Robust nonlinear cascade control strategy for permanent magnet synchronous motor drives

Boualem SALHI[1]

***Abstract*— This paper presents a robust nonlinear cascade control strategy for angular velocity trajectory tracking of a permanent magnet synchronous motor under parametric uncertainties, unknown time-varying load torque, and unknown initial rotor position. The control strategy is synthesized based on the recursive gain technique. The proposed controller does not suffer from the "explosion of complexity" problem encountered in robust backstepping-based controllers, nor does it require the implementation of a disturbance observer, unlike controllers based on Active Disturbance Rejection Control. Simulation results are provided to illustrate the effectiveness of the proposed control strategy.**

I. INTRODUCTION

Due to its excellent characteristics, such as high efficiency, low noise level, high power density, and ease of maintenance, the permanent magnet synchronous motor (PMS motor) has gained a prominent place in industrial sectors such as electric vehicles, aviation, and robotics. Despite all these advantages, the control performance of the PMS motor, like that of other electrical motors, can be degraded by variations in its parameters around their nominal values. Furthermore, the controller must compensate for unknown load torque disturbances. Another very important issue is the lack of exact knowledge of the initial rotor position, which is required to determine the angular position of the motor magnetic flux.

Taking into account parametric variations and load torque disturbances, or only the latter, several authors have investigated the robust control problem of the PMS motor, where the initial rotor position is considered under different scenarios. Using incremental encoders, several robust control strategies have been proposed, such as those based on active disturbance rejection control [1], [2], [3]. By considering the initial rotor position as an uncertain motor parameter, a robust observer-based adaptive controller for the PMS motor was developed in [4]. Estimating the rotor position from speed measurements using a speed sensor, while taking into account the unknown initial rotor position, was addressed in robust control designs based on active disturbance rejection control and nonsingular terminal sliding mode in [5], [6].

This article presents a robust speed sensor-based control design for a PMS motor that takes into account parameter uncertainties and unknown load torque disturbances. Moreover, it investigates the effect of the lack of exact knowledge of the initial rotor position on the motor operation. Based on the recursive gain technique [7], [8], the proposed robust controller is shown to be simpler in design and easier to implement than other existing approaches.

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II. MOTOR UNCERTAIN MATHEMATICAL MODEL

In this section, we introduce the uncertain mathematical model that will be used to design the control strategy for the permanent magnet synchronous motor. First, we recall that, in the rotating *d-q* reference frame, this motor is governed by the following differential equations [5]:

$$\begin{aligned}
\frac{d}{dt} i_d &= -\frac{R}{L}\ i_d + p\ \Omega\ i_q + \frac{p\ K_e}{L}\ \Omega\ \sin(p\ \theta_0) + \frac{1}{L}\ u_d \\
\frac{d}{dt} i_q &= -\frac{R}{L}\ i_q - p\ \Omega\ i_d - \frac{p\ K_e}{L}\ \Omega\ \cos(p\ \theta_0) + \frac{1}{L}\ u_q \\
\frac{d}{dt}\Omega &= \frac{1}{J}\ T_{em} - \frac{f}{J}\ \Omega - \frac{1}{J}\ T_l
\end{aligned} \tag{1}$$

with the motor electromechanical torque T_{em} given as

$$T_{em} = p\ K_e\ [i_q\ \cos(p\ \theta_0) - i_d\ \sin(p\ \theta_0)] \tag{2}$$

where (u_d, u_q) are the control input voltages, (i_d, i_q) are the measured stator currents, Ω is the measured angular velocity, T_l is the unknown load torque, and p is the number of pole pairs. The parameters R, L, J and f denote the stator resistance and inductance, the motor inertia, and the viscous damping coefficient, respectively. The angle θ_0 represents the unknown initial mechanical rotor position, which is related to the unknown absolute rotor position θ_r by

