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INDUCTION MOTOR CONTROL BASED ON ADAPTIVE PASSIVITY

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ABSTRACT

In this paper two new schemes for induction motor control are proposed and compared. Both approaches are based on the concept of adaptive passivity. First, a technique using the scheme of field oriented control (FOC) is proposed, and by means of an adaptive state feedback, a passive equivalent system is obtained. Furthermore, making use of the novel torque-flux control principle (TFCP), the proposed scheme is greatly simplified. Second, a technique based on energy shaping approach, which does not make use of the FOC scheme, is proposed. The technique is based on interconnection and damping assignment (IDA) control transforming the original system into a passive one. Since this technique does not use the FOC scheme, it gives more flexibility in the implementation. Both techniques are then implemented at laboratory level and compared from experimental viewpoint using as benchmark the standard FOC scheme with PI controllers.

Key Words: Induction motor, adaptive control, field oriented control, passivity based control, adaptive passivity feedback, interconnection and damping assignment.

I. INTRODUCTION

During the past three decades control of induction motors has attracted the attention of researchers all over the world. This interest is mainly due to the advances in power electronics together with the statement of the ‘Field Orientation Principle’ developed in 1969 [1], that allows an independent torque and speed control for induction motors, similar to the case of continuous current motors with independent excitement.

The nonlinear nature of induction machines has motivated the use of several control techniques including classical (fixed parameters), adaptive and nonlinear. In the last years several modern control techniques, taking into account the nonlinear dynamics of the induction motor, have been developed, including adaptive control, geometric control, predictive control [2] and robust control. Perhaps the most interesting attempts used lately are related to the concept of passivity based control (PBC) [3–5] and to the concept of energy shaping [6, 7].

The first approach deals with the study of passive systems and their properties, mainly the case of passive equivalence by state feedback [3]. With this technique it is possible to stabilize a nonlinear equivalent system in a simple way, using only a proportional controller. In order to make this passivity technique more robust, the adaptive passivity case was thoroughly studied in [5, 8–10] for the case of unknown parameters, where an equivalent passive system is

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obtained by means of an adaptive feedback. Passivity concepts have been applied to permanent magnet stepper motors [11] and to motion control for robot manipulators [12].

In parallel with development of passivity based techniques, a new control methodology based on handling the system's energy was developed [6, 7]. This technique is applicable in a simple way to mechanical and electromagnetic systems, and aims to control the system from an energy point of view, modeling the problem as a function of energy transfers and interconnections using a Lagrangian or Hamiltonian formulation of the system. This methodology allows a great flexibility when designing a control strategy, with the inconvenience that the state of the system in this new formulation does not necessarily possess a physical meaning.

In this paper, two new control schemes based on adaptive passivity applied to induction motors are presented and compared from experimental viewpoint. The first is an adaptive passivity feedback developed in [5, 8–10] that was applied later to the case of controlling induction motors [13–17]. This technique is based on the field oriented control (FOC) scheme, where using an adaptive state feedback, a passive equivalent system is obtained and then controlled. Furthermore, using a novel torque-flux control principle (TFCP) [13, 15], the proposed scheme is greatly simplified.

The second scheme makes use of the concept of energy shaping referred as IDA-PBC (Interconnection and Damping Assignment—Passivity-Based Control), which was developed in [18–20]. The technique is based on an energy shaping approach and does not make use of the FOC scheme. This technique uses interconnection and damping assignment (IDA) control transforming the original system into a passive one. Since this technique does not use the FOC scheme, it gives more flexibility in the implementation.

Both strategies are compared, with the standard FOC strategy using fixed PI controllers. This standard strategy presents two loops arranged in cascade, one inner loop for torque control and an outer loop for speed control. The control of the inner loop is carried out using a proportional controller and the outer one uses a proportional-integral controller. This strategy will be referred from now on as the basic control strategy (BCS).

The comparison of these techniques from simulation point of view can be found in [21, 22] and were developed previous to the experimental comparison done in the present paper.

II. PASSIVITY BASED CONTROL

2.1 Adaptive passivity feedback (APF)

Let us consider a system Σ in the normal form

$$\Sigma : \begin{cases} \dot{y}(t) = a(y,z) + b(y,z)u(t) \\ \dot{z}(t) = f_o(z) + p(y,z) \end{cases} \tag{1}$$

where $y(t), u(t) \in \Re^m$, $z(t) \in \Re^{n-m}$, $a(y,z) \in \Re^m$, $b(y,z) \in \Re^{m\times m}$, $f_0(z) \in \Re^{n-m}$. $f_0(z)$ is called the zero dynamics of the system. Let us assume first that all the parameters are constant and known, and that the origin is an equilibrium state. According to [3], Σ is locally equivalent to a passive system with storage function V positive definite if and only if Σ has unity relative degree at $x = 0$ and there exists a positive definite function $W_0(z)$ such that

$$\nabla_z W_0 f_0 \le 0 \tag{2}$$

The necessary state feedback to achieve the proposed objective is [3]

$$u = b^{-1}(y,z)[-a(y,z) - p(y,z)\nabla_z W_0(z) + v] \tag{3}$$

where the new input to the system is v. Thus the resulting system is C^2 passive from v to y. It is shown [3] that using feedback (3) system Σ can be transformed into a C^2 passive system with storage function

$$V = W_0(z) + \tfrac{1}{2} y^T y \tag{4}$$

Then, the equivalent system to be controlled is shown in Fig. 1.

For the case when parameters are unknown the system is written in the normal form defined in [9, 10], having the following representation

$$\begin{aligned} \dot{y} &= \Lambda_a A(y,z) + \Lambda_b B(y,z)u \\ \dot{z} &= \Lambda_0 f_o(z) + P^T(y,z)\Lambda_p y \end{aligned} \tag{5}$$

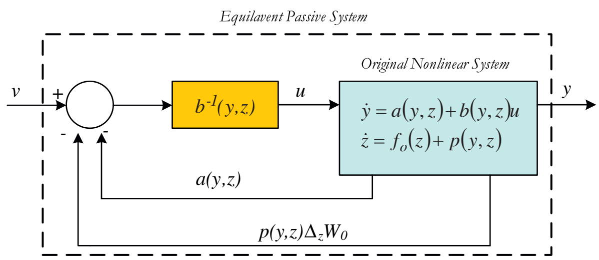

Fig. 1. Equivalent passive system by means of a state feedback. Case of known parameters.

where $\Lambda_a, \Lambda_b \in \Re^{m\times m}$, $\Lambda_0 \in \Re^{(n-m)\times(n-m)}$, $\Lambda_p \in \Re^{p\times m}$ and $A(y,z) \in \Re^m$, $B(y,z) \in \Re^{m\times m}$, $P(y,z) \in \Re^{p\times(n-m)}$. Then, the feedback required to achieve the equivalent passive system when parameters Λ_a, Λ_b, Λ_p are unknown is [9, 10]

$$u = B^{-1}(y,z)\Lambda_b^{-1}[-\Lambda_a A(y,z) -\Lambda_p^T P(y,z)\nabla_z W_0(z)+v] \tag{6}$$

Since parameters Λ_a, Λ_b, Λ_p are unknown, the following state feedback is proposed in [9, 10] for the case of Λ_b diagonal

$$u(y,z,\theta_j) = B^{-1}(y,z)\lfloor\theta_1(t)A(y,z)+\theta_2(t) P(y,z)\nabla_z W_0(z)+\theta_3(t)v\rfloor \tag{7}$$

with adaptive laws

$$\begin{aligned}
\dot\theta_1(t) &= -\frac{\Gamma_1^{-1}(t)}{\alpha(t)}y(t)A^T(y,z)\mathrm{sign}(\Lambda_b)\\
\dot\theta_2(t) &= -\frac{\Gamma_2^{-1}(t)}{\alpha(t)}y(t)\nabla_z W(z)^T P^T(y,z)\mathrm{sign}(\Lambda_b)\\
\dot\theta_3(t) &= -\frac{\Gamma_3^{-1}(t)}{\alpha(t)}y(t)v^T(t)\mathrm{sign}(\Lambda_b)
\end{aligned} \tag{8}$$

where $\alpha(t)=\sqrt{1+\mathrm{Trace}(\Gamma_1^{-2}(t)+\Gamma_2^{-2}(t)+\Gamma_3^{-2}(t))}$ is a normalization factor and the adaptive gains vary in the following way

$$\begin{aligned}
\dot\Gamma_1(t) &= -\Gamma_1(t)A^T(y,z)\Gamma_1(t)\\
\dot\Gamma_2(t) &= -\Gamma_2(t)P(y,z)\nabla_z W(z)\nabla_z W(z)^T P^T(y,z)\Gamma_2(t)\\
\dot\Gamma_3(t) &= -\Gamma_3(t)v(t)v(t)^T\Gamma_3(t)
\end{aligned} \tag{9}$$

$\mathrm{sign}(\Lambda_b)$ is a diagonal matrix on whose diagonal the sign of each element of matrix Λ_b are located. The adaptive case with fixed gains, can also be considered by choosing constant matrices $\Gamma_i=\Gamma_i^T>0$. The advantage of choosing time-varying adaptive gains is that better transient responses for the overall adaptive system can be obtained. The equivalent passive system to be controlled is shown in Fig. 2.

