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Comparative Evaluation of Predictive Control Schemes for Three-Phase Buck-Type PFC Rectifiers

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***Abstract*—This paper presents a comparative evaluation of different control schemes based in Model Predictive Control (MPC). These control schemes are compared with a classical control scheme based on pulse width modulation and linear controllers. MPC-based schemes with and without the inclusion of a modulation stage are studied in this work. Simulation results show that although MPC-based schemes present superior dynamic performance, the ripple in the input currents is much higher than in the classical PWM-based schemes operating at the same switching frequency. This problem is solved by the inclusion of a modulation stage in the MPC control scheme.**

I. Introduction

Three-phase buck-type rectifiers, also known as current source rectifiers, present several advantages that make them suitable for a wide range of applications such as power supplies and front-end for AC drive applications. The systems operates with sinusoidal input currents, in phase with the grid voltages, and provide a wide output voltage range, which can be extended by the addition of a DC/DC boost converter at the output. The power circuit of the converter is shown in **Fig. 1**, including a two-stages input filter and a LC output filter. The specifications of the system are listed in **Table I**. A picture of a laboratory prototype of the three-phase buck-type rectifier is shown in **Fig. 2**.

Several modulation and control schemes have been proposed for this converter. The optimization of the modulation method has been studied in [1], [2]. In general terms, control schemes can be divided in two types:

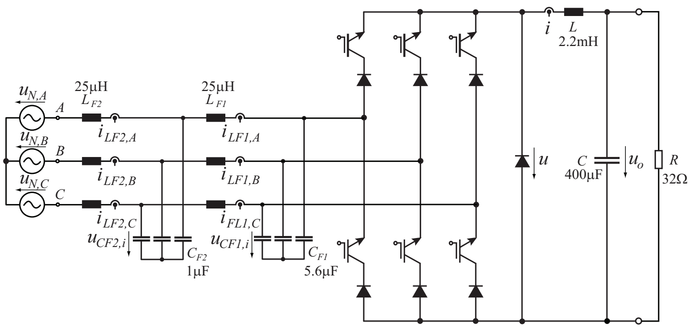

Fig. 1 Three-phase 6-switch buck rectifier with two-stage input filter.

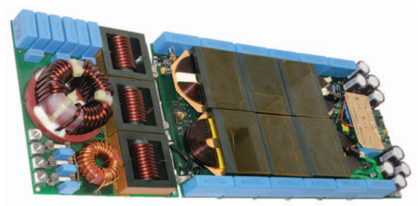

Fig. 2 Three-phase 6-switch buck rectifier prototype.

TABLE I Specifications for the Buck Rectifier.

Nominal output power	5 kW
DC-link voltage	400 V
Nominal input voltage (rms, line to neutral)	230 V
Input frequency	50 Hz
Switching frequency	36 kHz

- The ones based on an inner DC current control (without control of the input currents) [3], [4].
- The ones based on the control of the AC currents [5], [6].

A comparative evaluation of these two types of control schemes is discussed in [7].

The controller design must consider that disturbances from the load side as well as from the grid side can be present, and that the input and output variables are highly coupled. If the input and output are simultaneously controlled, the response to input and output disturbances can be improved. However, both controllers must act on the same converter and in some cases the duty cycles or switching states required by each controller can be conflicting.

In order to achieve a fast dynamic response on the DC side and provide a fast response for disturbances in the grid voltage, different control strategies for simultaneous control of the input and output variables are presented and compared in this paper. The proposed control schemes are based on the ideas of Model Predictive Control (MPC), which have already been applied in several power electronics and drives applications [8], [9], including current source rectifiers at low switching frequency [10]. The use of MPC allows to overcome the problem of having conflicting actuations when simultaneous control of the input and output is required. With the use of a cost function the actuation decision is based on

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the minimization of the input and output errors.

II. Three-Phase Buck-Type PFC Rectifiers

Current source rectifiers (CSR) present important differences to the voltage source rectifiers (VSR) that make the control more difficult. The most important characteristic is the low energy storage capability of the DC-side energy storage, i.e. DC-side inductor which introduces a highly dynamic interaction between the input and output quantities. In a VSR the output voltage presents a much lower time constant with respect to the input currents time constant. This is because a relative large capacitor is typically used in the DC-link. In order to illustrate the difference in the relation of the input and output dynamics, the converter topologies and values shown in **Fig. 3** are considered. A rated power of 5 kW will be assumed for the following estimations.