$$\theta_r = \theta + \theta_0 \tag{3}$$

where, for $\Omega(0) = 0$, the mechanical rotor position calculated from the angular velocity measurement is given by

$$\theta = p\ \int_0^t \Omega(\tau)\ d\tau \tag{4}$$

As explained in [5], the PMS motor $d-q$ model given in (1) is derived from the $\alpha-\beta$ model by using the known rotor position θ instead of the unknown absolute rotor position θ_r. It is important to note that there is a difference between the model given in (1) and the one presented in [5]. Indeed, in the latter reference, the effect of the unknown initial rotor position on the electromagnetic torque is neglected. Moreover, for $\theta_0 = 0$, the traditional model in the rotating $d-q$ reference frame is recovered.

In addition to the unknown initial rotor position and load torque, we assume that the motor parameters vary over time. Parameter variations are taken into account by assuming that each parameter is equal to its perfectly known nominal value plus an unknown variation term, as follows:

$$
\begin{aligned}
R &= R_n + \delta R & J &= J_n + \delta J \\
L &= L_n + \delta L & f &= f_n + \delta f
\end{aligned} \tag{5}
$$

where R_n, L_n, J_n, and f_n are the known nominal motor parameters. Moreover, for a small initial rotor position such that $p\ \theta_0 \ll 1$, the Taylor series expansions of $\cos(p\ \theta_0)$ and $\sin(p\ \theta_0)$ in a neighbourhood of 0 are:

$$
\begin{aligned}
\cos(p\ \theta_0) &= 1 + \Sigma_{i=1}^{n/2}\ (-1)^i\ \frac{[p\ \theta_0]^{2i}}{(2\ i)!} + R_{cos}(p\ \theta_0) \\
\sin(p\ \theta_0) &= \Sigma_{i=0}^{(n-1)/2}\ (-1)^i\ \frac{[p\ \theta_0]^{2i+1}}{(2\ i+1)!} + R_{sin}(p\ \theta_0)
\end{aligned} \tag{6}
$$

Substituting (5) and (6) into Eqs. (1) and (2), and using the following relation, which holds for all $x, y \in \mathbb{R}$

$$
\frac{y}{x} = \frac{y_n + \delta y}{x_n + \delta x} = \frac{y_n}{x_n} + \frac{\delta y}{x} - \frac{y_n\ \delta x}{x\ x_n} \tag{7}
$$

for parameter variations, then the resulting uncertain mathematical model of the PMS motor is given by the following three first-order systems:

$$
\frac{d}{dt}\Omega = \frac{1}{J_n}\ T_{em,n} - \frac{f_n}{J_n}\ \Omega + \Delta_1 \tag{8}
$$

$$
\frac{d}{dt}i_q = \frac{1}{L_n}\ u_q - \frac{R_n}{L_n}\ i_q - p\ \Omega\ i_d - \frac{p\ K_e}{L_n}\ \Omega + \Delta_2 \tag{9}
$$

$$
\frac{d}{dt}i_d = \frac{1}{L_n}\ u_d - \frac{R_n}{L_n}\ i_d + p\ \Omega\ i_q + \Delta_3 \tag{10}
$$

with the nominal motor electromechanical torque

$$
T_{em,n} = p\ K_e\ i_q \tag{11}
$$

and where the uncertain terms are found as:

$$
\begin{aligned}
\Delta_1 &= \frac{\delta T_{em}}{J} - \frac{T_{em,n}\ \delta J}{J\ J_n} + \left[\frac{f_n\ \delta J}{J\ J_n} - \frac{\delta f}{J}\right]\Omega - \frac{1}{J}\ T_L \\
\Delta_2 &= -\frac{\delta L}{L\ L_n}\ u_q + \left[\frac{R_n\ \delta L}{L\ L_n} - \frac{\delta R}{L}\right]\ i_q \\
&\quad + p\ K_e\ \Omega\ \left[\frac{\delta L}{L\ L_n} - \frac{GS}{L}\right] \\
\Delta_3 &= -\frac{\delta L}{L\ L_n}\ u_d + \left[\frac{R_n\ \delta L}{L\ L_n} - \frac{\delta R}{L}\right]\ i_d + \frac{p\ K_e}{L}\ \Omega\ \sin(p\ \theta_0)
\end{aligned} \tag{12}
$$

