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NONLINEAR ADAPTIVE BACKSTEPPING CONTROL OF PERMANENT MAGNET SYNCHRONOUS MOTOR (PMSM)

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ABSTRACT

In this paper, a nonlinear adaptive speed controller for a permanent magnet synchronous motor (PMSM) based on a newly developed adaptive backstepping approach is presented. The exact, input-output feedback linearization control law is first introduced without any uncertainties in the system. However, in real applications, the parameter uncertainties such as the stator resistance and the rotor flux linquage and load torque disturbance have to be considered. In this case, the exact input-output feedback linearization approach is not very effective, because it is based on the exact cancellation of the nonlinearity. To compensate the uncertainties and the load torque disturbance, the input-output feedback linearization approach is first used to compensate the nonlinearities in the nominal system. Then, nonlinear adaptive backstepping control law and parameter uncertainties, and load torque disturbance adaptation laws, are derived systematically by using adaptive backstepping technique. The simulation results clearly show that the proposed adaptive scheme can track the speed reference s the signal generated by a reference model, successfully, and that scheme is robust to the parameter uncertainties and load torque disturbance.

Keywords: ***Feedback input-output linearization – Adaptive Backstepping control***, ***Permanent Magnet Synchronous Motor (PMSM)***

1. INTRODUCTION

Permanent magnet synchronous motors (PMSM) are widely used in high performance servo applications due to their high efficiency, high power density, and large torque to inertia ratio [1], [2]. However, PMSMs are nonlinear multivariable dynamic systems and, without speed sensors and under load and parameter perturbations, it is difficult to control their speed with high precision, using conventional control strategies.

Linearization and/or high-frequency switching based nonlinear speed control techniques, such as feedback linearization control and sliding mode control, have been implemented for the PMSM drives [13]. However, it is more efficient to use a nonlinear control method that is based on minimizing a cost function and allows one to tradeoff between the control accuracy and control effort.

The adaptive backstepping design offers a choice of design tools for accommodation of uncertainties an nonlinearities. And can avoid wasteful cancellations. In addition, the adaptive backstepping approach [11][16] is capable of keeping almost all the robustness properties of the mismatched uncertainties. The adaptive backstepping is a systematic and recursive design methodology for nonlinear feedback control. The basic idea of backstepping design is to select recursively some appropriate functions of state variables as pseudo control inputs for lower dimension subsystems of overall system. Each backstepping stage results in a new pseudo control design, expressed in terms of the pseudo control designs from preceding design stages. When the procedure terminates, a feedback design for the true control input results in achieving of a final Lyapunov function by an efficient original design objective. The latter is formed by summing up the Lyapunov functions associates with each individual design stage [3][6][9].

In this paper, the backstepping approach is used to design state feedback non linear control, first under an assumption of known electrical parameters, and then with the possibility of

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parametric uncertainly and bounded disturbances included. The steady state performances of the backstepping based controller and the problem of the disturbance rejection are enhanced via the introduction of an integral action in the controller.

2. MODEL OF THE PMSM

The dynamic model of a typical surface-mounted PMSM can be described in the well known (d-q) frame through the park transformation as follows [5][7][11]:

$$
\begin{aligned}
\frac{di_d}{dt} &= -\frac{R_s}{L_d} i_d + \frac{L_q}{L_d} p\Omega i_q + \frac{v_d}{L_d} \\
\frac{di_q}{dt} &= -\frac{R_s}{L_q} i_d - \frac{L_d}{L_q} p\Omega i_d - \frac{\Phi_f}{L_q} p\Omega + \frac{v_d}{L_d} \\
\frac{d\Omega}{dt} &= \frac{3p}{2J}(\Phi_f i_q + (L_d - L_q) i_d i_q) - \frac{f}{J}\Omega + \frac{T_L}{J}
\end{aligned} \quad (1)
$$

Where

v_d, v_q	Direct-and quadrature-axis stator voltages
i_d, i_q	Direct-and quadrature-axis stator currents
L_d, L_q	Direct -and quadrature-axis inductance
P	Number of poles
R_s	Stator resistance
Φ_f	Rrotor magnet flux linkage
Ω	Eelectrical rotor speed
ω	Mechanical rotor speed ($\omega = p.\Omega$)
f	Viscous friction coefficient
T_L	Load torque
J	Moment of Inertia

