Current Harmonic Mitigation Using a Multi-Vector Solution for MPC in Six-Phase Electric Drives

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ABSTRACT Multiphase machines offer some interesting features to develop more-competitive electric drives. However, the control complexity increases due to the higher number of freedom degrees. Specifically, the regulation of secondary currents becomes critical to avoid an unacceptable harmonic distortion of phase currents. In this regard, standard finite-control-set model predictive control (FCS-MPC) is characterized by a limited capability to provide a suitable current quality, since a single switching state is applied per control cycle. In order to reduce the current harmonic content and retain the well-known FCS-MPC advantages, the use of a multi-vector approach has been recently explored in the field of multiphase electric drives. Following this trend, this work develops a simplified FCS-MPC version with a new generation of virtual voltage vectors (VV) that are used as control actions. The switching states that form the proposed VV provide null average $x-y$ voltages and reduced instantaneous injection of these secondary components. The capability of the suggested VV-based FCS-MPC strategy to mitigate the $x-y$ injection is experimentally tested and compared to field-oriented control (FOC) using carrier-based pulse width modulation (CB-PWM) and diverse FCS-MPC schemes.

INDEX TERMS Finite-control-set model predictive control, multiphase electric drives, virtual voltage vectors.

I. INTRODUCTION Multiphase electric drives have allowed the implementation of novel operating modes for industrial applications [1], [2]. To cite one example, the independent regulation of the active/reactive power contribution for two microgrids can be carried out with a single six-phase energy conversion system, hence avoiding the use of interface inductors [3]. Apart from novel operating modes, multiphase systems also provide a better fault tolerance, an enhanced power distribution and an augmented torque production than conventional three-phase electric drives [4]. These favorable features have promoted the use of multiphase systems in industrial applications where higher reliability and/or efficiency are required [1].

Since the exploitation of these advantages requires a high-performance control scheme, different well-known three-phase regulation strategies have been extended to multiphase systems [1], [4]. Regardless of the selected control scheme, the impedance of the secondary $x-y$ subspace is typically low in distributed-winding machines [5]. For this reason, the phase current quality becomes highly dependent on how the $x-y$ current regulation is performed [6]. Focusing on the case of the asymmetrical six-phase machine, all the active switching states produce an inherent $x-y$ injection, but some of them have a rather low production in this critical subspace [6]. Then, these switching states become ideal candidates to avoid an unacceptable harmonic distortion of phase currents.

With that in mind, the control designer should ideally have the capability to directly discard voltage vectors with a high $x-y$ generation. In this regard, for instance, the utilization of field-oriented control (FOC) using carrier-based pulse-width modulation (CB-PWM) shows a limited flexibility to define the selectable voltage vectors to synthetize the desired reference voltage [7]–[11]. On the contrary, the discrete nature of finite-control-set model predictive control (FCS-MPC) schemes provides a higher flexibility and permits a pre-selection of the most advantageous switching states. In fact, in FCS-MPC the control designer can select in advance the set of voltage vectors employed to compose the
voltage output. Despite this capability, in the case of standard FCS-MPC the harmonic distortion is higher than in the case of FOC with CB-PWM [11] because a single switching state is applied during the entire sampling period.

Fortunately, the current quality limitation of standard FCS-MPC strategies can be significantly overcome with the utilization of virtual voltage vectors (VV) [6], [12]–[19]. These multi-vector techniques are characterized by the use of couples of voltage vectors as control actions to minimize the $x - y$ injection. Different sets of offline [12]–[17], [19] and online estimated [6], [18] VVs have been employed to simultaneously satisfy the flux/torque production and the harmonic injection reduction. Focusing on the critical mitigation of the harmonic distortion according to the operating point. The developed strategy is founded on the use of a new generation of LVVs (termed in this case as LVVs) is considered a satisfactory solution thanks to their behavior in the $x - y$ plane [17], [18]. Unfortunately, two adjacent large voltage vectors cannot provide null average $x - y$ voltages within the control cycle. Consequently, some remaining current harmonics appear if only LVVs are employed to mitigate these secondary components. Although the $x - y$ production can be reduced if the LVVs are combined with a null voltage vector in each sampling period, the cancellation of the average $x - y$ voltages cannot be fully achieved [17], [18].