with

$$
\begin{aligned}
\delta T_{em} &= p\ K_e [GS\ i_q - \sin(p\ \theta_0)\ i_d] \\
GS &= \Sigma_{i=1}^{n/2}\ (-1)^i\ \frac{[p\ \theta_0]^{2i}}{(2\ i)!} + R_{cos}(p\ \theta_0)
\end{aligned} \tag{13}
$$

III. ROBUST CONTROL DESIGN

The cascade robust control strategy used in this study is depicted in Fig. 1. The control strategy consists of three control blocks, each of which deals with one of the three uncertain first-order dynamics that model the PMS motor given by Eqs. (8), (9), and (10), respectively. Therefore, for uncertain first-order dynamical system, a modified design based on the recursive gain technique is first developed.

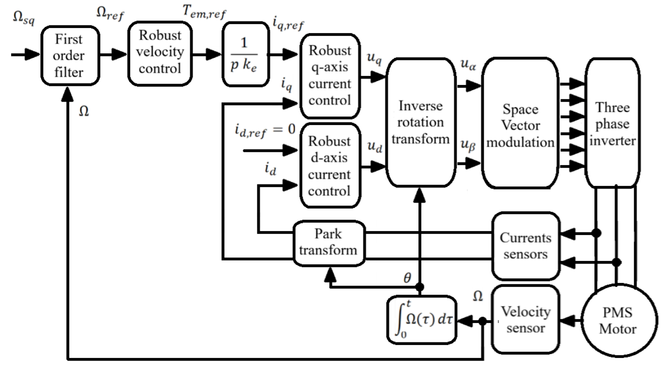


Fig. 1: Block diagram of the cascade control strategy.

Thus, consider the following first-order nonlinear system with uncertainty:

$$
\dot{y} = a\ u + f(y, \theta, \eta) + \Delta(u, y, \eta, \delta a, \delta\theta, \gamma) \tag{14}
$$

where $y \in \mathbb{R}$ is the system output to be controlled, $u \in \mathbb{R}$ is the control input, and $\eta \in \mathbb{R}^m$ is the state of any other nonlinear system connected to system (14). The scalar constant $a \neq 0$ and the vector $\theta \in \mathbb{R}^n$ are the known nominal high-frequency gain and system parameters, respectively. The scalar function $f(y, \theta, \eta)$ represents the nominal nonlinear part of the system, assumed to be perfectly known and bounded. Parametric uncertainties δa and $\delta\theta$, together with the disturbance $\gamma \in \mathbb{R}$, are lumped into the uncertain scalar term $\Delta(u, y, \eta, \delta a, \delta\theta, \gamma)$, which is assumed to be bounded by an unknown positive constant $\rho > 0$, as follows:

$$
|\Delta(u, y, \eta, \delta a, \delta\theta, \gamma)| \leq \rho \tag{15}
$$

The robust control design for the uncertain nonlinear system (14) is based on the recursive gain technique. Therefore, the convergence analysis is carried out for the errors $\tilde{y}$, $\tilde{\alpha}$ and e, which are defined as follows:

$$
\begin{aligned}
z &= y - y_{ref} = \tilde{y} + \tilde{\alpha} \\
e &= \int_0^t z(\tau) d\tau
\end{aligned}
\quad \text{with} \quad
\begin{aligned}
\tilde{y} &= y - \phi \\
\tilde{\alpha} &= \phi - y_{ref}
\end{aligned} \tag{16}
$$

where the desired reference signal y_{ref} is bounded and continuous, with a bounded first derivative.

