# Predictive Control of an Indirect Matrix Converter

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***Abstract*—This paper presents the implementation of a predictive control scheme for an indirect matrix converter. The control scheme selects the switching state that minimizes the reactive power and the error in the output currents according to their reference values. This is accomplished by using a prediction horizon of one sample time and a very intuitive control law. Experimental results with a 6.8-kVA indirect matrix converter prototype are provided in order to validate the proposed control scheme. The converter uses standard digital signal processor operating at a sampling frequency of 20 $\mu$s. It is shown that the idea of controlling this converter topology with a predictive approach can be implemented simply and input currents with unity power factor and a total harmonic distortion lower than 5% can be obtained.**

***Index Terms*—AC motor drives, current control, predictive control.**

## I. Introduction

MATRIX converters feature several advantages compared with standard two-level converters, such as a bidirectional power flow and the reduced size of the reactive components [1]. Different kinds of converters without dc link have been presented in the technical literature. They are classified into three main groups: 1) the cycloconverters in the high-power range; 2) the standard matrix converters; and 3) indirect matrix converters [2], all of them in the low-power range, as it is shown in Fig. 1. The conventional indirect matrix converter is similar to a back-to-back inverter but includes bidirectional switches in the rectifier and has no dc-link capacitor, as shown in Fig. 2. References to this topology can be traced back to 20 years ago, when the concept was first introduced using a converter with gate turnoff (GTO) thyristors [3].

Over the last few years, research on indirect matrix converters has benefited from advances in semiconductor technologies, which have mainly contributed toward enhancing efficiency. Compared with the standard matrix converter, this topology can use a simpler modulation scheme and does not need an additional overvoltage protection system. Conversely, the current path from input to output produces higher power losses. This can be partly mitigated by the use of semiconductors such as reverse blocking insulated-gate bipolar transistors (RB-IGBTs) [4], which have already been used in conventional matrix converters [5], [6]. Additionally, some studies have focused on sparse matrix topologies, which can improve power density by reducing the number of semiconductors at the expense of functionality. This is the case of the sparse matrix [7], [8], the very sparse matrix [9], and the ultrasparse matrix converter [10], which feature 15, 12, and 9 switches, respectively.

Indirect matrix converters use complex pulsewidth modulation (PWM) schemes to achieve the goal of unity power factor and sinusoidal output current. However, since power converters have a discrete nature, the application of predictive control constitutes a promising and better suited approach as compared with standard schemes that use mean values of the variables [11]. Furthermore, predictive control utilizes a very intuitive control law that can easily deal with multivariable cases and the treatment of constraints and compensate for the dead time [12].

Early practical applications of predictive control can be found in the 1980s for the case of two-level inverters [13]. Nowadays, it is possible to find applications in machine control [14], [15], active rectifiers [16], [17], matrix converters [18], [19], and even multilevel converters [20], [21]. Among the broad family of predictive control methods, those with a prediction horizon of one sample time are particularly attractive due to the simplicity of implementation. More advanced control techniques with a prediction horizon greater than one have been proposed for simple converter topologies such as dc–dc converters and two-level inverters [22], [23]. These types of schemes offer some advantages in terms of faster dynamics and reduced harmonic distortion at the expense of an increased calculation burden. In order to solve this problem, some methods have been proposed to move part of the calculations' effort offline [23]. Unfortunately, matrix converters operate at a very fast sampling rate and feature a high number of possible switching states.

