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Research Article

A Single Nonlinear Current Control for PWM Rectifier Robust to Input Disturbances and Dynamic Loads

Nancy Visairo-Cruz,[1] Ciro Núñez-Gutiérrez,[1] Eliseo Alcázar,[2] and Elías Rodríguez[3]

[1] *Faculty of Engineering, Universidad Autónoma de San Luis Potosí, 78290 San Luis Potosí, SLP, Mexico*
[2] *Schweitzer Engineering Laboratories, S.A. de C.V., 78395 San Luis Potosí, SLP, Mexico*
[3] *Instituto Tecnológico de Celaya, Celaya, GTO, Mexico*

Correspondence should be addressed to Nancy Visairo-Cruz; nvisairoc@uaslp.mx

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The requirements of PWM rectifiers for delivering power to motor drives include power factor correction and output voltage regulation even when strong variations such as voltage sags and dynamic load transients occur simultaneously. To achieve these objectives, the classic approach is to use a two-loop controller with its d-q model. In this paper, the authors propose a simplified approach to address that problem by using a feedback linearization-based nonlinear controller using only a single-loop current control and avoiding d-q modeling to reduce processing stages. To demonstrate the feasibility of this approach, several simulations are presented considering a 1.5 kW PWM rectifier.

1. Introduction

The analysis of the PWM rectifiers is a very well-studied topic in both single-phase and three-phase applications due to its many advantages from the power electronics standpoint [1–3]. So, there is a lot of technical background related to it and most documented papers use a two-loop control scheme to obtain an adequate system performance, where the first loop is meant to control current with fast dynamics and the second one is meant to control output voltage with slow dynamics [4–12].

Also, several papers report the use of synchronous d-q frame transformation in order to calculate the compensation references [13, 14]. Besides, it is also usual to use d-q modeling for the control design. One powerful reason is that the tracking problem required in time domain is converted into a regulation problem in a d-q synchronous frame [15], which is a common practice, although it implies a certain degree of delay due to the transformation itself.

Considering the aforementioned, several nonlinear control schemes have been proposed in the literature in the last two decades in order to obtain the desired benefits of the topology, which include power factor correction, harmonic compensation, and, as an extended property, voltage sag ride through [16–22], in spite of the fact that these techniques are as old as feedback control [23].

Nevertheless, this paper analyzes the PWM rectifier with a different approach. Firstly, the d-q transformation is not used either to obtain references or to convert the system model from time domain to d-q domain. When a controller is tuned in d-q domain, the system should be perfectly synchronized with the AC mains and each stage dedicated to the d-q conversion and to the d-q-inverse conversion should maintain this synchronization. If AC mains have frequency variations of ±3%, the d-q conversion can fail even with smaller variations. However, with the controller proposed in this paper, this synchronization is avoided in each stage and the tracking problem is solved directly in time domain. Secondly, the use of two-loop control is avoided, considering only a single-loop current control by using a nonlinear controller based on its input-output linearization via feedback. The two above-mentioned points are considered in order to facilitate a fast dynamic response and to avoid overprocessing.

Table 1: Approaches reported in literature versus proposed approach.

Concept	Reported approaches	Proposed approach
d-q transformation	√	×
Control theory	Linear and nonlinear	Nonlinear
Current control loop	√	√
Voltage control loop	√	×
Load dynamics	Usually resistive with load steps	Nonlinear, different patterns of current consumption, with load steps
Robustness for source sags	√	√
Analysis complexity	Medium	Medium

Table 2: Switching combination for the single-phase PWM rectifier circuit.

Mode	Q_1	Q_2	Q_3	Q_4	sw_1	sw_2	$V_{a'b'}$	I_{DC}
I	0	1	0	1	1	1	0	0
II	0	1	1	0	0	1	$-V_{DC}$	$-i_L$
III	1	0	0	1	1	0	$+V_{DC}$	$+i_L$
IV	1	0	1	0	0	0	0	0

Now, the control objectives are to regulate the output voltage, to maintain power factor near to unity, and to achieve a sinusoidal current waveform even when voltage sags and dynamic load transients operate simultaneously, with an unexpected pattern. To highlight the new characteristics of the proposed control approach in this paper, Table 1 presents its main characteristics compared against those approaches reported in the literature related to the use of the d-q transformation and the control schemes with two loops.

