**ADAPTIVE SPEED CONTROL OF PMSMs WITH UNKNOWN LOAD TORQUE**

**F. N. Koumboulis[*], N. D. Kouvakas[*], G. E. Panagiotakis[*] and A. G. Pantelios[+]**

[*]*Department of Automation, Halkis Institute of Technology, Greece*

[+]*Department of Physics, University of Ioannina, Greece*

Abstract: The problem of speed control of a permanent magnet synchronous motor (PMSM) with unknown parameters and unknown external load torque is formulated and solved as an adaptive control problem that meets the requirements of asymptotic command following with simultaneous asymptotic disturbance attenuation. The key point is to identify the system without knowledge, estimation or measurement of the unknown disturbance. The satisfactory performance of the results is illustrated through application of the proposed nonlinear dynamic controller to a PMSM for three cases: first, no external load torque; second, smoothly starting steady external torque; third, unknown noisy and fast varying load torque. 

Keywords: permanent magnet motors, AC machines, adaptive control, model following control, disturbance localization, dynamic output feedback, nonlinear control systems, speed control, industrial control, robust performance

NOMENCLATURE

| | |
|---|---|
| $\omega_m$ | Angular velocity |
| $T_L$ | External torque |
| $u_d$ and $u_q$ | Stator voltages in direct and quadrature $d-q$ frame |
| $i_d$ and $i_q$ | Stator currents in $d-q$ frame |
| $L$ | $d-q$ phase inductances |
| $\Phi_v$ | Magnetic flux |
| $R_s$ | Phase resistance |
| $J$ | Moment of inertia |
| $B$ | Viscous load torque coefficient |
| $p$ | Number of pole pairs |

## 1. INTRODUCTION

PMSMs have attracted significant attention due to their wide range of applications (CNC machine tools, industrial robots, elevators etc) as well as their advantages (high efficiency, high torque to inertia ratio, superior power density) as compared to other types of electrical motors (Lessmeier, *et al.*, 1986). In literature, many control algorithms have been developed to satisfy various performance requirements for the variables (speed, position, torque and current) of a PMSM model. For the problem at hand, namely speed control of a PMSM, there is a long list of results (Baik, *et al.*, 1998; Cerruto, *et al.*, 1995; Chang, *et al.*, 1994; Lin and Lin, 1999; Liu and Liu, 1990; Liu and Cheng, 1994; Pillsy and Krishnan, 1990; Rahman and Hoque, 1998; Sepe and Lang, 1991; Sepe and Lang, 1992; Xu, *et al.*, 1998; Zhu, *et al.*, 2000 and the references therein). Here, we focus on adaptive techniques (Baik, *et al.*, 1998; Cherruto, *et al.*, 1995; Liu and Cheng, 1994; Rahman and Hoque, 1998; Sepe and Lang, 1991; Sepe and Lang, 1992; Xu, *et al.*, 1998). Particularly, in Cherruto, *et al.* (1995), a robust controller based on the model reference adaptive control approach is constructed to compensate the variation of the system parameters and a disturbance torque observer is employed to balance the required load torque. In Sepe and Lang (1991a) a digital adaptive controller is constructed through a linear least square estimator. This controller is of integral – proportional (IP) type while only viscous and coulombic types of load torque are considered. In Liu and Cheng (1994), three adaptive speed controller, without shaft sensor, is proposed. Self-tuning, model following and model reference adaptive controllers are applied separately for sensorless guidance of the PMSM. In Baik, *et al.* (1998), a model reference adaptive control scheme using Lyapunov stability theory is developed. Furthermore, in order to improve the robustness and performance, a boundary layer integral sliding mode controller is designed. In Rahman and Hoque (1998), speed control is achieved using an on-line self-tuned artificial neural network (ANN) that is based on the motor dynamics and the nonlinear load characteristics. The weights and biases of the ANN scheme are adjusted both off-line and on-line. In Sepe and Lang (1992), a speed control scheme based on a sensorless full-state observer is studied theoretically and experimentally. Finally, in Sepe and Lang (1991b), a fully digital adaptive speed controller is developed and the following issues where examined: discretization and global linearization of the nonlinear motor system, nonlinearities in the inverter, nonminimum phase behavior due to sampling, sampling rate restrictions, robust mechanical state estimation, persistent excitation of the mechanical states, mechanical parameter estimation and validation of these estimates and unmodeled dynamics.

