Robust Model Predictive Control for AFE-Inverter Drives With Common Mode Voltage Elimination

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ABSTRACT This paper proposes a novel and robust version of Model Predictive Control scheme for AC drives based on Voltage Source Inverter (VSI) with Active Front End (AFE). The main feature of the proposed MPC is elimination of Common Mode Voltage (CMV) without imposing a penalty on the corresponding term in the cost function, but rather by a smart utilisation of the restricted set of switching states in a computationally efficient algorithm. Furthermore, the paper proposes to split the conventional MPC scheme into separate Control and Modulation stages, and to enhance the Control stage by integral action, and the Modulation stage - by a Feedback Quantizer. The resulting AC drive scheme provides high tracking accuracy over the full speed range, robustness to disturbances and parameters error, coupled with practically zero CMV and consequently - very low levels of conducted and radiated electromagnetic interference (EMI). This makes the proposed scheme a competitive alternative to the existing AC drive solutions in the most challenging industrial applications. The benefits of the proposed scheme are validated by simulation and experiment.

INDEX TERMS Predictive control, MPC, common mode voltage, cost function, feedback quantizer, AC motor drive, back-to-back converter, active front end, induction motor.

I. INTRODUCTION
AC drives used in various industrial environments need to meet a host of performance expectations. These typically include excellent tracking and dynamic performance, high tolerance to parameter uncertainty and disturbances, and low noise emissions. In some environments, minimization of the Common Mode Voltage (CMV) can be also critical [1].

Problems associated with CMV include: (a) common mode currents due to parasitic capacitive coupling between motor neutral and earth [2]; (b) EMI with other equipment within the same power system or neighborhood [3]; (c) extra cost and weight associated with EMI filters [4]; (d) generation of shaft voltages that are destructive to motor bearings [5]; and (e) safety issues in some industrial environments [1].

Existing CMV mitigation solutions can be classified into those based on common mode filters; and those based on modified modulation strategies. Traditional solutions based on passive common mode filters [6], [7] are easy-to-use but bulky. Their effectiveness reduces when using power converters based on fast-switching devices [8]. Active common mode filters explored in [9], [10] have size advantages but add extra complexity. Both strategies add to power losses and cost.

As an alternative, modulation based CMV solutions have been proposed for various converter topologies, including two-level [11], [12], multilevel [13], [14], modular multilevel [15] and matrix [16] converters. The literature shows that some degree of CMV reduction is achievable for practically every converter topology [17]. However, complete elimination of CMV is only possible for converter topologies that have sufficient redundancy of switching states.

Model Predictive Control (MPC) can be seen as a special type of modulation-based solution [18]. CMV mitigation can be included in MPC by adding a corresponding term to the cost function [19]; by employing sequential predictive control principle [20]; or by excluding the switching states with higher CMV from the Finite Control Set (FCS).
The current paper focuses on a back-to-back converter topology comprising a voltage source inverter and an active front end (AFE). Its suitability for CMV-critical industrial environments has recently been investigated in [21]. An AC drive based on an AFE-inverter is shown in Fig. 1. The total CMV generated by the AFE-inverter can be calculated as [22]:

$$V_{cm} = \frac{V_{dc}}{3} (S_a + S_b + S_c - S_a - S_v - S_w)$$  \hspace{1cm} (1)$$

where $S_a$, $S_b$ and $S_c$ are switching functions of the inverter; and $S_v$, $S_s$ and $S_w$ are switching functions of the AFE.

CMV levels corresponding to different AFE and inverter states are given in Table 1. It follows from Table 1 that it is possible to completely eliminate the total CMV if: (a) any odd index vector on one side is combined with any odd index vector on the other side; (b) any even index vector on one side is combined with any even index vector on the other side; (c) the same zero vector (either $v_0$ or $v_7$) is applied on both sides.

To maintain the CMV at zero, switching events at both sides should be synchronized. Whenever a switching state changes on one side, an appropriate change should occur on the other, leading to one of the diagonal combinations listed in Table 1. A strategy for synchronization of AFE and inverter PWM sequences has been proposed in [22]. FCS-MPC options with reduced, or zero, CMV have been explored in [23], [24].

The current paper makes a further contribution to these developments. Its major contribution is to develop a high performance AC drive based on AFE-inverter topology, with CMV elimination. AC drive control is based on the MPC principle, with embellishments, including integral action and Feedback Quantizer. This leads to superior tracking, robustness and low harmonics. An important aspect of the proposed solution is minimal number of terms in the MPC cost function. In AC drive applications, weighting of such terms is complicated due to variable operating conditions. Instead, it is proposed to address different control objective at different levels, in a decoupled manner. In this sense, the proposed control can be termed as “decoupled MPC”. The resulting AC drive reliably operates at any speed, has low sensitivity to parameter errors and disturbances, and produces essentially zero CMV.

