Predictive-space vector PWM current control method for asymmetrical dual three-phase induction motor drives

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Abstract: The interest in multiphase drives has re-emerged in the last decade, being the asymmetrical dual three-phase induction motor drive one of the most popular options. Predictive control techniques, already implemented in three-phase drives, have been recently adapted to the multiphase case. Schemes proposed so far have demonstrated high performance at the expenses of a higher degree of computational cost and ill-defined switching frequency. In this study, a predictive space vector PWM (SVPWM) current control technique with fixed switching frequency is proposed for asymmetrical dual three-phase AC drives. Fast torque and current response are achieved similar to those obtained using conventional predictive current control techniques. Electrical noise suppression is favoured as in PWM current control methods. Experimental results are provided to examine the benefits of the proposed control method.

1 Introduction

Although the analysis and application of multiphase machines goes back to the late 1920s [1], the development and research in this area has been scarce until the beginning of the 21st century. Multiphase electrical drives have been recently proposed for applications where some specific advantages (lower torque pulsations, less DC link current harmonics, higher overall system reliability, better power distribution per phase) can be better exploited [2]. For instance, high-power applications take advantage of the power splitting whereas high security applications benefit from the robustness provided by the redundancy of multiphase machines [3]. Another niche of industrial applications includes electrical and hybrid vehicles [4], ship propulsion [5–7] or wind power systems [8], showing a good prospect for the applicability of multiphase drives.

Among multiphase drives, a very interesting and much discussed in multiphase solution literature is the dual three-phase induction machine which has two sets of three-phase windings spatially shifted by 30 electrical degrees with isolated neutral points (also called asymmetrical dual three-phase AC machine) [9]. Different control techniques, already developed and implemented in three-phase machines, have been extended to the multiphase case. The most frequent control structure for asymmetrical six-phase drives is a cascaded scheme with an inner loop for current control and an outer loop for flux and speed control. The well-known field-oriented control (FOC) and direct torque control (DTC) techniques have been successfully applied to asymmetrical six-phase machines, dealing with problems associated with machine and converter asymmetries [10–13]. Controllers are usually of proportional–integer (PI)-type while current control is achieved by means of carrier based or space vector PWM (SVPWM) schemes [14–16].

When fast torque and current response are required, predictive control technique is one of the most widely used control techniques. The predictive control theory, developed
at the end of the 1970s, determines the optimal set of voltage source inverter (VSI) switching states during a prediction horizon based on a model of the real system. The predicted control schemes have been recently extended to the current control of asymmetrical dual three-phase drives, avoiding the use of PI controllers and PWM schemes which are difficult to design in multiphase drives and offering a high flexibility control design [17–19]. The main shortcoming of predictive control for AC drives is the intensive calculations involved in determining the optimal set of VSI switching states to be applied in the next sampling time. However, the increase in computing power of microprocessors has made predictive control plausible for controlling multiphase power converter and electrical drives. Another important drawback of predictive control is the switching frequency that depends on the torque reference and the load parameters. Power converters that operate at a fixed switching frequency have desirable electrical noise characteristics. For instance, the fixed frequency enables the use of simple filters and blanking techniques to suppress any electrical noise emitted from the converter. In this case, enhanced versions of predictive control techniques are presented for conventional three-phase AC drives [20], selecting more than one VSI switching states during a computation period to fix the switching frequency in the VSI.

This paper considers the use of predictive and SVPWM schemes for the current control of multiphase drives when fast current and torque responses are required. The asymmetrical dual three-phase AC machine is used as a case example because of its interest, and the performance of the proposed control technique is studied for quasi-balanced drive operation. By using the SVPWM with predictive control, lower ripples in the current are achieved while the fast dynamic response is maintained. The difficulties in the generalisation of the control method are analysed, and its performance is studied, comparing the obtained results with conventional predictive current control strategies.