In a VSR, the effect of applying the maximum actuation voltage to the inductance of the filter can be estimated in a rough first step as:

$$\Delta i_{LF,A} = \frac{\Delta t \cdot u_o}{\frac{3}{2} L_F}, \tag{1}$$

and the variation of the output voltage can be estimated as

$$\Delta u_o = \frac{\Delta t \cdot i}{C}, \tag{2}$$

considering that the corresponding switching state is applied for a complete PWM period, for a switching frequency of 36 kHz and a DC-link voltage u_o of 700 V. (For estimation under the worst case conditions the input voltages and the load current are assumed to be zero. It is also assumed that one phase is connected to the positive bar and the other two are connected to the negative bar of the DC-link.) The variation in the input current is $\Delta i_{LF,A}$ =1.3 A, which corresponds to a 25% of the rated current. The maximum variation of the DC-link voltage is approximately Δu_o =1.6 V, which corresponds to a 0.2% of the rated voltage. This example shows that a switching action in a VSR has a pronounced effect on the input currents, in contrast to the effect on the output voltage. In this way, large difference in time constants allows a decoupled control of the input and output of the VSR.

Applying the same analysis, the changes in the input capacitor voltage and output current of the CSR can be estimated as

$$\Delta u_{CF,A} = \frac{\Delta t \cdot i}{C_F} \tag{3}$$

and

$$\Delta i = \frac{\Delta t \cdot u_{CF,ij}}{L}, \tag{4}$$

considering an output current i =12.5 A, with the application of the corresponding switching state for a complete PWM period. For the worst case, the grid current and the load voltage are assumed to be zero. The capacitor voltage presents a variation of $\Delta u_{FC,A}$ =64 V which is 19% of its rated amplitude. On the other side, the output current can change by the application of the line-to-line voltage that has the maximum amplitude over the output filter inductor, resulting in a maximum variation of Δi =7.1 A, which represents a 56.8% of the rated output current. It can be observed that the changes in the input and output quantities due to a switching action is in the same order of magnitude, with respect to their respective rated values. This clearly shows the highly dynamic interaction between the input and output quantities present in a CSR like the three-phase buck rectifier presented in this paper.

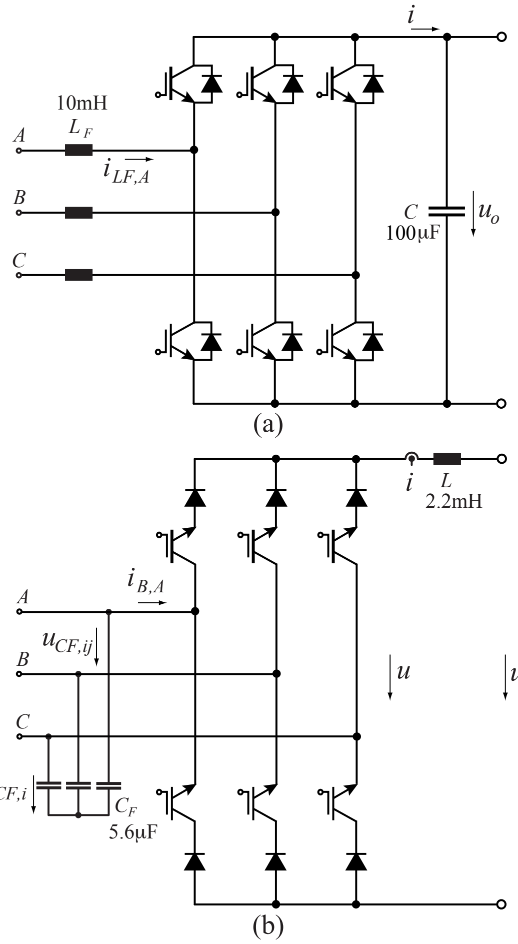


Fig. 3 Power circuit of voltage source and current source rectifiers. (a) Voltage source rectifier. (b) Current source rectifier.

For this reason, a simultaneous control of the input and output sides of the three-phase buck-type rectifier can result in a better performance of the system in a wide range of operating conditions such as reference changes, load changes and input voltage variations. In order to overcome the conflict of having two different control targets, the use of MPC is proposed. This paper presents and compares several control schemes exploring this idea.

