Predictive Current Control of an Induction Machine Fed by a Matrix Converter With Reactive Power Control

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Abstract—A different approach to perform the control of an induction machine fed by a matrix converter (MC) is presented in this paper. The proposed technique is based on predictive control and effectively controls input and output variables to the power converter, as expected from an MC. The method allows the use of all valid switching states, including rotating vectors that are not considered in most control techniques, as space vector modulation or direct torque control for induction machines fed by MCs. Experimental results show the excellent performance of the proposed approach, with low-distortion input currents, adjustable power factor, sinusoidal output currents with smooth frequency transitions, and good speed control in motoring and regeneration conditions, even working under an unbalanced power supply. The implementation and comprehension of the method should be considered simple compared to other control strategies with similar features. The high computational effort required should not be a problem considering recent progresses in digital signal processors—and even less in years to come.

Index Terms—AC–AC power conversion, ac motor drives, current control, induction motor drives, matrix converters (MCs), predictive control, reactive power.

I. INTRODUCTION

The matrix converter (MC) is a single-stage power converter capable of feeding an \(n\)-phase load from an \(m\)-phase source (\(n \times m\) MC) without energy storage devices [1]. As drive for electric ac machines [2], it represents an alternative to back-to-back converters, particularly in cases where size and the absence of large capacitors or inductances to store energy are relevant issues [3]–[5]. Several modulation techniques have been developed to control an MC, which can be classified into two main groups: scalar and vectorial methods [6]–[11]. The high number of switching states, the direct interaction between source and load, and the presence of rotating vectors introduce an important complexity in the analysis and control of an electric machine through an MC [12]–[17]. Ideally, the MC should feed the load with sinusoidal currents while it generates sinusoidal input currents with controlled power factor (PF) to the mains [18]–[20]. The behavior of this drive under polluted ac voltage supplies has called the attention of several works that usually face the challenge incorporating modifications to previous modulation techniques [21]–[27].

Predictive control is a control theory which was developed at the end of the 1970s [28]. Variants of this type of control strategy, associated with modulation techniques, have been used for current control [29], drives [30]–[32], PF correction [33], and active filters [34], [35]. Another variant directly selects the switching state, evaluating the error of a predicted vector for current and torque control [36], [37]. The method is not based on the evaluation of a quality function.

Recently, model-based predictive control (MPC) has been introduced for current control [38]–[40], power control [41], [42], and torque and flux control from a voltage-source inverter [43], [44]. A quality function is evaluated, based on predictions from a model of the system, over a finite receding horizon. No modulation or linear controllers are required.

In this paper, the MPC approach is applied to obtain a method to control the current of an induction machine and, at the same time, allow the control of the input current and reactive power to the system, in order to obtain low-distortion input currents, controlled PF, and excellent drive performance even if the energy source presents disturbances.

II. MC MODULATION AND CONTROL

A. MC and Power Circuit of the System

The power circuit of the drive under study is shown in Fig. 1. A three-phase induction machine, represented in the figure as a three-phase resistive-inductive-active load, is fed by a \(3 \times 3\) MC. With this load model, it is also possible to characterize a wide range of applications, including passive loads and grid-connected converters, even though, throughout this document, the system is used as drive for an induction machine.

The converter directly connects each output line to one of the three input phases, by means of an array of bidirectional switches. Those switches, as shown in Fig. 1, are built based on diodes and insulated gate bipolar transistors (IGBTs) in a common-emitter configuration, although other alternatives are
feasible to be implemented [1]. The presented drive allows power flow from and to the mains. The converter is connected to the mains through an input filter that avoids overvoltages and filters high harmonic content in the input current [1]–[4].

