Research Article

Robust Multiobjective Model Predictive Control with Computation Delay Compensation for Electric Vehicle Applications Using PMSM with Multilevel Inverter

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The Three-Level Neutral-Point-Clamped (3L-NPC) inverter fed Permanent Magnet Synchronous Motor (PMSM) drive is an attractive configuration for high performance Electric Vehicle (EV) applications. For such configuration, due to their high performances, the Finite-Control-Set Model Predictive Control (FCS-MPC) is a very attractive control solution. The FCS-MPC scheme is based on the prediction of the future behavior of the controlled variables using the dynamic model of PMSM and the discrete nature of the 3L-NPC inverter. However, the parametric uncertainties and time-varying parameters affect the FCS-MPC algorithm performances. In this paper, robust FCS-MPC controls based on “dynamic error correction” (DEC) and “modified revised prediction” (MRP) are proposed to improve the FCS-MPC robustness without affecting the controller performances and complexity. The proposed strategies are improved also by multiobjective (MO) algorithm optimization and computation delay compensation. The simulation results included prove the performance in robustness and efficiency of the proposed robust FCS-MPC-DEC.

1. Introduction

The Electric Vehicle (EV) has received more attention as an alternative solution of energy crisis and environmental challenges [1]. In the next years, with the research efforts and the constant improvement of the electrical drive systems and control strategies, the penetration of EVs will increase exponentially [2]. The electrical drive part of EVs is constituted by electric machine, power converter, and drive train [3]. Currently, a great variety of electrical drive technologies can be used on EVs: two/multilevel power converter and induction machines (IM)/Permanent Magnet Synchronous Motor (PMSM) electric machine. For the electric machine part, the PMSM is considered as the potential candidates for EVs applications with potential high energy density, high efficiency, easier control scheme, robust operations, and

reduced maintenance [4]. Generally, the two-level inverters are widely used to control the PMSM in EVs applications [5]. To deal with this structure, several control strategies are proposed: linear controls with the modulator (e.g., field-oriented control and direct torque control) and nonlinear controls without the modulator (e.g., switching table based) [6]. In different aspects, other classification include adaptive control (AC), sliding mode control (SMC), and active disturbance rejection control (ADRC) [7–10]. In the recent years, many Model Predictive Control (MPC) approaches are developed to improve the control performances for the power converter applications [11]. Based on the mathematical system model, MPC predict the future behavior of the system. The optimal control is obtained using predefined optimization criteria. The MPC presents several advantages: simple to apply in variety of processes and multiple-input-

multiple-output (MIMO) systems, the system constraints and nonlinearities are considered in the control system. Furthermore, several control loops can be incorporated into one control law. Particularly, the Finite-Control-Set Model Predictive Control (FCS-MPC) is considered as attractive alternative predictive control [12]. Taking into account the discrete nature of the power converter, FCS-MPC uses the possible output voltage vectors to formulate the control algorithm and not require any external modulator stage. Furthermore, FCS-MPC provides good dynamic-state and steady-state performances [13]. In EVs' applications, research efforts have been made to improve the efficiency of FCS-MPC algorithms. Due to the parameter variation and road conditions, the EVs' process is considered as a time-varying system. The dynamic behavior of EVs, model uncertainties and inexactitude, unmodelled dynamics (motor and inverter), and unmeasurable parameters lead to a mismatch between the actual parameter values and their nominal values used in the prediction model and degrade the control robustness. There are many research studies accorded to robustness of the MPC controller [14]. A novel-simplified MPC scheme combined with the Lyapunov-based concept for interior PMSM is proposed in [15] to eliminate the flux-current mapping from the MPC formulation. Stumper et al. [15] propose a robust extension for predictive torque control (PTC) to improve the steady-state performances against parametric uncertainties for PMSM based on two-degree-of-freedom (2DoF) control. Then, the switching frequency and the ripples are eliminated without affecting the dynamics-state performances. In [16], the authors design a new discrete-time robust predictive current controller for PMSM. Using the static error correction (SEC) principle, the robustness of predictive current control of PMSM against the effect of the inaccurate of motor behavior is improved and implemented by a DSP model TMS320F2812 practically [17]. Nevertheless, the two-level inverters limit the set of control in eight possible combinations which lead to high output current/voltage total harmonic distortion (THD) and high switching losses. Furthermore, a high dv/dt derivative causes high stress in the semiconductor devices. To overcome these challenges, in the last few years, several multilevel inverter structures are recognized as a novel approach of high power converter performances [18]. Among them, the three-level neutral-point-clamped (3L-NPC) is the most interesting multilevel converter topology due to economy, complexity of control, and performances [19]. However, the computational burden will increase exponentially with the level of the power converter. This issue is addressed in [20] and the computational burden is reduced with 12.31%. The robustness of the FCS-MPC control for the multilevel converter fed PMSM high performances (e.g., EVs' applications) is still an open research topic. In [6], based on one-horizon simple revised prediction (SRP), the system robustness against parameters variations is improved by reducing the control variables ripples. The SRP principle has been implemented on a fully FPGA-based platform (NI-cRIO 9082 system). Typically, the existing research works are based on mono-objective problem optimization, without computational delay compensation, and adopted for two-

level power converters. The main contributions in this work are as follows:

- (i) Improved the predictive control robustness against parameters variation and uncertainties using "dynamic error correction" (DEC) and "modified revised prediction" (MRP) principles without affecting the system complexity and control dynamic-state performances.
- (ii) A multiobjective problem optimization: current tracking control, neutral-point voltage control, common-mode voltage control, and reduction of switching frequency. These control objectives are controlled using single loop and one control action without internal/external loops and additional tables.

