\Phi}_{\theta_{1i}} \left|\Lambda_{bi}^*\right| \right\} \\
&\quad + \left\{ \Phi_{\theta_{(n+1)i}} \mu_i y_i \Lambda_{bi}^* + \Phi_{\theta_{(n+1)i}} \dot{\Phi}_{\theta_{(n+1)i}} \left|\Lambda_{bi}^*\right| \right\} \le 0
\end{aligned}
\right\}
$$

for $i = 1, 2$. (20)

If the rotor flux components are eliminated there results $\left(\frac{\partial W_o(z)}{\partial z}\right) = z = 0$, and the corresponding factors will vanish without needing adaptive laws. Then $\theta_{(j+1)i} = \dot{\theta}_{(j+1)i} = 0$, $j = 1, 2$, can be imposed. So it is verified from (20) that the controller will still make the system (1) equivalent to a C^2-passive system.

From Eqs. (10) and (11) it can be seen that

$$\begin{aligned} &A_1(y_1, z) = \begin{bmatrix} y_1 & \omega_g y_2 & z_1 & \omega_r z_2 \end{bmatrix}^{\mathrm{T}} \\ &\left(\frac{\partial W_0(z)}{\partial z}\right) = z = \begin{bmatrix} z_1 & z_2 \end{bmatrix}^{\mathrm{T}} \\ &B(y_1, z) = 1 \end{aligned} \tag{21}$$

and

$$\begin{aligned} &A_2(y_2, z) = \begin{bmatrix} y_2 & \omega_g y_1 & z_2 & \omega_r z_1 \end{bmatrix}^{\mathrm{T}} \\ &\left(\frac{\partial W_0(z)}{\partial z}\right) = z \\ &B(y_2, z) = 1. \end{aligned} \tag{22}$$

Now if we do not consider (discard) $z = [z_1 \; z_2]^{\mathrm{T}}$ variables, by making $z = 0$ in Eqs. (12) and (13) the following simplified controllers are obtained for Subsystems 1 and 2:
Subsystem 1:

$$\left.\begin{aligned} &u_1(y_1, z, \theta_h) = \theta_1^{\mathrm{T}} \begin{bmatrix} y_1 \\ \omega_g y_2 \end{bmatrix} + \theta_4 \mu_1 \\ &\dot{\theta}_1 = -y_1 \begin{bmatrix} y_1 \\ \omega_g y_2 \end{bmatrix} \\ &\dot{\theta}_4 = -y_1 \mu_1 \end{aligned}\right\} \quad h = 1, 4. \tag{23}$$

Subsystem 2:

$$\left.\begin{aligned} &u_2(y_2, z, \theta_h) = \theta_1^{\mathrm{T}} \begin{bmatrix} y_2 \\ \omega_g y_1 \end{bmatrix} + \theta_4 \mu_2 \\ &\dot{\theta}_1 = -y_2 \begin{bmatrix} y_2 \\ \omega_g y_1 \end{bmatrix} \\ &\dot{\theta}_4 = -y_2 \mu_2 \end{aligned}\right\} \quad h = 1, 4. \tag{24}$$

On the other hand, a still more simplified controller is obtained by setting the controller directly feeding the motor in the stator coordinate system. This means that $\omega_g = 0$, so the following SISO controller applicable to each subsystem (8) is finally obtained:

$$\left.\begin{aligned} &u_i(y_i, z, \theta_{hi}) = \theta_{1i} y_i + \theta_{4i} \mu_i \\ &\dot{\theta}_{1i} = -y_i^2 \\ &\dot{\theta}_{4i} = -y_i \mu_i \end{aligned}\right\} \quad \begin{aligned} &i = 1, 2 \\ &h = 1, 4. \end{aligned} \tag{25}$$

4.3. Application to the controller CTVAG

In the same way as in Section 4.2 it can be verified that adaptive laws designed for **CTVAG** are such that the storage functions V_{2i} for $i = 1, 2$ satisfy the following condition:

$$\begin{aligned} \dot{V}_{2i} = &\left\{ \sum_{j=1}^{n-1} \left[\frac{1}{\sqrt{1 + \frac{1}{\gamma_i^{\mathrm{T}} \gamma_i}}} \Phi_{\theta_{(j+1)i}} P_{ji}(y_i, z) \left(\frac{\partial W_o(z)}{\partial z_j}\right) y_i \Lambda_{bi}^* \right.\right. \\ &\left.\left. + \gamma_{(j+1)i} \Phi_{\theta_{(j+1)i}} \dot{\Phi}_{\theta_{(j+1)i}} \left|\Lambda_{bi}^*\right| \right] \right\} \\ &+ \left\{ \frac{1}{\sqrt{1 + \frac{1}{\gamma_i^{\mathrm{T}} \gamma_i}}} \Phi_{\theta_{1i}} A_i^{\mathrm{T}}(y_i, z) y_i \Lambda_{bi}^* \right. \\ &\left. + \gamma_{i1} \Phi_{\theta_{1i}} \dot{\Phi}_{\theta_{1i}} \left|\Lambda_{bi}^*\right| \right\} \\ &+ \left\{ \frac{1}{\sqrt{1 + \frac{1}{\gamma_i^{\mathrm{T}} \gamma_i}}} \Phi_{\theta_{(n+1)i}} \mu_i \, y_i \Lambda_{bi}^* \right. \\ &\left. + \gamma_{(n+1)i} \Phi_{\theta_{(n+1)i}} \dot{\Phi}_{\theta_{(n+1)i}} \left|\Lambda_{bi}^*\right| \right\} \\ &+ \frac{1}{\sqrt{1 + \frac{1}{\gamma_i^{\mathrm{T}} \gamma_i}}} \left(\frac{\partial W_o(z)}{\partial z}\right) \Lambda_o f_o(z) \le 0. \end{aligned}$$

Now, if the rotor flux components are eliminated from Eqs. (14) and (15), i.e. $\left(\frac{\partial W_o(z)}{\partial z}\right) = z = 0$, for $i = 1, 2$, and the corresponding factors depending on z will vanish without employing any adaptive law, so $\theta_{(j+1)} = \dot{\theta}_{(j+1)} = \gamma_{(j+1)} = \dot{\gamma}_{(j+1)} = 0$, $j = 1, 2$ can be imposed. Besides, from (18) and (19), by setting the controller in the stator coordinate system, where $\omega_g = 0$, a more simplified controller applicable to each subsystem (8) is obtained from (14)–(16) as follows:

$$\left.\begin{aligned} &u_i(y_i, z, \theta_{hi}) = \theta_{1i} y_i + \theta_{4i} \mu_i \\ &\dot{\theta}_{1i} = -\mathrm{sign}(\Lambda_{bi}^*) \frac{\gamma_1^{-1}}{\sqrt{1 + \frac{1}{\gamma_i^{\mathrm{T}} \gamma_i}}} y_i^2 \\ &\dot{\theta}_{4i} = -\mathrm{sign}(\Lambda_{bi}^*) \frac{\gamma_4^{-1}}{\sqrt{1 + \frac{1}{\gamma_i^{\mathrm{T}} \gamma_i}}} \mu_i y_i \\ &\dot{\gamma}_{1i} = -(\gamma_{1i} y_i)^2 \\ &\dot{\gamma}_{4i} = -(\gamma_{4i} \mu_i)^2 \end{aligned}\right\} \quad \begin{aligned} &i = 1, 2 \\ &h = 1, 4 \end{aligned} \tag{26}$$

4.4. Practical considerations in the design

The designs obtained up to here are based on Assumption 1 and use both desired and real variables. Next we will explain how to apply these controllers in practice. In the control scheme we will use a typical cascade control for induction motors as shown in Fig. 3, obtaining the controller output variables. Another assumption on these variables is made.

Assumption 3. The desired current and flux variables, i^*_{sx}, i^*_{sy} and ψ^*_{rx}, ψ^*_{ry}, obtained by applying a desired input voltage u^*_{sx}, u^*_{sy} (Assumption 1) are the same variables as are obtained from the speed controller output (Fig. 3(a)) and the flux controller output (Fig. 3(b)) respectively.