The control input u is defined as:

$$
u = \frac{1}{a}\ \left[-k\ y - f(y, \theta, \eta) + k\ y_{ref} - \sigma\ e\right] \tag{17}
$$

while the fictitious signal ϕ is given by:

$$
\dot{\phi} + k\ \phi = k\ y_{ref} - \sigma\ e + \Delta(u, y, \eta, \delta a, \delta\theta, \gamma) \tag{18}
$$

where the constants $k > 0$ and $\sigma > 0$ are positive control gains.

Theorem 1: For the uncertain first-order nonlinear system (14), satisfying the previously stated boundedness assumptions, the feedback control law (17) ensures global tracking of the desired reference signal y_{ref} if, for the chosen control gains $k > 0$ and $\sigma > 0$, there exist a scalar constant $\beta > 0$ and a positive-definite matrix $P = P^T > 0$ such that the Lyapunov inequality

$$A^T\ P + P\ A + P\ P \leq -\beta\ P$$

has a solution. In addition, all signals are bounded such that $z,\ y,\ u \in L_\infty$.

Proof: The proof of the above theorem is carried out in the following two steps.

Step 1

In this step, we analyse the convergence of the error $\tilde{y}$. To this end, consider the following Lyapunov function:

$$V_1 = \frac{1}{2}\ \tilde{y}^2 \tag{19}$$

whose time derivative is given by:

$$\dot{V}_1 = \tilde{y}\ \dot{\tilde{y}} = \tilde{y}\ (\dot{y} - \dot{\phi}) \tag{20}$$

thus, using (14) and (18) we obtain:

$$\dot{V}_1 = \tilde{y}\ (a\ u + f(y,\theta,\eta) + k\ \phi - k\ y_{ref} + \sigma\ e) \tag{21}$$

and by substituting the control u given by (17), we have that:

$$\dot{V}_1 = \tilde{y}\ (-k\ y + k\ \phi) = -k\ \tilde{y}^2 \tag{22}$$

using (19), we finally get:

$$\dot{V}_1 = -2\ k\ V_1 \tag{23}$$

Eq. (23) shows that, for any $k > 0$, the Lyapunov function V_1 and, consequently, the error $\tilde{y}$ converge exponentially to zero. Moreover, it can be readily concluded that $\tilde{y} \in L_\infty \cap L_2$.

Step 2

The convergence analysis of the errors $\tilde{\alpha}$ and e is carried out in this second step. Therefore, we consider the following second-order system:

$$\begin{aligned} \dot{\tilde{\alpha}} &= \dot{\phi} - \dot{y}_{ref} \\ \dot{e} &= z = \tilde{y} + \tilde{\alpha} \end{aligned} \tag{24}$$

by substituting the fictitious signal (18) in (24), we obtain:

$$\begin{aligned} \dot{\tilde{\alpha}} &= -k\ \phi + k\ y_{ref} - \sigma\ e + \Delta(u,y,\eta,\delta a,\delta\theta,\gamma) - \dot{y}_{ref} \\ \dot{e} &= \tilde{\alpha} + \tilde{y} \end{aligned} \tag{25}$$

or equivalently

$$\begin{aligned} \dot{\tilde{\alpha}} &= -k\ \tilde{\alpha} - \sigma\ e + \Delta(u,y,\eta,\delta a,\delta\theta,\gamma) - \dot{y}_{ref} \\ \dot{e} &= \tilde{\alpha} + \tilde{y} \end{aligned} \tag{26}$$

which is rewritten in matrix form as:

$$\begin{bmatrix} \dot{\tilde{\alpha}} \\ \dot{e} \end{bmatrix} = \begin{bmatrix} -k & -\sigma \\ 1 & 0 \end{bmatrix} \begin{bmatrix} \tilde{\alpha} \\ e \end{bmatrix} + \begin{bmatrix} \Delta(u,y,\eta,\delta a,\delta\theta,\gamma) - \dot{y}_{ref} \\ \tilde{y} \end{bmatrix} \tag{27}$$

