

Adaptive Nonlinear Control of an Electric Motor

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Abstract

This paper presents an adaptive nonlinear controller applied to a three-phase permanent-magnet synchronous motor. The designed controller combines the nonlinear Input-Output linearization technique with linear adaptive control techniques. It takes into account the uncertainties in the stator inductance and the rotor moment of inertia which are difficult to measure with accuracy in practice.

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1 Introduction

Nonlinear control system design has traditionally relied on linearized approximations of models treated by linear control methods. However, when the system in question is inherently nonlinear and/or operates over a large dynamical regime, linear approximations may be uninformative and designs based on them will exhibit unsatisfactory performance. This is the case in large angle, rapid slewing attitude control problems, high performance flight control systems, robot manipulator design, and certain chemical processes. The design approach used in this work is based on the geometric theory of nonlinear systems. A comprehensive geometric theory for linear systems was developed by

W.M. Wonham [16] and others. In recent years, a geometric theory of nonlinear control systems has been developed which retains much of the conceptual and technical flavor of the linear geometric theory. The work of R. Hermann [18] and C. Lobry [17] mainly used the language of differential geometry. From this, the interaction between mathematicians and engineers working in nonlinear control theory produced numerous results, first on qualitative concepts such as controllability, observability and realizability. Subsequent design methods of significant power were developed for control problems such as disturbance decoupling, input-output decoupling and feedback linearization. The technique of input-output feedback linearization is known by now. It gives a good solution for tracking control problems. Several accessible references ([1], [2],[4] and [11]) which describe its constructions are available. This nonlinear control technique has been successfully applied to the control of AC motors that are known to have highly nonlinear models and are, therefore, difficult to control (synchronous motors [7],[13], [14], [15], induction motors [9], and switched reluctance motors [10]). It has been shown that the nonlinear controller can handle the nonlinearities of the motor' model and gives good results as long as the PMSM parameters are known with high accuracy and remain constant. When these parameters are not well known, adaptive techniques are considered. An important line of research that appeared in the last decades is the development of an adaptive nonlinear methodology which combines feedback linearization techniques with linear adaptive control techniques [4], [11] and has been applied to several types of motors with uncertainty on the parameters ([9], [10], [11], [12], and [14]). The nonlinear adaptive technique used in this paper follows the results given in [3] and [11]. It shows some results in the case when the parameters uncertainty concerns mainly the stator inductance L and the rotor moment of inertia J which are, in fact, difficult to measure exactly.

2 Design of the adaptive nonlinear controller

In this paper we develop an adaptive nonlinear controller which combines feedback linearization with adaptive control. Several results on this adaptive nonlinear control have also appeared. The nonlinear systems treated are subject to uncertainties parameterized by fixed but unknown parameters. Our objective is to design an adaptive nonlinear controller for use with a PMSM having constant but unknown L and J . We start by designing a nonlinear input-output linearizing controller for the nominal model of the PMSM. We can rewrite the system of equations in a form that suggests an adaptive scheme for the estimation of L and J :

$$x' = f_0(x) + \sum_{i=1}^2 g_i(x)u_i + \sum_{j=1}^2 \delta_j f_{\delta_j}(x) + \sum_{k=1}^2 \delta_k \aleph_{\delta_k}(x) \quad (1)$$

where:

$$f_0(x) = \begin{bmatrix} -\frac{R}{L_n}x_1 + px_2x_3 \\ -\frac{R}{L_n}x_2 - px_1x_3 - \frac{p\Phi_v}{L_n}x_3 \\ \frac{1}{J}(k_Tx_2 - Bx_3 - T_L) \end{bmatrix} = \begin{bmatrix} f_1(x) \\ f_2(x) \\ f_3(x) \end{bmatrix} \quad (2)$$

with:

$$\begin{aligned} g_1(x) &= \begin{bmatrix} \frac{1}{L_n} & 0 & 0 \end{bmatrix}^T \\ g_2(x) &= \begin{bmatrix} 0 & \frac{1}{L_n} & 0 \end{bmatrix}^T \\ f_{\delta_1}(x) &= \begin{bmatrix} -Rx_1 & -(Rx_2 + px_3\Phi_v) & 0 \end{bmatrix}^T \\ f_{\delta_2}(x) &= \begin{bmatrix} 0 & 0 & k_Tx_2 - Bx_3 - T_L \end{bmatrix}^T \\ \aleph_1(x) &= \begin{bmatrix} 1 & 0 & 0 \end{bmatrix}^T \\ \aleph_2(x) &= \begin{bmatrix} 0 & 1 & 0 \end{bmatrix}^T \end{aligned} \quad (3)$$