The remainder of the paper is organized as follows. Section II presents a critical evaluation of the MPC current control approach when applied to AC drives. Section III describes various improvements, which make MPC competitive in AC drive applications. Section IV explains the proposed CMV elimination strategy and illustrates its effectiveness by simulation results. Section V describes the experimental setup used in this study and provides experimental results. Finally, conclusions are presented in Section VI.

II. MPC CURRENT CONTROL FOR AN AC DRIVE

Application of traditional MPC current control [25] to an AFE-inverter drive is shown in Fig. 2 and is described below.

On one side, the drive interacts with an induction motor. A traditional PI speed control scheme produces current references $i_{ds}$ and $i_{qs}$. To match these, the induction motor model is described in a rotating $(dq)$ frame aligned with the rotor flux linkage [26]:

$$\begin{align*}
V_{ds} &= R_s i_{ds} + \sigma L_s \frac{di_{ds}}{dt} - \omega_s \omega_s L_d i_{qs} \\
V_{qs} &= R_s i_{qs} + \sigma L_s \frac{di_{qs}}{dt} + \omega_s \omega_s L_d i_{ds}
\end{align*}$$  \hspace{1cm} (2)$$

where standard notation has been used for stator voltages ($V_{ds}$, $V_{qs}$), stator currents ($i_{ds}$, $i_{qs}$), stator resistance ($R_s$), synchronous frequency $\omega_s$, and inductances: stator ($L_s$), magnetizing ($L_m$) and rotor ($L_r$). Leakage constant $\sigma = 1 - \frac{L_m}{L_r}$, and rotor flux linkage $\psi_r = L_m i_{ds}$ (in steady state).

For the AFE side, similar equations can be derived in the $dq$-frame aligned with the grid voltage $V_g$:

$$\begin{align*}
V_{df} &= R_f i_{df} + L_f \frac{di_{df}}{dt} - \omega_s L_f i_{qf} + V_{gd} \\
V_{qf} &= R_f i_{qf} + L_f \frac{di_{qf}}{dt} + \omega_s L_f i_{df}
\end{align*}$$  \hspace{1cm} (3)$$

where the subscripts “$f$” and “$g$” stand for “filter” and “grid,” and $V_{gq} = 0$ due to the assumed frame alignment. Both models (2) and (3) have the same form:

$$V_{dq} = R_i i_{dq} + \frac{di_{dq}}{dt} + E_{dq}$$  \hspace{1cm} (4)$$

where $E_{dq}$ is the back-emf (or voltage on the other side of the RL-load), including $dq$-axes coupling terms.

It is assumed that back-emfs are compensated separately via a feedforward decoupling circuit. This leads to very simple RL-plant model, decoupled between $d$- and $q$-axes. The plant can be controlled on each axis by using, for example, a typical FCS-MPC current control, implemented in a stationary $(\alpha\beta)$ frame [25]. Such a control scheme is based on a discrete-time

![Figure 1. An AC drive based on inverter with AFE.](image)
model, calculated separately for $\alpha$- and $\beta$-axes:

$$i_{j,\alpha\beta}(k+1) = a_i\alpha\beta(k) + b_j\alpha\beta(k)$$  \hspace{1cm} (5)

where $v_j$ are voltage choices from FCS ($j = 0 \ldots n - 1$); $a = e^{-\frac{RT}{V}}$; $b = \frac{1}{T_s} \left( 1 - e^{-\frac{RT}{V}} \right)$; and $T_s$ is sampling interval.

Predicted values, $\hat{i}_{j,\alpha\beta}(k+1)$, are then compared to the corresponding reference currents $i_{\alpha\beta}^*(k+1)$. The following cost function is used for each $v_{j,\alpha\beta}$:

$$g_j = (\hat{i}_{ja}(k+1) - i_{a}^*(k+1))^2 + (\hat{i}_{jb}(k+1) - i_{b}^*(k+1))^2$$  \hspace{1cm} (6)

Finally, $v_{j,\alpha\beta}^{opt}$, which achieves the minimum value of the cost function (6), is selected and applied.

