

Article

Dynamic Extension IDA-PBC for an Active Switched-Inductor High-Gain Power Converter

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Abstract

This paper presents a nonlinear control scheme for regulating the output voltage of a high-gain active switched-inductor boost converter. The proposed converter topology extends the conventional boost converter by incorporating an additional inductor and a power semiconductor device, thereby enhancing the voltage conversion ratio. The control objectives are twofold: precise regulation of the output voltage and stabilization of the inductor current. To achieve these goals, an interconnection and damping assignment passivity-based control strategy is developed. A dynamic extension is further introduced to compensate for steady-state errors caused by unmodeled parasitic resistances in the system components. In addition, a reference current generator with proportional–integral action is implemented to provide an appropriate current reference. The effectiveness of the proposed controller is validated both numerically and experimentally under three operating scenarios: load step changes, input voltage variations, and output voltage reference transitions.

Keywords: high-gain power converters; output voltage regulation; passivity-based control; port-controlled Hamiltonian systems



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1. Introduction

The electrical industrial sector has been motivated to increase the integration of renewable energy due to the high environmental effects brought on by the excessive use of pollutant energy sources like oil, coal, and natural gas. Conditioning and storing energy generated by renewable sources is crucial; however, the output voltage from the vast majority of renewable sources is low and uncontrolled, meaning that it cannot be utilized in a wide range of applications. For this reason, the adoption of renewable energies at domestic and industrial levels requires the development of power electronic systems as well as control strategies to operate under reliable conditions [1].

The well-known DC-DC boost converter (BC) is a functional topology for increasing the output voltage of DC renewable sources such as photovoltaic (PV) and fuel cell (FC) systems; however, the BC's low efficiency for a high-duty ratio results in a limited voltage boost for applications dealing with high output power [2]. A solution to this problem involves

employing transformerless high-gain BCs that use voltage-boosting techniques [3], such as switched capacitors (charge pump) [4], switched-inductor networks [5], voltage multiplier cells [6], voltage lift techniques [7], magnetic coupling [8], and multistage converters [2]. These high-gain non-isolated converters are low cost, energy efficient, and compact compared to converters with high-frequency transformers. Moreover, to further enhance the voltage conversion ratio, some combinations of these voltage-boosting techniques can be implemented, such as switched inductors with switched capacitors [9], switched inductors with magnetic coupling [10], and switched inductors with voltage lift [11].

For instance, in [5], three active switched-inductor (ASL) BCs are presented. The structure of these ASL converters is based on the conventional BC; however, the main difference is that an inductor and a power semiconductor device have been added to the topology, which enhances the conversion ratio of the output voltage while keeping a duty cycle within acceptable limits when compared with the conventional BC. Some advantages of these ASL-BC are low-voltage stress in semiconductors, high-voltage gain, and the use of equal-value inductors to simplify the analysis. Additionally, from a dynamic performance and feedback point of view, topologies focused on renewable energy require advanced control schemes that regulate the output voltage, achieve current tracking for a secure operation, and even mitigate the effect of uncertainties and disturbances. In addition, power converters with boosting capability exhibit a non-minimum phase behavior of the output voltage with respect to the duty cycle, which may cause instability and poor dynamic performance. These challenges are effectively addressed by the well-known current-mode control (CMC) where the output voltage is driven by a direct control of the inductor current. To this end, CMC employs a multi-loop control strategy in which the inner loop is dedicated to current regulation and the outer loop is responsible for output voltage control. From a stability standpoint, both loops can be designed independently, each admitting its own stability proof. Several control strategies have been reported in the literature, regarding CMC. For example, ref. [12] uses classical control theory to design a CMC switching regulator by using a quadratic BC to operate an FC system. In this case, the inner loop is based on a high-gain compensator, and the outer loop uses a proportional–integral (PI) controller. Furthermore, in [13], for the high-gain ASL-BC considered in this work as well, three controllers based on multi-loop PI actions are designed and experimentally validated. These controllers use the time-scale separation principle between the current and voltage dynamics to construct independent inner and outer loops which simplifies the control design stage. However, this approach does not provide formal stability guarantees, as no rigorous stability proof is offered for either the inner or outer loop. Furthermore, a CMC considering Euler–Lagrange (EL) standard passivity-based control (PBC) and a PI action for the inner and outer loops, respectively, is detailed in [14,15], to control a double-inductor BC and an FC/BC system, respectively. However, this approach requires the online solution of four additional differential equations to reconstruct the parasitic parameters and compute the control signal. In particular, the PBC approach has been widely used to control power electronics [16–19] due to its main characteristic: reformulation of energy balancing, i.e., system energy can be driven to a desired energy level. Under this idea, three main branches have been reported: EL standard PBC [20], and PBC for port-controlled Hamiltonian (PCH) systems, which in turn can be divided into interconnection and damping assignment (IDA) [21] and PI-PBC [22]. The IDA-PBC control approach has been considered in several applications such as DC microgrid systems [23,24], robotics, mechanical and electromechanical systems [25–29], and for power conversion [30–33]. Explicitly, the IDA-PBC approach is based on assigning a desired energy function, interconnection, and damping using the control input. Different from other nonlinear control strategies that focus on the signal approach, IDA-PBC exploits the