where $(x = [x_1 \ x_2 \ x_3]^T)$ is the state vector and $(x_1 = i_d)$, $(x_2 = i_q)$ and $(x_3 = \omega_r)$ are the direct and quadrature currents and the rotor speed. ϕ_v is the rotor flux, R and L are the stator resistance and inductance, T_L is the load torque, u_d and u_q are the stator voltages. Finally, B is the damping coefficient and J is the rotor moment of inertia. The vector δ represents the error between the inverse of the uncertain parameters (L and J) and their nominal values, i.e:

$$\delta = \begin{bmatrix} \delta_1 & \delta_2 \end{bmatrix}^T = \begin{bmatrix} \frac{1}{L} - \frac{1}{L_n} & \frac{1}{J} - \frac{1}{J_n} \end{bmatrix}^T \quad (4)$$

where the index n denotes nominal values.

2.1. Non-adaptive version of the controller

We start by designing the non-adaptive version of the controller, where δ is assumed to be zero. The control goal is twofold, first to regulate the rotor

speed ($x_3 = \omega_r$) and second to force the d-component of the stator currents ($x_1 = i_d$) to be zero to insure a maximum torque operation, hence the output vector is:

$$\begin{bmatrix} y_1 \\ y_2 \\ y_3 \end{bmatrix} = \begin{bmatrix} \phi_1 \\ \phi_2 \\ \phi_3 \end{bmatrix} = \begin{bmatrix} i_d \\ \omega_r \\ \dot{\omega}_r \end{bmatrix} \quad (5)$$

then:

$$\begin{bmatrix} \dot{y}_1 \\ \dot{y}_2 \\ \dot{y}_3 \end{bmatrix} = \begin{bmatrix} L_{f_0}\phi_1 + L_{g_1}\phi_1 u_d + L_{g_2}\phi_1 u_q \\ L_{f_0}\phi_2 + L_{g_1}\phi_2 u_d + L_{g_2}\phi_2 u_q \\ L_{f_0}^2\phi_2 + L_{g_1}L_{f_0}\phi_2 u_d + L_{g_2}L_{f_0}\phi_2 u_q \end{bmatrix} \quad (6)$$

where:

$$\begin{aligned} L_{f_0}\phi_1 &= \frac{\partial \phi_1}{\partial x} f_0(x) = \begin{bmatrix} 1 & 0 & 0 \end{bmatrix} f_0(x) = f_1(x) \\ L_{g_1}\phi_1 &= \begin{bmatrix} 1 & 0 & 0 \end{bmatrix} g_1(x) = \frac{1}{L_n} \\ L_{g_2}\phi_1 &= L_{g_1}\phi_2 = L_{g_2}\phi_2 = L_{g_1}L_{f_0}\phi_2 = 0 \\ L_{f_0}\phi_2 &= f_3(x) \\ L_{f_0}^2\phi_2 &= L_{f_0}(L_{f_0}\phi_2) = \frac{k_T}{J_n} f_2(x) - \frac{B}{J_n} f_3(x) \\ L_{g_2}L_{f_0}\phi_2 &= \frac{k_T}{L_n J_n} \end{aligned} \quad (7)$$

Recall that the Lie derivative of ϕ along the vector field f is defined by:

$$\frac{\partial \phi}{\partial x} f(x) \quad (8)$$

The feedback linearization technique which uses a nonlinear change of coordinates and feedback to transform the nonlinear system (1) into a decoupled linear one is given by:

$$[z] = [z_1 \ z_2 \ z_3]^T = [i_d \ \omega_r \ \dot{\omega}_r]^T \quad (9)$$

Then:

$$[\dot{z}_1 \ \ddot{z}_3]^T = \xi(x) + D(x) \begin{bmatrix} u_d \\ u_q \end{bmatrix} \quad (10)$$

where

$$\xi(x) = \begin{bmatrix} -\frac{R}{L_n}x_1 + px_2x_3 \\ \frac{k_T}{J_n}f_2(x) - \frac{B}{J_n}f_3(x) \end{bmatrix} \quad (11)$$

$D(x)$ is an invertible matrix given by:

$$D(x) = \begin{bmatrix} \frac{1}{L_n} & 0 \\ 0 & \frac{k_T}{J_nL_n} \end{bmatrix} \quad (12)$$

with its determinant satisfying the condition:

$$\det(D(x)) = \frac{k_T}{J_nL_n^2} \neq 0 \quad (13)$$

for all operation points. The linearizing control law is then:

$$u = \begin{bmatrix} u_d \\ u_q \end{bmatrix} = D(x)^{-1}(-\xi(x) + \bar{v}) \quad (14)$$

where \bar{v} is the new system input vector to be determined. Classical pole placement scheme can be used to insure speed tracking:

$$\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} = \begin{bmatrix} -k_1(x_1 - x_{1ref}) \\ \ddot{x}_{3ref} - k_2(x_3 - x_{3ref}) - k_3(\dot{x}_3 - \dot{x}_{3ref}) \end{bmatrix} \quad (15)$$

The matrix gain k is determined by choosing some poles which give a good behavior to the motor.