For RL plant model with decoupled $\alpha\beta$-components, prediction horizon 1, and quadratic cost function (6), an alternative and fully equivalent way to implement FCS-MPC is to calculate the reference voltage $v_{\alpha\beta}^*(k)$ by inverting the model (5) as:

$$v_{\alpha\beta}^*(k) = \frac{1}{b} \left( i_{\alpha\beta}(k+1) - ai_{\alpha\beta}(k) \right)$$  \hspace{1cm} (7)

and then to compare $v^*(k)$ to the available states using the following cost function:

$$g'_j = (v_{ja}(k) - v_{a}^*(k))^2 + (v_{jb}(k) - v_{b}^*(k))^2$$  \hspace{1cm} (8)

Equivalence of the cost function forms (6) and (8) can be readily proven by expressing $i_{\alpha\beta}^*(k+1)$ from (7) and substituting, together with the prediction (5), into (6), yielding

$$g_j = (bv_{ja}(k) - bv_{a}^*(k))^2 + (bv_{jb}(k) - bv_{b}^*(k))^2$$  \hspace{1cm} (9)

This shows that the two cost functions (6) and (8) relate via a constant coefficient: $g_j = b^2 g'_j$. Therefore, using the current-based cost function (6) or the voltage-based cost function (8) at every sampling instant, will result in identical choices for $v_{j,\alpha\beta}^{opt}$. To maintain this equivalence when other cost terms are added, weighting coefficient $b^2$ may need to be applied.

In Fig. 2, application of the alternative cost function $g_j'$ is assumed, due to its computational efficiency [24] and transparency. The result is a choice, from the finite set, of the voltage that is nearest (in the Euclidean distance sense) to the desired voltage $v^*(k)$. Therefore, application of the cost function $g_j'$ given by (8) can be interpreted as a Nearest Neighbour Quantizer (NNQ).

Cost functions in the form (8) would normally be evaluated independently for the motor and the AFE sides. Then the optimal switching state would be applied at each side independently. However, if CMV minimization is desired, then synchronization between the two sides is necessary. This can be achieved by using the following combined cost function:

$$g_{cm}^{total} = \lambda \left( v_{\alpha,j}(k) - V_{\alpha,j}^*(k) \right)^2 + \lambda \left( v_{\beta,j}(k) - V_{\beta,j}^*(k) \right)^2 + (v_{\alpha,j}(k) - V_{\alpha,j}^*(k))^2 + (v_{\beta,j}(k) - V_{\beta,j}^*(k))^2 + \lambda_{cm} V_{cm,j,i}^2$$  \hspace{1cm} (10)

where $\lambda$ is weighting coefficient for AFE side voltage; and $\lambda_{cm}$ is weighting coefficient for CMV.

As explained earlier in relation to (9), to have equal weights on the motor and AFE side currents, coefficient $\lambda$ in (10) has to be selected as $\lambda = (b_f/b_s)^2$, where $b_f$ and $b_s$ are the $b$-coefficients in the AFE and motor side models (5). CMV reduction can be controlled by choosing an appropriate value of $\lambda_{cm}$. If $\lambda_{cm} = \infty$ then CMV is completely eliminated.

Although the traditional scheme shown in Fig. 2 achieves its goals, it has certain drawbacks. Firstly, Fig. 2 reveals that the conventional FCS-MPC current control does not possess integral action. If the model parameters $a$ and $b$ in (5) are not precisely known, then a steady state current tracking error will result. Without integral action, rejection of disturbances, such as residual back-emf errors, may not be effective. The lack of
integral action is a known issue in MPC [27], [28]. In drive applications, this issue may become critical.

Back-emf compensation in AC drives can also be a source of another problem. Under rated motor conditions, back-emf is a major contributor to the motor side voltage $V_{s,\alpha\beta}$. When a sinusoidal reference is quantized by the NNQ, a "six-step" voltage waveform results. In the described scheme, back-emf (that is practically sinusoidal) is injected inside the closed-loop current control. This leads to both the motor-side voltage reference $V_{s,\alpha\beta}^*$ and the resulting voltage $V_{s,\alpha\beta}$ attaining a step-wise shape, as illustrated by experimental plots shown in Fig. 4. The corresponding motor currents are highly distorted.

This problem can be addressed by using a different method of back-emf compensation. For example, in Field Oriented MPC presented in [20], [29], the back-emf is implicitly compensated by using stator and/or rotor flux predictive models. However, this adds to the MPC implementation complexity and increases its sensitivity to modeling errors. The intent of this work is to retain the simplicity of feedforward compensation but to avoid its negative effects. Therefore, the basic Model Predictive current control with back-emf feedforward compensation is used in this paper as the base case.

III. PROPOSED FINITE SET FEEDBACK QUANTIZER MODULATION (FBQM) FOR AC DRIVES

To address the problems described above, the current paper proposes several enhancements to the traditional MPC-based current control. These changes lead to a robust AC drive scheme, with MPC as its core principle, but without the associated drawbacks. A block diagram of the proposed scheme is shown in Fig. 3. Its main features are described below.

