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Robust output feedback learning control for induction motor servo drives

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SUMMARY

This paper addresses the output feedback tracking control problem for induction motor servo drives with mechanical uncertainties: rotor angle, rotor speed and stator currents are assumed to be available for feedback. A robust adaptive learning control is designed under the assumption that the reference profile for the rotor angle is periodic with known period: it 'learns' the periodic disturbance signal by identifying the Fourier coefficients of any truncated approximation; $\mathscr{L}_2$ and $\mathscr{L}_\infty$ transient performances are guaranteed in the 'learning phase'. It is shown that, for any motor initial condition belonging to an arbitrary given compact set, by properly setting the control parameters: (i) the rotor position and flux modulus tracking errors exponentially converge to residual sets, which may be arbitrarily reduced by increasing the number of terms in the truncated Fourier series; (ii) when the unknown periodic disturbance can be represented by a finite Fourier series, the rotor position and flux modulus tracking errors exponentially converge to zero. As in field oriented-control, the control algorithm generates references for the magnetizing flux component and for the torque component of the stator current leading to significant simplifications for current-fed motors. Copyright © 2008 John Wiley & Sons, Ltd.

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1. INTRODUCTION

Induction motors are widely utilized in servo applications as an alternative to hydraulic or pneumatic drives, thanks to their reliability, ruggedness and ability to turn large mechanical loads. Nevertheless, induction motor control constitutes a rather difficult problem due to the highly nonlinear and multivariable nature of its model and to the non-measurability of all state variables: while rotor

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position and speed along with stator currents can be easily measured, flux sensors are not available in standard induction motors. Solutions to the problem of controlling robot manipulators actuated by induction motors are presented in [1–5]. A local partial state feedback link position tracking controller is proposed in [1], while global output feedback position tracking is obtained in [2]: exact model knowledge is required. By assuming linear parameterization of manipulator inertia matrix determinant, global asymptotic link position tracking is guaranteed in [3] despite parametric uncertainties in the mechanical subsystem. Robust control algorithms for uncertain robot manipulators driven by current-fed induction motors are proposed in [4, 5]: an arbitrarily small link position tracking error can be obtained by properly setting the design parameters. In the case of periodic reference trajectories, system parameter uncertainties may cause periodic disturbances in motor servo drives so that learning control techniques may be successfully applied. Even though several (finite and infinite interval) learning control strategies have been developed to compensate for periodic disturbances in robot applications requiring repetitive tasks (see [6–8] and references therein), the servo drive electromagnetic dynamics have been only considered in [9, 10]: backstepping, robust and repetitive learning control techniques are applied in [9] to design an output feedback control scheme (formally requiring infinite memory) for rotor angle and flux modulus tracking of current-fed induction motor servo drives with uncertain mechanical model, while a global output feedback rotor angle and flux modulus tracking control for current-fed induction motor servo drives with mechanical uncertainties is presented in [10].

In this paper, the rotor angle and flux modulus tracking control problem by output feedback is addressed for uncertain voltage-fed induction motor servo drives under the hypothesis that the reference profile for the rotor angle is periodic with known period: no simplifying assumptions on the stator current dynamics are invoked. The non-structured unknown periodic disturbance due to mechanical uncertainties is developed in Fourier series and the estimates of its coefficients are continuously adapted using the robust adaptive control techniques proposed in [11]. For any motor initial condition in an arbitrary given compact set, by properly setting the control parameters, the guaranteed rotor position and flux modulus tracking errors are reduced as the number of such Fourier coefficients is increased, while exponential rotor position and flux modulus tracking are obtained when the unknown periodic disturbance is represented by a finite Fourier series expansion. $\mathscr{L}_2$ and $\mathscr{L}_\infty$ transient performances on the tracking errors are guaranteed during the learning phase. When compared with the commonly used Internal Model Approach (see for instance [12]), the adaptive design proposed in this paper (inspired by Del Vecchio *et al.* [13]) allows to deal with periodic disturbances/uncertainties admitting infinite Fourier series expansion and to solve the problem of the controller robustness with respect to the truncation error. On the other hand, even when the tracking problem with unstructured uncertainties can be reduced to a tracking problem with linearly parameterized uncertainties (so that standard adaptive techniques can be applied), persistency of excitation and exponential tracking are, in general, not guaranteed (in [14] only asymptotic convergence is assured when the external disturbance can be modeled as the output of a finite-dimensional linear exosystem). Simulation results for a digital implementation of the control algorithm illustrate the effectiveness of the proposed solution. A preliminary version of this work was presented in [15].

2. SYSTEM DYNAMIC MODEL

Assuming linear magnetic circuits, the electromagnetic dynamics of a balanced non-saturated induction motor with one pole pair are given in a fixed reference frame attached to the stator by

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(see, for instance, [16] for its derivation and [17] for modeling assumptions)

$$\begin{aligned}
\frac{\mathrm{d}\psi_{ra}}{\mathrm{d}t} &= -\alpha\psi_{ra}-\omega\psi_{rb}+\alpha M i_{sa}\\
\frac{\mathrm{d}\psi_{rb}}{\mathrm{d}t} &= -\alpha\psi_{rb}+\omega\psi_{ra}+\alpha M i_{sb}\\
\frac{\mathrm{d}i_{sa}}{\mathrm{d}t} &= -\left(\frac{R_s}{\sigma}+\beta\alpha M\right)i_{sa}+\frac{1}{\sigma}u_{sa}+\beta\alpha\psi_{ra}+\beta\omega\psi_{rb}\\
\frac{\mathrm{d}i_{sb}}{\mathrm{d}t} &= -\left(\frac{R_s}{\sigma}+\beta\alpha M\right)i_{sb}+\frac{1}{\sigma}u_{sb}+\beta\alpha\psi_{rb}-\beta\omega\psi_{ra}
\end{aligned} \tag{1}$$

where (ψ_{ra},ψ_{rb}) are the rotor fluxes, (i_{sa},i_{sb}) are the stator currents, ω is the rotor speed and (u_{sa},u_{sb}) are the stator voltages. The electromagnetic constant parameters are: rotor and stator windings resistances (R_r,R_s) and inductances (L_r,L_s), mutual inductance M. To simplify notations we use the reparameterization: $\alpha=R_r/L_r,\beta=M/\sigma L_r,\sigma=L_s(1-M^2/L_sL_r)$. The motion of the induction motor rotor shaft connected to a generic one-degree of freedom mechanical system is described by the Euler–Lagrange equation (see [9, 18])

$$\begin{aligned}
\frac{\mathrm{d}\theta}{\mathrm{d}t} &= \omega\\
J(\theta)\frac{\mathrm{d}\omega}{\mathrm{d}t} &= m(\psi_{ra}i_{sb}-\psi_{rb}i_{sa})-\frac{1}{2}\frac{\partial J(\theta)}{\partial\theta}\omega^2-b(\theta)\omega
\end{aligned} \tag{2}$$