By assumption, the uncertain term $\Delta(u,y,\eta,\delta a,\delta\theta,\gamma)$ is bounded by an unknown constant ρ, the error $\tilde{y}$ is bounded by $\tilde{y}(0)$, as established in Step 1, and the reference signal y_{ref} is chosen as a bounded continuous signal with a bounded first derivative such that there exist an unknown constant $c > 0$ verifying

$$|\dot{y}_{ref}| \leq c \tag{28}$$

then system (27) is as follows:

$$\begin{bmatrix} \dot{\tilde{\alpha}} \\ \dot{e} \end{bmatrix} \leq \begin{bmatrix} -k & -\sigma \\ 1 & 0 \end{bmatrix} \begin{bmatrix} \tilde{\alpha} \\ e \end{bmatrix} + \begin{bmatrix} \rho + c \\ \tilde{y}(0) \end{bmatrix} \tag{29}$$

Thus, consider the following vector:

$$\xi = \begin{bmatrix} \tilde{\alpha} \\ e \end{bmatrix} \tag{30}$$

whose dynamic is given by:

$$\dot{\xi} \leq A\ \xi + B \tag{31}$$

with

$$A = \begin{bmatrix} -k & -\sigma \\ 1 & 0 \end{bmatrix} \qquad B = \begin{bmatrix} \rho + c \\ \tilde{y}(0) \end{bmatrix}$$

Consider the following Lyapunov function, where $P = P^T > 0$ is a positive-definite matrix:

$$V_2 = \xi^T\ P\ \xi \tag{32}$$

Differentiating V_2 with respect to time yields:

$$\begin{aligned} \dot{V}_2 &= \dot{\xi}^T\ P\ \xi + \xi^T\ P\ \dot{\xi} \\ \dot{V}_2 &\leq\ \xi^T\ \left[A^T\ P + P\ A\right]\ \xi + 2\ \xi^T\ P\ B \end{aligned} \tag{33}$$

Then, applying Young's inequality gives:

$$\dot{V}_2 \leq\ \xi^T\ \left[A^T\ P + P\ A + P\ P\right]\ \xi + B^T\ B \tag{34}$$

If, for the chosen control gains k and σ, there exists a positive constant $\beta > 0$ such that the following Lyapunov inequality

$$A^T\ P + P\ A + P\ P \leq -\beta\ P$$

admits a solution, then the Lyapunov derivative in (34) is equivalent to

$$\begin{aligned} \dot{V}_2 &\leq\ -\beta\ \xi^T\ P\ \xi + B^T\ B \\ &\leq\ -\beta\ V_2 + (\rho + c)^2 + \tilde{y}(0)^2 \end{aligned} \tag{35}$$

This shows that the Lyapunov function V_2 converges exponentially. Hence, the error vector ξ also converges exponentially and satisfies the following bound:

$$\xi \leq \sqrt{\frac{(\rho + c)^2 + \tilde{y}(0)^2}{\beta\ \lambda_{min}(P)}} \tag{36}$$

Thus, we have $\xi \in L_\infty$ and, consequently, $\tilde{\alpha}, e \in L_\infty$. In addition, since $\tilde{\alpha}, \tilde{y} \in L_\infty$, it follows that $z \in L_\infty$, which further implies that $y \in L_\infty$. Therefore, it can be concluded that the control input u is bounded, i.e., $u \in L_\infty$.

Moreover, by considering the steady-state condition of the second dynamic in system (26), namely

$$0 = \tilde{\alpha}_{eq} + \tilde{y}$$

and using the result from Step 1, which guarantees the exponential convergence of $\tilde{y}$ to zero, we obtain $\tilde{\alpha}_{eq} = 0$.