In the present paper, a digital dynamic and nonlinear indirect adaptive controller is developed to achieve speed control of PMSM. In particular, for the rotor's angular velocity, asymptotic command following with simultaneous asymptotic disturbance (load torque) attenuation will be achieved while the direct phase stator current will be shown to follow exactly the respective command. The discretized PMSM model, upon which the controller is constructed, is produced using forward differences. The parameters of the nonlinear model come from RLS identification. Taking into account, the difficulty of measuring the external load torque, the respective parameter has not been included in the identification algorithm. Consequently, the influence of the torque is incorporated inside the rest parameters of the model in the sense that the identified model parameters converge far from their real values. The strong point in our approach that permits identification without knowledge of the external load is that the proposed nonlinear dynamic controller is enough robust to satisfy the design requirements even in the case where the estimated model parameters diverge from their original values. The advantages of the above proposed scheme lies on the fact that no knowledge, measurement, estimation or observer of the load torque is required to achieve attenuation of the influence of the unknown load torque to the rotor's speed. Additionally, it is important to mention that the direct phase stator current is decoupled from the external load torque.

## 2. MATHEMATICAL MODEL OF A PMSM

Permanent Magnet Synchronous Motors (PMSMs) are modeled as follows (see f.e. Leonard, 1996):

$$\frac{d\omega_m}{dt} = \frac{3p\Phi_v}{2J} i_q - \frac{B}{J}\omega_m - \frac{1}{J}T_L \tag{1a}$$

$$\frac{di_d}{dt} = -\frac{R_s}{L} i_d + p i_q \omega_m + \frac{1}{L} u_d \tag{1b}$$

$$\frac{di_q}{dt} = -\frac{R_s}{L} i_q - p i_d \omega_m - \frac{p\Phi_v}{L}\omega_m + \frac{1}{L} u_q \tag{1c}$$

Using forward differences discretization method, the discrete time model of the PMSM takes on the form

$$A(q)\begin{bmatrix}\omega_m(k)\\ i_d(k)\\ i_q(k)\end{bmatrix} = P(q)\begin{bmatrix}\omega_m(k)i_q(k)\\ \omega_m(k)i_d(k)\end{bmatrix} + B(q)\begin{bmatrix}T_L(k)\\ u_d(k)\\ u_q(k)\end{bmatrix} \tag{2}$$

where

$$A(q) = \begin{bmatrix} q + a_{1,1} & 0 & a_{1,3} \\ 0 & q + a_{2,2} & 0 \\ a_{3,1} & 0 & q + a_{3,3}\end{bmatrix}$$

$$P(q) = \begin{bmatrix} 0 & 0 \\ p_{2,1} & 0 \\ 0 & p_{3,2}\end{bmatrix}, \; B(q) = \begin{bmatrix} b_{1,1} & 0 & 0 \\ 0 & b_{2,2} & 0 \\ 0 & 0 & b_{3,3}\end{bmatrix}$$

$$a_{1,1} = BTJ^{-1} - 1, a_{1,3} = -1.5p\Phi_v TJ^{-1},$$
$$a_{2,2} = R_s L^{-1}T - 1, \; a_{3,1} = L^{-1}p\Phi_v T,$$
$$a_{3,3} = R_s L^{-1}T - 1, \; p_{2,1} = pT, \; p_{3,2} = -pT,$$
$$b_{1,1} = -J^{-1}T, \, b_{2,2} = b_{3,3} = L^{-1}T$$

and where $T$ is the sampling period and $q$ the forward shift operator.

## 3. SPEED CONTROL WITH KNOWN DATA

In the present paper, the design goal is accurate speed control of the PMSM independently from the unknown load torque. To satisfy this requirement, the design scheme of asymptotic command following with simultaneous asymptotic disturbance attenuation (ACFADA) (see Koumboulis, 1999, Koumboulis and Kouvakas, 2002) will be applied to control a PMSM with unknown load torque and known parameters.

To satisfy the requirement of ACFADA for the model (2), the following nonlinear dynamic controller solving the problem is proposed

$$u_d(k) = \frac{a_{2,2}}{b_{2,2}} i_d(k) - \frac{p_{2,1}}{b_{2,2}}\omega_m(k) i_q(k) + \frac{1}{b_{2,2}} r_d(k) \tag{3a}$$

$$u_q(k) = -\frac{p_{3,2}}{b_{3,3}}\omega_m(k) i_d(k) + \frac{a_{3,1}}{b_{3,3}}\omega_m(k) + $$
$$+\frac{a_{3,3}}{b_{3,3}} i_q(k) + \frac{1}{b_{3,3}}\left(-\frac{F(q)}{C(q)}\omega(k) + \frac{G(q)}{C(q)} r_q(k)\right) \quad (3b)$$

where $r_d(k)$ and $r_q(k)$ are external commands and

$$F(q) = \sum_{i=0}^{7} f_i q^{7-i}, \; C(q) = q^7 + \sum_{i=1}^{7} c_i q^{7-i} \quad (4a)$$
$$G(q) = \sum_{i=0}^{7} g_i q^{7-i} \quad (4b)$$