Then, the development of the paper is organized as follows: Section 2 is dedicated to the development of the model of the PWM rectifier and showing the capabilities of the system. After that, in Section 3, the closed-loop system using the proposed control scheme is analyzed. In Section 4, simulations of the system considering two nonlinear loads with different profiles of current consumption appearing at different times and unexpected voltage sags are developed in order to show the benefits of this proposal. Finally, in Section 5, the authors state their conclusions of this research.

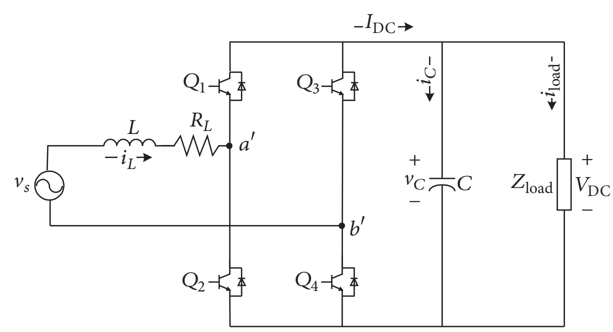

Figure 1: Single-phase PWM rectifier circuit.

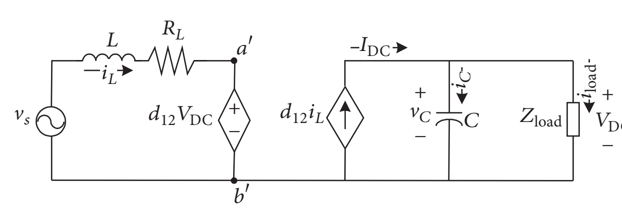

Figure 2: Simplified equivalent circuit of the single-phase PWM rectifier.

2. Single-Phase PWM Rectifier Modeling

In this section, the single-phase PWM rectifier model is presented. Figure 1 shows a typical single-phase rectifier circuit, with power factor correction capability.

Table 2 summarizes all the possible switching combinations for the single-phase PWM rectifier and their corresponding created full-bridge voltage, $V_{a'b'}$. Switches sw_1 and sw_2 are created to simplify the model analysis and besides it is considered that $V_{a'b'} = V_{DC}(sw_1 - sw_2)$; $I_{DC} = i_L(sw_1 - sw_2)$; and $sw_1, sw_2 \in \{1, 0\}$.

Based on previous considerations, analysis leads to the equivalent circuit shown in Figure 2, where $d_{12} = (d_1 - d_2)$ with $d_1 \in [1, 0]$, $d_2 \in [1, 0]$, and $d_{12} \in [-1, 1]$.

By doing the circuit analysis, the mathematical model of the system can be obtained:

$$\begin{aligned} \frac{di_L}{dt} &= -\frac{R_L}{L} i_L - \frac{1}{L} d_{12} v_C + \frac{1}{L} v_s, \\ \frac{dv_C}{dt} &= \frac{1}{C} d_{12} i_L - \frac{1}{C Z_{load}} v_C. \end{aligned} \tag{1}$$

Table 3: System test conditions.

Parameter	Value
$v_s(t)$	$180\sin(\omega t)$
V_{DC}	200 V
P_{load}	1500 W
Power factor	Near to unity
Maximum ripple in steady state	10%
Maximum overshoot in transient state	15%
Switching frequency	20 kHz
Load	Two inverters connected at different time: the first one of 750 W, switching frequency of 1 kHz, and output frequency of 60 Hz with 0° of phase shift and the second one of 750 W, switching frequency of 1.5 kHz, and output frequency of 50 Hz with 30° of phase shift

Then, considering the state variables,

$$\begin{bmatrix} i_L \\ v_C \end{bmatrix} = \begin{bmatrix} x_1 \\ x_2 \end{bmatrix}, \quad d_{12} = u. \tag{2}$$

Next, the state space system representation is attained:

$$\begin{aligned} \dot{x}_1 &= -\frac{R_L}{L}x_1 - \frac{1}{L}ux_2 + \frac{1}{L}v_s, \\ \dot{x}_2 &= \frac{1}{C}ux_1 - \frac{1}{CZ_{load}}x_2, \end{aligned} \tag{3}$$

where $Z_{load} = V_{DC}/i_{load}$.