The main abovementioned contributions are adopted by the multilevel power converters structures with computational delay compensation based on the proposed modified two-step horizon of prediction approach.

The remainder of the paper is structured as follows. The problem statement is given in Section 2. The three principle elements in the MPC strategies are discussed. The prediction model is stated in Section 3. Cost function and optimization algorithm are illustrated in Section 4. In Section 5, the robustness of the proposed algorithms is verified by simulation using different scenarios.

2. Problem Statement

The MPC strategy is based on the explicit use of the system model to select the optimal action $s_{abc}^{\text{opt}} = \{s_a, s_b, s_c\}$. Then, the performances of the MPC controller depend on the accuracy of the system model and its parameters. In this section, the effects of the variable system uncertainties, computational delay, and standard FCS-MPC principle are presented.

2.1. Impact of Variable System Uncertainties. The MPC controller established the optimal control action based on information about the dynamic behavior of the system. The system performances are affected with the accuracy of the information about the inductance L_{dq} (time-varying variable) and the permanent magnet flux linkage ψ_{pm} (unmeasurable variable). A variation in ψ_{pm} value only affects the feed-through matrix, $H_d(k)$; on the contrary, a variation in L_{dq} value affects both the system and control matrices, $A_d(k)$ and B_d (equation (15)).

2.2. Impact of Computational Delay. In the ideal case, due to negligibility of the calculation time, the optimal control action s_{abc}^{opt} that minimizes the cost function f at instant $k+1$ is applied at time k . Consequently, the controlled variable reaches their predicted value at $k+1$ time. On the contrary, in the real case, a significant calculation time t_k is required between the instants k when the system state $x_m(k)$ is measured and the generation of the optimal action s_{abc}^{opt} at $k+t_k$. The computational delay affects the system

performances and lead to the oscillation of the controlled variables around their references values [21].

2.3. Standard FCS-MPC. In this work, four control objectives (COs) are considered:

- (i) (CO1) Fast and accurate reference current tracking to reduce the mechanical component stress of system and acoustic pollution
- (ii) (CO2) Capacitor voltage balancing between upper and lower DC-link capacitors to reduce the sensibility of output inverter voltage to the ripples in DC-link voltage
- (iii) (CO3) Limit the peak of common-mode voltage to reduce the overvoltage stress and increase the safety and reliability of electrical machine part of EVs
- (iv) (CO4) Minimize the switching frequency to reduce the switching losses

These control objectives are incorporated in a single cost function and can be designed as follows:

$$f = f_i + f_{DC} + f_{cmv} + f_{Nc}, \quad (1)$$

$$f_i = \lambda_i |i^{\text{ref}}(k+1) - i^p(k+1)|, \quad (2)$$

$$f_{DC} = \lambda_{DC} |u_z^p(k+1)|, \quad (3)$$

$$f_{cmv} = \lambda_{cmv} |u_{cmv}^p(k+1)|, \quad (4)$$

$$f_{Nc} = \lambda_{Nc} N_c, \quad (5)$$

where f_i , f_{DC} , f_{cmv} , and f_{Nc} are the cost functions for current tracking control, neutral-point voltage control, common-mode voltage control, and reduction of switching frequency, respectively. $i^{\text{ref}}(k+1)$ and $i^p(k+1)$ represent the reference and predicted currents at instant $k+1$, respectively. $u_z^p(k+1)$ and $u_{cmv}^p(k+1)$ denote predicted neutral point and common-mode voltages, respectively. λ_i , λ_{DC} , λ_{cmv} , and λ_{Nc} are the weightings factors. It is worth mentioning that the weighting factors values have a direct influence on the controlled variables. The good performances of multiobjective optimization-based predictive control are guaranteed by the choice of the optimal weighting factors. In this work, the weighting factors λ_i , λ_{DC} , λ_{cmv} , and λ_{Nc} are tuned using “trial-and-error” manner.