Clearly, the rational functions $F(q)/C(q)$ and $G(q)/C(q)$ are causal. To apply the controller (3) to the system (1), a Zero Order Hold D/A converter will be used to produce the continuous time input of the system. Application of the controller (3) to the system (1), results in a nonlinear hybrid (continuous time – discrete time) system. If the controller (3) is applied to the approximate discretized system (2), the following closed loop system will be derived

$$\omega_m(k) = \frac{B(q)G(q)}{A(q)C(q) + B(q)F(q)} r_q(k) + $$
$$+\frac{D(q)}{A(q)C(q) + B(q)F(q)} T_L(k) \quad (5a)$$
$$i_d(k) = q^{-1} r_d(k) \quad (5b)$$
$$i_q(k) = \frac{A(q)G(q)}{q\left[A(q)C(q) + B(q)F(q)\right]} r_q(k) - $$
$$-\frac{D(q)F(q)}{qC(q)\left[A(q)C(q) + B(q)F(q)\right]} T_L(k) \quad (5c)$$

The polynomials $F(q)$ and $C(q)$ will be used to place the closed loop system poles while the polynomial $G(q)$ will be used to set the external input to output transfer function gain to 1. Note that the controller (3) results in a closed loop system being in I/O decoupled form, i.e. for zero external torque $i_d$ is driven only by $r_d$ while $i_q$ is driven only by $r_q$. So, the angular velocity is controlled only by $r_q$ while $i_d$ is controlled only by $r_d$. Let the desired characteristic polynomial be of the form

$$A_D(q) = (q+\varepsilon)^9 \quad (6)$$

where $-1 < \varepsilon < 1$. To satisfy this requirement the coefficients of $F(q)$ have to be

$$f_0 = \left(c_2 + 9a_{1,1}\varepsilon - 36\varepsilon^2 - a_{1,1}^2\right) a_{1,3}^{-1}$$
$$f_1 = \left(a_{1,1}c_2 + c_3 - 84\varepsilon^3\right) a_{1,3}^{-1},$$
$$f_2 = \left(a_{1,1}c_3 + c_4 - 126\varepsilon^4\right) a_{1,3}^{-1},$$
$$f_3 = \left(a_{1,1}c_4 + c_5 - 126\varepsilon^5\right) a_{1,3}^{-1},$$
$$f_4 = \left(a_{1,1}c_5 + c_6 - 84\varepsilon^6\right) a_{1,3}^{-1},$$
$$f_5 = \left(a_{1,1}c_6 + c_7 - 36\varepsilon^7\right) a_{1,3}^{-1}, \; f_6 = \left(a_{1,1}c_7 - 9\varepsilon^8\right) a_{1,3}^{-1}$$
$$f_7 = -\varepsilon^9 a_{1,3}^{-1}$$

while the coefficient $c_1$ has to be

$$c_1 = 9\varepsilon - a_{1,1} \quad (7)$$

It can readily be verified that if one chooses $\varepsilon > 0$ the requirement of disturbance attenuation is satisfied. Furthermore, if the coefficients of $G(q)$ are chosen to be $g_i = -\left(8a_{1,3}\right)^{-1}(1+\varepsilon)^9, \; i = 0,\ldots,7$ unity amplitude of the closed loop transfer function that maps the external command $r_q$ to the speed $\omega_m$ is derived. Hence asymptotic command following with simultaneous disturbance attenuation is achieved.

For the closed loop system and the controller to remain stable the polynomial $C(q)$ will be chosen to be of the following special form

$$C(q) = \prod_{m=0}^{6}\left(q - \rho_c - ma\right) \quad (8)$$

where $a$, $\rho_c \in \mathbb{R}$. The controller polynomial form (8) implies that its roots are real and equally spaced at a distance $a$. Hence, the controller is stable if and only if the following conditions hold

$$-1 < \rho_c < 1, \; -1 < \rho_c + 6a < 1 \quad (9)$$

Taking into account relation (7), the pole $\rho_c$ can be chosen to be of the form

$$\rho_c = \left(a_{1,1} - 21a - 9\varepsilon\right)/7 \quad (10)$$

Thus the inequalities in (9) are reduced to

$$\left(a_{1,1} - 9\varepsilon - 7\right)/63 < a < \left(a_{1,1} - 9\varepsilon + 7\right)/63 \quad (11)$$

It is clear that the inequalities in (11) can always be satisfied while a possible choice for $a$ could be $a = \left(a_{1,1} - 9\varepsilon + 7\right)/126$ with $\varepsilon = 0.1$. Observe that since $BTJ^{-1} > 0$ then the parameter $a_{1,1}$ is grater than $-1$. Also observe that for $0 < \varepsilon < 2/3$ it holds that $a > 0$.

