Robust Nonlinear Control of AC Brushless Motor for Electric Vehicles Application

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Abstract

This article proposes a robust nonlinear control based on Lyapunov's redesign, whose purpose is to deal with parametric uncertainty in the resistance of the motor windings. The robust controller design is based on the passivity properties of the motor, as well as energy shaping and damping injection. The application of this control technique is focused on electric vehicles mainly formed by a battery bank, a power inverter, an AC brushless motor and the mechanical transmission. The sine PWM technique is used to trigger the switching devices of inverter. The results were obtained from simulation, where is shown that robust control makes a proper tracking of electromagnetic torque.

Key Words: Electric vehicles, Lyapunov's stability, Passivity, Robust nonlinear control, Sine PWM

I. INTRODUCTION

The electric vehicle (EV) does not damage our environment because it does not emits gases. The global need for oil and strong emissions to the atmosphere by combustion cars makes electric transportation an effective solution. In recent years the development of the EV has increased, also the development of electric motors capable of satisfying the traction requirements. The AC brushless motor (ACBM) has significant advantages over other electric machines: there is no slip, maintenance free because there is no slip rings, permanent magnet in the rotor requires no electricity to the field excitation, high ratio of power/weight, robustness and efficiency [3]. The ACBM is also used in power generation systems and as servo-systems actuator [12].

Several types of controllers were designed in order to achieve their control objectives under specific conditions. In [17], a sensorless control technique is proposed using an observer (easily implementable) for both angular position and speed of the PMSM rotor. The design of this observer is based on immersion and invariance technique. For example, [20] proposes a control technique to perform vector control of a Permanent Magnet Synchronous Motor (PMSM) without measurement of the rotor position and speed using high frequency voltage injection into stator terminals. Also, [1] proposes a sensorless control for a synchronous motor in a single piston rotary compressor, this technique is focused to reduce speed ripples, as well as, mechanical noise and

vibration. In [8] a robust adaptive backstepping controller is proposed for the speed control of PMSM in pure electrical vehicles applications under several electrical and mechanical parameters uncertainties. Another good reference in EV speed control is [7], this shows the road estimation and speed control of a car with four PMSM mounted in each wheel. The advantage of this configuration is that each wheel torque can be controlled completely and independently from the others. [2] and [14] are two good references of Passivity Based Control for nonlinear systems.

The next sections presents: three-phase and two-phase Euler-Lagrange (E-L) models for the ACBM, EV model, nominal control design based on passivity and robust control design based on Lyapunov's redesign to deal with parametric uncertainty in the resistance of ACBM windings. A controller is designed to achieve the torque tracking objective in all EV operation zone, ensuring signals bounding and internal stability of the entire system. Simulations are presented to illustrate the performance of this controller taking into account variations in electrical either mechanical parameters.

II. SYSTEM MODEL

This section presents the mathematical model development proposed for the EV, which is usually formed by an electrical subsystem (battery bank, three-phase power inverter and motor) and a mechanical subsystem (motor, mechanical transmission and wheels), both illustrated in figure 1. The ACBM is part of both subsystems, because is responsible to transform the electric energy into mechanical energy and also provides the necessary traction force for EV motion. The battery model is considered an ideal voltage source capable of delivering the motor current demanded from it.

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Fig. 1. EV diagram.

Fig. 2. Inverter Topology.

A. Power Inverter Model Using Sine PWM Technique

The DC-AC power electronic converter is also called inverter and the general purpose of this topology is to control amplitude, phase and frequency of the generated three-phase voltages. The presented model is reported in [19] and only considers ideal switching devices ([15] and [18] presents an electro-thermal model for switching devices and a high frequency IGBT model, respectively). Fig. 2 shows the general inverter topology formed by six switching devices (S_1, S_2, S_3, S_4, S_5 and S_6) and six free-wheeling diodes. The carrier signal used in sine PWM modulation is a triangular wave form represented by $g(t) = \frac{2}{\pi} \arcsin(\sin(2\pi f_s t + \frac{\pi}{2}))$, where f_s is the carrier frequency [16]. Modulating signals are sine control signals (M_a, M_b, M_c) and are also shown in figure 3.

Sine PWM technique compares the modulating signals with the carrier signal to produce binary trigger variables a, b and c . For example, if at any time the modulating signal M_a is bigger than carrier signal $g(t)$, then a is a logic one so S_1 is on and S_2 is off. In the case that M_a is lower than $g(t)$, then a is a logic zero so S_1 is off and S_2 is on. The equation (1) relates the binary variables a, b and c with voltage v_i to generate the phase voltages v_a, v_b and v_c , where v_i represents the output voltage of the battery bank.

$$\begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} = \frac{v_i}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} a \\ b \\ c \end{bmatrix} \quad (1)$$

For the case of close-loop system operation, three-phase signals from nonlinear controller are divided by the voltage bus v_i to produce modulating signals. [10] is a good reference that proposes a passivity-based control strategy for single-phase inverters, this approach is an alternative to ours.

Fig. 3. Carrier and modulating signals.