We assume :

$$a_1 = -\frac{R_s}{L_d} \; ; a_2 = p\frac{L_q}{L_d} ; \quad a_3 = \frac{1}{L_d}$$

$$b_1 = -\frac{R_s}{L_q} \; ; b_2 = -p\frac{L_d}{L_q} \; ; b_3 = -p\frac{\Phi_f}{L_q}$$

$$b_4 = \frac{1}{L_q} \; ; c_1 = -\frac{f}{J} \; ; c_2 = \frac{3p}{2J}(L_d - L_q)$$

$$c_3 = \frac{3p}{2J}.\Phi_f \quad \text{et} \quad c_4 = -\frac{1}{J}$$

The (1) equation become:

$$
\begin{aligned}
\frac{di_d}{dt} &= a_1 i_d + a_2 i_q \Omega + a_3 v_d \\
\frac{di_q}{dt} &= b_1 i_q + b_2 i_d \Omega + b_3 \Omega + b_4 v_q \\
\frac{d\Omega}{dt} &= c_1 \Omega + c_2 i_d i_q + c_3 i_q + c_4 T_L
\end{aligned} \quad (2)
$$

3. PMSM NONLINEAR STATE REPRESENTATION [11],[12]

The vector $[X] = \begin{bmatrix} i_d \\ i_q \\ \Omega \end{bmatrix}$ choice as state vector is justified by the fact that currents and speed are measurable and that the control of the instantaneous torque can be done comfortable via the currents i_{sd} and/or i_{sq}. Such representation of the nonlinear state corresponds to:

$$
\begin{aligned}
\frac{d[X]}{dt} &= F[X] + [G].[U] \\
[y] &= H[X]
\end{aligned} \quad (3)
$$

With

$$[X] = \begin{bmatrix} x_1 \\ x_2 \\ x_3 \end{bmatrix} = \begin{bmatrix} i_d \\ i_q \\ \Omega \end{bmatrix} \text{ Three order state vector} \quad (4)$$

$$[U] = \begin{bmatrix} v_d \\ v_q \end{bmatrix} \quad \text{Control vector} \quad (5)$$

$$[y] = \begin{bmatrix} y_1 \\ y_2 \end{bmatrix} = \begin{bmatrix} i_d \\ \Omega \end{bmatrix} \quad \text{Output vector}$$

$$F[X] = \begin{bmatrix} f_1(x) \\ f_2(x) \\ f_3(3) \end{bmatrix} = \begin{bmatrix} a_1.x_1 + a_2.x_2.x_3 \\ b_1.x_2 + b_2.x_1.x_3 + b_3.x_3 \\ c_1.x_3 + c_2.x_1.x_2 + c_3.x_2 + c_4.C_r \end{bmatrix} \quad (6)$$

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$$\begin{bmatrix} f_1(x) \\ f_2(x) \\ f_3(x) \end{bmatrix} = \begin{bmatrix} a_1.i_d + a_2.i_q.\Omega \\ b_1.i_q + b_2.i_d.\Omega + b_3.\Omega \\ c_1.\Omega + c_2.i_d.i_q + c_3.i_q + c_4.C_r \end{bmatrix} \quad (7)$$

Nonlinear functions of our model:

$$[G] = \begin{bmatrix} g_{11} & g_{12} \\ g_{21} & g_{22} \\ g_{31} & g_{32} \end{bmatrix} = \begin{bmatrix} a_3 & 0 \\ 0 & b_4 \\ 0 & 0 \end{bmatrix} \text{ Control matrix} \quad (8)$$

$$H[X] = \begin{bmatrix} h_1(x) \\ h_2(x) \end{bmatrix} = \begin{bmatrix} x_1 \\ x_3 \end{bmatrix} = \begin{bmatrix} i_d \\ \Omega \end{bmatrix} \text{ Output vector} \quad (9)$$

The model set up in (2) can be represented by the following functional diagram (Figure1)

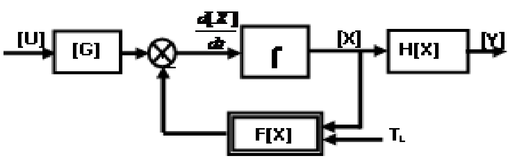

Figure 1: Nonlinear diagram block of PMSM in d-q frame.

4. INPUT-OUTPUT FEEDBACK LINEARIZATION

The input-output feedback linearization control for a PMSM is introduced first. The control objective is to make the mechanical speed follow the desired speed asymptotically.