Three-phase variables are characterized as complex vectors by means of a 2-D representation. Throughout this paper, this representation is considered when vectors are used. From Fig. 1, the output or load voltage space vector and the input current space vector are defined as

\[
\mathbf{v}_\alpha = \frac{2}{3}(v_u + \alpha v_b + \alpha^2 v_c)
\]

\[
\mathbf{i}_e = \frac{2}{3}(i_{eu} + \alpha i_{ev} + \alpha^2 i_{ew})
\]

where \(\alpha = e^{j(2\pi/3)}\), \(v_x\)’s, with \(x \in \{u, v, w\}\), are output phase voltages, and \(i_{xy}\)’s, with \(x \in \{u, v, w\}\), are input phase currents of the MC. A similar definition can be applied to obtain the source current vector \(i_x\), the source voltage vector \(v_s\), and the load or output current vector \(i_o\).

In Fig. 1, bidirectional switches are associated with variables defined as \(S_{xy}\) with \(x \in \{u, v, w\}\) and \(y \in \{a, b, c\}\). \(S_{xy}\) is also known as switching function for the switch \(xy\). \(S_{xy} = 1\) implies that the switch \(xy\) is on, closed, or conducting, and \(S_{xy} = 0\) implies that the switch \(xy\) is off, open, or blocking. Taking into account that the load should not be in an open circuit, due to its inductive nature, and that phases of the source should not be connected in short circuit, switching functions should at all times fulfill the following equation:

\[
S_{ux} + S_{uy} + S_{wy} = 1 \quad \forall y \in \{a, b, c\}.
\]

The previous restriction allows this topology to have 27 valid switching states. These 27 switching states are classified into three groups, according to the kind of output voltage and input current vector that each switching state generates.

1) All three output phases connected to the same input phase. Space vectors from this group have zero amplitude and will be identified by the symbol ♦.

2) Two output phases connected to a common input phase and the third connected to a different input phase. This group generates stationary space vectors with varying amplitude and fixed direction, identified by □.

3) Each output phase connected to a different input phase. Space vectors have constant amplitude, but its angle varies at the supply angular frequency. The symbol to identify vectors from this group will be □.

To expose graphically the behavior of these space vectors, Fig. 2 shows the trajectories of all output voltage space vectors feasible to be generated by a \(3 \times 3\) MC under balanced three-phase input voltages of 230 V rms per phase.

Most control methods developed for MCs consider only switching states within groups 1 and 2, leaving aside group 3. That is the case of modulation strategies, like most space vector modulation (SVM) methods [1], [6], [8]–[11], of machine control techniques that use these modulation strategies [4], [14] and of direct torque control of induction machines fed by MCs [13]. The reason for that is the difficulty of dealing with rotating vectors with those control techniques. This issue has been overcome by modulation schemes based on detecting the location of the input voltage and output current vectors, selecting the switching state accordingly [45]. The method presented in this paper considers all 27 valid switching states, including the ones that generate rotating vectors, by means of a conceptually simple strategy.

B. Control Scheme

The proposed control strategy counts with a predictive stage that performs predictive current control (PCC) and a classic stage that handles speed, flux, and torque control by means of field oriented control (FOC) [38], which generates reference currents for the predictive stage. The block diagram of the entire control strategy is shown in Fig. 3.

A proportional-integral (PI) controller receives the speed error and generates the reference torque. The reference amplitude of the stator flux and reference torque are used, by means
of FOC, to generate the output reference current to the PCC segment of the control strategy, emphasized in Fig. 3. The PCC strategy replaces, in this approach, linear current controllers and modulation techniques of classic methods [4], [14]. The predictive algorithm handles with the objectives of controlling and modulation techniques of classic methods [4], [14]. The strategy replaces, in this approach, linear current controllers of FOC, to generate the output reference current to the PCC.

![Block diagram of the control strategy.](image)

III. PREDICTIVE CURRENT CONTROL (PCC)

Power converters have a certain number of valid switching states. For a three-phase inverter, it is possible to apply eight different switching states [39]. On an MC, there are 27 valid combinations [1]. The PCC method discussed in this paper consists of choosing, at fixed sampling intervals, the best possible switching state of the converter, based on an evaluation criterion and predictions of the behavior of the system. For that purpose, the algorithm performs a quality function minimization by means of predictions of variables obtained from a model of the system. The nonlinear optimization problem is solved in real time by means of an exhaustive search process, i.e., simply evaluating the quality function for each of the 27 valid switching states. Briefly, the method presents two main aspects: the model, from which predictions are obtained, and a quality function that represents the evaluation criterion.