A. INTEGRAL ACTION

Integral action is introduced in the proposed scheme by generating the voltage references $V_{s,\alpha\beta}^*$ using an optimal, in the MPC sense, current controller. This controller replaces the model inversion as per (7). The differences can be appreciated by comparing the MPC blocks in Fig. 3 and the Current Control Loop blocks in Fig. 2. In Fig. 2 the optimal MPC current control blocks are denoted as "Optimal PI". The reason for that becomes clear from the design procedure presented below.

The design is performed in the generalised MPC-based framework introduced in [30]. To drive tracking error to zero and reject disturbances, the linear system given by (4)–(5) is complemented by a disturbance model. It then corresponds to the following general description, in state space form:

$$
x_{k+1} = A x_k + B u_k + n_k; \quad d_{k+1} = A_d d_k + \omega_k; \quad y_k = C x_k + C_d d_k + \nu_k
$$

where $x$, $u$ and $y$ are state, input and output vectors, respectively; $n$, $\omega$ and $\nu$ are Gaussian white noise sequences; $d$ is a periodic disturbance; $A$, $A_d$, $C$, $C_d$ and $B$ are matrices of appropriate dimensions. One can then design a corresponding steady state Kalman observer, in polynomial form, as [31]:

$$
A(z)D(z)y_k = B(z)D(z)u_k + E(z)\epsilon_k
$$

where $A(z) = \det(zI - A)$; $D(z) = \det(zI - A_d)$; $B(z) = \frac{B(z)}{A(z)} = C(zI - A)^{-1}B$; $\epsilon$ is a white noise sequence with zero mean;
and $E(z)$ includes optimal gains which can be found using standard Kalman filtering theory.

Interpreting this result in the context of the given plant model, it can be shown that $A(z) = 1 - az^{-1}$ and $B(z) = bz^{-1}$. Polynomial $D(z)$ is nulling operator for the disturbance $d$. In the $dq$-frame, tracking errors and major disturbances (those related to parameter errors, cross-coupling, back-emfs) appear as constants. For constant disturbance, $D(z) = 1 - z^{-1}$. One practical choice of Kalman filter gains yields $E(z) = 1 - az^{-1}$. Then, using standard polynomial manipulations, equation (13) can be expressed as:

$$y_k = z^{-1}y_k + \frac{bz^{-1}(1 - z^{-1})}{1 - az^{-1}}u_k + \varepsilon_k$$

(14)

By advancing both sides of (14) by one step, a one-step ahead prediction of the output $y_k$ can be obtained as:

$$\hat{y}_{k+1} = y_k + b\left[1 - \left(1 - a\right)z^{-1}\right]u_k$$

(15)

In (15) prediction of white noise $\varepsilon_k$ is set to its mean ($= 0$), and the term associated with $u_k$ is split into two terms. Next, a standard MPC quadratic cost function for the tracking error is applied:

$$J = \left[\hat{y}_{k+1} - y^*\right]^2$$

(16)

The minimal cost value of (16), $J_{\text{min}} = 0$, clearly corresponds to $\hat{y}_{k+1} = y^*$. Note that $z^{-1}$ term on the right-hand side of (15) corresponds to the control, $u_{k-1}$, that has already been applied and is known. The only unknown is $u_k$, i.e. the control to be applied. The unconstrained optimal control, which brings cost (16) to $J_{\text{min}} = 0$, can be determined from (15) as:

$$u_k^{\text{unc}} = \frac{1}{b}(y^* - y_k) + \frac{1 - a}{1 - az^{-1}}u_{k-1}$$

(17)

Finally, applying the constraint on the control set, leads to the following constrained optimal policy: $u_k^{\text{con}} = \left[\text{Nearest } u_k \text{ to } u_k^{\text{unc}}\right]$. In (17), the last term corresponds to the actually applied (constrained) past control, i.e. $u_{k-1} = u_k^{\text{con}}$. The control law (17) is illustrated in Fig. 5. It is readily recognizable as a discrete-time PI control in anti-windup form with gain $K_p = \frac{1}{b}$ and time constant $\tau_i$ such that $a = e^{-\frac{\tau_i}{\tau}}$.

To summarize, the control law described by (17) has been designed using a one-step ahead prediction, cost function minimization, and applying an arbitrary (convex or finite set) constraint on the control set. Therefore, the anti-windup representation of a PI controller with the following parameters:

$$K_p^{\text{opt}} = \frac{1}{b} = R\left(1 - e^{-\frac{\tau_i}{\tau}}\right)^{-1} \approx \frac{L}{\tau_i}; \quad \tau_i^{\text{opt}} = \frac{L}{R}$$

is, indeed, optimal control in the MPC sense.