In this case, the first dynamic of system (26) at steady state becomes:

$$0 = -\sigma \ e_{eq} + \Delta(u, y, \eta, \delta a, \delta\theta, \gamma) - \dot{y}_{ref} \tag{37}$$

from which it follows that:

$$|e_{eq}| \leq \frac{\rho + c}{\sigma} \tag{38}$$

Therefore, the integral of the tracking error z satisfies:

$$|\int_0^{\infty} z(\tau) d\tau| \leq \frac{\rho + c}{\sigma} \tag{39}$$

Moreover, for sufficiently large control gain σ, and more specifically when the reference signal y_{ref} is constant such that $\dot{y}ref = c = 0$, we obtain:

$$|\int_0^{\infty} z(\tau) d\tau| \to \frac{\rho}{\sigma} \tag{40}$$

This shows that the tracking error z satisfies $z \in L_1$. In addition, from the errors definition in (16), we have $\dot{z} = \dot{y}$. Since all signals are bounded, it follows then that $\dot{z}, \dot{y} \in L_\infty$. Therefore, from the facts that $z \in L_1$ and $\dot{z} \in L_\infty$, Barbalat's lemma implies that $z \to 0$ as $t \to \infty$. ■

IV. CONTROL AND SIMULATION RESULTS

The effectiveness of the proposed robust cascade control strategy is investigated through simulations using the nominal PMS motor parameters given in Table I.

TABLE I: Nominal motor parameters.

Rated power P_n	1100 [W]
Rated speed	1500 [$r.p.m$]
Nominal stator resistance R_n	2.875 [Ω]
Nominal stator inductance L_n	8 [mH]
Permanent magnet flux linkage λ_{pm}	0.175 [Wb]
Poles pair number p	2
Nominal Machine inertia J_n	0.001 [$Kg.m^2$]
Nominal Viscous friction coefficient f_n	0.00038 [Nm sec/rad]
Nominal load torque $T_{l,n}$	3 [N.m]

Applying the proposed control algorithm in Eq. (17) to the motor mechanical dynamic (8) yields the electromechanical torque reference $T_{em,ref}$ as follows:

$$\begin{aligned} \dot{e}_\Omega &= \Omega - \Omega_{ref} \\ T_{em,ref} &= J_n \ \left[-k_\Omega \ \Omega + \frac{f_n}{J_n} \ \Omega + k_\Omega \ \Omega_{ref} - \sigma_\Omega \ e_\Omega \right] \end{aligned} \tag{41}$$

Then, using (11), the quadrature-axis current reference is determined as:

$$i_{q,ref} = \frac{1}{p \ K_e} \ T_{em,ref} \tag{42}$$

Therefore, the q-axis voltage control law u_q is obtained by again applying the proposed control algorithm in Eq. (17) to the q-axis current dynamics (9), yielding:

$$\begin{aligned} \dot{e}_{i_q} &= i_q - i_{q,ref} \\ u_q &= L_n \begin{bmatrix} -k_{i_q} \ i_q + \frac{R_n}{L_n} \ i_q + p \ \Omega \ i_d \\ + \frac{p \ K_e}{L_n} \ \Omega + k_{i_q} \ i_{q,ref} - \sigma_{i_q} \ e_{i_q} \end{bmatrix} \end{aligned} \tag{43}$$

Similarly, the d-axis voltage control law u_d is obtained by applying the proposed control algorithm in Eq. (17) to the d-axis current dynamics (10), where the d-axis current reference is chosen as $i_{d,ref} = 0$, yielding:

$$\begin{aligned} \dot{e}_{i_d} &= i_d \\ u_d &= L_n \left[-k_{i_d} \ i_d + \frac{R_n}{L_n} \ i_d - p \ \Omega \ i_q - \sigma_{i_d} \ e_{i_d} \right] \end{aligned} \tag{44}$$