A. Model of the System

To predict the behavior of the system, the method requires models. It is necessary to evaluate the output voltage and input current that each switching state would produce and the future effect of those variables in the output current and reactive input power.

1) Model of the MC: Referenced to a common point, the relation between load or output voltages and input voltages of the MC can be expressed as

$$\begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} = \begin{bmatrix} S_{ua} & S_{va} & S_{wa} \\ S_{ub} & S_{vb} & S_{wb} \\ S_{uc} & S_{vc} & S_{wc} \end{bmatrix} \begin{bmatrix} v_{eu} \\ v_{ev} \end{bmatrix}. \tag{4}$$

Output voltages applied to the load can be considered then as dependent variables of the switching functions, present in matrix $T$, and the input voltages. The relation between input currents and output or load currents is expressed as

$$\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \begin{bmatrix} S_{ua} & S_{va} & S_{wa} \\ S_{ub} & S_{vb} & S_{wb} \\ S_{uc} & S_{vc} & S_{wc} \end{bmatrix} \begin{bmatrix} i_e \end{bmatrix} \tag{5}$$

Due to the inductive nature of the load, output or load currents will present slow dynamics. That is the reason to consider the dependence implicit in (5). In other words, input currents can be considered to depend on the output currents and the switching state of the MC.

2) Model of the Load: The objective in this case is to obtain an equation to predict the future value of the load current in the next sampling instant for each possible switching state. The load model is based on the equation of a three-phase resistive-inductive-active load (Fig. 1), which fulfills

$$L \frac{di_o(t)}{dt} = v_o(t) - Ri_o(t) - e(t) \tag{6}$$

where $R$ and $L$ are the load resistance and inductance, respectively, and $e$ is the electromotive force (EMF) vector of the load. This load model covers a wide variety of possible applications, including passive loads and grid-connected converters. The attention is placed on an induction machine. The following analysis of the dynamic equations of an induction machine [46] will be presented in order to relate it with the model expressed in (6). The machine’s stator and rotor voltage equations in fixed stator reference frame can be written as

$$v_o(t) = R_s i_o(t) + \frac{d\psi_s(t)}{dt} \tag{7}$$

$$v_r(t) = R_r i_r(t) + \frac{d\psi_r(t)}{dt} - j\omega \psi_n(t) = 0 \tag{8}$$

where $R_s$ and $R_r$ are the stator and rotor resistances, $\psi_s$ and $\psi_r$ are the stator and rotor fluxes, and $\omega$ is the rotor angular frequency. The stator and rotor fluxes are related with the stator and rotor currents by

$$\psi_s(t) = L_s i_o(t) + L_m i_r(t) \tag{9}$$

$$\psi_r(t) = L_m i_o(t) + L_s i_r(t) \tag{10}$$

where $L_s$, $L_r$, and $L_m$ are the motor’s self- and mutual inductances. These machine’s equations and model are widely used in literature [12], [14]–[17], [43], [44]. Replacing (9) in (7), we obtain

$$L_s \frac{di_o(t)}{dt} = v_o(t) - R_s i_o(t) - L_m \frac{di_r(t)}{dt}. \tag{11}$$
Comparing (11) with the presented load model (6), it is possible to observe that, considering an induction machine as load, \( L = L_s \) (stator self-inductance), \( R = R_s \) (stator resistance), and the EMF or back-EMF of the machine is generated by the interaction with the rotor through the mutual inductance, thus \( e(t) = L_m(t) (d\lambda(t))/dt \). This analysis connects the well-known dynamic equations of an induction machine to the load model presented in (6).