This work develops a FCS-MPC that employs control actions characterized by a null average $x - y$ voltage, where the applied switching states also minimize the instantaneous $x - y$ injection according to the operating point. The developed strategy is founded on the use of a new generation of VVs formed by pre-defined trios of adjacent large voltage vectors (TV) and a specific null voltage vector. On the one hand, three is the minimal number of adjacent large voltage vectors to achieve a null average $x - y$ voltage, whereas, on the other hand, the null voltage vectors allow the minimization of the harmonic distortion according to the operating situation. To summarize, in the proposed multi-vector approach, four voltage vectors per sampling period are applied in the designed FCS-MPC (termed MV4-MPC from now on) to enhance the current quality and, consequently, reduce the stator copper losses. Searching for an extra reduction of the switching frequency losses, the developed FCS-MPC selects the null voltage vector with the minimum number of switch changes for each trio of large voltage vectors. This selection process is done offline to avoid increasing the computational cost. Attending to this issue, a single cost function is employed to choose the optimal trio of large voltage vectors and an analytical expression allows the calculation of the corresponding duty cycle of the null voltage vector. While the features that have popularized model predictive control in the field of electric drives are retained [20], the proposed MV4-MPC provides a current quality that is in the same range as in FOC with CB-PWM and better than in previous multi-vector FCS-MPC versions.

This manuscript is structured in the following manner. Section II describes the generalities of the tested asymmetrical six-phase induction motor drive. In Section III the proposed FCS-MPC is developed point-by-point. Section IV explores the performance of the implicit modulator of the considered MV4-MPC whereas Section V shows its capability to achieve a lower harmonic injection than FOC using CB-PWM or previous FCS-MPC schemes based on different sets of virtual voltage vectors. Finally, Section VI summarizes the main conclusions of this work.

II. SIX-PHASE INDUCTION MOTOR DRIVE

The electrical drive under study is an asymmetrical six-phase induction machine (IM) with two sets of three-phase distributed windings that are spatially shifted 30 degrees. The IM is fed by a two-level dual three-phase voltage source converter (VSC) as shown in Figure 1. The VSC is connected to a single dc-link providing $2^6 = 64$ switching states. Stator phase voltages ($v_i$) can be obtained from the dc-link voltage ($V_{DC}$) and the VSC switching states ($S_{ij}$) as follows:

$$
\begin{bmatrix}
    v_{i1} \\
    v_{i2} \\
    v_{i3} \\
    v_{i4} \\
    v_{i5} \\
    v_{i6}
\end{bmatrix} = \frac{V_{DC}}{3} \begin{bmatrix}
    2 & -1 & -1 & 0 & 0 & 0 \\
    -1 & 2 & -1 & 0 & 0 & 0 \\
    -1 & -1 & 2 & 0 & 0 & 0 \\
    0 & 0 & 0 & 2 & -1 & -1 \\
    0 & 0 & 0 & -1 & 2 & -1 \\
    0 & 0 & 0 & -1 & -1 & 2
\end{bmatrix} \begin{bmatrix}
    S_{11} \\
    S_{12} \\
    S_{13} \\
    S_{14} \\
    S_{15} \\
    S_{16}
\end{bmatrix}
$$

where $S_{ij}$ provides a binary value, related to the switching state of each VSCs leg. If the upper switch of the leg is ON and the lower switch is OFF $S_{ij} = 1$, and $S_{ij} = 0$ if the opposite occurs.

Phase variables can be used to describe the behavior of the IM, but different reference frames and transformation have been extensively used in the literature to facilitate the understanding and control of the IM. One of the most popular choices is the vector space decomposition (VSD), where phase variables are transformed into a stationary reference frame [21]. These variables are expressed in three orthogonal subspaces when the amplitude-invariant Clarke transformation is used as follows:

$$
[C] = \frac{1}{3} \begin{bmatrix}
    1 & -\frac{1}{2} & -\frac{1}{2} & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & 0 \\
    0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & \frac{1}{2} & \frac{1}{2} & -1 \\
    1 & -\frac{1}{2} & -\frac{1}{2} & -\frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & 0 \\
    0 & -\frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} & \frac{1}{2} & \frac{1}{2} & 1 \\
    1 & 1 & 1 & 0 & 0 & 0 \\
    0 & 0 & 0 & 1 & 1 & 1
\end{bmatrix},
$$

