Model Predictive Control of an Inverter With Output LC Filter for UPS Applications

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***Abstract*—The use of an inverter with an output LC filter allows for generation of output sinusoidal voltages with low harmonic distortion, suitable for uninterruptible power supply systems. However, the controller design becomes more difficult. This paper presents a new and simple control scheme using predictive control for a two-level converter. The controller uses the model of the system to predict, on each sampling interval, the behavior of the output voltage for each possible switching state. Then, a cost function is used as a criterion for selecting the switching state that will be applied during the next sampling interval. In addition, an observer is used for load-current estimation, enhancing the behavior of the proposed controller without increasing the number of current sensors. Experimental results under linear and nonlinear load conditions, with a 5.5-kW prototype, are presented, verifying the feasibility and good performance of the proposed control scheme.**

***Index Terms*—Power conversion, predictive control, uninterruptible power systems.**

I. Introduction

THE CONTROL of inverters with an output LC filter has a special importance in applications where a high-quality voltage is needed. Such applications include distributed generation, energy-storage systems, stand-alone applications based on renewable energy, and uninterruptible power supplies (UPSs) [1], [2]. In these systems, it is desired, particularly for stand-alone applications and UPS systems [3], to achieve a good output-voltage regulation with any kind of load, being very important that the functionality of the system does not deteriorate under nonlinear loads, such as diode rectifiers.

The inclusion of an LC filter at the output of the inverter makes more difficult the controller design and controller parameters' adjustment. Several control schemes have been proposed for this converter, including deadbeat control [4]–[6], multiloop feedback control [7]–[9], adaptive control based on bank resonant filters [10], [11], and repetitive-based controllers [12], [13]. Some tuning strategies have been presented, including H-infinity control design [14]. In most of these schemes, the output voltage and one of two currents are used by a cascaded control considering outer and inner control loops, with linear or nonlinear controllers, to generate the reference voltages and by an external modulator to generate the firing pulses for the power semiconductors.

Predictive control appears as an attractive alternative for the control of power converters due to its fast dynamic response [15]. Several control algorithms have been presented under the name of predictive control, as presented in [16]. The most well-known scheme is deadbeat control, and it has been applied to current control in inverters [17], [18], rectifiers [19], active filters [20], [21], power-factor preregulators [22], and UPSs [6], [23]. When implemented in a digital system, deadbeat control needs to be modified in order to improve robustness. Several modifications have been proposed in the last years [24]–[27] considering adaptive schemes, neural networks, and other changes that make the controller robust but more complex. Another approach is the model predictive control (MPC), also known as receding-horizon control; it uses a model of the system to predict the behavior of the variables until a certain horizon of time, and a cost function is used as criterion to select the optimal future actions [28]–[31]. Several works have introduced the models for power converters [32], [33], which can be used to develop new model predictive controllers, improving the performance of the system. MPC is a very flexible control scheme that allows the easy inclusion of system constraints and nonlinearities in the design stage of the controller. In MPC, different formulations of the cost function are possible, considering different norms and including several variables and weighting factors [15]. It is also possible to consider different prediction horizons, as shown in [34]. The inputs of the system can be considered continuous, by using a modulator to apply the optimal voltages, as presented in [28] and [35]. In order to simplify the implementation of MPC, the converter can be modeled as a system with a finite number of switching states, and only one time step horizon can be considered for the optimization, as presented for the current control in a matrix converter in [36] and [37], a three-phase inverter in [38] and [39], an active front-end rectifier in [40], a multilevel inverter in [41], and a flying capacitor converter in [42]. This way, all possible switching states can be evaluated online; then, the one that minimizes the cost function is selected.

This paper proposes a new and simple MPC scheme for a three-phase inverter with output LC filter. The controller uses

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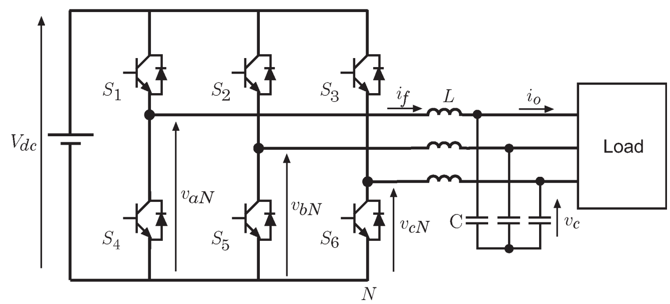

Fig. 1. Three-phase inverter with output LC filter.

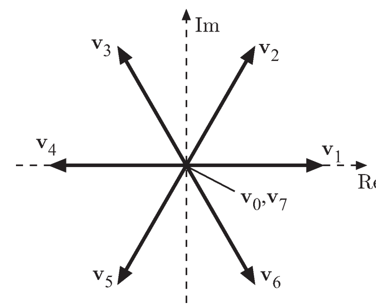

Fig. 2. Possible voltage vectors generated by the inverter.