B. AC Brushless Motor Model in Euler-Lagrange Formulation

The AC BM is a Permanent Magnet Synchronous Motor in which the rotor angular velocity is related to the angular velocity of the rotating magnetic field in the stator and there is no slip between this two velocities. The stator contains multiple coils per phase winding distributed in slots around it. The rotor consists of one or more permanent magnets with the aim of generating the magnetic field of the rotor. For our analysis it is assumed that the stator windings are sinusoidally distributed. The theory to obtain the model is available at [13]. The E-L equations for both, electrical and mechanical subsystems are described by:

$$\begin{aligned} \Sigma_e : D_e \ddot{q}_e + W_1 \dot{q}_m \dot{q}_e + W_2 \dot{q}_m + R_e \dot{q}_e &= M_e u \text{ and} \\ \Sigma_m : D_{MB} \ddot{q}_m - \tau_{em} + R_m \dot{q}_m &= -\tau_L. \end{aligned} \quad (2)$$

The electromagnetic torque is defined as:

$$\tau_{em} = \frac{1}{2} \dot{q}_e^T W_1 \dot{q}_e + \dot{q}_e^T W_2. \quad (3)$$

For electrical subsystem Σ_e the generalized coordinates are defined as $q_e = [q_1 \ q_2 \ q_3]^T$, which represents the electric charges in motor windings. The generalized velocities $\dot{q}_e = [\dot{q}_1 \ \dot{q}_2 \ \dot{q}_3]^T = [i_a \ i_b \ i_c]^T$ are the currents in the motor windings. For mechanical subsystem Σ_m the generalized coordinate is defined as the mechanical angular position q_m . The mechanical angular velocity is $\dot{q}_m$. $R_e = \text{diag}\{r_e, r_e, r_e\}$ is the resistance matrix of the stator windings. M_e is a 3×3 dimensions identity matrix and the control input is $u = [u_1 \ u_2 \ u_3]^T$, corresponding to the stator voltages. D_{MB} , R_m and τ_L are the ACBM inertia, the viscous friction coefficient and load torque respectively. The elements of the inductance matrix D_e are defined in [11] by

$$D_e = \begin{bmatrix} l_{aa} & l_{ab} & l_{ac} \\ l_{ba} & l_{bb} & l_{bc} \\ l_{ca} & l_{cb} & l_{cc} \end{bmatrix}, \quad (4)$$

where each element is represented by:

$$\begin{aligned} l_{aa} &= L_{ls} + \bar{L}_m - L_{\Delta m} \cos 2(n_p \theta_m) \\ l_{bb} &= L_{ls} + \bar{L}_m - L_{\Delta m} \cos 2(n_p \theta_m - 2\pi/3) \\ l_{cc} &= L_{ls} + \bar{L}_m - L_{\Delta m} \cos 2(n_p \theta_m + 2\pi/3) \\ l_{ab} &= l_{ba} = -\bar{L}_m/2 - L_{\Delta m} \cos 2(n_p \theta_m - \pi/3) \\ l_{ac} &= l_{ca} = -\bar{L}_m/2 - L_{\Delta m} \cos 2(n_p \theta_m + \pi/3) \\ l_{bc} &= l_{cb} = -\bar{L}_m/2 - L_{\Delta m} \cos 2(n_p \theta_m + \pi). \end{aligned} \quad (5)$$

L_{ls} is the leakage inductance, $\bar{L}_m$ is the average value of magnetizing inductance and $L_{\Delta m}$ is the peak value of magnetizing inductance. $W_1 = \partial D_e / \partial q_m$ and $W_2 = \partial \mu / \partial q_m$, where μ is the flux linkages vector due to the permanent magnets and is represented as:

$$\mu = \lambda_m \begin{bmatrix} \sin(n_p q_m) \\ \sin(n_p q_m - 2\pi/3) \\ \sin(n_p q_m + 2\pi/3) \end{bmatrix}. \quad (6)$$

λ_m is the flux linkages amplitude and is determined by the magnetic material in the rotor, n_p is the pole pair number of the motor. The angular electric position θ_e is related with the angular mechanic position q_m by $\theta_e = n_p q_m$.

C. Motor Model in Fixed Rotor Reference Frame

The mathematical model of the electrical subsystem Σ_e given by the equation (2), has coefficients which depend on the rotor position. The Park's transformation is used to eliminate this dependence. This transformation allows refer all electric variables to the fixed rotor reference frame using $f_{qd0} = k_s^r f_{abc}$ where the Park's transformation matrix k_s^r is given by [12]:

$$k_s^r = \frac{2}{3} \begin{bmatrix} \cos n_p q_m & \cos(n_p q_m - \frac{2\pi}{3}) & \cos(n_p q_m + \frac{2\pi}{3}) \\ \sin n_p q_m & \sin(n_p q_m - \frac{2\pi}{3}) & \sin(n_p q_m + \frac{2\pi}{3}) \\ 1/2 & 1/2 & 1/2 \end{bmatrix} \quad (7)$$

This transformation is only used for the electrical subsystem in equation (2), this equation now is expressed in function of $(qd0)$ variables, in other words, $\dot{q}_e = (k_s^r)^{-1} \dot{q}_e^r$, $D_e = (k_s^r)^{-1} D_e^r$, $\mu = (k_s^r)^{-1} \mu^r$, $\dot{q}_e = (k_s^r)^{-1} \dot{q}_e^r$ and $u = (k_s^r)^{-1} u^r$:

$$D_e(k_s^r)^{-1} \ddot{q}_e^r + \frac{\partial(k_s^r)^{-1} D_e^r}{\partial q_m} \dot{q}_m \dot{q}_e + \frac{\partial(k_s^r)^{-1} \mu^r}{\partial q_m} \dot{q}_m \dots \dots + R_e(k_s^r)^{-1} \dot{q}_e^r = M_e(k_s^r)^{-1} u^r. \quad (8)$$

Now all terms are multiplied by k_s^r :

$$k_s^r D_e(k_s^r)^{-1} \ddot{q}_e^r + k_s^r \frac{\partial(k_s^r)^{-1} D_e^r}{\partial q_m} \dot{q}_m \dot{q}_e + k_s^r \frac{\partial(k_s^r)^{-1} \mu^r}{\partial q_m} \dot{q}_m \dots \dots + k_s^r R_e(k_s^r)^{-1} \dot{q}_e^r = k_s^r M_e(k_s^r)^{-1} u^r. \quad (9)$$