where θ is the rotor angle, $J(\theta)$ is the equivalent total inertia, $b(\theta)\omega$ is the equivalent total linear viscous friction torque and $m=3M/2L_r$ with $J(\cdot),b(\cdot)$ typically being the uncertain functions. The output to be controlled is $y=[\theta,\sqrt{\psi_{ra}^2+\psi_{rb}^2}]^{\mathrm{T}}$ while the measured output is $y_m=[\theta,\omega,i_{sa},i_{sb}]^{\mathrm{T}}$. If we introduce an angle $\varepsilon_0(t)$, whose dynamics $\mathrm{d}\varepsilon_0(t)/\mathrm{d}t=\omega_0(t)$ will be later defined ($\varepsilon_0(0)$ is an arbitrary initial condition) we can equivalently consider the vectors $[\psi_{rd},\psi_{rq}]^{\mathrm{T}},[i_{sd},i_{sq}]^{\mathrm{T}},[u_{sd},u_{sq}]^{\mathrm{T}}$, which are obtained multiplying the corresponding (a,b) vectors $[\psi_{ra},\psi_{rb}]^{\mathrm{T}},[i_{sa},i_{sb}]^{\mathrm{T}},[u_{sa},u_{sb}]^{\mathrm{T}}$ by the matrix

$$\mathscr{R}(\varepsilon_0)=\begin{bmatrix}\cos\varepsilon_0 & \sin\varepsilon_0\\ -\sin\varepsilon_0 & \cos\varepsilon_0\end{bmatrix} \tag{3}$$

They represent the direct and quadrature components of rotor flux, stator current and stator voltage vectors, respectively, with respect to a time-varying (d,q) reference frame rotating at speed $\omega_0(t)$ and identified by the angle $\varepsilon_0(t)$ in the fixed (a,b) reference frame. In the new state coordinates $(\theta,\omega,\psi_{rd},\psi_{rq},i_{sd},i_{sq})$ and new control coordinates (u_{sd},u_{sq}) the motor dynamics become (see [9]):

$$\begin{aligned}
\frac{\mathrm{d}\theta}{\mathrm{d}t} &= \omega\\
J(\theta)\frac{\mathrm{d}\omega}{\mathrm{d}t} &= m(\psi_{rd}i_{sq}-\psi_{rq}i_{sd})-\frac{1}{2}\frac{\partial J(\theta)}{\partial\theta}\omega^2-b(\theta)\omega\\
&\doteq m(\psi_{rd}i_{sq}-\psi_{rq}i_{sd})-f(\theta,\omega)
\end{aligned}$$

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$$\begin{aligned}
\frac{\mathrm{d}\psi_{rd}}{\mathrm{d}t} &= -\alpha\psi_{rd} + (\omega_0 - \omega)\psi_{rq} + \alpha M i_{sd} \\
\frac{\mathrm{d}\psi_{rq}}{\mathrm{d}t} &= -\alpha\psi_{rq} - (\omega_0 - \omega)\psi_{rd} + \alpha M i_{sq} \\
\frac{\mathrm{d}i_{sd}}{\mathrm{d}t} &= -\left(\frac{R_s}{\sigma} + \beta\alpha M\right) i_{sd} + \frac{1}{\sigma} u_{sd} + \omega_0 i_{sq} + \beta\alpha\psi_{rd} + \beta\omega\psi_{rq} \\
\frac{\mathrm{d}i_{sq}}{\mathrm{d}t} &= -\left(\frac{R_s}{\sigma} + \beta\alpha M\right) i_{sq} + \frac{1}{\sigma} u_{sq} - \omega_0 i_{sd} + \beta\alpha\psi_{rq} - \beta\omega\psi_{rd}
\end{aligned} \tag{4}$$

3. PROBLEM STATEMENT AND PRELIMINARY COMPUTATIONS

Consider the induction motor (4) with the non-structured uncertain scalar function $f(\theta, \omega)$ under the following assumptions:

(A.1) $m, \alpha, M, R_s, \sigma, \beta$ are known constant parameters;
(A.2) the functions $J(\theta)$ and $b(\theta)$ are of class $\mathscr{C}^{s_j}$ ($s_j \geqslant 6$) and $\mathscr{C}^{s_b}$ ($s_b \geqslant 5$), respectively;
(A.3) there exist known positive reals J_m, J_M, k_{ji}, k_{bi} ($i = 1, 2, 3$) such that for all $\theta \in \mathbb{R}$ [$i = 1, 2, 3$]: (i) $J(\theta) \in [J_m, J_M]$; (ii) $|\partial^i J(\theta)/\partial\theta^i| \leqslant k_{ji}$; (iii) $|\partial^{i-1} b(\theta)/\partial\theta^{i-1}| \leqslant k_{bi}$.

We address the problem of designing an output feedback control for system (4) in order to track an output reference signal vector $y_r(t) = [y_{r1}(t), y_{r2}(t)]^{\mathrm{T}}$ which is assumed to belong to the following class:

(A.4) $y_{r1}(t) = \theta^*(t)$ is periodic with known period T, of class $\mathscr{C}^{s_\theta}$ ($s_\theta \geqslant 7$) with bounded time derivatives $|y_{r1}^{(i)}(t)| \leqslant c_{\theta i}$ ($i = 1, 2, 3, 4$) for all $t \in [0, T)$;
(A.5) $y_{r2}(t) = \psi^*(t) \geqslant c_\psi > 0$ is of class $\mathscr{C}^{s_\psi}$ ($s_\psi \geqslant 2$) with bounded time derivative $|y_{r2}^{(i)}(t)| \leqslant c_{\psi i}$ ($i = 1, 2$) for all $t \geqslant 0$.

Define the tracking errors (k_θ is a positive control parameter)

$$\tilde{\theta} = \theta - \theta^*$$

$$\tilde{\omega} = \omega - \omega^* \doteq \omega + k_\theta\tilde{\theta} - \dot{\theta}^*$$

so that

$$\dot{\tilde{\theta}} = -k_\theta\tilde{\theta} + \tilde{\omega}$$

and express $f(\theta, \omega)$ as

$$f(\theta, \omega) = q_0(\theta^*, \dot{\theta}^*, \ddot{\theta}^*) - J(\theta)\ddot{\theta}^* + g(\tilde{\theta}, \tilde{\omega}, \omega) + J(\theta)k_\theta\tilde{\omega} - J(\theta)k_\theta^2\tilde{\theta}$$

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with

$$q_0(\theta^*,\dot\theta^*,\ddot\theta^*)=\frac{1}{2}\frac{\partial J(\theta)}{\partial\theta}\bigg|_{\theta=\theta^*}\dot\theta^{*2}+b(\theta^*)\dot\theta^*+J(\theta^*)\ddot\theta^*$$

$$\begin{aligned}g(\tilde\theta,\tilde\omega,\omega)=&-J(\theta)k_\theta\tilde\omega+J(\theta)k_\theta^2\tilde\theta+\frac{1}{2}\frac{\partial J(\theta)}{\partial\theta}\bigg|_{\theta=\theta^*}[k_\theta^2\tilde\theta^2-2k_\theta\tilde\theta\dot\theta^*+\tilde\omega^2\\&+2\dot\theta^*\tilde\omega-2k_\theta\tilde\theta\tilde\omega]+\frac{1}{2}\left[\frac{\partial J(\theta)}{\partial\theta}-\frac{\partial J(\theta)}{\partial\theta}\bigg|_{\theta=\theta^*}\right]\omega^2\\&+b(\theta^*)(\tilde\omega-k_\theta\tilde\theta)+[b(\theta)-b(\theta^*)]\omega+[J(\theta)-J(\theta^*)]\ddot\theta^*\end{aligned}$$