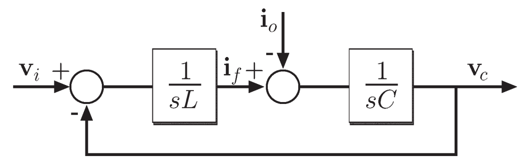

Fig. 3. LC filter model.

a model of the system to predict, on each sampling interval, the behavior of the output voltage for each possible switching state, and then, a cost function is used as a criterion for selecting the switching state that will be applied during the next sampling interval. There is no need of internal current-control loops and no modulators; the gate-drive signals are generated directly by the control. In addition, an observer is used for load-current estimation, enhancing the behavior of the proposed controller without increasing the number of current sensors.

II. Converter Model

The three-phase inverter with output LC filter considered in this paper is shown in Fig. 1. The converter and filter models are presented here, and the load is assumed unknown.

The switching states of the converter are determined by the gating signals S_a, S_b, and S_c as follows:

$$S_a = \begin{cases} 1, & \text{if } S_1 \text{ on and } S_4 \text{ off} \\ 0, & \text{if } S_1 \text{ off and } S_4 \text{ on} \end{cases} \tag{1}$$

$$S_b = \begin{cases} 1, & \text{if } S_2 \text{ on and } S_5 \text{ off} \\ 0, & \text{if } S_2 \text{ off and } S_5 \text{ on} \end{cases} \tag{2}$$

$$S_c = \begin{cases} 1, & \text{if } S_3 \text{ on and } S_6 \text{ off} \\ 0, & \text{if } S_3 \text{ off and } S_6 \text{ on} \end{cases} \tag{3}$$

and can be expressed in vectorial form by

$$\mathbf{S} = \frac{2}{3}\left(S_a + \mathbf{a}S_b + \mathbf{a}^2 S_c\right) \tag{4}$$

where $\mathbf{a} = e^{j(2\pi/3)}$.

The output-voltage space vectors generated by the inverter are defined by

$$\mathbf{v}_i = \frac{2}{3}\left(v_{aN} + \mathbf{a}v_{bN} + \mathbf{a}^2 v_{cN}\right) \tag{5}$$

where v_{aN}, v_{bN}, and v_{cN} are the phase voltages of the inverter, with respect to the negative terminal of the dc-link N (see Fig. 1). Then, the load voltage vector $\mathbf{v}_i$ can be related to the switching state vector $\mathbf{S}$ by

$$\mathbf{v}_i = V_{\mathrm{dc}}\mathbf{S} \tag{6}$$

where V_{dc} is the dc-link voltage.

Considering all the possible combinations of the gating signals S_a, S_b, and S_c, eight switching states and, consequently, eight voltage vectors are obtained. Note that $\mathbf{v}_0 = \mathbf{v}_7$, resulting in only seven different voltage vectors, as shown in Fig. 2.

Using modulation techniques like pulsewidth modulation, the inverter can be modeled as a continuous system. Nevertheless, in this paper, the inverter is considered as a nonlinear discrete system with only seven different voltage vectors as possible outputs.

Using vectorial notation, the filter current $\mathbf{i}_f$, the output voltage $\mathbf{v}_c$, and the output current $\mathbf{i}_o$ can be expressed as space vectors and are defined as

$$\mathbf{i}_f = \frac{2}{3}(i_{fa} + \mathbf{a}i_{fb} + \mathbf{a}^2 i_{fc}) \tag{7}$$

$$\mathbf{v}_c = \frac{2}{3}(v_{ca} + \mathbf{a}v_{cb} + \mathbf{a}^2 v_{cc}) \tag{8}$$

$$\mathbf{i}_o = \frac{2}{3}(i_{oa} + \mathbf{a}i_{ob} + \mathbf{a}^2 i_{oc}). \tag{9}$$

The LC filter is modeled as shown in the block diagram in Fig. 3. This model can be described by two equations, one that describes the inductance dynamics and the other describing the capacitor dynamics.

The equation of the filter inductance expressed in vectorial form is

$$L\frac{d\mathbf{i}_f}{dt} = \mathbf{v}_i - \mathbf{v}_c \tag{10}$$

where L is the filter inductance.

The dynamic behavior of the output voltage can be expressed by the following:

$$C\frac{d\mathbf{v}_c}{dt} = \mathbf{i}_f - \mathbf{i}_o \tag{11}$$

where C is the filter capacitance.

