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IDA-PBC CONTROLLER FOR DC-DC POWER ELECTRONIC CONVERTERS WITH NONLINEAR LOAD ENHANCED WITH ADDITIVE DISTURBANCE ESTIMATION

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ABSTRACT

This work presents the design of stabilising controllers for the DC-DC boost and buck-boost power electronic converters using a passivity-based approach. The first step in the controller design is the definition of a convenient transformation of the state vector. The first variable of the transformation is the flat output of the converter, and the second is a bijective function of the charge of the output capacitor. This alternative to a previous work by the authors, which also considers flat outputs for the state vector transformation, ensures the bijectivity of the complete transformation. The disadvantage is that the designer is not allowed to choose a closed loop energy function, thus having to solve a partial differential equation to find one. A nonlinear state feedback control law is finally obtained. Disturbance rejection is addressed using a dynamic estimator of the load current, using a technique from the literature. The controller performance is validated via digital simulation.

Keywords: DC-DC power electronic converters, passivity-based control, port-Hamiltonian systems, flatness-based control, Bond Graphs.

1. INTRODUCTION

Due to their versatility, high efficiency, controllable behaviour, fast dynamics and wide-range of power management, Power Electronic Converters (PEC) are ubiquitous and pervade most of the cutting-edge engineering application areas. Indeed, they can be found in electrical drives, switched-mode power supplies, battery chargers, uninterrupted power supplies, all type of mobile devices, distributed generation and renewable energy conversion systems, embedded in electric/hybrid vehicles (cars, trains and airplanes), etc. Closed loop control design of PEC is a key topic, as for high performance applications not only asymptotic stability must be assured but performance too. The challenge in the coming years lies in developing new techniques at the lowest possible cost, size and weight for emerging applications (Ojo, 2019). This motivates a new approach of solving control problems in PEC feeding nonlinear loads.

PEC are highly nonlinear dynamical systems whose dynamics can be represented by means of averaged models, where the control input is the duty cycle of the PWM controlled electronic switch. For the second order boost and buck-boost converters models, the duty cycle has relative degree one with respect to both system states, making the controller synthesis a difficult task. This obstacle can be overcome through a conveniently designed coordinate transformation using the flat variable of the converter and choosing the other variable as a bijective function of the output capacitor charge to ensure output voltage regulation.

The rationale of the controller design relies on finding a state feedback control law in the new coordinates using interconnection and damping assignment passivity-based control (IDA-PBC), such that the closed loop can be written in port-Hamiltonian (pH) form. This allows using the closed loop Hamiltonian as a Lyapunov candidate function to analyze stability. As the coordinate transformation is bijective, asymptotic stability of the output voltage is achieved via regulation of the equivalent equilibrium of the transformed coordinates.

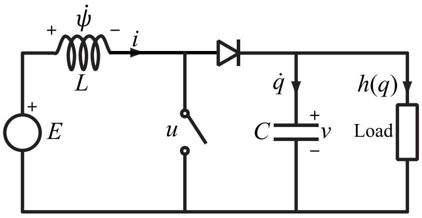
The design is enhanced using a disturbance dynamic estimator (He, et al., 2018) allowing the system to reject constant additive disturbances on the load side.

The remainder of this article is organized as follows: Section 2 presents the averaged models of the PEC and the control problem formulation. Section 3 introduces the general concepts of IDA-PBC and presents the controller design for the PEC. Disturbance estimators are developed in Section 4. Section 5 validates the controller designs via digital simulation, and conclusions are given in Section 6.

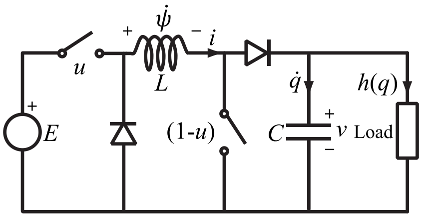
2. AVERAGED MODELS OF PEC

In this section the averaged models of the Boost and the Buck-Boost converters are presented. Figure 1 shows the equivalent circuits of the PEC under study. The inductance and the capacitor are assumed linear components with known parameters L and C. The averaged duty cycle of the electronic switch is represented by the continuous control signal $u \in (0,1)$. It is assumed that the PEC operate in the continuous conduction mode (CCM), and the state variables magnetic flux and electric charge $(\psi, q) > (0,0)\ \forall\, t$. Further, $h(q)$ is the current absorbed by a truly dissipative nonlinear load satisfying

$\{h(q) > 0 \,\forall\, q > 0\}$. The load and capacitor terminal voltages being the same allows to write the load current in terms of the capacitor charge. The reader is referred to (Mohan, Undeland, & Robbins, 1995) for further details on PEC averaged models.



(a)



(b)

Figure 1: Equivalent Circuits of (a) the boost converter, and (b) the buck-boost converter.