This paper presents an application of predictive control in indirect matrix converters. The switching state is selected by minimizing a quality function that considers the instantaneous reactive power in the input, the current error in the output, and the generation of a positive voltage in the dc link. Feasibility, implementation details, and advantages and disadvantages are also discussed. Due to the limited capacity of the controller, the use of a prediction horizon of one sample time will be a design criterion. This paper is organized in the following manner. Section II presents the converter topology. In Sections III and IV, the control scheme is explained in detail. Section V describes the commutation sequence for the rectifier and the

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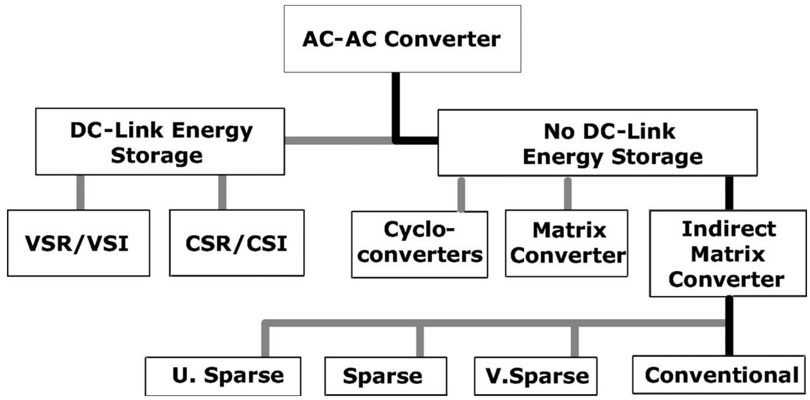


Fig. 1. Classification of power converters.

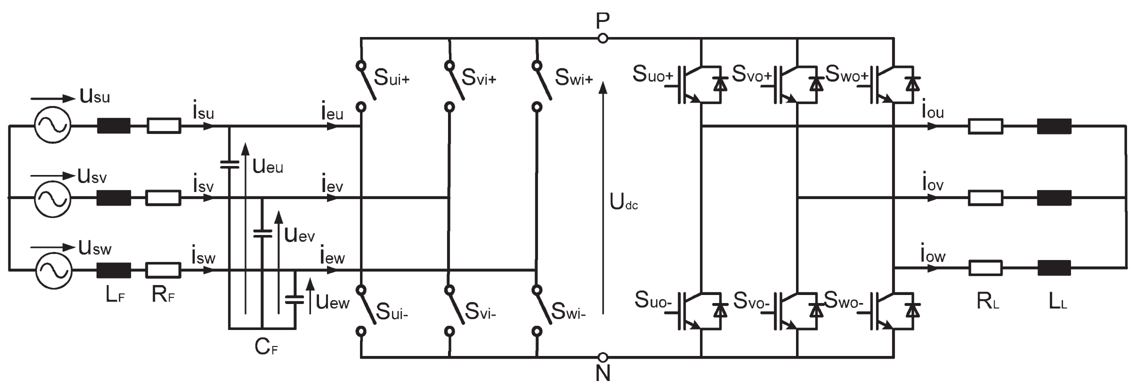


Fig. 2. Simplified scheme of the indirect matrix converter.

inverter. Section VI describes the selection of the weight factors for the predictive controller. Finally, the experimental setup and results are explained, showing the feasibility of the proposed method in Section VII.

## II. Power Circuit of the Indirect Matrix Converter

The indirect matrix converter consists of an array of power semiconductors that is very similar to the ac/dc/ac back-to-back converter, as shown in Fig. 2. The converter synthesizes a positive voltage in the dc link by selecting a switching state in the rectifier that connects one phase to point $P$ and the other phase to point $N$. Additionally, the rectifier includes an $LC$ filter in the input which is needed to provide a path for the phase current which is momentarily not connected to the dc link. It should be noted that the indirect matrix converter topology includes as many switches as the standard matrix converter, but the former features an extra freedom degree that alleviates the complexity of the commutation sequence.

An indirect matrix converter with RB-IGBTs is considered in this paper. The conduction losses of this kind of semiconductor are lower than a bidirectional switch consisting of a standard IGBT and a diode connected in series [7], [8].