These equations can be rewritten as a state-space system as

$$\frac{d\mathbf{x}}{dt} = A\mathbf{x} + B\mathbf{v}_i + B_d\mathbf{i}_o \tag{12}$$

where

$$\mathbf{x} = \begin{bmatrix} \mathbf{i}_f \\ \mathbf{v}_c \end{bmatrix} \tag{13}$$

$$\mathbf{A} = \begin{bmatrix} 0 & -1/L \\ 1/C & 0 \end{bmatrix} \tag{14}$$

$$\mathbf{B} = \begin{bmatrix} 1/L \\ 0 \end{bmatrix}$$

$$\mathbf{B}_d = \begin{bmatrix} 0 \\ -1/C \end{bmatrix}. \tag{15}$$

Variables $\mathbf{i}_f$ and $\mathbf{v}_c$ are measured, while $\mathbf{v}_i$ can be calculated using (6), and $\mathbf{i}_o$ is considered as an unknown disturbance. In this paper, the value of V_{dc} is assumed fixed and known.

The output of the system is the output voltage $\mathbf{v}_c$ and written as a state equation

$$\mathbf{v}_c = [0 \quad 1]\mathbf{x}. \tag{16}$$

A. Discrete-Time Model of the Filter

A discrete-time model of the filter is obtained from (12) for a sampling time T_s and is expressed as

$$\mathbf{x}(k+1) = \mathbf{A}_q\mathbf{x}(k) + \mathbf{B}_q\mathbf{v}_i(k) + \mathbf{B}_{dq}\mathbf{i}_o(k) \tag{17}$$

where

$$\mathbf{A}_q = e^{\mathbf{A}T_s} \tag{18}$$

$$\mathbf{B}_q = \int_0^{T_s} e^{\mathbf{A}\tau}\mathbf{B}d\tau \tag{19}$$

$$\mathbf{B}_{dq} = \int_0^{T_s} e^{\mathbf{A}\tau}\mathbf{B}_{\mathbf{d}}d\tau. \tag{20}$$

These equations are used as the predictive model in the proposed predictive controller.

In order to predict the output voltage using (17), the output current $\mathbf{i}_o$ is needed, but usually, this current is not measured, and the load is unknown. A simple estimation of the load current can be calculated from filter-current and output-voltage measurements using the following equation obtained from (11):

$$\mathbf{i}_o(k-1) = \mathbf{i}_f(k-1) - \frac{C}{T_s}\left(\mathbf{v}_c(k) - \mathbf{v}_c(k-1)\right). \tag{21}$$

However, this estimation is very sensitive to noise in the measurements, because it is based on the derivative of the output voltage, so it will be preferred to use an observer such as the one presented in the next section.

III. Load-Current Observer

The load current depends on the load connected at the output of the filter which is unknown. However, some considerations can be taken in order to build an appropriate observer. These considerations consist of assuming a certain dynamic behavior of the load current.

A simple consideration is to assume that the load current is changing very slowly, compared to the sampling frequency. Lets assume that the load current can be approximated as a constant, so its behavior is described by the following differential equation:

$$\frac{d\mathbf{i}_o}{dt} = 0. \tag{22}$$

Then, including this load-current model in the filter model, the system is described by the following state-space equations:

$$\frac{d}{dt}\underbrace{\begin{bmatrix} \mathbf{i}_f \\ \mathbf{v}_c \\ \mathbf{i}_o \end{bmatrix}}_{\mathbf{x}} = \underbrace{\begin{bmatrix} 0 & -\frac{1}{L} & 0 \\ \frac{1}{C} & 0 & -\frac{1}{C} \\ 0 & 0 & 0 \end{bmatrix}}_{\mathbf{A}} \begin{bmatrix} \mathbf{i}_f \\ \mathbf{v}_c \\ \mathbf{i}_o \end{bmatrix} + \underbrace{\begin{bmatrix} \frac{1}{L} \\ 0 \\ 0 \end{bmatrix}}_{\mathbf{B}} \mathbf{v}_i. \tag{23}$$

The output of this system are the two measured variables, the filter current and the output voltage, and is defined by the following equation:

$$\mathbf{y} = \underbrace{\begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix}}_{\mathbf{C}} \begin{bmatrix} \mathbf{i}_f \\ \mathbf{v}_c \\ \mathbf{i}_o \end{bmatrix}. \tag{24}$$

A full-order observer for the system can be used to estimate the state vector $\mathbf{x}$. An observer is an open-loop model of the system which includes a correcting term based on the measured output. This is

$$\frac{d\hat{\mathbf{x}}}{dt} = \mathbf{A}\hat{\mathbf{x}} + \mathbf{B}\mathbf{v}_i + \mathbf{J}(\mathbf{y} - \hat{\mathbf{y}}) \tag{25}$$

where $\hat{\mathbf{y}} = \mathbf{C}\hat{\mathbf{x}}$ and $\mathbf{J}$ is the so-called observer gain [43].