Applying a sampling period \( T_s \), the derivative form \( d\lambda(t)/dt \) can be approximated by

\[
\frac{d\lambda(t)}{dt} \approx \frac{i_\lambda(k) - i_\lambda(k-1)}{T_s}.
\]

Replacing (12) in (6) and shifting the discrete time one step forward, the relation between the discrete-time variables can be described as

\[
i_\lambda(k+1) = \frac{T_s}{T_s + L} L_i_\lambda(k) + v_\lambda(k+1) - e(k+1).
\]

Equation (13) is used to obtain predictions for the future value of the load current \( i(k+1) \) for each voltage vector \( v_\lambda(k+1) \) generated by valid switching states. The corresponding voltage vector for each switching state can be calculated by means of (4) and (1) evaluated at time \( k+1 \).

The future load back-EMF \( e(k+1) \) can be estimated using a second-order extrapolation from present and past values or considering \( e(k+1) \approx e(k) \), depending on the sampling time and the platform used for implementation. For a sufficiently small sampling time, no extrapolation is needed. Present and past estimations of \( e \) can be obtained from the load equation (13) shifted backward in time as

\[
\hat{e}(k) = v_\lambda(k) + \frac{L}{T_s} i_\lambda(k-1) - \frac{RT_s + L}{T_s} i_\lambda(k).
\]

3) Model of the Input Filter: The input filter model, based on Fig. 1, can be described by the following continuous-time equations:

\[
v_\alpha(t) = Rf i_\alpha(t) + Lf \frac{di_\alpha(t)}{dt} + v_e(t)
\]

\[
i_\alpha(t) = i_\alpha(t) + C_f \frac{dv_e(t)}{dt}
\]

where \( L_f \) and \( R_f \) are the joint inductance and resistance of the line and filter, respectively, and \( C_f \) is the filter’s capacitance. This continuous-time system can be rewritten as

\[
\begin{bmatrix}
0 & \frac{1}{L_f} \\
-1 & -R_f/L_f
\end{bmatrix}
\begin{bmatrix}
0 \\
1/L_f
\end{bmatrix}
\begin{bmatrix}
v_\alpha(t) \\
i_\alpha(t)
\end{bmatrix}
+ \begin{bmatrix}
0 \\
1/L_f
\end{bmatrix}
\begin{bmatrix}
u_e(t) \\
i_e(t)
\end{bmatrix}
= \begin{bmatrix}
\frac{1}{L_f} \\
0
\end{bmatrix}
u(t)
\]

\[
A_c x(t) + B_c u(t) = \begin{bmatrix}
v_\alpha(t) \\
i_\alpha(t)
\end{bmatrix}
\]

A discrete state space model can be derived when a zero-order hold input is applied to a continuous-time system described in state space form as (17). Considering a sampling period \( T_s \), the discrete-time system derived from (17) is

\[
x(k+1) = A_q x(k) + B_q u(k)
\]

with

\[
A_q = e^{A_c T_s} \\
B_q = \int_0^{T_s} e^{A_c (T_s - t)} B_c dt.
\]

For more details about sampled-data systems and the theory employed in this analysis, see [47, ch. 12, pp. 340 and 341]. The discrete-time variables will match the continuous-time variables at sampling intervals. A convenient way of obtaining the discrete model is through the MATLAB’s function \( e^{2d()} \): conversion of continuous-time models to discrete time. To predict the mains’ current, it is necessary simply to solve \( i_\alpha(k+1) \) from (19)

\[
i_\alpha(k+1) = A_q(2,1)v_e(k) + A_q(2,2)i_\alpha(k) + B_q(2,1)v_\alpha(k) + B_q(2,2)i_\alpha(k).
\]

At this point, the method counts with a model to predict the value of \( i_\alpha \) depending on \( i_\alpha \). Equations (5) and (2) must be used to calculate \( i_\alpha \) for each switching state.

