Reduction of common-mode voltage using a simplified FSC-MPC for a five-phase induction motor drive

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Abstract: This paper presents a cascaded closed-loop control of a five-phase induction motor (IM) drive system with common-mode voltage reduction capability. The outer loop uses standard PI regulators for speed reference tracking, while the inner loop control is implemented using a reduced order finite-state model-based current predictive approach. Establishing certain constraints, different combinations of switching state vectors are obtained, leading to the generation of different magnitudes of common-mode voltages. As a result, precise speed and current control is achieved in addition to the reduction of common-mode voltage and computational burden. Experimental results are presented to validate the concept.

1 Introduction

The increasing interest in applications such as electric vehicle propulsion and wind energy generation systems has motivated the research on electrical drives that are capable of providing higher resiliency and power capacity. As a result, the use of different topologies, types of machines, and voltage source inverters (VSIs) [1] and more recently multiphase systems (with more than three-phases) [2–4] have been extensively addressed in the literature. Electric drives are commonly constituted by VSIs based on semiconductor switching devices that operate at high frequency [1–5]. The high switching frequency of these devices leads in turn to the appearance of common-mode voltages that generate ground leakage currents that flow out of the motor frame through the grounding wire to the system ground or bearing currents that flow through the non-drive end to the drive-end bearings to the motor shaft [6, 7]. As a consequence, the supplied motors are subjected to overvoltage stress in the machines windings insulation, electromagnetic interference, and bearing failure, which may lead to the complete drive malfunction [8, 9]. Three different components can be identified in bearing currents [10], (i) displacement bearing current that results from \( \frac{dv}{dt} \) in the common-mode voltage, (ii) electric discharge machining bearing current originated due to dielectric breakdown of lubricating material between inner and outer bearing races, and (iii) circulating bearing current flowing between the motor frame through two bearings and caused by high-frequency flux around the motor shaft.

Several methods have been proposed to eliminate common-mode voltages using passive and active methods. Passive methods are based on designing and implementing different types of filters (i.e. differential, sine-wave and common-mode filters), in conjunction with the electrical drive [11]. Although, passive methods are effective in dealing with the common-mode voltage, they add to cost, weight, and volume [11]. On the contrary, the active approach emerged as a better alternative, eliminating the need to implement extra hardware. In this case, the pulse pattern is altered by modifying the pulse width modulation (PWM) scheme [11–14]. Several space vector PWM techniques (SVPWM) are presented in the literature [15, 16], based on the fact that the use of zero-voltage vectors generates the highest amount of common-mode voltage and hence must be avoided when synthesising the reference voltages. Although this approach is effective in reducing the common-mode voltage, the voltage gain or modulation index is sacrificed and as a consequence the achievable output voltage is reduced. To cope with this effect, a SVPWM was proposed recently for a five-phase drive, where instead of using the zero-voltage vectors, two opposing (180° phase shifted) space vectors are employed to generate the output voltages that lead to reduced common-mode voltage without sacrificing the modulation index [17].

The use of finite-state model-based predictive control (FCS-MPC) in multiphase drives has been studied in recent years due to their large control bandwidth and the capacity to include several control features by modifying the optimisation cost function, offering a precise and flexible control strategy. Nevertheless, as a natural rule, some trade-off does exist in this strategy due to the high computational cost that FCS-MPC require to solve the optimisation problem online, which further increases when several control variables are included. Model predictive current control has been investigated for five-phase drives [9, 18, 19], six-phase drives [20, 21], and seven-phase drives [22] with the aim of obtaining sinusoidal output voltages by eliminating the unwanted x-y plane vectors using a tuning factor in the cost function. Furthermore, in order to cope with the high computational cost to solve the optimisation problem online, some modified predictive schemes have been proposed in the literature considering the dismissal of certain voltage vectors resulting in optimal and suboptimal FCS-MPC strategies or restricting the available voltage vectors to be used by the predictive model using specific predefined constraints [23]. Moreover, the use of FCS-MPC with common-mode voltage reduction capability has also been studied for a five-phase drive [9] in the literature, implementing a modified predictive torque control (PTC) whose cost function considers torque, flux, and the magnitude of the applied voltage vectors as the main control constraints.