III. Inner DC current control

As a reference point for comparisons, the control scheme shown in **Fig. 4.(a)** is considered. This control scheme is explained in detail in [7]. An outer control loop regulates the output voltage u_o and its output is the reference for the inner DC current control loop. The current controller

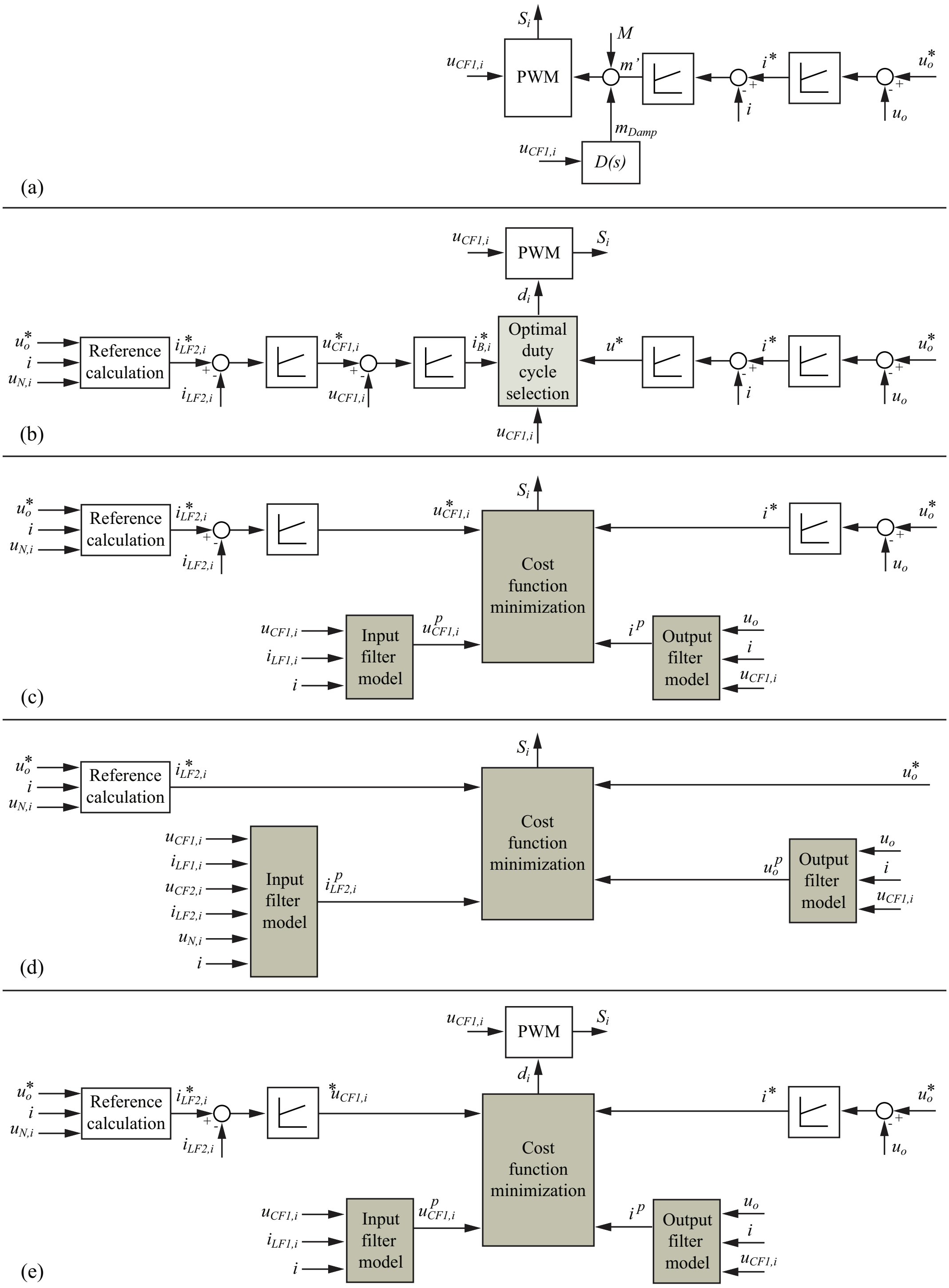


Fig. 4 Control schemes for the buck rectifier. (a) Inner DC current control scheme. (b) Inner DC current and AC capacitor voltage control scheme. (c) Inner DC current and AC capacitor voltages control using MPC. (d) MPC scheme for output DC voltage and input AC currents control. (e) Inner DC current and AC capacitor voltages control using MPC and PWM.

determines the required modulation index m and the pulse-width modulator (PWM) generates the corresponding gate signals for the IGBTs.