In the case that matrix Λ_b is not diagonal a similar scheme can also be derived and the reader is referred to [9, 10] for more details. In this study the induction motor model leads to a normal form with Λ_b diagonal.

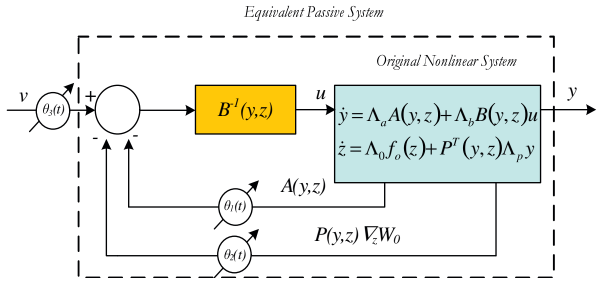

Fig. 2. Equivalent passive system by means of an adaptive feedback. Case of unknown parameters.

2.2 Energy shaping control

Let us consider a system described in the form called Port-Controlled Hamiltonian (PCH) [4],

$$\Sigma_{PCH}: \begin{cases} \dot x = [\mathcal{J}(x)-\mathcal{R}(x)]\nabla H + g(x)u \\ y = g^T(x)\nabla H \end{cases} \tag{10}$$

where $x\in\Re^n$ is the state, and u, $y\in\Re$ are the input and the output of the system. H represents the system's total stored energy, $\mathcal{J}(x)$ is a skew-symmetric matrix ($\mathcal{J}(x)=-\mathcal{J}^T(x)$) called the interconnection matrix and $\mathcal{R}(x)$ is a symmetric positive definite matrix ($\mathcal{R}(x)=\mathcal{R}(x)^T\geq 0$) called the damping matrix.

Let us assume [6, 7] that there exist matrices $g^\perp(x)$, $\mathcal{J}_d(x)=-\mathcal{J}_d^T(x)$, $\mathcal{R}_d(x)=\mathcal{R}_d^T(x)\geq 0$ and a function $H_d:\Re^n\to\Re$, such that

$$g^\perp(x)[\mathcal{J}(x)-\mathcal{R}(x)]\nabla H = g^\perp(x)[\mathcal{J}_d(x)-\mathcal{R}_d(x)]\nabla H_d \tag{11}$$

where $g^\perp(x)$ is the full-rank left annihilator of $g(x)$ ($g^\perp(x)g(x)=0$) and $H_d(x)$ is such that

$$x^* = \arg\min_{x\in\mathbb{R}}(H_d) \tag{12}$$

Then, applying the control $\beta(x)$ defined as

$$\beta(x) = [g^T(x)g(x)]^{-1}g^T\{[\mathcal{J}_d(x) -\mathcal{R}_d(x)]\nabla H_d - [\mathcal{J}(x)-\mathcal{R}(x)]\nabla H\} \tag{13}$$

the overall system under control can be written as

$$\dot x = [\mathcal{J}_d(x)-\mathcal{R}_d(x)]\nabla H_d \tag{14}$$

where x^* is a locally Lyapunov stable equilibrium. That is to say applying control (13) to (10) the dynamics of the system is changed to (14).

In order to find control (13) there exist two ways to do it. The first one consists on fixing the topology of the system (by fixing $\mathcal{J}_d$, $\mathcal{R}_d$ and $g^\perp$) and solve the differential equation (11). The second method consists on fixing H_d (the initial geometrical form of the desired energy) and then (11) becomes an algebraic system that has to be solved for $\mathcal{J}_d$, $\mathcal{R}_d$ and $g^{\perp}$ [6, 7].

III. INDUCTION MOTOR MODEL

The adaptive passivity feedback of Section 4.1 will be derived for the standard induction motor model referred to an arbitrary reference system of $x-y$ coordinates rotating at a generic speed ω_g. This model can be written in the following form [13, 21, 23–25]

$$\begin{aligned} \dot{x} &= f(x)+g(x)u \\ y &= h(x) \end{aligned} \tag{15}$$

where x, $f(x)$, $g(x)$, $h(x)$ and u are defined as follows

$$x = (i_{sx} \quad i_{sy} \quad \psi_{rx} \quad \psi_{ry} \quad \omega_r)^T$$

$$f(x) = \begin{pmatrix} -\left(\frac{R_s}{\sigma L_s}+\frac{L_m^2 R_r}{\sigma L_s L_r^2}\right) i_{sx}+\omega_g i_{sy} \\ +\frac{L_m}{\sigma L_s L_r^2} R_r \psi_{rx}+\frac{L_m}{\sigma L_s L_r}\omega_r \psi_{ry}, \\ -\left(\frac{R_s}{\sigma L_s}+\frac{L_m^2 R_r}{\sigma L_s L_r^2}\right) i_{sy}-\omega_g i_{sx} \\ +\frac{L_m}{\sigma L_s L_r^2} R_r \psi_{ry}-\frac{L_m}{\sigma L_s L_r}\omega_r \psi_{rx}, \\ R_r\frac{L_m}{L_r} i_{sx}-\frac{R_r}{L_r}\psi_{rx}+(\omega_g-\omega_r)\psi_{ry}, \\ R_r\frac{L_m}{L_r} i_{sy}-\frac{R_r}{L_r}\psi_{ry}-(\omega_g-\omega_r)\psi_{rx}, \\ (T_{em}-T_c)\frac{1}{J}-\frac{B_p}{J}\omega_r \end{pmatrix} \tag{16}$$

$$g(x) = \begin{pmatrix} \frac{1}{\sigma L_s} & 0 & 0 & 0 & 0 \\ 0 & \frac{1}{\sigma L_s} & 0 & 0 & 0 \end{pmatrix}^T$$

$$h(x) = (i_{sx} \;\; i_{sy} \;\; \omega_r)^T$$

$$u = (u_{sx} \;\; u_{sy})^T$$

The meaning of variables and parameters is as follows; i_{sx}, i_{sy} are stator currents, ψ_{rx}, ψ_{ry} are rotor fluxes, ω_r is the rotor speed and u_{sx}, u_{sy} are 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 stator and rotor resistances respectively. J is the rotor inertia, T_{em} is the electromagnetic torque produced by motor, T_c is the load torque and B_p is the mechanical viscous damping coefficient.

We define $\sigma = 1-L_m^2/L_s L_r$ as the leakage or coupling factor, $R_s' = R_s + R_r L_m^2/L_r^2$ as the stator transient resistance and σL_s as the stator transient inductance. Furthermore, the electromagnetic torque is given by

$$T_{em} = \frac{3}{2}\frac{p}{2}\frac{L_m}{L_r}(\psi_{rx} i_{sy}-\psi_{ry} i_{sy}) \tag{17}$$

To design the adaptive passivity feedback controller in Section 4.1, an $x-y$ reference system fixed to the stator machine will be chosen, so that $\omega_g = 0$.

For IDA-PBC scheme developed in Section 4.2, the induction motor model previously stated should be expressed in the form called Port-Controlled Hamiltonian (PCH) [4], which has the general form shown in (10). In this study it will be assumed a load torque proportional to rotor speed ($T_c = B\omega_r$) which represents typically the case when the motor is moving a kind of fan. In this particular case the PCH model of the induction motor, assuming also that the speed of the $x-y$ reference system is synchronized to electrical frequency ($\omega_g = \omega_s$), has the form [18, 19]

$$\begin{aligned} \dot{x} &= \begin{bmatrix} -R_s & 0 & 0 & 0 & 0 \\ 0 & -R_r & 0 & 0 & -x_4 \\ 0 & 0 & -R_s & 0 & 0 \\ 0 & 0 & 0 & -R_r & x_2 \\ 0 & x_4 & 0 & -x_2 & -B' \end{bmatrix} \nabla H \\ &\quad + \begin{bmatrix} 1 & 0 & x_3 \\ 0 & 0 & x_4 \\ 0 & 1 & -x_1 \\ 0 & 0 & -x_2 \\ 0 & 0 & 0 \end{bmatrix} \begin{pmatrix} u_{sx} \\ u_{sy} \\ \omega_s \end{pmatrix} \\ y &= \begin{bmatrix} 1 & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 \\ x_3 & x_4 & -x_1 & -x_2 & 0 \end{bmatrix} \nabla H \\ &= \begin{bmatrix} i_{sx} \\ i_{sy} \\ 0 \end{bmatrix} \\ x &= [\psi_{sx} \;\; \psi_{rx} \;\; \psi_{sy} \;\; \psi_{ry} \;\; J\omega_r]^T \\ &= [x_{12}^T \;\; x_{34}^T \;\; x_5]^T \\ H &= \tfrac{1}{2} x_{12}^T L^{-1} x_{12} + \tfrac{1}{2} x_{34}^T L^{-1} x_{34} + \tfrac{1}{2} J^{-1} x_5^2 \end{aligned} \tag{18}$$

where ψ_{sx}, ψ_{sy} are the stator fluxes and $B' = B_p + B$. Matrix L is defined as

$$L = \begin{bmatrix} L_s & L_m \\ L_m & L_r \end{bmatrix}$$

In general, when using PCH representation, the state variables of this representation are not necessarily the best variables for analysis and additional measurement/estimation may be needed in the controller implementation.