In order to make the actual speed follow The desired one and avoid the zero dynamic, w and isd are chosen as the control outputs. Then the new variables are defined as follows:

$$\begin{aligned} y_1 &= i_d \\ y_2 &= \Omega \end{aligned} \quad (10)$$

Then the new state equations in the new state coordinates can be written as follows:

$$\begin{aligned} \dot{y}_1 &= \dot{i}_d = f_1(x) + a_3.v_d \\ \dot{y}_2 &= \dot{\Omega} = f_3(x) \\ \ddot{y}_2 &= \ddot{\Omega} = c_1 f_3(x) + c_2 i_q (f_1(x) + a_3 v_d) + \\ &\quad (c_2 i_d + c_3)(f_2(x) + a_4.v_q) \end{aligned} \quad (11)$$

To linearize the equations, define the new control variable as follows:

$$\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} = \begin{bmatrix} \dot{y}_1 \\ \ddot{y}_2 \end{bmatrix} = [\xi(x)] + [D(x)].\begin{bmatrix} v_d \\ v_q \end{bmatrix} \quad (12)$$

Where

$$[\xi(x)] = .\begin{bmatrix} f_1(x) \\ c_2 x_2 f_1(x) + (c_2 x_1 + c_3) f_2(x) + c_1 f_3(x) \end{bmatrix}$$

$$[D(x)] = .\begin{bmatrix} a_3 & 0 \\ a_3 c_2 x_2 & b_4 (c_2 x_1 + c_3) \end{bmatrix}$$

Now, the standard pole-assignment technique can be adopted to design the state feedback control output as follows:

$$\begin{aligned} v_1 &= k_i(i_{dref} - i_d) + \frac{d}{dt} i_{dref} \\ v_2 &= \frac{d^2}{dt^2}\Omega_{ref} + k_{\Omega 1}.\frac{d}{dt}(\Omega_{ref} - \Omega) + k_{\Omega 2}.(\Omega_{ref} - \Omega) \end{aligned} \quad (13)$$

Where the i_{dref} and Ω_{ref} are the commands for the speed and the d-axis current, respectively.

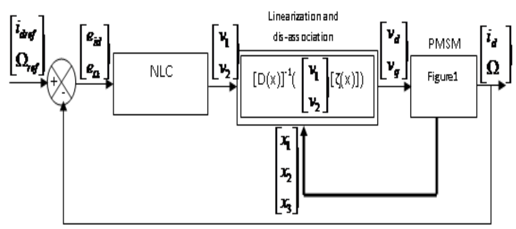

Figure2 : simulation scheme of input-output feedback linearization control

5. NONLINEAR BACKSTEPPING CONTROLLER

5.1 Non adaptive case

It is assumed that the engine parameters are known and invariant.

By choosing $[i_d \quad i_q \quad \Omega]^T$ as variable states and equation (1) the mathematical model of the machine. The objective is to regulate the speed to its reference value . We begin by defining the tracking errors:

$$\begin{aligned} e_w &= \Omega_{ref} - \Omega \\ e_d &= i_{dref} - i_d \quad \text{with } i_{dref} = 0 \\ e_q &= e_{qref} - i_q \end{aligned} \quad (14)$$

And its dynamics derived from (1):

$$\frac{de_w}{dt} = \frac{d\Omega_{ref}}{dt} - \frac{d\Omega}{dt} = -\frac{d\Omega}{dt} \quad (15)$$

We replace (1) in (15), we obtain:

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$$\frac{de_w}{dt} = -\frac{3p}{2J}[(L_d - L_q)i_d i_q + \Phi_f i_q] + \frac{f}{J}\Omega + \frac{T_L}{J} \qquad (16)$$

We define the following quadratic function:

$$V_1 = \frac{1}{2}e_w^2 \qquad (17)$$

Its derivative along the solution of (16), is given by:

$$\dot{V}_1 = e_w \dot{e}_w$$
$$= e_w\left[-\frac{3p}{2J}[(L_d - L_q)i_d i_q + \Phi_f i_q] + \frac{f}{J}\Omega + \frac{T_L}{J}\right] \qquad (18)$$

Utilizing the backstepping design method, we consider the d-q axes currents components i_d and i_q as our virtual control elements and specify its desired behavior, which are called stabilising function in the backstepping design terminology as follows:

$$i_{dref} = 0$$
$$i_{qref} = \frac{2}{3p\Phi_f}\left(f\Omega + T_L + k_w J e_w\right) \qquad (19)$$