4. FOURIER SERIES EXPANSION AND PROJECTION ALGORITHM

By virtue of assumption (A.4), $q(t)=q_0(\theta^*(t),\dot\theta^*(t),\ddot\theta^*(t))$ is periodic of known period T and can be approximated according to Fourier approximation theory (see [13, 19, 20]). Let $\rho[N]=[\rho_0,\ldots,\rho_{N-1}]^{\mathrm T}\in\mathbb{R}^N$ be the vector of the first N Fourier coefficients of the function $q(t)\in\mathscr{C}^p$ ($p=\min\{s_\theta-2,s_j-1,s_b\}$): for any $N>1$ (N is an odd number) we can write

$$q(t)=\sum_{l=0}^{N-1}\rho_l\varphi_l(t)+\varepsilon(t)\quad\text{with }|\varepsilon(t)|\leqslant\varepsilon_N$$

where ($l=1,\ldots,(N-1)/2$, $p\geqslant2$)

$$\varphi_0(t)=1,\quad \varphi_{2l}(t)=\sqrt{2}\cos\left(lt\frac{2\pi}{T}\right),\quad \varphi_{2l-1}(t)=\sqrt{2}\sin\left(lt\frac{2\pi}{T}\right)\tag{5}$$

$$\varepsilon_N=\left[\left(\frac{2\pi}{T}\right)^p(N-1)^{p-1/2}\right]^{-1}(2^{p-1/2}B_{qp})\tag{6}$$

with B_{qp} being the upper bound on $|\mathrm{d}^pq(t)/\mathrm{d}t^p|$. Moreover, from assumptions (A.3) and (A.4), a bound for $q(t)$ is given by

$$|q(t)|\leqslant\tfrac{1}{2}k_{j1}c_{\theta1}^2+k_{b1}c_{\theta1}+J_Mc_{\theta2}\doteq B_q\tag{7}$$

so that by virtue of the Parseval identity, we obtain

$$\sum_{l=0}^{N-1}\rho_l^2\equiv\|\rho[N]\|^2\leqslant\frac{1}{T}\int_0^T q^2(\tau)\,\mathrm{d}\tau\leqslant B_q^2$$

Hence, for any choice of N, we have

$$\|\rho[N]\|\leqslant B_q\tag{8}$$

Define an estimate $\hat q[N](t)$ of the Fourier approximation for $q(t)$ as $\hat q[N](t)=\sum_{l=0}^{N-1}\hat\rho_l(t)\varphi_l(t)\doteq\hat\rho[N](t)^{\mathrm T}\Phi_N(t)$ in which $\Phi_N(t)=[\varphi_0(t),\ldots,\varphi_{N-1}(t)]^{\mathrm T}$ and $\hat\rho[N](t)=[\hat\rho_0(t),\ldots,\hat\rho_{N-1}(t)]^{\mathrm T}$ with $\hat\rho_l(t)$ being the estimate of the lth Fourier coefficient ρ_l in $q(t)$, $0\leqslant l\leqslant N-1$.

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Before stating the main result of this paper, we recall the projection operator $\mathrm{Proj}[\xi, \hat{\rho}[N]]$ defined in [21], which will be used in the adaptation algorithm. We will define $\dot{\hat{\rho}}[N] = \mathrm{Proj}[\xi, \hat{\rho}[N]]$ with ξ being a suitable function and $\mathrm{Proj}[\xi, \hat{\rho}[N]]$ being the Lipschitz continuous function given by

$$\mathrm{Proj}[\xi, \hat{\rho}[N]] = \begin{cases} \xi & \text{if } \mathscr{C}_{*1} \\ \xi & \text{if } \mathscr{C}_{*2} \\ \left[I - \dfrac{s(\hat{\rho}[N])\mathrm{grad}[s(\hat{\rho}[N])]\mathrm{grad}[s(\hat{\rho}[N])]^{\mathrm{T}}}{\|\mathrm{grad}[s(\hat{\rho}[N])]\|^2} \right] \xi & \text{if } \mathscr{C}_{*3} \end{cases}$$

$$s(\hat{\rho}[N]) = \frac{\|\hat{\rho}[N]\|^2 - B_q^2}{v^2 + 2vB_q}$$

$$\mathscr{C}_{*1} : s(\hat{\rho}[N]) \leqslant 0$$

$$\mathscr{C}_{*2} : s(\hat{\rho}[N]) > 0 \quad \text{and} \quad \langle \mathrm{grad}[s(\hat{\rho}[N])], \xi \rangle \leqslant 0$$

$$\mathscr{C}_{*3} : s(\hat{\rho}[N]) > 0 \quad \text{and} \quad \langle \mathrm{grad}[s(\hat{\rho}[N])], \xi \rangle > 0$$

in which v is an arbitrary positive real and B_q is the radius of the closed ball $\overline{\mathscr{B}(0, B_q)}$ in $\mathbb{R}^N$ (with center the origin) in which $\rho[N]$ is constrained. If $\hat{\rho}[N](0) \in \overline{\mathscr{B}(0, B_q)}$, then the following properties hold: (1) $\|\hat{\rho}[N](t)\| \leqslant B_q + v, \ \forall t \geqslant 0$; (2) $\|\mathrm{Proj}[\xi, \hat{\rho}[N]]\| \leqslant \|\xi\|$; (3) $(\rho[N] - \hat{\rho}[N])^{\mathrm{T}} \mathrm{Proj}[\xi, \hat{\rho}[N]] \geqslant (\rho[N] - \hat{\rho}[N])^{\mathrm{T}} \xi$.

5. ROBUST ADAPTIVE DESIGN AND STABILITY ANALYSIS

5.1. Control algorithm

Define the stator current tracking errors:

$$e_d = i_{sd} - i_{sd}^*$$

$$e_q = i_{sq} - i_{sq}^*$$

in which i_{sd}^* and i_{sq}^* are the reference signals for the stator current vector (d, q) components. The proposed control algorithm, whose design relies on back-stepping and robust adaptive techniques (involving, as we shall see, the choice of suitable Lyapunov functions), consists of: (i) a modified indirect field-oriented control (see, for instance, [16])

$$\begin{aligned} i_{sd}^* &= \frac{\psi^*}{M} + \frac{\dot{\psi}^*}{\alpha M} - \frac{m i_{sq} \tilde{\omega}}{\lambda \alpha M} \\ \dot{\varepsilon}_0 = \omega_0 &= \omega + \frac{\alpha M}{\psi^*} i_{sq} - \frac{m i_{sd} \tilde{\omega}}{\lambda \psi^*} \end{aligned} \tag{9}$$

