Global adaptive linear control of the permanent-magnet synchronous motor
Antonio Loria, G. Espinosa-Pérez, Sofía Avila-Becerril

To cite this version:

HAL Id: hal-00831451
https://hal.archives-ouvertes.fr/hal-00831451
Submitted on 7 Jun 2013

HAL is a multi-disciplinary open access archive for the deposit and dissemination of scientific research documents, whether they are published or not. The documents may come from teaching and research institutions in France or abroad, or from public or private research centers.

L’archive ouverte pluridisciplinaire HAL, est destinée au dépôt et à la diffusion de documents scientifiques de niveau recherche, publiés ou non, émanant des établissements d’enseignement et de recherche français ou étrangers, des laboratoires publics ou privés.
Global Adaptive Linear Control of the Permanent-Magnet Synchronous Motor

Antonio Loria\textsuperscript{a}, Gerardo Espinosa–Pérez\textsuperscript{b}\textsuperscript{*} and Sofía Avila–Becerril\textsuperscript{b}

\textsuperscript{a} CNRS, at LSS–Supelec, 3 Rue Joliot Curie, Gif sur Yvette France. E-mail: loria@lss.supelec.fr

\textsuperscript{b} FI – UNAM, A.P. 70-256, 04510 México D.F., MEXICO. gerardoe@unam.mx, soavbec@gmail.com

SUMMARY

We contribute with a linear time-varying controller for the permanent magnet synchronous motor. We solve the open problem of speed-tracking control by measuring only stator currents and the rotor angular positions, under parametric uncertainty. Integral action is used to compensate for the effects of the unknown load-torque and adaptation is employed to estimate the unknown parameters. In the case that parameters are known (except for the load) we show that the origin of the closed-loop system is uniformly globally exponentially stable. For the case of unknown parameters we prove uniform global asymptotic stability hence, we establish parametric convergence. In contrast to other adaptive control schemes for electrical machines, we use a reduced-order adaptive controller. Indeed, adaptation is used only for the electrical dynamics equations. Moreover, not surprisingly, the closed-loop system has a structure well-studied in adaptive-control literature. Performance is illustrated in a numerical setting. Copyright © 2012 John Wiley & Sons, Ltd.

1. INTRODUCTION

Control of electrical machines has been deeply studied both from a practical and from a theoretical perspective for a long time –see \cite{15}, \cite{22}. On one hand, field-oriented control is the preferred scheme in industrial applications, due to its structure based on nested proportional-integral loops \cite{11}. On the other hand, academic contributions which have better captured the attention of practitioners, are those obtained from a passivity-based perspective \cite{17} and those using feedback linearization \cite{5}. In \cite{7} the authors present an injection-and-damping-assignment controller for the permanent-magnet synchronous motor in Hamiltonian coordinates; the control design is carried out following a procedure designed for general Hamiltonian systems to which integral action is added. In \cite{25} the authors use a feedback linearizing controller and then, design a Luenberger-type observer and apply a certainty-equivalence controller; (local) asymptotic stability is established via Lyapunov’s first method. In \cite{19} it has been shown that there exists a downward compatibility between a passivity-based control for induction motors and its corresponding field-oriented control. Some proportional-integral control tuning rules for field-oriented control of induction motors have been proposed in \cite{4} by exploiting its passivity properties, and in \cite{14} a feedback linearization controller was proposed based on the stability properties of field-oriented control of induction motors. The article \cite{24} presents a locally exponentially stabilizing controller without velocity nor

\textsuperscript{*}Correspondence to: FI – UNAM, A.P. 70-256, 04510 México D.F., MEXICO. gerardoe@unam.mx.

\textsuperscript{†}Part of this work was developed during a sabbatical stay of G. Espinosa–Perez at LSS–SUPELEC, France sponsored by Fondation Supelec. Part of this work was supported by DGAPA–UNAM (IN111211).
position measurements however, it relies on parametric knowledge (except for the load torque) and internal viscous friction.

Control of the permanent-magnet synchronous motor under parametric uncertainty and partial state feedback has also been addressed via a range of complex nonlinear control approaches – see [23, 3] for sensorless schemes and [2, 8] for adaptive controllers. The authors of [2] solved a problem of adaptive control without velocity measurements and established asymptotic convergence of tracking errors (for induction motors). In [8] an adaptive controller for permanent-magnet synchronous motors is proposed using measurement of currents and positions only. Although in [8] exponential stability is claimed, it relies on a condition of persistency of excitation along the closed-loop trajectories which is (clearly) impossible to verify. Furthermore, as the main proof in [8] relies on tools for linear time-varying systems, it is implicitly required that the closed-loop trajectories are bounded. Although this is proved and not assumed, exponential stability can be established only on compact subsets of the state-space that is, it may be established that the origin is globally asymptotically stable and locally exponentially stable. Since the analysis does not establish that stability and convergence are uniform in the initial times robustness cannot be guaranteed\(^\dagger\).

In this note we solve the speed-tracking problem without velocity measurement and under parametric uncertainty, for permanent-magnet synchronous motors via (adaptive) PID control. Our controller is composed of two parts conceived separately: on one hand, a PI\(^2\)D controller –see [16], for the rotor dynamics and on the other, a linear time-varying “tracking” for the stator current dynamics. The PI\(^2\)D controller consists in a proportional (to the position error) term, a derivative term in which velocities are replaced by approximate differentiation and integral feedback of the position errors and the approximate derivative. The integral action, and not adaptive control, compensates for the effects of the unknown constant torque-load.

The application of PI\(^2\)D control for the permanent-magnet synchronous motor relies on the structural properties of the machine –mainly passivity. Indeed, the PI\(^2\)D controller forms an outer control loop which acts on the rotor (the mechanical part of the machine). This control law enters as a virtual input through a reference trajectory purposely designed for one of the stator currents. An inner loop is composed of control laws to drive the stator currents to the appropriate operating point.

We address two cases: first, with known parameters but unknown constant load-torque and second, with all unknown parameters. In the first case we establish uniform global exponential stability. In the second we establish uniform global asymptotic stability under a persistency of excitation condition on (a function of) the reference trajectories and for sufficiently large control gains (independent of the initial conditions). The technical tools to establish our main results are tailored for nonlinear time-varying systems –see [20]; in particular, we establish that the unknown parameters are asymptotically estimated if and only if the aforementioned persistency-of-excitation condition holds.

We emphasize that for non-autonomous systems the uniform stability properties that we establish guarantee local input to state stability hence, our main results supersede others in the literature by guaranteeing robustness to bounded additive disturbances. Furthermore, using converse Lyapunov functions, our main results may be extended to address the sensorless problem via certainty-equivalence output feedback. Last but not least, our controllers (PID control for the mechanical variables and linear time-varying for the electrical part) are comparably much simpler than others in the literature.