The design described above can be applied to the motor side (where $L = \sigma L_s$, $R = R_s$) and to the AFE side (where $L = L_f$, $R = R_f$). In a practical drive, considering a one-control-cycle implementation delay, a modified optimal PI control with transport delay compensation may be preferred [30].

B. BACK-EMF COMPENSATION

Back-emf compensation in the proposed scheme is achieved in exactly the same manner as in traditional FOC [26]. That is, the cross-coupling terms $E_{ds} = -\omega_{syn}\sigma L_s i_{qf}^d$ and $E_{df} = \omega_{syn}\left(\frac{L_f}{L_s}\psi_r + \sigma L_s i_{qf}^d\right)$ are added to the motor side current control output (see green block in Fig. 3), while the steady state rotor flux linkage is assumed constant: $\psi_r = L_s i_{qf}^d$. Similarly, the cross-coupling terms $E_{df} = -\omega_{qf} L_f i_{qf}^d + V_{pd}$ and $E_{qf} = \omega_{qf} L_f i_{qf}^d$ are added to the AFE side current control output (see grey block in Fig. 3).

This is different to (and much simpler than) the practices adopted in other FOC-MPC schemes [20], [29], where back-emf compensation involves stator and rotor flux models, while the rotor flux is assumed to be time varying.

C. FEEDBACK QUANTIZER

To overcome the problem of the “six-step” voltage distortions and improve the overall harmonic performance, the traditional Nearest Neighbor Quantizer (NNQ) is replaced by a Feedback Quantizer (FBQ). This is achieved by feeding back the voltage quantization error $q_e(z)$ to the NNQ input. The difference between NNQ and FBQ can be seen by comparing Fig. 6(a) and Fig. 6(b).

FBQ has been first introduced by the authors in [32]. FBQ, in its simplest form, uses a feedback filter $H(z) = z^{-1}$. In this case, the quantization error $q_e$, calculated in the previous step, is simply subtracted from the voltage reference $v^*$ in the current step. It can be easily shown that the output (quantized) voltage of the FBQ is then given by:

$$v^0(k) = v^*(k) - q_e(k) = v^*(k) - q_e(k - 1) + q_e(k)$$

$$= v^*(z) + S(z) q_e(z)$$

(18)

where $S(z) = 1 - z^{-1}$ is the corresponding noise sensitivity function that shapes the FBQ quantization noise. Compared to the NNQ output noise $q_e(z)$, the spectrum of the FBQ output noise $S(z) q_e(z)$ is shifted towards higher frequencies. Also, according to (18), the voltage reference $v^*(z)$ is reproduced at the FBQ output with gain 1.
The main effect of replacing NNQ with FBQ is that the feedback of the voltage error leads to the quantizer output be PWM-like rather than step-like. Additionally, by using a specially designed feedback filter, $H(z)$, it is further possible to shape the spectrum of the converter quantisation noise in a desired manner [33].

### D. CMV MINIMIZATION

One prominent advantage of using the cost function $g_i''$, as in (8), is its computational efficiency. Indeed, once $v_{αβ}^a(k)$ is known, it is possible to determine which sector it belongs to. The three voltage vectors in the corners of this sector are the only three voltages to be considered for the purpose of the cost function minimization. This reduces the number of associated comparisons, from 7 to 3, on each side of the AFE-inverter.

However, if the objective of CMV minimization is added, by using the $\Delta V_{in}$ term in cost function (10), this computational advantage can no longer be exploited. It is unknown which combination of motor side voltage $v_{αβ,i}$ and AFE side voltage $v_{αβ,j}$ will minimize the total cost function (10). Hence an exhaustive comparison of all possible options is required. Zero vectors $v_{0}$ and $v_1$ need to be considered separately, due to their different effects on CMV. As a result, for the AFE-inverter drive, $8 \times 8 = 64$ full cost function evaluations of the type (10) need to be performed in each cycle.

In the proposed scheme, a different and very computationally efficient approach is used for the CMV minimization. Firstly, rather than minimizing the CMV, the scheme aims to completely eliminate it. This leaves only the 20 switching combinations in the Finite Control Set, that appear as diagonal elements in Table 1. This is a natural decision if the AC drive protection design relies on that nonzero CMV cannot be generated even occasionally, which is the case for industrial environments with IT-earthing system [1]. This corresponds to setting the CMV weight $\lambda_{CM}$ in (10) to infinity.