In the cost function (5), N_c penalize the switching state variation as follows:

$$N_c = \sum_{x=a,b,c} |s_x(k+1) - s_x(k)|. \quad (6)$$

3. System Dynamic

3.1. PMSM. The dynamic model of PMSM in the d - q rotating reference frame is given by the differential equation [22]:

$$\frac{di_d}{dt} = -\frac{R_s}{L_d} i_d(t) + \frac{L_q}{L_d} \omega_e(t) i_q(t) + \frac{1}{L_d} u_s^d(t), \quad (7)$$

$$\frac{di_q}{dt} = -\frac{R_s}{L_q} i_q(t) + \frac{L_d}{L_q} \omega_e(t) i_d(t) - \frac{1}{L_q} \omega_e(t) \psi_{pm} + \frac{1}{L_d} u_s^q(t), \quad (8)$$

$$\frac{\omega_e(t)}{dt} = \frac{Z_p}{J_m} T_e - \frac{B_v}{J_m} \omega_e(t) - \frac{Z_p}{J_m} T_L, \quad (9)$$

where u_s^{dq} is stator dq -voltages, V; i^{dq} is stator dq -currents, A; R_s is stator winding resistance, Ω ; L_d and L_q are stator dq -inductances, H; ω_e is electric rotatory frequency, rad/s; ψ_{pm} is permanent magnet flux, Wb; T_e and T_L electromagnetic and load torques, Nm; Z_p is the number of pole pairs; J_m is moment of inertia, $\text{kg} \cdot \text{m}^2$; B_v is friction coefficient, Nm.s; and T_s sampling interval, s. The electromagnetic torque of PMSM is obtained by [23]

$$T_e = \frac{3}{2} Z_p (\psi_{pm} i_q(t) + (L_d - L_q) i_d(t) i_q(t)). \quad (10)$$

Remark 1. Due to the uniform air-gap in surface mounted PMSM (SPMSM), the stator dq -inductance are identical to each other: $L_d = L_q = L_s$. Opposed to SPMSM, the internal PMSM (IPMSM) is specified by the nonuniform air-gap flux. Thus, the stator q -inductance L_q is smaller than stator d -inductance L_d . The approximation $L_d = L_q = L_s$ of SPMSM is considered in the rest of the paper.

Using the dq -current dynamic and the switching state of 3L-NPC inverter, the neutral-point voltage dynamic is [24]

$$\frac{du_z}{dt} = \frac{1}{4C} ((2|s_a| - |s_b| - |s_c|) i_d + \sqrt{3}(|s_b| - |s_c|) i_q), \quad (11)$$

where $C = C_1 = C_2$ is the DC-link capacitor (Figure 1(a)).

3.2. 3L-NPC Inverter. The 3L-NPC inverter consists of three identical branches, each branch is composed with four switches and can be described by three switching states [1], [0], and [-1] (Table 1). Thus, there are 27 possible combinations of switching states corresponding to 19 inverter output voltages obtained in $\alpha\beta$ -coordinate as follows [20]:

$$u_s^{\alpha\beta} = U_{DC} ((2s_a - s_b - s_c) + j\sqrt{3}(s_b - s_c)), \quad (12)$$

where U_{DC} denotes the DC-link voltage.

In abc coordinate, the common-mode voltage u_{CMV} generated by the 3L-NPC inverter can be presented by [24]

$$u_{CMV} = \frac{u_{AZ} - u_{BZ} - u_{CZ}}{3}. \quad (13)$$

u_{AZ} is defined as the voltage between the phase (A) and the neutral-point voltage (Z) (Figure 1(a)). Table 2 shows the switching states and corresponding u_{CMV} [25]. Simplified electrical circuit of 3L-NPC inverter-fed PMSM drives are presented in Figure 1(a).

FIGURE 1: (a) Simplified electrical circuit of the system (b) Block diagram of standard FCS-MPC-ICF control scheme.

TABLE 1: Operating principle of 3L-NPC inverter.

Switching state	Switches				One-phase inverter voltage
$s_x, x \in \{a, b, c\}$	s_{x1}	s_{x2}	s_{x3}	s_{x4}	$u_{xz}, x \in \{A, B, C\}$
1	ON	ON	OFF	OFF	$U_{dc}/2$
0	ON	OFF	OFF	ON	0
-1	OFF	OFF	ON	ON	$-U_{dc}/2$

TABLE 2: u_{CMV} corresponding for every switching state combination.

Switching state combination	u_{cmv}
111	$U_{dc}/2$
-1-1-1	$-U_{dc}/2$
110, 011, 101	$U_{dc}/3$
0-1-1, -10-1, -1-10	$-U_{dc}/3$
100, 010, 001, 11-1, -111, 1-11	$U_{dc}/6$
00-1, -100, 0-10, 1-1-1, -11-1, -1-11	$-U_{dc}/6$
000, 10-1, 01-1, -110, -101 0-11, 1-10	0

4. FCS-MPCs Algorithms

4.1. Standard FCS-MPC with Improved Cost Function (FCS-MPC-ICF). Considering a sampling period T_s , the discrete state space model of PMSM is obtained by employing the forward Euler method for equations (7) and (8):

$$x_m(k+1) = A_d(k)x_m(k) + B_d u_m(k) + H_d(k), \quad (14)$$

where $x_m(k) = [i_m^d \ i_m^q]^T$ and $u_m(k) = [u_s^d \ u_s^q]^T$. $A_d(k)$, B_d , and $H_d(k)$ are the system, input, and feed-through matrices. They are given by

$$\begin{aligned} A_d(k) &= \begin{bmatrix} 1 - \frac{R_s T_s}{L_s} & T_s \omega_e(k) \\ -T_s \omega_e(k) & 1 - \frac{R_s T_s}{L_s} \end{bmatrix}, \\ B_d &= \begin{bmatrix} \frac{T_s}{L_s} & 0 \\ 0 & \frac{T_s}{L_s} \end{bmatrix}, \\ H_d(k) &= \begin{bmatrix} 0 \\ -\frac{T_s \psi_{pm}(k)}{L_s} \omega_e(k) \end{bmatrix}. \end{aligned} \quad (15)$$