The rest of the paper is organized as follows. In the following section we formulate the problem and discuss its solution, under the assumption that the parameters (except load torque) are known; a formal statement is made in Section 3. Then, this result is extended to the case of unknown parameters in Section 4. Numerical simulations are presented in Section 5. We conclude with some remarks in Section 6.

\(^\dagger\)This is a drawback encountered in numerous articles on Model Reference Adaptive Control (MRAC) where classic linear systems theory is inappropriately used to analyze nonlinear systems. The readers are invited to see [20] for further discussions.
2. PROBLEM FORMULATION AND ITS SOLUTION

2.1. The model

Consider the well–known dq model of the non–salient permanent-magnet synchronous motor – [5, 21].

\[
\begin{align*}
L \frac{di}{dt} &= -Ri - \omega \Phi J \dot{\theta} - \omega JL i + U \\
J \dot{\omega} &= n_p \Phi i_q - \tilde{\tau}_L \\
\dot{\theta} &= \omega
\end{align*}
\]

where \(i = [i_d, i_q]^T\) and \(U = [u_1, u_2]^T\) are the stator currents and voltages respectively, \(\theta\) and \(\omega\) are the mechanical (position and speed) variables, \(\tilde{\tau}_L\) is the load torque, \(L\) is the proper inductance of the stator windings, \(R\) corresponds to the stator resistance, \(\Phi\) is the magnetic field, \(J\) is the moment of inertia and \(n_p\) is the number of pole pairs and

\[
J = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}, \quad \varrho = \begin{bmatrix} 1 \\ 0 \end{bmatrix}.
\]

For the purpose of control design and analysis we introduce the state \(x \in \mathbb{R}^4, x = [x_1, x_2, x_3, x_4]\) with \([x_1, x_2] = i, x_3 = \omega, x_4 = \theta\) and the parameters

\[
\tau_L = \frac{\tilde{\tau}_L}{J}, \quad \sigma = \frac{n_p \Phi}{J}.
\]

Then,

\[
\begin{align*}
L \dot{x}_1 &= -Rx_1 + Lx_2 x_3 + u_1 \quad (1a) \\
L \dot{x}_2 &= -Rx_2 - Lx_1 x_3 - \Phi x_3 + u_2 \quad (1b) \\
\frac{\dot{x}_3}{\sigma} &= \frac{x_2 - \frac{\tau_L}{\sigma}}{x_2} \quad (1c) \\
\dot{x}_4 &= x_3 \quad (1d)
\end{align*}
\]

Given any rotor angular velocity reference \(x_3^* : \mathbb{R}_{\geq 0} \to \mathbb{R}\), twice differentiable, bounded and with bounded derivatives (almost everywhere), the control goal is to design a dynamic position-feedback controller for the system (1) with measurable states \([x_1, x_2, x_4]^T\) and unknown (constant) perturbation \(\tau_L\), such that

\[
\lim_{t \to \infty} (x_3 - x_3^*) = 0
\]

and Lyapunov stability is ensured.

2.2. Ideal state-feedback control

The control design method is reminiscent of backstepping control; it also exploits the structure of the model (1) and the natural properties of the motor, passivity in particular. Firstly, we regard \(x_2\) as a virtual control input to Equation (1c) and introduce the PI\(^2\)D controller for the mechanical dynamics (1c), (1d). The PI\(^2\)D controller in closed loop with the rotor dynamics defines a passive map. Then, we design a tracking controller \((u_1, u_2)\) for the electrical dynamics i.e., Equations (1a) and (1b).

For clarity of exposition we depart from an ideal control scheme that could be used if all the parameters were known and the velocities were available from measurement. Although these are restrictive conditions which we do not assume to hold in our main statements, by describing such scenario we identify the difficulties that arise in the context of angular-position feedback control and parametric uncertainty.
Consider Equations (1c), (1d) and let us regard $x_2$ as a virtual control input. Let

$$\begin{align*}
x_2^* &= \frac{\tau_L}{\sigma} + \frac{\dot{x}_2^*}{\sigma} + v_3 \\
v_3 &= -k_pe_4 - k_de_3, \quad k_p, k_d > 0
\end{align*}$$

(2a)

where $x_2^*$ is the velocity reference, hence $\dot{x}_2^* = x_2^*$; let the tracking errors be defined as $e_i = x_i - x_i^*$. Using $x_2 = e_2 + x_2^*$ in (1c) and subtracting $\dot{x}_2^* = x_3^*$ from (1d) we obtain

$$\begin{align*}
\frac{1}{\sigma}\dot{e}_3 &= -k_pe_4 - k_de_3 + e_2 \\
\dot{e}_4 &= e_3
\end{align*}$$

(3a) (3b)

whose origin is globally exponentially stable for any positive $k_p$ and $k_d$, provided that $e_2 = 0$. Moreover, this system is input to state stable from the input $e_2$ and defines an output-strictly-passive map $e_2 \mapsto e_3$. In other words, $x_2^*$ stabilizes (1c), (1d) at the equilibrium $^1 x_4 = x_4^*$ and $x_3 = x_3^*$.

With this in mind, we proceed to design the second part of the controller, in order to steer $e_2 \to 0$ asymptotically. Under ideal conditions (parameter knowledge and state measurement feedback) this is achieved via the control laws

$$\begin{align*}
u_1^* &= Rx_3^* + L\dot{x}_1^* - Lx_2x_3^* - k_1'e_1, \quad k_1' > 0 \\
u_2^* &= \Phi x_3^* + Lx_1x_3^* + Rx_2^* + L\dot{x}_2^* - k_2'e_2, \quad k_2' > 0
\end{align*}$$

(4a) (4b)

with $x_3^*$ as in (2a) and $x_1^*$ being a continuously differentiable function, given as reference for $x_1$; typically, $x_1^*$ is chosen constant. Under these conditions the closed-loop electrical system is

$$\begin{align*}
L\dot{e}_1 - Lx_2e_3 + Re_1 + k_1'e_1 &= 0 \\
L\dot{e}_2 + Lx_1e_3 + \Phi e_3 + Re_2 + k_2'e_2 &= 0.
\end{align*}$$

(5a) (5b)

The interest of these equations relies in their structure; notice that, defining $e_{12} := [e_1 \ e_2]^T$, the previous equations have the form

$$\dot{e}_{12} = Ae_{12} + B(e_3)e_{12} + Ce_3$$

(6)

where the poles of $A$ depend on the control gains $k_1'$ and $k_2'$ hence it may be easily rendered Hurwitz, $B(e_3)$ is a skew-symmetric matrix such that $B(0) = 0$ and $C$ depends on $x_1^*$ and $x_3^*$. Therefore, using the quadratic Lyapunov function $V = |e_{12}|^2$ one may easily establish global exponential stability for

$$\dot{e}_{12} = Ae_{12} + B(e_3)e_{12}.$$ 

Furthermore, provided that $C$ in (6) is bounded, exponential stability of $e_{12} = 0$ is conserved if $e_3 \to 0$ exponentially. The latter may be established invoking output-injection arguments—see [1]. In addition to this, input-to-state stability of (6) with input $e_3$ may be established using Lyapunov theory. Thus, stability of the origin of the overall closed-loop equations (3), (5) follows invoking small-gain arguments. The proof of these claims follows as a corollary of our results, presented in Section 3 below.