2.1. Boost Converter

The averaged state equations of the boost converter are given in (1):

$$\begin{aligned} \dot{\psi} &= E - u\frac{q}{C} \\ \dot{q} &= u\frac{\psi}{L} - h(q) \end{aligned} \quad ; \text{ with } \{u \in \mathbb{R} \mid u \in (0,1)\} \tag{1}$$

2.2. Buck-Boost Converter

The averaged state equations of the boost converter are given in (2):

$$\begin{aligned} \dot{\psi} &= uE - (1-u)\frac{q}{C} \\ \dot{q} &= (1-u)\frac{\psi}{L} - h(q) \end{aligned} \quad ; \text{ with } \{u \in \mathbb{R} \mid u \in (0,1)\} \tag{2}$$

2.3. Control Problem Formulation

The control problem for the converters consists in finding a map ξ such that the state feedback controller:

$$u = \xi(\psi, q) \tag{3}$$

stabilizes the output voltage to a desired set-point $v^\star$ whilst ensuring internal stability. Since $v = q/C$, then output voltage regulation is equivalent to regulation of capacitor charge to the set point $q^\star = Cv^\star$. Internal stability is ensured if the equilibrium $(\psi^\star, q^\star)$ is stable.

3. PASSIVITY-BASED CONTROL

In this section, the general ideas of passivity-based control (PBC) for pH systems are first summarized and then the main result of the article is presented. Prior to the controller design, the coordinate transformation for each PEC is introduced. This will end up in a very similar matching equation for both converters reducing the effort needed to solve it.

3.1. General Concepts of IDA-PBC

Consider a dynamical system

$$\dot{x} = f(x) + g(x)u \tag{4}$$

with $x \in \mathbb{R}^n$ and $u \in \mathbb{R}^m$, the problem of stabilizing (4) using IDA-PBC consist on finding a mapping $\xi: \mathbb{R}^n \to \mathbb{R}^m$ such that the system (4) in closed loop with the controller $u = \xi(x)$ can be written in pH form as follows:

$$\dot{x} = [\boldsymbol{J}(x) - \boldsymbol{R}(x)]\frac{\partial H(x)}{\partial x} \tag{5}$$

where the matrices $\boldsymbol{J} = -\boldsymbol{J}^T$ and $\boldsymbol{R} = \boldsymbol{R}^T \geq 0$ describe the interconnection and dissipation structure, and the function $H: \mathbb{R}^n \to \mathbb{R}$ is the Hamiltonian representing the total energy stored in the system. Let $x^\star$ be the minimizer of the Hamiltonian: $x^\star = \arg\min\{H(x)\}$, then $x^\star$ is a stable equilibrium point of the system (5). Moreover, under some detectability conditions, the equilibrium is asymptotically stable. If the following matching equation has a solution:

$$g(x)^\perp f(x) = g(x)^\perp[\boldsymbol{J}(x) - \boldsymbol{R}(x)]\frac{\partial H(x)}{\partial x} \tag{6}$$

with $g(x)^\perp$ is the full-rank left annihilator of $g(x)$, then the control law can be synthesized as follows

$$u = [g^T g]^{-1} g^T \left[f - (\boldsymbol{J} - \boldsymbol{R})\frac{\partial H(x)}{\partial x}\right] \tag{7}$$

See (Ortega & García-Canseco, 2004) for further details on IDA-PBC method.

3.2. Rationale of the Design Method

In the sequel, prior to the controller design a coordinate transformation for each converter will be introduced to ease the effort needed for the controller synthesis. The main methodological contribution of this paper consist in providing:

- A bijective state transformation $(\psi, q) \to (y, z)$ ensuring that driving $(y, z) \to (y^\star, z^\star)$ is equivalent to drive $(\psi, q) \to (\psi^\star, q^\star)$

- A matching equation with a similar structure for both converters, which can be easily solved by integration

- Finding a similar control law for both converters under the proposed coordinate transformation.

3.3. Boost Converter

3.3.1. Coordinate transformation

Consider the following coordinate transformation:

$$\begin{aligned} y &= \frac{\psi^2}{2L} + \frac{q^2}{2C} \\ z &= \frac{q^2}{2C} \end{aligned} \tag{8}$$

The original variables, are related to the new ones as follows:

$$\begin{aligned} \psi &= \sqrt{2L}\,(y-z)^{\frac{1}{2}} \\ q &= \sqrt{2C}\, z^{\frac{1}{2}} \end{aligned} \tag{9}$$

Time derivation of y and z leads to the following dynamics:

$$\begin{aligned} \dot{y} &= E\frac{\psi}{L} - \frac{q}{C}\, h(q) \\ \dot{z} &= \frac{q}{C}u\frac{\psi}{L} - \frac{q}{C}h(q) \end{aligned} \tag{10}$$

A new control input is defined as $m = u\frac{q}{C}\frac{\psi}{L}$. In terms of the new coordinates and the load power $P(q) = \frac{q}{C}h(q)$, the boost converter state space model can be written in the form $[\dot{y} \quad \dot{z}]^\top = f(y,z) + g\, m$:

$$\begin{bmatrix} \dot{y} \\ \dot{z} \end{bmatrix} = \begin{bmatrix} E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - P_z(z) \\ -P_z(z) \end{bmatrix} + \begin{bmatrix} 0 \\ 1 \end{bmatrix} m \tag{11}$$

where $P_z(z) = P\big(q(z)\big)$ stands for the load power, in terms of the coordinate z. As expected, the new states equations are power balances. Recall that y stands for the total energy stored in the system and z for the energy stored in the capacitor. As the averaged model is valid for $(\psi, q) > (0,0)$, then $(y-z) > 0, \forall\, (y,z)$.