$$
[v_a, v_b, v_c, v_y, v_z, v_{i2}]^T = [C][v_{a1}, v_{b1}, v_{c1}, v_{a2}, v_{b2}, v_{c2}]^T,
$$

$$
[i_a, i_b, i_c, i_y, i_z, i_{i2}]^T = [C][i_{a1}, i_{b1}, i_{c1}, i_{a2}, i_{b2}, i_{c2}]^T,
$$

where $\alpha - \beta$ currents are related to the flux/torque production and $x - y$ currents only produce stator copper losses in a distributed-winding machine with negligible
spatial harmonics. The zero-sequence currents $z_1 - z_2$ present a null value due to the use of two isolated neutral points.

Applying (2) to each switching state, it is possible to map phase voltages onto VSD variables. Each voltage vector is identified using a decimal number equivalent to the binary number of the switching state vector $[S] = [S_{01}, S_{02}, S_{03}, S_{12}, S_{23}, S_{31}]$. Their location, in $\alpha - \beta$ and $x - y$ subspaces, is shown in Figure 2. Attending to this issue, the available voltage states can be sorted into five different groups: large ($C_l$), medium-large ($C_{ml}$), medium ($C_m$), small ($C_s$) and null voltage vectors ($C_n$). The quality index $R_{a\beta}$ proposed in [6] has been included in Table 1 to analyse the capability of these sets of voltage states to provide a suitable performance in both planes.

III. PROPOSED MULTI-VECTOR MPC

A. CONTROL ACTIONS

FCS-MPC can be classified as a regulation strategy with an implicit modulator because the gate signals (switching states) are directly evaluated as control actions. In the developed MV4-MPC, the voltage output in each control cycle is formed by a trio of adjacent large voltage vectors combined with a specific null voltage vector. In addition, a completely null voltage output is also evaluated as an available control action in the proposed control scheme. Due to the high impact of the secondary subspace on the phase current quality in asymmetrical six-phase machines, the control actions have been defined to maximize the performance in this plane. With that in mind, three requirements have been established:

i) To use active voltage vectors with an acceptable ratio $R_{a\beta}$.

ii) To obtain a null average $x - y$ voltage during the control cycle.

iii) To minimize the application time $t_a$ of active voltage vectors at different operating points.

According to the information from Table 1 and Figure 2, large voltage vectors can be considered the most suitable active control actions, since this set of switching states provides the lower $x - y$ injection and the better value of $R_{a\beta}$.

Focusing on the second constraint, three large voltage vectors is the minimal number of voltage vector to satisfy requirement ii). The per-unit duty cycles of any trio of large voltage vectors can be estimated as follows:

$$
\begin{bmatrix}
V^x_{11} & V^x_{12} & V^x_{13} \\
V^y_{11} & V^y_{12} & V^y_{13}
\end{bmatrix}
\begin{bmatrix}
t_1 \\
t_2 \\
t_3
\end{bmatrix}
= \begin{bmatrix} 0 \\ 1 \end{bmatrix},
$$

where $V^i_{ij}$ is the projection of each large voltage vector in the $x - y$ plane, $t_1$, $t_2$ and $t_3$ subscripts denote three adjacent large voltage vectors (being $l_2$ located between $l_1$ and $l_3$), and $t_1$, $t_2$ and $t_3$ the respective application times of these voltage vectors.

From (3), the following time solution is obtained:

$$
t_1 = 0.2679, \quad t_2 = 0.4642 \quad t_3 = 0.2679.
$$

These per-unit duty cycles can be extended to any TV based on symmetry considerations and thanks to their location in the $x - y$ plane (see Figure 2).

To minimize the application time $t_a$ of active voltage vectors during the control cycle, requirement iii), and, consequently, to mitigate the $x - y$ injection, the null voltage vector must also be employed. This application time $t_a$ can be obtained with an analytical expression related to the operating conditions (reference $q$-current $i_{qs}^*$ and rated $i_{qs,\text{max}}$):

$$
t_a = \frac{i_{qs}^*}{i_{qs,\text{max}}},
$$

where the $q$-component has been obtained applying the Park transformation [$D$] to VSD variables [19].