Where k_w is a positive constant

Substituting (19) in (18) the derivative of V_1 becomes:

$$\dot{V}_1 = -k_w e_w^2 \qquad (20)$$

Substituting (14) and (19) into (16) the dynamics of the tracking error of the velocity becomes:

$$\frac{de_w}{dt} = \frac{3p\Phi_f}{2J}e_q + \frac{3p}{2J}(L_d - L_q)e_d i_q - k_w e_w \qquad (21)$$

Furthermore the dynamics of the stabilizing errors $(e_d,\ e_q)$ can be computed as:

$$\frac{de_d}{dt} = \frac{di_{dref}}{dt} - \frac{di_d}{dt} = -\frac{di_d}{dt} = -\frac{v_d}{L_d} + \frac{R_s}{L_d}i_d - \Omega\frac{L_q}{L_d}i_q \qquad (22)$$

$$\frac{de_q}{dt} = \frac{di_{qref}}{dt} - \frac{di_q}{dt} =$$
$$\frac{2(k_w J - f)}{3p\Phi_f}\left[\frac{3p\Phi_f}{2J}e_q + \frac{3p}{2J}(L_d - L_q)e_d i_q - k_w e_w\right] -$$
$$\frac{v_q}{L_q} + \frac{R_s}{L_q}i_q + \Omega\frac{L_d}{L_q}i_d + \Omega\frac{\Phi_f}{L_q} \qquad (23)$$

To analyze the stability of this system we propose the following Lyapunov function:

$$V_2 = \frac{1}{2}(e_w^2 + e_d^2 + e_q^2) \qquad (24)$$

Its derivative along the trajectories (21) (22) and (23) is:

$$\dot{V}_2 = e_w\dot{e}_w + e_d\dot{e}_d + e_q\dot{e}_q \qquad (25)$$
$$= -k_w e_w^2 - k_d e_d^2 - k_q e_q^2 +$$
$$e_d\left[k_d e_d - \frac{v_d}{L_d} + \frac{R_s}{L_d} - \Omega\frac{L_q}{L_d}i_q + \frac{3p}{2J}(L_d - L_q)e_w i_q\right]$$
$$+ e_q\left[k_q e_q + \frac{2(k_w J - f)}{3p\Phi_f}\left(\frac{3p\Phi_f}{2J}e_q + \right.\right.$$
$$\left.\frac{3p}{2J}(L_d - L_q)e_d i_q - k_w e_w\right) + \frac{3p\Phi_f}{2J}e_w - \frac{v_q}{L_q} +$$
$$\left.\frac{R_s}{L_q}i_q + \Omega\frac{L_d}{L_q}i_d + \Omega\frac{\Phi_f}{L_q}\right]$$

The expression (25) found above requires the following control laws:

$$v_d = k_d L_d e_d + R_s i_d - \Omega L_q i_q + \frac{3pL_d}{2J}(L_d - L_q)e_w i_q$$
$$v_q = \frac{2L_q(k_w J - f)}{3p\Phi_f}\left(\frac{3p\Phi_f}{2J}e_q + \frac{3p}{2J}(L_d - L_q)e_d i_q - k_w e_w\right) +$$
$$\frac{3p\Phi_f L_q}{2J}e_w + R_s i_q + \Omega L_d i_d + \Omega\Phi_f + k_q L_q e_q \qquad (26)$$

With this choice the derivatives of (24) become:

$$\dot{V}_2 = -k_w e_w - k_d e_d - k_q e_q \le 0 \qquad (27)$$

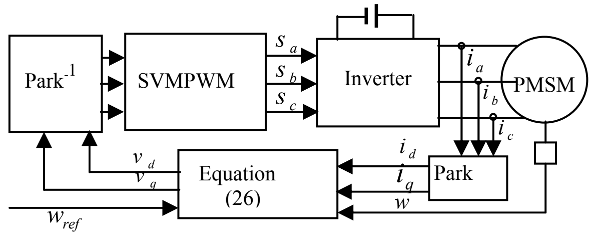

Figure3: Simulation scheme of non-adaptive case

5.2 Adaptive case

In the controller development in the previous section, it was assumed that all the system parameters are known. However, this assumption is not always true. The flux linkage varies nonlinearly with the temperature rise and, also, with the external fields produced by the stator current due to the nonlinear demagnetization characteristics of the magnets. The winding resistance may vary due to heating. In addition, working condition changes

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such as load torque and inertia mismatch inevitably impose parametric uncertainties in control system design. Hence, it becomes necessary to account for all these uncertainties in the design of high performance controller. We will see how we efficiently handle these uncertainties through step-by-step adaptive backstepping design and parameter adaptations.