3.3.2. Controller Design

We present here the main result for the boost converter, under the assumption of a known load VA (Volt-Ampère) characteristic.

Proposition 1: Consider the following controller for system (11):

$$\begin{aligned} m = {} & E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - P_z(z^\star) + K_y(y-y^\star) + \\ & P_z(z) + r\left(E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - P_z(z)\right) \end{aligned} \tag{12}$$

with $K_y > 0$ and $r > 0$. $z^\star = {q^\star}^2/(2C)$ is the desired equilibrium value for the capacitor energy, and $y^\star = \frac{{\psi^\star}^2}{2L} + \frac{{q^\star}^2}{2C}$ is the equilibrium for the total stored energy in the boost converter. It is assumed that $\frac{\partial P_z(z)}{\partial z} \geq 0\ \forall\, z > 0$. System (11) in closed loop with the controller given in (12) has the following properties:

P1. The closed loop dynamics can be written in pH form as given in (13) with the Hamiltonian in (14):

$$\begin{bmatrix} \dot{y} \\ \dot{z} \end{bmatrix} = \begin{bmatrix} 0 & -1 \\ 1 & -r \end{bmatrix} \cdot \begin{bmatrix} \frac{\partial H(y,z)}{\partial y} \\ \frac{\partial H(y,z)}{\partial z} \end{bmatrix} \tag{13}$$

$$\begin{aligned} H(y,z) = {} & E\frac{2}{3}\sqrt{\frac{2}{L}}(y-z)^{\frac{3}{2}} + \int_0^z P_z(\alpha)d\alpha + \\ & \frac{K_y}{2}\left(y - \frac{P_z(z^\star)}{K_y} - y^\star\right)^2 \end{aligned} \tag{14}$$

$H(y,z)$ is a positive definite function, as $P\big(q(z)\big) = P_z(z)$ is also a positive definite function $\forall\, z > 0$.

From (13), the closed loop interconnection and dissipation matrices are the following constant matrices:

$$\boldsymbol{J} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix} \quad \text{and} \quad \boldsymbol{R} = \begin{bmatrix} 0 & 0 \\ 0 & r \end{bmatrix}$$

P2. The controller ensures the asymptotic stability of the equilibrium $(y^\star, z^\star)$.

Proof: The claim in P1 is easily proved taking into account that:

$$\begin{bmatrix} \frac{\partial H(y,z)}{\partial y} \\ \frac{\partial H(y,z)}{\partial z} \end{bmatrix} = \begin{bmatrix} E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - K_y(y-y^\star) - P_z(z^\star) \\ -E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} + P_z(z) \end{bmatrix} \tag{15}$$

and matching $\dot{y}$ given by (11) and (13). The next step is to use (12) in (11) to obtain:

$$\begin{aligned} \dot{z} = {} & E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - P_z(z^\star) + K_y(y-y^\star) + \\ & +r\left(E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - P_z(z)\right) \end{aligned} \tag{16}$$

$$\dot{z} = \frac{\partial H(y,z)}{\partial y} - r\,\frac{\partial H(y,z)}{\partial z} \tag{17}$$

completing the proof of the claim in P1. Asymptotic stability will be proved using the closed loop Hamiltonian (14) as a Lyapunov candidate function. First recall that $\sqrt{(2/L)}\,(y-z)^{\frac{1}{2}} = E\,(\psi/L)$:

$$E\sqrt{\frac{2}{L}}\,(y^\star - z^\star)^{\frac{1}{2}} = E\frac{\psi^\star}{L} = P(q^\star) = P_z(z^\star) \qquad (18)$$

Then, we conclude:

$$\begin{bmatrix} \frac{\partial H(y,z)}{\partial y} \\ \frac{\partial H(y,z)}{\partial z} \end{bmatrix}_{(y^\star,z^\star)} = \begin{bmatrix} 0 \\ 0 \end{bmatrix} \qquad (19)$$

Integrability condition is fulfilled as:

$$\frac{\partial}{\partial z}\left(\frac{\partial H(y,z)}{\partial y}\right) = \frac{\partial}{\partial y}\left(\frac{\partial H(y,z)}{\partial z}\right) = -E\frac{1}{2}\sqrt{\frac{2}{L}}\,(y-z)^{\frac{-1}{2}} \qquad (20)$$