After taking into account the three requirements, it is possible to define the average voltage output $\bar{v}_{out}$ of each control action as follows:

$$
\begin{bmatrix}
\bar{v}_{\alpha}^x \\
\bar{v}_{\beta}^x \\
\bar{v}_{\alpha}^y \\
\bar{v}_{\beta}^y \\
\bar{v}_{\alpha}^z \\
\bar{v}_{\beta}^z
\end{bmatrix}
= \begin{bmatrix}
V^x_{11} & V^x_{12} & V^x_{13} \\
V^y_{11} & V^y_{12} & V^y_{13} \\
V^z_{11} & V^z_{12} & V^z_{13} \\
V^x_{21} & V^x_{22} & V^x_{23} \\
V^y_{21} & V^y_{22} & V^y_{23} \\
V^z_{21} & V^z_{22} & V^z_{23}
\end{bmatrix}
\begin{bmatrix}
t_a \cdot t_1 \\
t_a \cdot t_2 \\
t_a \cdot t_3 \\
(1 - t_a)
\end{bmatrix},
$$

FIGURE 1. Asymmetrical six-phase induction machine fed by a dual three-phase two-level voltage source converter.

FIGURE 2. Voltage vectors in $\alpha - \beta$ and $x - y$ subspaces.

TABLE 1. Ratio of production $v_{a\beta}$ and $v_{xy}$.

| Parameters   | $C_l$ | $C_{ml}$ | $C_m$ | $C_s$ | $C_n$ |
|--------------|------|---------|------|------|------|
| $v_{a\beta}$ | 100  | 64      | 47   | 33.33| 17   | 0    |
| $v_{xy \cdot 100}$ | 17  | 47      | 33.33| 64   | 0    |
| $R_{a\beta} = |v_{a\beta}|/|v_{xy}|$ | 3.8 | 1       | 1    | 0.3  | 0    |

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where a trio of large voltage vectors is combined with a specific null voltage vector \( V_{null} \). The proposed combination of voltage states and their duty cycles allows a simultaneous tangential and radial response in the \( \alpha - \beta \) plane:

- Tangential \( \alpha - \beta \) voltage requirements: twelve different trios of adjacent large voltage vectors can be selected (Figure 3).
- Radial \( \alpha - \beta \) voltage requirements: the duty cycle of the null voltage vector has been adapted to the operating conditions.

To address the reduction of VSC switching losses, it is convenient to select an optimal null voltage vector that minimizes the switching changes for each TV. Without lack of generality, the optimization process is exemplified when \( TV_2 \) (\( V_{36}, V_{52} \) and \( V_{54} \)) is selected as the optimal active voltage action. There are four different null voltage states in a dual three-phase two-level VSC and all of them can be combined with the selected TV. However, there is an optimal null voltage state to reduce the number of switch changes in the second transition of the synthesized voltage output (Figure 4). For example, if \( V_0 \) is employed as the null voltage vector, this solution implies four switch changes in the transition between \( V_{54} \) and \( V_0 \) (Figure 4a). Fortunately, the switching losses can be reduced if \( V_{63} \) is designated as the optimal null voltage vector, since the transition between \( V_{63} \) and \( V_{54} \) only involves two switch changes in this case (Figure 4b). This estimation is done offline for each TV, hence, the computational burden is also reduced.

**B. PROPOSED CONTROL SCHEME STRUCTURE**

The standard structure of FCS-MPC strategies typically includes an outer speed loop with a proportional-integer (PI) controller and a two-stage inner current loop based on a discretized mathematical predictive model of the machine (Figure 5). The outer PI controller provides the \( q \)-current reference, whereas the \( d \)-current reference is set to provide rated stator magnetic flux in the base-speed region. In the implemented control scheme, the \( q \)-current presents a double role: on the one hand it satisfies the reference speed tracking condition of the synthetic voltage output for the sampling period (6). Finally, the signal production of the selected control actions in the secondary plane.

