PASSIVITY–BASED PI CONTROL OF SWITCHED POWER CONVERTERS

Marcelo Pérez*, **Romeo Ortega**† **and José Espinoza***

* Electrical Engineering Department
Universidad de Concepción,
CHILE
tel.:(56)-41-203512
fax:(56)-41-246999
mperez@die.udec.cl
jespinoz@die.udec.cl

† Laboratorie des Signaux et Systèmes
CNRS-SUPELEC
Gif–sur–Yvette 91192
FRANCE
tel.:(33)-01-69-85-17-65
fax:(33)-01-69-85-17-66
Romeo.Ortega@lss.supelec.fr

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Abstract

This paper addresses the practically important question of when a nonlinear system can be asymptotically stabilized with a linear PI control, and show that it is possible if the system can be passified with constant control actions. This result is used to propose a constructive methodology to design PI schemes for switched power converters, which are viewed from a novel perspective: with the switches acting as effective "port variables" that establish a passive mapping with suitably defined outputs. Detailed calculations are given for a three–phase rectifier and simulations with a realistic model compares the performance of the proposed PI with other linear and nonlinear designs.[1]

1 Introduction

Passive systems constitute a very important class of dynamical systems for which the stored energy cannot exceed the energy supplied to them by the external environment. In view of this energy–balancing feature, it is clear that passivity is intimately related with the property of stability, a *sine qua non* condition for any controller design. Furthermore, invoking the universal principle of energy conservation, it may be argued that all physical systems are passive with respect to some suitably defined port variables that couple the system with the environment. It is not surprising then that, since the introduction of the first passivity–based controller (PBC) more than two decades ago [1, 2], we have witnessed an ever increasing popularity of passivity as a building block for controller design for all classes of physical systems.

One of the main practical advantages of PBC is that it allows to explain why, and more importantly identify when, a simple control scheme, like PID, can stably regulate the behavior of a coarsely defined (possibly complicated nonlinear) plant. In [3], it is shown that passive systems with a detectable output can be stabilized with strictly passive controllers. In this paper we prove, a simple slight variation of this basic result, namely that if an input–affine nonlinear system is passifiable via a constant control action, then it is stabilizable with a PI controller that does not require the knowledge of the constant term. We use this fact to propose a methodology to design PI controllers for a large class of switched power converters. The procedure is constructive an identifies the passifiable output–which fed–back through the PI controller will ensure stability. If this output turns out to be detectable then stability is, furthermore, asymptotic.

We use the methodology to derive, for the standard nonlinear model of a power rectifier, a novel PI controller that exhibits remarkable robustness and transient performance properties. Interestingly, one of the passive outputs that we identify is related with the difference between supplied and extracted instantaneous active powers (of a suitably scaled representation) of the rectifier. In this way we make a nice connection with the widely popular PQ Instantaneous Power controllers of [4, 7], where an outer PI loop around the output voltage is used to generate a reference for an inner PI loop acting on the aforementioned power difference. Another contribution of our paper is a detailed simulation study comparing the proposed PI with classical schemes as well as linearizing controllers.

2 A class of nonlinear systems stabilizable via PI control

In this section we prove the simple fact that systems that can be passified with constant controls are PI stabilizable.

Proposition 1. *Consider the system*

$$\dot{x} = f(x) + g(x)u, \tag{1}$$

with state $x \in \mathbb{R}^n$ *and control* $u \in \mathbb{R}^m$*, and an admissible equilibrium point* $x_\star \in \mathbb{R}^n$ *to be stabilized, such that:*

(H1) (Passifiability via constant control)

$$\left(\frac{\partial V}{\partial x}\right)^\top g(x) = h^\top(x) \tag{2}$$

$$\left(\frac{\partial V}{\partial x}\right)^\top [f(x) + g(x)\theta] \leq 0. \tag{3}$$

[1]This is an abridged version of the full paper which is available upon request to the authors.