In ( ) , we do not know exact value of the load torque T_L; then , it is necessary to estimate them adaptively, and replace them with their estimates $\widehat{T}_L$.

Or $\hat{i}_{sqref} = \dfrac{2}{3p\Phi_f}\left(f\omega + \widehat{T}_L + k_\omega J e_\omega\right)$ (28)

So from ( ) and ( ) the following speed error dynamics can be deduced :

$$\frac{de_\omega}{dt} = \frac{1}{J}\left(-\widetilde{T}_L + \frac{3p\Phi_f}{2}e_q + \frac{3p}{2}(L_d - L_q)e_d i_q - k_\omega J e_\omega\right)$$

where $\widetilde{T}_L = \widehat{T}_L - T_L$ (29)

As a result, the d-q currents errors dynamics can be rewritten as :

$$\frac{de_d}{dt} = -\frac{di_d}{dt} = -\frac{v_d}{L_d} + \frac{R_s}{L_d}i_d - \Omega\frac{L_q}{L_d}i_q \quad (30)$$

$$\frac{de_q}{dt} = \frac{di_{qref}}{dt} - \frac{di_q}{dt} =$$
$$\frac{2(k_w J - f)}{3p\Phi_f}\left[\frac{3p\Phi_f}{2J}e_q + \frac{3p}{2J}(L_d - L_q)e_d i_q - k_w e_w\right] -$$
$$\frac{v_q}{L_q} + \frac{R_s}{L_q}i_q + \Omega\frac{L_d}{L_q}i_d + \Omega\frac{\Phi_f}{L_q} + \frac{2}{3p\Phi_f}\left(\frac{f}{J} - k_\omega\right)\widetilde{T}_L \quad (31)$$

Let us define a new Lyapunov function the closed loop system as follows:

$$V_2 = \frac{1}{2}\left(e_w^2 + e_d^2 + e_q^2 + \frac{\widetilde{T}_L^2}{\gamma_1} + \frac{\widetilde{R}_s^2}{\gamma_2} + \frac{\widetilde{\Phi}_f^2}{\gamma_3}\right) \quad (32)$$

Where the stabilizing errors e_ω , e_d, e_q, the load torque estimation error $\widetilde{T}_L$, the resistance estimation error $\widetilde{R}_s$ ($\widetilde{R}_s = \hat{R}_s - R_s$ and the rotor magnetic flux linkage estimation error $\widetilde{\Phi}_f$ ($\widetilde{\Phi}_f = \hat{\Phi}_f - \Phi_f$) are all included. And γ_1, γ_2 and γ_3 are an adaptation gain. Tracking the derivative of (32) and substing (29) , (30) and (31) into this derivative we can obtain:

$$\dot{V}_2 = e_w\dot{e}_w + e_d\dot{e}_d + e_q\dot{e}_q +$$
$$\frac{1}{\gamma_1}\widetilde{T}_L\dot{\widetilde{T}}_L + \frac{1}{\gamma_2}\widetilde{R}_s\dot{\widetilde{R}}_s + \frac{1}{\gamma_3}\widetilde{\Phi}_f\dot{\widetilde{\Phi}}_f \quad (33)$$
$$= -k_w e_w^2 - k_d e_d^2 - k_q e_q^2 +$$
$$e_d\left[k_d e_d - \frac{v_d}{L_d} + \frac{R_s}{L_d} - \Omega\frac{L_q}{L_d}i_q + \frac{3p}{2J}(L_d - L_q)e_w i_q\right]$$
$$+ e_q\left[k_q e_q + \frac{2(k_w J - f)}{3p\hat{\Phi}_f}\left(\frac{3p\hat{\Phi}_f}{2J}e_q +\right.\right.$$
$$\left.\frac{3p}{2J}(L_d - L_q)e_d i_q - k_w e_w\right) + \frac{3p\hat{\Phi}_f}{2J}e_w - \frac{v_q}{L_q} +$$
$$\left.\frac{R_s}{L_q}i_q + \Omega\frac{L_d}{L_q}i_d + \Omega\frac{\hat{\Phi}_f}{L_q}\right] + \widetilde{T}_L\left[\frac{1}{\gamma_1}\dot{\widetilde{T}}_L - \frac{2k_\omega e_q}{3p\hat{\Phi}_f} + \frac{2fe_q}{3pJ\hat{\Phi}_f} - \frac{e_\omega}{J}\right]$$
$$+ \widetilde{R}_s\left[\frac{1}{\gamma_2}\dot{\widetilde{R}}_s + \frac{1}{L_d}i_d e_d - \frac{1}{L_q}i_q e_q\right] + \widetilde{\Phi}_f\left[\frac{1}{\gamma 3}\dot{\widetilde{\Phi}}_f - \frac{3p}{2J}e_\omega e_q -\right.$$
$$\left.\frac{(k_\omega J - f)}{J\hat{\Phi}_f}e_q^2 - \frac{1}{Lq}\Omega e_q\right]$$