The predictive reference current is calculated using the Lagrange method [20]:

$$i^{ref}(k+1) = 3i^{ref}(k) - 3i^{ref}(k-1) + i^{ref}(k-2). \quad (16)$$

The evolution of neutral-point voltage is defined as follows:

$$u_z^p(k+1) = u_z(k) + \frac{T_s}{4C} (2|s_a(k)| - |s_b(k)| - |s_c(k)|)i_\alpha(k) + \frac{\sqrt{3}T_s}{4C} (2|s_b(k)| - |s_c(k)|)i_\beta(k). \quad (17)$$

To compensate the computational delay, according to long horizon of prediction approach, we propose to use two-step horizon of prediction. In the 3L-NPC inverter case, combination of 27 switching states are available, a discrete set of $27^2 = 729$ possible trajectories of the switching states is required to evaluate the cost function led to a large number of cost computations which makes the real challenge to the actual digital system implementation. To reduce the number of real time evaluation, this paper proposes to use a modified two-step horizon prediction by applying the same voltage vector in the two-step horizon. The improved cost function f^{dcom} is evaluated at time $k+2$ to generate the optimal switching state at time $k+1$:

$$f^{\text{dcom}} = f_i^{\text{dcom}} + f_{DC}^{\text{dcom}} + f_{cmv}^{\text{dcom}} + f_{Nc}^{\text{dcom}}, \quad (18)$$

$$f_i^{\text{dcom}} = \lambda_i |i^{\text{ref}}(k+1) - i^p(k+1)|, \quad (19)$$

$$f_{DC}^{\text{dcom}} = \lambda_{DC} |u_z^p(k+1)|, \quad (20)$$

$$f_{cmv}^{\text{dcom}} = \lambda_{cmv} |u_{cmv}^p(k+1)|, \quad (21)$$

$$f_{Nc}^{\text{dcom}} = \lambda_{Nc} N_c^{\text{dcom}}, \quad (22)$$

$$N_c^{\text{dcom}} = \sum_{x=a,b,c} |s_x(k+2) - s_x(k)|. \quad (23)$$

The reference current $i^{\text{ref}}(k+2)$ is extrapolated by using the second-order Lagrange extrapolation [20]:

$$i^{\text{ref}}(k+2) = 6i^{\text{ref}}(k) - 8i^{\text{ref}}(k-1) + 3i^{\text{ref}}(k-2). \quad (24)$$

The evolution of x_m and u_z^p at instant $k+2$ are expressed based on equations (14) and (17):

$$x_m(k+2) = A_d(k+1)x_m(k+1) + B_d u_m(k+1) + H_d v(k+1), \quad (25)$$

$$u_z^p(k+2) = u_z(k+1) + \frac{T_s}{4C} (2|s_a(k+1)| - |s_b(k+1)| - |s_c(k+1)|)i_\alpha(k+1) + \frac{\sqrt{3}T_s}{4C} (2|s_b(k+1)| - |s_c(k+1)|)i_\beta(k+1). \quad (26)$$

In order to guarantee the stability, a Lyapunov candidate function (LCF) can be used to ensure that the stator current

i_{dq} are maintained according to their reference value i_{dq}^{ref} as follows [26]:

$$V(\tilde{i}_d, \tilde{i}_q) = \frac{1}{2} \Upsilon_d \tilde{i}_d^2 + \frac{1}{2} \Upsilon_q \tilde{i}_q^2. \quad (27)$$

With

$$\Delta V = V(\tilde{i}_d^p(k+1), \tilde{i}_q^p(k+1)) - V(\tilde{i}_d(k), \tilde{i}_q(k)), \quad (28)$$

where Υ_d and Υ_q are the positive gains, $\tilde{i}_{dq} = [\tilde{i}_d, \tilde{i}_q]$ is the state error of the stator current defined as follows:

$$\tilde{i}_{dq} = \begin{bmatrix} \tilde{i}_d^m(k) \\ \tilde{i}_q^m(k) \end{bmatrix} - \begin{bmatrix} i_d^{\text{ref}}(k) \\ i_q^{\text{ref}}(k) \end{bmatrix}. \quad (29)$$

The optimal switching state s_{abc}^{opt} is given by the following equation:

$$s_{abc}^{\text{opt}} = \arg \left\{ \min_{s_{abc}^{\text{opt}}(k) \in \{-1, 0, 1\}^3, s_{abc}^{\text{opt}}(k+1) \in \{-1, 0, 1\}^3} f^{\text{dcom}} \right\},$$

subject to $((V(k) > \varrho_v) \& (\Delta V < 0)) \vee ((V(k) < \varrho_v) \& (V(k+1) < \varrho_v))$
 $s_{abc}(k) = s_{abc}(k+1).$ (30)

With the aim to ensure the stability of the closed loop, two cases are considered [26]: the first one corresponds to $V(k) > \varrho_v$ and the second one $V(k) < \varrho_v$. If $V(k) > \varrho_v$, only switches obtaining $\Delta V < 0$ are considered and if $V(k) < \varrho_v$, only switches assuring $V(k+1) < \varrho_v$ are considered. Hence, ϱ_v can be used as a stabilizing constraint, which guarantees feasibility and stability for the optimization problem.