Besides the stability properties that it ensures, the controller (2)–(4) is attractive due to its simplicity; note that the right-hand side of (2) corresponds to a simple Proportional Derivative controller with load compensation and a feedforward term, while the control laws (4) are linear time-varying and ensure tracking control for the stator current dynamics with reference $x_2^*$. However, simplicity comes at the price of conservatism; on one hand, for the mechanical part (2) the controller requires the exact value of load-torque as well as velocity measurements. On the other hand, to compute $u_2^*$ in (4a) one requires the time derivative of $x_2^*$ in (2a). In the next section we relax these conservative assumptions and present a modification of the state-feedback controller previously discussed. The controller is of PID type and the velocities are replaced with approximate derivatives.

---

1Note that the definition of $x_2^*$ is of little importance in regards to the control objective.
2.3. Position–feedback control with unknown load–torque

Inspired by the ideal structure (2) we redefine the virtual rotor controller as

\[
\begin{align*}
x_2^* &:= \nu + \frac{\dot{x}_3^*}{\sigma} + v_3 \\
v_3 &= -k_pe_4 - k_d\vartheta, \quad k_p, k_d > 0.
\end{align*}
\]

(7a)
(7b)

As before, \(v_3\) is of Proportional “Derivative” type except for the fact that the unavailable velocity \(x_3\) is replaced by the variable \(\vartheta\) which is the output of the approximate-differentiation filter,

\[
\dot{q}_c = -a(q_c + be_4) \quad (8a)
\]
\[
\vartheta = q_c + be_4, \quad a, b > 0 \quad (8b)
\]

whose output \(\dot{\vartheta}\) satisfies

\[
\dot{\vartheta} = -a\vartheta + be_3. \quad (9)
\]

The effect of the load-torque, that is the constant perturbation \(\tau_L/\sigma\) in (1c), is compensated for by integral action that is, \(\nu\) in (7a) is defined as

\[
\dot{\nu} = -k_i(e_4 - \vartheta) \quad (10)
\]

where \(k_i\) is a constant positive gain. Thus, the uncertainty in the load-torque can be coped with by modifying the current reference \(x_2^*\).

With the redefinition of the reference \(x_2^*\) the stator control laws (4) would remain unchanged if not because the new definition of \(x_2^*\), in (7a), implies that

\[
\dot{x}_2^* = [\dot{\nu} + \frac{\dot{x}_3^*}{\sigma} + ak_d\vartheta] - (k_p + bk_d)e_3 \quad (11)
\]

where (recalling that the load-torque is assumed constant) we introduced \(\dot{\nu} = \nu - \tau_L/\sigma\). From here, it is clear that \(\dot{x}_2^*\) may not be part of \(u_2^*\) since it depends on the unmeasurable velocity errors \(e_3\).

Therefore, in the control implementation we shall only use the first three terms in brackets on the right-hand-side of (11), that is, let

\[
\begin{align*}
u_2^* &= \Phi x_3^* + Lx_1x_3^* + Rx_2^* + v_2 + \rho - k_2^e_2 \\
\rho &= L[\dot{\nu} + \frac{\dot{x}_3^*}{\sigma} + ak_d\vartheta] \\
v_2 &= -\varepsilon(e_4 - \vartheta).
\end{align*}
\]

(12a)
(12b)
(12c)

The role of \(v_2\) shall become clear from the proof; it corresponds to a Lyapunov-redesign term –see [10] which is added to enhance negative semi-definiteness of the time derivative of a Lyapunov function candidate –see next section.

Remark 2.1

It is important to stress that contrary to other works in the literature, we do not use adaptive control to estimate the constant load-torque. The controller relies on the much simpler approach, commonly used in control practice, of compensating the perturbation via integral action. The fact that velocities are replaced by dirty derivatives, requires that the Proportional Integral controller involves integration of angular positions \(e_4\) and filter outputs \(\vartheta\). For simplicity, we use a unique gain for both integrators. See [18] for an analysis of the PI²D controller from a passivity viewpoint.

3. PI²D CONTROL WITH KNOWN PARAMETERS

Proposition 3.1

Consider the system (1) in closed loop with \(u_1 = u_1^*, u_2 = u_2^*,\) given by Equations (4a), (12), (7), (8) and (10). The origin of the closed-loop system with state \(\varsigma := [e_1 e_2 e_3 e_4 \vartheta \dot{\nu}]^T\), is uniformly globally exponentially stable, for sufficiently large gains.
The proof is established in three essential steps. Firstly, we derive the closed-loop equations then, we introduce a Lyapunov function and finally, we evaluate uniform integrability conditions of the trajectories, which lead to the conclusion of exponential stability. More precisely, according to [20, Lemma 3] uniform global exponential stability is equivalent to the existence of constants $c_1$, $c_2 > 0$, $p \in [1, \infty)$ such that for all $t_0 \in \mathbb{R}_{\geq 0}$, $x_0 \in \mathbb{R}^n$, all solutions $x(\cdot, t_0, x_0)$ satisfy the

\begin{align}
\text{uniform } L_\infty \text{ bound: } & \quad \sup_{t \geq t_0} |x(t)| \leq c_1 |x_0|; \\
\text{uniform } L_p \text{ bound: } & \quad \left( \lim_{t \to \infty} \int_{t_0}^{t} |x(t)|^p \right)^{1/p} \leq c_2 |x_0|.
\end{align}

Note that the previous conditions are reminiscent of boundedness and $p$-integrability, commonly used in adaptive control to establish (using Barbalăt’s lemma) convergence to zero. The bounds (13) are more conservative since they are uniform in the initial times and have linear growth in the normalized initial states however, they imply uniform global exponential stability which clearly is a much stronger property than convergence. The essence of the proof consists in verifying the inequalities (13) for $\varsigma(t)$.

3.1. The closed-loop equations

From (11) we see that $\rho = L [\dot{x}_2^* + (k_p + bk_d)e_3]$. Using this in (12a) and $u_2^*$ (as defined in the latter) in (1b) we obtain

\begin{align}
L \dot{e}_1 & = -k_1 e_1 + L x_2 e_3; \quad k_1 = k'_1 + R \\
L \dot{e}_2 & = -k_2 e_2 + \Delta e_3 + v_2; \quad k_2 = k'_2 + R \\
\Delta & = L(k_p + bk_d) - (\Phi + L x_1).
\end{align}

The introduction of $\Delta$ is motivated by the possibility of writing the equations as a nominal $e_{12}$–dynamics perturbed by the ‘input’ $e_3$, as done in Section 2.2.