If we consider that $\mu^r = k_s^r \mu(q_m) = [0 \ \lambda_m \ 0]^T$ and developing all the indicated operations in equation (9) then the model for the electrical subsystem Σ_e in fixed rotor reference frame is:

$$D_e^r \ddot{q}_e^r + W_1^r \dot{q}_m \dot{q}_e^r + W_2^r \dot{q}_m + R_e^r \dot{q}_e^r = u^r, \quad (10)$$

with

$$D_e^r = \begin{bmatrix} L_q & 0 & 0 \\ 0 & L_d & 0 \\ 0 & 0 & L_{ls} \end{bmatrix}, \quad W_1^r = n_p \begin{bmatrix} 0 & L_d & 0 \\ -L_q & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix},$$

$$W_2^r = \begin{bmatrix} n_p \lambda_m \\ 0 \\ 0 \end{bmatrix} \quad \text{and} \quad R_e^r = \begin{bmatrix} r_e & 0 & 0 \\ 0 & r_e & 0 \\ 0 & 0 & 0 \end{bmatrix}.$$

Fig. 4. EV transmission.

The control signal is now defined as $u^r = [u_1^r \ u_2^r \ u_3^r]^T = [v_q \ v_d \ v_0]^T$ and current vector as $\dot{q}_e^r = [\dot{q}_1^r \ \dot{q}_2^r \ \dot{q}_3^r]^T = [\dot{i}_q \ \dot{i}_d \ \dot{i}_0]^T$. The new expression of electromagnetic torque is:

$$\tau_{em} = \frac{3}{2} n_p \dot{q}_1^r ((L_d - L_q) \dot{q}_2^r + \lambda_m). \quad (11)$$

The q and d inductances are: $L_q = L_{ls} + \frac{3}{2} \bar{L}_m - \frac{3}{2} L_{\Delta m}$ and $L_d = L_{ls} + \frac{3}{2} \bar{L}_m + \frac{3}{2} L_{\Delta m}$.

When stator windings resistances have the same value, the resistance matrix of the stator is:

$$R_e = \begin{bmatrix} r_e & 0 & 0 \\ 0 & r_e & 0 \\ 0 & 0 & r_e \end{bmatrix} \quad (12)$$

and satisfies

$$R_e^r = k_s^r R_e (k_s^r)^{-1} = R_e, \quad (13)$$

in other words, despite the transformation, the resistance matrix remains exactly as before to apply this transformation. In the other hand, when stator windings resistances have different value, the resistance matrix of the stator is represented by single phase resistance value:

$$R_e = \begin{bmatrix} r_a & 0 & 0 \\ 0 & r_b & 0 \\ 0 & 0 & r_c \end{bmatrix}, \quad (14)$$

and the transformation to fixed rotor reference frame is:

$$R_{ex}^r = k_s^r R_{ex} (k_s^r)^{-1} = \begin{bmatrix} r_{11} & r_{12} & r_{13} \\ r_{21} & r_{22} & r_{23} \\ r_{31} & r_{32} & r_{33} \end{bmatrix}, \quad (15)$$

where $r_{11} = \frac{2}{3} r_a + \frac{1}{6} r_b + \frac{1}{6} r_c$, $r_{12} = \frac{\sqrt{3}}{6} (r_b - r_c)$, $r_{13} = \frac{2}{3} r_a - \frac{1}{3} r_b - \frac{1}{3} r_c$, $r_{21} = \frac{1}{6} (r_b - r_c)$, $r_{22} = \frac{1}{2} (r_b + r_c)$, $r_{23} = \frac{\sqrt{3}}{3} (r_b - r_c)$, $r_{31} = \frac{1}{3} r_a - \frac{1}{6} r_b - \frac{1}{6} r_c$, $r_{32} = -\frac{\sqrt{3}}{6} (r_b - r_c)$, $r_{33} = \frac{1}{3} (r_a + r_b + r_c)$.

The subscript R_{ex} refers to an explicit resistances matrix and this is used in the robust nonlinear controller design.

D. Electric Vehicle Model

When a simple transmission like in figure 4 is considered [5] [6], the equation used to modelling this transmission is:

$$\tau = \frac{r}{\eta_g G} F_{te}. \quad (16)$$

Where r is the radius of the wheel, η_g is the trasmission efficiency, G is the reduction ratio of the angular speed of the transmission, τ is the torque given in the motor shaft and F_{te} is the traction force which makes possible the EV motion. In figure 4 we obtain the relationship between angular speed

Fig. 5. Forces acting on the car.

$\dot{q}_m$ and lineal speed v of the EV with:

$$\dot{q}_m = G \frac{v}{r}. \quad (17)$$

The forces acting on the car are shown in figure 5. The force F_{te} drives the vehicle and is provided by the electric motor through the transmission. The friction force F_{rr} between the wheels and ground is given by $F_{rr} = \mu_{rr}mg \cos(\psi)$, where μ_{rr} is a friction coefficient, m is the total mass of the EV in Kg, $g = 9.8 \text{ m/s}^2$ is the gravity constant and ψ is the inclination angle.

The force given by the air is $F_{ad} = \frac{1}{2}\rho AC_d v^2$, where $\rho = 1.25 \text{ kg/m}^3$ is the air density, A is the frontal area of the car in m^2 , C_d is the aerodynamic coefficient and v is the lineal speed in m/s . The force F_{hc} is a function of inclination angle ψ and is represented by $F_{hc} = mg \sin(\psi)$.