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containing stabilizing feedback terms; (ii) two outer rotor angle and speed cascaded control loops

$$
\begin{aligned}
i_{sq}^{*} &= \frac{1}{m\psi^{*}}\left[-k_{\omega}\tilde{\omega}-k_{v}\tilde{\theta}+\hat{\rho}[N]^{\mathrm{T}}\Phi_{N}-\left(\frac{\bar{k}_{\theta}}{2}+\frac{g_{1}}{2k_{\theta}k_{v}}+\frac{g_{2}}{2\bar{k}_{\theta}}+\frac{k}{4}+\frac{\bar{k}_{\omega}}{4}+\frac{k_{j1}^{2}}{4\bar{k}_{\omega}}\omega^{2}\right)\tilde{\omega}\right] \\
\dot{\hat{\rho}}[N] &= \mathrm{Proj}[-\mu\Phi_{N}\tilde{\omega},\hat{\rho}[N]], \quad \|\hat{\rho}[N](0)\|\leqslant\frac{B_{q}}{\sqrt{N}} \\
\mu &= \frac{\min\{2^{-1}k_{\theta},k_{\omega}J_{M}^{-1},4^{-1}\alpha,2k_{e}\}\mu_{0}k}{2}
\end{aligned}
\tag{10}
$$

containing the estimate $\hat{\rho}[N](t)$ of the Fourier coefficient vector $\rho[N]$ and suitable robustifying terms taking into account the state-dependent disturbance due to mechanical uncertainties; (iii) an inner stator current vector control loop

$$
\begin{aligned}
u_{sd} &= \sigma\left[\left(\frac{R_{s}}{\sigma}+\alpha M\beta\right)i_{sd}-\omega_{0}i_{sq}-\alpha\beta\psi^{*}+\Gamma_{0d}-k_{e}e_{d}-v_{d1}e_{d}-v_{d2}+\frac{m\beta}{\lambda\alpha M}\tilde{\omega}\omega\psi^{*}\right] \\
u_{sq} &= \sigma\left[\left(\frac{R_{s}}{\sigma}+\alpha M\beta\right)i_{sq}+\omega_{0}i_{sd}+\beta\omega\psi^{*}+\Gamma_{0q}-k_{e}e_{q}-v_{q1}e_{q}-v_{q2}\right]
\end{aligned}
\tag{11}
$$

in which the terms $(\Gamma_{0d},\Gamma_{0q},v_{d1},v_{q1},v_{d2},v_{q2})$ containing feed-forward actions and suitable stabilizing and robustifying feedback terms are given by

$$
\begin{aligned}
\Gamma_{0d} &= \frac{\dot{\psi}^{*}}{M}+\frac{\ddot{\psi}^{*}}{\alpha M}+\frac{mi_{sq}}{\lambda\alpha M}\dot{\omega}^{*}-\frac{m\tilde{\omega}}{\lambda\alpha M}\left[-\left(\frac{R_{s}}{\sigma}+\alpha M\beta\right)i_{sq}-\omega_{0}i_{sd}+\frac{1}{\sigma}u_{sq}\right] \\
\Gamma_{0q} &= \frac{1}{m\psi^{*}}\left[k_{\omega}\dot{\omega}^{*}-k_{v}\tilde{\omega}+k_{v}k_{\theta}\tilde{\theta}+\dot{\hat{\rho}}[N]^{\mathrm{T}}\Phi_{N}+\hat{\rho}[N]^{\mathrm{T}}\dot{\Phi}_{N}+\left(\frac{\bar{k}_{\theta}}{2}+\frac{g_{1}}{2k_{\theta}k_{v}}+\frac{g_{2}}{2\bar{k}_{\theta}}+\frac{k}{4}\right.\right. \\
&\quad \left.\left.+\frac{\bar{k}_{\omega}}{4}+\frac{k_{j1}^{2}}{4\bar{k}_{\omega}}\omega^{2}\right)\dot{\omega}^{*}-\left(\frac{\gamma_{0}}{2k_{\theta}k_{v}}+\frac{\tilde{\gamma}_{0}}{2\bar{k}_{\theta}}\right)\tilde{\omega}\right]-\frac{\dot{\psi}^{*}}{\psi^{*}}i_{sq}^{*} \\
v_{d1} &= \frac{1}{\alpha\lambda}\left[\left(\alpha\beta-\frac{m\beta}{\lambda\alpha M}\tilde{\omega}\omega\right)^{2}+\left(\beta\omega+\frac{m\beta}{\lambda M}\tilde{\omega}\right)^{2}\right]+\Gamma_{1d}^{2}\left(\frac{k}{2J_{m}^{2}}+\frac{2}{k_{\omega}}\Gamma_{\omega1}+\frac{4}{k_{\theta}k_{v}}\Gamma_{\omega2}\right. \\
&\quad \left.+\frac{2}{\alpha\lambda}(\Gamma_{\omega3}+\Gamma_{\omega4})+\frac{1}{2}\Gamma_{\omega5}+\frac{k}{2J_{m}^{2}}\|\Phi_{N}\|^{2}+\frac{4g_{1}}{k_{\theta}k_{v}J_{m}^{2}}+\frac{2g_{2}}{k_{\omega}J_{m}^{2}}\right)+2\alpha M^{2}\lambda \\
v_{q1} &= \frac{1}{\alpha\lambda}[(\alpha\beta)^{2}+(\beta\omega)^{2}]+\Gamma_{1q}^{2}\left(\frac{k}{2J_{m}^{2}}+\frac{2}{k_{\omega}}\Gamma_{\omega1}+\frac{4}{k_{\theta}k_{v}}\Gamma_{\omega2}+\frac{2}{\alpha\lambda}(\Gamma_{\omega3}+\Gamma_{\omega4})\right. \\
&\quad \left.+\frac{1}{2}\Gamma_{\omega5}+\frac{k}{2J_{m}^{2}}\|\Phi_{N}\|^{2}+\frac{4g_{1}}{k_{\theta}k_{v}J_{m}^{2}}+\frac{2g_{2}}{k_{\omega}J_{m}^{2}}\right)+1 \\
v_{d2} &= -\Gamma_{1d}\dot{\omega}^{*} \\
v_{q2} &= -\Gamma_{1q}\dot{\omega}^{*}+m\psi^{*}\tilde{\omega}
\end{aligned}
\tag{12}
$$