physical structure and considers the energy relationship of the system to design stabilizing control laws with practical feasibility [32]. Thus, this work details the design of a nonlinear control scheme under CMC to regulate the output voltage of a high-gain ASL BC. The IDA-PBC strategy is developed for the inner current loop, while a model-based reference current generator (RCG) to provide the current reference is designed as an outer voltage loop. In addition, a dynamic extension is constructed to cope with the parasitic effects of the converter, such as the non-modeled inductor resistance and parameter uncertainty, thus eliminating steady-state error. The following are the main contributions of this research regarding the output voltage regulation of the ASL-BC system:

- The port-controlled Hamiltonian formulation for the ASL-BC.
- The nonlinear control scheme design for the ASL-BC is based on interconnection and damping assignment passivity-based control (IDA-PBC) with an integral action to eliminate the effect of non-modeled parasitic resistances and parameter uncertainty in the system. No restrictions are imposed over the control tuning gains, nor does the control law exhibit singularities; therefore, asymptotic stability is ensured for the operation range of the ASL-BC.
- The combination of a nonlinear control strategy for the inner current loop and a model-based reference current generator to avoid an intricate approximation of the system equilibrium. This approach requires the online solution of only two additional differential equations to compute the duty cycle.
- Numerical and experimental validation of the closed-loop system under several scenarios.

The rest of the paper is organized as follows: In Section 2, the operation and modeling of the ASL-BC in the PCH formulation approach are analyzed. Section 3 details the design of the proposed IDA-PBC controller along with the dynamic extension. Section 4 shows the evaluation of the proposed control law with numerical and experimental results. Finally, Section 5 addresses the conclusion and future work.

2. System Description

The ASL-BC under study is shown in Figure 1. The topology consists of a DC source v_i used as a power supply to the converter, the ASL network formed by two active switches S_{w1} and S_{w2} that operate in synchrony, and inductors L_1 and L_2 . This switching array allows the inductors to be charged in parallel and discharged in series. Additionally, a diode D , an output capacitor C , and a resistor R_o used to model the output nominal load are considered. By virtue of its topology, the ASL-BC achieves a high voltage conversion ratio without requiring an extreme duty cycle, making it particularly suitable for applications such as fuel cell systems, photovoltaic systems, and other energy conversion systems that demand a high DC-DC step-up voltage gain [5].

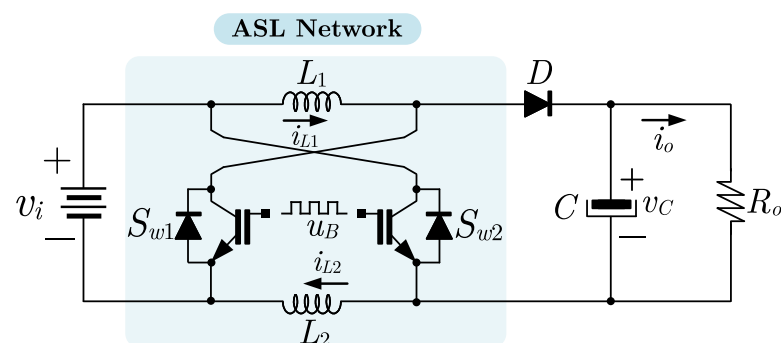


Figure 1. ASL-BC configuration.