## III. Predictive Control Method

In the following, it will be assumed that the three phase quantities of the converter are symmetrical and, hence, can be represented by the well-known 2-D space vector. For example, the phase current components $i_{\mathrm{su}}$, $i_{\mathrm{sv}}$, and $i_{\mathrm{sw}}$ will be described by the complex space vector

$$\underline{i}_{\mathrm{s}} = i_{\mathrm{s}\alpha} + j \cdot i_{\mathrm{s}\beta} \tag{1}$$

which is defined as

$$\left.\begin{aligned} i_{\mathrm{s}\alpha} &= \tfrac{1}{3}(2i_{\mathrm{su}} - i_{\mathrm{sv}} - i_{\mathrm{sw}}) \\ i_{\mathrm{s}\beta} &= \tfrac{1}{\sqrt{3}}(i_{\mathrm{sv}} - i_{\mathrm{sw}}) \end{aligned}\right\}. \tag{2}$$

This space vector is referred to a stationary reference frame that will be denoted as an $\alpha\beta$ reference frame.

Predictive control aims to select the converter switching state that leads the controlled variables closest to their respective references at the end of the sampling period. In order to meet this requirement, the load current and input voltages are measured, and predicted values of the input and output currents are generated for each possible switching state.

Three main conditions must be fulfilled for the converter to properly operate. First, the line side of the rectifier must deliver active power. Second, the load current must follow the reference with high accuracy, and third, the dc-link voltage must be positive. This last condition differentiates the proposed control law from the method presented in [18]. The first condition is accomplished by minimizing the predicted value of the instantaneous reactive power

$$q^{k+1} = \left| u_{\mathrm{s}\alpha}^{k+1} \cdot i_{\mathrm{s}\beta}^{k+1} - u_{\mathrm{s}\beta}^{k+1} \cdot i_{\mathrm{s}\alpha}^{k+1} \right|. \tag{3}$$

On the other hand, the second condition requires a minimum error between the predicted load currents and its references

$$\Delta i_{1}^{k+1} = \left|i_{1\alpha}^{*} - i_{1\alpha}^{k+1}\right| + \left|i_{1\beta}^{*} - i_{1\beta}^{k+1}\right| \tag{4}$$

where $i_{1\alpha}$ and $i_{1\beta}$ denote the load current in $\alpha\beta$ coordinates and $i_{1\alpha}^{*}$ and $i_{1\beta}^{*}$ denote their respective references.

The third condition necessary for the operation of the rectifier is to discard those states that produce negative dc-link voltages. In order to fulfill this condition, the following function is defined:

$$h^{k+1} = \begin{cases} 0, & u_{\mathrm{dc}}^{k+1} > 0 \\ M, & u_{\mathrm{dc}}^{k+1} \leq 0 \end{cases} \tag{5}$$

where $M$ is the maximum positive number that can be generated by the arithmetic unit of the controller. Equations (3)–(5) are merged into a single so-called quality function

$$g^{k+1} = q^{k+1} + A \cdot \Delta i_{1}^{k+1} + h^{k+1}. \tag{6}$$

The control method operates as follows. At each sampling time, all possible switching states are used to calculate (6). The switching state that produces the minimum value of $g$ is selected to be applied for one sampling period.

## IV. Calculation of Predicted Values

A mathematical model of the input filter and the load provides the basis for the prediction of the values of the input and output currents, which are needed for evaluating the quality function.

The line side of the rectifier consists of a second-order system described by

$$L_{\mathrm{f}} \frac{d\underline{i_{\mathrm{s}}}}{dt} = \underline{u_{\mathrm{s}}} - \underline{u_{\mathrm{e}}} - R_{\mathrm{f}} \cdot \underline{i_{\mathrm{s}}} \tag{7}$$

$$C_{\mathrm{f}} \frac{d\underline{u_{\mathrm{e}}}}{dt} = \underline{i_{\mathrm{s}}} - \underline{i_{\mathrm{e}}} \tag{8}$$

where $L_{\mathrm{f}}$ comprises the mains and filter inductances and $R_{\mathrm{f}}$ represents the mains and filter damping resistances. The prediction of the input current and capacitor voltages are computed from a first-order difference equation, as is described in [18]