This equation can be rewritten as

$$\frac{d\hat{\mathbf{x}}}{dt} = \mathbf{A}_{\mathrm{obs}}\hat{\mathbf{x}} + [\mathbf{B} \quad \mathbf{J}] \begin{bmatrix} \mathbf{v}_i \\ \mathbf{i}_f \\ \mathbf{v}_c \end{bmatrix} \tag{26}$$

where $\mathbf{A}_{\mathrm{obs}} = \mathbf{A} - \mathbf{JC}$. The output of the observer is the estimated load current

$$\hat{\mathbf{i}}_o = [0 \quad 0 \quad 1]\hat{\mathbf{x}}. \tag{27}$$

Note that the observer can be understood just as a filter which gives an estimate of the (unknown) load current $\hat{\mathbf{i}}_o$, based on measurements of the filter current $\mathbf{i}_f$, the output voltage $\mathbf{v}_c$, and the inverter voltage $\mathbf{v}_i$.

Matrix gain $\mathbf{J}$ will define the observer dynamics. As a design parameter, it must take into account the tradeoff between bandwidth and noise rejection. In fact, such gain can be chosen to be *optimal* if we characterize the statistical properties of the noise, and then, the observer is designed as a steady-state Kalman filter [44]. A simpler alternative is to choose the observer gain such that the poles of the observer give dynamics several times faster than the open-loop system dynamics.

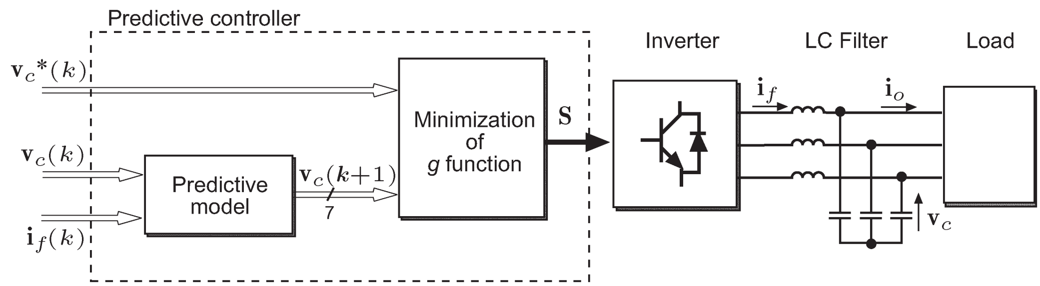

Fig. 4. Block diagram of the predictive controller.

IV. Proposed Predictive Controller

The use of MPC to control the power converter is proposed in this paper. MPC presents several advantages that make it suitable to control this kind of system: concepts are easy to understand, constraints and nonlinearities of the system can be easily included, and the desired behavior of the system is formulated as a cost function to be minimized. This paper takes into account an important restriction of the inverter, it can generate only seven different output-voltage vectors, and takes advantage of this restriction, making it possible to solve online the optimization problem of MPC. In these control schemes, an open-loop model is used for prediction and selection of the optimal actuations, but the use of a receding horizon provides the feedback to the control. This means that only the first element of the optimal actuation sequence is applied, and all the optimization is calculated again each sampling time.

The block diagram of the proposed predictive control for a three-phase inverter with output LC filter is shown in Fig. 4. Here, measurements of the output voltage $\mathbf{v}_c(k)$ and the filter current $\mathbf{i}_f(k)$ are used to predict, using (17), the value of the output voltage at the next sampling instant $\mathbf{v}_c(k+1)$ for all the possible voltage vectors that the inverter generates.

Note that in order to obtain the prediction $\mathbf{v}_c(k+1)$ from (17), an estimate of the (unmeasured) output current $\mathbf{i}_o(k)$ is required. In fact, a comparison is presented in Section V with results obtained using the simple derivative approximation in (21) and the load-current observer presented in the previous section.

To choose the *optimal* voltage vector $\mathbf{v}_i$ to be applied by the inverter, the seven predictions obtained for $\mathbf{v}_c(k+1)$ are compared using a cost function g. The voltage vector $\mathbf{v}_i$ that minimizes this function is then chosen and applied at the next sampling instant.

In this paper, we choose a cost function g expressed in orthogonal coordinates and defines the desired behavior of the system: To minimize the error in the output voltage

$$g = (v_{c\alpha}^* - v_{c\alpha})^2 + \left(v_{c\beta}^* - v_{c\beta}\right)^2 \tag{28}$$

where $v_{c\alpha}^*$ and $v_{c\beta}^*$ are the real and imaginary parts of the output-voltage reference vector $\mathbf{v}_c^*$, while $v_{c\alpha}$ and $v_{c\beta}$ are the real and imaginary parts of the predicted output-voltage vector $\mathbf{v}_c(k+1)$.

This cost function has been chosen in order to obtain the lowest voltage error. However, additional constraints can be considered in this function, such as current limitation, switching-frequency reduction, and spectrum shaping.