Degradation of the waveform quality, expected from limiting the switching choices, is mitigated, to a large extent, by using FBQ. FBQ produces the output voltage $v(z)$ that accurately follows its reference $v^*(z)$, even when using a very restricted set of the switching states.

The second salient feature of the proposed control algorithms is that, starting from the very first switching combination, the switching states always maintain synchronisation on both sides, so as to maintain zero CMV at all times.

This can be better explained using a specific example. Say that, at a sampling instant $k - 1$, the zero vectors $v_{mot}^{m00}$ and $v_{afe}^{a00}$ were applied to both sides (where superscripts “mot” and “afe” indicate “motor” and “AFE” side, respectively). At the next sampling instant $k$, the following sequence is performed.

**Step 1.** Voltage references $v_{αβ}^{mot}(k)$ and $v_{αβ}^{afe}(k)$ are separately determined by the respective current controllers.

**Step 2.** As per Fig. 6(b), the modified voltage references are then calculated as

\[
\begin{align*}
    v_{αβ}^{mot}(k) &= v_{αβ}^{mot}(k) - q_{αβ}^{mot}(k - 1) \\
    v_{αβ}^{afe}(k) &= v_{αβ}^{afe}(k) - q_{αβ}^{afe}(k - 1)
\end{align*}
\]

where $q_{αβ}^{mot}(k)$ and $q_{αβ}^{afe}(k)$ are the previous quantization errors on the respective side. The modified references, $v_{αβ}^{mot}$ and $v_{αβ}^{afe}$, will drive the respective NNQ on each side.

**Step 3.** On each side the corresponding voltage sectors are determined. Say that $v_{αβ}^{mot}$ belongs to sector $1$, then the three candidate voltage vectors are: $v_{αβ}^{mot}(000)$, $v_{αβ}^{mot}(001)$ and $v_{αβ}^{afe}(011)$. Say that $v_{αβ}^{afe}$ belongs to sector $4$, then the three candidate voltage vectors are: $v_{αβ}^{afe}(000)$, $v_{αβ}^{afe}(100)$ and $v_{αβ}^{afe}(110)$. Note that the order of the candidate vectors is such that the zero vector goes first, the vector with one on-state goes second, and the vector with two on-states goes last.

**Step 4.** On each side the cost functions are determined for the three respective candidate vectors:

\[
\begin{align*}
    g_i^{mot} &= (v_{ia}^{mot} - v_{ia}^{mot}(k))^2 + (v_{ib}^{mot}(k) - v_{ib}^{mot}(k))^2 \\
    g_i^{afe} &= (v_{ia}^{afe} - v_{ia}^{afe}(k))^2 + (v_{ib}^{afe}(k) - v_{ib}^{afe}(k))^2
\end{align*}
\]

where $i = 0, 1, 2$. In this example, it follows that $v_{1}^{mot} \in \{v_{0}^{mot}, v_{1}^{mot}, v_{2}^{mot}\}$ and $v_{1}^{afe} \in \{v_{0}^{afe}, v_{1}^{afe}, v_{2}^{afe}\}$.

**Step 5.** The total cost function is calculated, based on weighting the motor and AFE side costs, e.g.:

\[
g_i^{total} = \lambda g_i^{afe} + (1 - \lambda) g_i^{mot}
\]

Calculations according to (21) are performed for the three respective combinations: $g_{0}^{mot}$ with $g_{1}^{afe}$, $g_{2}^{mot}$ with $g_{1}^{afe}$, and $g_{2}^{afe}$ with $g_{0}^{afe}$. Note that, for each combination, the two sides have either: the same zero vectors (e.g. 000 with 000), vectors with one on-state (e.g. 001 with 100), or vectors with two on-states (e.g. 011 with 110). According to Table 1, in all such cases the total CMV is zero.

**Step 6.** The three cost function values are compared, and the pair of voltage vectors, $v_{mot}^{min}$ and $v_{afe}^{min}$, that deliver the minimum cost, is selected and applied. From the above description it follows that zero CMV continues to be maintained.

In order to minimise switching transitions and to achieve equal utilisation of $v_{0}(000)$ and $v_{1}(111)$, on average, on both sides, the following logic is applied.

**Step 7.** If at step 6 the best pair of vectors, delivering minimal cost function, are zero vectors, i.e. $v_{mot}^{min} = v_{afe}^{min} = 000$, then the previous switching state is considered. If the previously applied vectors had 0 or 1 on-state (e.g. 000 with 000, or 001 with 100) then $v_{0}(000)$ is applied on both sides. If the previously applied vectors had 2 or 3 on-states (e.g. 011 with 110, or 111 with 111) then $v_{1}(111)$ is applied on both sides.