On the basis of Assumption 3 we can write

$$\begin{aligned} u_1 &= u^*_{sx} - u_{sx}, \quad \text{with } u^*_{sx} = \mu_1 \\ u_2 &= u^*_{sy} - u_{sy}, \quad \text{with } u^*_{sy} = \mu_2. \end{aligned} \tag{27}$$

From (27) we obtain the real voltages to be applied:

$$\begin{aligned} u_{sx} &= \mu_1 - u_1 \\ u_{sy} &= \mu_2 - u_2. \end{aligned}$$

Besides, it is known that a proportional controller of the form $\mu_i = -ky_i$ for $i = 1, 2$ turns a passive system asymptotically stable [9,10]. This is exactly the kind of feedback that will be used here (P controller).

5. Simulation results

In order to verify the advantages of the proposed controllers a comparison between the proposed controllers and a traditional current regulated PWM induction motor drive with PI loop controllers was carried out. In the simulations there was considered a squirrel-cage induction motor whose nominal parameters are: 15 [kW] (20 [HP]), 220 [V], $fp = 0.853$, 4 poles, 60 [Hz], $R_s = 0.1062$ [Ω], $X_{ls} = X_{lr} = 0.2145$ [Ω], $x_m = 5.8339$ [Ω], $R_r = 0.0764$ [Ω], $J = 2.8$ [kg m^2] and $B_p = 0$. These parameters were taken from [12]. The reference speed will increase in a ramp fashion up to 0.5 s. and after that time it will remain constant. On the other hand the load torque will be fixed equal to the nominal value. All the simulations were made using the software package MATLAB SIMULINK with the ODE 15s (stiff/NDF) integration method and a variable step size.

In Fig. 4 some steady state motor characteristics are shown, when a nominal voltage is supplied. It can be seen from Fig. 4 that even though the nominal torque is equal to 69.5 [N m], this motor can supply a torque of 230 [N m]. Besides, in spite the fact that the nominal current is 49.7 [A] it can reach values of 280 [A] at each phase. This is very important because in some cases the motor should be able to develop such high torque values in transient periods, for example to satisfy high performance.

The control schemes used in this study are shown in Figs. 5 and 6, where in the proposed control scheme simple proportional controllers are employed.

The control scheme shown in Fig. 5 was taken from [12], where it is important to note that the "Torque_Flux Control" block depends on the exact values or the estimates of parameters X_r, X_m, R_r. Besides, the "Field_orient" block depends on the exact values or the estimates of X_m and T_r. The

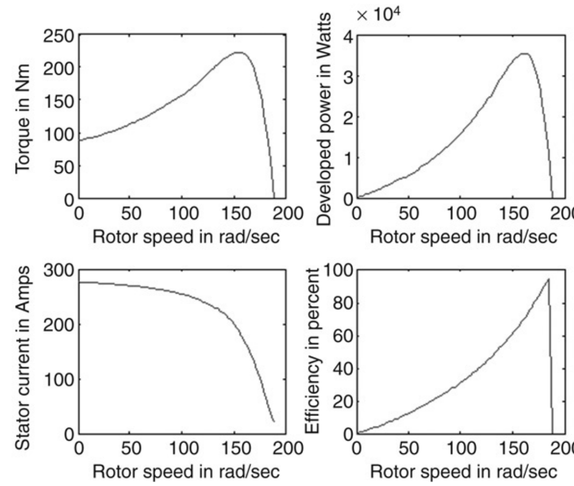

Fig. 4. Steady state characteristics of the 20 HP motor employed in this study.

PI speed controller is tuned with $P = 30$ and $I = 10$ (see Ref. [12]). The other control loops (flux and current) are not PI controllers.

The proposed control scheme is shown in Fig. 6. The two proposed controllers **CFAG** and **CTVAG** were developed and tested in the "Proposed Controller" block. It is important to observe that the speed, rotor flux, and stator current controllers are simple proportional (P) gains. These proportional gains were tuned such that the current, flux, speed and voltage were in suitable ranges.

This control scheme just needs the exact values or the estimates of parameters X_m and T_r for the "Field_orient" block. No other parameter or state estimations are needed.

Fig. 7 shows the information used to compare the two control schemes. There are shown the variations of the reference speed ω^*_r (Fig. 7(a)), the variations in load torque (Fig. 7(b)), the variation of about 30% in the stator and rotor resistance (Fig. 7(c) and Fig. 7(d)), the linear increasing up to the double the load inertia during the motor operation (Fig. 7(e)) and the variations in the viscous friction coefficient (Fig. 7(f)). For both proposed control strategies and the selected traditional control, five comparative tests considering the variations shown in Fig. 7 were carried out. These tests will allow us to study the behavior of the schemes in the following situations:

Situations:

1. Initially all the schemes are simulated with the motor nominal data. The reference of speed is increased as a ramp from 0 to 190 [RPM] in 0.5 [s] and the load torque is fixed at the nominal value 69.5 [N m].

2. Variations of the load torque, as indicated in Fig. 7(b).

3. Variations of the speed reference, as shown in Fig. 7(a).

4. Variation of the motor resistances, as shown in Fig. 7(c) and 7(d).

5. Variation of the load parameters, as indicated in Fig. 7(e) and 7(f).

6. Changes in the controller parameters (P and I) of the control loops.

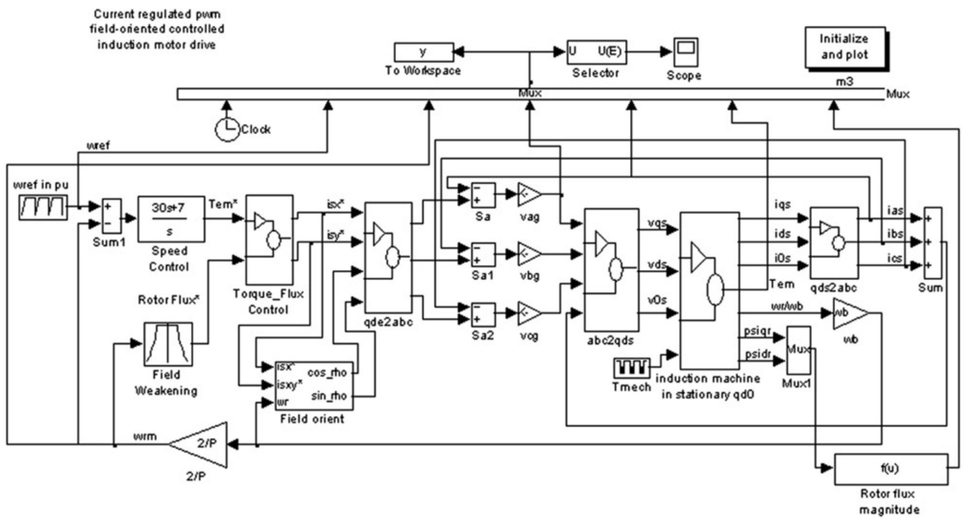

Fig. 5. Traditional current regulated PWM induction motor drive with PI controllers.

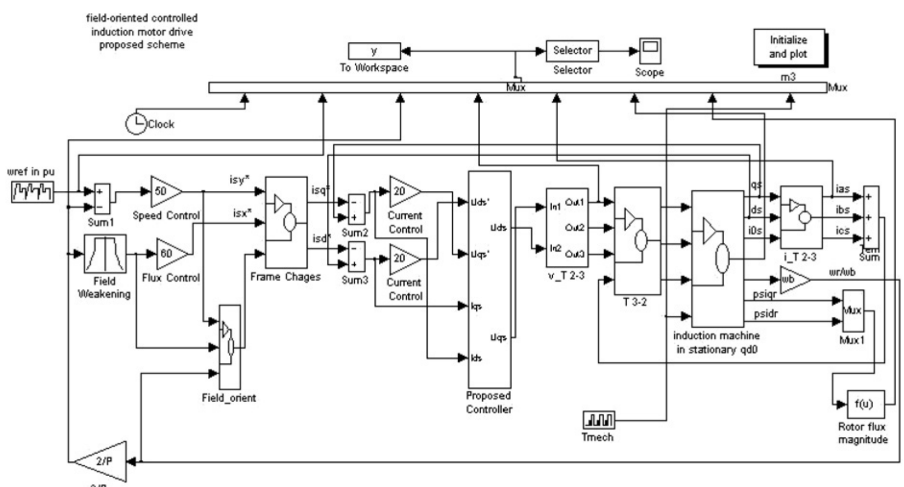

Fig. 6. Proposed control scheme with field oriented block.