$$\underline{i_{\mathrm{s}}^{k+1}} = c_{1i} \cdot \underline{u_{\mathrm{s}}^{k}} + c_{2i} \cdot \underline{u_{\mathrm{e}}^{k}} + c_{3i} \cdot \underline{i_{\mathrm{s}}^{k}} + c_{4i} \cdot \underline{i_{\mathrm{e}}^{k}} \tag{9}$$

$$\underline{u_{\mathrm{s}}^{k+1}} = c_{1u} \cdot \underline{u_{\mathrm{s}}^{k}} + c_{2u} \cdot \underline{u_{\mathrm{e}}^{k}} + c_{3u} \cdot \underline{i_{\mathrm{s}}^{k}} + c_{4u} \cdot \underline{i_{\mathrm{e}}^{k}}. \tag{10}$$

The real coefficients $c_1$, $c_2$, $c_3$, and $c_4$ are defined so that the obtained values for the predicted currents correspond to those of the continuous-time system after one sampling time. They can be calculated by representing (7) and (8) by a state-space system with state variables $\underline{i}_{\mathrm{s}}$ and $\underline{u}_{\mathrm{e}}$

$$\begin{bmatrix} \underline{\dot{u}_{\mathrm{e}}} \\ \underline{\dot{i}_{\mathrm{s}}} \end{bmatrix} = \begin{bmatrix} 0 & 1/C_{\mathrm{f}} \\ -1/L_{\mathrm{f}} & R_{\mathrm{f}} \end{bmatrix} \begin{bmatrix} \underline{u_{\mathrm{e}}} \\ \underline{i_{\mathrm{s}}} \end{bmatrix} + \begin{bmatrix} 0 & -1/C_{\mathrm{f}} \\ 1/L_{\mathrm{f}} & 0 \end{bmatrix} \begin{bmatrix} \underline{u_{\mathrm{s}}} \\ \underline{i_{\mathrm{e}}} \end{bmatrix} \tag{11}$$

and expressing the system in its discrete form as

$$\begin{bmatrix} \underline{u_{\mathrm{e}}}^{k+1} \\ \underline{i_{\mathrm{s}}}^{k+1} \end{bmatrix} = \Phi \begin{bmatrix} \underline{u_{\mathrm{e}}}^{k} \\ \underline{i_{\mathrm{s}}}^{k} \end{bmatrix} + \Gamma \begin{bmatrix} \underline{u_{\mathrm{s}}}^{k} \\ \underline{i_{\mathrm{e}}}^{k} \end{bmatrix} \tag{12}$$

where

$$\Phi = e^{A \cdot T_{\mathrm{s}}} \quad \Gamma = A^{-1}(\Phi - I)B \tag{13}$$

with

$$A = \begin{bmatrix} 0 & 1/C_{\mathrm{f}} \\ -1/L_{\mathrm{f}} & R_{\mathrm{f}} \end{bmatrix} \quad B = \begin{bmatrix} 0 & -1/C_{\mathrm{f}} \\ 1/L_{\mathrm{f}} & 0 \end{bmatrix}. \tag{14}$$

The load model is obtained similarly. Assuming an inductive–resistive load as shown in Fig. 2, after representing the three-phase system in $\alpha{-}\beta$ coordinates, the following equation describes the behavior of the load:

$$L_{1} \frac{d\underline{i_{1}}}{dt} = \underline{u_{1}} - R_{1} \cdot \underline{i_{1}} \tag{15}$$

which is discretized as follows:

$$\underline{i_{1}^{k+1}} = d_{1}\underline{u_{1}^{k}} + d_{2}\underline{i_{1}^{k}} \tag{16}$$

where $d_1$ and $d_2$ depend on $L_1$, $R_1$, and the sample time.

## V. Dead-Time Compensation

The previously explained method does not deal with the dead time generated by the calculations. In fact, the inverter and the rectifier can generate eight and nine valid switching states, respectively, which, altogether, produce 72 possible switching combinations that must be taken into account in the quality function $g$. A simplification is possible if $h$ is *a priori* evaluated using each of the nine possible rectifier states and the present voltage values of the capacitor filter. In this way, three of the nine possible rectifier states are selected beforehand, and just 24 switching combinations have to be taken into account for the evaluation of $g$. The dead time associated to these calculations cannot be neglected if a standard low-cost digital signal processor (DSP) controller is used.