As a result of the first stage, predicted currents, \( \hat{i}_{a\beta,s,k+1} \), are obtained and used in a second stage where available control actions are again evaluated in (7) to obtain \( \hat{i}_{a\beta,s,k+2} \) currents. The predicted currents for each control action and the reference currents, \( i_{a\beta,s,k+2} \), are assessed in a cost function. In an analogous manner as in (7), the cost function can be expressed only in terms of \( \alpha - \beta \) currents, without the inclusion of \( x - y \) components since their regulation is done in open-loop mode with the use of VVs (TV and null voltage vector).

\[
J = (i_{a,s,k+2} - \hat{i}_{a,s,k+2})^2 + (i_{\beta,s,k+2} - \hat{i}_{\beta,s,k+2})^2.
\]

The control action that achieves the minimum value of (9) provides the optimum active switching states. Then the TV and its corresponding null voltage vector are combined according to the estimated value of \( t_n \) in order to generate the voltage output for the sampling period (6). Finally, the signal gates and their duty cycles are sent to the VSC.
IV. EXPLORING THE MV4-MPC IMPLICIT MODULATOR

A. CLASSIFICATION AND GENERALITIES OF GATE SIGNAL GENERATION METHODS

Control methods for electric drives can be divided according to how they generate the gate signals:

- Control methods with pulse-width modulation.
- Control methods with an implicit modulator.

Attending to the control methods with a PWM stage, FOC scheme using a CB-PWM is one of the most widespread regulation techniques in electric drives at industry. This control strategy estimates the reference voltages via a set of inner current PI controllers. Subsequently, the modulation stage provides the VSC gate signals to synthesize the desired voltage references (Figure 6). Among the control methods with an implicit modulator, FCS-MPC strategies have played an interesting role in the last years thanks to their inherent advantages [20]. Schemes with an implicit modulator show two significant differences compared to control methods based on PWM: there are no reference voltages and the switching states are directly assessed in the control scheme as available control actions (Figure 6). The behavior of these two regulation techniques is related to their capability to generate the suitable control actions at each sampling period. Namely, the outputs of their respective modulator. This section provides an in-depth analysis of the proposed MV4-MPC, taking an asymmetrical six-phase drive as a case study and using CB-PWM as a benchmark to validate the goodness of the proposed control scheme.

B. GENERATION OF THE VOLTAGE OUTPUT

This section describes the nature of the considered modulation methods to select the gate signals in the synthesisization process of the voltage output. In the case of CB-PWM the working principle is based on the variation of the duty cycles to obtain the voltage references on average during the sampling period. Nevertheless, the suitable average voltage production cannot ensure the best harmonic mitigation if some selected voltage vectors present a high degree of \(x - y\) injection [17]–[20]. An extreme example of the average assumption is provided in [22], where two opposite large voltage vectors are applied with the same time ratio in order to obtain an \(a - \beta\) null voltage vector with minimum common mode voltage. Although the average voltage value is obviously null, it is obtained with an active production of each individual voltage vector and therefore the resulting harmonic distortion can be highly increased [22]. In this regard, CB-PWM stages follow an average approach without any flexibility to exclude voltage vectors with a high \(x - y\) voltage production. As highlighted in [20], this approach can produce an undesired large amplitude of \(x - y\) currents because the voltage vectors generated by the CB-PWM do not guarantee minimum projections on the \(x - y\) plane. The obtained gate signals in a sampling period, when a CB-PWM is implemented in an asymmetrical six-phase machine, are employed to exemplify the previous approach. In this case, an up-down triangle signal is used as the carrier-based signal [23]. Figure 7 shows the gate signals of each VSC leg and the voltage vectors created by these switching states for an examined sampling period. Concerning Figure 7, twelve switching changes appear (two per phase) as expected with the employed carrier signal. On account of the generated gate signals, six different voltage vectors are applied in the considered sampling period: \(V_{63}, V_{45}, V_{41}, V_{33}, V_1\) and \(V_0\). The average \(a - \beta\) and \(x - y\) production for these voltage

![FIGURE 5. Implemented MV4-MPC scheme.](image-url)
vectors can be estimated as follows (Table 2):

$$\begin{bmatrix}
\bar{v}_\alpha \\
\bar{v}_\beta \\
\bar{v}_x \\
\bar{v}_y
\end{bmatrix} = \begin{bmatrix}
V_{\alpha 63}^x & V_{\alpha 45}^x & V_{\alpha 41}^x & V_{\alpha 33}^x & V_{\alpha 31}^x & V_{\alpha 0}^x \\
V_{\beta 63}^x & V_{\beta 45}^x & V_{\beta 41}^x & V_{\beta 33}^x & V_{\beta 31}^x & V_{\beta 0}^x \\
V_{\alpha 63}^y & V_{\alpha 45}^y & V_{\alpha 41}^y & V_{\alpha 33}^y & V_{\alpha 31}^y & V_{\alpha 0}^y \\
V_{\beta 63}^y & V_{\beta 45}^y & V_{\beta 41}^y & V_{\beta 33}^y & V_{\beta 31}^y & V_{\beta 0}^y
\end{bmatrix} \begin{bmatrix}
t_{63} \\
t_{45} \\
t_{41} \\
t_{33} \\
t_{31} \\
t_{0}
\end{bmatrix}, \tag{10}
$$