Then, for all initial conditions $(x(0), z(0))$, the trajectories of the system (1) in closed–loop with the PI controller

$$\begin{array}{rcl} y &=& h(x) \\ u &=& -K_p y + z \\ \dot{z} &=& -K_i y \end{array} \tag{4}$$

with $K_p, K_i \in \mathbb{R}^{m\times m}$ symmetric positive definite matrices, are bounded and

$$\lim_{t\to\infty} y(t) = 0.$$

Furthermore, $lim_{t\to\infty} x(t) = x_\star$, if additionally we have that

(H2) (Detectability) *The equilibrium $x_\star$ is locally detectable from the output. That is, for any solution $x(t)$ of the system (1) which belongs to some open neighborhood of the equilibrium for all $t \geq 0$, the following implication is true:*

$$h(x(t)) \equiv 0 \Rightarrow \lim_{t\to\infty} x(t) = x_\star.$$

Remark 1. *Adopting an adaptive control perspective, the integral action of the PI, plays the role of an estimator that "identifies" the unknown parameter vector θ.*

3 Passifiability of switched power converters

Proposition 1 is of little practical relevance for general nonlinear systems because of the need to solve the partial differential equation (2) to determine the storage function. We will investigate in this section the application of this proposition to switched power converters, which are in essence switched RLC circuits with voltage and current sources.

3.1 Model

It has been shown in [6] that a large class of power converters can be described in port–controlled Hamiltonian form as

$$\dot{x} = \left(J_0 + \sum_{i=1}^{m} J_i u_i - R \right) \frac{\partial H}{\partial x}(x) + G \tag{5}$$

where u denotes the duty ratio of the switches, $H(x) : \mathbb{R}^n \to \mathbb{R}$ is the total energy stored in inductors and capacitors, $J_i = -J_i^\top \in \mathbb{R}^{n\times n},\ i = 0, \ldots, m$ are the interconnection matrices, $R \in \mathbb{R}^{n\times n}, R = R^\top \geq 0$ is the dissipation matrix and the vector $G \in \mathbb{R}^n$ contains the external voltage and current sources. To be consistent with engineering practice, here we will take instead inductor currents and capacitor voltages, and assume that these elements are linear with total energy function

$$H(x) = \frac{1}{2} x^\top Q x, \quad Q = Q^\top > 0. \tag{6}$$

Remark 2. *The converter representation (5) highlights the role of the control action as a regulator (via circuit topology modifications) of the internal energy exchanges—an interconnection perspective that is well suited for the application of the methodology advanced in [5]. See Remark 4.*

3.2 Main result

As pointed out above the applicability of Proposition 1 is stymied by the difficulty associated with the solution of (2). Proposition 2 below shows that for power converters the PDE is obviated and, moreover, the verification of (3) reduces to a test of positivity of a constant matrix which is parameterized in terms of a positive constant and two n–dimensional vectors. It turns out that fixing this vector is necessary to satisfy $h(x_\star) = 0$, which is imposed by the detectability requirement—hence fixing this vector is quite natural.

To simplify the presentation of the proposition we find convenient to explain before how the condition $h(x_\star) = 0$ is verified. Towards this end, we note that Proposition 2 below identifies as admissible passifiable outputs the linear functions

$$h(x) = \left[J_1^\top \rho, J_2^\top \rho, \cdots, J_m^\top \rho \right]^\top Q x = D(\rho) x \tag{7}$$

where $\rho \in \mathbb{R}^n$ is a free vector. Therefore, we must select this vector ρ so that

$$h(x_\star) = \left[J_1^\top \rho, J_2^\top \rho, \cdots, J_m^\top \rho \right]^\top Q x_\star = 0. \tag{8}$$