Remark 1. Assumptions 1–3 represent standard idealizations commonly adopted in the passivity-based control literature for power converters [34], yielding a tractable averaged model while preserving the essential dynamic behavior for control design. The unmodeled parasitic resistances in 2–3 motivate the dynamic extension, which compensates for the resulting steady-state error without requiring explicit knowledge of their values, as demonstrated in Section 4. Regarding 1, unknown stepwise changes in v_i are explicitly considered in Scenario ii. More generally, the integral action x_c converges to a non-zero steady-state value that offsets any unmodeled constant parameters or disturbances, as established in Proposition 1.

To define the physical dynamical behavior of the ASL-BC, it is necessary to consider the following idealizing assumptions:

Assumption 1. The input voltage source v_i is considered ideal, i.e., provides a constant voltage.

Assumption 2. The semiconductors S_{w1} , S_{w2} , and the diode D are considered energy lossless.

Assumption 3. Energy losses in inductors due to unknown constant parasitic resistances are not considered in the average system model; however, real-world parasitics are compensated by the dynamic extension proposed in this work.

Assumption 4. The continuous conduction mode (CCM) is considered for the physical operation of the system. This mode means that the inductor current never reaches zero value during operation. The discontinuous conduction mode (DCM), which is when the inductor current reaches zero value during operation, is not considered due to the high input current ripple that may affect the power supply [5].

Assumption 5. Furthermore, L_1 , L_2 , C , and R_o are known constant parameters.

As a result of the synchronous operation of the active switches S_{w1} and S_{w2} , the operation of the ASL-BC is described as follows. During turn ON mode the switches are activated, and the inductors are charged in parallel. The stored energy in the capacitor C is delivered to the load R_o . Consequently, the diode D is reverse-biased and therefore isolates the load from the power supply. During turn OFF mode the switches are disabled, and the stored energy in the inductors is discharged in series to the capacitor C . Since the diode D is forward-biased, it allows current to circulate from the inductors and source to the nominal load R_o . Consequently, based on the operation modes described above and Kirchhoff's current and voltage laws, the dynamic average model of the system is determined as

$$\begin{aligned}\dot{x}_1 &= \frac{1}{2L} \left(-(1-u)x_2 + (1+u)v_i \right), \\ \dot{x}_2 &= \frac{1}{C} \left((1-u)x_1 - \frac{x_2}{R_o} \right),\end{aligned}\tag{1}$$

where $x = [x_1 \ x_2]^\top = [i_L \ v_C]^\top$ is the state vector of the system. Note that i_L and v_C represent the current flowing through inductors L_1 and L_2 and the output capacitor voltage, respectively. The input control u indicates the duty cycle to trigger the active switches through a PWM function. As seen in Figure 1, three energy storage devices are observed; therefore, a third-order model can be expected. However, considering $L_1 = L_2 = L$, it is possible to reduce the order of the system from third to second. Consequently, the current dynamics in (1) presents the time constant $2L$. In addition, it is assumed all state variables are measurable. Furthermore, coherent with the physical meaning, x and the control signal u are considered to be bounded from below to ensure CCM and from above for safety

reasons. In steady state, considering that the control signal remains constant, i.e., $u = \bar{u}$, the equilibrium points of (1) are found to be

$$\begin{aligned}\bar{x}_1 &= \frac{(1 + \bar{u})v_i}{R_o(1 - \bar{u})^2} \\ \bar{x}_2 &= \frac{(1 + \bar{u})v_i}{(1 - \bar{u})}.\end{aligned}\quad (2)$$

It is worth noting that the ASL-BC topology achieves a higher voltage conversion ratio than both the conventional BC and the interleaved boost converter, which are limited to a ratio of $\bar{x}_2 = v_i/(1 - \bar{u})$, while the ASL-BC attains this enhancement by simply incorporating an additional switched-inductor branch into the conventional topology.

Remark 2. To compute the equilibrium point for the inductor current, complete knowledge of the load resistance is required, which is subject to unknown variations in practice. Additionally, as presented in [5,14,35], the transfer function relating the output voltage to the control signal exhibits a non-minimum phase behavior; therefore, implementing a controller to regulate the output voltage directly is not suitable, since it may cause instability; instead, a CMC is preferred.