On the other hand, the dynamics equation for the rotor speed can be written as

\[ \frac{1}{\sigma} (\dot{x}_3 - \dot{x}_3^*) = x_2 - \frac{\tau L}{\sigma} - \frac{\dot{x}_3}{\sigma} + x_2^* - x_2^* \]

which is equivalent to

\[ \frac{1}{\sigma} \dot{e}_3 = e_2 + \tilde{\nu} - k_p e_4 - k_d \theta. \]

Furthermore, we define

\[ z = \tilde{\nu} - \frac{k_i}{\varepsilon} e_4, \quad 0 < k_i < \varepsilon \ll 1 \]

\[ k'_p = k_p - \frac{k_i}{\varepsilon} \]

so that

\[ \frac{1}{\sigma} \dot{e}_3 = e_2 - k'_p e_4 - k_d \theta + z \]

\[ \dot{z} = -k_i (e_4 - \theta) - \frac{k_i}{\varepsilon} e_3. \]

Let $x = [e_1, e_2, e_3, e_4, \theta, z]^\top$; note that stability of $\{x = 0\}$ is equivalent to that of $\{\varsigma = 0\}$. Thus, in the sequel we seek to establish (13) for the trajectories generated by Equations (14), (15), (20), (21) and $\dot{e}_4 = e_3$. 

3.2. Stability analysis

Consider the following functions which are proposed after [16],

\[
V_1 = \frac{L}{2}(e_1^2 + e_2^2) \tag{22a}
\]
\[
V_2 = \frac{1}{2} \left( \frac{e_3^2}{\sigma} + \frac{k_p e_2^2}{b} + \frac{k_d \varrho^2 + \varepsilon e_4^2}{k_i} \right) \tag{22b}
\]
\[
V_3 = \frac{1}{\sigma} \varepsilon e_3(e_4 - \varrho), \quad \varepsilon \ll 1 \tag{22c}
\]
\[
V := V_1 + V_2 + V_3. \tag{22d}
\]

We establish that \( V \) is positive definite and its derivative is negative semidefinite.

3.2.1. Positivity of \( V \). Using the triangle inequality we see that

\[
-\frac{1}{2\sigma} \varepsilon (e_3^2 + e_4^2 + \varrho^2) \leq V_3 \leq \frac{1}{2\sigma} \varepsilon (e_3^2 + e_4^2 + \varrho^2)
\]

therefore, given any control gains, there always exists a constant \( 1 \gg \varepsilon > 0 \) and for such \( \varepsilon \), there exist positive reals \( \alpha_1, \alpha_2 \) such that the function \( V \) satisfies

\[
\alpha_1 |x|^2 \leq V(x) \leq \alpha_2 |x|^2 \quad \forall x \in \mathbb{R}^6.
\]  

3.2.2. Negativity of \( \dot{V} \). The total time derivative of \( V_1 \) along the trajectories of (14), (15) yields

\[
\dot{V}_1 = -k_1 e_1^2 + L x_2 e_1 e_3 - k_2 e_2^2 + \Delta e_2 e_3 + v_2 e_2
\]

the derivative of \( V_2 \) is given by

\[
\dot{V}_2 = e_2 e_3 - \frac{k_d a}{b} \varrho^2 - \varepsilon (e_4 - \varrho)
\]

where we used \( \dot{e}_4 = e_3. \) Finally, the derivative of \( V_3 \) satisfies

\[
\dot{V}_3 = \frac{k_d a}{2b} \varrho^2 - \frac{\varepsilon b'}{2\sigma} e_3^2 - \frac{\varepsilon k_p}{2} e_2^2 + \varepsilon (e_4 - \varrho)(z + e_2)
\]

where we used \( b' := b - 1. \) The matrix above is positive semidefinite if

\[
k_d a \geq \varepsilon \left[ 2k_d + \frac{a^2}{\sigma b'} + \frac{(k_p' - k_d)^2}{k_p'} \right]
\]

which holds for sufficiently small values of \( \varepsilon. \) Note, from (18), that this restricts the choice of \( k_1 \) but not of the other control gains. Next, we use (12c) to see that, under the condition (27), the total time derivative of \( V \) satisfies

\[
\dot{V} \leq -\frac{1}{2} \left( k_1 e_1^2 + k_2 e_2^2 + \frac{k_d a}{b} \varrho^2 + \frac{\varepsilon b'}{2\sigma} e_3^2 + \varepsilon k_p' e_4^2 \right)
\]

\[
-\frac{1}{2} \begin{bmatrix} e_1^T \begin{bmatrix} k_1 & 0 & L x_2 \\ 0 & k_2 & (\Delta + 1) \\ L x_2 & (\Delta + 1) & \frac{k_p'}{\sigma} \end{bmatrix} e_2 \end{bmatrix} \cdot \begin{bmatrix} e_1 \\ e_2 \\ e_3 \end{bmatrix}.
\]  

(28)
The matrix in the expression above is positive semidefinite for positive values of all the control gains and if $k_1k_2\varepsilon b' \geq 2\sigma (\Delta + 1)^2k_1 + 2\sigma L^2k_2x^2_2$, which holds if
\[
\begin{align*}
k_2\varepsilon b' & \geq 4\sigma (\Delta + 1)^2 \\
k_1\varepsilon b' & \geq 4\sigma L^2x^2_2.
\end{align*}
\] (29a, 29b)

Note that the expressions above impose that $k_1$ and $k_2$ depend on measurable states hence, these may be chosen as strictly positive functions of $x_2$ and $x_1$ respectively.

We conclude that there exists a constant $\alpha_3 > 0$ such that, defining $y^T := [e_1, e_2, e_3, e_4, \dot{y}]$,
\[
\dot{V}(x) \leq -\alpha_3 |y|^2 \leq 0 \quad \forall x \in \mathbb{R}^6.
\] (30)

That is, $V$ is negative semidefinite and the origin of the system, $\{x = 0\}$, is uniformly globally stable i.e., the solutions are uniformly globally bounded and the origin is Lyapunov stable –see [9]. It is left to show uniform global exponential convergence of the error trajectories.

**Remark 3.1**
Note that $V$ is negative semidefinite. One may not invoke Lasalle’s principle because the system is non-autonomous. Now, (30) implies that $y \in L_2$ and all signals are bounded. From this, it follows from Barbalat’s lemma that $y(t) \to 0$. However, on one hand, this argumentation does not lead to the convergence of $z$ hence of the integrator variable $\dot{v}$. On the other, from Barbalarit’s lemma, the convergence may not be guaranteed to be uniform in the initial conditions hence our purpose to verify the uniform bounds in (13).