Proposition 2. *Consider the power converter model (5) with linear inductors and capacitors and total energy function (6). Assume there exist a positive constant η_0 and a constant vector $\eta \in \mathbb{R}^m$ such that*

$$\begin{bmatrix} Q^\top R Q & w_1 + \eta_0 w_2 + w_3 \eta \\ (w_1 + \eta_0 w_2 + w_3 \eta)^\top & v_0 \eta_0 \end{bmatrix} \geq 0 \tag{9}$$

where $w_1, w_2 \in \mathbb{R}^n, w_3 \in \mathbb{R}^{n\times m}$ and $v_0 \in \mathbb{R}$ are fixed functions of the matrices J_i, R, Q and the vectors ρ, G. Then, the converter is passifiable with constant control, with the output (7) and the storage function

$$V(x) = \frac{1}{\eta_0} H(x) + \rho^\top x.$$

Remark 3. *We should underscore that the PI controller resulting from application of Propositions 1 and 2 substantially differs from standard PIs. Indeed, while the latter operate on linear error signals, $x - x_\star$, in our PI the information about the equilibrium enters through the vector ρ in (7), which (in general) will depend nonlinearly on $x_\star$.*

3.3 Extensions

The proposed passivity framework can be used to analyze the stability, and in particular to tune the parameters, of some linear control schemes. Also, an adaptive version of the PI scheme is easily obtained replacing ρ by some estimate $\hat{\rho}$ that depends on the uncertain parameters of the converter.

4 Application to a voltage source rectifier

In the remaining of the paper we will apply the technique suggested in Propositions 1 and 2 to derive a PI controller for a

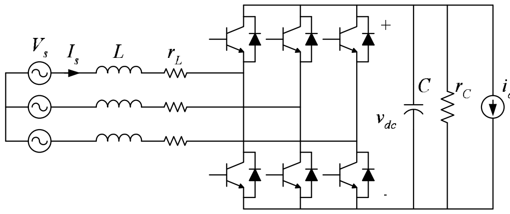


Fig. 1: Three–phase rectifier

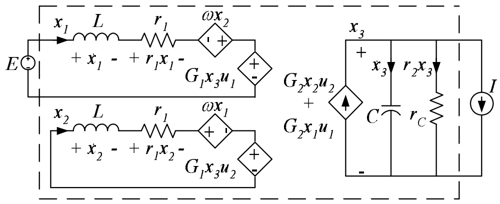


Fig. 2: Circuit representation of the scaled rectifier (1), (14)

voltage source rectifier, whose model is given in Subsection 4.1. It is clear from these propositions that the two critical steps for the success of the design are:

1. Select a vector ρ such that the candidate output function, $D(\rho)x$, verifies the detectability condition (H2).

2. Given ρ, check the viability of the LMI (9) to verify if (3) holds for all $x \in \mathbb{R}^3$. If not, take into account that the practical operation of the converter is restricted to a subset of $\mathbb{R}^3$, hence we only need to check (3) for this subset.

4.1 Model

The device is a three–phase (boost–type) rectifier feeding a load that is modelled as a pure constant current source, as schematically depicted in Fig. 1. After the transformation and a rotation to dq reference frame we obtain the model

$$\dot{\widehat{i}}_{sd} = -\frac{r_L}{L} i_{sd} + \omega i_{sq} + \frac{1}{L} v_{sd} - \frac{\gamma_{ac}}{L} m_d v_{dc} \tag{10}$$

$$\dot{\widehat{i}}_{sq} = -\omega i_{sd} - \frac{r_L}{L} i_{sq} - \frac{\gamma_{ac}}{L} m_q v_{dc} \tag{11}$$

$$\dot{v}_{dc} = \frac{\gamma_{ac}}{C} m_d i_{sd} + \frac{\gamma_{ac}}{C} m_q i_{sq} - \frac{1}{C r_c} v_{dc} - \frac{1}{C} i_{dc}, \tag{12}$$

where i_{sd}, i_{sq} are the direct and quadrature input currents, v_{dc} is the DC output voltage, m_d, m_q are the direct and quadrature modulation indices, v_{sd} is the constant direct supply voltage, i_{dc} is the constant load output current, and L, C, r_L, r_C, ω, γ_{ac} are positive model parameters.

The control objective for this device is the regulation of the capacitor voltage v_{dc} to a given constant value. Moreover, it is desirable that (in steady state) the power factor be as close as possible to one, an objective that in the present formulation is achieved driving i_{sq} to zero.