IV. Proposed Control Schemes

A. *Inner DC current and AC capacitor voltage control scheme*

In this control scheme, as shown in **Fig. 4.(b)**, an outer DC voltage control loop generates the reference value for the inner DC current. For the input side of the converter, an outer current control loop generates the reference values for the inner capacitor voltage control loop. The input current reference is generated by consideration of the input-output power balance. An inner DC current control loop and AC capacitor voltage control loop generate the reference values for the output voltage u_o^* and converter input currents $i_{B,i}$, respectively. Subindex B denotes the currents at the input of the buck rectifier. The corresponding duty cycles d_A ,d_B and d_C are selected in order to minimize the output voltage and input current errors, defined by the following function:

$$g = (u^* - u(d_A, d_B, d_C))^2 + K_{AC} \sum_{i=A,B,C} (i_{B,i}^* - i_{B,i}(d_A, d_B, d_C))^2, \quad (5)$$

where the weighting factor K_{AC} allows to adjust the importance of each term of this function. Then, the PWM algorithm generates the gate drive signals according to the required duty cycles.

B. Inner DC current and AC capacitor voltages control using MPC

This control scheme presents a similar structure than the previous one, shown in **Fig. 4.(b)**, but the inner DC current and AC capacitor voltages are controlled using MPC, as shown in **Fig. 4.(c)**. The PWM stage is also replaced by the MPC-based controller, which directly generates the gate signals for the IGBTs. A model of the input filter is used for calculation of predictions of the capacitor voltage values for all possible switching states of the buck rectifier. In the same way, the model of the output filter is used for prediction of the values of the DC current for all switching states. Then, the switching state that minimizes a given cost function is selected and applied for a complete sampling period. The cost function is defined as a function of the DC current error and the capacitor voltages error:

$$g = (i^* - i^p)^2 + K_{AC}[(u_{CF1,\alpha}^* - u_{CF1,\alpha}^p)^2 + (u_{CF1,\beta}^* - u_{CF1,\beta}^p)^2] \quad (6)$$

where K_{AC} is a weighting factor that allows to adjust the importance of each term in the cost function. The superindex p denotes the predicted values. In order to reduce the number of calculations, the measured capacitor voltages $u_{CF1,A}$, $u_{CF1,B}$ and $u_{CF1,C}$ are transformed to the orthogonal coordinates system (α, β). The MPC algorithm is then implemented using the transformed capacitor voltages $u_{CF1,\alpha}$ and $u_{CF1,\beta}$.

C. MPC scheme for output DC voltage and input AC currents control

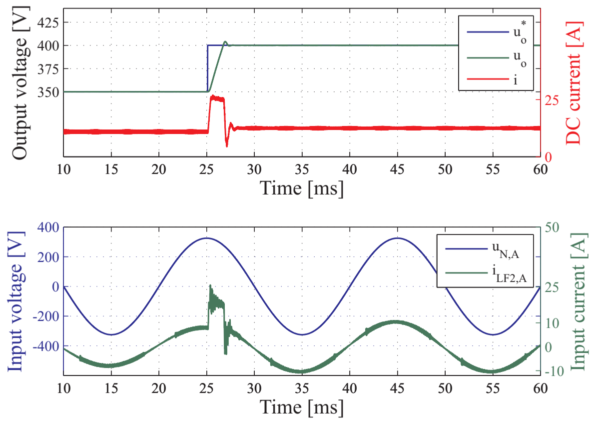
A control scheme using MPC for the complete system is shown in **Fig. 4.(d)**. Here, the system model is used for calculation of the predicted values of the output voltage and input currents for all possible switching states of the buck rectifier. The optimal switching state is selected by minimization of the following cost function:

$$g = \sum_{i=1}^{N} (u_o^* - u_o^p)^2 + K_i \left(\int_0^{k+N} (u_o^*(t) - u_o^p(t))dt \right)^2 + K_{AC} \sum_{i=1}^{N} [(i_{LF2,\alpha}^* - i_{LF2,\alpha}^p)^2 + (i_{LF2,\beta}^* - i_{LF2,\beta}^p)^2] \quad (7)$$