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with

$$\gamma_0 = 4k_{j1}^2 k_\theta^4 \tilde{\theta}(\tilde{\omega} - k_\theta \tilde{\theta}) - 16k_{j1}^2 k_\theta^2 \tilde{\omega}\dot{\omega}^*, \quad \tilde{\gamma}_0 = -2k_{j1}^2 \tilde{\omega}\dot{\omega}^*$$

$$\Gamma_{1d} = -\frac{m i_{sq}}{\lambda \alpha M}$$

$$\Gamma_{1q} = \frac{1}{m\psi^*}\left[-k_\omega - \frac{\bar{k}_\theta}{2} - \frac{g_1}{2k_\theta k_v} - \frac{g_2}{2\bar{k}_\theta} - \frac{k}{4} - \frac{\bar{k}_\omega}{4} - \frac{k_{j1}^2}{4\bar{k}_\omega}\omega^2 - \left(\frac{\gamma_1}{2k_\theta k_v} + \frac{\tilde{\gamma}_1}{2\bar{k}_\theta} + \frac{k_{j1}^2}{2\bar{k}_\omega}\omega\right)\tilde{\omega}\right]$$

$$\gamma_1 = 16k_{b2}^2\omega + 8k_{j2}^2\omega^3 + 16k_{j1}^2 k_\theta^2 \tilde{\omega}, \quad \tilde{\gamma}_1 = 2k_{j1}^2\tilde{\omega}$$

$$g_1 = 8J_M^2 k_\theta^4 + 2k_{j1}^2 k_\theta^4 \tilde{\theta}^2 + 8k_{j1}^2 k_\theta^2 c_{\theta 1}^2 + 8k_{j1}^2 k_\theta^2 \tilde{\omega}^2 + 2k_{j2}^2\omega^4 + 8k_{b1}^2 k_\theta^2 + 8k_{b2}^2\omega^2 + 8k_{j1}^2 c_{\theta 2}^2$$

$$g_2 = 4J_M^2 k_\theta^2 + k_{j1}^2\tilde{\omega}^2 + 4k_{j1}^2 c_{\theta 1}^2 + 4k_{b1}^2$$

$$\Gamma_{\omega 1} = \frac{1}{J_m^2}\left[-k_\omega - \frac{\bar{k}_\theta}{2} - \frac{g_1}{2k_\theta k_v} - \frac{g_2}{2\bar{k}_\theta} - \frac{k}{4} - \frac{\bar{k}_\omega}{4} - \frac{k_{j1}^2}{4\bar{k}_\omega}\omega^2\right]^2$$

$$\Gamma_{\omega 2} = \frac{k_v^2}{J_m^2}, \quad \Gamma_{\omega 3} = \frac{m^2 i_{sq}^2}{J_m^2}, \quad \Gamma_{\omega 4} = \frac{m^2 i_{sd}^2}{J_m^2}, \quad \Gamma_{\omega 5} = \frac{m^2\psi^{*2}}{J_m^2}$$

The proposed control law depends on: the arbitrary odd integer $N > 1$, the orthogonal basis functions $\Phi_N(t) = [\varphi_0(t), \ldots, \varphi_{N-1}(t)]^{\mathrm{T}}$ in (5), the positive control parameters $k_\theta, k_\omega, k_v, \bar{k}_\theta, \bar{k}_\omega, k, \lambda, \mu_0, k_e$ with $\mu_0 \geqslant 1$, the positive reals $J_m, J_M, k_{ji}, k_{bi}, c_{\theta i}$ $(i = 1, 2)$ in assumptions (A.3), (A.4) and B_q defined in (7).

5.2. *Stability analysis*

Let $z(t) = [z_1(t)^{\mathrm{T}}, z_2(t)^{\mathrm{T}}]^{\mathrm{T}}$ be the solution of the closed loop system with[‡] $z_1 = [\tilde{\theta}, \tilde{\omega}, \psi_{rd} - \psi^*, \psi_{rq}, i_{sd} - i_{sd}^*, i_{sq} - i_{sq}^*]^{\mathrm{T}}$ and $z_2 = \rho[N] - \hat{\rho}[N]$. Let $\overline{\mathscr{B}(0; m_{\mathscr{A}})}$ the closed ball in $\mathbb{R}^6$ with center the origin and arbitrary radius $m_{\mathscr{A}} > 0$. Let

$$c_1 = \max\{k_v, J_M, \lambda, 1\}(\min\{k_v, J_m, \lambda, 1\})^{-1}$$

$$c_2 = \min\{2^{-1}k_\theta, k_\omega J_M^{-1}, 4^{-1}\alpha, 2k_e\}, \quad c_3 = 6(c_2 \min\{k_v, J_m, \lambda, 1\})^{-1}$$

$$c_4 = c_1 c_2^{-1}, \quad c_5 = c_3(3c_2)^{-1}, \quad c_6 = 3^{-1}(2c_3)$$

be positive reals. The main result of the paper is stated in the following theorem.

[‡]According to the field-oriented control strategy, the control objective

$$\lim_{t\to\infty}[\sqrt{\psi_{ra}^2(t) + \psi_{rb}^2(t)} - \psi^*(t)] = 0$$

is achieved by guaranteeing

$$\lim_{t\to\infty}[\psi_{rd}(t) - \psi^*(t)] = 0$$

$$\lim_{t\to\infty}[\psi_{rq}(t)] = 0$$

which imply that the flux vector $(\psi_{ra}(t), \psi_{rb}(t))$ to asymptotically rotate at speed $\omega_0(t)$ and the d-axis of the (d, q) reference frame (rotating at speed $\omega_0(t)$) to be asymptotically coincident with the rotating flux vector.

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Theorem 1
Consider system (4) satisfying assumptions (A.1), (A.2), (A.3), an output reference signal $y_r(t)=[y_{r1}(t), y_{r2}(t)]^{\mathrm{T}}=[\theta^*(t), \psi^*(t)]^{\mathrm{T}}$ satisfying assumptions (A.4), (A.5) and the output feedback control (9)–(12). Then

- $z_2(t)$ satisfies, for any $t>0$, the following inequality: $\|z_2(t)\|\leqslant\sqrt{2B_q^2+2(B_q+v)^2}$;
- $z_1(t)$ satisfies, for any $t>0$, the following $\mathscr{L}_\infty$ and $\mathscr{L}_2$ inequalities [ε_N is the approximation bound in (6)]:

  (P1) $\|z_1(t)\|^2\leqslant c_1\|z_1(0)\|^2\mathrm{e}^{-c_2t}+(c_3/k)[\varepsilon_N^2+\max_{0\leqslant\tau\leqslant t}\{\|z_2(\tau)\|^2\}]$;
  (P2) $\int_0^t\|z_1(\tau)\|^2\,\mathrm{d}\tau\leqslant c_4\|z_1(0)\|^2+(c_5/k)\|z_2(0)\|^2+(c_6/k)\int_0^t[\varepsilon_N^2+\|z_2(\tau)\|^2]\,\mathrm{d}\tau$;

- for any initial condition $[\theta(0), \omega(0), \psi_{rd}(0), \psi_{rq}(0), i_{sd}(0), i_{sq}(0)]^{\mathrm{T}}\in\overline{\mathscr{B}(0; m_{\mathscr{A}})}$, there exists an explicitly computable positive real $k_e^*(N, m_{\mathscr{A}})$ such that for any control parameter $k_e>k_e^*(N, m_{\mathscr{A}})$, $z(t)$ satisfies the following properties [$p=\min\{s_\theta-2, s_j-1, s_b\}$]:

  (P3) $z(t)$ is exponentially attracted, as $t\to+\infty$, into a closed ball of radius $r(N)$, with $r(N)=O\left(N^{2-p/2}\right)$ for $N\to+\infty$;
  (P4) if there exists an odd integer N^* such that $q(t)=\sum_{l=0}^{N^*-1}\rho_l\varphi_l(t)$, then the control algorithm with $N\geqslant N^*$ guarantees that $\|z(t)\|$ exponentially tends to zero.