We make at this point the important observation that, due to physical considerations not captured by the model, the system is restricted to evolve in the subspace

$$\mathcal{A} = \{(i_{sd}, i_{sq}, v_{dc}) \mid i_{sd} > 0,\ v_{dc} > 0\} \subset \mathbb{R}^3. \tag{13}$$

Introducing the state vector $x = [i_{sd}, i_{sq}, v_{dc}]^\top \in \mathbb{R}^3$, the control vector $u = [m_d, m_q]^\top \in \mathbb{R}^2$, and defining the new parameters

$$r_1 = \frac{r_L}{L}, r_2 = \frac{1}{r_C C}, \gamma_1 = \frac{\gamma_{ac}}{L}, \gamma_2 = \frac{\gamma_{ac}}{C}, E = \frac{v_{sd}}{L}, I = \frac{i_{dc}}{C}$$

we can write the model in the standard form (1) with

$$f(x) = \begin{bmatrix} \omega x_2 - r_1 x_1 + E \\ -\omega x_1 - r_1 x_2 \\ -r_2 x_3 - I \end{bmatrix},\ g(x) = \begin{bmatrix} -\gamma_1 x_3 & 0 \\ 0 & -\gamma_1 x_3 \\ \gamma_2 x_1 & \gamma_2 x_2 \end{bmatrix} \tag{14}$$

A circuit representation of (1), (14) is given in Fig. 2, where we have "pulled–out" the external sources to highlight the (physical) external port variables.

Remark 4. *The system (1), (14) can be written in port–controlled Hamiltonian form (5), with the (scaled) energy function*[2]

$$H(x) = \frac{1}{2} x_1^2 + \frac{1}{2} x_2^2 + \frac{\gamma_1}{2\gamma_2} x_3^2, \tag{15}$$

the interconnection matrices

$$J_0 = \begin{bmatrix} 0 & \omega & 0 \\ -\omega & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix},\ J_1 = \begin{bmatrix} 0 & 0 & -\gamma_2 \\ 0 & 0 & 0 \\ \gamma_2 & 0 & 0 \end{bmatrix},\ J_2 = \begin{bmatrix} 0 & 0 & 0 \\ 0 & 0 & -\gamma_2 \\ 0 & \gamma_2 & 0 \end{bmatrix}$$

and the dissipation matrix and external sources vector

$$R = \begin{bmatrix} r_1 & 0 & 0 \\ 0 & r_1 & 0 \\ 0 & 0 & r_2 \frac{\gamma_2}{\gamma_1} \end{bmatrix}, \quad G = \begin{bmatrix} E \\ 0 \\ -I \end{bmatrix},$$

respectively. The power–balancing equation takes then the form

$$\dot{H} = -r_1(x_1^2 + x_2^2) - \frac{r_2 \gamma_1}{\gamma_2} x_3^2 + E x_1 - \frac{\gamma_1}{\gamma_2} I x_3. \tag{16}$$

4.2 Detectability

The admissible equilibrium set of (1), (14) is easily computed as all points such that the energy is constant. Hence, from (16), we get the set of admissible equilibria as

$$\mathcal{E} = \{\bar{x} \in \mathcal{A} \mid (r_1 \bar{x}_1 - E)\bar{x}_1 + r_1 \bar{x}_2^2 + \frac{\gamma_1}{\gamma_2}(r_2 \bar{x}_3 + I)\bar{x}_3 = 0\}.$$

Now, it is clear that a necessary condition to fulfill the detectability assumption (H2) is that

$$x_\star \in \mathcal{E} \cap \{x \in \mathcal{A} \mid h(x) = 0\}. \tag{17}$$

We have that candidate linear outputs that satisfy (17) are

$$\begin{aligned} h_1(x) &= M x_3 - x_1 \\ h_2(x) &= -x_2 \end{aligned} \tag{18}$$

[2] This function is L times the actual energy of the circuit.

where M is a positive constant chosen such that the line $h_1(x) = 0$ intersects the admissible equilibria at the desired point $x_{1\star}, x_{3\star} \in \mathbb{R}^2_+$. Since the slope of the line tangent to this circle at the origin is $\frac{E\gamma_2}{I\gamma_1}$, it is clear that M should be restricted to

$$0 < \frac{1}{M} < \frac{E\gamma_2}{I\gamma_1}. \tag{19}$$

Using the notation of (7) the outputs (18) correspond to the choice $\rho = \left[-\frac{M}{\gamma_1}, 0, -\frac{1}{\gamma_2}\right]^\top$.