Finally, it must be ensured that $(y^\star, z^\star)$ is a minimum of $H(y,z)$:

$$\begin{aligned} \frac{\partial}{\partial y}\left(\frac{\partial H(y,z)}{\partial y}\right)\bigg|_{(y^\star,z^\star)} &= \frac{E}{2}\sqrt{\frac{2}{L}}\,(y^\star - z^\star)^{-\frac{1}{2}} + K_y y^\star \\ \frac{\partial}{\partial y}\left(\frac{\partial H(y,z)}{\partial y}\right)\bigg|_{(y^\star,z^\star)} &> 0 \end{aligned} \qquad (21)$$

and the determinant of the Hessian of $H(y,z)$ evaluated in the equilibrium:

$$\nabla^2 H(y,z)|_{(y^\star,z^\star)} = \frac{E}{2}\sqrt{\frac{2}{L}}(y^\star - z^\star)^{-\frac{1}{2}}\left(K_y\, y^\star + \frac{\partial P_z(z)}{\partial z}\bigg|_{z^\star}\right) + \left(K_y\, y^\star \cdot \frac{\partial P_z(z)}{\partial z}\bigg|_{z^\star}\right) \qquad (22)$$

under the assumption that $\frac{\partial P(q)}{\partial q} \geq 0$ $\left(\Rightarrow \frac{\partial P_z(z)}{\partial z} \geq 0\ ;\ \forall\, q \geq 0\right)$ and recall $K_y > 0$ then $\nabla^2 H(y,z)|_{(y^\star,z^\star)} > 0$. This completes the proof. ■

Remark 1: $P_z(z) = P(q(z))$ and $\frac{\partial P_z(z)}{\partial z} = \frac{\partial P(q(z))}{\partial q}\frac{\partial q(z)}{\partial z}$. $\frac{\partial q(z)}{\partial z} > 0$, this is given by the coordinate transformation. Then, if $\frac{\partial P(q)}{\partial q} \geq 0 \Rightarrow \frac{\partial P_z(z)}{\partial z} \geq 0.$

Remark 2: Equilibrium $(\psi, q) = (\psi^\star, q^\star)$ is achieved through the equilibrium of the energy variables $(y,z) = (y^\star, z^\star)$.

3.4. Buck-Boost Converter

3.4.1. Coordinate transformation

Recall the dynamics of this converter given by system (2). The following convenient coordinate transformation is proposed:

$$\begin{aligned} y &= \frac{\psi^2}{2L} + \frac{q^2}{2C} + Eq \\ z &= \frac{q^2}{2C} + Eq \end{aligned} \qquad (23)$$

The original variables then, can be written in terms of y and z as:

$$\begin{aligned} \psi &= \sqrt{2L}\,(y-z)^{\frac{1}{2}} \\ q &= C\left(\left(E^2 + \frac{2}{C}z\right)^{\frac{1}{2}} - E\right) \end{aligned} \qquad (24)$$

Time derivation of y and z leads us to the following dynamics:

$$\begin{aligned} \dot{y} &= E\frac{\psi}{L} - \left(\frac{q}{C} + E\right) h(q) \\ \dot{z} &= \left(E + \frac{q}{C}\right)\frac{\psi}{L}(1-u) - \left(E + \frac{q}{C}\right) h(q) \end{aligned} \qquad (25)$$

For this case the new control input defined is $m = (1-u)\left(\frac{q}{C} + E\right)\frac{\psi}{L}$, and the current absorbed by the load in terms of z is written as $h(q(z)) = h_z(z)$. It is convenient to define the following quantity $P_z(z) = h_z(z)\left(E^2 + \frac{2}{C}z\right)^{\frac{1}{2}}$, which has units of power. Then, the state space dynamics of the buck boost converter can be written in the form $[\dot{y} \quad \dot{z}]^\top = f(y,z) + g\, m$:

$$\begin{bmatrix} \dot{y} \\ \dot{z} \end{bmatrix} = \begin{bmatrix} E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - P_z(z) \\ -P_z(z) \end{bmatrix} + \begin{bmatrix} 0 \\ 1 \end{bmatrix} m \qquad (26)$$

As expected, once again, the new state equations are power balances. Recall that y stands for the total energy stored in the converter plus the term Eq, and the variable z is the output capacitor energy plus the term Eq. The averaged model is valid for $(\psi, q) > (0,0)$ implying that $y > z > 0$, then $(y - z) > 0,\ \forall\,(y,z)$. Recalling $h_z(q) > 0$ then $P_z(z) > 0$.

3.4.2. Controller Design

The controller is designed for system (26), assuming a known load VA characteristic.

Proposition 2: Consider the following controller for system (26):

$$m = E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - P_z(z^\star) + K_y(y - y^\star) + P_z(z) + r\left(E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - P_z(z)\right) \qquad (27)$$

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with $K_y > 0$ and $r > 0$. $z^\star = \frac{q^{\star 2}}{2C} + Eq^\star$ and $y^\star = \frac{\psi^{\star 2}}{2L} + \frac{q^{\star 2}}{2C} + Eq^\star$. It is assumed that $\frac{\partial P_z(z)}{\partial z} \geq 0 \; \forall \, z > 0$.

Remark 3: Properties P1 and P2 are guaranteed by the control law (27), since the coordinate transformation defined by (23) has the same structure as (8) defined for the boost converter. Hence the proof will be omitted.

4. DISTURBANCE ESTIMATION

A constant disturbance estimation method is presented in this section. Constant current disturbances in the output stage of the two converters under study are considered. The method is based on the design for unknown constant power load in (He, et al., 2018) and is an Immersion & Invariance design.