With the Newton's second law we obtain:

$$F_{te} - F_{rr} - F_{ad} - F_{hc} = ma, \quad (18)$$

where a is the car acceleration in m/s^2 . Solving for the traction force F_{te} in equation (18) and considering all expressions of the forces acting on the car, the next expression is obtained:

$$F_{te} = ma + \mu_{rr}mg \cos(\psi) + \frac{1}{2}\rho AC_d v^2 + mg \sin(\psi) \quad (19)$$

In this last equation (19), the AC brushless motor and the transmission have not been considered. In the mechanical subsystem Σ_m given by equation (2), the electromagnetic torque is:

$$\tau_{em} = D_{MB}\ddot{q}_m + R_m\dot{q}_m + \tau_L. \quad (20)$$

The total inertia is the sum of the BM inertia and the EV inertia:

$$D_m = D_{MB} + D_{VE}. \quad (21)$$

The inertia of the EV is $D_{VE} = \frac{1}{2}m \frac{r^2}{G^2}$. The equations (16) and (21) are replaced in (20) and the next expression is obtained:

$$\tau_{em} = D_m\ddot{q}_m + R_m\dot{q}_m + \frac{r}{\eta_g G} F_{te}. \quad (22)$$

When the equation (19) is replaced in (22), results:

$$\begin{aligned} \tau_{em} = D_m\ddot{q}_m + R_m\dot{q}_m + \frac{r}{\eta_g G} \left(m \frac{dv}{dt} + \mu_{rr}mg \cos(\psi) \dots \right. \\ \left. \dots + \frac{1}{2}\rho AC_d v^2 + mg \sin(\psi) \right). \end{aligned} \quad (23)$$

The equation (23) contains both angular speed $\dot{q}_m$ and lineal speed v . Finally the mechanical subsystem model is

obtained when the equation (23) is only in function of $\dot{q}_m$ using equation (17) and solving for $\ddot{q}_m$:

$$\begin{aligned} \ddot{q}_m = \frac{\eta_g G^2}{D_m \eta_g G^2 + m r^2} \left(\tau_{em} - R_m \dot{q}_m - \frac{\mu_{rr} m g r \cos(\psi)}{\eta_g G} \dots \right. \\ \left. \dots - \frac{\rho A C_d r^3}{2 \eta_g G^3} \dot{q}_m^2 - \frac{m g r \sin(\psi)}{\eta_g G} \right). \end{aligned} \quad (24)$$

This equation (24) represents all the mechanical subsystem of the EV given by figure 1, the mechanical dynamics of the BM is also included.

III. ROBUST CONTROLLER DESIGN

A. Nominal Control Design Based on Passivity

The nominal control design for the ACBM is divided in two parts: a controller for mechanical subsystem and a controller for electrical subsystem. For the mechanical subsystem the controller design begin with equation (20) and its desired dynamics:

$$\begin{aligned} D_m \ddot{q}_m + R_m \dot{q}_m + \tau_L = \tau_{em} \text{ and} \\ D_m \ddot{q}_{md} + R_m \dot{q}_{md} + \tau_L = \tau_{emd}. \end{aligned} \quad (25)$$

The subscript d refer to the desired trajectories. Now the speed and acceleration errors are defined as $\tilde{q}_m = \dot{q}_m - \dot{q}_{md}$ and $\tilde{\ddot{q}}_m = \ddot{q}_m - \ddot{q}_{md}$ respectively. The error dynamic equation is obtained by subtract equations in (25):

$$D_m \ddot{\tilde{q}}_m + R_m \dot{\tilde{q}}_m = \tau_{em} - [D_m \ddot{q}_{md} + R_m \dot{q}_{md} + \tau_L]. \quad (26)$$

The second member of the last equation is the input to the error dynamic equation and is defined as:

$$\sigma_m = \tau_{em} - [D_m \ddot{q}_{md} + R_m \dot{q}_{md} + \tau_L], \quad (27)$$

hence, if $\tau_{em} = \tau_{emd}$, then, $\sigma_m = 0$. The next Lyapunov's function is proposed to design the controller for the mechanical subsystem:

$$V(\dot{\tilde{q}}_m) = \frac{1}{2} D_m \dot{\tilde{q}}_m^2, \quad (28)$$

as $D_m \geq 0$ and also $\dot{\tilde{q}}^2 > 0$, it can say that $V(\dot{\tilde{q}}_m) > 0$, also, $V(0) = 0$, $V(\dot{\tilde{q}}_m) \neq 0$ and $V(\infty) \rightarrow \infty$.

The derivative of the candidate function is:

$$\dot{V} = \dot{\tilde{q}}_m D_m \ddot{\tilde{q}}_m. \quad (29)$$

This derivative must be negative definite to ensure stability. The $D_m \ddot{\tilde{q}}_m$ term in equation (26) is replaced in equation (29), so we obtain:

$$\dot{V} = \dot{\tilde{q}}_m (\tau_{em} - [D_m \ddot{q}_{md} + R_m \dot{q}_{md} + \tau_L]). \quad (30)$$

To force $\dot{V}$ be a defined negative function is considered:

$$\tau_{em} - [D_m \ddot{q}_{md} + R_m \dot{q}_{md} + \tau_L] = -\Gamma_m \dot{\tilde{q}}_m, \quad \Gamma_m > 0. \quad (31)$$

So, $\dot{V}$ is now negative definite and its expression takes de form:

$$\dot{V} = -\dot{\tilde{q}}_m \Gamma_m \dot{\tilde{q}}_m = -\Gamma_m \dot{\tilde{q}}_m^2. \quad (32)$$