The paper is organised as follows. First, the asymmetrical dual three-phase induction motor drive is described in Section 2. Section 3 details the conventional predictive current control technique and its application to the asymmetrical dual three-phase induction machine. The general principles of the predictive SVPWM current control method are shown in Section 4. Then, Section 5 compares the experimental results obtained using the proposed current control technique with a conventional predictive current control method and a standard PWM strategy. Finally, the conclusions are given in the last section.

2 Asymmetrical dual three-phase AC drive

The system under study consists of an asymmetrical dual three-phase AC machine supplied by a six-phase VSI and a DC link. A detailed scheme of the drive is provided in Fig. 1.

Figure 1 General scheme of an asymmetrical dual three-phase AC drive

This six-phase machine is a continuous system which can be described by a set of differential equations. Machine modelling follows two different paths: the double $d$–$q$ winding approach and the vector space decomposition (VSD) approach. According to the first approach [21], the machine can be represented with two pairs of $d$–$q$–$o$ windings corresponding to the two three-phase stator windings. From this point of view, the analytical model of the asymmetrical dual three-phase induction machines is an extension of the conventional three-phase induction machine model. The $d$–$q$–$o$ reference frame transformation decomposes the original three-dimensional vector space into the direct sum of a $d$–$q$ subspace and a zero sequence subspace which is orthogonal to $d$–$q$, decoupling the components that produce rotating magnetomotive force (m.m.f.) and the components of zero sequence. According to the second approach [14], VSD, the machine can be represented with three stator–rotor pairs of windings in orthogonal subspaces. One stator–rotor pair engages with electromechanical energy conversion ($\alpha$–$\beta$ subspace in what follows), whereas the others do not. The first stator–rotor pair represents the fundamental supply component plus supply harmonics of the order $12n \pm 1$ ($n = 1, 2, 3, \ldots$). The second stator–rotor pair represents supply harmonics of the order $6n \pm 1$ ($x$–$y$ subspace with $n = 1, 3, 5, \ldots$), whereas the zero sequence harmonic components can exist only if there is a single neutral point and they then belong to the third pair.

Although it can be modelled using the double $d$–$q$ approach as an extension of the $d$–$q$ approach of three-phase machines, the most popular option is the use of the VSD approach because it explains the physical phenomena in the machine better. According to the VSD approach, the machine can be modelled, using an amplitude invariant
The electromechanical energy conversion variables are mapped in the \((\alpha, \beta)\) subspace, meanwhile the non-electromechanical energy conversion variables can be found in the other subspaces.

The current components in the \((x, y)\) subspace do not contribute to the airgap flux, and are limited only by the stator resistance and stator leakage inductance, which is usually small. These currents will only produce losses in the stator resistance and stator leakage inductance, which is limited only by the machine parameters. From the motor model, the following conclusions should be emphasised here:

1. The electromechanical energy conversion variables are mapped in the \((\alpha, \beta)\) subspace, meanwhile the non-electromechanical energy conversion variables can be found in the other subspaces.

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3. Predictive control for asymmetrical dual three-phase AC machines: general principles

When fast torque responses are required, high-performance current control is requested. Among these current control techniques, predictive current control techniques show very good performance compared to classical methods like DTC or vector control techniques [17–19].

Predictive control uses a model of the physical system (the VSI and the machine). This model, called predictive model, is used at each sampling period to predict the machine state vector evolution for each possible VSI state. The machine equations (1) and (2) can be written in state space taking the VSI and the machine). This model, called predictive model, is used at each sampling period to predict the machine state vector evolution for each possible VSI state. The machine equations (1) and (2) can be written in state space taking

