A Nonlinear Control Scheme for Single-Phase PWM Multilevel Rectifier

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***Abstract*— A single phase active rectifier is commonly used to address two issues in electric power delivery: DC voltage regulation and AC power factor improvement through current tracking. In this paper we propose a nonlinear controller that solves the DC voltage regulation and AC current tracking for the single phase active rectifier. The proposed controller is based on a partial input–output feedback linearization and an adaptive indirect control of the zero dynamics. Numerical simulations are included to evaluate the performance of the proposed controller.**

I. INTRODUCTION

Electrical energy quality is a broad term that covers a great variety of electrical energy signal disturbances. Low quality energy, highly disturbed electrical energy signals, may have negative effects in electricity distribution networks. The broad use of electronic equipment drawing highly nonlinear currents has attracted the attention on power quality for domestic, industrial and commercial electric energy consumers (Bollen, 2000). Power quality is concerned with deviations of current and voltage from the ideal waveform and differences in their phase. The ideal waveform is a single frequency sine wave of constant frequency and magnitude.

Some years ago, electric equipment was very simple so that it was more robust and insensitive to ac mains variations. Nowadays, the use of electronic devices has increased. Most electronic devices use a DC source that first rectifies the ac voltage with a diode bridge followed by a big capacitor (Maswood, et al, 2006). Some advantages of this sources are low cost, simple structure, robustness and control absence, however, they have a low power factor and inject harmonic currents into the ac mains.

A solution is to use active rectifiers whose advantages are power factor improvement, harmonic current distortion reduction and DC voltage regulation. There are several works about this kind of rectifiers. For instance, in (Bor-Ren, et al, 1999)-(Choi, 2005) a hysteresis based current control with load changes is presented. Two control techniques are discussed in (Salaet, et al, 2004), however, the DC regulation is not satisfactory because DC voltage vary when ac mains voltage vary. A proportional current control scheme and a PI voltage regulator are used in (Bor-Ren, et al, 2004), showing just steady state results. In (Joong, et al, 1997), a PWM modulation technique which allows an adequate DC voltage regulation in steady state is presented. A PI control scheme is used in (Bor-Ren, et al, 2003) and (Cichowlas, et al, 2005) for current and voltage control, it presents load change tests.

In this paper we propose a nonlinear controller to solve the DC link regulation and current tracking problems for a single-phase PWM multilevel rectifier. The proposed controller is based on a partial input–output feedback linearization and the indirect control of the zero dynamics. Numerical simulations are included to evaluate the performance of the proposed controller. This paper is organized in the following manner. The single-phase active rectifier model is exposed in section II. The control scheme is described in section III. The simulation results are discussed in Section IV, followed by some concluding remarks in Section V.

II. MULTILEVEL RECTIFIER DYNAMICS

The single-phase PWM multilevel active rectifier topology studied in this paper is shown in fig. 1. The differential equations that describe the rectifier dynamics, according to

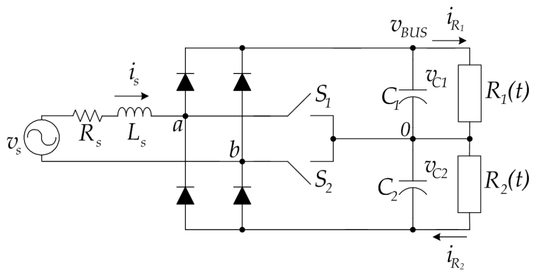

Figure 1. Single-phase active rectifier topology.

(Bor-Ren, et al, 1999), are

$$\begin{aligned} L_s\frac{d}{dt}i_s &= -v_{ab}+v_s \\ C_1\frac{d}{dt}v_{C_1} &= i_1-i_{L_1} \\ C_2\frac{d}{dt}v_{C_2} &= i_2-i_{L_2} \end{aligned} \tag{1}$$

where

$$\begin{aligned} v_{ab} &= \tfrac{1}{2}\left[\text{sgn}(i_s)+1\right]\left[v_{C_1}(1-S_1)+v_{C_2}(1-S_2)\right] \\ &\quad +\tfrac{1}{2}\left[\text{sgn}(i_s)-1\right]\left[v_{C_1}(1-S_2)+v_{C_2}(1-S_1)\right] \\ i_1 &= \tfrac{1}{2}\left[\text{sgn}(i_s)+1\right](1-S_1)\,i_s \\ &\quad +\tfrac{1}{2}\left[\text{sgn}(i_s)-1\right](1-S_2)\,i_s \\ i_2 &= \tfrac{1}{2}\left[\text{sgn}(i_s)+1\right](1-S_2)\,i_s \\ &\quad +\tfrac{1}{2}\left[\text{sgn}(i_s)-1\right](1-S_1)\,i_s \end{aligned}$$