The required torque for reference speed tracking comes from τ_{em} in equation (31). This torque and its derivative are renamed as τ_{emd} and $\dot{\tau}_{emd}$ and are defined by:

$$\begin{aligned}\tau_{emd} &= -\Gamma_m \dot{q}_m + D_m \ddot{q}_m + R_m \dot{q}_m + \tau_L \text{ and} \\ \dot{\tau}_{emd} &= -\Gamma_m \ddot{q}_m + D_m \dddot{q}_m + R_m \ddot{q}_m + \dot{\tau}_L.\end{aligned}\quad (33)$$

The load torque and its derivative are defined in the EV model as:

$$\begin{aligned}\tau_L &= \frac{r}{\eta_g G} \left(\frac{mr}{G} \ddot{q}_m + \mu_{rr} mg \cos(\psi) + \frac{\rho A C_d r^2}{2G^2} \dot{q}_m^2 \dots \right. \\ &\quad \left. \dots + mg \sin(\psi) \right), \\ \dot{\tau}_L &= \frac{r}{\eta_g G} \left(\frac{mr}{G} \dddot{q}_m - \mu_{rr} mg \dot{\psi} \sin(\psi) + \frac{\rho A C_d r^2}{G^2} \dot{q}_m \ddot{q}_m \dots \right. \\ &\quad \left. \dots + mg \dot{\psi} \cos(\psi) \right).\end{aligned}\quad (34)$$

It is important to say that the nominal controller design for the electrical subsystem of the motor takes into account that $\dot{q}_3^r = i_0$ has a non-zero dynamic, because the windings resistances are unbalanced, they have not the same value. Also, the control signal has three elements, $u^r = [u_1^r \ u_2^r \ u_3^r]^\top = [v_q \ v_d \ v_0]^\top$. In the controller design, the input trajectories are defined by the desired torque and its derivative in equation (33). The desired currents are obtained from the equation:

$$\tau_{emd} = \frac{3}{2} n_p \dot{q}_{1d}^r ((L_d - L_q) \dot{q}_{2d}^r + \lambda_m). \quad (35)$$

To emulate the torque in a DC motor, is considered $\dot{q}_{2d}^r = 0$, then the last expression takes the form:

$$\tau_{emd} = \frac{3}{2} n_p \lambda_m \dot{q}_{1d}^r, \quad (36)$$

also it is desired that the zero phase current be $\dot{q}_{3d}^r = 0$. Solving for $\dot{q}_{1d}^r$ in the last equation, the desired current vector and its derivative is:

$$\dot{q}_{ed}^r = \begin{bmatrix} \frac{2\tau_{emd}}{3n_p\lambda_m} \\ 0 \\ 0 \end{bmatrix} \quad \text{and} \quad \ddot{q}_{ed}^r = \begin{bmatrix} \frac{2\dot{\tau}_{emd}}{3n_p\lambda_m} \\ 0 \\ 0 \end{bmatrix}. \quad (37)$$

Now, the equations for the electrical subsystem and its desired dynamics are:

$$\begin{aligned}D_e^r \ddot{q}_e^r + W_1^r \dot{q}_m \dot{q}_e^r + W_2^r \dot{q}_m + R_e^r \dot{q}_e^r &= u^r \text{ and} \\ D_e^r \ddot{q}_{ed}^r + W_1^r \dot{q}_m \dot{q}_{ed}^r + W_2^r \dot{q}_m + R_e^r \dot{q}_{ed}^r &= u_d^r.\end{aligned}\quad (38)$$

The electric error and its derivative are defined as:

$$\begin{aligned}e &= \dot{q}_e^r - \dot{q}_{ed}^r = \begin{bmatrix} \dot{q}_1^r \\ \dot{q}_2^r \\ \dot{q}_3^r \end{bmatrix} - \begin{bmatrix} \dot{q}_{1d}^r \\ \dot{q}_{2d}^r \\ \dot{q}_{3d}^r \end{bmatrix} \quad \text{and} \\ \dot{e} &= \ddot{q}_e^r - \ddot{q}_{ed}^r = \begin{bmatrix} \ddot{q}_1^r \\ \ddot{q}_2^r \\ \ddot{q}_3^r \end{bmatrix} - \begin{bmatrix} \ddot{q}_{1d}^r \\ \ddot{q}_{2d}^r \\ \ddot{q}_{3d}^r \end{bmatrix}.\end{aligned}\quad (39)$$

Subtracting the systems dynamics and the desired dynamics, the electric error dynamic equation is obtained:

$$D_e^r \dot{e} + W_1^r \dot{q}_m e + R_e^r e = \sigma_e, \quad (40)$$

with the input defined as:

$$\sigma_e = u^r - [D_e^r \ddot{q}_{ed}^r + W_1^r \dot{q}_m \dot{q}_{ed}^r + W_2^r \dot{q}_m + R_e^r \dot{q}_{ed}^r]. \quad (41)$$

For design purposes, the resulting term in $W_1^r \dot{q}_m$ requires accomplish with skew-symmetrical property, that is to say, a transposed matrix is equal than the original matrix ($A = -A^\top$). To accomplish this property, the matrix Z_1 is proposed as:

$$Z_1 = \begin{bmatrix} 0 & (L_d - L_q)n_p \dot{q}_m & 0 \\ 0 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix}. \quad (42)$$

Adding and subtracting Z_1 to the error dynamic equation, results:

$$D_e^r \dot{e} + [W_1^r \dot{q}_m - Z_1]e + [R_e^r + Z_1]e = \sigma_e. \quad (43)$$

Then:

$$\begin{aligned}C_e^r &= W_1^r \dot{q}_m - Z_1 = n_p \dot{q}_m \begin{bmatrix} 0 & L_q & 0 \\ -L_q & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \quad \text{and} \\ \bar{R}_e^r &= R_e^r + Z_1 = n_p \begin{bmatrix} r_e & (L_d - L_q)n_p \dot{q}_m & 0 \\ 0 & r_e & 0 \\ 0 & 0 & r_e \end{bmatrix},\end{aligned}\quad (44)$$

where C_e^r accomplish with skew-symmetry property. Under this new transformation, the dynamic error equation is:

$$D_e^r \dot{e} + C_e^r e + \bar{R}_e^r e = \sigma_e, \quad (45)$$

where $\sigma_e = u^r - [D_e^r \ddot{q}_{ed}^r + C_e^r \dot{q}_{ed}^r + W_2^r \dot{q}_m + \bar{R}_e^r \dot{q}_{ed}^r]$.