A simple way to consider the dead time is by calculating the quality function at the end of the next sampling period, i.e., $g^{k+2}$. Thus, the selected switching state can be applied at $t^{k+1}$, and a period of time equivalent to one sampling period is available for calculations. The aforementioned compensation requires the calculation of $u_{\mathrm{s}}^{k+1}$, $u_{\mathrm{e}}^{k+1}$, $i_{\mathrm{s}}^{k+1}$, and $i_{1}^{k+1}$ in order to have the basis for the calculation of $g^{k+2}$. These terms are obtained out of (9), (10), (16), and the current switching state $S^k$. Considering that the change in the input voltage is small in one sampling time, $u_{\mathrm{s}}^{k+1}$ will be considered equal to $u_{\mathrm{s}}^{k}$. A block diagram of the control scheme and the sequence of events for the dead-time compensation are shown in Fig. 3(a) and (b), respectively.

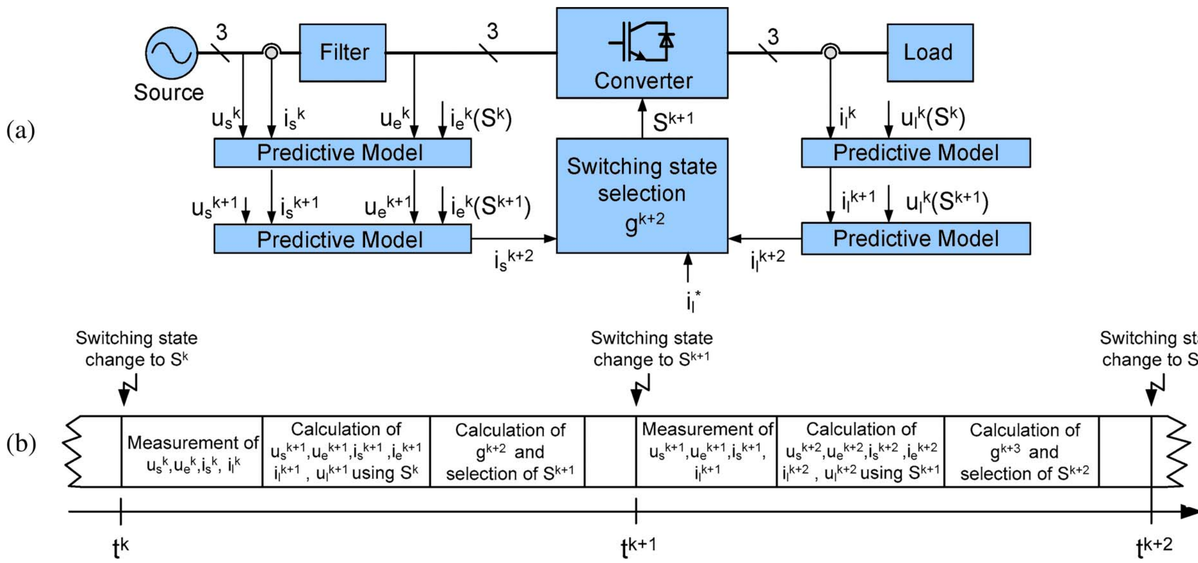


Fig. 3. Predictive controller. (a) Block diagram. (b) Time diagram of the algorithm.

## VI. Commutation Sequence of the Power Semiconductors

The converter requires a commutation sequence that allows a safe change of the rectifier switching state. Basically, this problem can be addressed by synchronizing the state changes in the rectifier with the application of a zero-voltage space vector in the inverter stage. Under this condition, no current circulates through the dc link, and the rectifier state can be changed without help of auxiliary commutation circuits [3].