The VSI has a discrete nature and has a total number of \(2^6 = 64\) different switching states defined by six switching functions corresponding to the six inverter legs \([S_a, S_b, S_c, S_d, S_e, S_f]\), where \(S_i \in \{0, 1\}\). The different switching states and the voltage of the DC link \((V_{dc})\) define the phase voltages which can in turn be mapped to the \((\alpha-\beta-\ x-y)\) space according to the VSD approach. Consequently, the 64 different on/off combinations of the six VSI legs lead to 64 space vectors in the \((\alpha-\beta-\ x-y)\) subspace. Fig. 2 shows the active vectors in the \((\alpha-\beta-\ x-y)\) subspaces, where each switching state is identified using the switching function by two octal numbers corresponding to the binary numbers \([S_a, S_b, S_c, S_d, S_e, S_f]\) and \([S_d, S_e, S_f]\), respectively. For the sake of conciseness, the 64 VSI switching vectors will be usually referred as voltage vectors, or just vectors, in what follows. It must be noted that the 64 possibilities imply only 49 different vectors in the \((\alpha-\beta-\ x-y)\) space. Nevertheless, redundant vectors should be considered as different vectors because they have a different impact on the switching frequency even though they generate identical torque and losses in the six-phase machine.
rotor current

\[ A(k) = I + T_m \]

\[
\begin{bmatrix}
-R_x \lambda_2 & \omega_t(k) L_m \lambda_3 & 0 & 0 \\
-\omega_t(k) L_m \lambda_3 & -R_y \lambda_2 & 0 & 0 \\
0 & 0 & -R_y \lambda_5 & 0 \\
0 & 0 & 0 & -R_y \lambda_5
\end{bmatrix}
\]

\[ B'(U(k)) = T_m \frac{V_{dc}}{9} U(k) \]

\[
\begin{bmatrix}
2 & -1 & -1 & 0 & 0 & 0 \\
-1 & 2 & -1 & 0 & 0 & 0 \\
-1 & -1 & 2 & 0 & 0 & 0 \\
0 & 0 & 0 & 2 & -1 & -1 \\
0 & 0 & 0 & -1 & 2 & -1 \\
0 & 0 & 0 & -1 & -1 & 2
\end{bmatrix}
\]

\[
C(k) = X(k) - X(k-1) - [A(k-1)X(k-1) + B(U(k-1))]
\]

being \( I \) the 4 \( \times \) 4 identity matrix, \( B' \) the transpose matrix of \( B \), \( c_i = \cos(i\pi/6) \), \( s_i = \sin(i\pi/6) \), \( \lambda_2 = L_r/(L_r - L_m) \), \( \lambda_3 = L_m/(L_r L_m - L_m^2) \) and \( \lambda_5 = 1/L_m \).

The control action is obtained solving an optimisation problem aimed at minimising a cost function. Different cost functions can be used to express different control criteria. For instance, the distance between the reference and the predicted stator currents can be used, defining the cost function \( J = ||\hat{i}_{rs} - i_{rs}|| \), where \( \hat{i}_{rs} \) is the reference stator current and \( i_{rs} \) is the predicted stator current which is computationally obtained using the predictive model. The direction of the stator current evolution can also be used to define a simple cost function. A cost function in the \( \alpha-\beta \) subspace \( J = |i_{\alpha}(k+1) - \hat{i}_{\alpha}(k+1)| + |i_{\beta}(k+1) - \hat{i}_{\beta}(k+1)| \), considering the 12 outermost vectors plus a zero one is a good alternative for the predictive current control implementation in the asymmetrical dual three-phase induction motor drive [17] if quasi-balanced operation of the drive is assumed. Notice that \( x-y \) stator current components should be taken into account in the cost function to obtain a good dynamic performance, if quasi-balanced operation of the drive is not assumed. The number of voltage vectors to evaluate the predictive model can be further reduced if only the 12 outer vectors (the largest ones) are considered. This assumption is commonly used in current control of the asymmetrical dual-three-phase AC machine [12, 14, 17].

In this way, the optimiser can be implemented using only 13 possible stator voltage vectors (12 active and 1 zero vectors), and the predictive current control with sinusoidal output voltage requires less computing time and favours the real-time implementation of the control algorithm. A detailed block diagram of the conventional predictive control technique for the asymmetrical dual three-phase induction motor drive is provided in Fig. 3, including a pseudo code of the control algorithm. The sequencer issues voltage vectors one at a time, whereas the minimiser chooses the one that provides the lower value of \( J \). The selected voltage vector is applied to the physical system (the AC machine) during a sampling time.