If $u^r = u_d^r$, then $\sigma_e = 0$, therefore the control signal in close-loop is:

$$u^r = D_e^r \ddot{q}_{ed}^r + C_e^r \dot{q}_{ed}^r + W_2^r \dot{q}_m + \bar{R}_e^r \dot{q}_{ed}^r. \quad (46)$$

To get a fast convergence, damping is added to the dynamic error equation, the product between a gain matrix K_1 and the current error e is included in both members of the equation:

$$D_e^r \dot{e} + C_e^r e + \bar{R}_e^r e + K_1 e = \sigma_e + K_1 e. \quad (47)$$

Under this damping injection, the nominal control law is:

$$u^r = D_e^r \ddot{q}_{ed}^r + C_e^r \dot{q}_{ed}^r + W_2^r \dot{q}_m + \bar{R}_e^r \dot{q}_{ed}^r - K_1 e. \quad (48)$$

The Lyapunov's stability analysis is used to find all elements of K_1 , the proposed candidate function is:

$$V = \frac{1}{2} e^\top D_e^r e. \quad (49)$$

V is positive definite because $D_e^r > 0$ and there is a quadratic product in error. The derivative of the candidate function is:

$$\dot{V} = e^\top D_e^r \dot{e} \quad (50)$$

From equation (47), when the input $\sigma_e + K_1 e = 0$, the $D_e^r \dot{e}$ term is:

$$D_e^r \dot{e} = -C_e^r e - [\bar{R}_e^r + K_1]e. \quad (51)$$

Then, replacing $D_e^r \dot{e}$ in (50), results:

$$\dot{V} = -e^\top C_e^r e - e^\top [\bar{R}_e^r + K_1]e = -e^\top [\bar{R}_e^r + K_1]e, \quad (52)$$

where $e^\top C_e^r e = 0$ because C_e^r is skew-symmetrical. $\dot{V}$ must forced to be a negative definite function, then $[\bar{R}_e^r + K_1]$ must be a positive definite. Therefore, K_1 is proposed as:

$$K_1 = \begin{bmatrix} k & -(L_d - L_q)n_p \dot{q}_m & 0 \\ 0 & k & 0 \\ 0 & 0 & k \end{bmatrix} \quad (53)$$

hence

$$\bar{R}_e^r + K_1 = \begin{bmatrix} r_e + k & 0 & 0 \\ 0 & r_e + k & 0 \\ 0 & 0 & r_e + k \end{bmatrix}, \quad k > 0. \quad (54)$$

This ensures that $\dot{V} < 0$.

B. Robust Control Design Based on Lyapunov's Redesign

Robust control is designed to deal with parametric uncertainty in the resistance of the stator windings [4]. The resistance can change due a lot of factor like temperature, age or inter-turn short-circuits. To cope with the effect of the resistance change, an additional term is designed and added to the nominal control law given in (48). To find this new term, we need the desired dynamic equation:

$$u_d^r = D_e^r \ddot{q}_{ed}^r + C_e^r \dot{q}_{ed}^r + W_2^r \dot{q}_m + \bar{R}_{ex}^r \dot{q}_{ed}^r, \quad (55)$$

where

$$\bar{R}_{ex}^r = R_{ex}^r + Z_1 = \begin{bmatrix} r_{11} & r_{12} + (L_d - L_q)n_p \dot{q}_m & r_{13} \\ r_{21} & r_{22} & r_{23} \\ r_{31} & r_{32} & r_{33} \end{bmatrix}. \quad (56)$$

Taking into account (37) and developing all terms in (55), the desired dynamic equation can be grouped together into:

$$u_d^{r0} = \begin{bmatrix} \frac{L_q 2\dot{\tau}_{emd}}{3n_p \lambda_m} + \lambda_m n_p \dot{q}_m \\ \frac{-\dot{q}_m L_q 2\tau_{emd}}{3\lambda_m} \\ 0 \end{bmatrix} + \begin{bmatrix} \frac{2}{3}\alpha & \frac{1}{6}\alpha & -\frac{1}{6}\alpha \\ 0 & \frac{1}{6}\alpha & -\frac{1}{6}\alpha \\ \frac{1}{3}\alpha & -\frac{1}{6}\alpha & -\frac{1}{6}\alpha \end{bmatrix} \begin{bmatrix} r_a^0 \\ r_b^0 \\ r_c^0 \end{bmatrix} = g_d^0 + B_d^0 \Phi^0, \quad (57)$$

where

$$\alpha = \frac{2\tau_{emd}}{3n_p \lambda_m}, \quad (58)$$

and g_d^0 is the vector that contains the independent terms to r_e without uncertainty, B_d^0 is a kind of regressor and contains terms that multiply r_e , Φ^0 contains the parameter with uncertainty. The superscript 0 indicates nominal value. The term $B_d^0 w$ is added to the nominal control law to compensate for the parameter uncertainty.:

$$u^r = u_d^{r0} - K_1 e + B_d^0 w. \quad (59)$$

The parametric error is defined as the difference between the nominal value and the actual value of r_e , that is to say,