Here, we consider a two-level PWM modulation technique. In this modulation technique, the DC link is controlled to be greater than v_{sp} and two ac power switches are closed and opened together, i.e., $S_1=S_2$. Thus, the PWM rectifier dynamics (1) becomes

$$\begin{aligned} L_s\frac{d}{dt}i_s &= -\text{sgn}(i_s)(1-S_1)(v_{C_1}+v_{C_2})+v_s \\ C_1\frac{d}{dt}v_{C_1} &= \text{sgn}(i_s)(1-S_1)i_s-i_{L_1} \\ C_2\frac{d}{dt}v_{C_2} &= \text{sgn}(i_s)(1-S_1)i_s-i_{L_2} \end{aligned} \tag{2}$$

A continuos form of the discontinous PWM rectifier dynamic model (2) has been introduced in (Flota, et al, 2009) replacing the discontinuos function sgn(i_s) by the continuos function $f(i_s)=\frac{2}{\pi}\arctan(ai_s)$ with a a constant parameter that modifies the slope of $f(i_s)$ around $i_s=0$. Using $f(i_s)$ the PWM rectifier model becomes

$$\begin{aligned} L_s\frac{d}{dt}i_s &= -f(i_s)(1-S_1)(v_{C_1}+v_{C_2})+v_s \\ C_1\frac{d}{dt}v_{C_1} &= f(i_s)(1-S_1)i_s-i_{L_1} \\ C_2\frac{d}{dt}v_{C_2} &= f(i_s)(1-S_1)i_s-i_{L_2} \end{aligned} \tag{3}$$

The control objective is to regulate the DC-link voltage to a desired value V_{DC} maintaining a unity power factor.

III. Control Design

This Section is devoted to control design. First, appealing to time scale separation we reduce the three dimensional model of the single phase rectifier (3) to a two dimensional set of differential equations based on the fact that the dynamics of the voltage difference $v_{C_1}-v_{C_2}$ is much faster than the inductor and the total voltage $v_{C_1}+v_{C_2}$ dynamics. Then, we perform an input–ouput feedback linearization taking as the output the inductor current and as the input the switch position. In this case, the zero dynamics is described by the total voltage dynamics. We command this dynamics indirectly through the inductor current reference.

Consider the following change of coordinates

$$\begin{aligned} V_T &= v_{C_1}+v_{C_2} \\ V_D &= v_{C_1}-v_{C_2} \\ u &= (1-S_1) \end{aligned}$$

Clearly, V_T is the total voltage and V_D is the voltage difference. In terms of the above coordinates, the rectifier dynamics is given by

$$\begin{aligned} L_s\frac{d}{dt}i_s &= -f(i_s)V_Tu+v_s \\ C_e\frac{d}{dt}V_T &= f(i_s)i_su-\tfrac{C_e}{C_1}i_{L_1}-\tfrac{C_e}{C_2}i_{L_2} \\ C_1C_2\frac{d}{dt}V_D &= (C_2-C_1)\,f(i_s)i_su-\tfrac{1}{C_2}i_{L_1}+\tfrac{1}{C_1}i_{L_2} \end{aligned} \tag{4}$$

with $\frac{1}{C_e}=\frac{1}{C_1}+\frac{1}{C_2}$. Note that

$$C_1C_2 << C_e < L_s$$

Thus, the V_D dynamics is faster than the i_s and V_T dynamics. As a consequence, for control design we consider the reduced model

$$\begin{aligned} L_s\frac{d}{dt}i_s &= -f(i_s)V_Tu+v_s \\ C_e\frac{d}{dt}V_T &= f(i_s)i_su-\tfrac{C_e}{C_1}i_{L_1}-\tfrac{C_e}{C_2}i_{L_2} \end{aligned} \tag{5}$$