Table 3 summarizes system specifications. It is worth mentioning that the initial analysis considers a maximum power capability of 1.5 kW as well as a highly dynamic load that does not exceed the maximum power limits. In order to obtain all the physical variables involved in the control loop, all sensors are considered available. However, these are the typical sensors used in power electronics applications, which means that there is no additional hardware with regard to other proposals.

3. Single-Phase PWM Rectifier Closed-Loop System

Before starting the design of the controller, it is important to establish clearly the control objectives, which always should be associated with the physical capabilities of the system [24]. From this point of view, the control objectives are the following:

(a) To regulate the output voltage v_C, which can be achieved considering the maximum power specified in Table 3. This condition should be maintained even when dynamic loads and input voltage sags occur simultaneously at unexpected times. The physical limitation is related to the amount of current that can be transferred by the PWM rectifier from the AC side to the DC side. Said amount of current is controlled by duty cycle

(b) To compensate the power factor

(c) To eliminate harmonic current, which means that the desired current waveform of i_L should be sinusoidal and in phase with the voltage v_s

3.1. Feedback Linearization-Based Nonlinear Control. Consider the single-input-single-output system

$$\begin{aligned} \dot{\mathbf{x}} &= f(\mathbf{x}) + g(\mathbf{x})u, \\ y &= h(\mathbf{x}), \end{aligned} \tag{4}$$

where f, g, and h are sufficiently smooth functions in a domain $D \in \mathfrak{R}^n$ and the mappings $f : D \rightarrow \mathfrak{R}^n$ and $g : D \rightarrow \mathfrak{R}^n$ are called vector fields on D [25, 26]. Consider also the relative degree ρ as the number of derivates of the output to find the input. If $\rho < n$, then a diffeomorphism restricted to a neighborhood N can be built as

$$\mathbf{z} = T(\mathbf{x}) = \begin{bmatrix} \phi_1(\mathbf{x}) \\ \vdots \\ \phi_{n-\rho}(\mathbf{x}) \\ \cdots \\ h(\mathbf{x}) \\ \vdots \\ L_f^{\rho-1}h(\mathbf{x}) \end{bmatrix} = \begin{bmatrix} \eta \\ \xi \end{bmatrix}, \tag{5}$$

where $\eta \in \mathfrak{R}^{n-\rho}$, $\xi \in \mathfrak{R}^{\rho}$, and ϕ_i for $i = 1, \ldots, n-\rho$ must satisfy

$$\frac{\partial \phi_i(\mathbf{x})}{\partial \mathbf{x}} g(\mathbf{x}) = 0, \quad \text{for } 1 \le i \le n-\rho, \ \forall x \in D_0 \tag{6}$$

with $D_0 \subset D$. Then, (4) can be expressed in its normal form given by

$$\begin{aligned} \dot{\eta} &= f_0(\eta, \xi), \\ \dot{\xi} &= A_c\xi + B_c\gamma(\mathbf{x})[u - \alpha(\mathbf{x})], \\ y &= C_c\xi, \end{aligned} \tag{7}$$

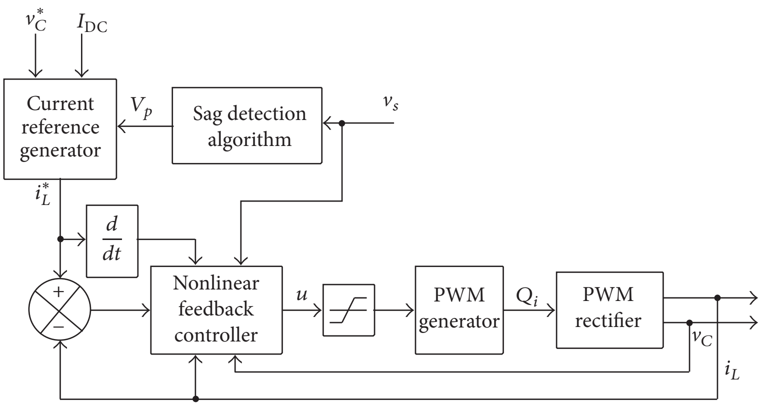

Figure 3: Proposed closed-loop control for the single-phase PWM rectifier.