To analyze the resulting effect on the reactive input power, it is necessary to consider the instantaneous power theory [48], [49]. The instantaneous reactive input power can be predicted, based on predictions of the input current, as

\[
Q(k+1) = \text{Im}\{v_\alpha(k+1)\bar{i}_\alpha(k+1)\} = v_{\alpha \beta}(k+1)i_{\alpha \beta}(k+1) - v_{\alpha \alpha}(k+1)i_{\alpha \alpha}(k+1)
\]

where \( \bar{i}_\alpha \) is the complex conjugate of vector \( i_\alpha \) and the subscripts \( \alpha \) and \( \beta \) represent real and imaginary components of the associated vector. Line voltages are low-frequency signals. Based on that, the method considers \( v_\alpha(k+1) \approx v_\alpha(k) \).

B. Evaluation Criterion: The Quality Function

The quality function \( g \) represents the evaluation criterion from which the method determines which switching state is the optimum to be applied during the next sampling time. The optimization process is performed by evaluating \( g \) for each valid state. However, prior to determining the equation that will represent the quality function, it is necessary to define the objectives that the converter must achieve. The MC must feed the load with currents close to the reference value and, at the same time, allow the control of the input current in order to have low harmonic distortion and regulated PF. Several other objectives can be included in the quality function, owing to the versatility of the presented approach, opening a wide range of possibilities for further research. To mention some examples, other objectives considered in previous works have been active and reactive power control in an active front end rectifier [41, balance in the dc link for multilevel inverters [40], the reduction of the switching frequency [40], and direct torque and flux control of an induction machine [43].
1) Control of the Output Current: As mentioned, one of the objectives reflected in the quality function must be the tracking of the reference current to the load. Switching states that generate closer values of the output current to the reference should be preferred. That goal is achieved by assigning cost or penalizing differences from the reference value. The alternative proposed to accomplish that objective is

\[ g_1 = |i_{oa}^* - i_{oa}| + |i_{oβ}^* - i_{oβ}|. \]  \hspace{1cm} (23)

Reference values are denoted by the superscript “*,” while predicted values can be identified by the superscript “p” and are obtained by the previously presented models.

2) Regulation of the Input PF: The MC can control, together with output signals, the phase of the input current from the mains. The amplitude is determined by the active power flow, since the MC does not store energy. Most modulation methods reach the objective of working with unity PF by means of relatively complex strategies [6], [11], [23]. In order to work with inductive or capacitive PF, the complexity increases considerably [6], [11], [23]. With the proposed predictive approach, the objective of controlling the reactive power \( Q \) can be easily achieved simply by penalizing switching states that produce predictions of \( Q \) distant from the reference value. That is

\[ g_2 = |Q^* - Q^p|. \]  \hspace{1cm} (24)

The value of the predicted reactive power \( Q^p \) is obtained for each valid state from (22). Its reference value \( Q^* \) is given externally, as stated in Fig. 3, to work with capacitive, inductive, or unity PF. Most applications require unity PF, hence \( Q^* = 0 \) and \( g_2 = |Q^p| \), but the method offers the alternative to control that variable with a very simple approach compared to other modulation strategies [6], [11], [23].

The resulting quality function that reflects both objectives, output current and reactive input power control, is obtained simply by adding \( g_1 \) and \( g_2 \), thus

\[ g = |i_{oa}^* - i_{oa}^p| + |i_{oβ}^* - i_{oβ}^p| + |Q^* - Q^p|. \]  \hspace{1cm} (25)

The weighing factor \( A \) handles the relevance that the designer assigns to each objective. In order to agree with other terms present in \( g \), \( A \) must have \( V^{-1} \) as unit.

A greater value of \( A \) implies higher relevance to control the PF or reactive input power over the output current reference tracking. The criterion to select the value of \( A \) is briefly treated in the Results section.

IV. RESULTS

The proposed control technique was experimentally tested on an 11-kW induction machine fed by an 18-kVA MC. Relevant parameters of the laboratory implementation are presented in Table I.

Prior to the experimental results, the method was simulated in order to verify the capability of the strategy to control the output current and input PF.