When implemented in a real system, the time needed for performing all the calculations of the control algorithm introduces a one sampling time delay that has to be compensated as explained in [45].

TABLE I
Parameters of the Experimental Setup

Parameter	Value
DC link voltage V_{dc}	520 [V]
Filter inductance L	2.4 [mH]
Filter capacitor C	40 [μF]
Sampling time T_s	33 [μs]

V. Experimental Results

The proposed predictive current-control strategy was tested experimentally using a Danfoss VLT5008 5.5-kW three-phase inverter with an output LC filter. The dc-link is fed by a three-phase diode-bridge rectifier. The inverter is controlled externally through an interface and protection card. A TMS320C6713 floating-point digital signal processor (DSP) was used for the control. A field-programmable-gate-array-based daughter card handles the analog-to-digital and digital-to-analog conversions and provides the digital outputs used as firing signals for power switches of the converter. The parameters of the system are shown in Table I.

In the implementation of the predictive-control algorithm, including the load-current observer, a minimum sampling time of $T_s = 33$ μs was achieved. However, further optimization in the programming is possible, but it is not the subject of this paper.

The behavior of the proposed predictive controller in steady-state operation for a resistive load of 20 Ω is shown in Fig. 5. The amplitude of the reference voltage is set to 200 V, and the frequency is 50 Hz. It is shown in the figure that the output voltages are sinusoidal with low distortion. The output voltages and currents with a different amplitude of the reference voltage of 150 V are shown in Fig. 6.

The output voltage, output current, and filter current in one phase are shown in Fig. 7. It is shown that, due to the resistive load, the load current is proportional to the output voltage while the filter current measured at the output of the converter presents high-frequency harmonics which are attenuated by the filter.

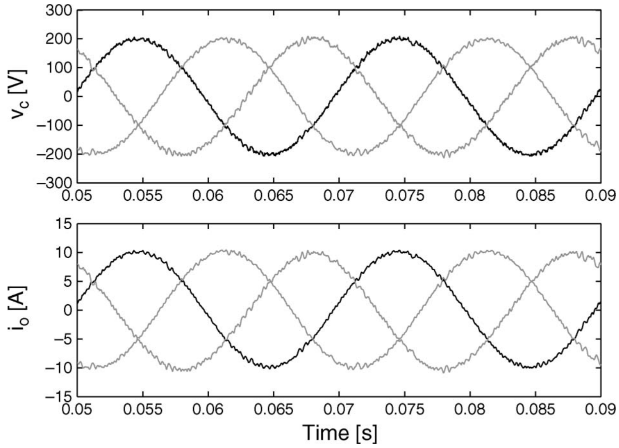

Fig. 5. Experimental results: output voltages and currents in steady state for a reference amplitude of 200 V with 20-Ω load. Voltage THD: 2.65%.

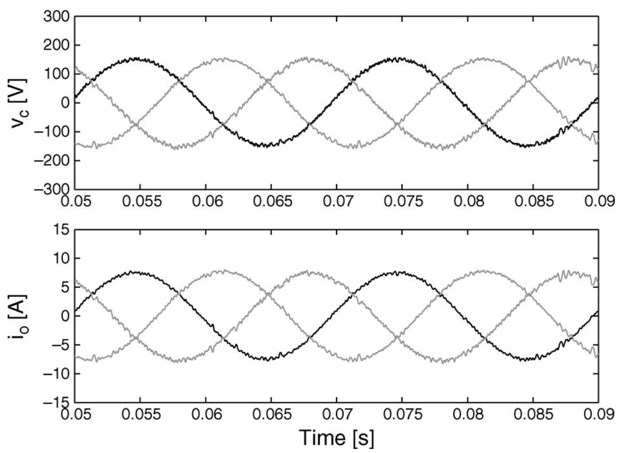

Fig. 6. Experimental results: output voltages and currents in steady state for a reference amplitude of 150 V with 20-Ω load. Voltage THD: 2.82%.

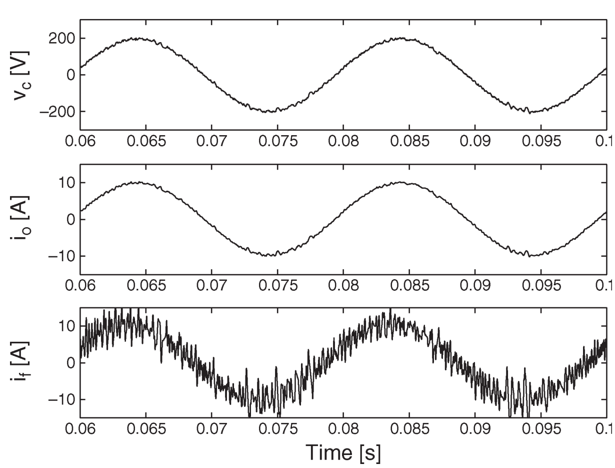

Fig. 7. Experimental results: output voltage, output current, and filter current in steady state.