$$\tilde{\Phi} = \Phi^0 - \Phi = \begin{bmatrix} r_a^0 \\ r_b^0 \\ r_c^0 \end{bmatrix} - \begin{bmatrix} r_a \\ r_b \\ r_c \end{bmatrix}, \quad (60)$$

and it must satisfy

$$\|\tilde{\Phi}\| \leq \rho_1, \quad \rho_1 \in \mathbb{R}^+. \quad (61)$$

ρ_1 is the maximum bound of the parameter error vector $\|\tilde{\Phi}\|$. With the Lyapunov's stability analysis w can be found, for this, $\pm B_d^0 \Phi$ is added to the equation (59) as:

$$u^r = g_d^0 + B_d^0 \Phi^0 - K_1 e + B_d^0 w \pm B_d^0 \Phi$$

$$u^r = (g_d^0 + B_d^0 \Phi) + B_d^0 (\tilde{\Phi} + w) - K_1 e \quad (62)$$

$$u^r - (g_d^0 + B_d^0 \Phi) = B_d^0 (\tilde{\Phi} + w) - K_1 e.$$

The term $u^r - (g_d^0 + B_d^0 \Phi)$ is the difference between real and desired behaviour, therefore, using (45):

$$D_e^r \dot{e} + C_e^r e + \bar{R}_e^r e = B_d^0 (\tilde{\Phi} + w) - K_1 e. \quad (63)$$

Again a candidate function is proposed to find w , this function is:

$$V = \frac{1}{2} e^\top D_e^r e. \quad (64)$$

It was shown before that V is positive definite. The derivative of the candidate function is:

$$\dot{V} = e^\top D_e^r \dot{e} \quad (65)$$

The term $D_e^r \dot{e}$ in equation (63) is replaced in the last equation, and this results in:

$$\dot{V} = -e^\top C_e^r e - e^\top [\bar{R}_e^r + K_1] e + e^\top B_d^0 (\tilde{\Phi} + w). \quad (66)$$

As mentioned before $-e^\top C_e^r e = 0$, also, it was shown that $e^\top [\bar{R}_e^r + K_1] e$ is a positive definite function. Now it only remains to show that $e^\top B_d^0 (\tilde{\Phi} + w)$ is a negative definite function or a bounded positive definite function. For this demonstration is proposed that:

$$w = -\frac{\rho_1^2 (B_d^0)^\top e}{\rho_1 \|e^\top B_d^0\| + \epsilon}. \quad (67)$$

Then

$$\begin{aligned} e^\top B_d^0 (\tilde{\Phi} + w) &= e^\top B_d^0 \left(\tilde{\Phi} - \frac{\rho_1^2 (B_d^0)^\top e}{\rho_1 \|e^\top B_d^0\| + \epsilon} \right) \\ &\leq \|e^\top B_d^0\| \left(\|\tilde{\Phi}\| - \frac{\rho_1^2 \|(B_d^0)^\top e\|}{\rho_1 \|e^\top B_d^0\| + \epsilon} \right) \\ &\leq \|e^\top B_d^0\| \left(\frac{\rho_1^2 \|e^\top B_d^0\| - \rho_1^2 \|(B_d^0)^\top e\| + \rho_1 \epsilon}{\rho_1 \|e^\top B_d^0\| + \epsilon} \right) \\ &= \frac{\rho_1 \|e^\top B_d^0\| \epsilon}{\rho_1 \|e^\top B_d^0\| + \epsilon} \\ &\leq \epsilon. \end{aligned} \quad (68)$$

Therefore, $\dot{V}$ is:

$$\dot{V} \leq -e^\top [\bar{R}_e^r + K_1] e + \epsilon, \quad (69)$$

and it can say that the robust control law (59) ensures a bounded error of speed and currents [9].

IV. SIMULATION

The AC brushless motor parameters (B26S model, HDT manufacturer), sine PWM parameters and EV parameters are listed in the table (I). The desired speed $\dot{q}_{md}$ is generated by a Bézier polynomials taking into account a simple driving cycle, acceleration $\ddot{q}_{md}$ and acceleration derivative $\dddot{q}_{md}$ are generated by a derivation and double derivation of the desired speed, respectively, using *MATLAB/Simulink* derivation block.

In the simulation a state filter is used to obtain the derivative of τ_L . This filter has a state representation as:

$$\begin{aligned} \dot{x}_1 &= x_2, \\ \dot{x}_2 &= -(2\pi f)^2 x_1 - 2^{2/3} \pi f x_2 + (2\pi f)^2 \tau_L, \end{aligned} \quad (70)$$

TABLE I
SIMULATION PARAMETERS

Parameters	Magnitude	Parameters	Magnitude
f_s	5 KHz.	r_e	0.121 Ω
v_i	600 V.	L_{ls}	1×10^{-5} H.
λ_m	0.262 Vs/rad.	$\bar{L}_m$	8×10^{-4} H.
R_m	0.00001 Nm/s/rad.	$L_{\Delta m}$	0 H.
n_p	4	D_m	0.022 Kg/m ² .
m	1366 Kg	μ_{rr}	0.015
g	9.8 m/s ²	ρ	1.25 Kg/m ³
A	2.66 m ²	C_d	0.23
ψ	0 rad	G	5.5
r	0.2876 m	η_g	0.95

Fig. 6. Parametric Variations.

with $f = 45$ Hz. The filter input is the load torque given by (34) and the filter output are the states x_1 and x_2 which are the load torque and its derivative respectively. This outputs are used in equations (33) with the intention of saving computational efforts. The nominal controller gains are $\Gamma_m = 9000$ and $k = 20$. For the robust controller, the gains are also $\Gamma_m = 9000$ and $k = 20$ plus $\rho_1 = 0.121$ and $\epsilon = 0.01$.