Any parameters variation or uncertainties compensation is included in the standard FCS-MPC-ICF algorithm control formulation. A simplified block diagram of standard FCS-MPC-ICF is presented in Figure 1(b). It is worth to be mentioned that the “error correction” and “revised prediction” principles are based on the same approach: modified prediction equation to take into account the time-varying behavior of the system. EC is based on the contribution of the controlled parameter error in the previous sampling period in the formulation of optimization problem. On the contrary, RP improves the robustness against the variations of controlled variables taking their different roles into account. These principles are developed in this work based on “dynamic error correction” and “modified revised prediction,” and they are implemented separately in robust FCS-MPC-DEC and robust FCS-MPC-MRP controls. In both proposed robust predictive controls, the parameters variation and uncertainties are explicitly included in the algorithms formulation.

4.2. Robust FCS-MPC-DEC. The future behavior of the controlled system is predicted using equation (12) for the 27 possible combinations in the 3L-NPC inverter for each prediction horizon. The accuracy of the prediction is directly affected by the quality of the parameter information. The EC

FIGURE 2: Behavior of standard FCS-MPC-ICF at normal conditions. (a) The stator current i_a . (b) The output voltage U_a . (c) Fast Fourier Transform of the current. (d) Fast Fourier Transform of the output voltage. (e) Performance of the neutral-point voltage U_z . (f) Performance of common-mode voltage U_{cmv} .

is based on the contribution of the weighted error between the predicted and measured controlled variable values in the previous sampling time $\eta_{dq}(k-1)$, in the prediction equation. In [15], the author used the previous current error without considering the optimal switching state combination applied in the next sampling interval. Due to the variation of the switching state applied in every sampling time, the direct consideration of the previous error is not suitable (Figure 2). Siami et al. [17] proposed to use a static weighted error for each switching state combination $j(j=1, \dots, 8)$. In this study, we proposed to use a dynamic DEC for every given switching combinations $j(j=1, \dots, 27)$. The weighted error is calculated for each possible switching state combination. The modified predicted state is

$$i_{dq,j}^{p,mod}(k+1) = i_{dq,j}^p(k+1) + \sigma_1 \eta_{dq,j}(k), \quad (31)$$

where σ_1 is a weighting factor between 0 and 1 and $\eta_{dq,j}$ is the discrete current error defined as follows:

$$\eta_{dq,j}(k) = \begin{bmatrix} i_{d,j}^p(k) \\ i_{q,j}^p(k) \end{bmatrix} - \begin{bmatrix} i_{d,j}^m(k) \\ i_{q,j}^m(k) \end{bmatrix}. \quad (32)$$

4.3. Robust FCS-MPC-MRP

4.3.1. Robustness against L_s Variation. To consider the mismatch between the actual value of the inductance and

their nominal value used in the controller, the system state is modified as follows:

$$x_m^{\text{mod}}(k) = \sigma_2 x_m(k) + (1 - \sigma_2) x_p(k), \quad (33)$$

where x_m and x_p are the measured and predicted values of the system state, respectively, σ_2 denotes a weighting factor between 0 and 1 according to the BIBO stability criteria (in Appendix). The tunable parameter will add a “weighted average” process to the measured and predicted values of the system. Therefore, the modified state of x_m^{mod} , which takes the benefits of this (“weighted average” process), will also add a smoothing and filtering effect to the system states whenever with or without parameter variations. Consequently, improved control variable quality is predictable.

4.3.2. Robustness against ψ_{pm} Variation. Introducing $L_s = L_{s0} + \Delta L_s$ and $\psi_{pm} = \psi_{pm0} + \Delta \psi_{pm}$, where L_{s0} and ψ_{pm0} are the nominal values and ΔL_s and $\Delta \psi_{pm}$ are the parametric uncertainties, and using equation (14), the relationship between the measured and the predicted current values is

$$i_q^m(k) = \frac{L_{s0}}{L_s} i_q^m(k) + \frac{\Delta L_s T_s \omega_e}{L_s} (k-1) i_d^p(k-1) + \frac{1}{L_s} T_s \omega_e (k-1) \Delta \psi_{pm}. \quad (34)$$

In the steady-state SPMSM operation ($i_d = 0$), the error between the predicted and measured q -axis current is