The aim of this paper, which is an extension of [24], is to devise a low computational cost FCS-MPC scheme to reduce the common-mode voltage, and consequently the common-mode currents, while ensuring the proper current and speed control of the multiphase drive. The proposed controller consists of an outer speed regulator and an inner simplified FCS-MPC current control that reduces the amount of online calculations by stabilising a subset of voltage vectors based on certain constraints, aiming to minimise common-mode voltage production and ensure proper operation. As the provided experimental tests show, the proposed method greatly enhances the performance of the five-phase drive by eliminating completely the displacement bearing current.
The common-mode voltage is given as:

\[ v_{\text{CM}} = -\frac{1}{3}(v_A + v_B + v_C + v_D + v_E) \]  

(2)

As it is seen from (2), it is not possible to obtain a zero common-mode voltage, nonetheless, its magnitude depends on the converter leg voltages.

To simplify the machine modelling equations, the five-phase drive can be mapped into the stationary reference frame using the power variant transformation [11]. This results in two orthogonal subspaces, namely, the \( \alpha - \beta \) and \( x-y \) planes and a zero sequence component. The phase-voltage space vectors obtained in the \( \alpha - \beta \) (3) and \( x-y \) (4) reference frames are shown in the left and the right side of Fig. 2, respectively. The different five-phase drive switching states yield 32 voltage vectors. Out of the 32 voltage vectors, 30 are active and two are zero vectors. At any instant of time, the inverter can produce only one space vector.

\[ v_{\alpha\beta} = \frac{2}{3}(v_a + a^0 v_b + a^0 v_e + a^1 v_d + a^1 v_e) \]  

(3)

\[ v_{xy} = \frac{2}{3}(v_a + a^0 v_b + a^0 v_e + a^1 v_d + a^1 v_e) \]  

(4)

where, \( a = e^{j(\pi/5)} \), \( a^0 = e^{j(2\pi/5)} \), \( a^1 = e^{j(3\pi/5)} \), \( a^1 = e^{j(4\pi/5)} \).

Notice in Fig. 2 that the voltage vectors are numbered in decimal values. When converted to their equivalent binary, each vector number will give the switching state of each leg, where 1 corresponds to the upper switch being ‘ON’ and 0 represents the lower switch being ‘ON’. The operation of each inverter leg is complementary in order to protect the DC-link, thus when one of the semiconductors is in ‘ON’ state, the other is in ‘OFF’ state.

The obtained voltage vectors can be grouped into three sets depending on their length, namely ‘large’ \((2/5)2\cos(\pi/5)V_{dc}\), ‘medium’ \((2/5)V_{dc}\), and ‘small’ \((2/5)2\cos(2\pi/5)V_{dc}\), forming three concentric decagons. The outer decagon space vectors of the \( \alpha - \beta \) plane (left side) map into the inner decagon of the \( x-y \) plane (right side), the inner most decagon of \( \alpha - \beta \) plane forms the outer decagon of the \( x-y \) plane, while the middle decagon space vectors

\[ v_{\alpha\beta} = \frac{2}{3}(v_a + a^0 v_b + a^0 v_c + a^1 v_d + a^1 v_e) \]  

\[ v_{xy} = \frac{2}{3}(v_a + a^0 v_b + a^0 v_c + a^1 v_d + a^1 v_e) \]  

(1)

\[ v_{\alpha\beta} = \frac{2}{3}(v_a + a^0 v_b + a^0 v_c + a^1 v_d + a^1 v_e) \]  

(3)

\[ v_{xy} = \frac{2}{3}(v_a + a^0 v_b + a^0 v_c + a^1 v_d + a^1 v_e) \]  

(4)
map into the same region. Furthermore, the above mapping shows that the phase sequence abc,dec,de of the α−β plane corresponds to α,c,e,b,d sequence of the x-y plane [11]. The common-mode voltage values for different switching combinations are given in Table 1. It is observed that large and small vectors provide the same $V_{CM}$ (0.4–0.2 $V_{dc}$), while medium length vectors provide a $V_{CM}$ of 0.8–0.2 $V_{dc}$. In consequence, the use of large and small voltage vectors results in lower common-mode voltage than when medium voltage vectors are used.