4 Proposed predictive SVPWM current control method

Conventional predictive control [17, 18] avoids the use of PI controllers and modulation techniques since a single switching vector is applied during the whole switching period. This procedure is somewhat similar to original DTC schemes and leads to variable switching frequency. Furthermore, the voltage harmonics are spread over the entire frequency spectrum instead of being concentrated around a certain predefined frequencies. Both the variable switching frequency and the spread harmonic characteristic are undesirable for two main reasons. Firstly, a constant switching frequency allows a better design and selection of the switching devices since the stress of the switches can be known a priori, which is not the case in conventional predictive control schemes. Secondly, the filtering of the current harmonics is easier as compared with the case when load voltage spectrum is spread over a wide range of frequencies, as it is shown in [22]. In order to overcome the problems of variable frequency, original DTC schemes
with hysteresis controllers and selection tables where modified in the 1990s for these reasons, including space vector PWM to have fixed switched frequency. New constant switching frequency DTC schemes have been successful, including the fast torque response of DTC and the fixed switching frequency of modulated vector control. This proposal follows a similar procedure but for the case of predictive control, achieving a constant switching frequency control scheme. In addition, the use of a single switching vector during the whole period in classical predictive current controllers makes it impossible to achieve null $x-y$ voltage components since every switching vector $S_i$ is associated with certain $x-y$ components that can be calculated using the VSD (Fig. 2).

The proposed predictive SVPWM current control method aims at improving these problems by combining various states of the VSI within one sample period. It was proposed in [20] that the use of a null vector together with the selected vector $S_k$ during the same sample period enhances the performance. The present proposal goes one step beyond, including a modulation process. The main idea of the proposed method is to use the predictive control to substitute the standard PI controllers but to maintain a conventional modulation technique. Following this procedure it is possible to achieve fixed switching frequency, adequate voltage harmonic spectrum profile and null $x-y$ voltage components, while the fast torque and current responses are also maintained thanks to the predictive control technique. The principle of operation is as follows. For a desired stator current vector $\vec{i}_s$, the proposed control scheme proceeds as in a conventional predictive current control method, using a predefined cost function to select the VSI switching states. The minimiser chooses the switching vector $S_{\text{optimum}}$ that provides the

**Table 1 Parameters of the asymmetrical dual three-phase induction machine**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>stator resistance</td>
<td>$R_s$ (Ω)</td>
</tr>
<tr>
<td>rotor resistance</td>
<td>$R_r$ (Ω)</td>
</tr>
<tr>
<td>stator inductance</td>
<td>$L_s$ (H)</td>
</tr>
<tr>
<td>rotor inductance</td>
<td>$L_r$ (H)</td>
</tr>
<tr>
<td>mutual inductance</td>
<td>$L_m$ (H)</td>
</tr>
<tr>
<td>inertia</td>
<td>$J$ (kg m$^2$/s)</td>
</tr>
<tr>
<td>pairs of poles</td>
<td>$P$</td>
</tr>
<tr>
<td>friction coefficient</td>
<td>$B$ (kg m$^2$/s)</td>
</tr>
<tr>
<td>nominal frequency</td>
<td>$\omega_0$ (Hz)</td>
</tr>
</tbody>
</table>
lower value of $J$. The selected vector provides the optimum solution $[u_{α}^{opt}, u_{β}^{opt}]$ in terms of currents error in the $α$–$β$ subspace, but the $x$–$y$ subspace is not considered in the cost function. Consequently, good dynamic performance is expected using conventional predictive control but at the expense of lower efficiency because of high $x$–$y$ Joule losses that do not generate any torque. Instead of applying the chosen voltage vector to the multiphase machine during the whole switching period, which is the procedure in conventional predictive control schemes, the proposed method uses $[u_{α}^{opt}, u_{β}^{opt}, 0, 0, 0]$ as the voltage references in the $α$–$β$–$x$–$y$–$z_1$–$z_2$ subspaces to solve a SVPWM problem. Since $x$–$y$ components are undesirable, the inputs for the SVPWM are only the $α$–$β$ component of the phase voltage, and the $x$–$y$ inputs are set to zero. It must be highlighted here that null $x$–$y$ components can only be achieved with the proposed method and not with conventional predictive control, being this an interesting feature of the present proposal. The selected VSI switching modulation indexes are then obtained from the mathematical expression of the phase voltages defined in the VSD theory \([14]\), as shown in (8), where the factor of $1/3$ corresponds to the amplitude-invariant criterion adopted in the transformation matrix. The modulation indexes ($τ_i$) associated to VSI phases have also been included in (8), being $τ_i$ scaled between 0 and 1.