Constructive Proof
The first part of the theorem follows on inequality (8) and property (1) of the projection algorithm. Define the rotor flux tracking errors:

$$\tilde{\psi}_{rd}=\psi_{rd}-\psi^*$$
$$\tilde{\psi}_{rq}=\psi_{rq}$$

so that we can write

$$J(\theta)\dot{\tilde{\omega}}=m\psi^*i_{sq}^*+m\psi^*e_q+m(\tilde{\psi}_{rd}i_{sq}-\tilde{\psi}_{rq}i_{sd})-\rho[N]^{\mathrm{T}}\Phi_N-\varepsilon-g(\tilde{\theta}, \tilde{\omega}, \omega)$$
$$\dot{\tilde{\psi}}_{rd}=-\alpha\tilde{\psi}_{rd}-\alpha\psi^*+(\omega_0-\omega)\tilde{\psi}_{rq}+\alpha Mi_{sd}^*+\alpha Me_d-\dot{\psi}^*$$
$$\dot{\tilde{\psi}}_{rq}=-\alpha\tilde{\psi}_{rq}-(\omega_0-\omega)\tilde{\psi}_{rd}-(\omega_0-\omega)\psi^*+\alpha Mi_{sq}$$

Design the reference signals i_{sd}^*, i_{sq}^* for the stator currents i_{sd}, i_{sq} and the speed ω_0 of the rotating (d, q) frame as in (9)–(10) so that the quadratic function

$$W=\frac{1}{2}\left(k_v\tilde{\theta}^2+J(\theta)\tilde{\omega}^2+\lambda(\tilde{\psi}_{rd}^2+\tilde{\psi}_{rq}^2)+\frac{\|z_2\|^2}{\mu}\right)$$

admits time derivative along the trajectories of the closed-loop system satisfying

$$\dot{W}\leqslant-\frac{k_vk_\theta}{2}\tilde{\theta}^2-k_\omega\tilde{\omega}^2-\alpha\lambda(\tilde{\psi}_{rd}^2+\tilde{\psi}_{rq}^2)+m\psi^*\tilde{\omega}e_q+\alpha M\lambda\tilde{\psi}_{rd}e_d+\frac{\varepsilon^2}{k}$$

Compute

$$\dot{e}_d=-\left(\frac{R_s}{\sigma}+\beta\alpha M\right)i_{sd}+\frac{1}{\sigma}u_{sd}+\omega_0i_{sq}+\beta\alpha\psi_{rd}+\beta\omega\psi_{rq}-\frac{\mathrm{d}i_{sd}^*}{\mathrm{d}t}$$

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$$\dot{e}_q = -\left(\frac{R_s}{\sigma} + \beta\alpha M\right) i_{sq} + \frac{1}{\sigma} u_{sq} - \omega_0 i_{sd} + \beta\alpha\psi_{rq} - \beta\omega\psi_{rd} - \frac{\mathrm{d}i_{sq}^*}{\mathrm{d}t}$$

and design the stator voltages u_{sd}, u_{sq} as in (11)–(12) so that the quadratic function

$$V = W + \tfrac{1}{2}(e_d^2 + e_q^2)$$

admits time derivative along the trajectories of the closed-loop system satisfying

$$\dot{V} \leqslant -\frac{k_v k_\theta}{4}\tilde{\theta}^2 - \frac{k_\omega}{2}\tilde{\omega}^2 - \frac{\alpha\lambda}{8}(\tilde{\psi}_{rd}^2 + \tilde{\psi}_{rq}^2) - k_e(e_d^2 + e_q^2) + \frac{2\varepsilon_N^2}{k} + \frac{\|z_2\|^2}{k}$$

$$\leqslant -c_2 V + \frac{2\varepsilon_N^2}{k} + \frac{2\|z_2\|^2}{k}$$

on which inequalities (P1) and (P2) follow.

The remaining part of the theorem can be proved by considering the quadratic function [a_p, s_* are suitable positive reals]

$$\mathscr{V} = W + \frac{1}{2} a_p \| Q(t) z_2 - \Omega(t)[\tilde{\theta}, \tilde{\omega}, \tilde{\psi}_{rd}, \tilde{\psi}_{rq}]^{\mathrm{T}} \|^2 + \frac{s_*}{2}(e_d^2 + e_q^2)$$

in which

$$\Omega^{\mathrm{T}} = \begin{bmatrix} 0 & \cdots & \cdots & 0 \\ -\dfrac{\varphi_0}{J(\theta)} & \cdots & \cdots & -\dfrac{\varphi_{N-1}}{J(\theta)} \\ 0 & \cdots & \cdots & 0 \\ 0 & \cdots & \cdots & 0 \end{bmatrix}$$

and $Q(t)$ is generated by the filter [I is the identity matrix]

$$\dot{Q}(t) = -Q(t) + \Omega(t)\Omega^{\mathrm{T}}(t), \quad Q(0) = \frac{T}{J_M^2} I$$

and by applying arguments similar to those used to prove the Persistency of Excitation Lemma in [13]. □

6. SIMULATION RESULTS

We tested by simulation a digital implementation of the proposed adaptive learning control algorithm with reference to a mechanism similar to the one reported in [9, 10] (see Figure 1), in which the rotor shaft of an one pole pair, 1.1 kW, 2850 rpm (50 Hz), 3.5 Nm induction motor is connected to a sliding mass. The equivalent total inertia $J(\theta)$ and linear viscous friction coefficient $b(\theta)$ are:

$$J(\theta) = J_0 + m_0\left[\frac{\partial x(\theta)}{\partial\theta}\right]^2 \quad \text{and} \quad b(\theta) = b_0 + b_v\left[\frac{\partial x(\theta)}{\partial\theta}\right]^2$$

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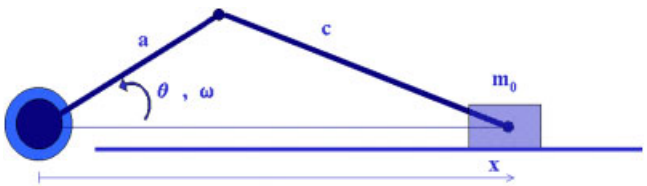


Figure 1. Mechanism.