where (A_c, B_c, C_c) is a canonical form representation and

$$f_0(\eta, \xi) = \frac{\partial \phi(\mathbf{x})}{\partial \mathbf{x}} f(\mathbf{x})\Big|_{\mathbf{x}=T^{-1}(\mathbf{z})},$$
$$\alpha(\mathbf{x}) = -\frac{L_f^{\rho} h(\mathbf{x})}{L_g L_f^{\rho-1} h(\mathbf{x})}, \tag{8}$$
$$\gamma(\mathbf{x}) = L_g L_f^{\rho-1} h(\mathbf{x}).$$

Equation (7) is formed by an external part ξ and an internal part η. The external part can be linearized by the controller

$$u = \alpha(\mathbf{x}) + \beta(\mathbf{x}) v \tag{9}$$

with $\beta = \gamma^{-1}(\mathbf{x})$ and $v = -\mathbf{k}\mathbf{x}$. Meanwhile, the zero dynamics

$$\dot{\eta} = f_0(\eta, 0) \tag{10}$$

must be asymptotically stable; that is, the system must be of minimum phase.

3.2. Design of the Nonlinear Controller. Then, according to the control objectives and the feedback linearization-based nonlinear control theory, a tracking problem of the current i_L will be the study objective. For this, the error dynamics of the system should be considered. So, the proposed closed-loop system to be designed is depicted in Figure 3, which consists of a sag detection algorithm, a reference generator, and the single-loop nonlinear controller.

Based on the state space representation (3),

$$f(\mathbf{x}) = \begin{bmatrix} -\frac{R_L}{L}x_1 + \frac{1}{L}v_s \\ -\frac{1}{CZ_{\text{load}}}x_2 \end{bmatrix},$$
$$g(\mathbf{x}) = \begin{bmatrix} -\frac{1}{L}x_2 \\ \frac{1}{C}x_1 \end{bmatrix}, \tag{11}$$

and it can be verified that the system is input-output linearizable with relative degree $\rho = 1$ for $y = x_1$ with

$$\dot{y} = \dot{x}_1 = -\frac{R_L}{L}x_1 - \frac{1}{L}ux_2 + \frac{1}{L}v_s. \tag{12}$$

Then, the following diffeomorphism can be built:

$$\mathbf{z} = \begin{bmatrix} T_1(\mathbf{x}) \\ T_2(\mathbf{x}) \end{bmatrix} = \begin{bmatrix} \phi_1(\mathbf{x}) \\ h(\mathbf{x}) \end{bmatrix} = \begin{bmatrix} \eta \\ \xi \end{bmatrix} = \begin{bmatrix} \frac{Lx_1^2 + Cx_2^2}{C} \\ x_1 \end{bmatrix}, \tag{13}$$

defined $\forall \mathbf{x} \in \mathfrak{R}^2 \mid x_2 \neq 0$, and it is possible to achieve its normal form (7) as

$$\dot{\eta} = -\frac{2R_L}{C}\xi^2 + \frac{2L}{C^2 Z_{\text{load}}}\xi^2 - \frac{2}{CZ_{\text{load}}}\eta + \frac{2}{C}v_s\xi,$$
$$\dot{\xi} = -\frac{x_2}{L}\left[u - \frac{v_s - R_L x_1}{x_2}\right], \tag{14}$$
$$y = \xi,$$

where $\eta \in \mathfrak{R}^1$, $\xi \in \mathfrak{R}^1$, and

$$f_0(\eta, \xi) = -\frac{2R_L}{C}\xi^2 + \frac{2L}{C^2 Z_{\text{load}}}\xi^2 - \frac{2}{CZ_{\text{load}}}\eta + \frac{2}{C}v_s\xi \tag{15}$$

with

$$\mathbf{x} = T^{-1}(\mathbf{z}) = \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} = \begin{bmatrix} \xi \\ \left(\frac{C\eta - L\xi^2}{C}\right)^{1/2} \end{bmatrix},$$
$$\alpha(\mathbf{x}) = \frac{v_s - R_L x_1}{x_2},$$
$$\gamma(\mathbf{x}) = -\frac{x_2}{L}, \tag{16}$$
$$A_c = [0],$$
$$B_c = [1],$$
$$C_c = [1].$$

Then, its internal dynamics is governed by

$$\dot{\eta} = -\frac{2}{CZ_{\text{load}}}\eta, \tag{17}$$

which means that the zero dynamics of the system is stable; so the nonlinear system is of minimum phase with positive Z_{load} and C.