Some simple calculations with the system (1), (14), (18) reveal that the zero dynamics is linear and described by the first order asymptotically stable system,

$$\dot{\xi} = -\frac{\gamma_2 M^2 r_1 + \gamma_1 r_2}{\gamma_1 + \gamma_2 M^2}\xi + \frac{\gamma_2 ME - \gamma_1 I}{\gamma_1 + \gamma_2 M^2} \tag{20}$$

with ξ describing the behavior of x_3 restricted to the set $\{x \in \mathbb{R}^3 \mid h(x) = 0\}$. This completes the verification of detectability of the outputs (18).

Remark 5. *The particular choice of the "signs" in (18) ensures that the decoupling matrix, which is given by*

$$\frac{\partial h}{\partial x} g(x) = \begin{bmatrix} \gamma_1 x_3 + M\gamma_2 x_1 & M\gamma_2 x_2 \\ 0 & \gamma_1 x_3 \end{bmatrix},$$

is not only nonsingular for all points in $\mathcal{A}$, but it is also positive definite, a property that is necessary for passivity.

4.3 Stabilizability with constant control

We will now study the inequality (3) for the rectifier model (1), (14) with storage function

$$V(x) = \alpha H(x) - \frac{M}{\gamma_1}x_1 - \frac{1}{\gamma_2}x_3, \tag{21}$$

where $H(x)$ is the total energy function defined in (15). Thus, we compute

$$\frac{\partial V}{\partial x}[f(x) + g(x)\theta] = -\alpha r_1(x_1^2 + x_2^2) - \alpha r_2 \frac{\gamma_1}{\gamma_2}x_3^2 +$$
$$\left(\frac{Mr_1}{\gamma_1} - \theta_1 + \alpha E\right)x_1 - \left(\frac{M\omega}{\gamma_1} + \theta_2\right)x_2$$
$$+\left(\frac{r_2}{\gamma_2} + M\theta_1 - \alpha\frac{\gamma_1 I}{\gamma_2}\right)x_3 + \frac{\gamma_1 I - M\gamma_2 E}{\gamma_1\gamma_2}$$

that we have to prove is non–positive. The sign–indefinite linear term in x_2 can be readily cancelled with the choice $\theta_2 = -M\omega/\gamma_1$. We prove now that, for some suitably defined $\alpha > 0$ and $\theta_1 \in \mathbb{R}$, the remaining quadratic function in x_1, x_3 is non–positive.

For, introducing the scaled variables

$$z_1 = x_1, \quad z_2 = \sqrt{\frac{\gamma_1 r_2}{\gamma_2 r_1}}x_3,$$

and the new function $\tilde{F}(z) = \frac{1}{\alpha r_1}F\left(z_1, \sqrt{\frac{\gamma_2 r_1}{\gamma_1 r_2}}z_2\right)$, to obtain,

$$\tilde{F}(z) = z_1^2 + z_2^2 - \left(\frac{M\eta_0}{\gamma_1} - \frac{\eta_1}{r_1} + \frac{E}{r_1}\right)z_1 -$$
$$\left(\frac{r_2\eta_0}{r_1\gamma_2} + \frac{M\eta_1}{r_1} - \frac{\gamma_1 I}{\gamma_2 r_1}\right)\sqrt{\frac{\gamma_2 r_1}{\gamma_1 r_2}}z_2 - \eta_0\frac{\gamma_1 I - M\gamma_2 E}{\gamma_1\gamma_2 r_1}$$

where we have denoted $\eta_0 = 1/\alpha$ and $\eta_1 = \theta_1/\alpha$. Some simple, but lengthy, calculations allows us to establish that

$$\max_{\substack{\eta_0 > 0 \\ \eta_1 \in \mathbb{R}}} \min_{z \in \mathbb{R}^2} \tilde{F}(z) = 0.$$

and $F \geq 0$ as desired. Indeed, denoting

$$z_\star = \arg\min_{z \in \mathbb{R}^2} \tilde{F}(z)$$

we have

$$z_\star = \frac{1}{r_1\gamma_1}\begin{bmatrix} \eta_0 M r_1 - \eta_1\gamma_1 + E\gamma_1 \\ (\eta_0 r_2 + \eta_1 M\gamma_2 - \gamma_1 I)\sqrt{\frac{r_1\gamma_1}{r_2\gamma_2}} \end{bmatrix},$$