2.2. Adaptive version of the controller

The dynamics of the outputs becomes in this case:

$$\begin{cases} \dot{y}_1 = L_{f_0}\phi_1 + L_{g_1}\phi_1u_d + L_{g_2}\phi_1u_q + \delta_1L_{f_{\delta_1}}\phi_1 + \delta_2L_{f_{\delta_2}}\phi_1 + \delta_1L_{\aleph_1}\phi_1u_d + \delta_1L_{\aleph_2}\phi_1u_q \\ \dot{y}_2 = L_{f_0}\phi_2 + L_{g_1}\phi_2u_d + L_{g_2}\phi_2u_q + \delta_1L_{f_{\delta_1}}\phi_2 + \delta_2L_{f_{\delta_2}}\phi_2 + \delta_1L_{\aleph_1}\phi_2u_d + \delta_1L_{\aleph_2}\phi_2u_q \\ \dot{y}_3 = L_{f_0}^2\phi_2 + L_{g_1}L_{f_0}\phi_2u_d + L_{g_2}L_{f_0}\phi_2u_q + \delta_1L_{f_{\delta_1}}L_{f_0}\phi_2 + \delta_2L_{f_{\delta_2}}L_{f_0}\phi_2 + \\ + \delta_1L_{\aleph_1}L_{f_0}\phi_2u_d + \delta_1L_{\aleph_2}L_{f_0}\phi_2u_q \end{cases} \quad (16)$$

with:

$$\begin{aligned}
 L_{f_{\delta_2}}\phi_1 &= L_{g_2}\phi_1 = L_{N_2}\phi_1 = L_{f_{\delta_1}}\phi_2 = L_{g_1}\phi_2 = 0 \\
 L_{g_2}\phi_2 &= L_{N_1}\phi_2 = L_{N_2}\phi_2 = L_{g_1}L_{f_0}\phi_2 = L_{N_1}L_{f_0}\phi_2 = 0 \\
 L_{f_{\delta_1}}L_{f_0}\phi_2 &= -\frac{k_T}{J_B}(Ri_q + p\omega_r\Phi_v) \\
 L_{f_{\delta_2}}L_{f_0}\phi_2 &= -\frac{B}{J_n}(k_Ti_q - B\omega_r - T_L) \\
 L_{g_2}L_{f_0}\phi_2 &= \frac{k_T}{J_nL_n} \\
 L_{N_2}L_{f_0}\phi_2 &= \frac{k_T}{J_n}
 \end{aligned} \tag{17}$$

then:

$$\begin{cases} \dot{y}_1 = L_{f_0}\phi_1 + L_{g_1}\phi_1u_d + \delta_1L_{f_{\delta_1}}\phi_1 + \delta_1L_{N_1}\phi_1u_d \\ \dot{y}_2 = L_{f_0}\phi_2 + \delta_2L_{f_{\delta_2}}\phi_2 \\ \dot{y}_3 = L_{f_0}^2\phi_2 + L_{g_2}L_{f_0}\phi_2u_q + \delta_1L_{f_{\delta_1}}L_{f_0}\phi_2 + \delta_2L_{f_{\delta_2}}L_{f_0}\phi_2 + \delta_1L_{N_2}L_{f_0}\phi_2u_q \end{cases} \tag{18}$$

Since δ is unknown, we replace it by its estimate $\hat{\delta}$:

$$\hat{\delta} = [\hat{\delta}_1 \quad \hat{\delta}_2]^T = [\frac{1}{L} - \frac{1}{L_n} \quad \frac{1}{J} - \frac{1}{J_n}]^T \tag{19}$$

The nonlinear change of coordinates (9) becomes:

$$\begin{cases} \hat{z}_1 = y_1 \\ \hat{z}_2 = y_2 \\ \hat{z}_3 = y_3 + \hat{\delta}_2L_{f_{\delta_2}}\phi_2 \end{cases} \tag{20}$$

satisfying:

$$\begin{cases} \dot{\hat{z}}_1 = \dot{y}_1 \\ \dot{\hat{z}}_2 = \dot{y}_2 \\ \dot{\hat{z}}_3 = \dot{y}_3 + \hat{\delta}_2L_{f_{\delta_2}}\phi_2 + \dot{\hat{\delta}}_2(L_{f_{\delta_2}}\phi_2)' \end{cases} \tag{21}$$