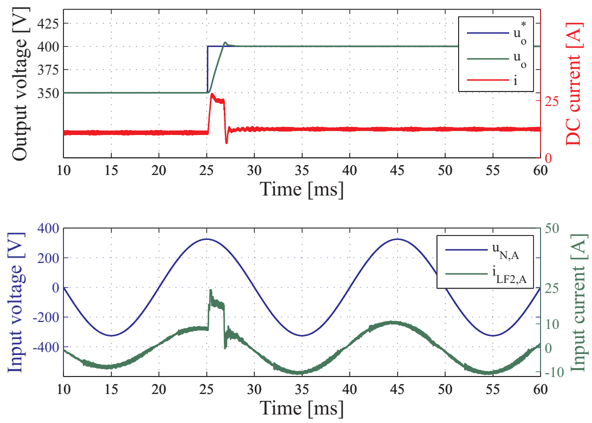
The input current references are calculated considering the input-output power balance, as in the previous case.

D. Model Predictive Control with PWM (MPC-PWM)

As it can be observed in the results presented later in this paper, the previous two control schemes present a high ripple in the controlled variables due to the absence of a modulation stage. In order to reduce the ripple a modulation stage is considered in the MPC scheme, as shown in **Fig. 4.(e)**. The MPC part of the control scheme is responsible for the control of the DC current and the AC capacitor voltages. The cost

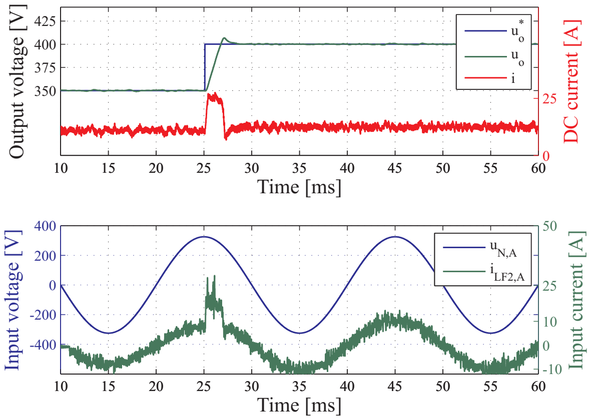


(a) Inner DC current control (cf. Fig. 4(a)).

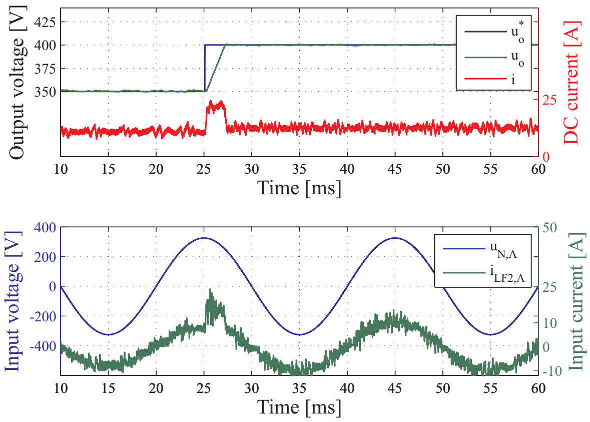


(b) Optimal duty cycles (cf. Fig. 4(b)).

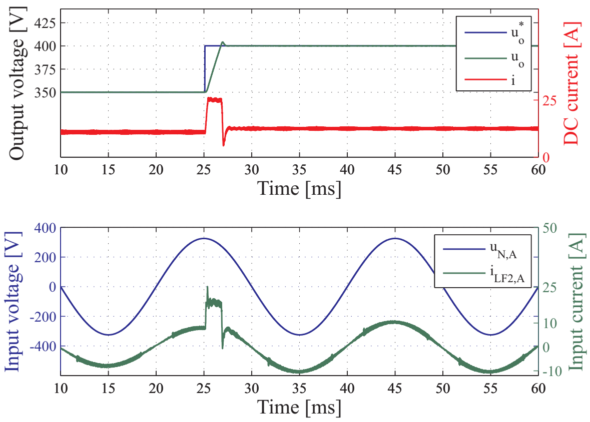
Fig. 5 Results for a step change in the reference voltage. PI-based schemes.



(a) Inner MPC (cf. Fig. 4(c)).