$$
\begin{bmatrix}
    u_α \\
    u_β \\
    u_x \\
    u_y \\
    u_{1z} \\
    u_{2z}
\end{bmatrix} = \frac{1}{3} \begin{bmatrix}
    1 & -\frac{1}{2} & -\frac{1}{2} & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & 0 \\
    0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & 1 & 1 & -1 \\
    1 & -\frac{1}{2} & -\frac{1}{2} & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & 0 \\
    0 & -\frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} & 1 & 1 & -1 \\
    1 & 1 & 1 & 0 & 0 & 0 \\
    0 & 0 & 0 & 1 & 1 & 1
\end{bmatrix} \begin{bmatrix}
    u_a \\
    u_b \\
    u_x \\
    u_y \\
    u_{1z} \\
    u_{2z}
\end{bmatrix}
$$

Once the references for the SVPWM are set, the calculation of the duty cycles for each VSI leg can be performed in a standard manner, and the modulation indexes are obtained using (8) as follows

$$
\begin{bmatrix}
    τ_a \\
    τ_b \\
    τ_x \\
    τ_y \\
    τ_{1z} \\
    τ_{2z}
\end{bmatrix} = T \frac{2}{3} \sqrt{2 + 3V_{dc}} \begin{bmatrix}
    τ_a \\
    τ_b \\
    τ_x \\
    τ_y \\
    τ_{1z} \\
    τ_{2z}
\end{bmatrix} = \frac{1}{2} + \frac{3}{2 \sqrt{2 + 3V_{dc}}} T^{-1} \begin{bmatrix}
    u_{α}^{opt} \\
    u_{β}^{opt} \\
    0 \\
    0 \\
    0 \\
    0
\end{bmatrix}
$$

A detailed block diagram of the proposed technique is provided in Fig. 4, including a pseudo code of the control algorithm. The proposed method is a hybrid solution between conventional predictive control and standard FOC,
maintaining interesting features of both schemes. Specifically, the use of predictive instead of PI controllers allows minimising the current error considering future values, avoids the tuning of PI controllers and provides enhanced flexibility through the definition of the cost function (which can minimise not only the current error but also the number of commutations, the switching losses or the DC link unbalance in multilevel converters [23, 24]). On the other hand, maintaining the SVPWM helps to obtain fixed switching frequency, adequate harmonic spectrum and lower \( x-y \) components. The results shown in the next section confirm these expected advantages.

5 Results

An experimental test rig has been designed for obtaining experimental results. The test rig is based on a conventional 36 slots, two pairs of poles, 10 kW three-phase induction machine whose stator has been rewound to construct a 36 slots, three pairs of poles, dual three-phase induction machine. Two sets of stator three-phase windings spatially shifted by 30 electrical degrees have been included.

A diagram and photos of the complete system are shown in Figs. 5 and 6. Table 1 shows the parameters of the machine used to obtain the experimental results. The control system is based on the TMS320LF2812 Texas Instruments digital signal processor (DSP) and the MCK2812 system. The control code is written in C, performing closed-loop current control, and using an optimised sampling frequency of 5 kHz, obtained after using specialised floating-point mathematical libraries and many source-code and compiler optimisations. A comparative study has been done. The conventional predictive current control technique proposed in [17, 18], and the proposed predictive SVPWM current control technique have been implemented at low electrical frequencies, where the proposed method better shows the obtained improvements. A series of tests are performed in order to examine the predictive SVPWM current control properties. Figs. 7–11 show the obtained results.