where $\bar{v}_i$ represents the average voltage production in the respective subspaces $i$ ($i = \alpha, \beta, x, y$), $V_j^i$ is the contribution of voltage vector $j$ on subspace $i$ and $t_j$ the per-unit duty cycle of voltage vector $j$.

Attending to the average voltage values of Table 2, the $x$–$y$ injection could be considered satisfactory. However, according to Figure 2 and Table 1, $V_{33}$ and $V_1$ are classified as voltage vectors with a high injection of $x$–$y$ components. Therefore, the applied gate signals for this sampling period cannot be considered optimal from the point of view of the $x$–$y$ mitigation, because the instantaneous $x$–$y$ production is not minimized.

The same average $\alpha$–$\beta$ production can be reached with a more appropriate $x$–$y$ injection. For that end, a null voltage state can be combined with a trio of adjacent large voltage vectors. Attending to the $\alpha$–$\beta$ voltage vectors obtained in Table 2, $V_{37}$, $V_{45}$ and $V_{41}$ are selected as the necessary TV. A similar $\alpha$–$\beta$ production can be obtained with the proposed control action if the duty cycle of the active couple is $62\%$ of the sampling period, whereas the null voltage vector is used during the rest of this period. Following the same approach as in (10), the average $\alpha$–$\beta$ and $x$–$y$ voltage vectors are calculated. As shown in Table 3, for a similar $\alpha$–$\beta$ production the average $x$–$y$ voltage is null. Additionally, the instantaneous $x$–$y$ injection has been minimized with this voltage solution because only voltage vectors with a suitable $R_{\alpha\beta}$ value have been employed to produce the required flux/torque, and therefore the current quality can be enhanced. Furthermore, this voltage output has been obtained with the application of three voltage states, resulting in a lower switching frequency than in the case of CB-PWM.

Taking advantage of the situation described in the preceding paragraph, the next section provides experimental results where the designed MV4-MPC strategy achieves a lower harmonic injection of the secondary components than in the case of FOC with CB-PWM. Additionally, two different FCS-MPC strategies using virtual voltage vectors [17], [19] are employed in the proposed analysis in order to confirm the goodness of the proposed model predictive control technique.

V. EXPERIMENTAL RESULTS

A. EXPERIMENTAL TEST BENCH

Experimental tests have been carried out in the test bench shown in Figure 8. The multiphase electric drive consists of a six-phase induction machine connected to a two-level dual three-phase voltage source converter (Semikron SKS22F) supplied by a single dc-link. The machine parameters have been obtained using AC time domain and still with inverter supply test [24], [25] and are summarized in Table 4. The control actions are performed by a digital signal processor from Texas Instrument (TMS320F28335) programmed using a J-TAG and the TI property Software (Code Composer). The current and speed values are measured using four hall-effect sensors (LEM LAH 25-NP) and an encoder (GNH510296R) respectively. The IM is loaded coupling to the shaft to a dc-machine working as a generator. The armature of the dc-link is connected to variable passive load that dissipates the power, being load torque speed dependent.

B. EXPERIMENTAL VALIDATION

This section includes experimental results to prove the MV4-MPC advantages over three control schemes: VV-MPC [19], LVV-MPC [17] and conventional FOC using CB-PWM [9]. For that purpose, three comparative tests have been carried out to confirm the desirable performance of the designed MV4-MPC. Tests 1 and 2 evaluate the steady-state performance of the considered control schemes whereas Test 3 shows their responses when speed variations occur. In addition, the dynamic behavior of the proposed MV4-MPC is also analyzed in two extra transient scenarios.