4.1. Boost Converter

Consider the state equation of the capacitor charge:

$$\dot{q} = u\frac{\psi}{L} - h_o(q) - \bar{\imath} \tag{28}$$

where $\bar{\imath}$ is the constant unknown current disturbance and $h_o(q)$ is the known original nonlinear load. The estimated current is defined as:

$$\hat{\imath} = \bar{\imath} + \tilde{\imath} \tag{29}$$

hence $\tilde{\imath}$ is the estimation error that must be driven to zero. To do so, consider the following proposition:

Proposition 3: Dynamic estimator for $\bar{\imath}$:

$$\begin{aligned} \hat{\imath} &= -\tfrac{1}{2}k_q\, q + k_q\alpha \\ \dot{\alpha} &= \tfrac{1}{2}\left(u\tfrac{\psi}{L} - h_o(q) - \hat{\imath}\right) \end{aligned} \tag{30}$$

where $\hat{\imath}$ converges to $\bar{\imath}$, with $k_q > 0$.

Proof: Computing the time derivative of $\hat{\imath}$ and then using the expression given for $\dot{\alpha}$ yields:

$$\begin{aligned} \dot{\hat{\imath}} &= -\tfrac{1}{2}k_q\, \dot{q} + k_q\dot{\alpha} \\ \dot{\hat{\imath}} &= -\tfrac{1}{2}k_q\left(u\tfrac{\psi}{L} - h_o(q) - \bar{\imath}\right) + k_q\tfrac{1}{2}\left(u\tfrac{\psi}{L} - h_o(q) - \hat{\imath}\right) = \tfrac{1}{2}k_q\bar{\imath} - \tfrac{1}{2}k_q\hat{\imath} \\ \dot{\hat{\imath}} &= -\tfrac{1}{2}k_q\tilde{\imath} \end{aligned} \tag{31}$$

As $\bar{\imath}$ is constant $\Rightarrow \dot{\bar{\imath}} = 0 \Rightarrow \dot{\hat{\imath}} = \dot{\tilde{\imath}}$. Finally, $\dot{\tilde{\imath}} = -\frac{1}{2}k_q\tilde{\imath}$, completing the proof. ■

The gain k_q will then be used to tune the time response of the estimator.

4.2. Buck-Boost Converter

Consider a constant unknown disturbance on the output stage of this converter. Then, the capacitor charge dynamics is:

$$\dot{q} = (1-u)\frac{\psi}{L} - h_o(q) - \bar{\imath} \tag{32}$$

where $\bar{\imath}$ is the constant unknown current disturbance and $h_o(q)$ is the known original nonlinear load.

Proposition 4: Dynamic estimator for $\bar{\imath}$:

$$\begin{aligned} \hat{\imath} &= -\tfrac{1}{2}k_q\, q + k_q\alpha \\ \dot{\alpha} &= \tfrac{1}{2}\left((1-u)\tfrac{\psi}{L} - h_o(q) - \hat{\imath}\right) \end{aligned} \tag{33}$$

with $k_q > 0$.

Remark 4: The convergence proof of the estimator is omitted as it follows the same procedure used for the boost converter estimator in the previous subsection.

4.3. Controller Enhancement

This section is intended to introduce the disturbance estimation into the control law previously designed for the PEC under the assumption of a perfectly known load VA characteristic and perform an enhancement that preserves the desired equilibrium of the output voltage. The approach the authors applied in (Tomassini, Donaire, Junco, & Pérez, 2017) for the Buck converter case is not applicable now as the equilibrium of the variable y is unknown in presence of a disturbance and both gradients of the Hamiltonian $H(y,z)$ depend on both y and z variables. This motivates the estimator design development we propose in this section. Taking into account the constant disturbance, the total current of the output stage of the converter is given by:

$$h(q(z)) = h_o(q(z)) + \bar{\imath} \tag{34}$$

where $\bar{\imath}$ is a constant unknown current, so consider the following load current estimation:

$$\hat{h}(q(z)) = h_o(q(z)) + \hat{\imath} = h_o(q(z)) + \bar{\imath} + \tilde{\imath} \tag{35}$$

Recall for both Boost and Buck-Boost converters, the term $P_z(z) = P\big(q(z)\big)$ needs the value of $\bar{\imath}$. The estimation of P_z:

$$\hat{P}_z(z) = P_z(z) + \tilde{P}_z(z) \tag{36}$$

with, for the Boost converter:

$$\hat{P}_z(z) = \frac{q(z)}{C}\, h(q(z)) + \frac{q(z)}{C}\tilde{\imath} \tag{37}$$

and for the Buck-Boost converter:

$$\hat{P}_z(z) = \left(E + \frac{q(z)}{C}\right) h(q(z)) + \left(E + \frac{q(z)}{C}\right) \tilde{\imath} \qquad (38)$$