First, a 2.5 A reference stator current at 12 Hz is established. Fig. 7 depicts the current tracking in the \( \alpha-\beta-x-y \) subspaces using the proposed method (left side) and the conventional predictive current control technique

**Figure 8** Total (measured) harmonic distortion content in the stator current

Experimental results for a 2.5 A reference stator current at 12 Hz, using the predictive SVPWM current control method (left side) and the conventional predictive current control technique proposed in [17, 18] (right side). The lower graph shows a zoom in the sampling frequency (5 kHz).

**Figure 9** Experimental results for a 2.5 A reference stator current at 12 Hz, using a PI-PWM current control technique in the \( \alpha-\beta \) subspace (left side) and a PI-PWM current control technique in the \( \alpha-\beta \) and \( x-y \) subspaces (right side)

Stator current tracking is shown in the \( \alpha-\beta \) and \( x-y \) subspaces. The tuning of the PI controllers has been performed using the approach presented in [18]. The total (measured) harmonic distortion content in the stator current is shown in the lower graphs.
Fig. 9 shows the current tracking using conventional PI–PWM methods. Better stator current tracking is obtained in the $\alpha$–$\beta$ subspace using the proposed method. However, the most important advantage of the proposed control technique against the conventional one is the obtained performance in the $x$–$y$ subspace where the stator current components highly decrease, as it is shown in Fig. 7 (figures in the middle side). The decrease in the $x$–$y$ stator current components implies a reduction in the harmonic components of the stator current, as it is shown in Fig. 8. The total (measured) harmonic distortion content is lowered using the proposed method. While the conventional predictive control technique spreads harmonics over the entire frequency spectrum (Fig. 8, right side), the harmonic frequencies are concentrated around the sampling frequency using the proposed method (Fig. 8, left side), simplifying their suppression.

Finally, simultaneous amplitude (from 2.5 to 3.5 A) and frequency (from 12 to 36 Hz) tracking is investigated. Figs. 10 and 11 show the obtained results. Again, good stator current tracking is obtained in the $\alpha$–$\beta$ subspace, being near zero the stator currents in the $x$–$y$ subspace.

The obtained results prove that the proposed predictive control method can be a good alternative in comparison to conventional ones; better stator current tracking is obtained in the $\alpha$–$\beta$ subspace using the proposed method when compared with conventional PI–PWM or predictive techniques, and the stator current performance in the $x$–$y$ subspace is also improved when compared with conventional predictive controller which simplifies the harmonic suppression. Notice also that the active vector selection using the proposed predictive control technique is not a time-consuming task because most of the required operations, (9), can be off-line evaluated. The increase in the computational cost is about 2% of the total computational cost.

6 Conclusions

The area of multiphase induction motor drives has experienced a substantial growth in recent years. Research has been conducted worldwide and numerous interesting developments have been reported in the literature, particularly in the VSI-driven asymmetrical dual three-phase AC machine and in the implementation of fast torque control schemes. Only the development of modern microprocessors allows the real-time implementation of predictive control techniques in conventional and multiphase drives with fast current and torque requirements, because of its high computational cost. It has been shown in this paper that it is possible to combine the use of predictive control together with modulation techniques in multiphase drives. Using predictive control as a substitute of conventional PI controllers but maintaining the PWM scheme, it is a hybrid scheme between a conventional modulation technique and a predictive control method, and it combines advantages of both approaches. On the one hand using predictive controllers to obtain the phase voltage references avoids the PI tuning, it provides enhanced performance by considering future current errors and exhibits higher design flexibility thanks to the use of a cost function. On the other hand, maintaining the modulation technique avoids the variable switching frequency inherent in conventional predictive controls, easing the switch
selection and providing a more adequate harmonic profile. The experimental results confirm the viability and capability of the proposed current control method.

7 Acknowledgment

The authors gratefully acknowledge support provided by the Spanish Ministry of Education and Science within the I + D + i national project with reference DPI2005/04438.

8 References


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IET Electr. Power Appl., 2010, Vol. 4, Iss. 1, pp. 26–34