As previously stated, one of the control objectives is the power factor correction. Therefore, the control problem is to track the reference i_L^*. Besides, consider the following:

(i) Reference current i_L^* and its derivative $\dot{i}_L^*$ are bounded for every $t \geq 0$ and $\dot{i}_L^*$ is piecewise continuous in t.

(ii) Signals i_L^* and $\dot{i}_L^*$ are available.

Then, considering its error dynamics, the tracking control law for the PWM rectifier system is given by

$$u = \alpha(\mathbf{x}) + \frac{1}{\gamma(\mathbf{x})}\left\{-k\left(T_2(\mathbf{x}) - i_L^*\right) + \dot{i}_L^*\right\}. \tag{18}$$

The value of $k = 13000$ is calculated such that $A_c - B_c k$ is to be Hurwitz for $v = -k(\xi - \xi^*)$. Then the tracking control law designed for (1) is

$$d_{12} = \frac{v_s - R_L i_L}{v_C} - \frac{L}{v_C}\left\{-13000\left(i_L - i_L^*\right) + \dot{i}_L^*\right\}. \tag{19}$$

So, the current tracking is achieved asymptotically for all t which can be seen in Section 4.

As a part of the design, a sensor that measures I_{DC} is included. Now, based on the fact that one of the control objectives is to regulate the DC output voltage in $200V_{\text{DC}}$, this measurement is used to instantaneously calculate the reference to be supplied to the controller. In this way, it is possible to use only one feedback loop $i_L - i_L^*$ as (19) specifies. Also consider that measurements of voltage and current are needed, as in any other control scheme.

The system parameters are $L = 1\,\text{mH}$, $R_L = 1\,\Omega$, and $C = 1200\,\mu\text{F}$. These values can be calculated using any of the procedures reported in power converters literature taking into account the maximum power and the switching frequency proposed in this paper.

3.3. Reference Current Generator. In order to generate the reference signal i_L^*, the AC power on the source voltage side is considered equal to the power on the DC side; from this, it can be obtained that

$$\begin{gathered} P_{\text{AC}} = \frac{V_p I_p}{2} = P_{\text{DC}} = V_{\text{DC}} I_{\text{DC}}, \\ \frac{V_p I_p}{2} = V_{\text{DC}} I_{\text{DC}}, \\ I_p^* = \frac{2 v_C^* I_{\text{DC}}}{V_p}, \end{gathered} \tag{20}$$

where V_p and I_p are the peak values of v_s and i_L, respectively, and $V_{\text{DC}} = v_C^*$.

Since the current is in phase with the source voltage v_s in order to have a unitary power factor,

$$\begin{aligned} i_L^* &= I_p^* \sin(2\pi f t), \\ \dot{i}_L^* &= 2\pi f I_p^* \cos(2\pi f t). \end{aligned} \tag{21}$$

So, based on (21), it can be concluded that i_L^* satisfies the considerations previously stated for it.

4. Simulation Results

In this section, a set of simulation tests is presented considering the operational conditions described in Table 3. The application of this nonlinear control scheme demonstrates how the system achieves the performance and control objectives imposed as a part of the design. Considering Figure 4, the tests were performed as follows:

(i) The simulator used was Simulink of MATLAB applying ODE15s as integration method with a maximum integration step of $1\,\mu$s.

(ii) The sequence of tests is as follows: from 0 s to 0.1 s, the PWM rectifier is functioning as an uncontrolled rectifier. This period of time is identified as Region I.

(iii) From 0.1 s to 0.2 s, the PWM rectifier operates as a power factor corrector and reaches steady state. During this time, the load is a 750 W power inverter with switching frequency of 1 kHz, modulation index $m = 0.9$, and output voltage at 60 Hz with zero degrees of phase shift. This period of time is identified as Region II.

(iv) A voltage sag to 75% occurs at 0.2 s. The time between 0.2 s and 0.3 s is identified as Region III.