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Read  $i_{dq}^{\text{ref}}(k)$  and extrapolate  $i_{dq}^{\text{ref}}(k+1)$  and  $i_{dq}^{\text{ref}}(k+2)$  by using equations (16) and (24);
Measure  $i_{dq}^m(k)$ ,  $\omega_e(k)$  and  $u_z(k)$ ;
% Initialize the values of  $s_{abc}^{\text{opt}}$  and  $f^{\text{dcom}}$ 
 $s_{abc}^{\text{opt}} = 1$ ;  $f^{\text{dcom}} = \text{zeros}(1: 27)$ ;
For  $i=1$  to 27 do
  Calculate  $u_s^{ab}$  based on equation (12);
  Predict  $i_{dq}^p(k+1)$  and  $i_{dq}^p(k+2)$  from equations (14) and (25);
  Evaluate the candidate Lyapunov function:  $V(k+1)$ ,  $V(k)$ ,  $\Delta V$  using equations (27) and (28);
  if  $((V(k) > \epsilon_v) \& (\Delta V < 0)) \vee ((V(k) < \epsilon_v) \& (V(k+1) < \epsilon_v))$  then
    Predict  $u_z^p(k+1)$  and  $u_z^p(k+2)$  based on equations (17) and (26);
    Calculate  $u_{cmv}^p(i)$  from Table 2;
    Calculate  $N_c^{\text{dcom}}$  from equation (23);
    Compute the cost functions  $f_i^{\text{dcom}}$ ,  $f_{DC}^{\text{dcom}}$ ,  $f_{cmv}^{\text{dcom}}$  and  $f_{Nc}^{\text{dcom}}$  based on equations (19)–(22);
    Evaluate the cost function  $f^{\text{dcom}}$  by using equation (18);
  End
End
Select and applied the optimal switching state  $s_{abc}^{\text{opt}} = \min(f^{\text{dcom}})$ ;

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ALGORITHM 1: Standard FCS-MPC-ICF.

$$\eta_q^m(k) = i_q^m(k) - i_q^p(k) \approx \frac{1}{L_s} T_s \omega_e(k-1) \Delta \psi_{pm}. \quad (35)$$

The permanent magnet flux uncertainties will contribute into the current error. To compensate the effect of the permanent magnet flux uncertainties into the equation prediction equation (35), a modified permanent magnet flux is proposed as follows:

$$\psi_{pm}^{\text{mod}}(k) = \psi_{pm}(k) + \sigma_3 \sum (\eta_q^m(k)), \quad (36)$$

where σ_3 is the weighted factor.

The proposed control algorithm for standard FCS-MPC-ICF, robust FCS-MPC-DEC, and robust FCS-MPC-MRP are described in Algorithms 1–3, respectively. Algorithm 1 shows the detailed instructions, equations, and parameters used in the simulation of standard FCS-MPC-ICF. The algorithms 1 and 2 can be summarized in the following steps: measurements, calculation, and extrapolation of references and prediction of control variables and minimization of the improved cost function. Both Algorithms 3 and 4 illustrated the specific instructions and parameters used in proposed robust FCS-MPC-DEC and robust FCS-MPC-MRP (e.g., modified system state and current error vector).

Remark 2. The factors σ_1 , σ_2 , and σ_3 are tuned in a “trial-and-error” manner.

5. Results and Analysis

In this section, the control performances of proposed robust FCS-MPC-DEC and robust FCS-MPC-MRP for 3L-NPC inverter fed PMSM for EVs’ applications are validated by simulations and compared with the control performance of

standard FCS-MPC-ICF under different scenarios both with and without parameters variations and uncertainties. Using Matlab/Simulink software and SimPowerSystems toolbox, the simulations are carried with the parameters, as indicated in Table 3.

With the aim to study better the robustness of the proposed robust FCS-MPCs controls, we proposed to use the two performances variables.

- (i) The mean absolute reference tracking error [21]:

$$\bar{e}_{\text{MARTE}}(x) = \frac{1}{N} \sum_{i=1}^N |x_{\text{ref}}(i) - x_{\text{mes}}(i)|. \quad (37)$$

N denotes the number of simulation iterations. x_{ref} and x_{mes} are the reference and measured controlled variables x , respectively.

- (ii) The average switching number per semiconductor:

$$\bar{f}_{\text{ASN}} = \frac{1}{N} \sum_{k=1}^N \frac{n_{ij}(k)}{12}, \quad (38)$$

where n_{ij} is the switching numbers of power semiconductors of phase $i \in \{a, b, c\}$ and order $j \in \{1, 2, 3, 4\}$ defined as follows:

$$n_{ij} = |s_{ij}(k+1) - s_{ij}(k)|. \quad (39)$$

The behavior of the standard FCS-MPC-ICF at the nominal conditions is illustrated in Figure 3. In the test, the good dynamic-state performances in fast and accurate reference current tracking are demonstrated (Figure 3(a)). The steady-state performances of the FCS-MPC-ICF are validated with low mean absolute reference tracking error and THD of current 0.0091 and 3.29 (Figure 3(c)), respectively. The output voltage is shown in Figure 3(b) with THD of

```

% Initialize the values of  $s_{abc}^{\text{opt}}$ ,  $f^{\text{dcom}}$  and current error  $\eta_{dq,j}$ ;
 $s_{abc}^{\text{opt}} = 1$ ;  $f^{\text{dcom}} = \text{zeros}(1:27)$ ;  $\eta_{(dq,j)} = \text{zeros}(2:27)$ ;
For  $i = 1$  to 27 do
    Calculate the current error  $\eta_{(dq,j)(:,i)}$  based on equation (32);
    Predict the modified  $i_{dq,j}^{p,\text{mod}}(k+1)$  and  $i_{dq,j}^{p,\text{mod}}(k+2)$  using the current
    error  $\eta_{dq,j}(:,i)$  and equations (31) and (25);
End
Select and applied the optimal switching state  $[\eta_{(dq,j)} s_{abc}^{\text{opt}}] = \min(f^{\text{dcom}})$ ;