Other types of load torque may also be considered in this analysis depending on the nature of the true load applied to the motor (e.g. constant, proportional to squared speed or cubed speed, etc.). In that case a little bit different PCH model will be obtained.

IV. PBC STRATEGIES FOR INDUCTION MOTOR

4.1 Adaptive passivity feedback

The adaptive passivity feedback (APF) scheme applied to the induction motor consists of two loops disposed on cascade, one inner loop to control torque and the outer loop to control velocity. The difference with respect to the standard FOC scheme is that now the inner torque loop does not act on the motor but on its passive equivalent. To obtain the passive equivalent of the induction motor an adaptive feedback developed in [13] is used, which is based on the work developed in [9, 10]. Besides, the application of the novel concept called the Torque-Flux Control Principle (TFCP) [13, 15], notably simplifies the feedback used. The TFCP is enunciated next:

'When operating in a field oriented control scheme the design of torque and flux controllers for the induction motor can be focused only on the control of the direct and quadrature components of stator current. It is useless to make efforts to control in a direct way the rotor flux or current components'. Notice that controller still guarantees a suitable control of torque and flux and it is then possible to disregard in the design all the terms concerning the rotor current or rotor flux. This simplifies the feedback necessary to obtain a passive equivalent for the induction motor.

Using the TFCP, the state feedback used is shown in (19)–(21). For further information see [13, 15] where a state feedback for the case where the TFCP is not used is also developed.

$$\begin{bmatrix} u_{sd} \\ u_{sq} \end{bmatrix} = \theta_1(t) \begin{bmatrix} e_{i_{sd}} \\ e_{i_{sq}} \end{bmatrix} + \theta_3(t) \begin{bmatrix} \varpi_{sd} \\ \varpi_{sq} \end{bmatrix} \tag{19}$$

$$\begin{aligned} \dot{\theta}_1(t) &= \frac{1}{\alpha(t)} \begin{bmatrix} e_{i_{sd}} \\ e_{i_{sq}} \end{bmatrix} [e_{i_{sd}} \;\; e_{i_{sq}}] \Gamma_1^{-1}(t) \\ \dot{\theta}_3(t) &= \frac{1}{\alpha(t)} \begin{bmatrix} e_{i_{sd}} \\ e_{i_{sq}} \end{bmatrix} [\varpi_{sd} \;\; \varpi_{sq}] \Gamma_3^{-1}(t) \end{aligned} \tag{20}$$

$$\begin{aligned} \dot{\Gamma}_1(t) &= -\Gamma_1(t) \begin{bmatrix} e_{i_{sd}} \\ e_{i_{sq}} \end{bmatrix} [e_{i_{sd}} \;\; e_{i_{sq}}] \Gamma_1(t) \\ \dot{\Gamma}_3(t) &= -\Gamma_3(t) \begin{bmatrix} \varpi_{sd} \\ \varpi_{sq} \end{bmatrix} [\varpi_{sd} \;\; \varpi_{sq}] \Gamma_3(t) \end{aligned} \tag{21}$$

$$\Gamma_1(0), \Gamma_3(0) > 0$$

where $e_{i_{sd}}(t) = i_{sd}^*(t) - i_{sd}(t)$, $e_{i_{sq}}(t) = i_{sq}^*(t) - i_{sq}(t)$ are the current errors with respect to their references. $\alpha(t) = \sqrt{1 + \mathrm{Trace}\{\Gamma_1^{-2}(t) + \Gamma_3^{-2}(t)\}}$ corresponds to a normalization factor and ϖ_{sd}, ϖ_{sq} correspond to the new set of inputs to the passive equivalent system. Notice that parameter θ_2 in (8) becomes null when applying the TFCP.

The resulting PBC scheme is shown in Fig. 3.

In the experimental implementation of this strategy the case of fixed adaptive gains ($\Gamma_i(t) > 0$ and constant $\forall t$) will be compared with the case of time-varying adaptive gains ($\Gamma_i(t)$ given by (21)).

For rotor flux control, the technique of field weakening is used. This consists of fixing the flux reference to a constant value (nominal value) for slower mechanical speeds, and weakening the field quadratically with the speed for higher speeds. The reason for using this technique is to maintain the total motor power constant not surpassing the limits of this and then to diminish the product speed-power. The technique of field weakening is summarized in Fig. 4.

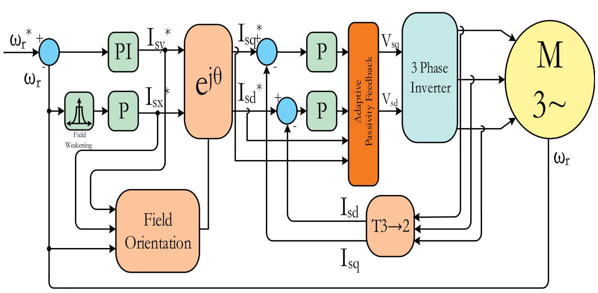

Fig. 3. APF control scheme.

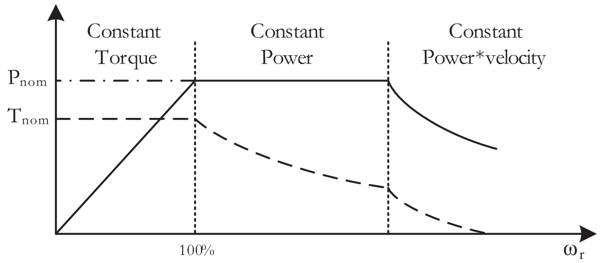

Fig. 4. Field weakening technique.

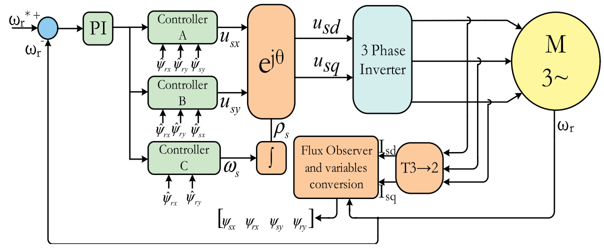

Fig. 5. IDA-PBC control scheme.

4.2 IDA-PBC strategy

The IDA-PBC strategy [6, 7] consists basically of assigning a new storage function to the closed-loop system, besides the possibility of changing the topology of the system, in terms of interconnections and energy transfers between states. For the case of induction motors [18, 19], the controller is defined by some feasible solution for k_1, k_2 and k_3 of the algebraic equation

$$
\begin{aligned}
&L^{-1}x_{12}+\begin{pmatrix} k_1 \\ \dfrac{x_4}{x_2^2+x_4^2}k_3 \end{pmatrix}=0\\
&L^{-1}x_{34}+\begin{pmatrix} k_2 \\ \dfrac{x_4}{x_2^2+x_4^2}k_3 \end{pmatrix}=0 \qquad (22)\\
&J^{-1}x_5+k_3=0
\end{aligned}
$$

From the third equation in (22), it is observed that an equilibrium point ω_r^* exists for ω_r defined as $\omega_r^*=-k_3$. For the other parameters (k_1,k_2) the solutions are given by [18, 19]

$$(k_1^2+k_2^2)L_m^2 \geq 2k_3 L_r B.$$

With the previous result the IDA-PBC controller is defined as

$$
\begin{aligned}
u_{sx}(x) &= -R_s k_1+\left(1+\frac{R_r B'}{x_2^2+x_4^2}\right)x_3 k_3\\
u_{sy}(x) &= -R_s k_2-\left(1+\frac{R_r B'}{x_2^2+x_4^2}\right)x_3 k_3 \qquad (23)\\
\omega_s(x) &= -\left(1+\frac{R_r B'}{x_2^2+x_4^2}\right)k_3
\end{aligned}
$$

States x_2 and x_4 correspond to rotor flux expressed in orthogonal coordinates (ψ_{rx},ψ_{ry}). The rotor flux will be zero if and only if the motor is at rest and without voltage applied. At $t=0$, to control the motor some tension has to be applied and therefore ψ_r becomes different from zero at $t=0$. Thus no undetermined values of the controller are obtained.