Using the power estimated functions, the control laws (12) and (27):

$$\begin{aligned} \hat{m} &= E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - \hat{P}_z(z^\star) + K_y(y-y^\star) + \\ &\hat{P}_z(z) + r\left(E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - \hat{P}_z(z)\right) \end{aligned} \qquad (39)$$

$$\begin{aligned} \hat{m} &= E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - P_z(z^\star) + K_y(y-y^\star) + \\ &P_z(z) + r\left(E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - P_z(z)\right) - \\ &\tilde{P}_z(z^\star) + \tilde{P}_z(z) - r\tilde{P}_z(z) = m + \tilde{m} \end{aligned} \qquad (40)$$

where the $\widetilde{(\cdot)}$ terms vanish as $\tilde{\imath} \to 0$. The closed loop dynamics in the (y, z) variables are:

$$\begin{aligned} \begin{bmatrix} \dot{y} \\ \dot{z} \end{bmatrix} &= \begin{bmatrix} E\sqrt{\frac{2}{L}}\,(y-z)^{\frac{1}{2}} - P_z(z) \\ -P_z(z) \end{bmatrix} + \begin{bmatrix} 0 \\ 1 \end{bmatrix} m + \\ &+ \begin{bmatrix} 0 \\ 1 \end{bmatrix} \tilde{m}(\tilde{\imath}) \end{aligned} \qquad (41)$$

$$\dot{\tilde{\imath}} = -\frac{1}{2} k_q \tilde{\imath} \qquad (42)$$

The dynamics given by (41) and (42) define a cascaded system. Proposition 4.1 of (Sepulchre, Jankovic, & Kokotovic, 2012) ensures asymptotic stability of system (41), (42) as the origin of the $\tilde{\imath}$-subsystem is asymptotically stable.

5. SIMULATION RESULTS

In this section the controller performance is tested via digital simulation on a Buck-Boost converter system. The model parameters are $L = 16mH$, $C = 1.2mF$ and $E = 50V$. The load VA characteristic is defined in terms of the capacitor charge in (43) and graphically shown in VA characteristic in Figure 2.

$$h(q) = \frac{q}{C}\frac{1}{51} - \left(\frac{q}{C}\frac{1}{51}\right)^3 + \left(\frac{q}{C}\frac{1}{68}\right)^5 + \operatorname{atan}\left(\frac{2}{3}\frac{q}{C}\right) \qquad (43)$$

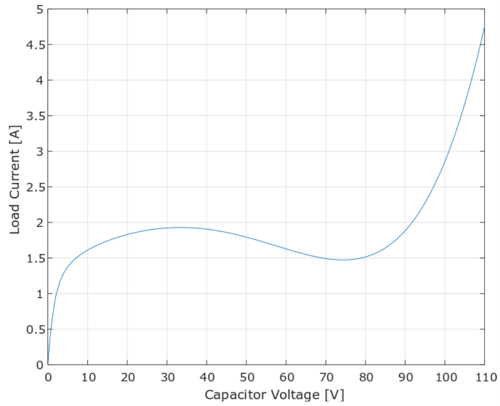
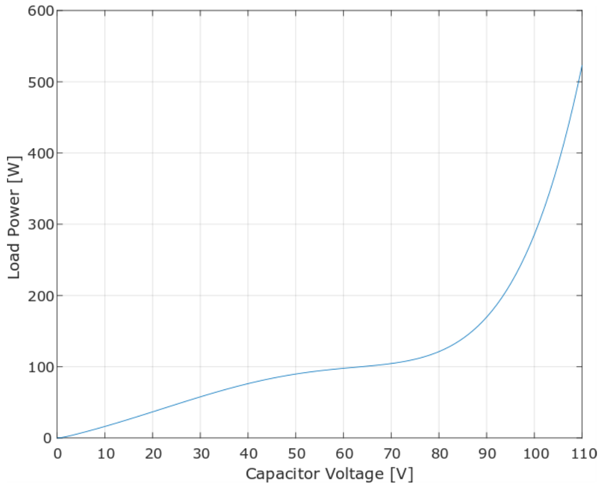
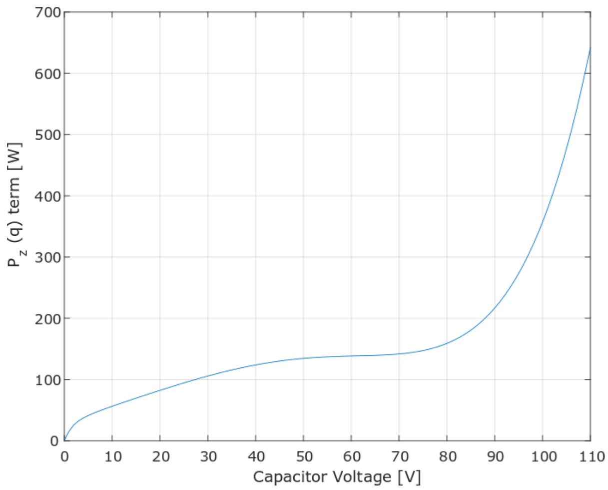


Figure 2: Load current, in terms of the output capacitor voltage.

where the capacitor voltage is $v_c = q/C$. The figure below shows the Load Power and the quantity $h(q)\left(E + \frac{q}{C}\right)$ where the last stands for the term $P_z(z)$ in terms of the capacitor voltage:



(a)



(b)

Figure 3: Load Power (a), and term $P_z(q)$ (b) in terms of the capacitor voltage.