<table>
<thead>
<tr>
<th>TABLE I</th>
<th>PARAMETERS OF THE POWER CIRCUIT AND CONTROL METHOD</th>
</tr>
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<tbody>
<tr>
<td>( C_f )</td>
<td>21 ( \mu )F</td>
</tr>
<tr>
<td>( L_f )</td>
<td>300 ( \mu )H</td>
</tr>
<tr>
<td>( R_f )</td>
<td>( \sim ) 0 ( \Omega )</td>
</tr>
<tr>
<td>Input Voltage (rms)</td>
<td>230 V</td>
</tr>
<tr>
<td>Input Frequency</td>
<td>50 Hz</td>
</tr>
<tr>
<td>( A )</td>
<td>0 or 0.0045 ( V^{-1} )</td>
</tr>
<tr>
<td>( T_s )</td>
<td>8 ( \mu )s</td>
</tr>
</tbody>
</table>

Fig. 4. Simulation results: Steady state at 1000 r/min and 30-N·m load torque without the proposed term to control input PF \(( A = 0 \ V^{-1})\).

Fig. 5. Simulation results: Steady state at 1000 r/min and 30-N·m load torque using the proposed term to control input PF \(( A = 0.0045 \ V^{-1})\).

A. Simulation Results

The method was simulated with the parameters exposed in Table I. The parameter \( A \), or reactive input power weighting factor, was tested with two different values, such as 0 and 0.0045 \( V^{-1} \), in order to expose the behavior of the method without and with reactive power control. The focus of this section is placed mainly on output and input currents. Further analysis of the motor variables, the performance working with inductive, unity, and capacitive PF, and the behavior of the system as a drive are included in the Experimental Results section. Simulated results are shown in Figs. 4 and 5 for \( A = 0 \ V^{-1} \) and \( A = 0.0045 \ V^{-1} \), respectively. The machine, running at 1000 r/min with 30-N·m load torque, requires from the MC output currents of 18-A amplitude and 35-Hz frequency. The reference reactive input power \( Q^* \) was set to 0 VAR in order to work with unity PF when \( A = 0.0045 \ V^{-1} \). As observed,
both cases accomplish good output current control, achieving sinusoidal currents to the motor. Input currents present high distortion and a phase shift with respect to input phase voltages when using $A = 0 \ V^{-1}$, as shown in Fig. 4. The presented method accomplishes good input currents, with low ripple and unity PF, when the weighting factor $A$ is set to $0.0045 \ V^{-1}$. A deeper analysis of the behavior of the system will be exposed based on experimental results in the corresponding section.

### B. Selection of Parameter $A$

The $A$ weighing factor is the only parameter from the predictive current controller to be selected by the designer. It replaces PI’s parameters from linear current controllers from traditional control structures. The adjustment of this kind of parameters is still an open topic for research. It is possible to find optimal values in cases where the system presents no constraints and under specific structures of the quality function [28]. For systems with a finite number of control actions, finite input sets, or state-alphabet [50], the proposed method to adjust the parameter is to simply evaluate the performance of specific system’s variables to determine the best value.

For the presented MC-based system, key variables to evaluate the behavior of the PCC method are the total harmonic distortion (THD) of the output and input or mains current. In order to perform the evaluation, an exhaustive search was carried out based on 400 simulations, each with a value of $A$ equidistant within the range $[0, 0.007] \ V^{-1}$. Figs. 4 and 5 show two of these simulations for $A = 0 \ V^{-1}$ and $A = 0.0045 \ V^{-1}$, respectively. As mentioned, the variables to observe and evaluate in order to select the weighing factor are the input and output current THDs. The result of this procedure is extracted in Fig. 6.

As expected, the input current’s THD drastically decreases as $A$ increases, reaching a value close to 5% near $A = 0.002 \ V^{-1}$. No further reduction is significant after that value of $A$. On the other hand, as a tradeoff, the output current’s THD increases as higher importance is placed on the reactive power in the quality function as $A$ increases. Although the output current’s THD is still low within the evaluated range—values from 0.09% to 0.14% are not considered to be high distortion—the ideal situation is to achieve low distortion on both currents. For that reason, the weighing factor $A$ was set at $A = 0.0045 \ V^{-1}$, to select a value far from the region where the input current’s THD drastically increases (under $A = 0.002 \ V^{-1}$) and not higher, in order to ensure low THD on the output current.