The transient behavior of the system for a load step from no load to full load is shown in Fig. 8. Here, a 20-Ω load is connected at time of 0.05 s. It can be seen in this result that the output voltage is not affected by this change in the load. A similar test is shown in Fig. 9 for a resistive–inductive load.

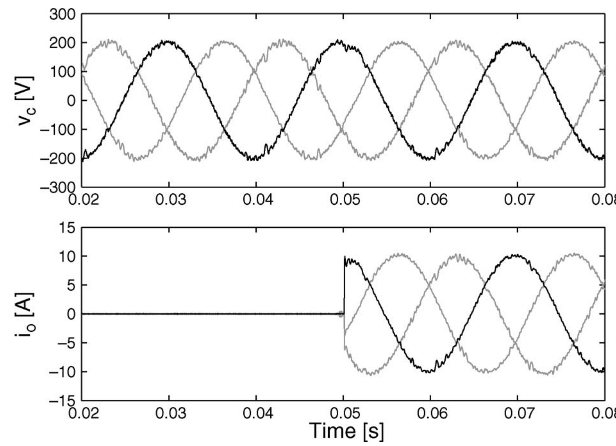

Fig. 8. Experimental results: output voltage and output current for a resistive load step from no load to full load.

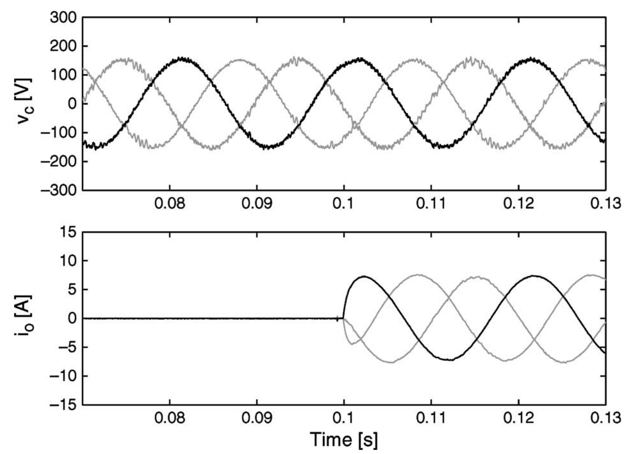

Fig. 9. Experimental results: output voltage and output current for a resistive–inductive load step.

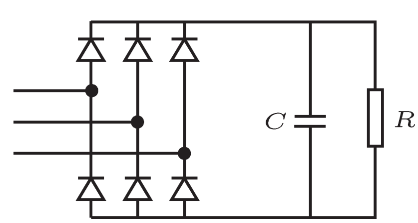

Fig. 10. Diode-bridge rectifier used as nonlinear load. $C = 3000\ \mu$F, $R = 60\ \Omega$.

The load, with values $R = 20\ \Omega$ and 10 mH, is connected at time 0.1 s. The amplitude of the reference voltage is 200 V for the resistive load and 150 V for the resistive–inductive load.

The diode-bridge rectifier shown in Fig. 10 was used as nonlinear load for the results shown in Fig. 11. Here, the output voltage presents a small distortion, but it is still sinusoidal despite the highly distorted load currents. A noticeable unbalance in the load currents is present in this result due to unbalanced voltages when a nonlinear load is connected. This result could be improved by using a higher sampling frequency, but this solution is difficult to implement due to hardware restrictions. Alternative solutions to improve the quality of the control for nonlinear loads are under consideration.

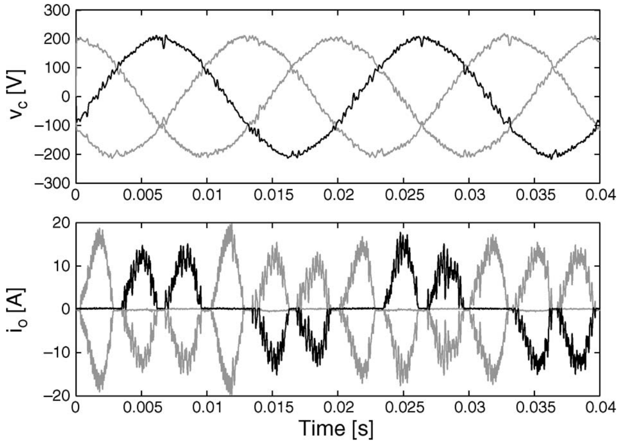

Fig. 11. Experimental results: output voltages and currents in steady state for a nonlinear load and a reference amplitude of 200 V. Voltage THD: 4.60%.