(b) Full MPC (cf. Fig. 4(d)).

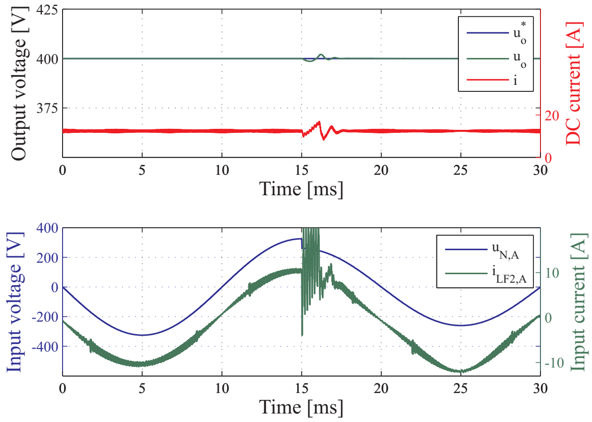


(c) Inner MPC+PWM (cf. Fig. 4(e)).

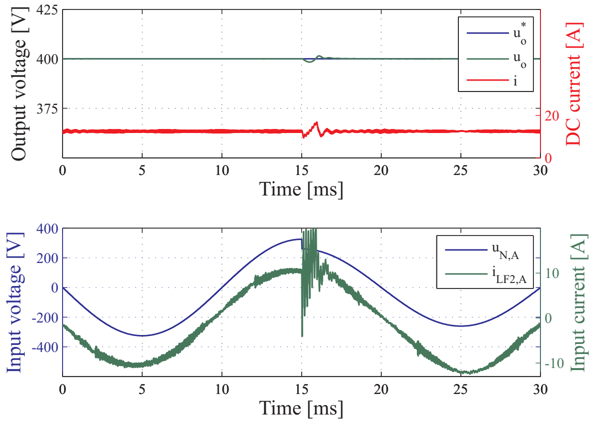
Fig. 6 Results for a step change in the reference voltage. MPC-based schemes.

function is the same as presented in (6).

In this case a continuous control set is considered for the duty cycles, with $d_i \in [0, 1]$. The values of the duty cycles



(a) Inner DC current control (cf. Fig. 4(a)).



(b) Optimal duty cycles (cf. Fig. 4(b)).

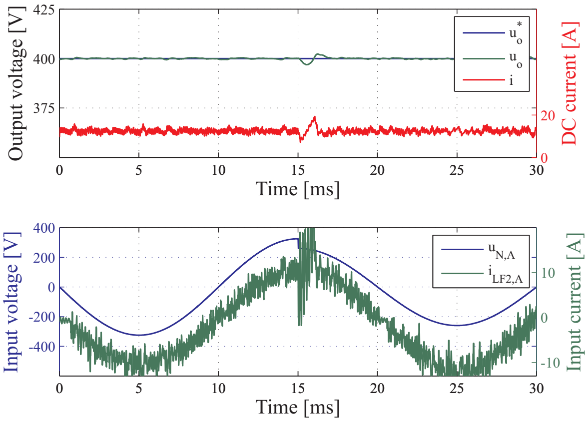
Fig. 7 Results for a step change in the amplitude of phase A grid voltage. PI-based schemes.

that minimize the cost function (6) are selected and then the PWM block generates the gate signals for the switches.

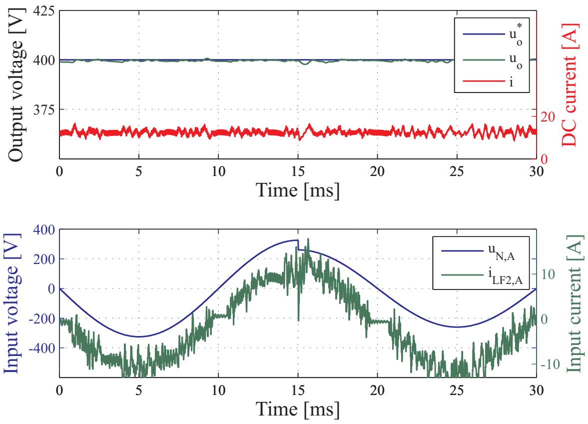
V. Comparative Results

In order to assess the behavior of the different control schemes, simulations results are presented. Results for a step change in the output voltage reference u_o^* from 350 to 400 V are presented in **Fig. 5** and **Fig. 6**. In all control schemes, the DC current is limited to 25 A (200% of the rated current). It can be observed that all control schemes present a similar dynamic behavior. However, the control schemes based on MPC without modulation show a very high ripple, compared to the PWM-based schemes.