### Table 1
Swiching combinations and common-mode voltages in a five-phase VSI

| Vector number | Switching combination | Common-mode voltage |
|---------------|------------------------|---------------------|
| large vectors (2/5)$2\cos(\pi/5)V_{dc}$ | 25, 28, 14, 7, 19 $\{11001, 11000, 01110, 00111, 10011\}$ | 0.6 $V_{dc}$ |
| medium vectors (2/5)$V_{dc}$ | 24,12,6,3,17 $\{11000, 01100, 00110, 00011, 10001\}$ | 0.4 $V_{dc}$ |
| small vectors (2/5)$2\cos(2\pi/5)V_{dc}$ | 16,8,4,2,1 $\{10000, 01000, 00100, 00010, 00001\}$ | 0.2 $V_{dc}$ |
| | 29,30,15,23,27 $\{11101, 11110, 10111, 10111, 11011\}$ | 0.8 $V_{dc}$ |
| | 9,20,10,5,18 $\{01001, 10100, 01010, 00101, 10010\}$ | 0.4 $V_{dc}$ |
| | 26,13,22,11,21 $\{11010, 01110, 10111, 01011, 11011\}$ | 0.6 $V_{dc}$ |

#### Fig. 3
Implemented control scheme. Outer PI based speed control loop and an inner Finite-Control Set Model-Based Current Controller

#### 3 Proposed control system

An outer closed-loop speed regulator and an inner predictive current controller form the five-phase drive control scheme (Fig. 3). The $d$-current reference is set constant, while the $q$-current reference is given by the error between the reference and the measured speed. Then, these $d$- and $q$-current references are mapped in the stationary reference frame by means of the inverse Park transformation. In order to minimise losses, the $x$- $y$ current references are set to zero. The set of $α−β$ and $x$-$y$ current references are provided to the predictive controller. In each sampling period ($k$), the measured speed and currents are updated in the predictive controller and the 32 available voltage vectors are used to estimate the current evolution in ($k+1$). The overall objective of the predictive controller is to correctly track the $α$, $β$, $x$, $y$ currents and consequently its cost function $J$ is defined as in (5). The selected voltage vector to be applied ($S_{opt}^{(k+1)}$) is the one that provides the minimum value of the cost function.

$$J = A|\vec{v}_a|^2 + B|\vec{v}_q|^2 + C|\vec{v}_d|^2 + D|\vec{v}_d|^2$$

where the terms $A$, $B$, $C$, $D$ are the weight factors of each $α−β$–$x$–$y$ current error term (6) and (7).

$$\vec{v}_a = \vec{i}_d(k + 1) - \vec{i}_d(k) - \hat{\vec{i}}_d\hat{i}(k + 1)$$

$$\vec{v}_q = \vec{i}_q(k + 1) - \vec{i}_q(k) - \hat{\vec{i}}_q\hat{i}(k + 1)$$

$$\hat{\vec{i}}_d\hat{i}(k + 1) = \left(\frac{cos(\theta)}{sin(\theta)}\right) \times \frac{\vec{i}_d}{\vec{i}_d}$$

$$\theta = \int \left(\frac{\vec{i}_d}{\tau_c \times \vec{i}_d}\right) \times dr$$

where the $d$-current reference ($\vec{i}_d$) and the parameter ($\tau_c$) are constants.

Analysing Table 1, it is foreseeable that by appropriately selecting the voltage vector to be applied to the multiphase drive, it is possible to indirectly control the magnitude of the obtained common-mode voltage. This fact has been previously considered in [9] where the use of a predictive torque scheme with common-mode voltage reduction capability was analysed. In this case, all the 32 available voltage vectors were considered in the predictive controller, where the common-mode voltage reduction constraint was included in the cost function, obtaining satisfactory results. Nonetheless, calculating the mathematical model considering the 32 voltage vectors and including the common-mode voltage reduction constraint in the cost function along with the control parameters, increments the computational burden. To solve this, the proposed predictive current control maintains the traditional cost function for current control, but reduces the number of voltage vectors provided to the predictive controller based on the following constraints:

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R1. Maximise the $\alpha - \beta$ components.
R2. Reduce the obtained $V_{CM}$.

The selection of the subsets of voltage vectors will highly affect the performance of the electrical drive in different operating points. Considering the aforementioned constraints and the effect that each voltage vector has on the magnitude of the common-mode voltage (Table 1), four subsets of voltage vectors are defined considering five medium, five large, ten medium, or ten large vectors. Notice that small vectors are not taken into account due to the fact that they do not comply with constraint (R1).

4 Experimental implementation

Next, several experimental tests were performed with the multiphase drive considering different subsets of voltage vectors. The five-phase induction machine is driven by a two-level VSI (Fig. 4), formed by using five phases of two off-the-shelf SEMIKRON converters that are controlled by a Texas Instruments TMS320F28335 digital signal processor (DSP). The rotor mechanical speed and phase currents are provided to the DSP by a GHM510296R/2500 digital encoder and current sensors included in the SEMIKRON converters, respectively. The $d$-axis current reference is set to 0.57 A, while the $q$-axis current reference is given by a PI controller based on the speed error. Moreover, different operating points were considered, providing further insight on the performance of the multiphase drive depending on the selected subset of voltage vectors. A DC-link voltage of 300 V is provided to the VSI by an external DC source. The sampling period of the predictive controller is set to 1/10 kHz, and the mechanical load is emulated by an independently controlled DC-Machine.