In all the simulation results for the proposed controllers shown in what follows, the initial conditions of all the controller parameters and adaptive gains were set equal to zero, that is to say, $\theta_{ik}(0) = \gamma_{ih}(0) = 0$, for $i = 1, 2$ and $h = 1, 4$.

Figs. 8(a), 8(b) and 8(c) show the results obtained for the three controllers under normal conditions (i.e. according to Situation 1).

In Fig. 8(a) it can be seen how the traditional control based on linear controllers exhibits a steady state velocity error tending to 0.75%. The transient response is slow and presents initial oscillations. We can observe how the machine is magnetized, with the flow expressed in volts. The torque during the ramp is large to compensate for the inertia. Something similar happens with the phase current. The other phase voltages and currents present a waveform similar to that shown in Fig. 8(a), although shifted by 120°.

It is observed from Fig. 8(b) that the controller **CFAG** presents a better transient response after 0.5 [s] and with a more accurate stationary state (with a velocity error less than 0.5%) than the traditional case. In Fig. 8(c) we can see that the **CTVAG** is as accurate as the **CFAG**. Nevertheless, a better transient response is observed, being as fast as the **CFAG** case but less oscillatory.

We see in Figs. 9(a), 9(b) and 9(c) how the different schemes behave under variations of the load torque, as described in Fig. 7(b). From Fig. 9(a) it can be seen that for lower load torque greater speed accuracy is obtained, tending to zero error for the case when there is no load torque. The error in steady state is about 1% for a nominal load and 0.45 % for half a nominal load.

In the case of the **CFAG**shown in Fig. 9(b), the error values are smaller than for the traditional case, being 0.5% for a nominal load and 0.22% for half a nominal load. We observe how this controller is less affected under abrupt variations of load torque.

The controller **CTVAG** presents a similar response to **CFAG**, but in all the cases the transient response is of better quality. This can be concluded from Fig. 9(c).

We see now in Figs. 10(a), 10(b) and 10(c) the behavior under variations of the speed reference at nominal load, according to the variations indicated in Fig. 7(a). In the case of the traditional controller (Fig. 10(a)) the error in stationary state

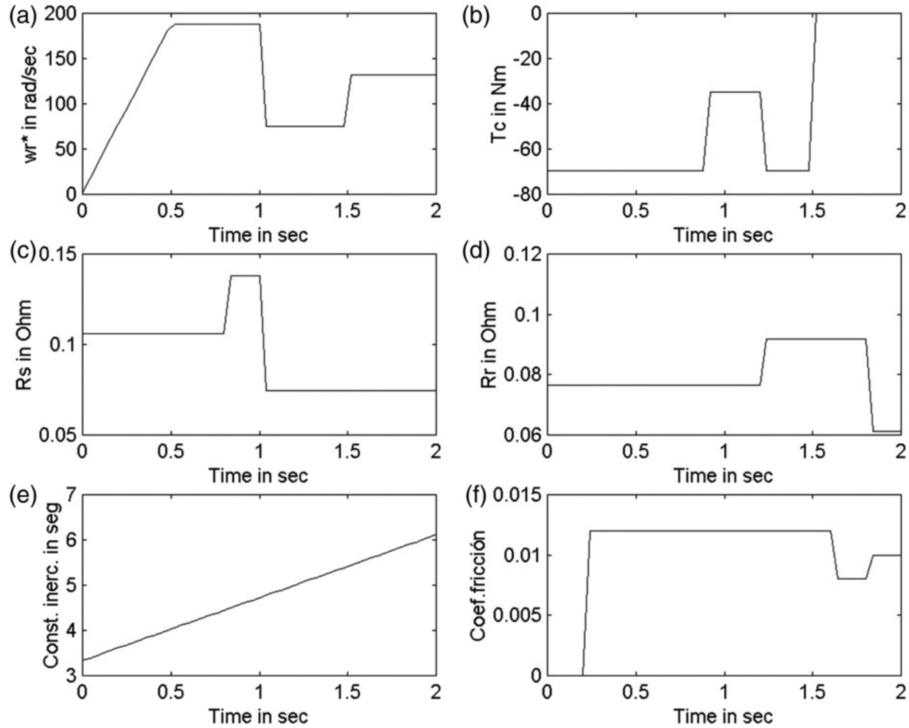

Fig. 7. Parameter and reference variations used in the set of comparative tests.

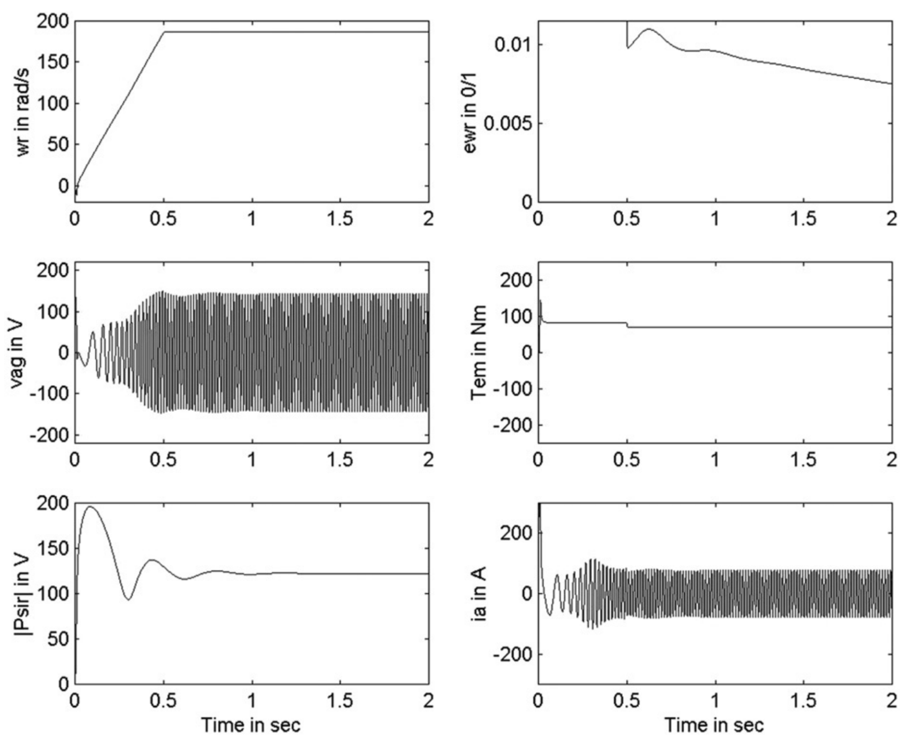

Fig. 8(a). Results of the traditional control scheme without considering parameter variations, under a speed ramp.

increases for lower speed. The error is about 1% at nominal speed and approximately double for half of nominal speed.

The results for proposed controllers **CFAG** and **CTVAG** are much better than the traditional ones, getting smaller errors, whereas the rest of the variables have a suitable behavior, similar to the one found for the traditional controller (see Figs. 10(b) and 10(c)). In these cases we have an error of about 0.5% for nominal speed and approximately 1.1% at half of the nominal speed.