The proposed algorithm maintains zero total CMV at all times. It requires only three separate cost calculations according to (20), and three total cost calculations according to (21),
TABLE 2. Parameters Used in Simulation and Experiment

| Parameter          | Value  | Parameter          | Value  |
|--------------------|--------|--------------------|--------|
| Power $P_{\text{rated}}$ | 11kW   | $d$-axis current $i_{ds}$ | 10A    |
| Torque $T_{\text{rated}}$ | 60 Nm  | Sampling $T_s$ | 20 µs  |
| Speed $n_{\text{rated}}$ | 1465 rpm | Grid $V_{g, L_s}$, rms | 415 V  |
| Pole Pairs $p_p$ | 2      | Fundamental $f$ | 50 Hz  |
| Current $i_{\text{rated}}$ | 20 A   | Input Filter $R_f$ | 0.4 Ω  |
| Stator $R_s$ | 0.8 Ω  | Input Filter $L_f$ | 5 mH   |
| Leakage $L_{1s}$ | 3.9 mH | DC Link $V_{dc}$ | 720 V  |
| Rotor $R_r$ | 0.1 Ω  | DC Link $C_{dc}$ | 20 mF  |
| Leakage $L_{1r}$ | 3.9 mH | Inertia $J$ | 0.077 K gm² |
| Magnetising $L_m$ | 86.4 mH | Friction $B$ | 5.75 $\times$ 10⁻⁴ |

at each sampling instant. Therefore, it is highly computationally efficient.

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IV. SIMULATIONS

To test the proposed scheme, simulation models of the AFE-inverter AC drive with an induction motor have been developed. Simulation parameters, listed in Table 2, match the parameters of the experimental setup, described later in Section V. The two main points of interest are the performance comparison of the proposed FBQ-MPC scheme against its traditional alternative; and the performance comparison of the proposed FBQ-MPC with and without CMV elimination.

Fig. 7 compares transient performance of the three schemes, including: (a) standard MPC that uses model inversion; (b) the proposed FBQ-MPC; (c) the proposed FBQ-MPC with CMV elimination. In all three cases, the motor is expected to reach the rated speed, to hold it until $t = 0.8$ s and to brake to a full stop. At $t = 0.6$ s, an additional load of 10 Nm is applied. Intentional 30% errors are introduced in back-emf ($E$), grid voltage ($V_g$) and resistor ($R_s$, $R_i$) estimations.

Figs. 7(a)–7(c) compare dynamics of speed $\omega_m$, torque $T_e$ and DC-link voltage $V_{dc}$. In all three cases, the correct speed is reached due to the PI speed control action. However, under MPC (Fig. 7(a)), higher torque variation, torque ripple and DC-link voltage variation are observed. Figs. 7(d)–7(f) compare the corresponding voltages and currents, in the order: motor side line-to-line voltage $V_{ab}$ and line current $I_a$; AFE side line-to-line voltage $V_{uv}$ and line current $I_u$; and total CMV. A complete elimination of the CMV is evident in Fig. 7(f).

The same voltages and currents, under steady-state conditions, are compared in Figs. 8(a)–8(c). As discussed in Section II, the standard MPC in Fig. 7(d) shows voltage step changes and severe current distortion, on both sides. This issue is overcome with the proposed FBQ-MPC in Figs. 8(b)–8(c), where the currents on both sides are sinusoidal. Comparing Fig. 8(c) to Fig. 8(b) shows that CMV elimination with FBQ-MPC is achieved at the price of somewhat higher harmonics on both sides, which is an unavoidable consequence of the limitations in the FCS. The corresponding harmonic contents of the motor and AFE side currents are illustrated in Figs. 8(d)–8(f).

A summary Table 3 helps appreciate the associated performance trade-offs. Under FBQ-MPC, the average switching frequency is approximately 50% higher than under MPC. However, the resulting improvements in the current THDs are much more significant. When FBQ-MPC is complemented by CMV elimination, the current THDs increase by factor $\lesssim 2$. This is better than when using the standard RFOC with PWM [1], where this factor has been found to be $\lesssim 3$. Dependence of the performance on the operating conditions is illustrated by including the current THDs for $1/2$ rated speed and torque.
Additionally, by varying the weighting $\lambda$ in (21), an appropriate trade-off between the motor and AFE side current THDs can be determined. The trade-off curves corresponding to the proposed FBQ-MPC with CMV elimination, for the rated and $1/2$ rated conditions, are shown in Fig. 9. In the AC drive application targeted in the paper, an acceptable choice was $\lambda = 0.5 \ldots 0.7$. The grid-side current distortion can be further mitigated, if necessary, by using larger grid-side inductors.