Now, the maximum of $F(z_\star)$ is achieved at

$$\eta_\star = \begin{bmatrix} \eta_{\star 0} \\ \eta_{\star 1} \end{bmatrix} = \frac{\gamma_1}{M^2\gamma_2 r_1 + r_2\gamma_1}\begin{bmatrix} EM\gamma_2 - I\gamma_1 \\ IMr_1 + Er_2 \end{bmatrix}$$

which, recalling (19), verifies the restriction $\eta_{\star 0} > 0$. Finally, replacing these arguments in the original function we get $\tilde{F}(z_\star(\eta_\star)) = 0$.

This concludes the proof of passifiability, via constant control, of the rectifier model (1), (14), (18).

4.4 PI control

We are in position to present the main result of the section: a new PI controller for the rectifier system that, for all initial conditions, asymptotically drives the state towards any point in the admissible equilibrium set $\mathcal{E}$ with $x_{2\star} = 0$.

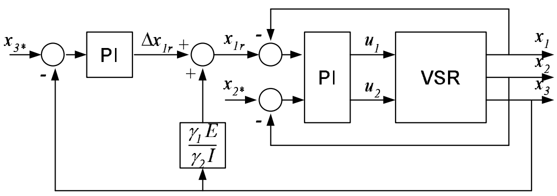
Proposition 3. *Consider the rectifier model (1), (14) and the desired equilibrium $x_\star = (x_{1\star}, 0, x_{3\star})$, where $x_{1\star} = Mx_{3\star}$, with*

$$M = \frac{1}{2r_1 x_{3\star}}\left(E - \sqrt{E^2 - 4\frac{r_1\gamma_1}{\gamma_2}x_{3\star}(x_{3\star}r_2 + I)}\right). \tag{22}$$

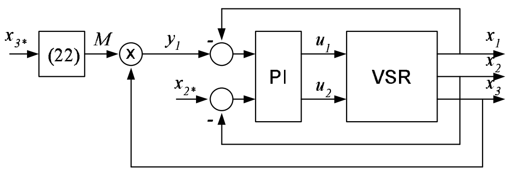
Let the control be given by

$$\begin{aligned} y &= \begin{bmatrix} Mx_3 - x_1 \\ -x_2 \end{bmatrix} \\ u &= -K_p y + z \\ \dot{z} &= -K_i y \end{aligned} \tag{23}$$

with $K_p, K_i \in \mathbb{R}^{m\times m}$ symmetric positive definite matrices. Then, for all initial conditions, the trajectories of the closed–loop system are bounded and $\lim_{t\to\infty} x(t) = x_\star$.



(a) Controller of [4].



(b) Passivity–Based PI controller.

Fig. 3: Block diagram representation of the controllers

4.5 A comparative simulation study

In this subsection we carry out a simulation study to compare the performance of the following controllers:

(A) PQ instantaneous power controller of [4];

(B) classical linearizing and decoupling controller ;

(C) proposed PI.

The following numerical values were used in the simulation $r_L = 0.01\Omega$, $r_C = 10K\Omega$, $L = 3mH$, $C = 470\mu F$, $v_{sd} = 400V$, $i_{dc} = 50A$, $\gamma_{ac} = 1$. The experiment consists of changes in the reference value $x_{3\star} = 1400 \rightarrow 1300$ and the resistance $r_1 = 3.33 \rightarrow 6.66$ at $t = 0.1, 0.3$, respectively.