(v) At time 0.3 s, an additional power inverter of 750 W with switching frequency of 1.5 kHz, modulation index $m = 0.6$, and output voltage at 50 Hz with thirty degrees of phase shift is connected. The voltage sag continues. The time between 0.3 and 0.4 s is identified as Region IV.

(vi) At time 0.4 s, the voltage sag ends, but the two inverters with 1.5 kW of consumption remain connected. The time between 0.4 and 0.5 s is identified as Region V.

Figure 5 shows the evolution of all regions mentioned above. The voltage supplied to the PWM rectifier v_s is shown on (a), while on (b) the reference current i_L^* and the AC current consumption of the system are plotted. It can be noticed that in Region I both signals are different because the system is working as a noncontrolled rectifier; however, once Region II is reached, the current consumption is sinusoidal and is in phase with the voltage during the rest of the test, which means that at all times power factor correction and harmonic elimination are achieved.

A general scope of the system performance is shown in Figure 6, where the current consumption of the nonlinear load based on power inverters connected as described for

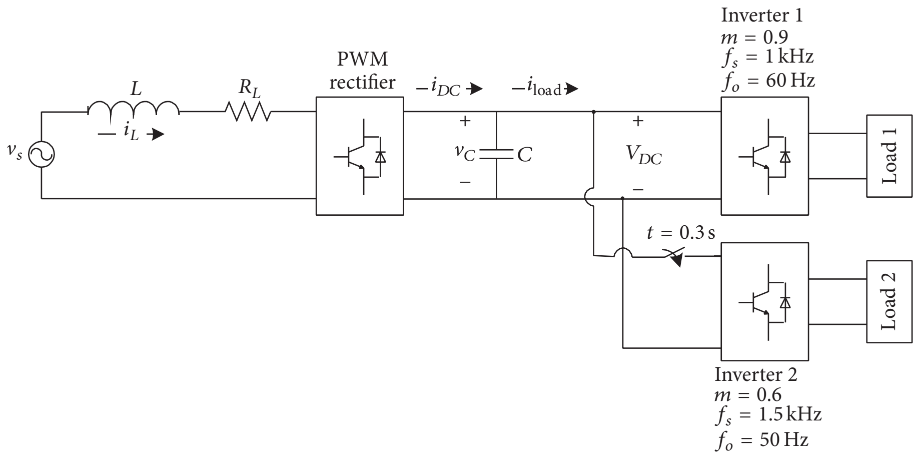

Figure 4: Test scheme to evaluate the performance of the closed-loop system.

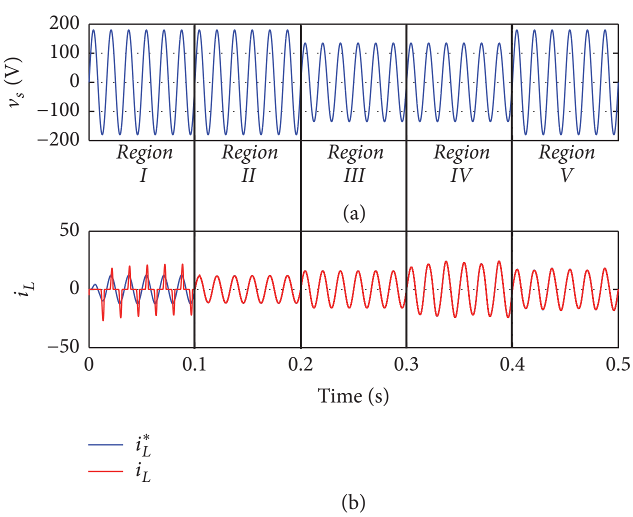

Figure 5: (a) Voltage delivered by the grid to the PWM rectifier. (b) Reference current and current consumption of the system.