![FIGURE 7. Example of applied switching states when a CB-PWM stage is employed. Being $T_m$ the sampling period.](image)

| TABLE 2. CB-PWM average voltage production. |
|---------------------------------------------|
| Voltage vector | Module | Argument |
|----------------|--------|----------|
| $v_{x\alpha}(p.u)$ | 0.371 | $-30.1^\circ$ |
| $v_{x\beta}(p.u)$ | 0.011 | $-96.21^\circ$ |

| TABLE 3. Implicit modulator average voltage production. |
|--------------------------------------------------------|
| Voltage vector | Module | Argument |
|----------------|--------|----------|
| $v_{x\alpha}(p.u)$ | 0.371 | $-45^\circ$ |
| $v_{x\beta}(p.u)$ | 0 | $-45^\circ$ |
1) TEST 1: HIGH-SPEED SCENARIO
Test 1 shows the performance of the aforementioned regulation techniques when the reference speed is set to 700 rpm and the load torque is equal to 3.5 Nm. Regardless of the implemented control scheme, the tracking of the reference speed is successfully carried out, as shown in Figure 9a. A suitable regulation of $d - q$ currents is provided by all methods (Figure 9b), with a lower $d$-current ripple in the case of FOC and MV4-MPC. This reduced current ripple appears because these regulation techniques can adapt the active voltage injection during the control cycle with the use of the null voltage vector. In other words, both schemes show a higher refinement in the $\alpha - \beta$ subspace than in the case of VV-MPC and LVV-MPC.

Focusing on the peak value of the $x - y$ currents (see Table 5), thanks to the nature of the control actions employed in the case of MV4-MPC this quality index is lower than in the others regulation techniques. In MV-MPC, the use of control actions formed by a large and a medium-large voltage vector provides a null average voltage production in the $x - y$ subspace. Nevertheless, medium-large voltage vectors also show a medium-large contribution in the secondary plane, increasing the harmonic injection and, thus, presenting a higher peak-to-peak value. Moreover, the injection of active voltages during the control cycle cannot be regulated. On the one hand, as previously exposed, the use of LVVs as control actions is considered a suitable solution to mitigate the $x - y$ production. Regrettably, the utilization during the entire control cycle of the considered virtual voltage vectors cannot produce a null $x - y$ injection, appearing certain harmonic content. Attending to FOC with CB-PWM, the mitigation of the secondary components shows a limited result because switching states with a higher $x - y$ contribution cannot be avoided. Hence the overall current harmonic injection is reduced in MV4-MPC compared to others control schemes. As a result of this $x - y$ current mitigation (Figure 9c), the phase currents of MV4-MPC show a total harmonic distortion $THD_{ph}$ of 11.12%, whereas this quality index increases up to 30.38%, 18.66%, and 16.21% for the cases of VV-MPC, LVV-MPC and FOC, respectively. The qualitative comparison of the phase current distortion can be observed in Figure 9d. All the quality indices from Test 1 are summarized in Table 5 and the switching frequency $f_{sw}$ of each regulation strategy has also been added in the caption of Figure 9.

2) TEST 2: LOW-SPEED SCENARIO
In Test 2, the reference speed is set to 500 rpm, and the load torque is equal to 3.2 Nm. The speed regulation is successfully regulated in both cases, as shown in Figure 10a. The tracking of $d - q$ currents is also suitable, showing the same $d$-current ripple trend as in Test 1 (Figure 10b). Analysing the waveform of $x - y$ currents, MV4-MPC outperforms all aforementioned control schemes with a lower injection of these components. In fact, the peak value of the $x - y$ currents with MV4-MPC is 1.78 A, whereas this quality index increases up to 3.98 A, 2.96 A, and 2.75 A for VV-MPC, LVV-MPC and FOC with CB-PWM, correspondingly. This reduction of the peak value implies a significant improvement over the considered control schemes (Figure 10c). This desirable reduction of harmonic distortion is obtained thanks to the proposed combination of a single TV with the null voltage vector and the suitable calculation of their duty cycles. The result of this new synthesis of voltage vectors is a lower THD of phase currents, as shown in Table 6. The switching frequency of each control scheme has been included in the caption of Figure 10.