According to the above equation (33) , the control laws are now designed as:

$$v_d = k_d L_d e_d + \hat{R}_s i_d - \Omega L_q i_q + \frac{3pL_d}{2J}(L_d - L_q)e_w i_q$$
$$v_q = \frac{2L_q(k_w J - f)}{3p\hat{\Phi}_f}\left(\frac{3p\hat{\Phi}_f}{2J}e_q + \frac{3p}{2J}(L_d - L_q)e_d i_q - k_w e_w\right) + \quad (34)$$
$$\frac{3p\hat{\Phi}_f L_q}{2J}e_w + \hat{R}_s i_q + \Omega L_d i_d + \Omega\hat{\Phi}_f + k_q L_q e_q$$

Consequently (33) is simplified to :

$$\dot{V}_2 = -k_w e_w^2 - k_d e_d^2 - k_q e_q^2 +$$
$$+ \widetilde{T}_L\left[\frac{1}{\gamma_1}\dot{\widetilde{T}}_L - \frac{2k_\omega e_q}{3p\hat{\Phi}_f} + \frac{2fe_q}{3pJ\hat{\Phi}_f} - \frac{e_\omega}{J}\right] + \quad (35)$$
$$\widetilde{R}_s\left[\frac{1}{\gamma_2}\dot{\widetilde{R}}_s + \frac{1}{L_d}i_d e_d - \frac{1}{L_q}i_q e_q\right] +$$
$$\widetilde{\Phi}_f\left[\frac{1}{\gamma 3}\dot{\widetilde{\Phi}}_f - \frac{3p}{2J}e_\omega e_q - \frac{(k_\omega J - f)}{J\hat{\Phi}_f}e_q^2 - \frac{1}{L_q}\Omega e_q\right]$$

The parameter adaptation laws are then chosen as :

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$$\dot{\tilde{T}}_L = \gamma_1 \left[\frac{2k_\omega e_q}{3p\hat{\Phi}_f} - \frac{2fe_q}{3pJ\hat{\Phi}_f} + \frac{e_\omega}{J} \right] \quad (36)$$

$$\dot{\tilde{R}}_s = \gamma_2 \left[-\frac{1}{L_d} i_d e_d + \frac{1}{L_q} i_q e_q \right] \quad (37)$$

$$\dot{\tilde{\Phi}}_f = \gamma_3 \left[\frac{3p}{2J} e_\omega e_q + \frac{(k_\omega J - f)}{J\hat{\Phi}_f} e_q^2 + \frac{1}{L_q} \Omega e_q \right] \quad (38)$$

Then (35) can be rewritten as follows

$$\dot{V}_2 = -k_w e_w^2 - k_d e_d^2 - k_q e_q^2 \leq 0 \quad (39)$$

Define the following equation

$$Q(t) = k_w e_w^2 + k_d e_d^2 + k_q e_q^2 \geq 0 \quad (40)$$

Furthermore, by using LaSalle Yoshizawa's theorem [11], its can be shown that Q(t) tends to zero as t →∞, ew, ed and eq will converge to zero as. Consequently, the proposed controller is stable and robust, despite the parameter uncertainties.

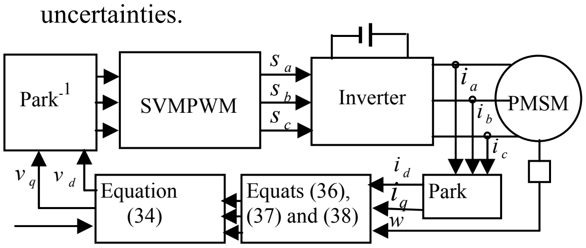

Figure4: simulation scheme of an adaptive case

6. Simulation Results

6.1 PMSM parameter's

TABLE I : PARAMETERS OF PMSM

parameter	value
Maximal voltage of food	300 v
Maximal speed	3000 tr/s to 150 Hz
Nominal Torque ;T_n	14.2 N.ms
R_s	0.4578 Ω
Number of pair poles :p	4
L_d	3.34 mH
L_q	3.58 mH
The moment of inertia J	0.001469 kg.m2s
Coefficient of friction viscous f	0.0003035 Nm/Rad/s
Flux of linquage Φ_f	0.171

6.2 Results

For a trajectory w_{ref}=200 rad/s at 0s, =100rad/s at t=0.15s, =200rad/s at t=0.25s, the following figures (1 to 7) show the performance of the input output linearization control.