### 3.2.3. Uniform global exponential stability

To complete the proof we show that the conditions (13) hold. From (30) we see that
\[
\dot{V}(x(t)) \leq -\alpha_3 |y(t, t_0, x_0)|^2 \leq 0 \quad \forall t_0 \in \mathbb{R}_{\geq 0}, \ x_0 \in \mathbb{R}^6
\]
which is equivalent to
\[
V(x(t)) - V(x(t_0)) \leq -\alpha_3 \int_{t_0}^{t} |y(s, t_0, x_0)|^2 ds \quad x(t_0) = x_0
\]

hence, in view of the positivity and boundedness of $V$ –see (24), we have
\[
\alpha_1 |x(t)|^2 \leq V(x(t)) \leq V(x(t_0)) \leq \alpha_2 |x_0|^2.
\] (31)

It follows that for all $t \geq t_0 \geq 0$ and all $x_0 \in \mathbb{R}^6$
\[
\int_{t_0}^{t} |y(s, t_0, x_0)|^2 ds \leq \frac{\alpha_2}{\alpha_3} |x_0|^2; \quad (32)
\]
\[
|x(t)| \leq c_1 |x_0| \quad c_1 := \sqrt{\frac{\alpha_2}{\alpha_1}} \quad (33)
\]
so (13a) holds. It is left to find a uniform $L_2$ bound on $z(t)$. For this, consider the function
\[
V_4 = \frac{1}{\sigma} e_3^2 z.
\] (34)

Its total time derivative along the closed-loop trajectories yields
\[
\dot{V}_4 = -z^2 + z(e_2 - k_d \vartheta - k_p e_4) - \frac{1}{\sigma} e_3 (-k_1 (e_4 - \vartheta) - \frac{k_i}{\varepsilon} e_3)
\] (35)

which after the triangle inequality, satisfies
\[
\dot{V}_4 \leq -z^2 + \frac{1}{2} [z^2 + \delta |y|^2]
\] (36)
for an appropriate but innocuous choice of $\delta$. Integrating on both sides of (36) along the trajectories, from $t_0$ to $t$ we obtain
\[
2e_3(t)z(t) - 2e_3(t_0)z(t_0) \leq -\sigma \int_{t_0}^{t} z(s)^2 ds + \sigma \int_{t_0}^{t} \delta|y(s)|^2 ds.
\] (37)

Now, even though they are of indefinite sign, the terms on the left hand side of the inequality are bounded by $2|x_o|^2$ hence, using (32) we obtain
\[
\int_{t_0}^{t} z(s)^2 ds \leq \left[ \frac{\delta\alpha_2}{\alpha_3} + 2\sigma \right] |x_o|^2 \quad \forall \, t \geq t_0 \geq 0.
\] (38)

Uniform global exponential stability follows from (32), (33) and (38) since the conditions (13) are satisfied.

**Remark 3.2**
It may be showed that the function $V + \varepsilon_2 V_4$ is a strict Lyapunov function for the closed-loop system with a negative definite derivative (for $\varepsilon_2 \ll 1$). However, this may be established under further restrictions on the control gains hence, it is beyond interest in this paper.

**Remark 3.3 (Robustness)**
The property of uniform global exponential stability cannot be overestimated. It implies (local) Input to State Stability that is, under the presence of relatively small additive disturbances, one recovers asymptotic stability of a residual compact set whose “size” depends on the disturbances’ magnitudes. In particular, consider the case of slowly-varying load torques; say $\tau_L = \tau_L^* + \Delta \tau_L(t)$ where $\tau_L^*$ is constant and $\Delta \tau_L$ is bounded with continuous and bounded first derivative. Then, a simple inspection shows that the closed-loop system takes the form
\[
\dot{x} = F(t, x) + d(t)
\] (39)

where $\dot{x} = F(t, x)$ corresponds to the closed-loop dynamics (14), (15), (20), (21) and (9), and the disturbance $d(t)$ consists in terms $\Delta \tau_L(t)$ and $\Delta \tau_L(t)$ which enter in equations (15), (20) and (21). From Proposition 3.1 and converse Lyapunov theory we deduce that (39) is input to state stable with input $d$.

### 4. CONTROL UNDER PARAMETRIC UNCERTAINTY

Now we relax the assumption that $R$, $L$, $\Phi$ and $\sigma(J, n_p)$ are known and we introduce an adaptive control law to estimate online, $R$, $L$ and $\Phi$. We establish a necessary and sufficient condition for these parameters to be estimated. In addition, we show that the controller is robust with respect to the uncertainty on $\sigma$. In the case that the reference acceleration $\ddot{x}^*_1$ is piece-wise constant, we establish uniform global asymptotic stability for the overall closed-loop system. The proof follows from the analysis in Section 3 and using familiar arguments based on persistency of excitation, to conclude parametric convergence. However, we stress the importance of using technical tools tailored for nonlinear systems –see [20], as opposed to classical adaptive control systems theory for linear systems.

Let $\theta_1 = L$, $\theta_2 = R$, $\theta_3 = \Phi$, $\theta_4 = 1/\sigma$ and $^\top \theta = [\theta_1 \, \theta_2 \, \theta_3] ^\top$. We denote by $\hat{\theta}$ the estimate of $\theta$ and the estimation errors by $\bar{\theta} = \hat{\theta} - \theta$. Then, Equations (1) become
\[
\begin{align*}
\theta_1 \dot{x}_1 &= -\theta_2 x_1 + \theta_1 x_1 x_3 + u_1 \\
\theta_1 \dot{x}_2 &= -\theta_2 x_2 - \theta_1 x_1 x_3 - \theta_3 x_3 + u_2 \\
\theta_4 \dot{x}_3 &= x_2 - \frac{\tau_L}{\sigma} \\
\dot{x}_4 &= x_3
\end{align*}
\] (40a-d)

$^\top \theta_4$ is not estimated online but is dealt with separately, hence the definition of $\theta$ and the explicit appearance of $\sigma$. 

and we introduce the certainty-equivalence adaptive controller

\[ u_1 = -\hat{\theta}_1 x_2 x_3^* + \hat{\theta}_2 x_1 + \hat{\theta}_1 \hat{x}_1^* - k_1 e_1 \tag{41a} \]

\[ u_2 = \hat{\theta}_3 x_3^* + \hat{\theta}_1 x_1^* + \hat{\theta}_2 x_2 - k_2 \hat{e}_2 + \hat{\theta}_1 \alpha + v_2 \tag{41b} \]

\[ \dot{\hat{\theta}} := -\gamma \Psi e_{12}, \quad \gamma > 0 \tag{41c} \]

where \( \hat{e}_2 := x_2 - \hat{x}_2^* \), \( e_{12} := [e_1, \hat{e}_2]^\top \),

\[ \hat{x}_2^* := \hat{\theta}_4 x_3^* + \nu + v_3, \quad \dot{\hat{\theta}}_4 = 0 \tag{42a} \]

\[ \alpha := -k_3 (e_4 - \theta) + k_4 \dot{a} \dot{\theta} \tag{42b} \]

\[ \Psi := \begin{bmatrix} x_1^* & -x_2 x_3^* & x_1 \\ \alpha + x_1 x_3^* & x_2 & x_3^* \end{bmatrix}. \tag{42c} \]

Note that \( \hat{\nu} = \nu + \dot{\hat{v}}_3 = \alpha - (k_p + k_d b) e_3 \) since \( \dot{\hat{\theta}}_4 = 0 \) and \( \hat{x}_3^* \) is (piece-wise) constant.