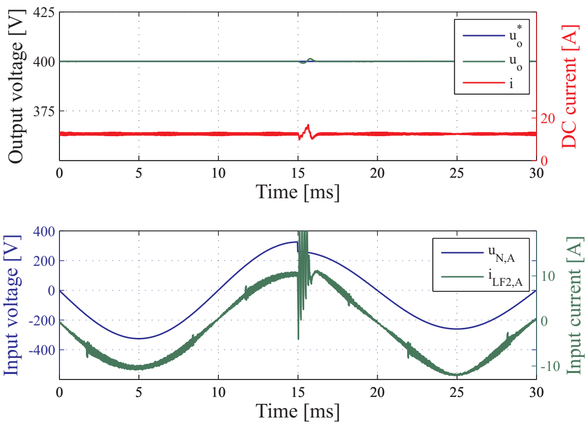
The behavior of the three-phase buck rectifier for a 20% step drop in the amplitude of one phase voltage of the grid is shown in **Fig. 7** and **Fig. 8**. The inner DC current control scheme presents a slower response, with a noticeable deviation of the output voltage u_o and the input current $i_{LF2,A}$, as it can be observed in **Fig. 7.(a)**. The behavior of the buck rectifier is improved by the control of the input side, with



(a) Inner MPC (cf. Fig. 4(c)).



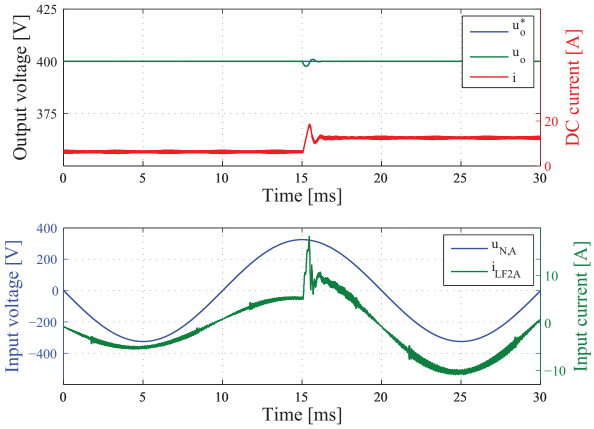
(b) Full MPC (cf. Fig. 4(d)).



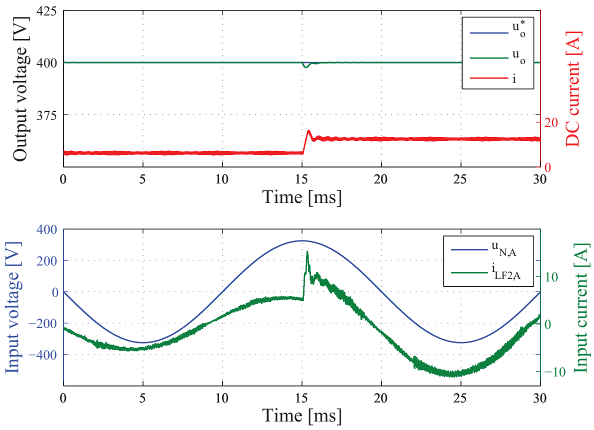
(c) Inner MPC+PWM (cf. Fig. 4(e)).

Fig. 8 Results for a step change in the amplitude of phase A grid voltage. MPC-based schemes.

best results for the control scheme with optimal duty cycles (**Fig. 7.(b)**), and the inner MPC with PWM (**Fig. 8.(c)**). From the other MPC-based schemes, the full MPC scheme presents better performance than the inner MPC scheme, as observed in **Fig. 8.(a)** and **Fig. 8.(b)**.



(a) Inner DC current control (cf. Fig. 4(a)).



(b) Optimal duty cycles (cf. Fig. 4(b)).