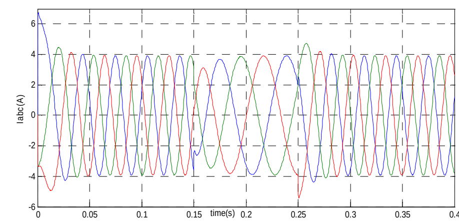
Figure3: i_{abc} currents without uncertainties

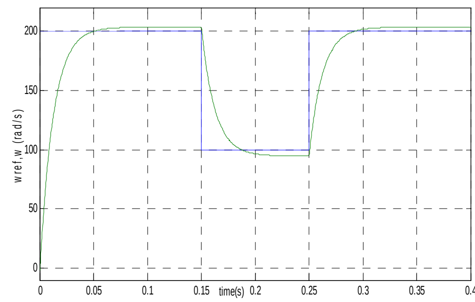
Figure4: speed response without uncertainties

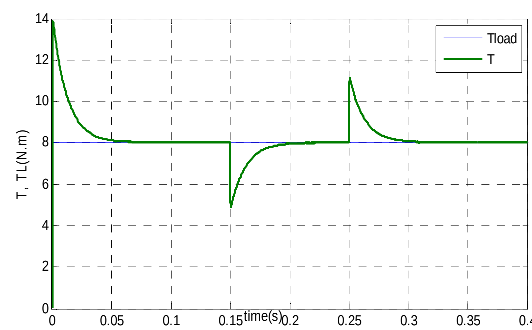

Figure5: Torque response without uncertainties

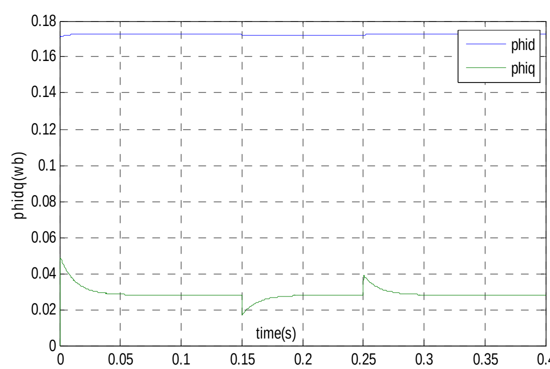

figure6: d-q axis flux without uncertainties

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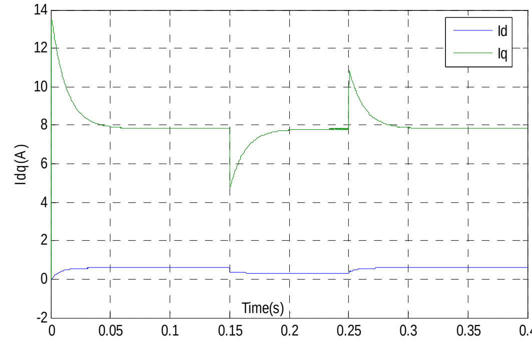

Figure7: d-q axis currents response without uncertainties

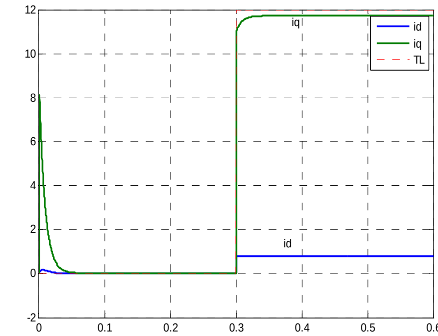

figure10: d-q axis currents with step load and without uncertainties.

The figures below (figures 8, 9 and 10) shows the evolution of speed, electromagnetic torque and d-q axis currents in the presence of a disturbance load torque.

The figures 10 to 13 show the robustness of adaptive backstepping controller compared with that of non-adaptive case and feedback input-output linearization.