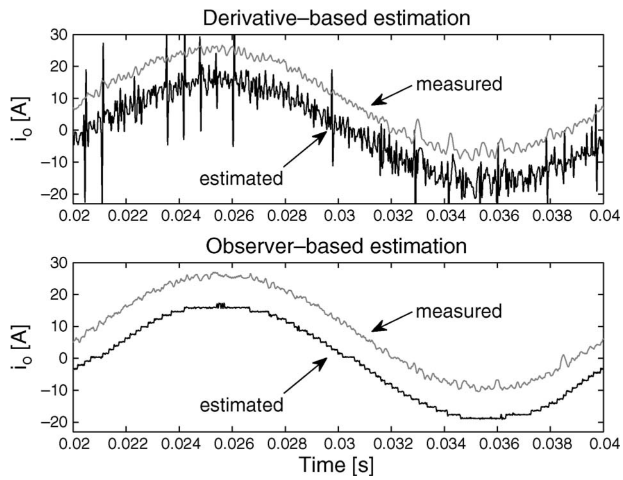

Fig. 12. Experimental results: derivative-based and observer-based load-current estimations.

Two estimation methods were implemented experimentally, the derivative-based estimation (21) and the observer-based estimation. All the results previously shown were obtained using the load-current observer. In Fig. 12, the performances of both estimators are shown. It can be noted that the derivative-based estimation is very noisy while the observer-based estimation is much cleaner due to the filtering action of the observer. The output voltages obtained using the proposed predictive controller with both estimation methods are shown in Fig. 13 for the same operating conditions. The use of a noisy estimation introduces errors in the calculation of the output-voltage prediction producing distorted voltages, as shown in Fig. 13(a). When comparing this result with the one shown in Fig. 13(b) using the observer, it is clear that the use of the observer enhances the behavior of the control, achieving a lower THD, without the use of additional current measurements. Results from Fig. 13 where obtained using a smaller filter, compared to the results shown in Fig. 5, and consequently present a higher THD value. When the nonlinear load is connected, the error introduced by the derivative-based estimator makes the controller fail to control the output voltage, and the UPS does not work. However, the load-current observer delivers a good estimation, as shown in Fig. 14, allowing for a good control of the output voltages.

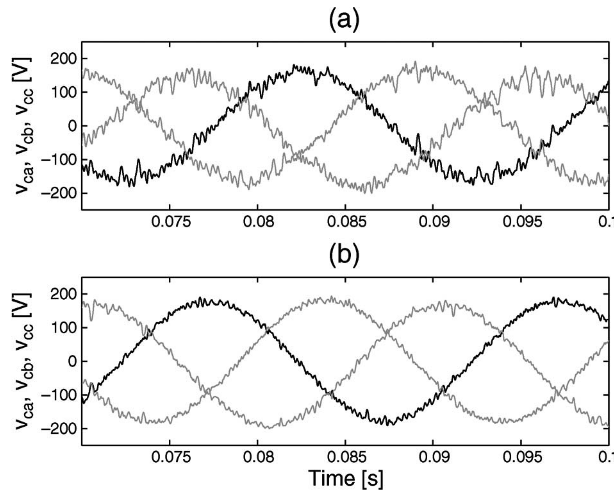

Fig. 13. Experimental results. (a) Output voltages using derivative-based load-current estimation. THD $= 7.8\%$. (b) Output voltages using observer-based load-current estimation. THD $= 3.8\%$.

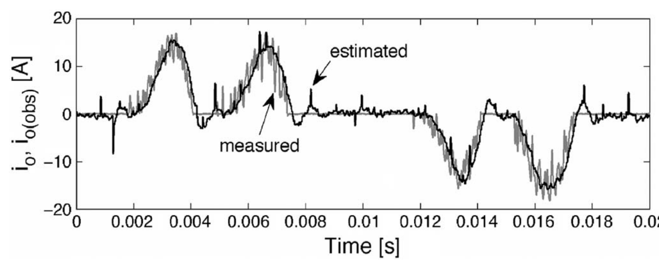

Fig. 14. Experimental results: observer-based load-current estimation for the nonlinear load. (An offset has been added to the measured signals.)

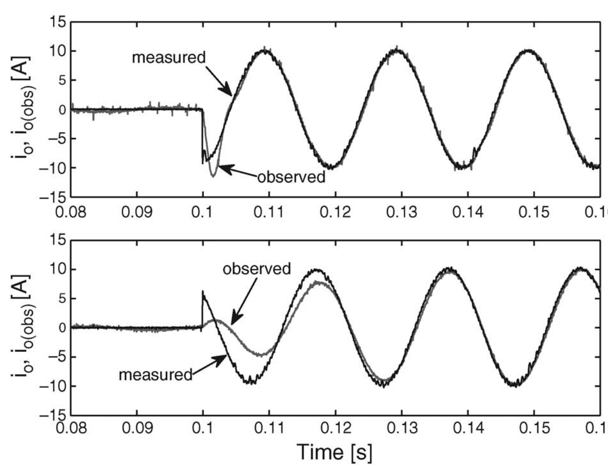

Fig. 15. Experimental results: load current and estimated load current using the observer for a load step. (Top) Using a fast observer. (Bottom) Using a slow observer.