By replacing z in (9) and u in (14) by \hat{z} and \hat{u} which depend on the parameter estimate vector (19). The nonlinear change of coordinates (20) gives:

$$\begin{bmatrix} \dot{\hat{z}}_1 \\ \dot{\hat{z}}_3 \end{bmatrix} = \xi(x) + \xi_\delta(x, \delta, \hat{\delta}) + D(x, \delta, \hat{\delta}) \begin{bmatrix} \hat{u}_d \\ \hat{u}_q \end{bmatrix} \tag{22}$$

Then, the linearizing control law becomes:

$$\begin{bmatrix} \hat{u}_d \\ \hat{u}_q \end{bmatrix} = D^{-1}(x, \delta, \hat{\delta}) \left\{ -\xi(x) - \xi_\delta(x, \delta, \hat{\delta}) + \hat{v} \right\} \quad (23)$$

where $\xi(x)$ is given by (11) and the linearizing part introduced by parameter uncertainties $\xi_\delta(x, \delta, \hat{\delta})$ is:

$$\begin{aligned} \xi_\delta(x, \delta, \hat{\delta}) &= \begin{bmatrix} \xi_{\delta_1}(x, \delta, \hat{\delta}) \\ \xi_{\delta_2}(x, \delta, \hat{\delta}) \end{bmatrix} \\ \xi_{\delta_1}(x, \delta, \hat{\delta}) &= -\hat{\delta}_1 R x_1 \\ \xi_{\delta_2}(x, \delta, \hat{\delta}) &= -\hat{\delta}_2 \frac{B}{J_n} \Xi(x) - \hat{\delta}_1 \left(\frac{k_T}{J_n} \right) \{ R x_2 + p x_3 \Phi_v \} \\ &\quad + \hat{\delta}_2 \Xi(x) + \hat{\delta}_2 \{ k_T f_2(x) - B f_3(x) \} \end{aligned} \quad (24)$$

where $f_2(x)$ and $f_3(x)$ are given by (2) and $\Xi(x) = (k_T x_2 - B x_3 - T_L)$. Here, $D^{-1}(x, \delta, \hat{\delta})$ is given by:

$$D(x, \hat{\delta}) = \begin{bmatrix} \frac{1}{L_n} + \hat{\delta}_1 & 0 \\ 0 & k_T \left(\frac{1}{J_n L_n} + \frac{\hat{\delta}_1}{JL} \right) \end{bmatrix} \quad (25)$$

with its determinant is non zero if $(\hat{\delta}_1 \neq -\frac{JL}{J_n L_n})$.

The new system inputs:

$$\hat{v} = \begin{bmatrix} \hat{z}_1 & \hat{z}_3 \end{bmatrix}^T = \begin{bmatrix} \hat{v}_1 & \hat{v}_2 \end{bmatrix}^T \quad (26)$$

are designed exactly as (15) where $z_3 = \dot{\omega}_r$ is replaced by \hat{z}_3 . In closed loop, the system (22) becomes:

$$\begin{aligned} \dot{\hat{z}} &= \xi(x) + \xi_\delta(x, \delta, \hat{\delta}) - D(x, \delta) D(x, \delta, \hat{\delta})^{-1} \left\{ \xi(x) + \xi_\delta(x, \delta, \hat{\delta}) \right\} \\ &\quad + D(x, \delta) D(x, \delta, \hat{\delta})^{-1} \hat{v} \end{aligned} \quad (27)$$

or in a compact form:

$$\dot{\hat{z}} = Az + W_1 \Delta \delta + W_2 \dot{\hat{\delta}} \quad (28)$$

2.3. Model Reference Adaptive Scheme

A standard adaptive law can be designed to insure that the augmented error approaches zero. In the design of the model reference adaptive controller, the reference model must be asymptotically stable and with relative degree larger than or equal to that of the dynamic systems (Electric motor).