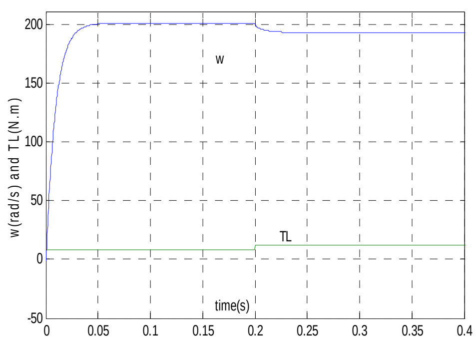

Figure8: response speed with step load at t= 0.2s

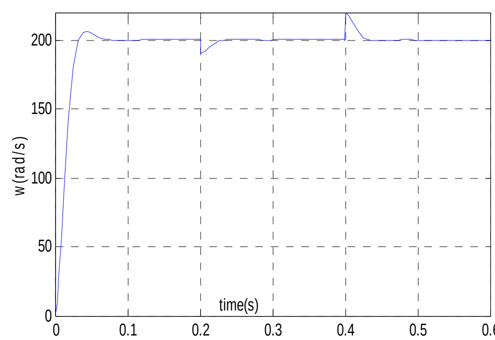

Figure11: speed response of the proposed adaptive controller with step load (+12Nm at 0.2s , +8Nm at 0.4s)

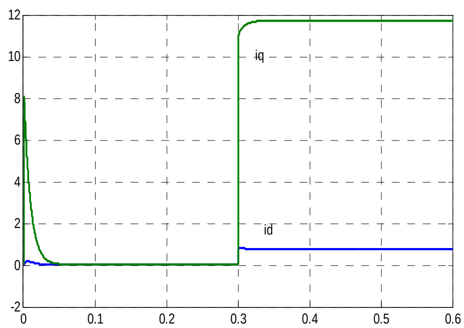

Figure9: electromagnetic torque with step load

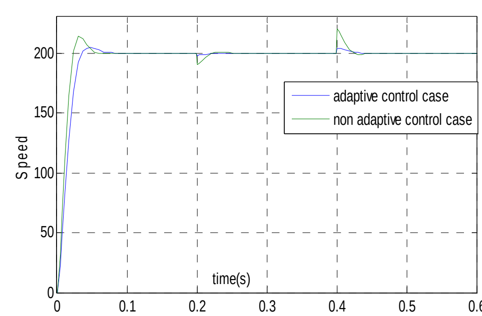

Figure12: speed response of the non-adaptive and adaptive controller with step load

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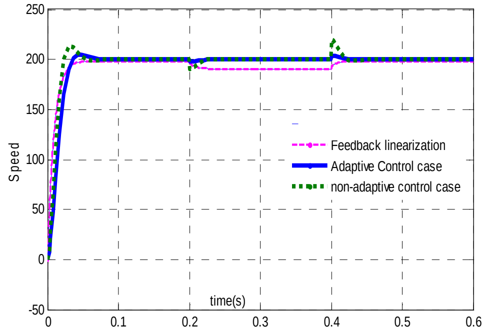

Figure13: speed response of the input-output linearization- non-adaptive and adaptive controller with variation of parameters ($R_s = R_{s0} \pm \Delta R_s$)

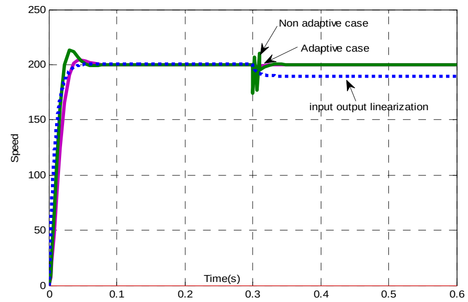

Figure14: speed response of the input-output linearization- non-adaptive and adaptive controller with variation of parameters ($\Phi_f = \Phi_{f0} \pm \Delta\Phi_f$) at t=0.3 s

6. CONCLUSION:

A method of nonlinear backstepping control has been proposed and used for the control of a PMSM. The simulation results show, with a good choice of control parameters, good performance obtained with the proposed control as compared with the non-linear control by state feedback. From the speed tracking simulations results, we can find that the nonlinear adaptive backstepping controllers have excellent performance. The direct axis current id is always forced to zero in order to orient all the linkage flux in the d-axis and to achieve maximum torque per ampere. The q-axis current will approach a constant when the actual motor speed achieves reference speed.
The show simulation results, fast response without overshoot and robust performance to parameter variations and disturbances throughout the system

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