Considering a current control strategy we define

$$\tilde{y}=i_s-i_s^*$$

whose time derivative along (5) is

$$\dot{\tilde{y}}=\frac{1}{L_s}\left(-f(i_s)V_Tu+v_s\right)-\frac{d}{dt}i_s^* \tag{6}$$

Note that in (6) it is possible to achieve input–output feedback linearization by a suitable selection of u. However, in this control design we just perform a partial linearization. Defining

$$u=f(i_s)\frac{L_sk\tilde{y}+v_s}{V_T} \tag{7}$$

we obtain

$$\dot{\tilde{y}}=-kf(i_s)^2\tilde{y}+\frac{v_s}{L_s}\left[1-f(i_s)^2\right]-\frac{d}{dt}i_s^*$$

Let us define $\mu = \frac{1}{k}$, then

$$\mu\dot{\tilde{y}} = -f(i_s)^2\tilde{y} + \mu\left\{\frac{v_s}{L_s}\left[1 - f(i_s)^2\right] - \frac{d}{dt}i_s^*\right\}$$

Thus, there exist a big enough k such that

$$\tilde{y} \to 0$$

arbitarily fast.

Consider now the second equation of (5) in closed-loop with (7), that is,

$$\begin{aligned} C_e\frac{d}{dt}V_T &= f(i_s)^2\frac{v_s}{V_T}i_s^* + f(i_s)^2\left(kL_si_s + v_s\right)\frac{\tilde{y}}{V_T} \\ &\quad -\frac{C_e}{C_1}i_{L_1} - \frac{C_e}{C_2}i_{L_2} \end{aligned} \tag{8}$$

By defining $\sigma = \frac{1}{2}C_eV_T^2$ the dynamics above can be written as follows

$$\dot{\sigma} = f(i_s)^2v_si_s^* + \zeta(t) \tag{9}$$

with

$$\zeta(t) = f(i_s)^2\left(kL_si_s + v_s\right)\tilde{y} - \sqrt{2\sigma C_e}\left(\frac{1}{C_1}i_{L_1} + \frac{1}{C_2}i_{L_2}\right)$$

Note that in equation (9), i_s^* has not been defined so that it can be considered as the control input and the disturbance $\zeta(t)$ is composed of vanishing terms and an unknown terms.

In order to determine i_s^* such that σ converges to $\sigma^* = \frac{1}{2}C_eV_{DC}^2$ we start by designing an estimator to obtain an estimated value of $\zeta(t)$. Define the estimation errors as follows

$$\begin{aligned} z_1 &= \zeta(t) - \rho_1 + \beta_1(\sigma) \\ z_2 &= \dot{\zeta}(t) - \rho_2 + \beta_2(\sigma) \\ z_3 &= \ddot{\zeta}(t) - \rho_3 + \beta_3(\sigma) \end{aligned}$$

The time derivative of the estimation errors is described by the following equations

$$\begin{aligned} \dot{z}_1 &= \frac{\partial\beta_1}{\partial\sigma}z_1 + z_2 \\ \dot{z}_2 &= \frac{\partial\beta_2}{\partial\sigma}z_1 + z_3 \\ \dot{z}_3 &= \frac{\partial\beta_3}{\partial\sigma}z_1 \end{aligned} \tag{10}$$

where we have supposed that $\zeta(t)^3 \approx 0$ and defined

$$\begin{aligned} \dot{\rho}_1 &= \rho_2 - \beta_2 + \frac{\partial\beta_1}{\partial\sigma}\left(f(i_s)^2v_si_s^* + \rho_1 - \beta_1\right) \\ \dot{\rho}_2 &= \rho_3 - \beta_3 + \frac{\partial\beta_2}{\partial\sigma}\left(f(i_s)^2v_si_s^* + \rho_1 - \beta_1\right) \\ \dot{\rho}_3 &= \frac{\partial\beta_3}{\partial\sigma}\left(f(i_s)^2v_si_s^* + \rho_1 - \beta_1\right) \end{aligned} \tag{11}$$

Assume now that

$$\frac{\partial\beta_i}{\partial\sigma} = k_i$$

then, the estimation error dynamics (10) can be written as

$$z_1^3 - k_1\ddot{z}_1 - k_2\dot{z}_1 - k_3z_1 = 0$$

it is clear that there exists $k_i, i = 1, 2, 3$ such that the above equation is exponentially stable and an asymptoyic estimate of $\zeta(t)$ is given by $\rho_1 - \beta_1(\sigma)$.