When analyzing Situation 4 (Fig. 7(c) and (d)) all of the controllers present a good behavior under changes of the stator resistance (see Figs. 11(a)–11(c)). Nevertheless, under changes

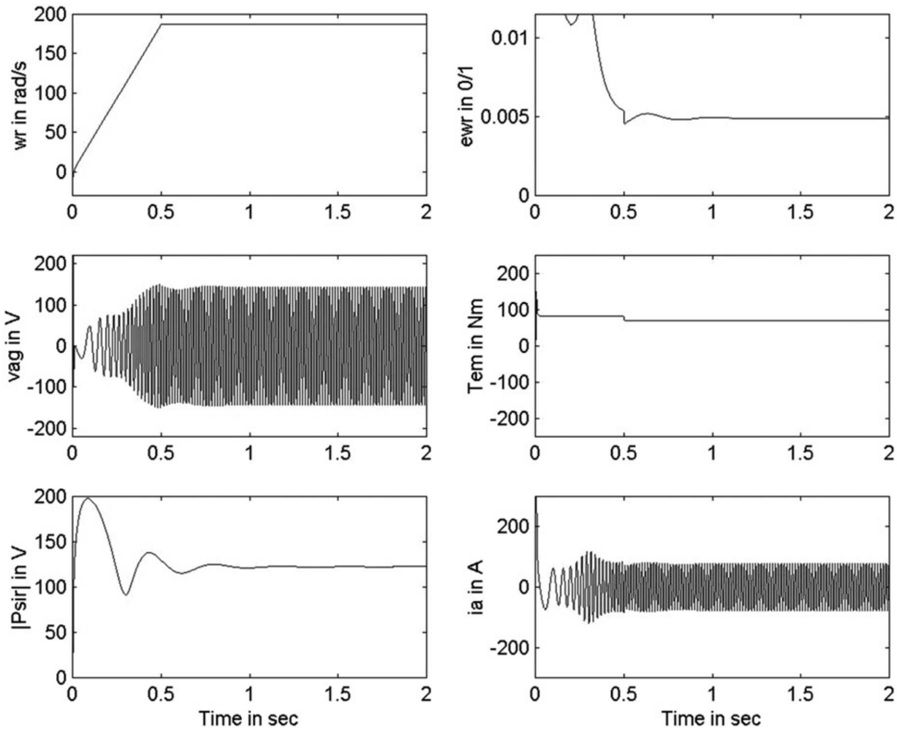

Fig. 8(b). Results using **CFAG**, without considering parameter variations, under a speed ramp.

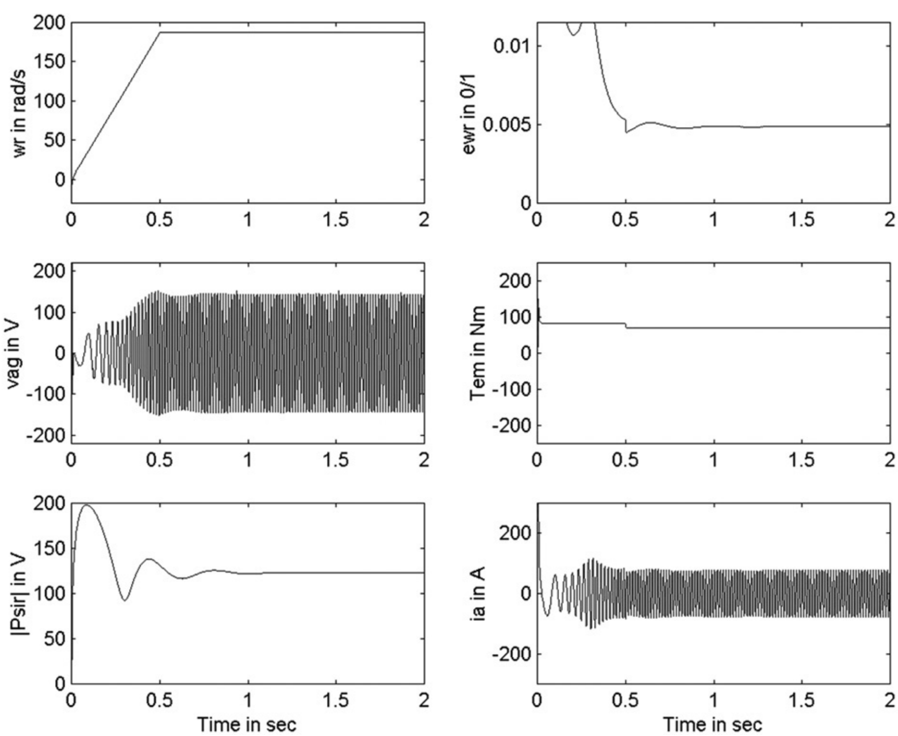

Fig. 8(c). Results using **CTVAG**, without considering variations of parameters, under a speed ramp.

of the rotor resistance the correct field orientation is lost and the speed response is considerably affected. Notice how the flow of the machine diminishes considerably.

For the proposed controllers shown in Figs. 11(b) and 11(c), the responses in all the cases are much more robust than the traditional controller (Fig. 11(a)). Both proposed controllers present lesser speed errors in the steady state than the classical scheme.

Considering now the variations of the load parameters according to Situation 5 (Fig. 7(e) and (f)), none of the three controllers under study was substantially affected, as shown in Figs. 12(a)–12(c).

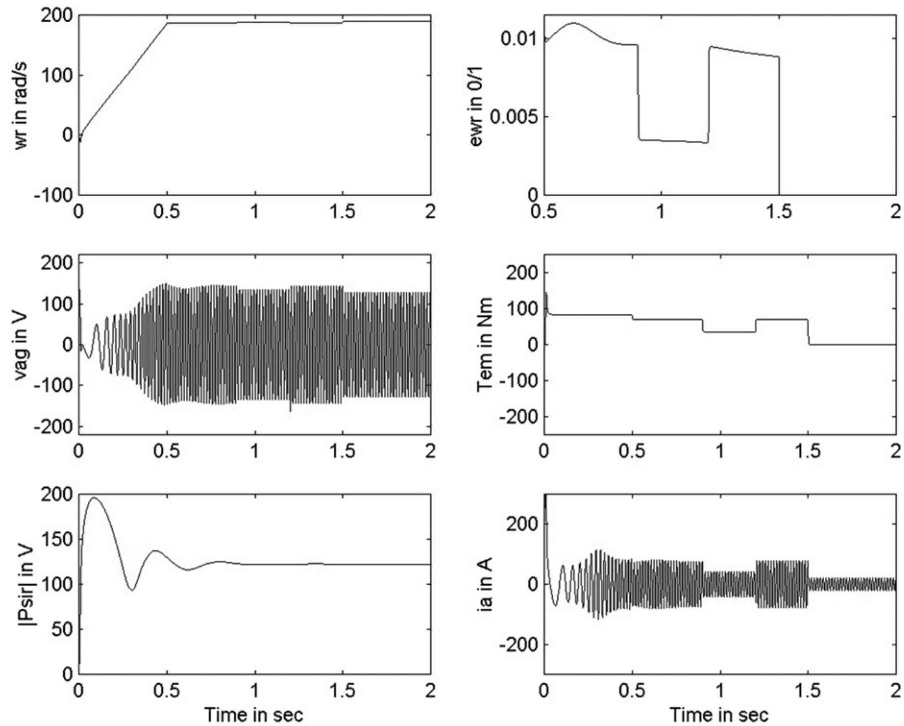

Fig. 9(a). Results of the traditional controller under variations of load torque.

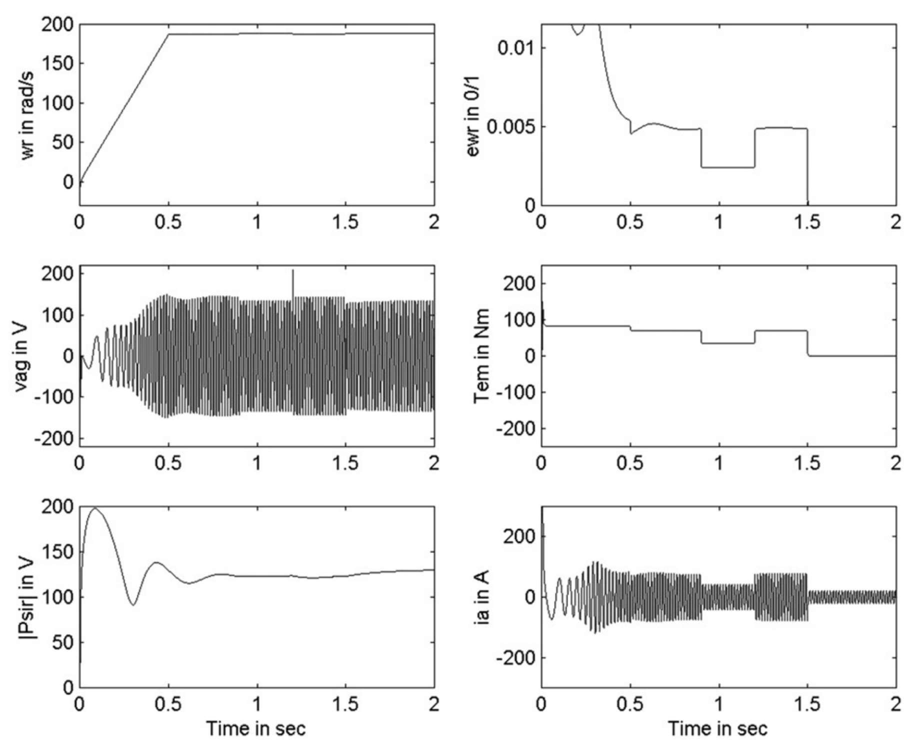

Fig. 9(b). Results using **CFAG**, under variations of the load torque.