Furthermore, to satisfy the continuity and boundedness conditions of the angular velocity reference, Ω_{ref} is generated by filtering a square-wave signal Ω_{sq} through the following first-order unit-gain filter with time constant $\tau = 0.025$:

$$\dot{\Omega}_{ref} = -\frac{1}{\tau} (\Omega_{ref} - \Omega_{sq}) \tag{45}$$

The actual motor parameters R, L, J, and f are varied according to the following laws:

$$\begin{matrix} R = R_n & & R = R_n \ 1.75 & \\ L = L_n & \text{if } t \leq 1.5s & L = L_n \ 0.25 & \text{if } t > 1.5s \\ J = J_n & & J = J_n \ 1.75 & \\ f = f_n & & f = f_n \ 1.75 & \end{matrix} \tag{46}$$

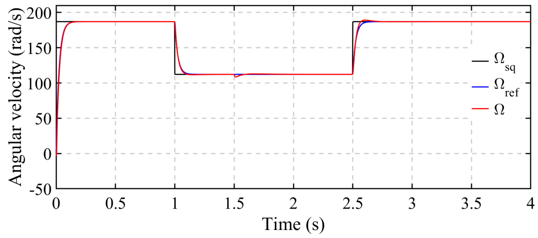
Moreover, the load torque T_l is considered as a piecewise-constant disturbance varying as follows:

$$\begin{cases} T_l = T_{l,n} & \text{if } \ t \leq 1s \\ T_l = 1.27 \ T_{l,n} & \text{if } \ 1 < \ t \leq 2.5s \\ T_l = 0.5 \ T_{l,n} & \text{if } \ t > 2.5s \end{cases} \tag{47}$$

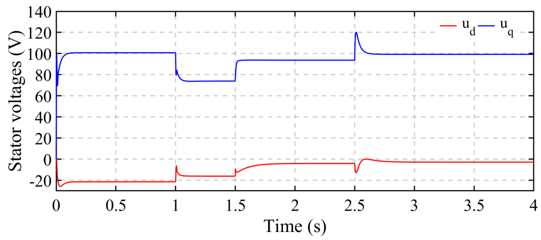
The simulation is carried out with the following control gains: $k_\Omega = 512$, $\sigma_\Omega = 6 \times 10^3$, $k_{i_q} = 510$, $\sigma_{i_q} = 5 \times 10^3$, $k_{i_d} = 515$, and $\sigma_{i_d} = 7.5 \times 10^3$. The simulation results for unknown rotor initial positions $\theta_0 = 0$, $\theta_0 = 0.7$ rad, and $\theta_0 = -0.5$ rad are shown in Fig. 2, Fig. 3, and Fig. 4, respectively. From these results, the first important observation is that the proposed robust nonlinear cascade control strategy is highly effective for controlling the PMS motor. Indeed, the angular velocity Ω tracks its reference Ω_{ref} very closely despite parameter variations and the presence of an unknown time-varying load torque. The second important observation is obtained from the comparison of stator voltages and currents for different rotor initial positions. It is clearly shown that, compared to the case $\theta_0 = 0$, the cases $\theta_0 = 0.7$ rad and $\theta_0 = -0.5$ rad lead to significantly larger values of stator currents and voltages. This indicates that large deviations from zero initial rotor position result in excessive control effort. This behaviour suggests that the PMS motor model given by Eqs. (8)–(10) becomes less effective for significantly uncertain rotor initial positions far from zero.

V. CONCLUSION

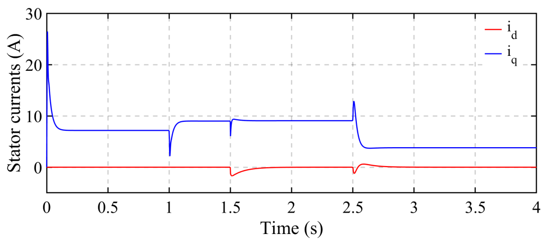
Based on the recursive gain technique, this paper has presented a robust nonlinear cascade control strategy for a PMS motor. The effectiveness of the proposed control strategy has been clearly established under varying motor parameters, unknown time-varying load torque, and small



(a) Evolution of angular velocity.