```

ALGORITHM 2: Robust FCS-MPC-DEC.

```

% Initialize the values of  $s_{abc}^{\text{opt}}$  and  $f^{\text{dcom}}$ 
 $s_{abc}^{\text{opt}} = 1$ ;  $f^{\text{dcom}} = \text{zeros}(1:27)$ ;
For  $i = 1$  to 27 do
    Predict the modified  $x_m^{\text{mod}}(k+1)$  and  $x_m^{\text{mod}}(k+2)$  using equation (33) and (25);
End
Select and applied the optimal switching state  $s_{abc}^{\text{opt}} = \min(f^{\text{dcom}})$ ;

```

ALGORITHM 3: Robust FCS-MPC-MRP against L_s variation.

```

% Initialize the values of  $s_{abc}^{\text{opt}}$ ,  $f^{\text{dcom}}$  and  $\sum \eta_q^m$ 
 $s_{abc}^{\text{opt}} = 1$ ;  $f^{\text{dcom}} = \text{zeros}(1:27)$ ;  $\sum \eta_q^m = 0$ ;
For  $i = 1$  to 27 do
    Calculate the current error  $\eta_q^m$  and the summation  $\sum \eta_q^m$  based on equation (36);
    Calculate the modified permanent magnet flux  $\psi_{pm}^{\text{mod}}$  using the equation (37);
End

```

ALGORITHM 4: Robust FCS-MPC-MRP against ψ_{pm} variation.

TABLE 3: System parameters.

Parameter	Value
PMSM power rate P_N (W)	1440
PMSM pole pairs Z_p [1]	2
PMSM resistance R_s [Ω]	2.89
PMSM inductance L_s (mH)	7
PM flux ψ (Wb)	0.125
Moment of inertia J_m [kg.m^2]	10^{-5}
Friction coefficient B_v (Nm.s)	11.10^{-5}
DC capacitance $C_1 = C_2 = C$ (μF)	1

22.79 (Figure 3(d)). The average switching number per semiconductor $\bar{f}_{\text{ASN}}$ is 9834. As shown in Figure 3(e), the deviation of DC-link voltage u_z is maintained around 6.66% of its nominal value. Figure 3(f) shows the evolution of u_{cmv} and the overvoltage stresses of PMSM are reduced to an average value of 0.2.

The evolution of current error $\eta_{dq}(k)$ in function of applied optimal switching state s_{abc}^{opt} is illustrated in Figure 2.

The application of dynamic DEC is motivated by the dynamic behavior of η_{dq} and its not constant for same s_{abc}^{opt} . Consequently, the proposition of fixed error values of each switching combination considered in static error correction is incorrect.

Figure 4 and Table 4 evaluates the robustness of standard FCS-MPC-ICF and proposed robust FCS-MPC-DEC and FCS-MPC-MRP controls against parameter system variations. The inductance L_s and the permanent magnet flux ψ_{pm} are varied to -50% and 200% of their nominal values. The performances of standard FCS-MPC-ICF are affected in presence of variations of L_s and ψ_{pm} , and the error and controlled variable ripples around their nominal values are highly increased. These control ripples run to the instability in case of high parameters variations (in case of 200% of parameter variation). In case of proposed robust FCS-MPC-DEC and FCS-MPC-MRP, the robustness against parameter variation is improved. The proposed robust FCS-MPC-DEC solution outperforms both the proposed robust FCS-MPC-MRP and standard FCS-MPC-ICF controls in terms of

FIGURE 3: The evolution of current error η_{dq} of standard FCS-MPC-ICF. (a) s_{abc}^{opt} . (b) The current error η_d . (c) The current error η_q .

FIGURE 4: Continued.

FIGURE 4: The evolution of current error η_{dq} : (a) $0.5L_s$, (b) $2L_s$, (c) $0.5\psi_{mg}$, and (d) $2\psi_{mg}$; coral: standard FCS-MP-ICF, red: robust FCS-MPC-DEC, and blue: robust FCS-MPC-MRP.

TABLE 4: Comparison of the control performances under different controls.