For the proposed controllers, differences found in the general behavior still remain. **CFAG** presents a lesser error in the steady state than the traditional controller. **CTVAG** exhibits a similar response to that of the **CFAG** but with a better transient behavior.

In Fig. 13(a) the results corresponding to the traditional controller are shown for variation of the proportional constant (P) in the speed loop, from a value of 30 according to [12] to a value of $P = 20$. That is, for a variation of 33.3%. The rest of the flux and current controller parameters and the

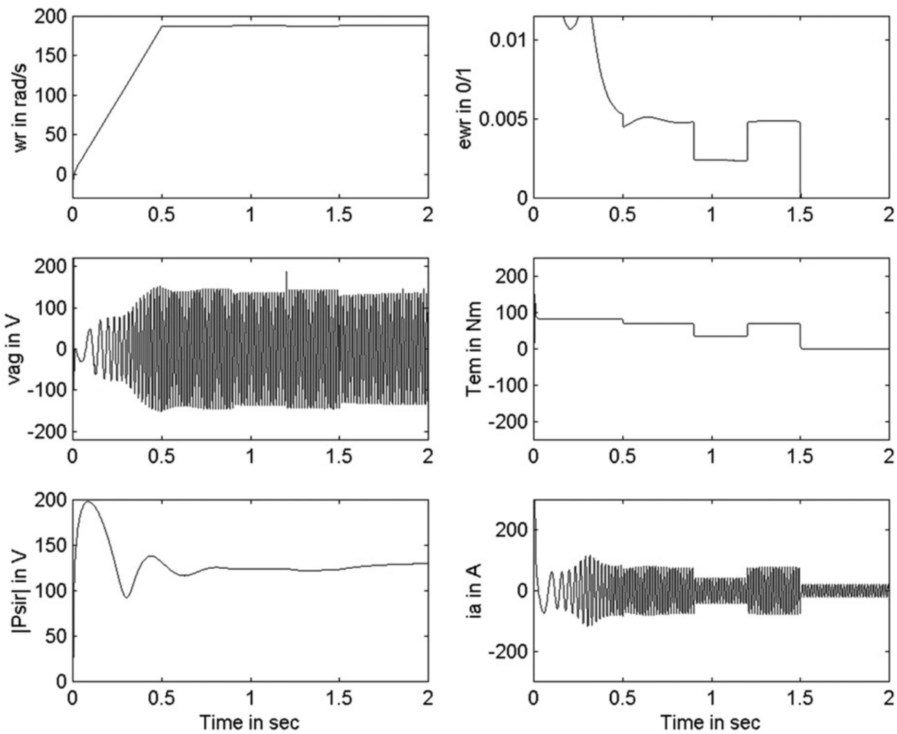

Fig. 9(c). Results using **CTVAG**, under variations of the load torque.

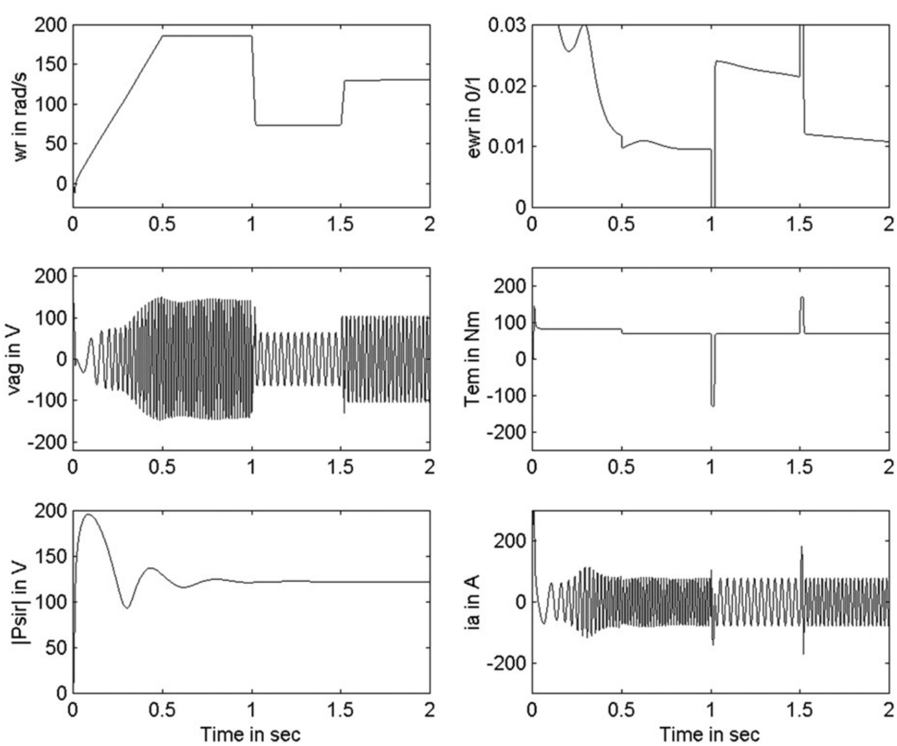

Fig. 10(a). Results of the traditional control under variations in the speed reference.

value of the integral constant (I) of the speed loop were not modified.

It important to point out that with the variation of only one controller parameter the control scheme does not perform properly. That is why it was not necessary to explore the influence of other parameter variations.

In the case of the proposed controllers, the proportional gains of all the control loops were changed while the integral

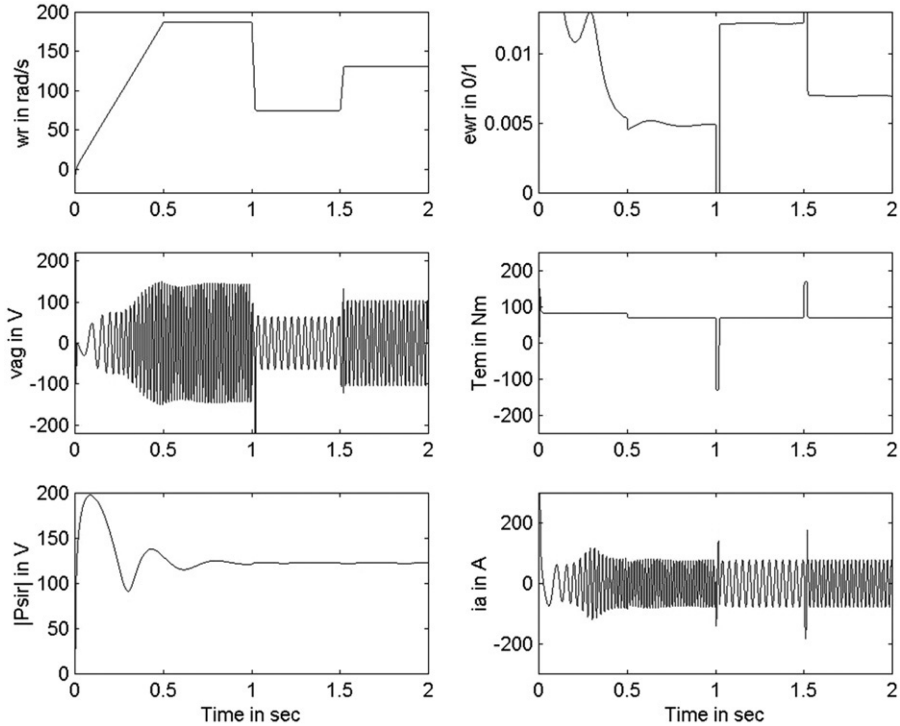

Fig. 10(b). Results using **CFAG** under variations in the speed reference.