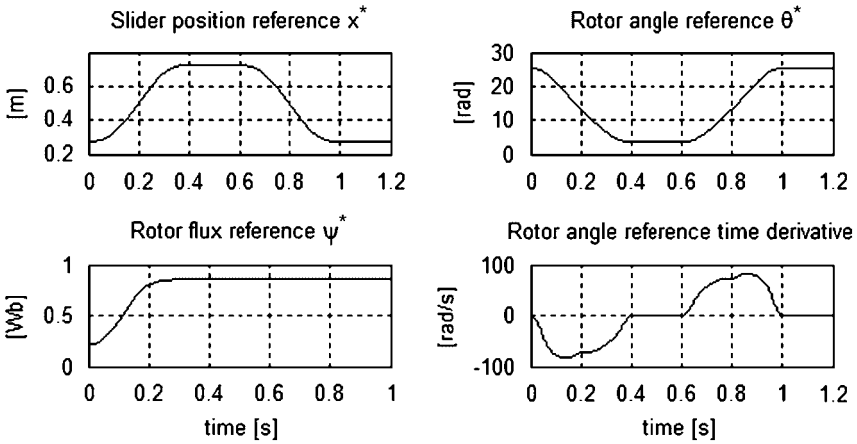


Figure 2. Slider position reference $x^*(t)$; rotor angle reference $\theta^*(t)$; rotor flux modulus reference $\psi^*(t)$; rotor angle reference time derivative $\dot{\theta}^*(t)$.

The kinematic relation $x(\theta)$ between the slider position x and the induction motor rotor angle θ is:

$$x(\theta) = a\cos\left(\frac{\theta}{r}\right) + \sqrt{c^2 - a^2\sin^2\left(\frac{\theta}{r}\right)}$$

The system parameters are: $R_s = 10.4\,\Omega$, $R_r = 4.5\,\Omega$, $L_s = L_r = 0.47\,\mathrm{H}$, $M = 0.434\,\mathrm{H}$, $a = 0.25\,\mathrm{m}$, $c = 0.5\,\mathrm{m}$, $r = 10$, $m_0 = 16\,\mathrm{Kg}$, $J_0 = 0.005\,\mathrm{kg\,m^2}$, $b_0 = 0.025\,\mathrm{Nms/rad}$, $b_v = 16\,\mathrm{Ns/m}$ (m_0, J_0, b_0, b_v are uncertain parameters with a $\pm 20\%$ uncertainty).

After a flux excitation phase (the rotor flux modulus reference starts from 0.21 Wb at $t = 0$ and grows up to the rated constant value 0.86 Wb), the slider mass displacement $x(\theta)$ is required to track a periodic (with period $T = 1.2\,\mathrm{s}$) reference signal $x^*(\cdot)$, which starts from $c - 0.9a = 0.275\,\mathrm{m}$ at $t = 0.4\,\mathrm{s}$ and grows up to $c + 0.9a = 0.625\,\mathrm{m}$ with maximum speed and acceleration modulus equal to $2\,\mathrm{m/s}$ and $15\,\mathrm{m/s^2}$, respectively. The rotor angle reference signal θ^* is computed by inverting the function $x^*(\theta^*)$. The reference signals are shown in Figure 2, while the torque disturbance and its Fourier coefficients are reported in Figure 3, showing that the harmonic content of function $q(t)$ is approximately up to the 13th harmonic (10.8 Hz).

A simplified discrete-time version of the proposed $(N+1)$ order dynamic adaptive learning control algorithm (9)–(12), in which the robustifying terms (v_{d1}, v_{q1}) are set to zero, is obtained by the Euler discretization method with a sampling time $T_s = 200\,\mu\mathrm{s}$ and then tested by simulation.

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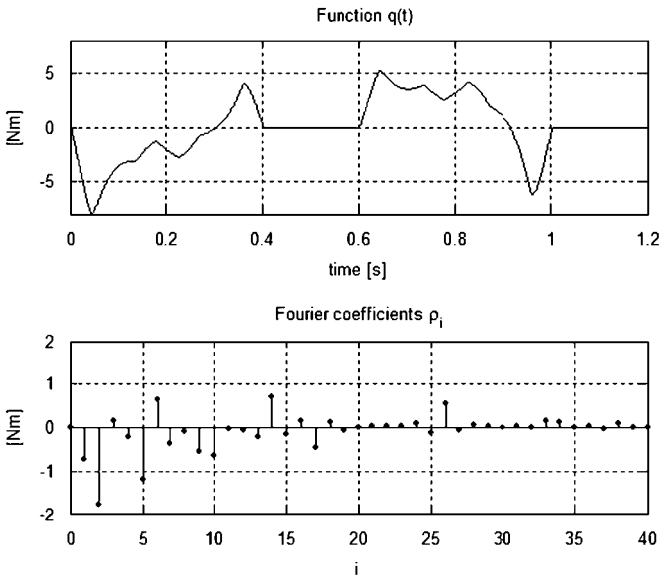


Figure 3. Function $q(t)$ and its Fourier coefficients ρ_i $(0 \leqslant i \leqslant 40)$.

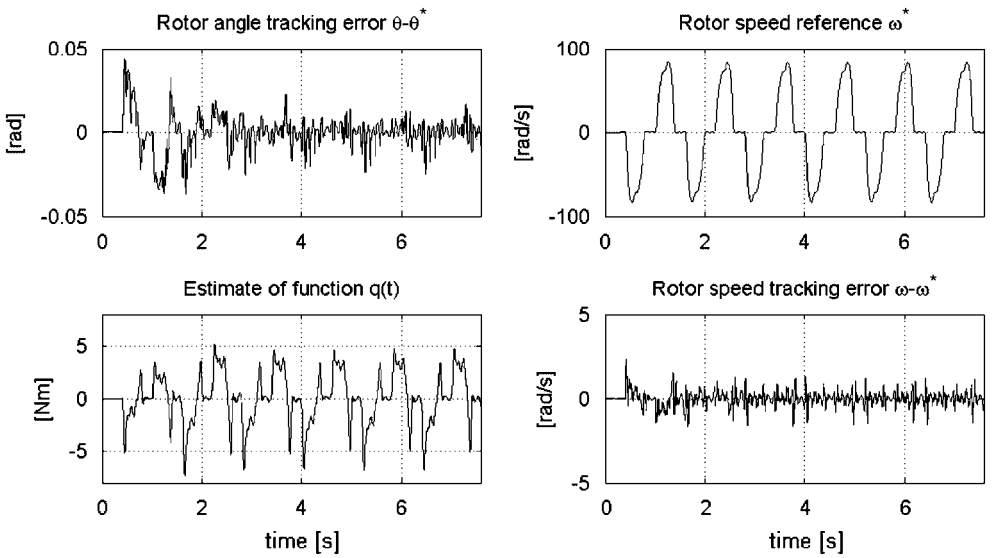


Figure 4. Rotor angle tracking error $\tilde{\theta}(t)$; rotor speed reference $\omega^*(t)$; estimate $\hat{q}$ [19](t) of the torque disturbance; rotor speed tracking error $\tilde{\omega}(t)$.