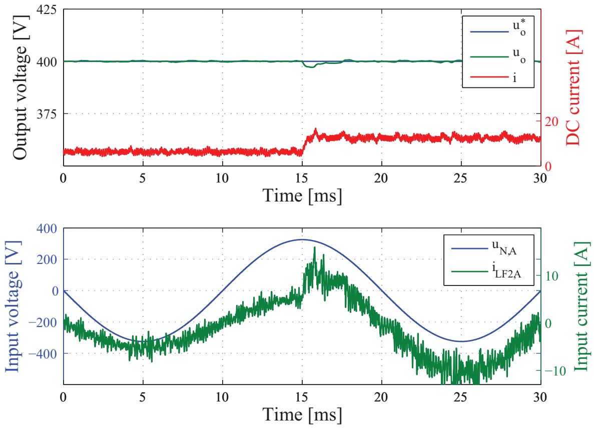
Fig. 9 Results for a step change in load resistance from 64 Ω to 32 Ω. PI-based schemes.

The response of the system for a step change in the load resistance is shown in **Fig. 9** and **Fig. 10**. The load resistance is chanced from 64 Ω to 32 Ω. It can be observed that the proposed control schemes present a lower overshoot in the input current, compared to the results obtained with the inner DC current control scheme, shown in **Fig. 9.(a)**.

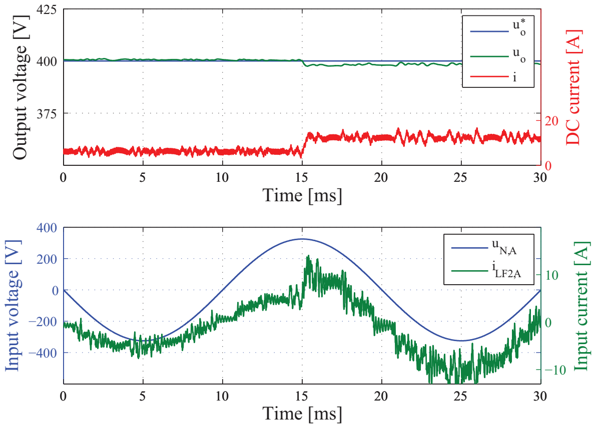
VI. Conclusions

Several control schemes for simultaneous control of the input and output sides of the three-phase buck rectifier are proposed in this paper. The proposed schemes are based on the ideas of MPC. The use of cost functions allow control both sides simultaneously and to adjust the importance of the control of the input and output variables by setting the appropriate value of the weighting factors. In this way, MPC solves the conflict of two different control targets.

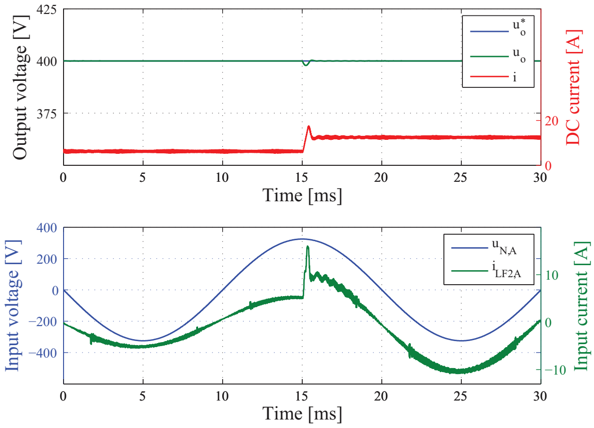
The use of MPC without a modulation stage allows to control the buck rectifier with fast dynamics. However, in the



(a) Inner MPC (cf. Fig. 4(c)).



(b) Full MPC (cf. Fig. 4(d)).



(c) Inner MPC+PWM (cf. Fig. 4(e)).

Fig. 10 Results for a step change in load resistance from 64 Ω to 32 Ω. MPC-based schemes.

switching frequency range of this application, the ripple in the currents is too high. The inclusion of a PWM stage in the MPC scheme provides the solution for reducing the ripple while keeping the fast dynamics.

In the course of future research the control concepts will be verified for a demonstrator system of the buck-type PWM rectifier. There a passive damping of the input filter will be employed which also would be present for an industrial system but has been omitted here in order to clearly show the different dynamic properties of the analyzed control schemes. This passive damping allows to avoid a ringing of the input filter also for standby mode operation and would also attenuate the oscillations occurring for a step change of the input voltage (cf. **Fig. 7** and **Fig. 8**). In addition to the comparative results shown in this paper, a more complete comparison will consider the frequency responses for the different control schemes, obtaining the following transfer functions: from grid voltage to grid current, from grid voltage to output voltage, from load current to grid current, and from load current to output voltage.

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