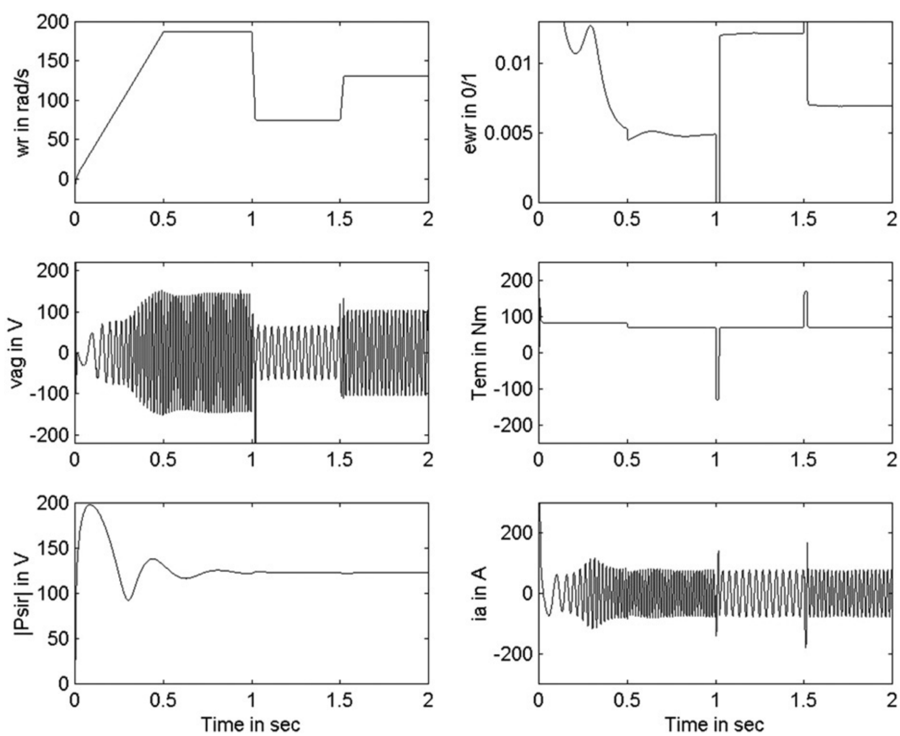

Fig. 10(c). Results using **CTVAG** under variations in the speed reference.

constant was kept fixed at the value 10. For both controllers **CFAG** and **CTVAG**, variations for the speed control parameters of 37.5% were applied (P from 80 to 50). The flux control parameter was varied by 13% (P from 69 to 60). The proportional gains in the current loops were varied by 33.3% (P from 30 to 20). From Figs. 13(b) and 13(c) it can be seen that in spite of these simultaneous variations the speed error continues to be less than 1% and the transient response after 0.5 [s] was

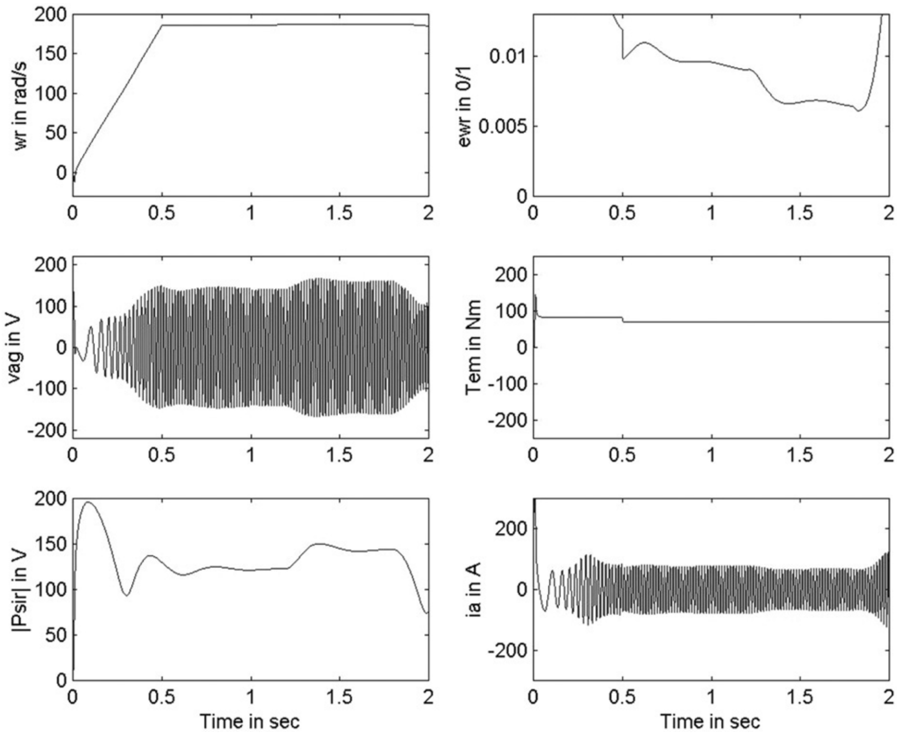

Fig. 11(a). Results of the traditional control scheme under variations of motor parameters.

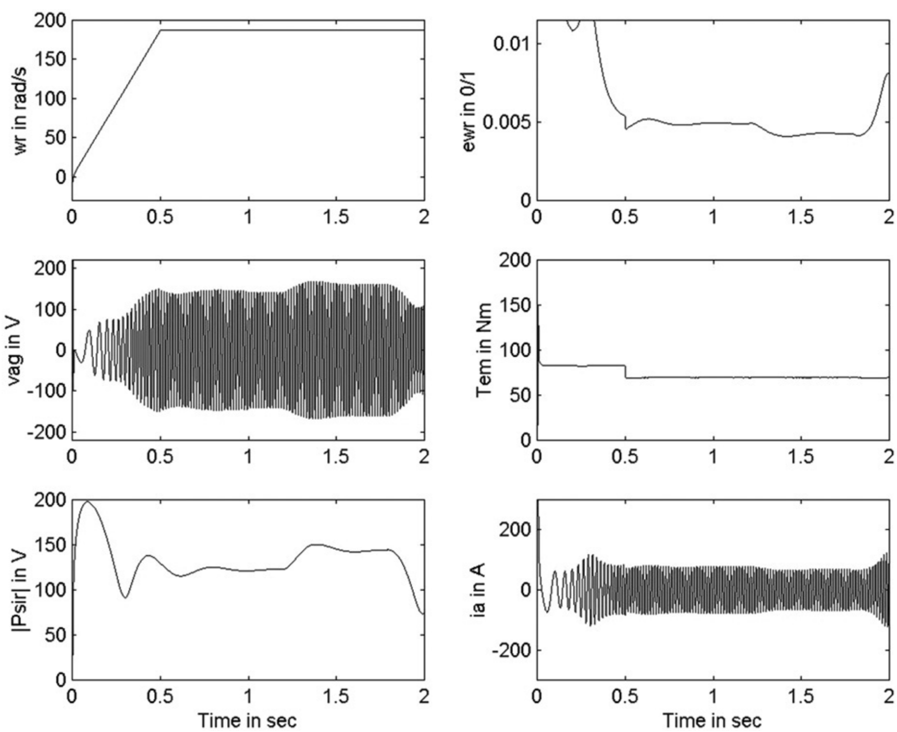

Fig. 11(b). Results using **CFAG**, for variations of motor parameters.

not affected. **CFAG** as well as **CTVAG** guarantee good results for wide ranges of variations of the proportional gains.

With the proposed control schemes wider parameter variations can be handled and a large number of parameters can be changed without an important deterioration of the overall system behavior.