The IDA-PBC scheme used in this paper was slightly modified. In principle, this strategy was developed to control the motor speed, not being robust with respect to load perturbations on the motor axis. This means that permanent errors in the mechanical speed were obtained. In order to solve this problem, a simple proportional integral loop was added for the speed error loop modifying the original IDA-PBC, scheme as shown in Fig. 5.

In general the rotor flux can not be measured in the great majority of induction motors, which is the reason why the implementation of a rotor flux observer for the experimental implementation of this strategy was necessary. The observer was implemented based on the voltage-current model of the induction motor, developed in [26–28].

V. EXPERIMENTAL RESULTS

Simulation results have already been illustrated in [21, 22] where these control strategies were compared by means of computational simulations, using the software MATLAB-SIMULINK. These results were obtained previous to the experimental implementation shown here.

In the experimental tests the control strategies were implemented in MATLAB-SIMULINK, using a fixed step of 10 ms and the solver ODE5 (Dormand-Prince). In the electronics, a vector modulation with a carrier frequency of 20 KHz was used.

In this section the experimental results obtained by applying APF and IDA-PBC strategies are presented. The tests carried out for each strategy to compare and analyze the control schemes are described in what follows.

Test 1 (Basic behavior). The speed reference is a ramp starting from zero at $t = 0$ until the nominal speed (146.08 rad/s) in 9 s. The load torque is proportional to the speed and was kept constant equal to the nominal value (100%) during the whole test. Initial conditions (IC) for controller parameters were all set to zero, except time-varying gains initial values that were chosen as $\Gamma_1(0) = \Gamma_3(0) = I$, where I is the 2×2 identity matrix.

Test 2 (Tracking). The speed reference is a ramp starting from rest at zero until the nominal speed (146.08 rad/s) in 9 s. Between $t = 40$ s and $t = 70$ s a pulse train reference of amplitude $0.1\omega_{r_{nom}}$ and frequency $\pi/10$ rad/s was added on top of the constant nominal value. Between $t = 80$ s and $t = 110$ s a sinusoidal reference of amplitude $0.1\omega_{r_{nom}}$ and frequency $\pi/10$ was added on top of the constant nominal value. A load torque proportional to the speed and equals to 50% of nominal value was kept constant during the whole test. IC of controller parameters were all set to zero, except time-varying gains initial values that were chosen as $\Gamma_1(0) = \Gamma_3(0) = I$, where I is the 2×2 identity matrix.

Test 3 (Regulation). The speed reference is a ramp starting from rest at zero until the nominal value (146.08 rad/s) in 9 s, then the reference is kept constant. Initial load torque was equal to 0% of the nominal value. Between $t = 40$ s and $t = 80$ s a torque perturbation equal to 50% of the nominal value is added. All IC were set to zero except time-varying gains initial values that were chosen as $\Gamma_1(0) = \Gamma_3(0) = I$, where I is the 2×2 identity matrix.

The three phase inverter used in the experiments was that designed and built in [18]. Communication to PC was done by means of MATLAB-SIMULINK software using an S-Function properly designed. The induction motor used in the experiments was a Siemens 1LA7080, 0.55 KW, $\cos(\phi) = 0.82$, 220 V, 2.5 A, 4 poles and 1395 RPM. From motor tests (no load and locked rotor) the estimated motor parameters used in the study are those shown in Table I.

In order to apply resistive torque on motor axis, the induction motor was mechanically coupled to a continuous current generator, Briggs & Stratton ETEK, having a permanent magnet field. The load to the generator was applied using a cage of discrete resistances connected to generator stator and manually controlled by switches. The magnitudes of the resistances were chosen such that maximum values of induction motor operation were not exceeded under any circumstances. The experimental assembly including the motor-generator group used in the experimental tests is shown in Fig. 6.

For experimental tests, the best values of PI controller parameters for inner and outer loops were chosen based on those obtained from computer simulations of control strategies making use of mathematical models of the induction motor [18, 22]. Values were first determined by Ziegler-Nichols criteria and modified later by simulations, until a good response was obtained. Taking into account these values, a fine tuning still was necessary in the experiments performing a small number of trail tests. The values finally chosen for the constants of control loops used in the BCS and in APF scheme are shown in Table II.

For the case of the IDA-PBC strategy, the values of constants k_1 and k_2 were determined based on simulations results reported in [22]. The chosen values were $k_1 = k_2 = -30$ and for proportional integral loop it was chosen $K_P = 3$ and $K_I = 0.5$.

The experimental results obtained by applying the techniques under study for Test 1, Test 2 and Test 3 already described are shown next. In Figs 7

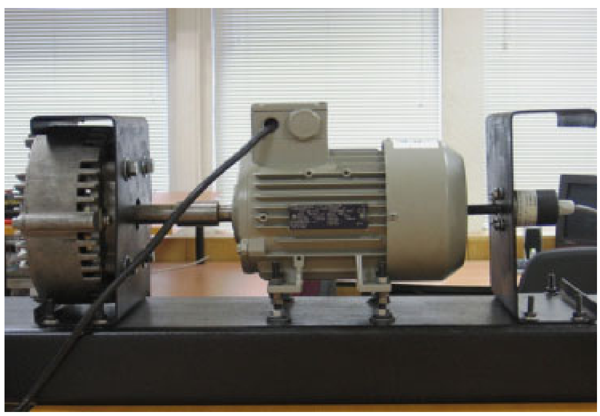

Fig. 6. Experimental assembly.

Table I. Induction motor estimated parameters.

Parameter	Value
R_s	14.7 Ω
R_r	5.5184 Ω
X_s	11.5655 Ω
X_r	11.5655 Ω
X_m	115.3113 Ω

Table II. Values of control loop constants used in experimental evaluation of BCS and APF scheme.

Outer Loop	Inner Loop
$K_P = 0.403$	$K_P = 45$
$K_I = 0.0189$	

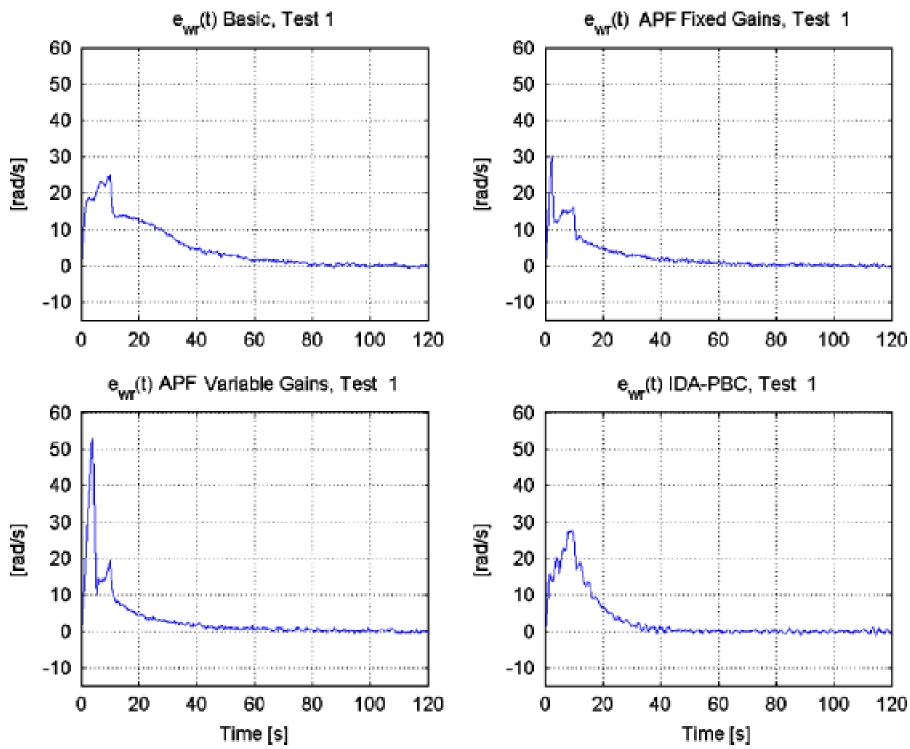

Fig. 7. Speed errors for experimental Test 1 with constant load torque.