3) TEST 3: COMPARATIVE DYNAMIC TEST
Test 3 (Figure 11) evaluates the performance of the four control strategies under comparison in a dynamic-situation. Reference speed changes in a ramp-wise form from 400rpm to 700rpm. Regardless of the selected control scheme, the tracking of the reference speed and $d - q$ currents is acceptably done (Figure 11a and 11b). From the point of view of the $x - y$ currents mitigation (Figure 11c), the phase currents of MV4-MPC show a total harmonic distortion $THD_{ph}$ of 11.12%, whereas this quality index increases up to 30.38%, 18.66%, and 16.21% for the cases of VV-MPC, LVV-MPC and FOC, respectively. The qualitative comparison of the phase current distortion can be observed in Figure 11d. All the quality indices from Test 1 are summarized in Table 5 and the switching frequency $f_{sw}$ of each regulation strategy has also been added in the caption of Figure 9.

![FIGURE 8. The employed test bench.](image-url)
vectors during the sampling period. On the other hand, IRFOC and MV4-MPC present control actions with a variable duty cycle of active switching states, reducing the $x - y$ voltage production as function of the operating conditions. However, the proposed FCS-MPC method offers an improved response thanks to the combination of the null voltage vector together with a single TV. This combination of voltage vectors satisfies the $\alpha - \beta$ requirements while the components of the secondary subspaces are minimized. As a summary, a suitable dynamic response is obtained with a lower harmonic distortion and a lower average value of the switching frequency (Figure 11).
4) TEST 4: LOAD TORQUE STEP SCENARIO
Test 4 analyzes the performance of the developed MV4-MPC when a load torque step is applied at \( t = 1.5s \), as shown in Figure 12. From the point of view of the dynamic response, the tracking of the speed and \( d-q \) currents (see Figure 12a and 12b). Attending to the current quality, the injection of secondary components increases due to the higher requirement of active voltage (Figure 12c). Nevertheless, a suitable ripple is obtained in the phase currents thanks to the nature of selected control actions.

5) TEST 5: SPEED REVERSAL TEST
Finally, the dynamic performance of the proposed MV4-MPC is assessed using a speed reversal test (see Figure 13).
For that purpose, the reference speed is set up from 500 rpm to −500 rpm. The response obtained with the considered MV4-MPC is acceptable with a suitable tracking of the reference speed/currents. On the other hand, the injection of the \( x \rightarrow y \) current is related to the active component requirements, as shown in Figure 13b and 13c. A zoom of the phase currents at the point of zero-speed crossing is depicted in Figure 13d.

Based on the presented results, MV4-MPC allows the improvement of the phase current quality obtained with previous multi-vector FCS-MPC approaches. Furthermore, it is worth noting that MV4-MPC enhances for the first time the harmonic distortion of FOC based on CB-PWM in the field of multiphase electric drives when an implicit modulator is employed. While predictive approaches are known to provide a good dynamic response and high flexibility for control designers [6], [18], [19], the higher steady-state phase current distortion was the Achilles heel of MPC strategies. The results from Tests 1 to 3 confirm the capability of MV4-MPC to improve the phase current quality retaining the well-known features of predictive approaches [20]. In addition, Tests 4 and 5 show the suitable dynamic response of the considered MV4-MPC.

### VI. CONCLUSION

Even though FCS-MPC schemes have been considered an interesting alternative to regulate multiphase machines, their capability to compete with FOC strategies has been questioned because standard predictive approaches show a limited capability to achieve a low harmonic distortion. However, this paper shows for the first time an empowered FCS-MPC outperforms other multi-vector FCS-MPC strategies and improves the total harmonic distortion of FOC based on CB-PWM for an asymmetrical six-phase machine.

The proposed MV4-MPC uses, as control actions, voltage vectors with an adaptive flux/torque production, a mitigated \( x \rightarrow y \) generation and reduced switching losses. The implicit modulator of the proposed MV4-MPC improves the performance provided by the CB-PWM, thanks to its flexibility to select only voltage vectors with a minimum \( x \rightarrow y \) contribution (e.g., THD is reduced over FOC by 30.59% in Test 1). At the same time, the use of 4 switching states with an optimized selection of the zero vector allows a lower switching frequency, and consequently, lower switching losses.

According to the experimental testing that is provided, MV4-MPC can be regarded as a proof of concept that shows how control methods with implicit modulators can be used without shame instead of regulation techniques founded on the use of a CB-PWM stage. Although MV4-MPC is specifically designed for six-phase electric drives, its main idea can be generalized for other number of phases. The proposal enables MPC to go one step beyond in the field of multiphase electric drives.

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