6. Conclusions

A principle called the "Principle of Torque–Flux Control" has been proposed in this paper; it is applied to the adaptive controllers based on passivity given in [9,10]. As a result, two simplified SISO control schemes have been obtained without state estimation (components of the rotor flux) and without

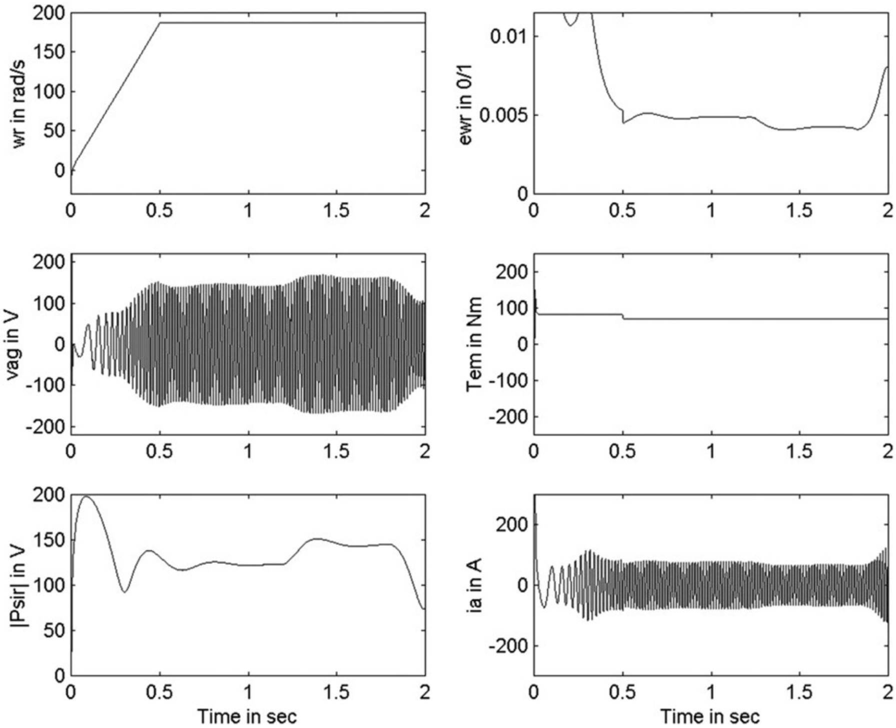

Fig. 11(c). Results using **CTVAG**, for variations of motor parameters.

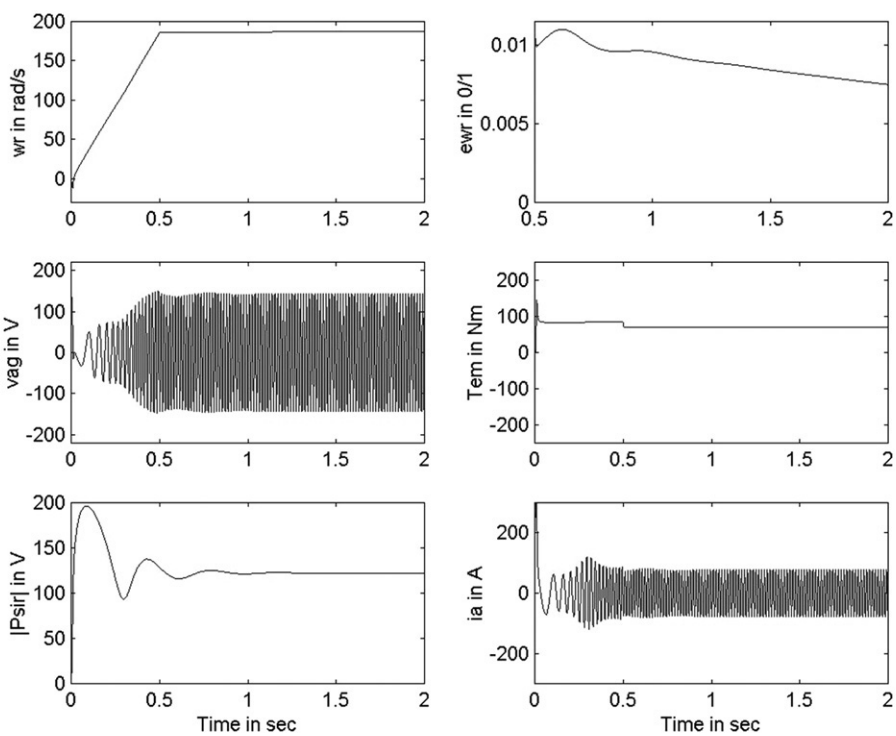

Fig. 12(a). Results of the traditional control, for variations of the load parameters.

parameter estimation (parameters of the motor–load set) except those used for the field orientation block. Their behaviors were studied and compared with that of a classical control scheme [12], and they were found suitable and robust.

Compared with other control schemes proposed in the literature such as those based on sliding modes [3–5], artificial intelligence [6,7] and non-adaptive passivity [5,8], we have presented two simple and novel SISO controllers. They have adaptive characteristics, they are robust in the presence of load parameter variations and they use simple proportional

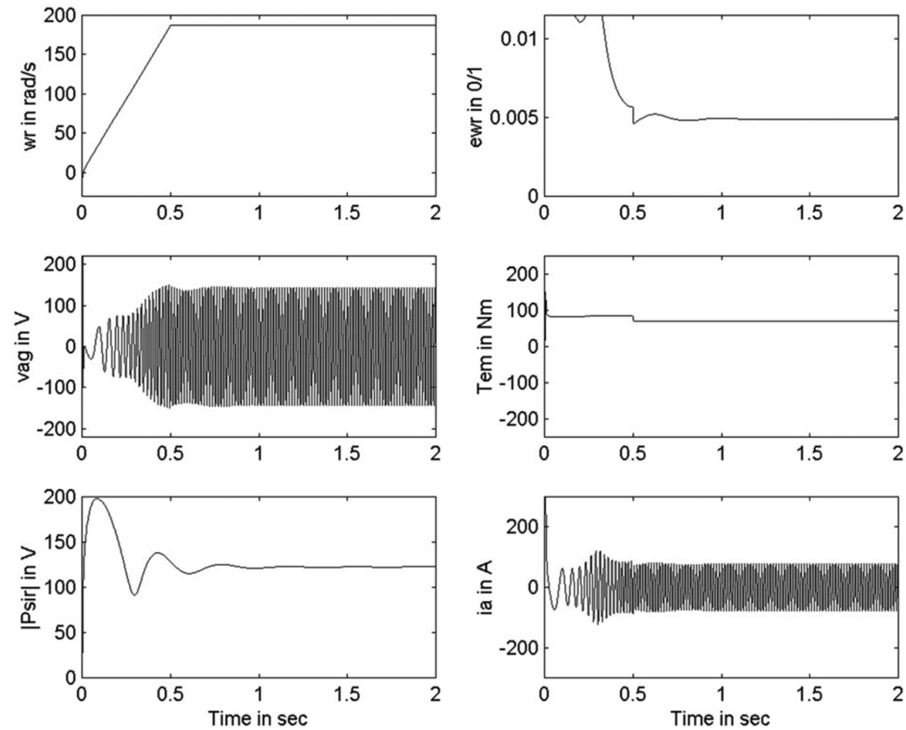

Fig. 12(b). Results of **CFAG** under variations of load parameters.

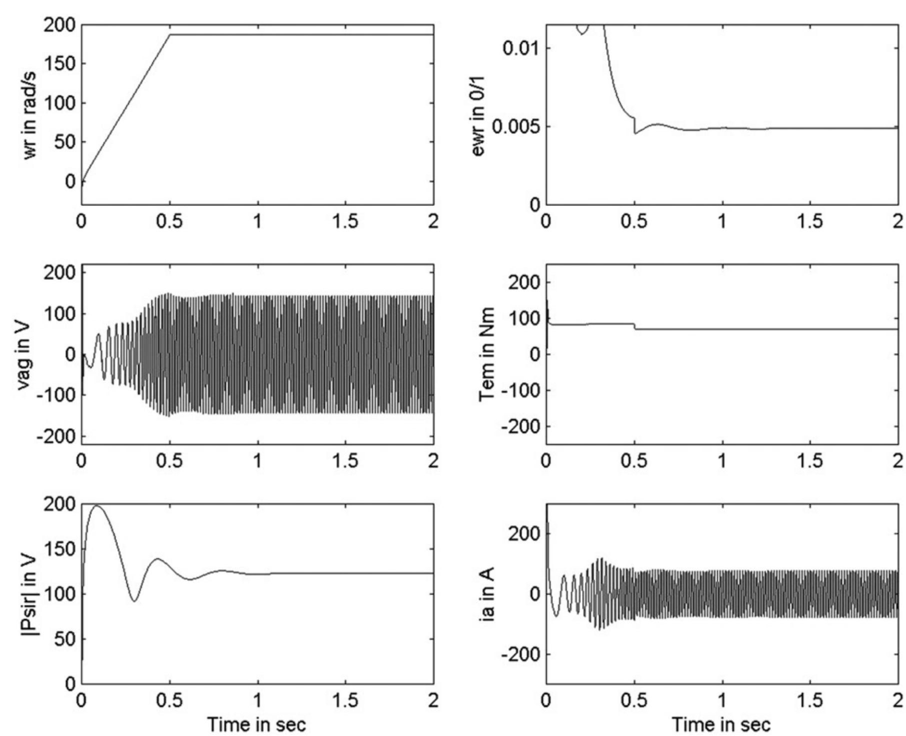

Fig. 12(c). Results of **CTVAG** under variations of load parameters.

controllers in the rotor speed, rotor flux and stator current control loops. They are also robust for a large range of proportional gain variations.