We consider the following linear (linearized) model:

$$\dot{z}_{ref} = A_r z_r + b_r r \quad z_r(0) = z_{r0} \quad (29)$$

The signal $r(t)$ is a uniformly bounded reference input and z_r is a prescribed reference trajectory we wish the state z to track. The spectrum of A_r lies in the negative left half plane and the pair (A_r, b_r) is assumed to be in controllable form. We define the vector of the states error by:

$$e = \hat{z} - z_r = [\hat{z}_1 - z_{1r} \quad \hat{z}_3 - z_{3r} \quad \hat{z}_3 - \dot{z}_{3r}]^T \quad (30)$$

The design objective is to find an adaptation law independent of the initial conditions, which assures asymptotic adaptation characterized by $\lim_{x \rightarrow \infty} e(t) = 0$ and a bounded parameter error ($\Delta \delta$). The error signal e satisfies:

$$\dot{e} = \dot{\hat{z}} - \dot{z}_r = Az + bv - A_r z_r + W_1 \Delta \delta + W_2 \dot{\hat{\delta}} \quad (31)$$

where $\Delta \delta$ is the vector of the error estimates, given by:

$$\Delta \hat{\delta} = [\Delta \delta_1 \quad \Delta \delta_2]^T = \begin{bmatrix} \frac{1}{L} - \frac{1}{\hat{L}} \\ \frac{1}{J} - \frac{1}{\hat{J}} \end{bmatrix} \quad (32)$$

and:

$$\begin{aligned}
 W_1 &= \begin{bmatrix} \hat{L}(\hat{v}_1 - px_3x_2) & 0 \\ 0 & \Xi(x) \\ \Upsilon(x) & \frac{B}{J_n}\Xi(x) \end{bmatrix} \\
 W_2 &= \begin{bmatrix} 0 & 0 \\ 0 & 0 \\ 0 & \Xi(x) \{1 + \delta\hat{L}\} \end{bmatrix}
 \end{aligned} \tag{33}$$

with:

$$\begin{aligned}
 \Upsilon(x) &= \Upsilon_1(x) + \Upsilon_2(x) + \Upsilon_3(x) + \Upsilon_4(x) + \hat{L}\hat{v}_2 \\
 \Upsilon_1(x) &= -\hat{L} \left\{ \left(\frac{k_T}{J_n}\right)f_2(x) - \left(\frac{B}{J_n}\right)f_3(x) \right\} \\
 \Upsilon_2(x) &= -\left(\frac{k_T}{J_n}\right)\left(\frac{\hat{L}(Rx_2 + px_3\Phi_v)}{L_n}\right) \\
 \Upsilon_3(x) &= \frac{\hat{\delta}_2\Xi(x)\hat{L}B}{J_n} \\
 \Upsilon_4(x) &= -\{k_T f_2(x) - B f_3(x)\} \hat{\delta}_2 \hat{L}
 \end{aligned} \tag{34}$$

The form (31) without the term W_2 is familiar in the linear adaptive control literature [19], [20]. The quantity multiplying the parameter error $(\Delta\delta, W_1)$ is often referred to as the regressor. The existence of the nonlinear term W_2 suggests the use of the concept of augmented error, defined in linear adaptive control [20,21, 22,23], to put the equation (31) into a more familiar form. A standard adaptive law can be designed to insure that the augmented error approaches zero.

We introduce the signal ε satisfying

$$\dot{\varepsilon} = k\varepsilon + W_2\dot{\delta} \tag{35}$$

and the augmented error η by $(\eta = e - \varepsilon)$ which satisfies

$$\dot{\eta} = k\eta + W_1\Delta\delta \tag{36}$$

We choose the candidate Lyapunov function V as:

$$V = \eta^T P\eta + (\Delta\delta)^T \Gamma(\Delta\delta) \tag{37}$$

Differentiating V we have:

$$\dot{V} = \eta^T(k^T P + Pk)\eta + 2(\Delta\delta)^T \left\{ W_1^T P\eta + \Gamma(\Delta\dot{\delta}) \right\} \quad (38)$$

In defining the parameter update law under the form:

$$\dot{\hat{\delta}} = -\Gamma^{-1} W_1^T P\eta \quad (39)$$

we can guarantee that $(\dot{V} = -\eta^T Q\eta \leq 0)$, where $(P = P^T > 0)$ is the solution of Lyapunov equation:

$$k^T P + Pk = -Q \quad (40)$$

where Q is a positive definite matrix and Γ represents a positive-definite adaptation gains matrix.

Finally, by using Theorem 4.1 ([19]), one can proof that:

$$\lim_{t \rightarrow \infty} \|e(t)\| = \lim_{t \rightarrow \infty} \|\hat{z}(t) - z_r(t)\| = 0 \quad (41)$$

3 Conclusion

This work concerns the construction of an adaptive nonlinear controller mainly for some nonlinearities cancelation of an electric motor. The method used considers parametric uncertainties and combine exact linearization techniques with adaptive linear techniques. This method assumes that the unknown parameters appear linearly in the nonlinear model and that the nonlinear system satisfies the linearizability conditions for all possible values of the unknown parameters.

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