4.1 Five Medium and large vectors

In the first place, the system behaviour considering only five medium and large vectors was considered. The results showed that the use of five medium vectors do not provide proper operation of the multiphase drive for the selected operating point. Furthermore, experimental tests performed at a lower operating point, considering a speed reference of 250 rpm and zero load torque (Fig. 5), show that neither the subset of five medium or five large voltage vectors are capable of providing the required power for a controlled operation of the electrical drive maintaining the speed reference, thus leading to distorted phase currents.

As a result, the subset of five voltage vectors is not considered in the following tests, where ten medium and large vectors are chosen to study the electrical drive performance under different operating points.

4.2 Using ten medium and large vectors

The effect of the selected voltage vectors on the common-mode voltage is shown in Fig. 6, where a no load test at 250 rpm is presented, considering 32 voltage vectors (left side plot), 10 medium vectors (middle plot), and 10 large vectors (right side plot). Notice that the obtained number of voltage levels and its magnitude change when the multiphase drive is provided with a different subset of vectors, obtaining the lowest value when only 10 large vectors are used.

Next, a series of tests were performed considering a transition from the 32 available voltage vectors and 10 large vectors (Figs. 7 and 8) at $t = 0.2$ s, under different working conditions (i, ii, iii, iv).

(i) $T_L = 0$ at 250 rpm (first row).
(ii) $T_L = 0$ at 500 rpm (second row).
(iii) $T_L = 0.5 \times T_n$ at 250 rpm (third row).
(iv) \( T_L = 0.5 \times T_n \) at 500 rpm (forth row).

where \( (T_n) \) represents the nominal torque of the machine.

It is possible to observe that the control is able to provide circular currents in the \( \alpha - \beta \) subspace, with near zero \( x-y \) currents, while appropriately tracking the speed, and \( q \)- and \( d \)-current references when 10 large vectors are used.

The experimental tests show that the proposed reduced order predictive current controller is capable of effectively minimising the common-mode voltage maintaining the conventional predictive current scheme simplicity. Furthermore, the computational cost of performing the optimisation problem online is reduced, being only necessary to calculate the subset of voltage vectors instead of the 32 available. Nonetheless, special consideration must be taken during the selection of the subset of voltage vectors to be used due to the impact it has on the achievable operating point. It is possible to state that the proposed method may not provide the optimal solution over a number of operating points, due to the constraints (R1 and R2) used to discard voltage vectors. However, the capability to reduce the common-mode voltage while reducing the calculation effort may make the proposed method a candidate for applications where high switching frequency is demanded and traditional predictive schemes are not able to perform the optimisation calculations online.

5 Conclusions

This paper presents a reduced-order finite-control set model-based predictive current controller with common-mode voltage minimisation capability. Considering the effect of the 32 voltage vectors available in a five-phase electrical drive in the common-mode voltage generation, four subsets of voltage vectors are defined and studied namely, five medium, five large, 10 medium, and 10 large voltage vectors. The proposed control is general in nature and can be adapted to different phase numbers of multiphase drive. In the proposed control technique, the cost function is chosen such as to minimise the effect of \( xy \) plane vectors and maximise the influence of \( \alpha \beta \) vectors in order to obtain sinusoidal output voltage and hence sinusoidal mmf. Provided experimental results demonstrate that five medium and large voltage vectors are not capable of maintaining the electrical drive operating even at the

Fig. 6 Obtained common-mode voltage under a no load test with 32 voltage vectors (left side plot), 10 medium voltage vectors (middle plot) and 10 large vectors (right side plot) with a reference speed of 250 rpm

Fig. 7 Phase currents (left side), \( \alpha - \beta \) (middle) and \( x-y \) (right side) plots, under different operating conditions. A transition from the 32 available voltage vectors to the subset of 10 large vectors is performed at \( t = 0.2 \) s
lowest operating point, while the use of 10 medium vectors maintains the electrical drive operating under low speed and low-load torque conditions. Finally, it is shown that the use of 10 large vectors provide a wider operating range, maintaining sinusoidal currents and following the controlled references, while reducing the common-mode voltage. As a result, the reduction of the computational burden in order to solve the optimisation problem online comes at the expense of the careful selection of the subsets of voltage vectors in order to ensure proper performance under different working points.

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7 References

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