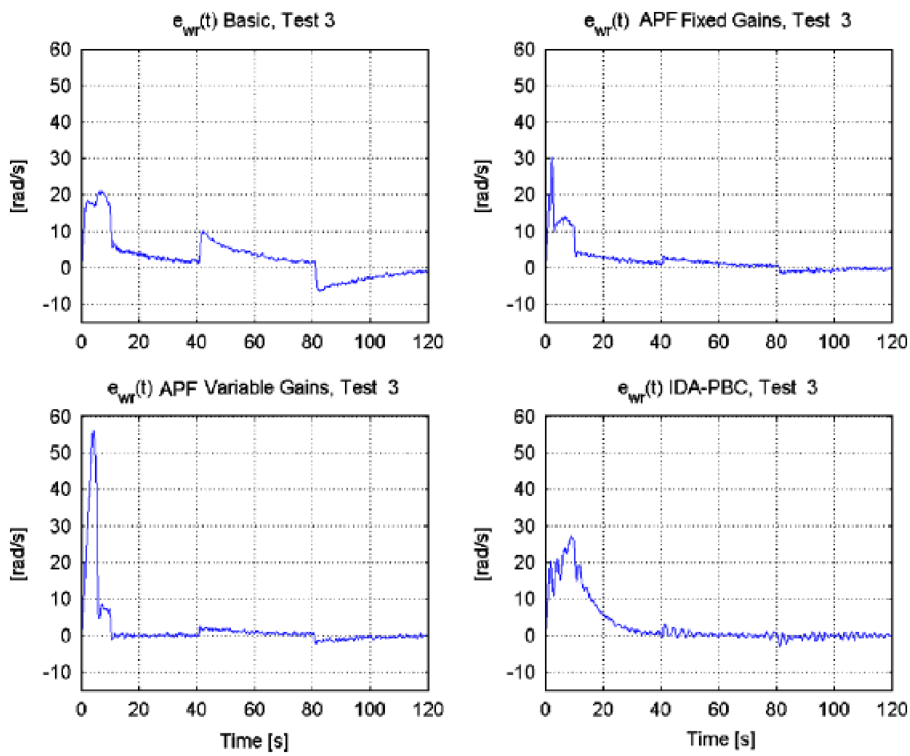

Fig. 8. Speed errors for experimental Test 2 for reference tracking.

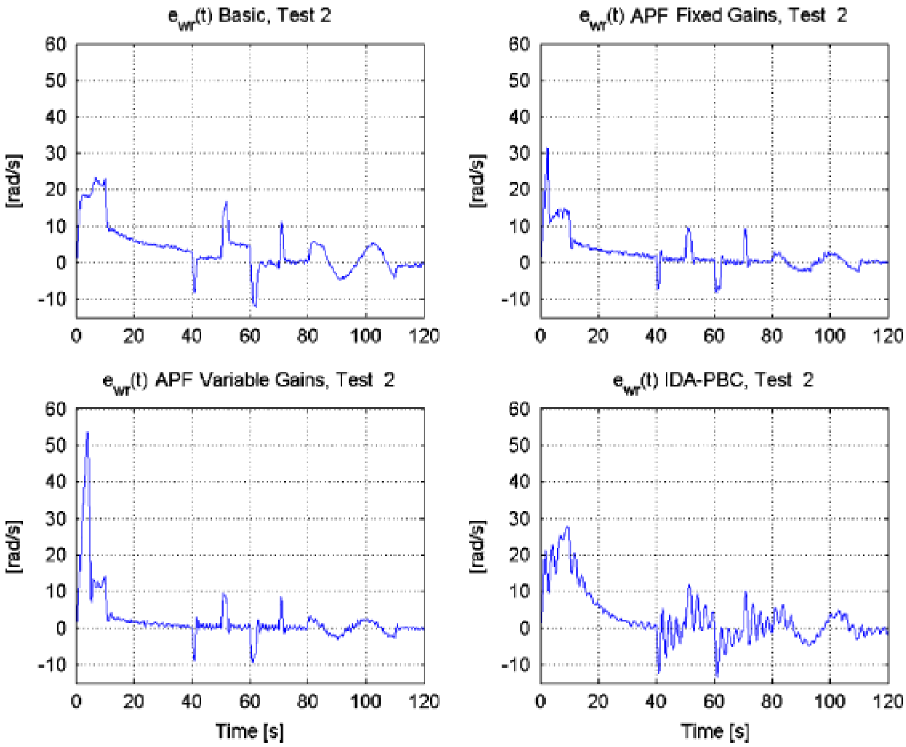

Fig. 9. Speed errors for experimental Test 3 for load torque perturbations.

to 9 the evolution of the speed errors are plotted for each strategy for each one of the three tests. The evolution of the remaining variables is shown in the Appendix.

It is observed from Fig. 7 that the fastest convergence of control error to zero, with a constant nominal load torque applied (Test 1), is obtained for the IDA-PBC strategy, with about 40 s. This has to be compared with 60 s and 80 s obtained with the BCS and APF strategies, respectively. However, in the IDA-PBC strategy an important oscillatory behavior of the control error is observed at the beginning. For more information about the behavior of other variables see Figs A1 to A4 in the Appendix.

From tracking viewpoint (Test 2), the best results are observed for APF strategies, which follow reference changes better than BCS (See Fig. 8). The IDA-PBC strategy is not able to properly follow reference changes, presenting an oscillatory behavior of speed error. This is mainly due to the fact that control error convergence to zero strongly depends on the tuning of outer proportional-integral loop designed for speed control. It is also important to point out that convergence of the control error to zero is influenced by rotor flux observer convergence, which necessarily adds a dynamic to the system affecting the global behavior of the overall system. For information about the evolution of other variables see Figs A5 to A8 in the Appendix.

When applying torque perturbations on motor axis (Test 3), it is observed from Fig. 9 that a quick stabilization is attained by APF strategies, without large oscillations. In the IDA-PBC strategy, however, although perturbations are quickly controlled, the behavior of control error presents oscillations. For the case of BCS, its response is quite slow with a larger error. This strategy is not robust with respect to perturbations on the mechanical subsystem. The evolution of other variables can be seen in Figs A9 to A12 in the Appendix. Numerous other experiments, not shown here for the sake of space, were carried out to analyze the influence of different parameters on BCS, APF and IDA-PCB strategies [21]. In particular the effects of initial conditions on APF strategies were analyzed, as well as the effects of using fixed and time-varying adaptive gains. It was observed, in general, that time-varying gains improve transient behavior and diminish initial control error. The complete results of application of all the strategies presented here can be consulted in [21].

As can be seen from (19)–(21), current signals are needed to implement the AFP adaptive laws.

In simulations, a small noise was added on these signals and the performance of the method was not importantly affected [18]. At experimental level, the influence of the normal noise present in the measurement of current signals during the test did not affect the behavior of the AFP, as observed in Figs 7–9. If the noise level is expected to be large some deterioration of the control system will certainly be observed. In this case robust adaptive laws should be used instead of the standard adaptive laws. It is recommended in this case to use the $\sigma\theta$-modification [29] to get good results.

VI. CONCLUSIONS

From experimental analysis performed on induction motor control some interesting conclusions can be drawn. In APF strategies an important simplification of the control scheme based on the FOC principle can be attained when using the TFCP, allowing an effective control of the system without the necessity of implementing a rotor flux observer to orientate the field. For APF strategy the use of time-varying adaptive gains noticeably improves the transient behavior of controlled system, both for tracking as well as for regulation.

In the case of an energy shaping strategy, IDA-PBC, a novel control scheme was studied and implemented. Since the original strategy was only designed for speed control, the addition of an outer speed loop of proportional-integral type, allowed certain robustness to be obtained with respect to torque perturbations. In this strategy it was necessary for the design and implementation of a rotor flux observer, adding a certain complexity to the complete system.

Since the BCS has fixed controller parameters its behavior is not as good as the adaptive strategies studied.

In conclusion, the two adaptive strategies studied present clear advantages with respect to the BCS used as basis of comparison. Amongst the adaptive schemes, the APF with time-varying adaptive gains is the one that behaves better.

APPENDIX A

The following shows, in detail, the evolution of electrical and mechanical variables, beside the speed error already discussed. For each of the four control strategies studied all main variables are plotted when each one of the three tests is applied to an induction motor.