All of the above mentioned characteristics for the proposed control schemes guarantee high performance control, such as high starting torque at low speed and during the transient, accuracy in steady state, wide range of speed control and good response under speed changes.

It was shown that the controller **CTVAG** presents better transient response than the controller **CFAG**, due to the time-

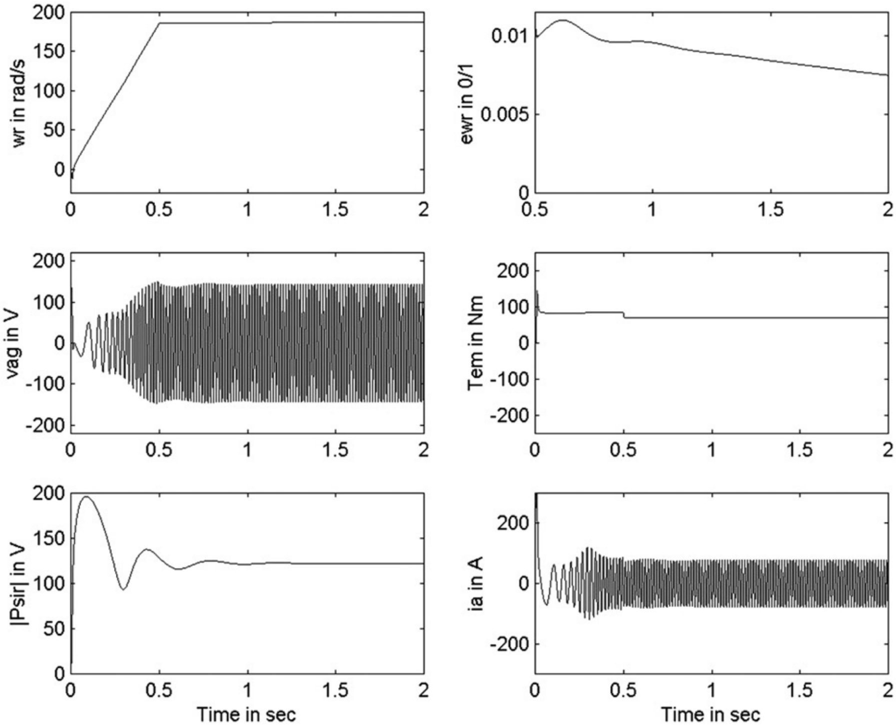

Fig. 13(a). Results for the traditional controller, under changes in the tuning of the proportional gain of the speed loop control.

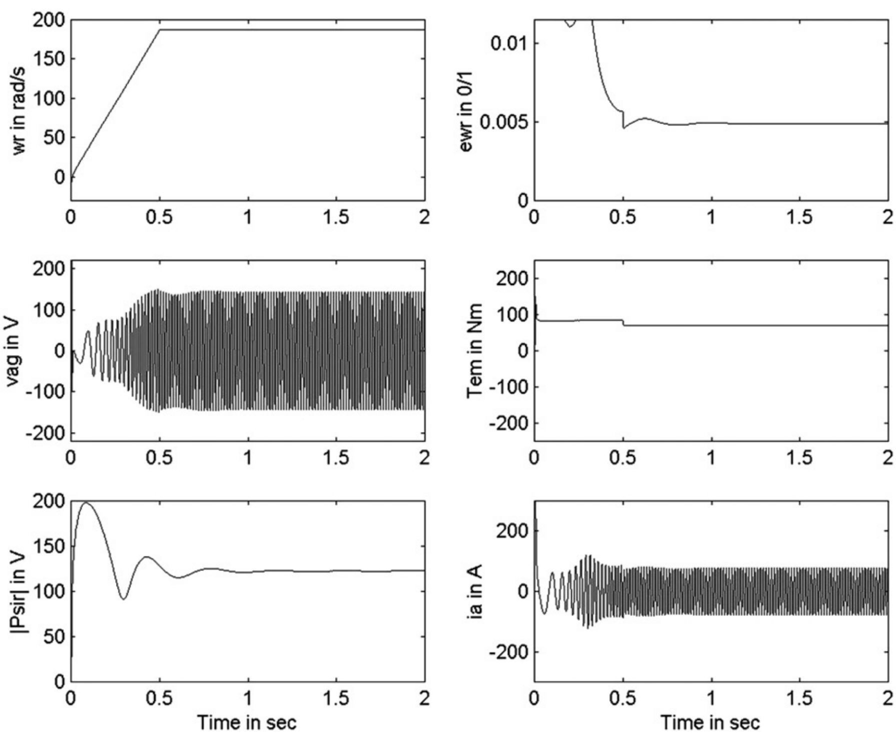

Fig. 13(b). Results using **CFAG** under changes in the tuning of the proportional gains.

varying adaptive gains included. Besides, it has some additional adjustment possibilities such as adjustment of the initial values of the controller parameters plus the initial values of time-varying adaptive gains.

Acknowledgements

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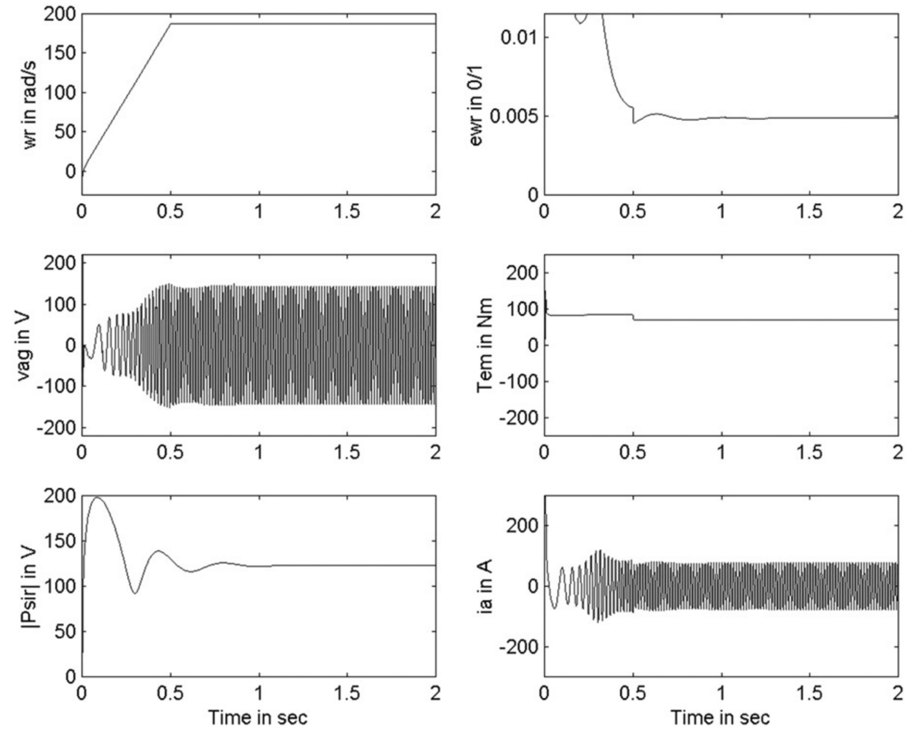

Fig. 13(c). Results using **CTVAG** under changes in the tuning of the proportional gains.

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