From the closed loop system (5), observe that, for a step external command $r_q(k)$ with amplitude $r_{q,0}$, the part of the speed depending upon the external command tends to the amplitude of the external command. Hence, the design requirement of asymptotic command following has been satisfied. Also, for a step load torque $T_L(k) = T_{L,0}$, the part of the speed depending on the load torque tends to $0.424097 b_{1,1} T_{L,0}$. Thus, the steady state value of the closed loop angular velocity is $\omega_0 = r_{q,0} + 0.424097 b_{1,1} T_{L,0}$. To illustrate the advantages of the latter result, consider the case of the PMSM in Zhu, *et al.* (2000), where $R_s = 1.2[\Omega]$, $J = 0.006[\mathrm{kgr\,m}^2]$, $\Phi_v = 0.18[\mathrm{V\,sec/rad}]$, $B = 0.0001[\mathrm{N\,m\,sec/rad}]$, $L = 0.011[\mathrm{H}]$ and $p = 3$. For this case, the poles of the closed loop system will be chosen to be $-0.1$ i.e. $\varepsilon = 0.1$. For the above data and for the case of $r_{q,0} = 80[\mathrm{rad/sec}]$ and $T = 0.001[\mathrm{sec}]$ it holds that $\omega_0 = 80 - 0.0706829 T_{L,0}$. Hence, asymptotic command following with asymptotic disturbance attenuation has practically been achieved via a 7th order controller yielding a 9th order closed loop characteristic polynomial. With regard to the behavior of the open loop system, it can easily be proven that for given direct phase stator current, let $i_{d,0}$, and rotor angular velocity, let $r_{q,0}$, and for zero external load torque, the required direct and quadrature stator voltages are equal to

$$u_d = i_{d,0} R_s - 2BL r_{q,0}^2 (3\Phi_v)^{-1} \tag{9}$$

$$u_q = \left[2BR_s + 3p^2 \Phi_v \left(L i_{d,0} + \Phi_v\right)\right] r_{q,0} (3p\Phi_v)^{-1} \tag{10}$$

Using the above motor data and for desired angular velocity $80\,\mathrm{rad/sec}$ and direct phase stator current $0.5\mathrm{A}$ it holds that $u_d = 0.573926\mathrm{V}$ and $u_q = 44.5319\mathrm{V}$. Using these voltages for different external load torques in the range of $(-1\ ,\ 1)[\mathrm{Nm}]$ the steady state of the angular velocity is presented in Figure 1 (dashed line). The respective steady state value for the closed loop system is also presented in Figure 1 (continuous line). According to Figure 1, the steady state of the angular velocity for the closed loop system is practically constant as a function of the external load torque while for the open loop system it changes dramatically.

## 4. INDIRECT ADAPTIVE CONTROL SCHEME

Consider the discrete time system (2). In general, the parameters $a_{1,1}$, $a_{1,3}$, $a_{2,2}$, $a_{3,1}$, $a_{3,3}$, $p_{2,1}$, $p_{3,2}$, $b_{1,1}$, $b_{2,2}$ and $b_{3,3}$ are not known. Taking into account the difficulty to measure the external torque, the respective coefficient $b_{1,1}$ cannot be identified. Hence, the vectors of the unknown parameters are

$$\hat{\theta}_1 = \begin{bmatrix} \hat{a}_{1,1} & \hat{a}_{1,3} \end{bmatrix}^T,\ \hat{\theta}_2 = \begin{bmatrix} \hat{a}_{2,2} & \hat{p}_{2,1} & \hat{b}_{2,2} \end{bmatrix}^T$$
$$\hat{\theta}_3 = \begin{bmatrix} \hat{a}_{3,1} & \hat{a}_{3,3} & \hat{p}_{3,2} & \hat{b}_{3,3} \end{bmatrix}^T$$

Identification data, namely the "measurements", come from the original continuous time model (1). RLS estimation of $\hat{\theta}_j\ (j = 1,2,3)$ at a particular time instant $N+1$ depends on samples of measurements of speed, rotor currents and rotor voltages through the following recursive relations (Ästrom and Wittenmark, 1989).