Furthermore, we remark that in contrast to other adaptive control schemes for electrical machines, we use a reduced-order adaptive controller. Indeed, adaptation is used only for the electrical dynamics equations. To obtain the closed-loop equations corresponding to the electrical dynamics, let \( \varphi = [L x_2 e_3 \ \Delta e_3 + v_2]^\top \) and \( K_{12} := \text{diag}[k_1 \ k_2] \) where the latter are functions of the state, according to (29) and (16). Then, using (41) and (42) we obtain

\[
\begin{bmatrix}
L \hat{e}_{12} \\
\dot{\hat{\theta}}
\end{bmatrix} =
\begin{bmatrix}
- K_{12}(t, \xi_1) & \Psi(t, \xi_1)^\top \\
-\gamma \Psi(t, \xi_1) & 0
\end{bmatrix}
\begin{bmatrix}
e_{12} \\
\dot{\hat{\theta}}
\end{bmatrix} + \begin{bmatrix}
\varphi \\
0
\end{bmatrix}.
\] \tag{43}

Not surprisingly, for \( \varphi = 0 \), the previous system (43) has the familiar structure of model-reference-adaptive control systems. The analysis of Equation (43) with \( \varphi = 0 \) is standard routine under the following observations: if \( \hat{\theta} = 0 \) the system reduces to that studied in the previous section otherwise, Barbalat’s lemma and standard signal-chasing arguments lead to the conclusion that \( e_{12} \to 0 \). Moreover, persistency of excitation of \( \Psi \) may be invoked to conclude that \( \dot{\hat{\theta}} \to 0 \). See for instance [2, 8]. Finally, a converse Lyapunov function for the system with \( \varphi = 0 \) might be used to analyze the perturbed dynamics (43).

The simplicity of such argumentation hides several technical fallacies which lead to wrong conclusions. Firstly, for notational simplicity \( \Psi \) is written in (42c) without arguments however, we emphasize that \( \Psi : \mathbb{R}_{\geq 0} \times \mathbb{R}^6 \to \mathbb{R}^{3 \times 2} \) i.e., it is a function of time through the functions \( t \mapsto x_1^* \), \( t \mapsto x_3^* \) and their derivatives, as well as of the state \( \xi_1 = [e_1 e_2 e_3 e_4 \varphi \dot{\varphi}]^\top \). More precisely, in place of \( x_2 \) one must read \( \hat{e}_2 + \hat{x}_2^* \), where \( \hat{x}_2^* \) is a function of the closed-loop state \( \xi_1 \). Correspondingly, one must read \( e_1 + x_1^*(t) \) in place of \( x_1 \), etc. Therefore, invoking standard results as for linear systems –such as those in [1], requires to impose persistency of excitation on \( \Psi(t, \xi_1(t)) \) that is, along closed-loop trajectories –cf.[23, 2, 6]. Not only this is un-necessary but clearly impossible to verify for all \( t \). Moreover, such reasoning may only lead to non-uniform attractivity which in turn, invalidates the invocation of converse Lyapunov theorems. To overcome these difficulties, we shall use the tools reported in [20], tailored for nonlinear-time-varying systems.

At this point we introduce the dynamics corresponding to the rotor variables; this may be computed as follows. We replace \( x_2 = \hat{e}_2 + \hat{x}_2^* \) in (1c) and use (42a) to obtain

\[ \theta_4 \dot{x}_3 = \hat{e}_2 + \hat{\theta}_4 \dot{x}_3^* + \nu + v_3 - \frac{\tau_L}{\sigma} \pm \theta_4 x_3^* \]

and we redefine \( \nu^* := \tau_L / \sigma - (\hat{\theta}_4 - \theta_4) \dot{x}_3^* \) where \( \dot{x}_3^* \) is (piece-wise) constant hence, so is \( \nu^* \).

Therefore,

\[ \theta_4 \dot{e}_3 = \dot{\hat{e}}_2 + \nu + v_3 \]

which is exactly of the form (17) and therefore, leads to (20) with the only difference laying in the redefinition of \( \nu^* \) hence, of \( z \). In other words, the additional uncertainty \( \hat{\theta}_4 = \theta_4 - \dot{\theta}_4 \) only changes
the equilibrium of the $\nu$-dynamics part of the closed-loop system. The latter is

$$\begin{bmatrix}
\dot{\theta}_1 \\
\dot{\theta}_2 \\
\dot{\vartheta} \\
\dot{z}
\end{bmatrix} = \begin{bmatrix}
0 & -k_p & -k_d & 1 \\
1 & 0 & 0 & 0 \\
b & 0 & -\alpha & 0 \\
-k_i/\varepsilon & -k_i & k_i & 0
\end{bmatrix} \begin{bmatrix}
\theta_1 \\
\theta_2 \\
\vartheta \\
z
\end{bmatrix} + \begin{bmatrix}
e_1 \\
e_2 \\
0 \\
0
\end{bmatrix}$$

(44)

in which it is clear that the control gains may be chosen appropriately to ensure that the matrix above is Hurwitz.

The system (44) corresponds to the tracking control dynamics for the mechanical variables, under the action of the virtual control $\hat{x}_2$. The absence of a term involving parametric uncertainty in (44) (as it might be expected) comes on one hand from the fact that we chose to maintain $\partial_1$ constant and $\hat{x}_2$ is assumed piece-wise constant and on the other hand, because $\nu$ is compensated for by integral action. Note that by replacing $\tilde{e}_2$ with $e_2$, system (44) also corresponds to the closed-loop equations (20), (21) and (9) hence, in the case that $\dot{e}_2 = 0$, the origin is exponentially stable for suitable values of the control gains; also, the dynamics is input to state stable with input $\tilde{e}_2$.

Thus, the overall closed-loop system (43), (44) may be regarded as the feedback interconnection of two input-to-state stable systems and uniform global asymptotic stability of the origin of (43), (44) may be inferred invoking a small-gain argument. This simple rationale hides the difficulty that $\varphi$ in (43) depends on all of $\xi_1$ and not only on the “mechanical” variables; in turn, this imposes the challenge of actually constructing a strict Lyapunov function for the overall system—see [12, 13]. Instead, in order to show that the origin of the closed-loop system is uniformly globally asymptotically stable, we follow a direct but rigorous method of proof which relies on [20, Proposition 3]. The latter is an output-injection statement for nonlinear systems, analogous to the well-known output-injection lemma for linear adaptive systems—cf. [1]. In words, the output-injection lemma in [1] establishes that uniform complete observability is invariant under an output injection provided that the output is in $L_2$. Its nonlinear “counterpart”, [20, Proposition 3], establishes that uniform global asymptotic stability is invariant under (a uniformly $L_2$) output injection.