In summary, the simulations have confirmed the capacity of the proposed FBQ-MPC to completely eliminate CMV,
while providing excellent tracking and acceptable level of harmonics.

V. EXPERIMENTAL RESULTS

The experimental setup used for validation of the proposed scheme, and equipment connections, are shown in Fig. 10. The drive control algorithms were implemented in C++ for a TMS320F28377D dual core processor. The AFE-inverter AC drive consists of two 20 kW Semikron IGBT stacks. The AC drive feeds an induction motor (parameters listed in Table 2) coupled to a DC motor of a similar size, on a test bed.

Fig. 11 illustrates transient performance of the proposed FBQ without (Fig. 11(a)) and with (Fig. 11(b)) CMV elimination. Transient plots are similar to those obtained by simulation (see Fig. 7), with slightly higher torque and DC-link voltage ripple for the CMV elimination option. With the standard MPC, it was not possible to bring the motor to the rated speed, hence no plots for the standard MPC are included.

Steady state performance of the three control schemes is illustrated by the experimental plots in Fig. 12. The bottom (brown) traces in Fig. 12 represent: at the motor side - CMV \( V_{cm} \), at the AFE side - voltage \( V_{NER} \) measured across a neutral-to-earth resistor (NER). Such a resistor, connected between the neutral point of the input side supply and earth, is required by the IT protection scheme used in industry [34]. In this study, \( V_{NER} \) serves as a measure of the conducted EMI caused by the drive. It is desired to minimise \( V_{NER} \).

Under the standard MPC, high distortions of the motor side and AFE side currents observed in Figs. 12(a) and 12(d), respectively, are consistent with the simulation results in Fig. 8. CMV elimination within this scheme was not pursued. Compared to MPC, the proposed FBQ-MPC delivers significant improvement of the motor side and the AFE side current quality, as shown in Fig. 12(b) and 12(e), respectively. Without mitigation, CMV observed in Fig. 12(b) reaches \( \pm V_{DC} \).

Application of the proposed CMV elimination strategy to FBQ-MPC is illustrated in Figs. 12(c) and 12(f). Voltages are more irregular and currents are slightly noisier than those in Fig. 12(b) and 12(e), due to the constraint on the FCS. However, CMV practically vanishes. A zoomed view, included in Fig. 12(c), reveals that the CMV is zero with very short (sub-\( \mu \)s length) pulses. These pulses are related to small mismatches between the inverter and the AFE switching instants, due to deadtime [23]. In these experiments, the dead-time effect has been minimized, but could not be completely eliminated [35].

Harmonic contents of the motor and AFE side currents are shown in Figs. 12(g)–12(i). They are consistent with the simulation results of Fig. 8. A summary of the performance characteristics is given in Table 4. It is clear that the proposed FBQ-MPC with CMV elimination dramatically reduces \( V_{cm} \) and \( V_{NER} \), at the cost of a relatively small increase in the current harmonics.

The last row of Table 4 presents the execution times for the three control schemes, in % to the sampling period (20 \( \mu \)s). Under the proposed, computationally efficient, FBQ-MPC with CMV elimination, the execution time is less than half the
cycle (allowing for reducing the sampling interval to 10 μs if necessary). Without the simplifications discussed in Section III-D, but rather using an exhaustive search, the CMV mitigation algorithm would take more than 100% of the cycle.

Two experimental points, obtained for 100% and 50% rated speed and torque, and $\lambda = 0.5$, are added to the THD trade-off plots in Fig. 9, for comparison. In a specific application, $\lambda$ can be further optimised. If necessary, the grid-side THD can be improved by increasing the size of the line inductors.

VI. CONCLUSION

This paper has proposed a novel control scheme for an AC drive based on an AFE-inverter topology and using MPC as its core principle. The traditional MPC scheme is enhanced by: using voltage domain cost function formulation; replacing model inversion by an optimal (in MPC sense) current control; providing integral action to reject steady state errors; and using Feedback Quantizer for improved harmonic performance. As shown in this paper, significant performance improvements result from these modifications.

Additionally, as desired in many industrial applications, the proposed AC drive scheme offers the option of CMV elimination. This is achieved without adding any additional terms in the MPC cost function, but rather by a smart utilisation of a restricted set of switching states by a computationally efficient algorithm. Despite the constraint on the control set, CMV elimination is achieved at the cost of only a small degradation of the input and output harmonic performance.

The findings of the paper have been validated by extensive simulations and experiments. The results have shown major advantages of proposed AC drive for industry applications.

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