$$\hat{\theta}_j(N+1) = \hat{\theta}_j(N) + \\ + K_j(N)[y_j(N) - \phi_j(N+1)\hat{\theta}_j(N)]$$
$$K_j(N) = Q_j(N)\phi_j^T(N+1) \times \\ \times \left[1 + \phi_j(N+1) Q_j(N) \phi_j^T(N+1)\right]^{-1}$$
$$Q_j(N+1) = \left[I_{m_j} - K_j(N)\phi_j(N+1)\right] Q_j(N)$$

where

$$\phi_1(N) = \begin{bmatrix} -\omega_m(N-1) & -i_q(N-1) \end{bmatrix}$$
$$\phi_2(N) = \begin{bmatrix} -i_d(N-1) & \omega_m(N-1) i_q(N-1) & \cdots \\ \cdots & u_d(N-1) \end{bmatrix}$$
$$\phi_3(N) = \begin{bmatrix} -\omega_m(N-1) & -i_q(N-1) & \cdots \\ \cdots & \omega_m(N-1) i_d(N-1) & u_q(N-1) \end{bmatrix}$$
$$y_1(N) = \omega_m(N),\ y_2(N) = i_d(N),\ y_3(N) = i_q(N)$$
$$m_1 = 2,\ m_2 = 3,\ m_3 = 4$$

Initialization of the RLS identification procedure requires an a priori choice for the initial values $\hat{\theta}_j(0)$ and $Q_j(0)$. There is no limitation for $\hat{\theta}_j(0)$. To preserve convergence, $Q_j(0)$ must be positive definite. According to the above algorithm, the variations of the external load torque influence the estimation of the system parameters and consequently the choice of the controller, through measurement of the speed and the currents. To derive an indirect adaptive scheme, the controller parameters will be computed from the results of the identification algorithm described in this section. This way the controller is as in (3) with the only difference that instead of the parameters $a_{i,j}$, $b_{i,j}$ and $p_{i,l}$ the estimates $\hat{a}_{i,j}$, $\hat{b}_{i,j}$ and $\hat{p}_{i,l}$ are used.

## 5. SIMULATION RESULTS

To illustrate the advantages of the proposed design scheme, three cases will be studied for the PMSM data given in Section 3. In the first case, the performance of the controller will be tested for a PMSM with zero external load torque. In the second case, a smoothly starting torque with a steady state of $0.8\mathrm{Nm}$ will be considered, i.e. $T_L(t) = 0.8\left(1+\eta(t)\right)^{-1}$, where $\eta(t) = e^{-\frac{t-1.1}{0.12}}$. In the third case an unknown noisy and fast varying load torque (see Figure 2) will be considered. Such a torque is usually met in machining (see f.e. Koumboulis, *et al.*, 2000, and the references therein). The external commands, for all cases, are chosen to be of the form $r_q(t) = 80\left(1+\eta(t)\right)^{-1}$ and $r_d(t) = 0.5\left(1+\eta(t)\right)^{-1}$. This type of external commands provide soft starting as well as smooth speed change. In practice, the parameters of the PMSM model in (1) are not known and an identification scheme must be applied. In all three cases presented in the previous paragraph, the RLS identification algorithm presented in Section 4 has been applied. Also, in practice, even though the PMSM parameters are not known, there is an a priori estimate coming from simple experiments, geometric characteristics, manufacturers manuals, guess or arbitrary choice. The initial estimates of the PMSM parameters, let $\tilde{R}_s$, $\tilde{J}$, $\tilde{L}$, $\tilde{\Phi}_v$ and $\tilde{B}$, differ significantly from their real values, i.e. $\tilde{R}_s = 0.6R_s$, $\tilde{J} = 0.8J$, $\tilde{L} = 1.35L$, $\tilde{\Phi}_v = 1.4\Phi_v$ and $\tilde{B} = 1.3B$. Hence, using a sampling frequency of $1[\mathrm{kHz}]$ the following initial vectors of unknown parameters are derived