Now we are in position to define the virtual control i_s^*. Note that the following definition

$$\begin{aligned} i_s^* &= \frac{v_s}{v_{s_p}^2}\left(-k_p\tilde{\sigma} - k_i\eta - \rho_1 + \beta_1(\sigma)\right) \\ \dot{\eta} &= \tilde{\sigma} \end{aligned}$$

with $\tilde{\sigma} = \sigma - \sigma^*$, gives the following closed loop dynamics

$$\begin{aligned} \dot{\tilde{\sigma}} &= f(i_s)^2\frac{v_s^2}{v_{s_p}^2}\left(-k_p\tilde{\sigma} - k_i\eta + z_1\right) \\ &\quad + \left(1 - f(i_s)^2\frac{v_s^2}{v_{s_p}^2}\right)\zeta(t) \\ \dot{\eta} &= \tilde{\sigma} \end{aligned} \tag{12}$$

Note that in equation (12) the time variant signal satisfies

$$1 \geq f(i_s)^2\frac{v_s^2}{v_{s_p}^2} \geq 0 \tag{13}$$

At this point we do not have a formal proof concerning the stability of the closed–loop dynamics.

IV. Simulation results

In order to verify the performance of the control scheme several simulation tests were done. The rectifier parameters are presented in table I.

TABLE I

Active rectifier parameters.

Description	Value
Power	$1\ kVA$
AC mains voltage	$127\ V_{RMS}$
DC voltage	$200\ V$ regulated
DC capacitors (C_1 and C_2)	$2200\ \mu F$
Inductor (L_S)	$5.0\ mH$
Inductor associated resistance (R_S)	$1.0\ \Omega$

The performance of the nonlinear controller with the active rectifier operating at steady-state is presented in figs. 2, 3, and 4. As it can be noted, the inductor current is almost sinusoidal (fig. 3). The DC voltage is shown in figure 4, it is regulated all the time with a 5.0% ripple with respect to the desired voltage ($200\ V$).

Another test to investigate the dynamic response of the control scheme was performed, the load change between $600 - 1000\ W$, $1000 - 600\ W$, $600 - 1000\ W$, and $1000 - 600\ W$ at 2.0, 4.0, 6.0, and $8.0\ s$, respectively, with a swell (20 %) and a sag (20 %) was tested and shown in figs. 5, 6, and 7. The swell occurs from 2.0 to $4.0\ s$ and sag from 6.0 to $8.0\ s$. The THD of i_s is less than 6% in this test.

V. Acknowledgements

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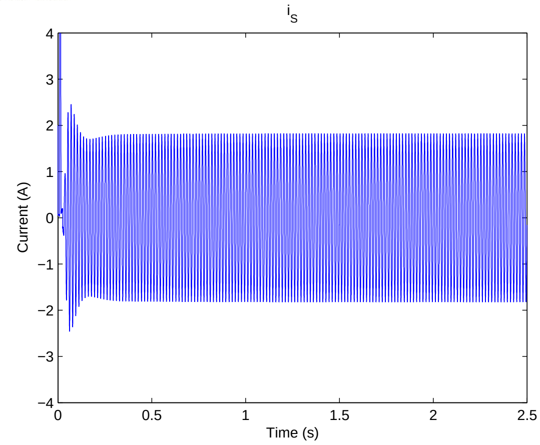

Figure 2. Inductor current at steady-state.

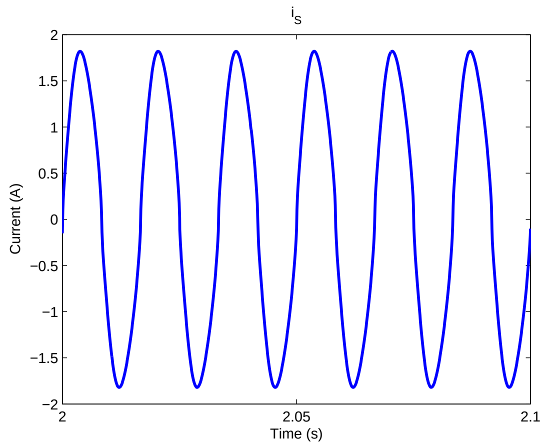

Figure 3. Inductor current at steady-state (zoom).