An exact determination of the switching frequency with the proposed scheme is not possible, considering the nondeterministic generation of switching pulses. In order to estimate the mean switching frequency of the switches of the inverter, the following assumptions were considered.

1) The rectifier stage changes at each sampling time.
2) If the inverter stage changes between two active vectors, i.e., those inverter stages that deliver to load a voltage different than zero, one extra switching in one phase is needed to provide the safe commutation of the rectifier. This is the case in the time instant $t^{k+3}$ in Fig. 4(a).
3) If the inverter stage has a zero space vector state or the last state was a zero vector, no extra switchings are necessary for the safe commutation of the rectifier. This is the case in the time instants $t_{k+1} y t^{k+2}$.

The characteristic curve shown in Fig. 4(b) depicts the mean switching frequency $\bar{f}_{\mathrm{igbt\_o}}$ of one of the IGBTs of the inverter stage, obtained by counting the number of switching transitions in several periods of the fundamental frequency. Other switches present the same behavior, which is why the curve corresponding to one semiconductor is shown. It can be observed that the mean switching frequency is lower than the sampling frequency and it decreases at a higher load current.

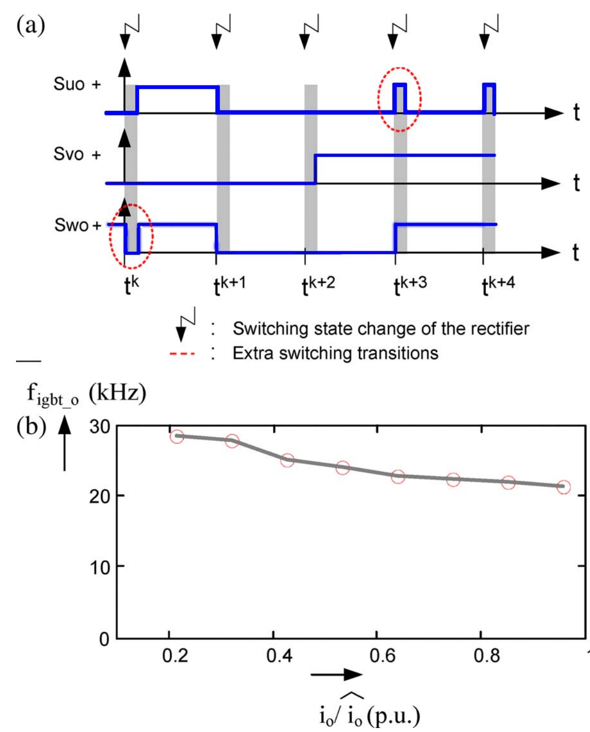


Fig. 4. Switching behavior of the inverter stage. (a) Commutation sequence of the inverter and extra switching transitions. (b) Mean switching frequency of one of the IGBTs of the inverter stage for a sampling time of 20 $\mu$s.

## VII. Selection of the Weight Factors

The weighting factor $A$ of the quality function (6) decides if the priority will be given to the control of the power factor or to the control of the output currents. This factor was adjusted empirically based on [24]. First, $A$ is set to a high value in order to prioritize the control of the output current. As such, the inverter can control the output currents while the input currents will be highly distorted. After that, factor $A$ is slowly decreased, thereby lending more importance to the control of the reactive power. The output currents will not follow the reference if the factor $A$ is too low, as shown in Fig. 5(a). Therefore, $A$ is selected as the minimum value such that the output current has no noticeable deviations with respect to the reference, as it is shown in Fig. 5(b). There is no specific design criterion