$$\hat{\theta}_1(0) = [-0.999973 \quad -0.23625]^T$$

$$\hat{\theta}_2(0) = [-0.951515 \quad 0.003 \quad 0.0673401]^T$$

$$\hat{\theta}_3(0) = [0.0509 \quad -0.9515 \quad -0.003 \quad 0.06734]^T$$

The initial values of $P_j(0)$ are chosen to be equal to $P_1(0) = I_2$, $P_2(0) = I_3$ and $P_3(0) = I_4$, where $I_m$ is the $m \times m$ identity matrix. For comparison reasons, the responses of the first two cases, namely the cases with nonzero external load torque, will be compared to the respective responses of the open loop system. The direct and quadrature stator voltages of the open loop system will be chosen to be $u_q(t) = 44.5319f(t)$ and $u_d(t) = 0.573926f(t)$ where $f(t) = \left(1+\eta(t)\right)^{-1}$. With regard to the response of the angular velocity of the closed loop system for the case of no external load torque (see Figure 3), it can be observed that the angular velocity follows accurately the external command while the overshoot is only $0.0002\left[\mathrm{rad/sec}\right]$. To illustrate how small is the error between the external command and the angular velocity, consider the so called relative error defined to be the $h_2$-norm of the error normalized by the $h_2$-norm of the external command. The relative error, computed from $0$ to $t_r$ (rise time of the external command), called rise relative error is equal to $0.915\%$ while the relative error from $t_r$ to $\infty$, called here steady state relative error, is equal to $0.006\%$. Similarly, for the second case namely the case of smoothly rising steady external load torque, the angular velocity follows accurately the external command (see Figure 3). The overshoot is less than $0.0001\left[\mathrm{rad/sec}\right]$. The rise relative error is $0.9153\%$ while the steady state relative error reduces to $0.0065\%$. Finally, for the case of unknown and fast varying external load torque, the angular velocity follows accurately the external command. The rise relative error is equal to $0.917\%$ while the steady state relative error is equal to $0.0785\%$ (Figure 3). In the presence of external load torque, the open loop system is totally unable to maintain the desired rotor angular velocity. With respect to the direct phase stator current, for the case of no external load torque (see Figure 4) observe that the current follows accurately the external command. There is no overshoot while the rise relative error is $0.316\%$ while steady state relative error is $0.0022\%$. Similarly, for the case of smooth starting steady external load torque, the current follows accurately the external command. Again, there is no overshoot while the rise relative error is computed to be $0.296\%$ while the steady state relative error is $0.0169\%$. Finally, for the case of unknown external load torque, the current follows accurately the external command (Figure 4). The rise relative error is $0.504\%$ while the steady state relative error reduces to $0.04\%$. In the presence of external load torque, the open loop system is unable to maintain the desired direct phase stator current (Figure 4). With respect to the quadrature phase stator current (see Figures 5 and 6), it can be observed that it lies within acceptable limits and in all cases, with or without the presence of external load torque, does not significantly vary. Similarly, the direct and quadrature phase stator voltages for all cases are within acceptable limits (see Figures 7 to 9). For all three cases, namely the case of no external torque, the case of smoothly starting external load torque and the case of unknown fast varying torque the maximum consumed electrical power is $34.86[\mathrm{W}]$, $58.37[\mathrm{W}]$ and $44.59[\mathrm{W}]$, respectively. As was expected, the identified model parameters for all cases are different than the original ones. However, this divergence does not obstruct the derivation of a satisfactory closed loop performance. This can be interpreted by the robust disturbance attenuation characteristics of the nonlinear dynamic controller proposed in Section 3. To verify this, consider Figures 10-11 where the steady responses of the speed and the direct phase stator current versus constant load torque in $[-1,1][\mathrm{Nm}]$ are presented, respectively. According to Figures 10-11, the speed and the direct phase stator current of the open loop system are dramatically affected by the load while in the adaptive closed loop cases where the controller depends upon the identified parameters they remain practically unaffected.

As was shown above, the performance of the closed loop system is satisfactory in all cases of external load torque. The controller is a time varying stable dynamic system being easily implementable. Indeed, the poles of the controller, namely the roots of $C(q)$, being functions of the identified parameters remain inside the unit disc and slightly change during time. Indicatively, see Figures 12 and 13 where the first pole $\rho_c$ and the distance $a$ are presented for the three load cases.

Before closing this section, it is important to comment on the selection of the sampling frequency. The good performance has been achieved using a rather large sampling period, $T = 0.001[\mathrm{sec}]$. It must be noted that, for all external load torque cases, extensive computational experiments have also been executed using larger sampling periods. It has been observed that in all cases the performance remains satisfactory till $T = 0.0022[\mathrm{sec}]$, i.e. till $455[Hz]$.

## 6. DISCUSSION

The problem of speed control of a PMSM with unknown parameters and unknown external load torque has been formulated as an adaptive control problem that meets the design requirement of asymptotic command following with simultaneous asymptotic disturbance attenuation. The proposed adaptive design scheme has been chosen to be indirect and based on RLS identification. The key point for disturbance attenuation was to identify the system without knowledge, estimation or measurement of the unknown disturbance as well as without using load observer. This way, the torque influence moves the model's identified parameters far from the model's parameters real values. The proposed design scheme has been tested via three computational experiments to a PMSM non-linear model. In the first, the proposed controller has been applied to the PMSM with no external load torque. In the second, a smoothly starting constant external torque has been applied to the PMSM, while in the third an unknown noisy and fast varying load torque (usually met in machining) has been applied. The responses of the last two cases including nonzero external load torque have been compared to the respective responses of the open loop system. In the cases of non-zero external load torquem, the open loop system was totally unable to maintain the desired performance. In all cases, the controller produces visually identical closed loop performances being almost independent from the external load torque. Before closing it is important to point out that the present design scheme is enough simple and computationally elegant thus offering itself for implementation to most of modern low-level controller architectures (DSPs, μCs, PLCs etc). Furthermore, the present results appear to contribute significantly to many PMSM industrial applications that demand high precision, for example to control distributed and flexible manufacturing systems, where precise synchronization and insensitivity to external disturbances appear to be indispensible.