**Proposition 4.1**

Let $t \mapsto x_1^r$ and $t \mapsto x_2^r$ be given bounded reference trajectories. Let $x_1^r$ be twice continuously differentiable with bounded derivatives and let $x_2^r$ be piece-wise continuous with $\dot{x}_2^r$ piece-wise constant. Consider the system (40) in closed loop with the adaptive controller (41), (42). Let $\sigma_m$, $\sigma_M$, $L_M$ and $\Phi_M$ be known constants such that $\sigma \in [\sigma_m, \sigma_M]$, $L_M \geq L$ and $\Phi_M \geq \Phi$. Then, all tracking errors $e_1, e_2, e_3, e_4, \tilde{v}$ and $\theta$ converge to zero asymptotically provided the control gains (dependent on $\sigma_m, \sigma_M, L_M$ and $\Phi_M$) are sufficiently large. Furthermore, define $\Psi : \mathbb{R}_0^+ \times \mathbb{R} \to \mathbb{R}^{3 \times 2}$ as

$$\Psi(t, \delta)^\top = \begin{bmatrix}
x_1^r(t) - \delta x_2^r(t) & x_1^r(t) \\
x_1^r(t) & x_2^r(t)
\end{bmatrix}$$

then, the origin of the closed loop system is uniformly globally asymptotically stable if and only if there exist $\mu > 0$ and $T > 0$ such that

$$M(t) := \int_t^{t+T} \Psi(s, 1)\Psi(s, 1)^\top ds \geq \mu I \quad \forall t \geq 0.$$  

(45)

**Proof of Proposition 4.1**: To apply [20, Proposition 3] we start by establishing uniform global stability hence, uniform boundedness of all trajectories $t \mapsto \xi$ where $\xi = [\xi_1^\top \ \theta^\top]^\top$. Following the developments of Section 3 we see that the total derivative of

$$V(\xi) = V(\xi_1) + \frac{1}{2\gamma} |\theta|^2$$

along the trajectories of (43), (44) yields $\dot{V}(\xi) = \dot{V}(\xi_1)$ where $\dot{V}(\xi_1)$ is upper-bounded by the right-hand side of (28) with $e_2$ replaced by $\dot{e}_2$. Define

$$\Delta_M = L_M(k_p + bk_d) + (\Phi_M + L_M[x_1])$$

Copyright © 2012 John Wiley & Sons, Ltd.


Prepared using acsauth.cls

DOI: 10.1002/acss
and let (29) and (27) hold with $\Delta_M$ in place of $\Delta$, with $\sigma_m$ in (27), $\sigma_M$ in (29) instead of $\sigma$ and with $L_M$ in place of $L$. Then, similarly to (30) we have

$$\dot{V}(\xi_1) \leq -\alpha_3|\dot{y}|^2 \leq 0,$$

hence, uniform global stability of the origin as well as uniform square integrability of $e_1$, $\dot{e}_2$, $e_3$, $e_4$ and $\dot{\theta}$ follow. Invoking standard signal-chasing arguments which involve Barbalat’s lemma, one obtains the convergence of $|\dot{y}(t)|$ to zero.

To establish uniform global asymptotic stability of the origin note that the analysis in Paragraph 3.2.3) holds, 

*mutatis mutandis* for the trajectories of (43)–(44) hence, from (32) and (46), there exists $c > 0$ such that

$$\int_{t_0}^{t} |\xi_1(s)|^2 ds \leq c|\xi_0|^2 \quad \forall t \geq t_0 \geq 0. \tag{47}$$

Next, note that the inequality in (45) holds if and only if there exist $T'$, $\mu' > 0$ such that

$$\int_{t}^{t+T'} \dot{\Psi}(s, \delta)\dot{\Psi}(s, \delta)^T ds \geq \mu I \quad \forall t \geq 0, \forall \delta \in \mathbb{R}; \tag{48}$$

this follows from the fact that the choice of $\delta$ does not modify the rank of $M(t)$. In what follows we set $\delta = \dot{x}_2^*(0)$ and observe that $\Psi(t, \dot{x}_2^*(0)) = \Psi(t, 0)$. Moreover, $K_{12}$ is a diagonal matrix bounded from below, say by the constant diagonal matrix $K_{12}^m > 0$ then, let $K_{12}'(t, \xi_1) := K_{12}(t, \xi_1) - K_{12}^m$ so the equations (43) may be rewritten as

$$\begin{bmatrix} \dot{e}_{12} \\ \dot{\theta} \end{bmatrix} = \begin{bmatrix} -K_{12}^m/L & \dot{\Psi}(t, \dot{x}_2^*(0))^T/L \\ -\gamma\dot{\Psi}(t, \dot{x}_2^*(0)) & 0 \end{bmatrix} \begin{bmatrix} e_{12} \\ \dot{\theta} \end{bmatrix} + K_1(t, \xi) \tag{49}$$

where the output-injection term

$$K_1(t, \xi) := \begin{bmatrix} \varphi \\ 0 \end{bmatrix} + \begin{bmatrix} K_{12}^m(t, \xi_1)/L & (\Psi(t, \xi_1)^T - \Psi(t, 0)^T)/L \\ -\gamma(\dot{\Psi}(t, \xi_1)^T - \Psi(t, 0)^T) & 0 \end{bmatrix} \begin{bmatrix} e_{12} \\ \dot{\theta} \end{bmatrix} \tag{50}$$

satisfies $K_1 \equiv 0$ if $\xi_1 \equiv 0$. Indeed, note that there exists $c > 0$ such that $|\Psi(t, \xi_1)^T - \Psi(t, 0)^T| \leq c|\xi_1|$ while $|\varphi| \leq c|\xi_1|(c + |\xi_1|)$ therefore $|K_1(t, \xi)| \leq c|\xi_1|(c + |\xi_1|)$. That is, $K_1$ is a vanishing output injection. More precisely, in view of (46) and (47) we have

$$\int_{t_0}^{t} |K_1(s, \xi_1(s))|^2 ds \leq \beta(|\xi_0|)|\xi_0|^2 \quad \forall t \geq t_0 \geq 0$$

where $\beta : \mathbb{R}_{\geq 0} \to \mathbb{R}_{\geq 0}$ is non-decreasing. To invoke [20, Proposition 3] it is only left to show uniform global asymptotic stability of the origin of (49), (44) under the condition that $K_1 \equiv 0$ in (49) and $\dot{e}_2 \equiv 0$ in (44). Note that in this case, the two dynamics are decoupled. Uniform global asymptotic stability of the origin of (44) follows from the analysis in Section 3 while uniform global exponential stability of the origin of (49) with $K_1 \equiv 0$ follows invoking standard theorems on linear time-varying adaptive control systems: observe that $K_{12}^m > 0$ is constant, $\dot{\Psi}(t, \dot{x}_2^*(0))$ and $\dot{\Psi}(t, \dot{x}_2^*(0))$ are bounded a.e. and the PE condition (48) holds. Necessity of the PE condition also follows. See [1].

\begin{flushright} \Box \end{flushright}

\begin{flushleft} \textbf{Remark 4.1} \end{flushleft}

We stress that the origin is also uniformly (locally) exponentially stable; this follows from [20, Proposition 4] however, global exponential stability under the PI²D control seems out of reach, in view of the output-injection terms which may be seen as a nonlinear non-globally-Lipschitz vanishing perturbation. Nonetheless, the robustness properties described in Remark 3.3 still hold, notably for the case that $\dot{x}_2^* \not\equiv 0$.