### C. Experimental Results

As mentioned, the proposed control technique was experimentally tested on an 11-kW induction machine fed by an 18-kVA MC, considering parameters observed in Table I. The experimental prototype of the MC was built based on 18 IXDN 55N120D1 discrete IGBTs from IXYS in a common-emitter configuration. Both the proposed control method and the FOC (Fig. 3) were implemented on a rapid prototyping board dS1103 from dSPACE, with a sampling period $T_s = 8 \ \mu s$ for the predictive method. Current-controlled commutation logic is performed by means of an Altera EPM 9320 complex programmable logic device, which generates the gate signals for the 18 IGBTs. External torque was applied from a dc machine.

In the first place, the system was tested under identical conditions as exposed in simulations, with the induction machine running in steady state at $1000 \ r/min$ with $30 \ N \cdot m$ load torque from the dc machine. Results can be observed in Fig. 7, for $A = 0 \ V^{-1}$, and Fig. 8 for $A = 0.0045 \ V^{-1}$.

Achieved experimental results confirm the observations made based on simulations. The method performs good current control to the load and significantly improves the input current control when considering $A = 0.0045 \ V^{-1}$, achieving almost sinusoidal currents from the mains and unity PF, as shown in Fig. 8. There is a remarkable resemblance between simulations and experimental results for both values of $A$. A slight
difference might be observed in the output voltage waveforms, due to expected and reasonable switching noise in the experimental implementation.

A theoretical limit for the maximum switching frequency of a particular IGBT within any bidirectional switch working with the presented method is given by the following equation:

\[
f_{\text{IGBT, max}} = \frac{1}{6} f_s = \frac{125 \text{ kHz}}{6} = 20.83 \text{ kHz}
\]  

where \( f_s = (1/T_s) \) is the sampling frequency. Note that this is a theoretical limit for the switching frequency of an IGBT [38]. In practice, most times, the method selects switching states that do not involve the maximum number of IGBTs’ commutations. As a result, the measured average switching frequency in both cases (Figs. 7 and 8) was significantly lower: 13.2 kHz.

Subsequently, the drive was tested under a speed reversal, changing the reference speed from 1000 to \(-1000\) r/min. The achieved measurements can be observed in Figs. 9 and 10. The reference reactive input power \( Q^* \) was set to 0 VAR, and the weighting factor \( A \) was set at 0 V \(^{-1}\) (Fig. 9) and 0.0045 V \(^{-1}\) (Fig. 10). Both cases presented similar behavior regarding output variables, including smooth speed and current–frequency transition.

Concerning the relation with the mains, in Fig. 9 \( A = 0 \text{ V}^{-1} \), the input current presents high distortion and different phase with the related input voltage. The ripple and distortion observed in the input current change during the speed reversal performed by the drive. This fact can also be observed in the behavior of the reactive input power \( Q \), included in Fig. 9.

In contrast, setting \( A = 0.0045 \text{ V}^{-1} \), the method achieves almost sinusoidal input current in phase with the source voltage during motoring operation of the drive (Fig. 10). The speed reversal performed also demands the drive to regenerate energy as the speed is reduced and the system decreases its kinetic energy. It can be observed that the method achieves, in those cases, sinusoidal currents with \( \pi \text{ rad} \) phase shift with the phase voltage, making the energy flow from the motor to the mains (regenerating) possible. The improved performance in terms of the quality of input currents can also be observed in the reactive input power, considerably lower than the obtained with \( A = 0 \text{ V}^{-1} \).