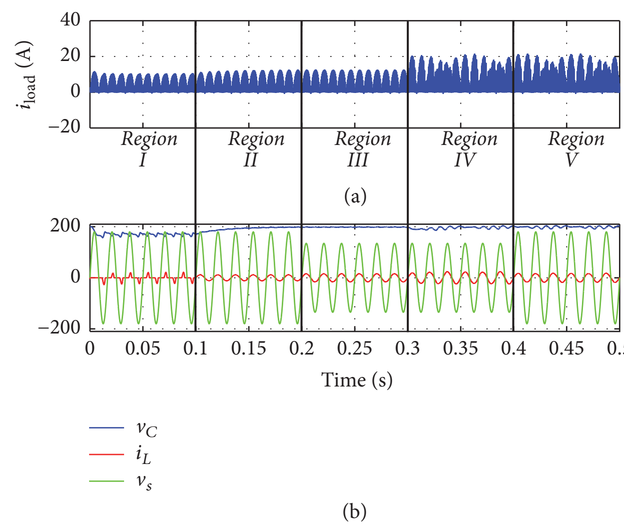

Figure 6: (a) Current consumption of the inverters. (b) DC voltage, AC current, and AC voltage.

each region can be observed on (a). On (b), the grid voltage supplied to the system with voltage sags events, the current consumption of the PWM rectifier, and the output DC voltage are plotted. It is important to notice that while in Region I the DC voltage is regulated at 180 V (because the system is working as a noncontrolled rectifier), when the PWM rectifier starts to work at 0.1 s, the DC voltage is regulated at 200 V because of its boost properties. It is also worth mentioning that, during the entire testing period, all the control and operative objectives set for the PWM rectifier were reached with only a one-loop controller, without the use of d-q transformation and considering realistic unexpected simultaneous disturbances.

The PWM rectifier has been designed considering power consumption of 1.5 kW with the tolerances shown in Table 3. Considering constant current consumption with a constant, ripple-free DC voltage, the operation point of the system corresponds to 7.5 A and 200 V, which provides 1.5 kW. However, the industrial applications impose dynamic loads and power quality problems. Therefore, it is normal to consider a certain margin of tolerance as a standard operation condition and then, only in terms of realistic performance, to establish an operation region in which the system will function as normal. Figure 7 shows a plot of v_C versus i_{load} considering all the scenarios shown in each region. As it can be observed, DC voltage is regulated inside the 10% tolerance, even during load transients or AC voltage sags. This is an important contribution of this work in terms of providing the system with robustness against undesirable events such as voltage sags, which are common in electrical networks.

5. Conclusion

In PWM rectifiers, it is very common to consider a two-loop control scheme in order to obtain a good dynamic response and output voltage regulation. At the same time, the d-q

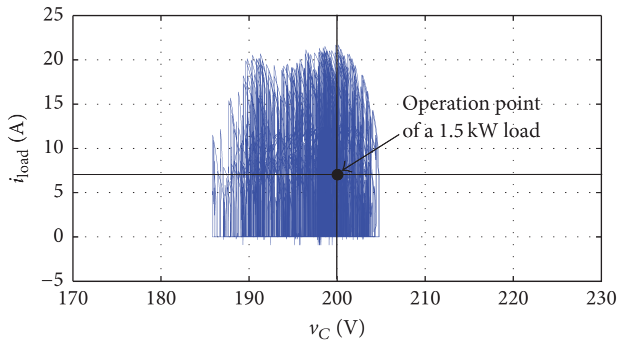

FIGURE 7: Operation region in terms of v_C versus i_{load}. The limits of the scale in abscissa axis correspond to a maximum variation of ±15% with respect to 200 V.

synchronous frame has been widely applied to these systems either to obtain references or to transform the mathematical model. Such considerations have given good results but require more processing, meaning time delays which can be critical, especially in transient situations.

Although the use of control in d-q frame has been reported with good results, in practical terms, frequency variations occur in electrical networks and they produce mathematical inconsistences in the transformation which generate a decreasing performance unless a perfect synchronization is carried out considering a computational effort. The control scheme proposed in this paper does not require any transformation; therefore performance is satisfactory even when frequency variations of the network occur.

This paper analyzes a different approach dependent on the feedback linearization-based nonlinear control and analyzing the stability of the system. Considering this approach, the authors explore the benefits of avoiding the d-q transformation applied to the model and they demonstrate the feasibility of providing the PWM rectifier with robustness and good dynamic response by using only a single-loop controller. Moreover, in order to show the capability of operating the system with strong variations, a set of simulation tests was performed considering realistic scenarios such as nonlinear dynamic loads and voltage sags. The results were very promising because the simulations revealed that, at any time, the control objectives initially established were reached.

Conflicts of Interest

The authors declare that they have no conflicts of interest.

Acknowledgments

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