The previous result may be compared with related literature, such as [2] which establishes convergence of the tracking errors but also [8] where the PE condition on the regressor is imposed on a regressor $\Gamma(t, x(t))$ that is, along reference trajectories.

\footnote{Without loss of generality $c$ denotes a generic constant whose value is unimportant.}
GLOBAL ADAPTIVE LINEAR CONTROL OF PMSM

5. SIMULATION RESULTS

We have performed some numerical simulations using the parameters reported in [8] with the
aim of evaluating the proposed controller and of comparing it with one of the existing reported
schemes. We use \( R = 3\Omega \), \( L = 0.006H \), \( \Phi = 0.33Wb \), \( n_p = 6 \) and \( J = 0.01Kgm^2 \). In order
to work under a more stringent condition than in [8] the damping coefficient is set to zero.
The evaluation consists in imposing a speed reference inspired by the signal profile proposed
as a benchmark, by the French Working Group Commande des Entraînements Electriques –
see http://www2.irccyn.ec-nantes.fr/CE2/. The desired motor speed starts at zero,
increases with a slope of 5.25 \( \frac{rad}{s} \) until it reaches the value 5.25 \( \frac{rad}{s} \), at \( t = 1s \). This value is kept
constant until \( t = 3s \) when it is increased again, with a slope of 3.675 \( \frac{rad}{s} \), up to 12.6 \( \frac{rad}{s} \) during 2
seconds. Then, the reference decreases (with a slope of 6.3 \( \frac{rad}{s} \)) and remains at zero for the rest of
the simulation time. The applied load torque is constant and equals 1Nm.

Following field-orientation control ideas, the desired value for \( x_1 \) is set to zero while the controller
gains are set to \( k_1 = 40 \), \( k_2 = 65 \), \( k_p = 5 \), \( k_d = 10 \), \( k_t = 0.005 \), \( a = 50 \), \( b = 50 \), \( \varepsilon = 0.02 \) and \( \gamma = 5 \);
the last one is used when parametric uncertainty is considered. It is assumed that the motor is at
stand-still at the beginning of the simulation, i.e., all the motor states are set to zero, and in a similar
way the initial value of the derivative filter state \( q_e \), the estimated load torque and the parameter
estimates are considered equal to zero. The estimated value for the unknown parameter \( \theta_1 \) was set
to \( \hat{\theta}_1 = 0.5\theta_1 \). Such is the worst-case scenario with respect to the uncertainty, both for the load
torque and the parameters.

Three different simulations were carried out. The first one illustrates the controller performance
under the conditions stated in Proposition 3.1 i.e., that all physical parameters except for the
load torque are known. The second is related to Proposition 4.1, under parametric uncertainty. In
this case, in addition to uncertainty on the model parameters, we added 400% uncertainty of the
nominal value of \( \tau_L \) i.e., it was set to 5Nm, from \( t = 10s \) to \( t = 15s \). One last simulation illustrates
the robustness property mentioned in Remark 3.3 by adding a slowly time-varying sinusoidal
perturbation of amplitude 2Nm and frequency 0.5Hz.

The simulation results for the first scenario (with known parameters) are showed in Figure 1. The
reference speed profile together with the actual speed response in Figure 1(a), while the currents and
speed error behaviors in Figure 1(b). It can be observed the remarkable performance of the PFtD
controller. The position error, \( e_4 \), is not showed to avoid graphical saturation and due to the fact that
its behavior is unimportant for achieving the control objective, although it must be recognized that
its rate of convergence to zero is considerably slower than that corresponding to the other errors.

In Figure 2(a) we show the required stator voltages. The noticeable spikes in the stator variables
are due to sudden changes in the first and second derivatives of the reference speed. These spikes are
avoided if a smooth reference is designed or a filter for the reference is included, but with the aim
at evaluating the controller under stringer conditions, the discontinuities are not avoided during this

\[ \text{(a) Actual (continuous line) and desired (dashed line) speeds} \]

\[ \text{(b) Currents and speed errors} \]

Figure 1. First scenario: with uncertainty in the load torque only and continuous speed reference

Prepared using acaauth.cls DOI: 10.1002/acs
simulation. In Figure 2(b) we depict the integral correction $\nu$ which compensates for the constant disturbance $\tau_L/\sigma$. As expected, the convergence of this variable to its steady-state is very slow due to the small value of the integral gain $k_i$.

In the second and third scenarios we use the adaptive controller (41) by filtering the reference $x_3^*$ using a second order filter of the form

$$G(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2}$$

where $\omega_n = 100$ while $\zeta = 20$. The numerical results obtained under uncertainty in both, the load torque and the motor parameters, are showed in Figures 3 and 4. It can be observed in Figure 3(a)
the capability of the control scheme to achieve the control objective even when, at $t = 10s$, the constant unknown disturbance in the load-torque is applied. As expected, the transient response at the beginning of the simulation exhibits some oscillations due to the error in the parameter values. This behavior is attenuated by using better values as initial conditions instead of zero. In Figure 3(b) is depicted the convergence of the state errors to zero, in spite of the perturbation. Figure 4(a) shows that the estimation errors converge to zero asymptotically and for completeness, in Figure 4(b) we depict the eigenvalues of $M(t)$ in (45). To induce the necessary excitation we have used $x_1^*(t) := \sin(2\pi t)$. Finally, Figure 5(a) shows the speed behavior under the aforementioned time-varying load torque disturbance. Input-to-state stability may be appreciated.

6. CONCLUDING REMARKS

We showed that a position-feedback controller of PID type solves the speed tracking control problem for permanent-magnet synchronous motors even in the case of unknown parameters. From a practical perspective, besides the remarkable dynamic performance achieved by the closed-loop system, the proposed design avoids the use of (noisy) speed sensor and does not rely on the knowledge of the load torque. The controller’s robustness established analytically is also evident from the numerical simulations.

REFERENCES