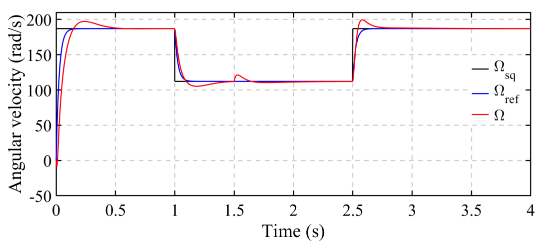


(b) Evolution of stator voltages.

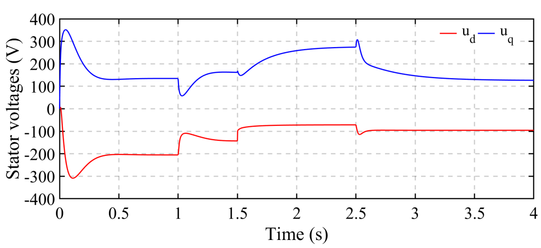


(c) Evolution of stator currents.

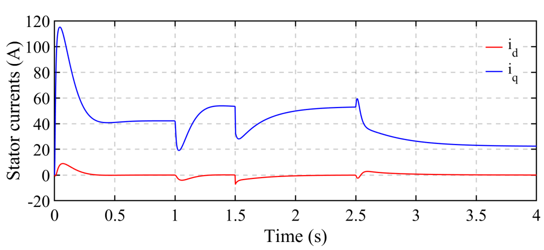
Fig. 2: Obtained results for initial position $\theta_0 = 0$ *rad*.



(a) Evolution of angular velocity.

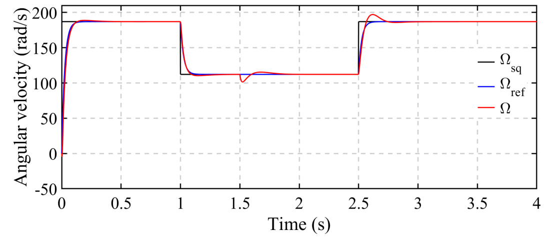


(b) Evolution of stator voltages.

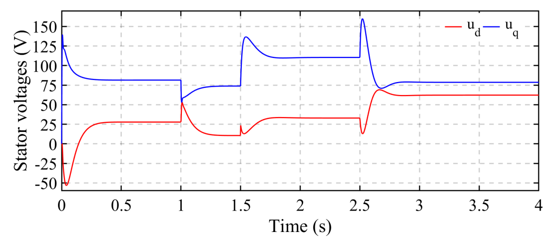


(c) Evolution of stator currents.

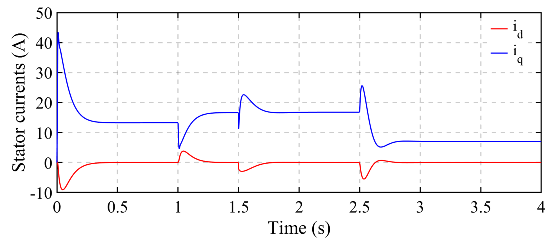
Fig. 3: Obtained results for initial position $\theta_0 = 0.7$ *rad*.



(a) Evolution of angular velocity.



(b) Evolution of stator voltages.



(c) Evolution of stator currents.

Fig. 4: Obtained results for initial position $\theta_0 = -0.5$ *rad*.

deviations in the unknown rotor initial position. Simulation results were also provided to illustrate the performance of the proposed control strategy.

Moreover, for large deviations of the unknown rotor initial position from zero, the proposed PMS motor model becomes less effective for motor control. Therefore, it would be of interest to combine the proposed robust nonlinear cascade control strategy with rotor initial position detection methods.

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