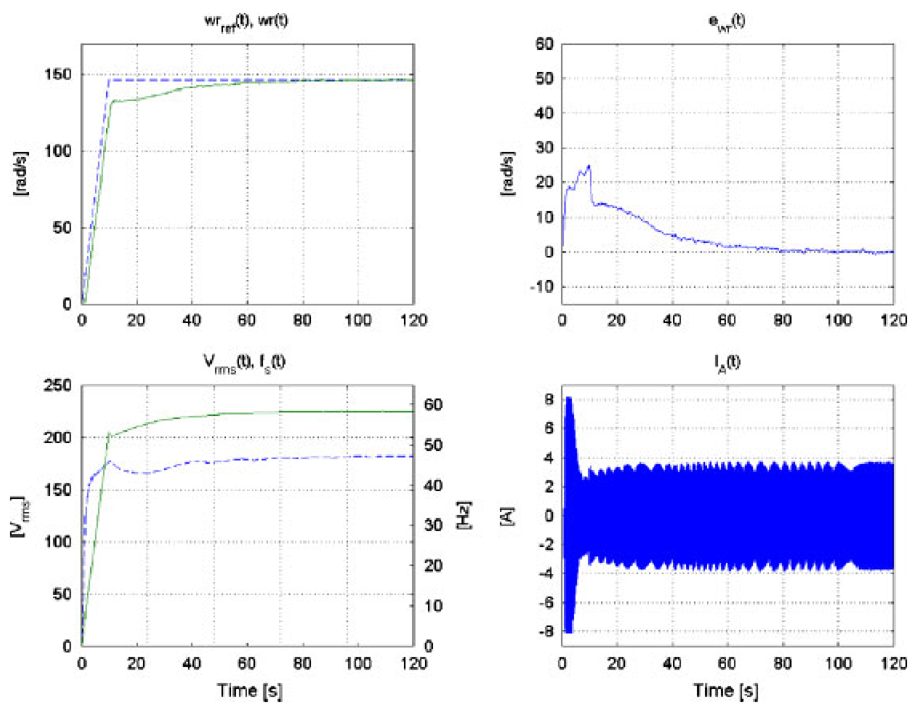

Fig. A1. Experimental results for BCS. Test 1, constant torque.

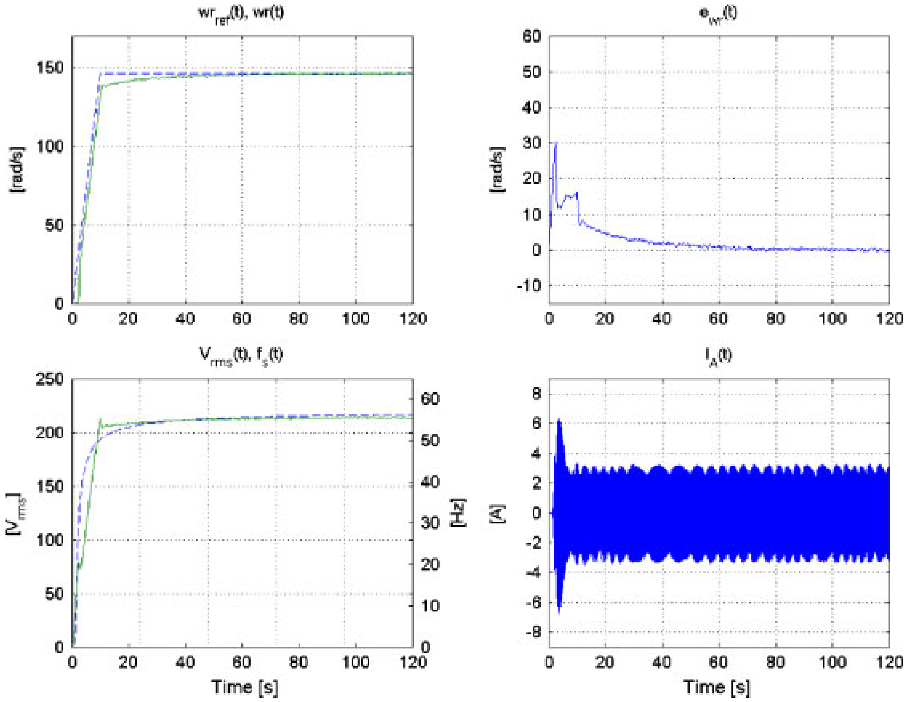

Fig. A2. Experimental results for PBC fixed gains strategy. Test 1, constant torque.

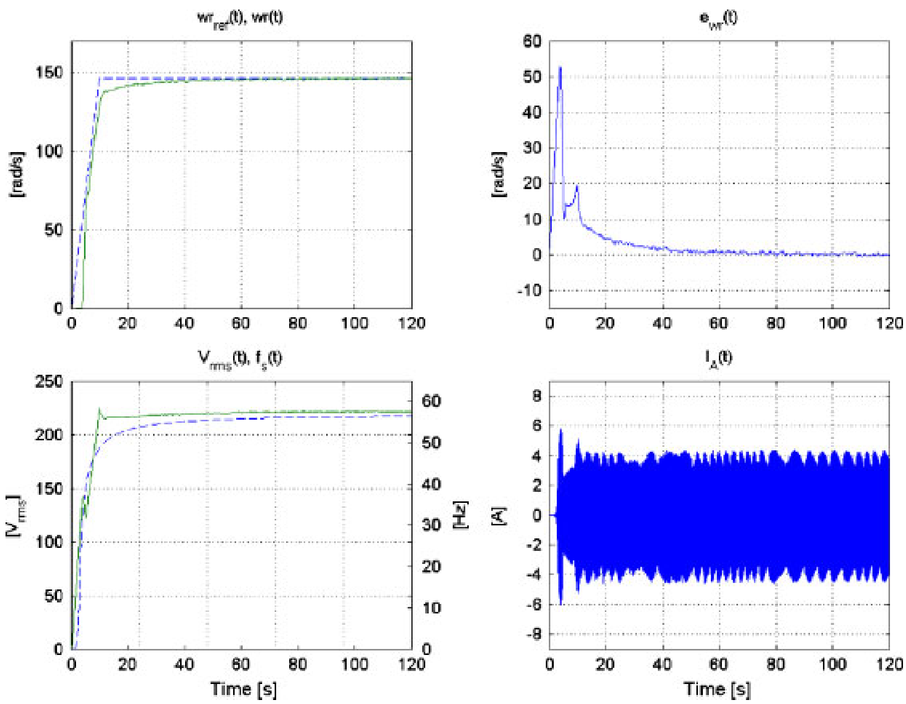

Fig. A3. Experimental results for PBC variable gains strategy. Test 1, constant torque.

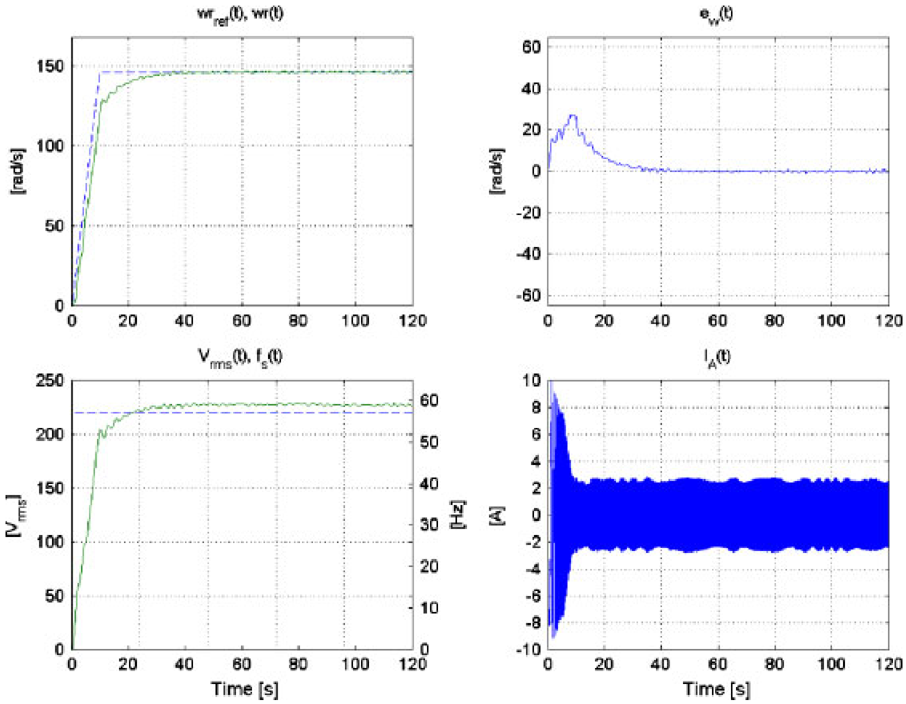

Fig. A4. Experimental results for IDA-PBC strategy. Test 1, constant torque.