To probe the robust controller, in the simulations abrupt changes in the winding resistances were considered to evaluate the performance of nominal and robust controllers under unbalanced operating conditions. It is assumed single phase variation in the resistance value. The figure 6 shows the variation in the nominal value of the resistance in $t = 15$ sec. Changes of 50%, 150% and 200% are considered.

On the other hand, to evaluate the controllers performance beyond the variations on the stator resistances, changes in ψ were considered. This variable represents the angle of the surface in which the vehicle moves and is related directly to mechanical parameters like the mass m of the car and gravity g . The figure 7 shows the variation in ψ , it is clear that until the instant $t = 23$ sec the car moves over a zero-slope surface, then the next 5 sec the vehicle moves over a road with angle ψ different of zero (ramp-road). At the instant $t = 28$ sec the ramp road is no more, so the angle ψ returns to zero.

V. RESULTS

A. Under Resistances Variations

Several simulations were performed, the table II shows the performance indices (Integral of Absolute Error IAE and the Integral of Time multiplied by Absolute Error ITAE) of the nominal and robust controllers for speed tracking under variations in single phase resistance.

In speed control performance, nominal controller shows a lower indices than robust. Here is necessary to say that robust control design was made to deal with parametric uncertainty in stator resistance, this involves the electrical subsystem directly.

Fig. 7. Variations of ψ .

TABLE II
NONLINEAR CONTROLLERS PERFORMANCE

Controller	Resistance	Variation	IAE	ITAE
Nominal	r_a, r_b, r_c	50%	1.132	17.63
Robust	r_a, r_b, r_c	50%	1.17	18.26
Nominal	r_a, r_b, r_c	150%	1.133	17.66
Robust	r_a, r_b, r_c	150%	1.175	18.36
Nominal	r_a, r_b, r_c	200%	1.133	17.65
Robust	r_a, r_b, r_c	200%	1.171	18.27

The next figures shows the results of both nominal and robust controllers under a 200% variation in r_c in $t = 15$ sec. The desired, nominal and robust speed trajectories are shown in figure 8. The maximum amplitude of desired trajectory is $265.6 \text{ rad/sec} = 50 \text{ Km/hr}$. It is clearly that a 200% variation in r_c do not affect the speed tracking. Figure 9 shows the electromagnetic torque tracking for both cases, the nominal case presents bigger maximum in $t = 27$ sec than the robust case. The advantage of robust control over the nominal is visible in the regulation of $\dot{q}_2^r$ and $\dot{q}_3^r$, this currents are desired equal to zero. Robust case presents lower currents than nominal case (figures 10 and 11). Lower amplitude currents extend the motor life-time because the heating is less. The table III clearly indicates that robust controller has lower indices for currents regulation than nominal controller.

B. Under Changes on ψ

The table IV shows the performance indices (Integral of Absolute Error IAE and the Integral of Time multiplied by Absolute Error ITAE) of the nominal and robust controllers for speed tracking under variations in the angle ψ . Also in this table, it is clearly that in speed tracking the nominal controller is slightly better than robust controller but, this slightly difference is not significant, we have to remember that EV's speed control application does not need a high level of accuracy. It is important to say now that we are interested to reduce the currents amplitude under resistance uncertainty and therefore extend the motor lifetime. The figure 12 shows the speed error for both controllers under a 20° variation in ψ . The amplitude of this error signals is short compared with the reference in figure 8. The both nominal and robust mechanical subsystems controllers are designed in the same manner, so there is no visible difference between these two signals. Because in this case resistance variations equals to zero, both nominal and robust produces the same currents in the motor stator. The figure 13 shows the current i_a produced due to both controllers, also it is clearly that in the instant $t = [23 \ 28]$ sec the current increased because ψ changed.

TABLE III
ITAE FOR i_d AND i_0 REGULATION

Controller/current	i_d	i_0
Nominal	2017	2643
Robust	1719	2107

Fig. 8. Angular speed tracking.

TABLE IV
NONLINEAR CONTROLLERS PERFORMANCE UNDER VARIATIONS IN ψ

Controller	Angle variation	IAE	ITAE
Nominal	$\psi = 10^\circ$	1.147	17.62
Robust	$\psi = 10^\circ$	1.148	17.65
Nominal	$\psi = 15^\circ$	1.132	17.29
Robust	$\psi = 15^\circ$	1.134	17.27
Nominal	$\psi = 20^\circ$	1.117	16.87
Robust	$\psi = 20^\circ$	1.119	16.91

VI. CONCLUSIONS

In this paper a EV model is presented. The use of sine PWM technique and power inverter in simulations allows results approximate to reality. An AC brushless motor (also called permanent magnet synchronous motor) model is presented in E-L formulation for two reference frames. Also a mechanical dynamic model for the EV is presented. A robust nonlinear control for AC brushless motors considering parameter uncertainty was presented. The Lyapunov redesign technique was used to obtain an additional feedback signal that is added to the nominal control law to guarantee the uniform ultimate stability in the presence of uncertainties. The parameters of the AC BM need not be precisely known a priori, only the uncertainty bound ρ_1 is needed to derive the compensation signals. The proposed control law is able to deal with the uncertainties in the winding resistances. The simulation results show that a good tracking behavior is obtained.

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Fig. 9. Electromagnetic torque tracking.

Fig. 10. $i_d = q_d^*$ current regulation.

Fig. 11. $i_q = q_q^*$ current regulation.

Fig. 12. Speed error for both nominal and robust controllers.

Fig. 13. Current i_a for both cases nominal and robust controllers.

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