The effect of two different pole placement for the load-current observer is shown in Figs. 15 and 16. A plot of the pole placement for the fast and slow observers is shown in Fig. 17. The load current in one phase of the load is shown in Fig. 15 when a resistive load step is applied at time 0.1 s. A fast observer is used in the upper figure where it is shown how the estimated value is equal to the measured value in a fraction of fundamental cycle. On the other hand, if a slow observer is used, the estimated value reaches the real value after a full

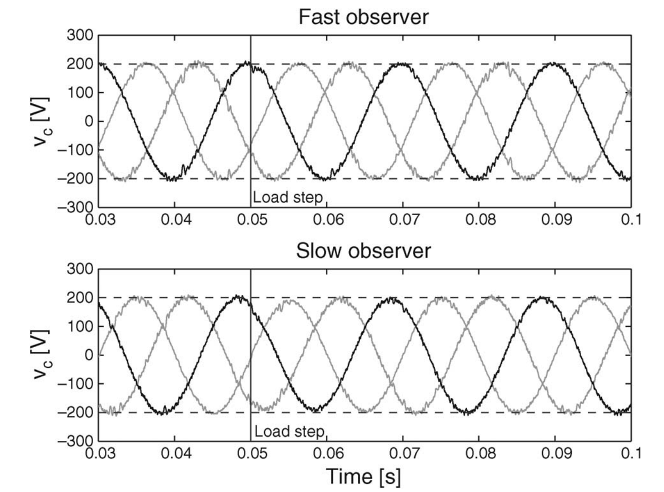

Fig. 16. Experimental results: output voltages for a load step. (Top) Using a fast observer. (Bottom) Using a slow observer.

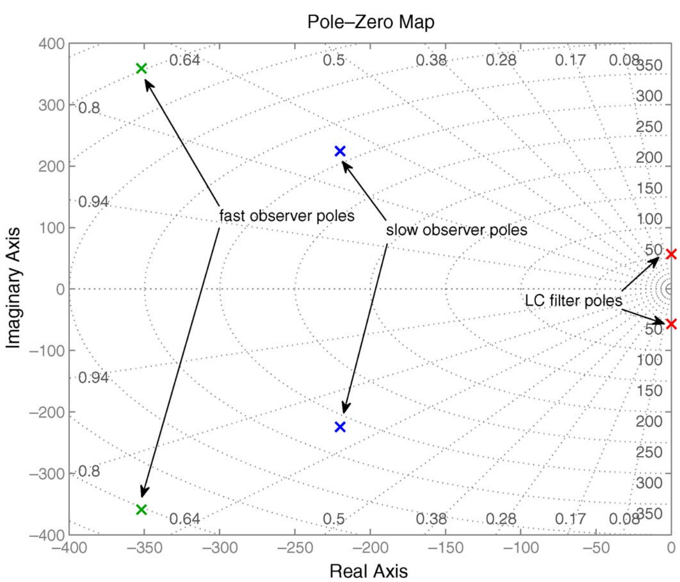

Fig. 17. Pole placement for fast and slow observers.

cycle. The effect of these differences in the estimation over the output voltage is shown in Fig. 16. Here, a resistive load step is applied at time 0.05 s. While with the fast observer the effect of the load step is a very short drop in the voltages, with the slow observer, there is a noticeable reduction of the voltage amplitude, with respect to the reference amplitude (dashed lines), between times 0.05 and 0.06 s.

VI. Conclusion

In this paper, a new and simple control scheme was presented for a three-phase inverter with output LC filter. The feasibility of the proposed predictive controller has been demonstrated by implementing it in a laboratory prototype. Results show that the proposed scheme achieves a good voltage regulation with linear loads as well as with nonlinear loads.

The proposed controller has no parameters to adjust; it needs a model of the system for calculating predictions of the controlled variables. The gate-drive signals are generated directly by the controller, so a modulator is not needed.

The output voltage is directly controlled, without using a cascaded control structure, with an inner current-control loop. This allows for a fast dynamic response of the voltage control.

It has been shown that the use of an observer allows for a good estimation of the unknown load current and improves the overall behavior of the system without the need of additional current measurements.

Predictive control presents a different approach for the control of power converters, taking into account the discrete nature of the converters and the microprocessors used for the control. In addition, the high calculation power of the current existing DSPs makes this method very attractive to control power converters.

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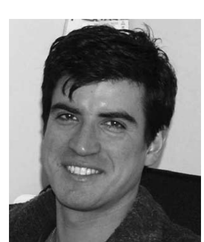

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