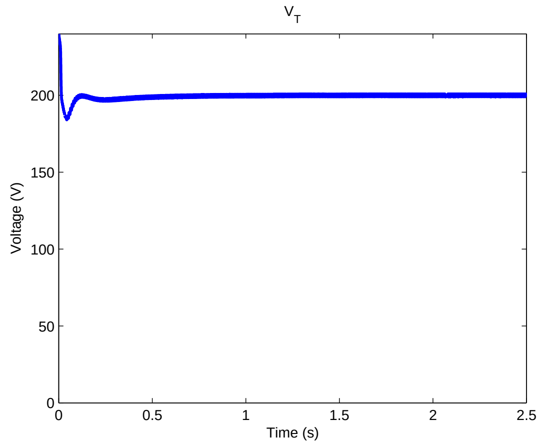

Figure 4. DC voltage at steady-state.

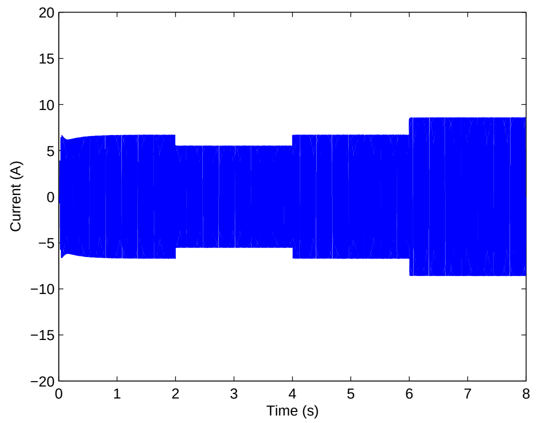

Figure 5. Inductor current when load change occurs between $600 - 1000\ W$ and $600 - 1000\ W$ with swell (20%) and sag (20%).

VI. Conclusions

We have proposed a nonlinear control strategy to solve the DC link voltage regulation and the current tracking problems in a single phase active rectifier. The proposed controller is based on a partial input–output feedback linearization and the adaptive indirect control of the zero dynamics. The proposed solution obviates the output power estimator. Numerical simulations illustrate the performance of the proposed solution.

An important point is left open in this paper: a formal proof concerning the stability of the closed–loop dynamics.

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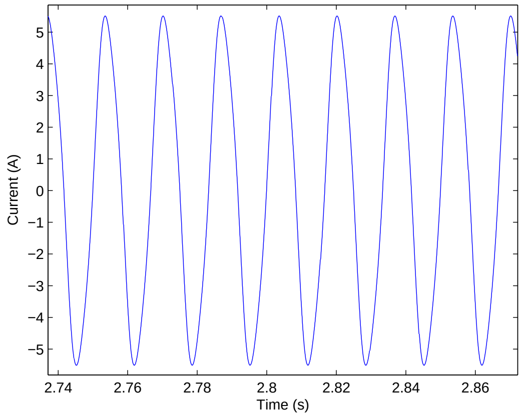

Figure 6. Inductor current when load change occurs between $600-1000\ W$ and $600-1000\ W$ with swell (20%) and sag (20%) (zoom).

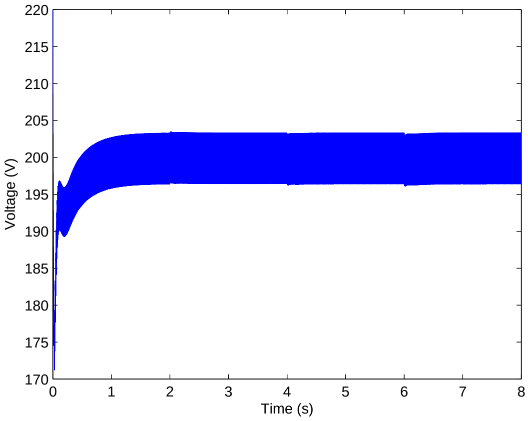

Figure 7. DC voltage when load change occurs between $600-1000\ W$ and $600-1000\ W$ with swell (20%) and sag (20%).

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