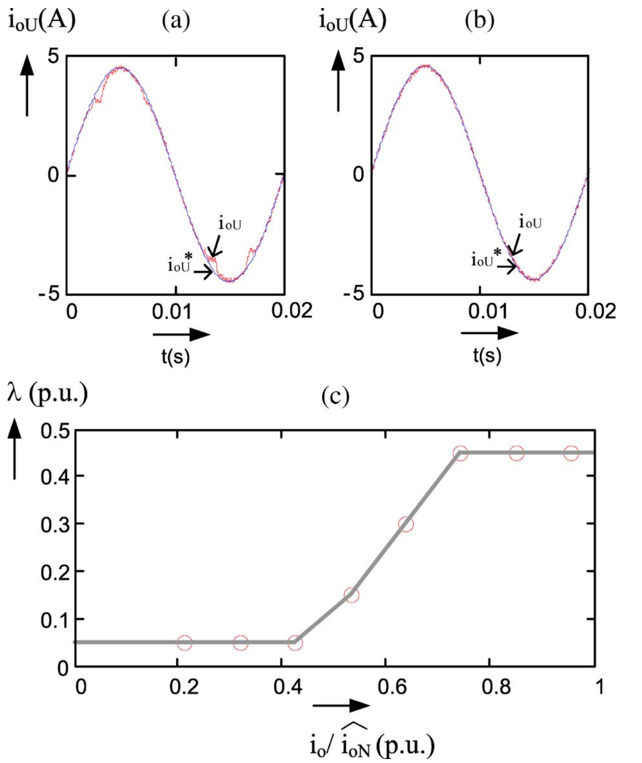


Fig. 5. Selection of the factor $A$. (a) Output current $A = 0.15$, 95% of the maximum output current. (b) Output current $A = 0.45$, 95% of the maximum output current. (c) Factor $A$ versus the load current.

to define the maximum allowable error in the output current, except that the maximum discrepancy must be below 5% of the maximum value of the reference. Since the adjustment of the parameter $A$ depends on the operating point, different values are obtained using this empiric method, as shown in Fig. 5(c).

## VIII. Results

A conventional indirect matrix converter, built by the Power Electronics Systems Laboratory of the Eidgenössische Technische Hochschule (ETH) Zürich, was used for the experimental evaluation. The converter features reverse blocking IGBTs of type IXRH40N120 for the bidirectional switch, standard IGBTs with antiparallel diodes of type FII50-12E for the inverter stage. A picture of the converter is shown in Fig. 6. The control scheme was implemented in a 160-MIPS fixed-point ADSP21991 DSP board which is also shown in Fig. 6. The processor board is connected to additional stacked boards that include a field-programmable gate array for the commutation sequence generation and the signal conditioning for the measurement of voltages and currents. The input filter is integrated in the converter board, as well as the current and voltage transducers. The original setup includes filter parameters of $L_f = 130$ $\mu$H and $C_f = 10$ $\mu$F. In this case, however, the capacity of the filter was increased to $C_f = 40$ $\mu$F to improve the stability of the input current control.

The sampling period of the control algorithm was set at $T_s = 20$ $\mu$s. It should be noted that the dead time necessary for a safe switching transition of the rectifier is approximately 10% of the aforementioned sampling period. Despite the fact that this dead time could be considered high for a low-power converter, better results can be expected from future generations of IGBTs.

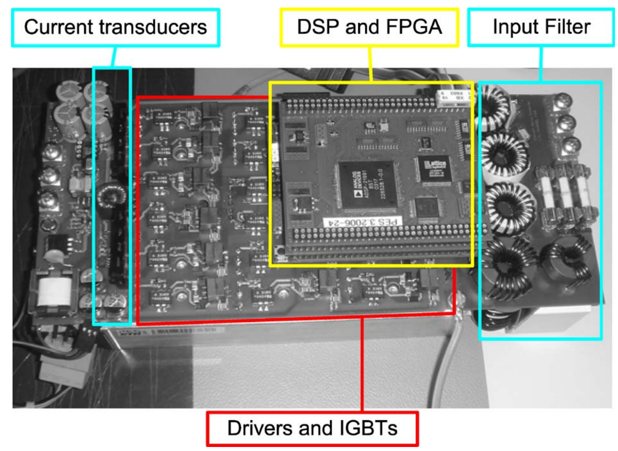


Fig. 6. Photograph of the indirect matrix converter.