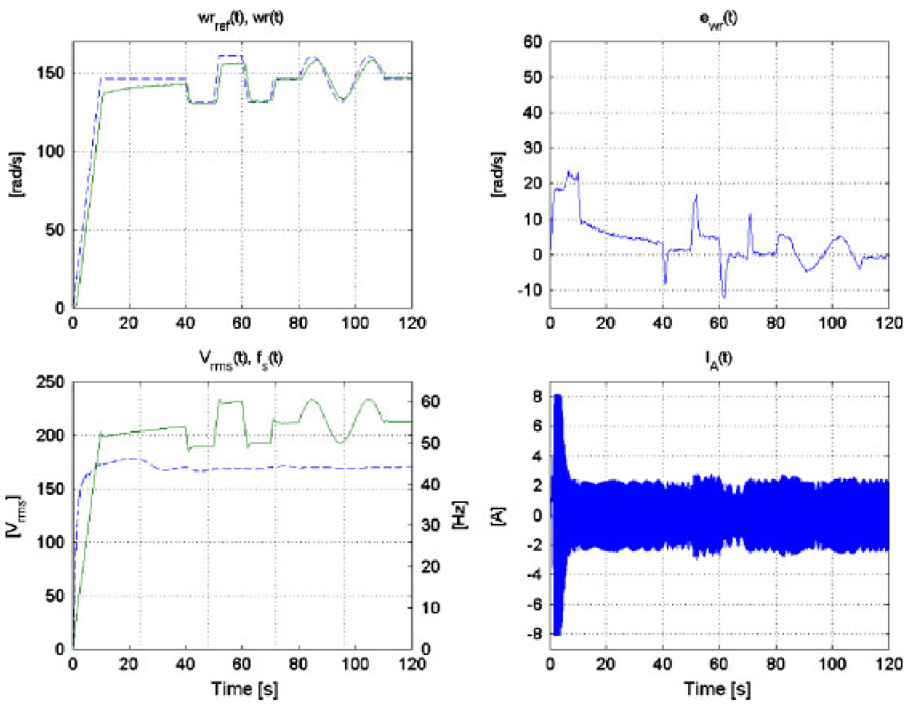

Fig. A5. Experimental results for BCS. Test 2, reference tracking.

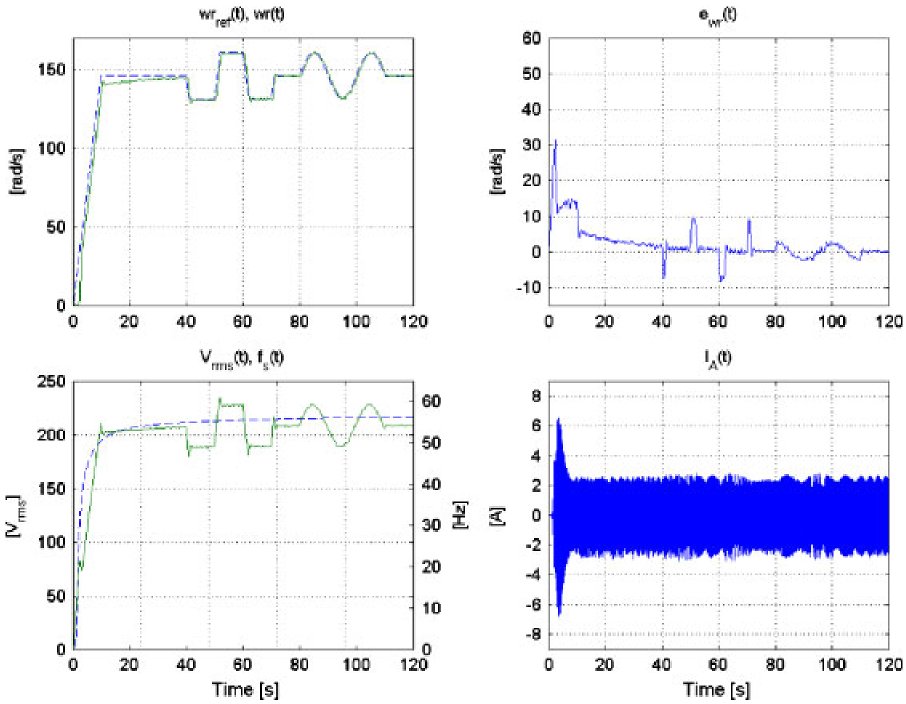

Fig. A6. Experimental results for PBC fixed gains strategy. Test 2, reference tracking.

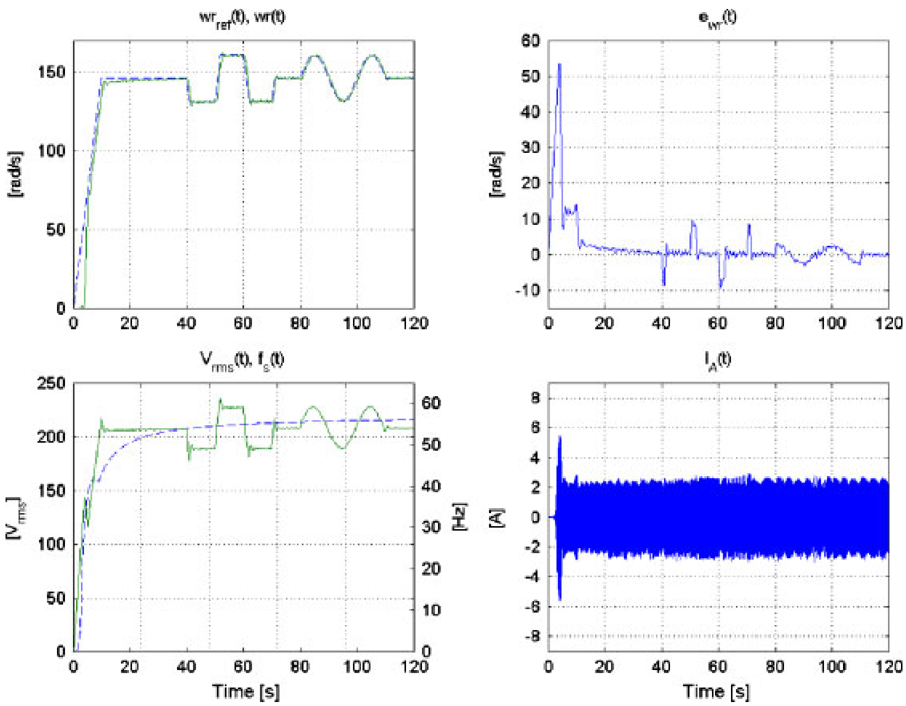

Fig. A7. Experimental results for PBC variable gains strategy. Test 2, reference tracking.

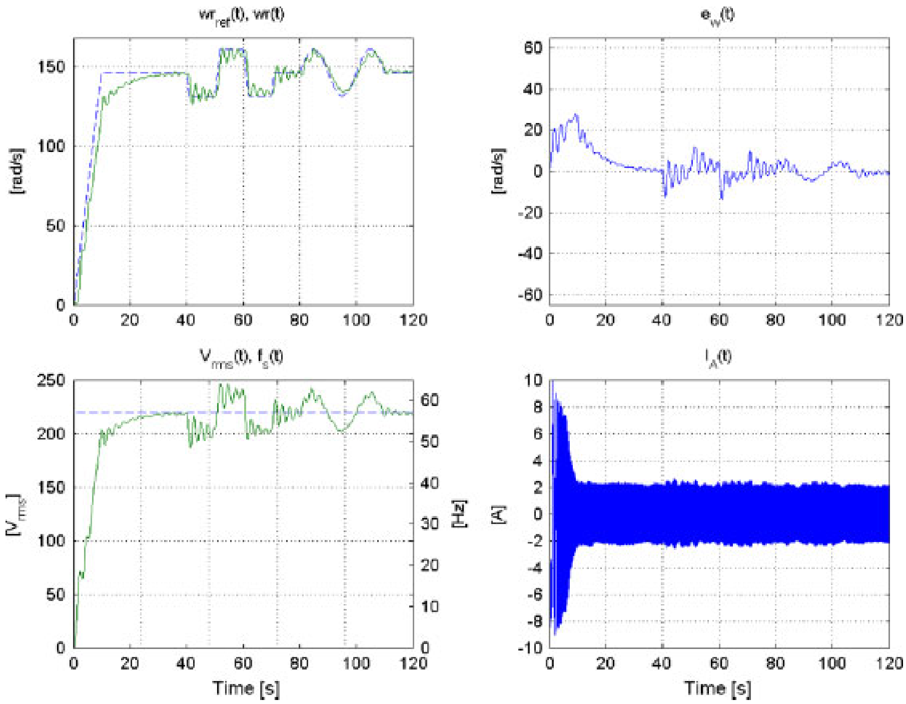

Fig. A8. Experimental Results IDA-PBC Strategy Test 2, Reference Tracking.