The control parameters, which were chosen according to suitable tuning rules for controllers with cascaded structure, are (all values are in SI units): $k_\theta = 20, k_\omega = 0.12, k_v = 30, \bar{k}_\theta = 0.25, \bar{k}_\omega = 0.01, k = 0.01, \lambda = 100, \mu = 5, k_e = 1000, N = 19$. All motor and controller initial conditions are set

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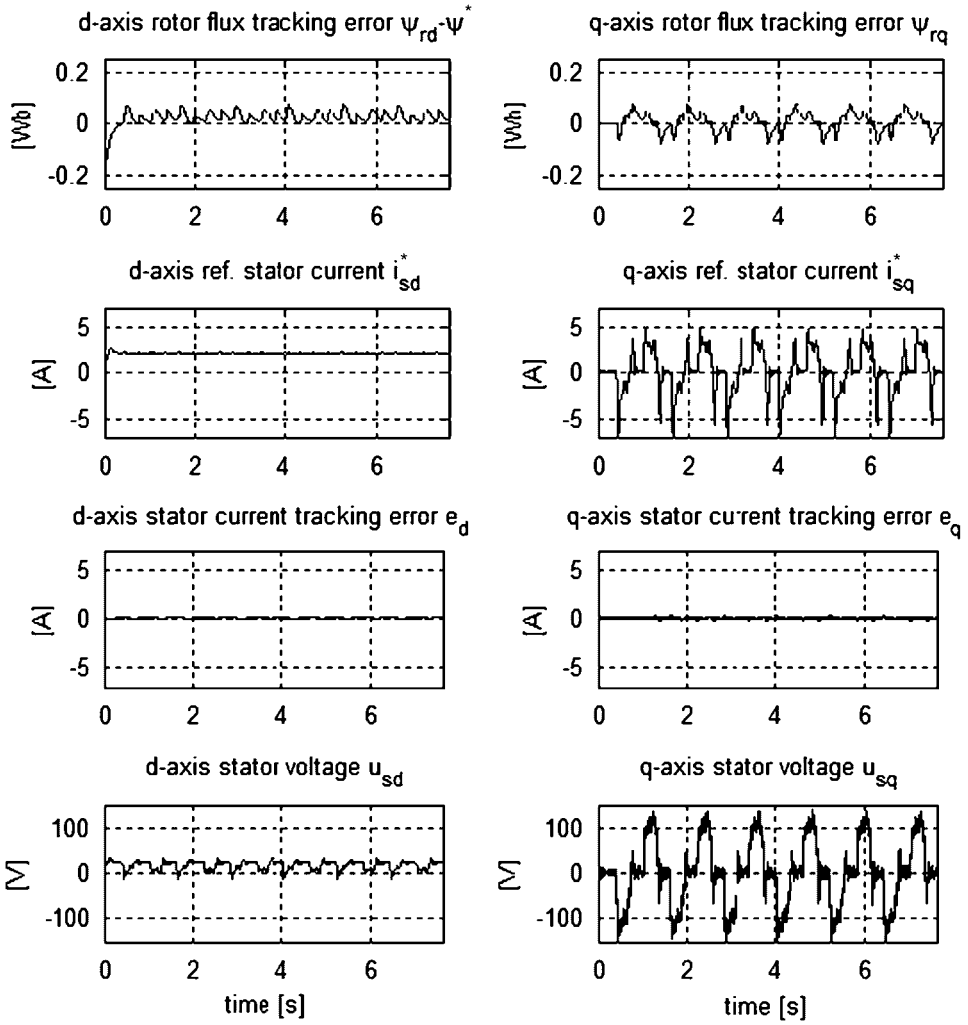


Figure 5. Rotor flux tracking errors $(\psi_{rd}(t)-\psi^*(t), \psi_{rq}(t))$; stator current references $(i^*_{sd}(t), i^*_{sq}(t))$; stator current vector (d,q) components tracking errors $(e_d(t), e_q(t))$; stator voltage vector (d,q) components $(u_{sd}(t), u_{sq}(t))$.

to zero excepting for $\theta(0)=\theta^*(0)=25.2$ rad. The rotor angle measurement is provided by a 5000 ppr encoder, while the rotor speed is obtained by using a discrete-time filtered time derivative of the measured rotor angle, with a 500 rad/s cut-off frequency of the first-order low-pass filter. The stator current signals are sampled by using a 12 bit ADC with 5 mA quantization amplitude. A time delay of one sampling period is introduced in order to take into account delays due to acquisition, computation and actuation. The tracking errors, the disturbance estimation, the controller internal signals along with the stator voltage vector (d,q) components are reported in Figures 4 and 5. The Fourier coefficients ρ_i and their estimates $\hat{\rho}_i, 0 \leqslant i \leqslant 3$, are shown in

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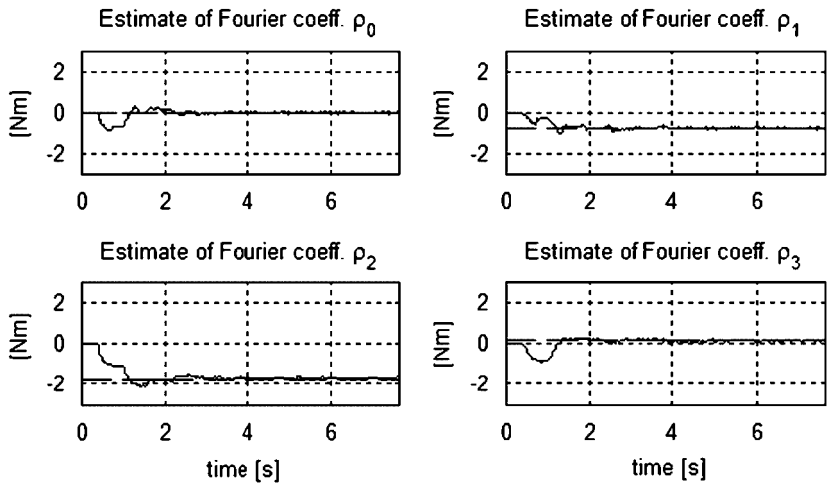


Figure 6. Fourier coefficients ρ_i and their estimates $\hat{\rho}_i(t)$ $(0 \leqslant i \leqslant 3)$.

Figure 6: good disturbance estimation and rotor position and flux modulus tracking precision are achieved while non-zero asymptotic tracking errors are due to truncation of the disturbance Fourier series, discrete-time implementation of the controller, rotor speed filtering, measurement noise, 200 μs time delay. High-frequency noise on the reference currents and stator voltages appears due to the encoder quantization error superimposed to the rotor angle and speed. The estimation of the torque disturbance and its compensation by means of feed-forward actions allow the controller to obtain satisfactory rotor angle and flux modulus tracking precision along with field orientation in the presence of mechanical uncertainties. The presented simulations for the digital version of the controller demonstrate that it is suitable for implementation on a real induction motor servo drive.

7. CONCLUSIONS

Under the assumption that the rotor angle reference signal is periodic with known period, an output feedback learning control has been designed for induction motor servo drives with mechanical uncertainties. $\mathscr{L}_2$ and $\mathscr{L}_\infty$ transient performances for the rotor position and flux modulus tracking errors are guaranteed in the learning phase while, for any motor initial condition in an arbitrary given compact set, by properly setting the control parameters the following properties hold: (i) the guaranteed rotor position and flux modulus tracking errors are reduced by increasing the number N in the truncated Fourier series; (ii) when the uncertain function $q(t)$ is represented by a finite Fourier series, it is exponentially reconstructed while the rotor position and flux modulus tracking errors exponentially converge to zero. Further research will be devoted to deal with rotor resistance uncertainties, which may have a detrimental effect on rotor flux tracking.

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