The controller tuning parameters are $k_y = 100$ and $r = 12$. In the first simulation set no disturbance is considered while in the latter a constant current is considered and consequently rejected using a dynamic estimator. The controller can incorporate (43) as a “lookup table”, with the necessary amount of breakpoints.

Remark 5: As for the controller tuning, considering that the closed loop is nonlinear, a set of parameters for K_y and r was obtained using linearization of the closed loop around the equilibrium $\frac{q^\star}{C} = 50V$, for an acceptable response time. Finally, noting that $\dot{H}(y,z) = -r\left(\frac{\partial H(y,z)}{\partial z}\right)^2$ it is concluded that the value of the gain r has a direct impact in the response time. The performance will be tested for different values of r in the next subsection.

5.1. Controller performance test

In this simulation experiment the controller performance is tested via an output voltage reference change. The initial conditions are set corresponding to $\frac{q}{C} = 50V$. The output voltage reference (dashed line) is changed to $35V$ at $t = 0.1s$, then changed to $60V$ at $t = 0.8s$ and finally to $85V$ at $t = 1.4s$. Notice that the two first reference values correspond to incremental negative resistance zone of the load VA characteristic of the load. Different response times are obtained by tuning the parameter r starting with $r = 12$ and reducing it.

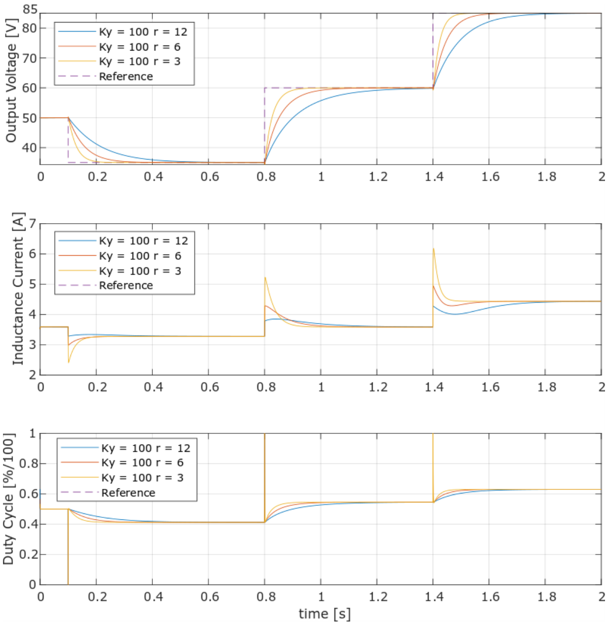


Figure 4: Time plots of the Buck-Boost converter original variables for different values of r.

And the time evolution of the transformed variables:

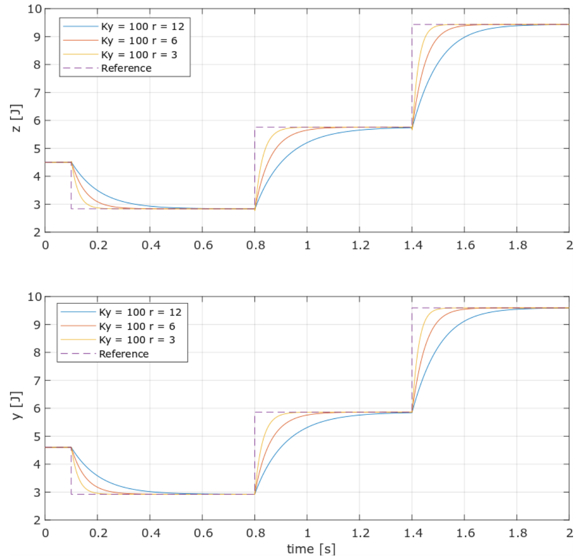


Figure 5: Time plots of the transformed variables for different values of r.

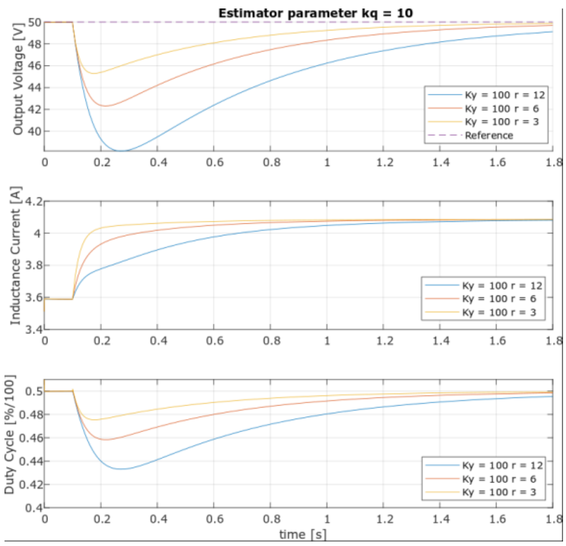
The controller achieves the control objective through the regulation of the variables y and z, for output reference voltages greater and lower than E; and including the negative incremental resistance zones of the load VA characteristic.