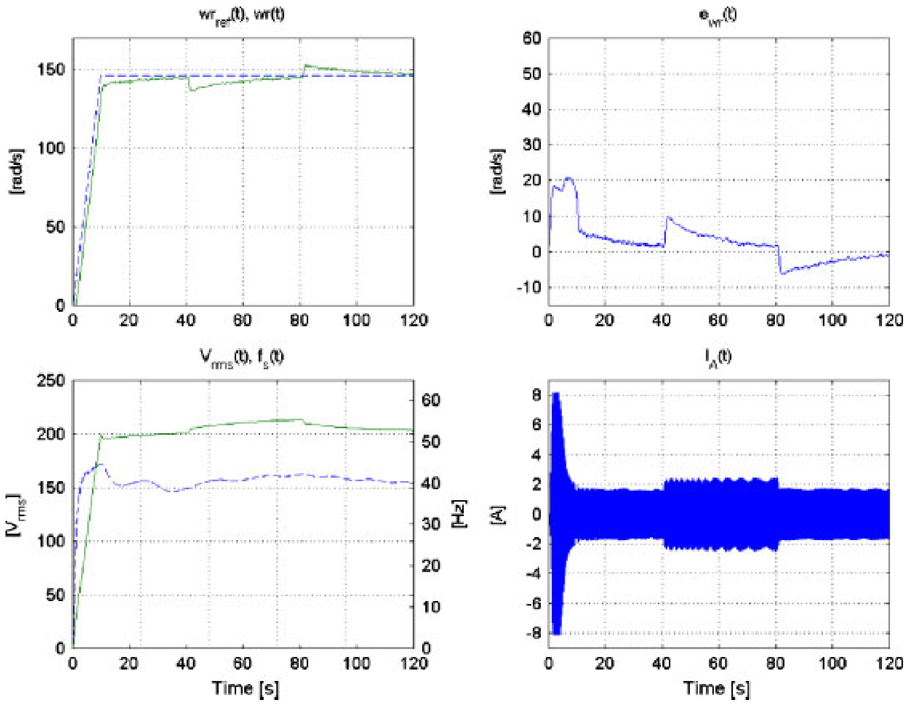

Fig. A9. Experimental results for BCS. Test 3, torque variations.

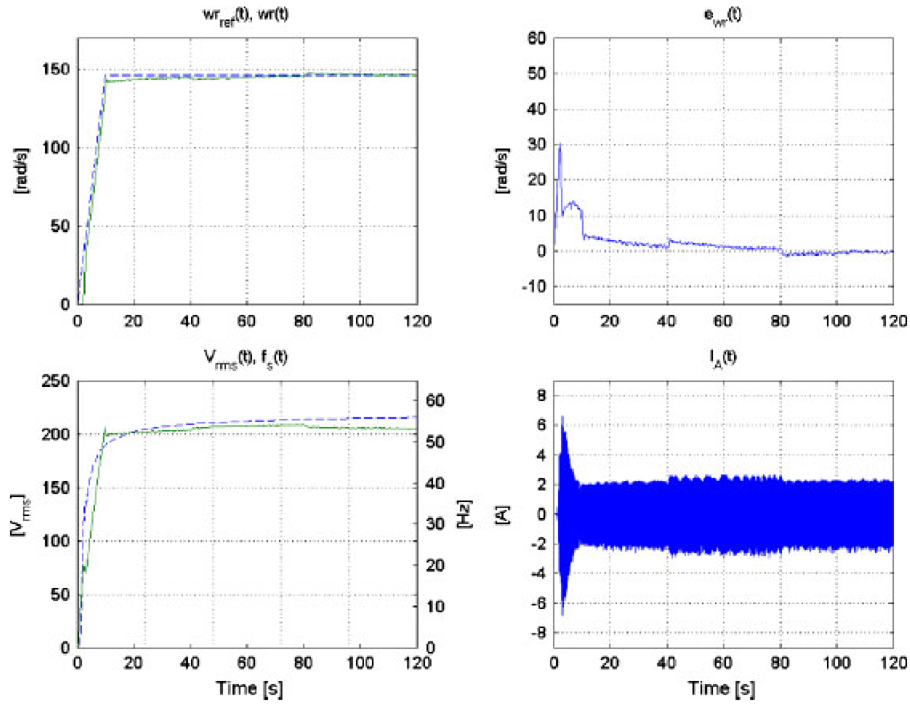

Fig. A10. Experimental results for PBC fixed gains strategy. Test 3, torque variations.

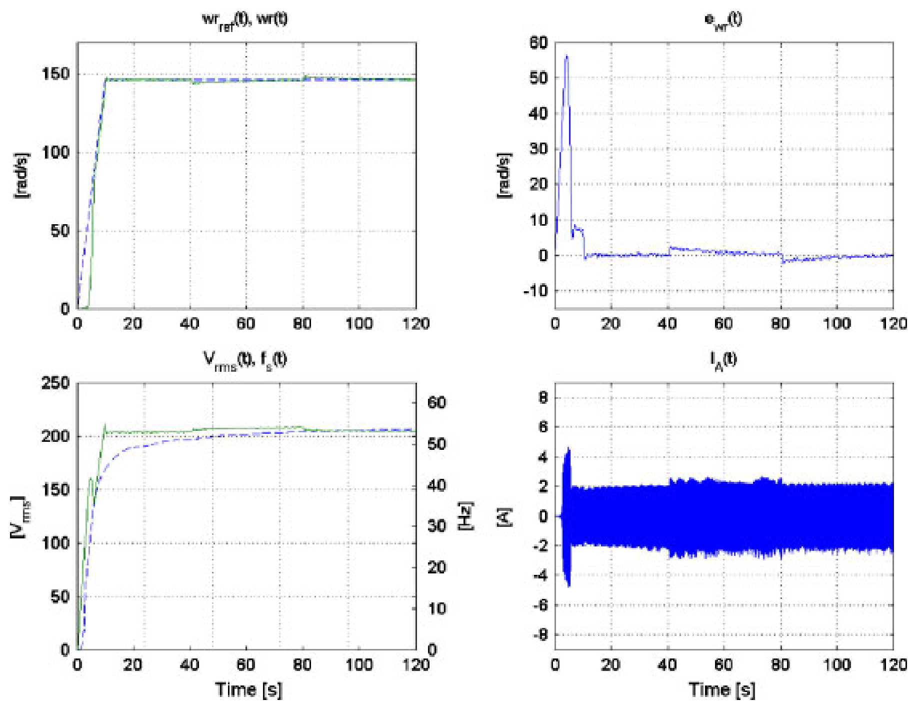

Fig. A11. Experimental results for PBC variable gains strategy. Test 3, torque variations.

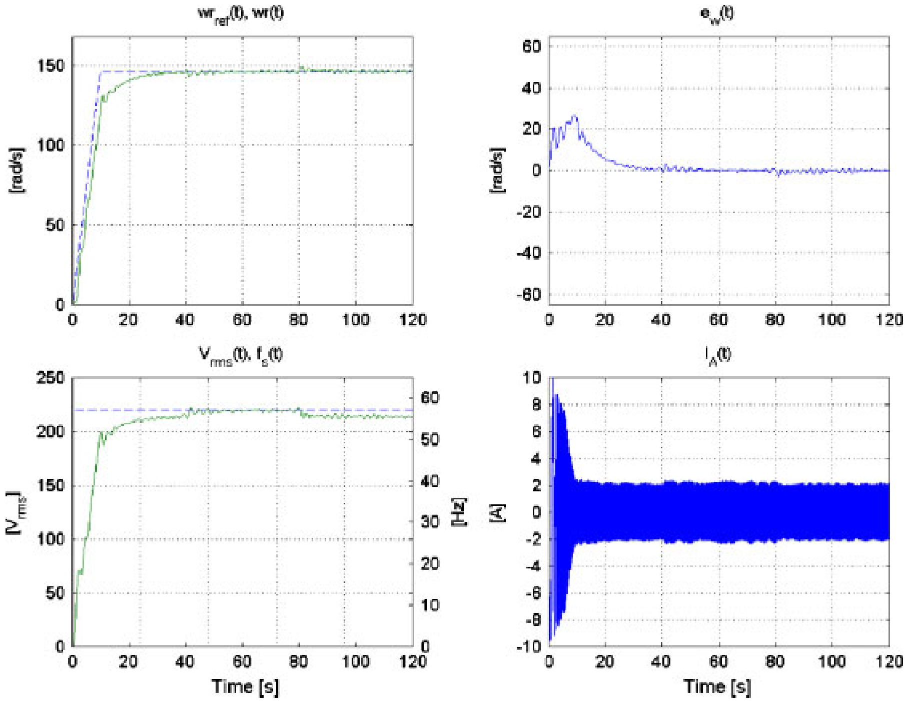

Fig. A12. Experimental results for IDA-PBC strategy. Test 3, torque variations.

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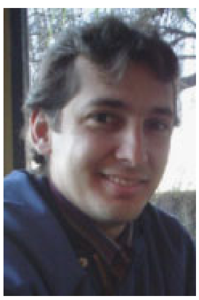

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