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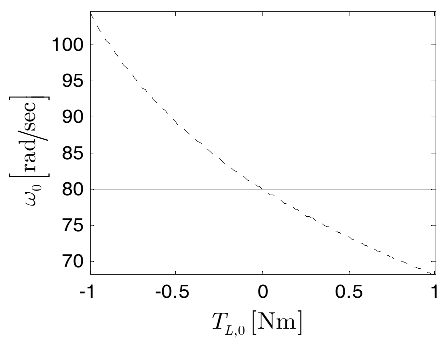


Fig. 1. Steady state of the rotor speed as a function of the external load torque (Dashed line: open loop, Solid line: closed loop)

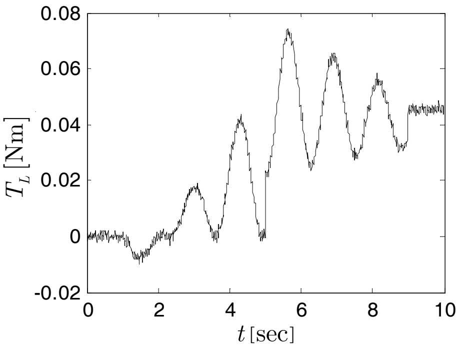


Fig. 2. Unknown and fast-varying external load torque

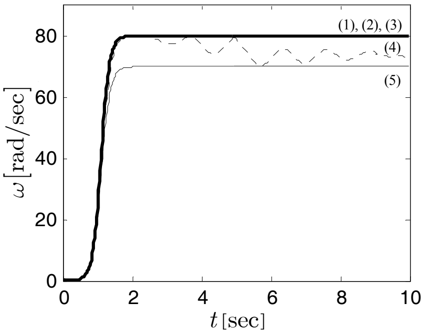


Fig. 3. Angular velocity of the rotor (1 - closed loop without load, 2 - closed loop with smoothly starting constant load, 3 - closed loop with fast varying load, 4 - open loop with fast varying load, 5 - open loop with smoothly starting constant load)

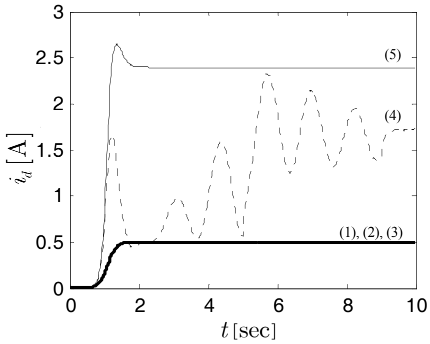


Fig. 4. Direct phase stator current (1 - closed loop without load , 2 - closed loop with smoothly starting constant load, 3 - closed loop with fast varying load, 4 - open loop with fast varying load, 5 - open loop with smoothly starting constant load)

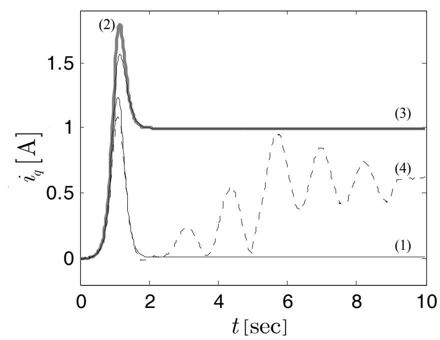


Fig. 5. Quadrature phase stator current (1 - closed loop without load, 2 - closed loop with smoothly starting constant load, 3 - open loop with smoothly starting constant load, 4 - open loop with fast varying load)

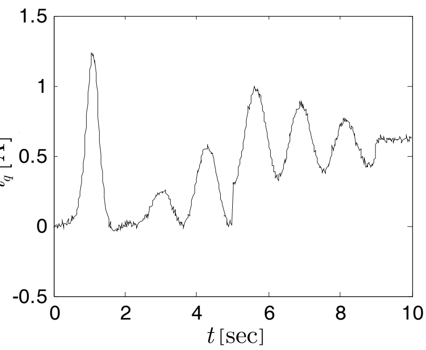


Fig. 6. Quadrature phase stator current of the closed loop system with fast varying load