Parameter	Variation	Control	$\bar{e}_{\text{MARTE}}(i_{dq})$	u_{cmv}	u_z (%)	$\bar{f}_{ASN}$	t_c	THD i_{dq}	u_s
ΔL_s	-50%	Standard FCS-MPC-ICF	0.4892	0.76	7.05	$1.60 \cdot 10^4$	21.96	4.63	29.06
		Robust FCS-MPC-DEC	0.0098	0.23	6.68	$9.83 \cdot 10^3$	14.70	3.68	27.58
		Robust FCS-MPC-MRP	0.0100	0.25	6.79	$1.22 \cdot 10^4$	18.80	3.73	28.26
	200%	Standard FCS-MPC-ICF	0.5740	4.28	8.31	$1.85 \cdot 10^4$	39.41	6.35	36.16
		Robust FCS-MPC-DEC	0.1297	0.75	7.16	$4.19 \cdot 10^3$	31.57	4.68	35.23
		Robust FCS-MPC-MRP	0.1306	0.76	7.33	$7.71 \cdot 10^3$	33.20	4.71	35.31
$\Delta \psi_{pm}$	-50%	Standard FCS-MPC-ICF	0.3199	0.75	6.95	$2.28 \cdot 10^4$	29.47	4.07	28.32
		Robust FCS-MPC-DEC	0.0124	0.28	6.81	$1.94 \cdot 10^3$	25.66	3.45	25.70
		Robust FCS-MPC-MRP	0.0136	0.38	6.95	$1.94 \cdot 10^3$	27.59	3.59	26.07
	200%	Standard FCS-MPC-ICF	0.5037	2.43	8.23	$2.31 \cdot 10^4$	47.80	6.01	70.84
		Robust FCS-MPC-DEC	0.0940	0.50	7.06	$9.83 \cdot 10^3$	34.21	4.45	32.55
		Robust FCS-MPC-MRP	0.1010	0.68	7.23	$1.18 \cdot 10^4$	36.69	4.51	32.71

smaller $\bar{e}_{\text{MARTE}}(i_{dq})$, u_{CMV} , u_z , $\bar{f}_{ASN}$, calculation time t_c , and THD of current and output inverter voltage.

6. Conclusion

A robust model predictive control for multilevel inverter fed PMSM for electrical vehicle application is proposed in this paper. The computational delay is compensated using a proposed modified two-step horizon prediction. A proposed improved multiobjective cost function f^{dcom} incorporates the control objectives and evaluates every sampling interval to generate the optimal switching state applied directly to the inverter without modulation stage. The time-varying behavior of the electrical vehicle system affects the good dynamic and steady-state performances of the standard finite-control-set model predictive control with improved cost function (FCS-MPC-ICF) control. Simulation results under different scenarios of parameters variation approve the effectiveness of the proposed robust FCS-MPC with “dynamic error correction” (DEC) and “modified revised prediction” (MRP) principles in terms of control robustness. The proposed robust FCS-MPC-DEC solution outperforms both the proposed robust FCS-MPC-MRP and standard FCS-MPC-ICF controls.

Appendix

Bounded-Input Bounded-Output (BIBO) Stability Criteria

With the assumption that the future “predicted” value will approach its “reference” $x_p(k+1) = x^{\text{ref}}(k+1)$ [27], the discrete transfer function between the measurement and the reference using equations (29) and (14) is obtained:

$$\frac{x_m}{x^{\text{ref}}} = \frac{(L_{s0}/L_s)z + (1 - (L_{s0}/L_s))\sigma_1}{(z - (1 - (L_{s0}/L_s))\sigma_1)} \stackrel{:=}{=} F(z). \quad (\text{A.1})$$

Based on “bounded-input bounded-output (BIBO)” stability criteria, the system is stable, if and only if all roots of $F(z) = z - 1 + \sigma_1(L_{s0}/L_s)$ are inside the unit circle, i.e.,

$$|z| = \left| \left(1 - \frac{L_{s0}}{L_s} \right) \sigma_1 \right| < 1 \iff 0 < \frac{L_{s0}}{L_s} < 1 + \underbrace{\frac{1}{\sigma_1}}_{\chi_{\max}}, \text{ where, } 0 < \sigma_1 \leq 1. \quad (\text{A.2})$$

The stability range of inductance value L_s in the controller is enlarged from the original/conventional solution $(0; 2L_s)$ to $(0; \chi_{\max}L_s)$.

In the original/conventional solution (i.e., the classical FCS-MPC solution), $\sigma_1 = 1$. So, the characteristic function becomes $F(z) = z - 1 + \sigma_1 (L_{s0}/L_s)$. Therefore, the BIBO stable requirement is $0 < (L_{s0}/L_s) < 2$.

The stability of a sampled closed loop system, with the discrete-time format of $C(z)/R(z) = G(z)/(1 - GH(z))$, can be determined by the location of its closed loop poles in z -plane, which are the roots of the characteristic equation $1 + GH(z) = 0$ as follows:

- (i) **System will be stable** if the closed loop poles or the roots of the characteristic equation will lie within the unit circle of $|z| = 1$ in the z -plane
- (ii) **System will be marginally stable** if a pole or a pair of complex conjugate poles lie(s) on the circle of $|z| = 1$
- (iii) **System will be unstable** in other cases

Data Availability

No data were used to support this study.

Conflicts of Interest

The authors declare that there are no conflicts of interest regarding the publication of this paper.

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