(A) The first controller consists of two nested linear PIs, where the outer one generates the "reference correction" of a second PI—see Fig 3(a). The outer loop PI acts on the error $x_3 - x_{3\star}$ as

$$\dot{z}_o = K_{io}(x_{3\star} - x_3) \quad (24)$$
$$\Delta x_{1r} = K_{po}(x_{3\star} - x_3) + z_o. \quad (25)$$

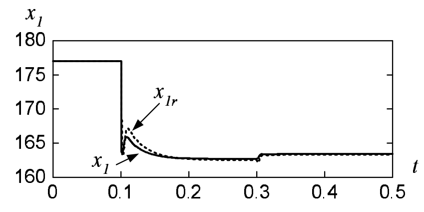
The inner loop PI is of the form

$$\dot{z}_i = K_{ii}\left(\begin{bmatrix} x_{1r} \\ 0 \end{bmatrix} - \begin{bmatrix} x_1 \\ x_2 \end{bmatrix}\right) + \begin{bmatrix} 0 & \omega \\ -\omega & 0 \end{bmatrix} z_i \quad (26)$$
$$u = K_{pi}\left(\begin{bmatrix} x_{1r} \\ 0 \end{bmatrix} - \begin{bmatrix} x_1 \\ x_2 \end{bmatrix}\right) + z_i. \quad (27)$$

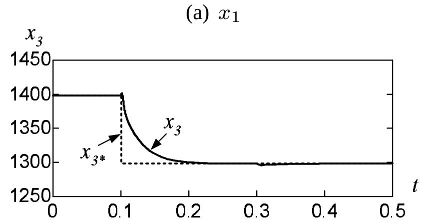
The reference for the current, x_{1r}, is computed as

$$x_{1r} = \frac{\gamma_1 I}{\gamma_2 E} x_3 + \Delta x_{1r}$$

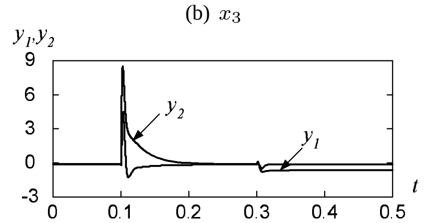
The simulations were carried out with $K_{po} = 0.4$, $K_{io} = 20$, $K_{pi} = 0.001I$, $K_{ii} = 0.1I$, and the results are depicted in Fig. 4. We can observe that the dinamics of x_3 is very good for changes in the reference and changes in the parameters converging always to the established value $x_{3\star}$.



(a) x_1



(b) x_3



(c) y_1, y_2

Fig. 4: Simulation results for controller (A).

(B) Given that the rectifier is minimum phase and has a well–defined relative degree—with respect to the outputs (18)—it is luring to try a linearizing and decoupling controller, which in this case yields the (obviously complicated) static state feedback

$$u_1 = \frac{1}{\gamma_1 x_3 + M\gamma_2 x_1}\left[(K_1 - r_1)x_1 + \omega x_2 - M(K_1 - r_2)x_3 + K_1 v_1 + E + MI\right] + \frac{M\gamma_2 x_2(\omega x_1 + r_1 x_2)}{(\gamma_1 x_3 + M\gamma_2 x_1)\gamma_1 x_3} \quad (28)$$
$$u_2 = \frac{1}{\gamma_1 x_3}\left[(K_2 - r_1)x_2 + K_2 v_2 - \omega x_1\right],$$

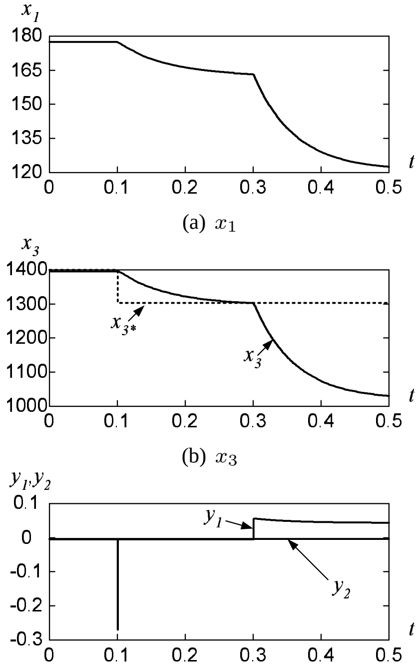
where we have added the coefficients $K_1, K_2 > 0$ to obtain in the closed–loop two decoupled first order linear systems of the form $\dot{y}_i = -K_i(y_i + v_i),\ i = 1, 2$, and we have chosen $K_1 = K_2 = 10000$. The behavior, shown in Fig. 5, exhibit excellent regulation of the outputs y, but a slow response in x_1 and x_3. We also observe the large steady state error that appears with the change of parameter in the output y and in the final value of x_1 and x_3.