The next test performed in order to evaluate the behavior of the drive was to set the reference speed at 1000 r/min and apply an accelerating torque of 30 Nm from the dc machine, forcing regeneration from the induction machine and the MC.
The method was tested with $A = 0$ V$^{-1}$ and $A = 0.0045$ V$^{-1}$, and results can be observed in Figs. 11 and 12, respectively. Both cases presented good control of the induction machine, maintaining the speed at the reference value and with low-ripple output currents. The difference is evident at input variables. While regeneration took place in both situations, using $A = 0.0045$ V$^{-1}$, the method achieved almost sinusoidal input currents in opposite phase with the associated phase voltage, presenting therefore a considerably better performance of the reactive input power.

An interesting feature of the proposed method is that it can maintain an appropriate control of the output variables even if the energy source presents disturbances. This characteristic is not trivial, considering that the MC has no energy storage. It is possible to find several works directed to achieve good performance of the MC for other modulation techniques working under polluted ac voltage supplies [21]–[27]. The predictive method achieves this objective without further modifications. To probe the performance of the control technique under abnormal supply, the MC was fed from unbalanced mains, with a 54-V difference in the amplitude of one phase. Two of the voltage input phases applied presented an rms value of 230 V, and the third presented an rms value of 176 V. The parameter $A$ was set to 0.0045 V$^{-1}$ in order to minimize the reactive input power, while the speed reference was set to 1000 r/min, and a load torque of 10 N·m was applied from a dc machine. Results can be observed in Fig. 13.

The output variables do not present signs of the distortion on the mains. The speed is successfully controlled at 1000 r/min, with very low torque ripple and almost sinusoidal output currents to the motor. In Fig. 13, it is also possible to observe the reduced input phase voltage and one of the unmodified voltages. The input current of the unbalanced phase presented low-frequency distortion but very low high-frequency distortion or ripple. All input currents maintained the phase of the corresponding phase voltage.

Finally, the capability of the method to control the input PF was tested, changing the value of $Q^*$. The reference speed was set to 1000 r/min, and the parameter $A$ was set to 0.0045 V$^{-1}$. A load torque of 10 N·m was applied from a dc machine. $Q^*$ was set at +900 V AR, achieving the results shown in Fig. 14.

Results for $Q^* = 0$ VAR (unity PF) can be observed in Figs. 5, 8, 10, and 12. No difference is observed in terms of the output behavior of the converter. The main difference, as expected, is observed in the behavior of the input current, presenting a capacitive PF of 0.81. Consequently, it is possible to control the input PF by changing the value of $Q^*$.

The effect of errors and uncertainties in the load’s parameters using PCC is treated in [39]. It is worth to mention that knowledge of the load parameters is a requirement on any control method in order to tune linear controllers and select the modulation frequency. PCC is not an exception in this sense. The presented experimental results show its effectiveness as control method applied to an MC-based motor drive with different load conditions.
V. CONCLUSION

The presented method for current control of an induction machine fed by an MC effectively controls the output current from the MC to the machine and the reactive input current to the system. The proposed strategy allows the regulation of the input PF by means of a simple and straightforward technique, controlling the phase of the input current in a way that the converter can work with capacitive, unity, or inductive PF, according to the requirements of the application. The proposed control method replaces a modulation stage—carrier-based, SVM or others—and the current controller in a classic control scheme. Joint with FOC, the strategy allows for an excellent control of an induction machine together with the input currents from the mains. The method was tested on speed reversals, different load conditions, and regenerating, revealing good performance in all cases. Moreover, the strategy was tested under unbalanced mains, presenting a satisfactory behavior without any further modification. Output variables, as currents to the machine and speed, remained unchanged when the unbalance is introduced to the grid. The current control strategy can also be applied to other loads that fulfill or can be represented by the proposed model.

The method can be easily implemented, taking advantage of the present technologies available in digital signal processors. The high sampling frequency required should not be a problem nowadays and even less in years to come. This control strategy uses, in a convenient way, the discrete nature of power converters and microprocessors used in their control. These results show that predictive control is a very powerful tool, with a conceptually different approach, that opens new possibilities in the control of power converters.

REFERENCES


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