5.2. Simulations with disturbance estimator

In this subsection the controller performance is tested with the disturbance estimator implemented. The first set of simulations shows only the disturbance rejection behaviour while the second set shows how the estimator degrades the performance of the controller presented in Subsection 5.1.

5.2.1. Disturbance rejection behaviour

Consider a disturbance of $\bar{\iota} = 0.25A$ applied at $t = 0.1s$ when the system is in steady state and regulating an output voltage of $50V$:



(a)

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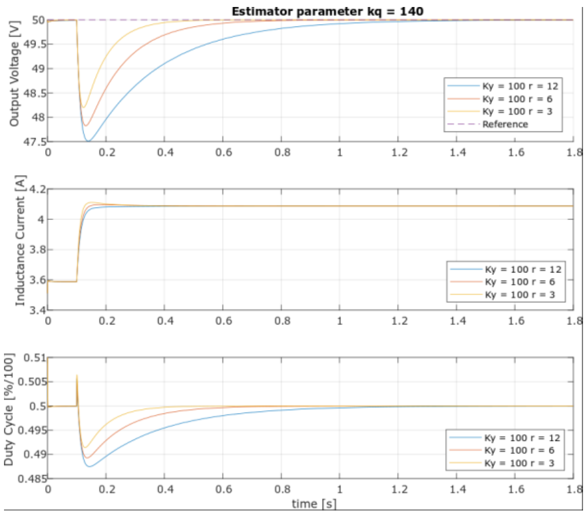

(b)

Figure 6: Disturbance rejection simulation results for estimator parameter $k_q = 10$ (a) and $k_q = 140$ (b).

The disturbance is satisfactorily rejected thanks to the estimator. Note that the response time of the system varies not only by means of k_q but also by r. This is because the estimator dynamics depends on the states (ψ, q) and on the duty cycle u.

5.2.2. Controller performance degradation

For this simulation set, consider the system in equilibrium corresponding to $\frac{q}{C} = 50V$ in presence of a disturbance of $\bar{\imath} = 0.25A$. Estimator parameter $k_q = 140$. The same voltage reference changes used in Subsection 5.1 are performed in this simulation set:

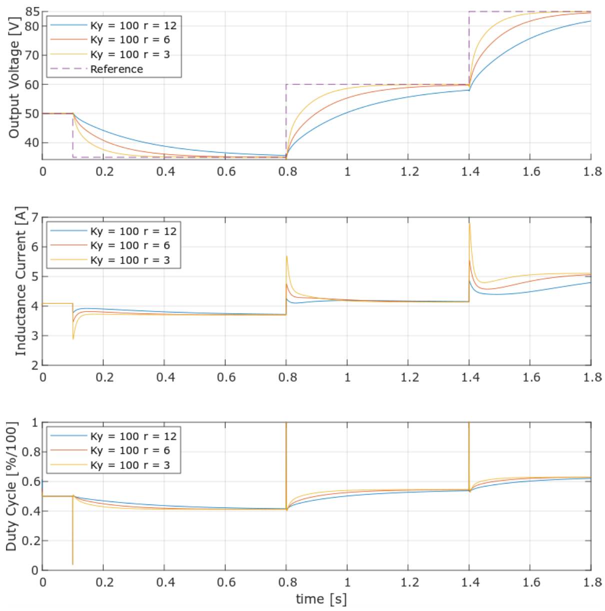

Figure 7: Controller performance with disturbance estimator implemented.

It is observed that the disturbance estimator degrades the transient performance of the whole system, as the response time increases and the inductance current overshoot is larger. On the other hand the estimator is needed to achieve the control objective in presence of a disturbance.

6. CONCLUSIONS

In this paper a control system design method for two DC-DC power electronics converters has been presented, where the control input acts through the system interconnection structure. Flatness–based coordinate transformation together with a state dependent input transformation allowed to write the open loop dynamics of both converters in linear affine form with a constant input matrix. This allow for an easy solution of the matching equation, usually the most difficult task in the IDA-PBC methodology, yielding a closed-loop in pH form with constant structure and dissipation matrices.

The main advantage of this solution is that it is not restricted to linear loads but works for a wide variety of nonlinear loads, including nonlinearities with negative incremental resistance in their Volt-Ampère characteristic. The resulting control law seems complicated in the transformed coordinates but in the original variables it can be easily understood as composed of a sum of power and energy terms. The controller implementation is also a simple task. The closed loop performance can be adjusted via experimental tuning (digital simulation) of a single parameter.

Instead of adding a PI for disturbance rejection - a technically difficult task in this case if intending to preserve the pH-form - the alternative of using a dynamic estimator was developed, which proved to preserve the equilibrium of the transformed variable, thus fulfilling the control objective. It is worthy to note that, adding the estimator has an impact on the original controller, increasing the response time and the current overshoot during transients. Increasing the speed of the estimator reduces the performance degradation of the controller, but demands a higher control effort. Depending on the application, a compromise relationship between response time and control effort must be found by the designer.

Future work will focus on extending the application of this control system design approach to DC electrical grids including energy storage and distributed sources, where the control of the power flow to satisfy the energy management requirements of the system is performed by these kind of converters.

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