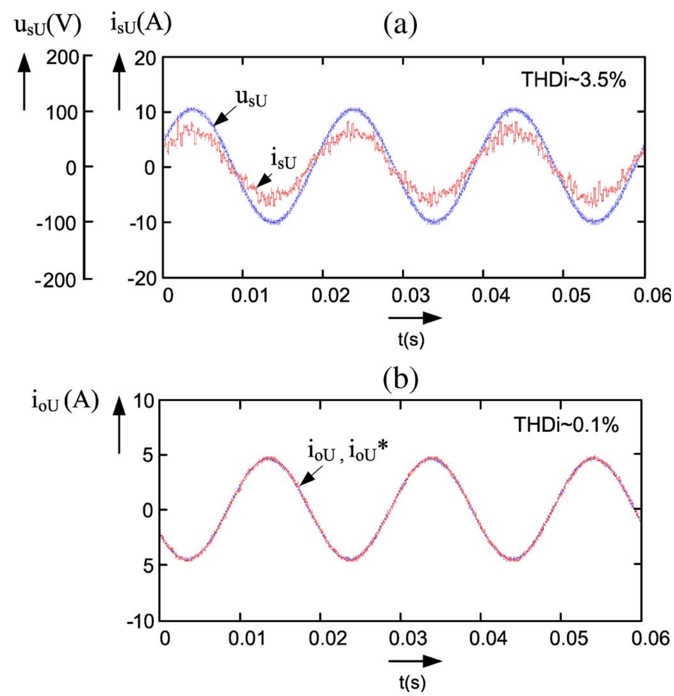


Fig. 7. Voltages and currents of the converter. (a) Input voltage $u_{sU}$ and input current $i_{sU}$. (b) Output current with its reference.

Fig. 7 shows the measured input current and voltage and output currents of phase U according to the parameters of the Appendix. As expected, the input current fulfils the condition of unitary power factor and presents an approximated total harmonic distortion (THD) of 3.5%. As it is shown in Fig. 7, the input current shows a ripple corresponding to the resonance frequency of the input filter. On the other hand, the output currents follow the reference accurately, with a THD lower than 1%. For the THD calculations, harmonics up to the sampling frequency were considered.

It is not possible to directly compare the predictive control scheme's performance with other methods that use a constant switching frequency. Results in [9] for a PWM-based control scheme operating at 25 kHz indicate that the proposed method performs similarly in terms of harmonic distortion. Better results can be expected by optimizing the input filter and adding active damping.

It must be acknowledged that the main advantage of the proposed control method is the simplicity in the implementation, since the controller does not need a complex modulation unit. This can reduce the overall cost of the complete system. The method also presents drawbacks as the quality function (6) is explicitly solved for each switching state. This can be a problem if a slow controller is used, as a higher sampling time could increase the harmonic distortion in the currents.

## IX. Conclusion

As a result of the advances in power semiconductors and processors, indirect matrix converter and predictive control schemes have recently emerged as feasible approaches. This paper takes advantage of these advances and proposes a predictive control scheme for an indirect matrix converter. The control scheme uses the predicted values of the input and output currents to evaluate the best suited converter state considering the output current error, the input power factor, and the generation of a positive dc-link voltage. Experimental results at a sampling time of $T_{\mathrm{s}} = 20\ \mu$s show that the unity power factor is obtained with input currents with THD of 3.5%. The output current follows the reference very accurately. The main advantage of the method is that, with a very simple control approach, results comparable with other PWM-based schemes can be obtained.

## Appendix

The parameters of the input filter and the load are as follows.

| Mains Filter: | Load: |
|---|---|
| $L_{\mathrm{f}} = 130\ \mu$H<br>$C_{\mathrm{f}} = 40\ \mu$F<br>$R_{\mathrm{f}} = 0.2\ \Omega$ | $R_1 = 20\ \Omega$<br>$L_1 = 10$ mH |

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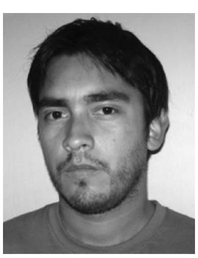

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