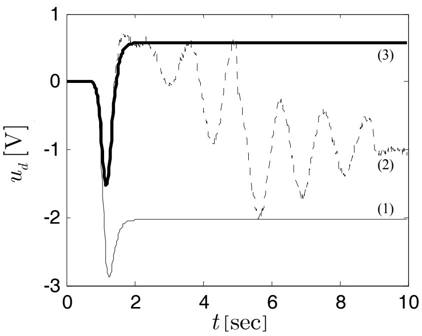


Fig. 7. Direct phase stator voltage (1 - closed loop with smooth load, 2 - closed loop with fast varying load, 3 - closed loop without load)

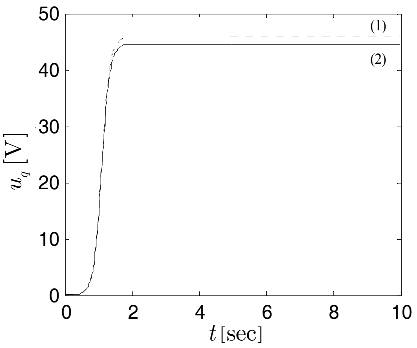


Fig. 8. Quadrature phase stator voltage (1 - closed loop with smooth load, 2 - closed loop without load)

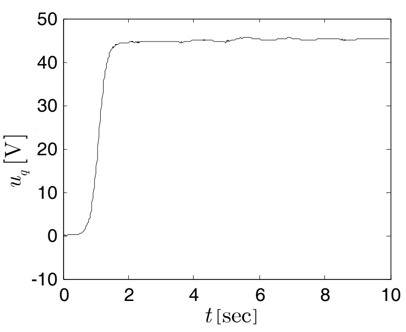


Fig. 9. Quadrature phase stator voltage of the closed loop system with fast varying load

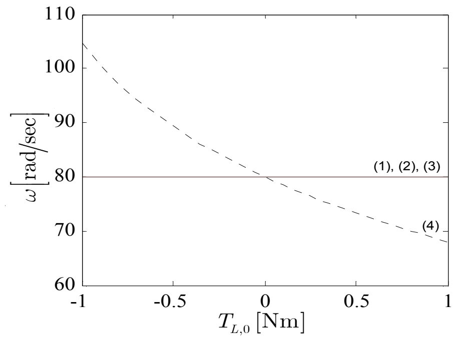


Fig. 10. Steady state angular velocity as a function of the external load torque (1 - closed loop without load, 2 - closed loop with smooth load, 3 - closed loop with fast varying load, 4 - Open loop)

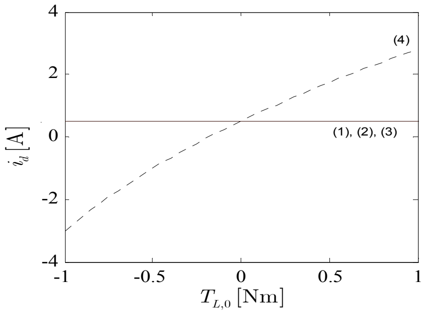


Fig. 11. Steady state direct phase stator current as a function of the external load torque (1 - closed loop without load, 2 - closed loop with smoothly starting constant load, 3 - closed loop with fast varying load, 4 - Open loop)

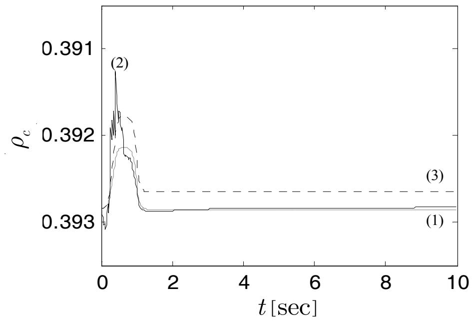


Fig. 12. Initial controller pole (1 - closed loop without load, 2 - closed loop with fast varying load, 3 - closed loop with smoothly starting constant load)

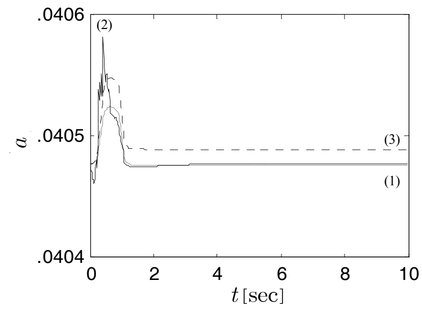


Fig. 13. Controller pole distance (1 - closed loop without load, 2 - closed loop with fast varying load, 3 - closed loop with smoothly starting constant load)