(C) The controller proposed in this paper is the linear PI (4), see Fig. 3(b). For simplicity, we have taken diagonal proportional and integral gain matrices, leading to the transfer matrix representation

$$\begin{bmatrix} u_1 \\ u_2 \end{bmatrix} = \begin{bmatrix} K_{p1} + K_{i1}\frac{1}{s} & 0 \\ 0 & K_{p2} + K_{i2}\frac{1}{s} \end{bmatrix} \begin{bmatrix} x_1 - Mx_3 \\ x_2 \end{bmatrix}$$

We have taken $K_{p1} = 0.01$, $K_{i1} = 10$, $K_{p2} = 0.01$, $K_{i2} = 0.1$, and introduced an estimator for r_1 that takes the form

$$\dot{\xi} = -\omega x_2 + E - \gamma_1 x_3 u_1 - x_1 \hat{r}_1 - \Lambda(\xi - x_1) \quad (29)$$
$$\dot{\hat{r}}_1 = \Gamma x_1(\xi - x_1) \quad (30)$$

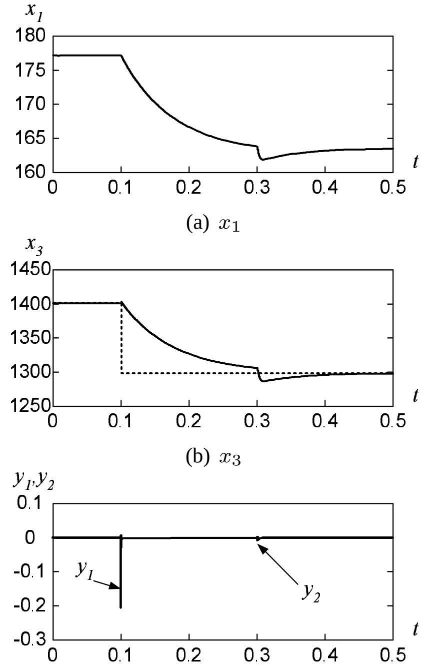


(a) x_1

(b) x_3

(c) y_1, y_2

Fig. 5: Simulation results for controller (B).



(a) x_1

(b) x_3

(c) y_1, y_2

Fig. 6: Simulation results for controller (C).

The transient performance of the adaptive scheme is shown in Fig. 6, where we observe the significant improvement in the converging values of x_1 and x_3. Therefore it is possible to obtain a general performance very similar to the PQ controller adding the quadrature current control capability.

5 Conclusions and outlook

PBCs have been designed for power converters for several years now, the conventional wisdom when it comes to practical applications is that an integral action is indispensable for adequate operation of these scheme. This feature can be naturally incorporated in PBCs in the form of adaptation schemes–but this considerably complicates the control laws.

Motivated by this concern we have explored in this paper the possibility of directly designing PI controllers using passivity principles. Towards this end, it turned out to be necessary to view power converters from a radically new perspective, with the switches appearing as port variables. It may be argued that this viewpoint is rather unnatural because, in general, the switch positions and the outputs are not conjugate variables—that is, their product does not have units of power. In spite of that, it turns out that this perspective yields some useful easily interpretable and implementable results that permits to establish some connections with existing schemes. In this paper detailed calculations are given only for a three–phase rectifier. The proposed methodology is also applicable to other switched power converters, and the results will be reported elsewhere.

Besides this stabilization–oriented line of research, we are currently investigating problems of reactive power compensation and impedance matching from a nonlinear viewpoint.

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