In order to design an IDA-PBC control scheme for the inner current loop, the physical behavior of the plant shown in Figure 1 is reformulated based on PCH systems as

$$\dot{x} = [\mathcal{J} - \mathcal{R}]\nabla_x \mathcal{H} + \mathcal{G}(x)u + \zeta, \quad (3)$$

where $\mathcal{J} = -\mathcal{J}^\top$ represent a skew-symmetry interconnection matrix containing the passive elements that store energy in the system, and the matrix $\mathcal{R} = \mathcal{R}^\top \geq 0$ refers to the dissipation matrix that contains the dissipating components, such as the nominal load R_o . The Hamiltonian (energy) function is represented by $\mathcal{H}(x)$ and consequently $\nabla_x \mathcal{H} = \partial \mathcal{H}(x)/\partial x$. The port characteristic matrix of the converter is denoted by $\mathcal{G}(x)$, and ζ is the input or disturbance vector. In this case, the associated matrices to (3) for the ASL-BC are defined as

$$\begin{aligned}\mathcal{J} &= \begin{bmatrix} 0 & -1/2LC \\ 1/2LC & 0 \end{bmatrix}, & \mathcal{R} &= \begin{bmatrix} 0 & 0 \\ 0 & 1/C^2 R_o \end{bmatrix}, \\ \mathcal{G}(x) &= \begin{bmatrix} (x_2 + v_i)/2L \\ -x_1/C \end{bmatrix}, & \zeta &= \begin{bmatrix} v_i/2L \\ 0 \end{bmatrix}.\end{aligned}$$

The Hamiltonian function of the ASL-BC is selected as the sum of the inductors and capacitor energy as

$$\mathcal{H}(x) = \frac{1}{2} (2Lx_1^2 + Cx_2^2), \quad (4)$$

and consequently the partial derivative of the energy with respect to the state is

$$\nabla_x \mathcal{H} = \frac{\partial \mathcal{H}(x)}{\partial x} = [2Lx_1 \quad Cx_2]^\top. \quad (5)$$

Problem formulation: Consider that the ASL-BC given in (3) fulfills Assumptions 1–5. It is required to design a dynamic state feedback control strategy such that the equilibrium of the extended inner closed-loop system is asymptotically stable.

In the following section, an IDA-PBC design with an integral action is detailed to operate the ASL-BC ensuring current tracking and output voltage regulation in the presence of parasitic, non-modeled resistances and capable of facing input voltage and load variations.

3. IDA-PBC Control Scheme Design

The main idea of the IDA-PBC strategy, based on PCH systems, is presented in [21] and relies on the designation of a desired storage energy function x_d , with a desired fixed equilibrium point, interconnection, and damping assignment via feedback. To ensure system stability, the desired energy storage function must have a minimum at the desired equilibrium point. Additionally, the damping assignment is performed through virtual resistances, which dissipate energy and impose the convergence rate to the desired point. Furthermore, the ideas above can be extended with an integral action to eliminate model errors while preserving the stability of the closed-loop system [36]. In this section, the design of an IDA-PBC scheme law is proposed for the inner current loop to generate the duty cycle for the ASL-BC, and then, a model-based reference current generator is implemented to provide a current reference to be tracked by the inner loop. Moreover, for the ASL-BC operating in a closed loop, the control objectives are described in detail below.

1. Input inductor current $x_1(t)$ tracking to a desired reference current $x_1^*(t)$, i.e.,

$$\lim_{t \rightarrow \infty} x_1(t) = x_1^*(t), \quad (6)$$

where the desired current value $x_1^*(t)$ is obtained from the model-based reference current generator.

2. Output voltage $x_2(t)$ regulation to a desired reference value x_2^*

$$\lim_{t \rightarrow \infty} x_2(t) = x_2^*, \quad (7)$$

where the reference voltage desired value x_2^* is defined by the nominal operation of the ASL-BC.

3.1. IDA-PBC Inner Current Loop

In this section, the CMC is designed based on the IDA-PBC theory for current tracking purposes. The resulting IDA-PBC scheme with dynamic extension is summarized in the following proposition and illustrated with the block diagram in Figure 2b.

Proposition 1. Consider the ASL-BC system (Figure 1) verifying Assumptions 1–5, with bounded state x and control signal u , then the dynamic nonlinear controller

$$u = \frac{1}{\left(\frac{x_2 + v_i}{2L}\right)^2 + \left(\frac{x_1}{C}\right)^2} \left[\frac{x_2 + v_i}{2L} \left(-\alpha \tilde{x}_1 + \frac{x_2}{2L} - \frac{v_i}{2L} \right) + \frac{x_1}{C} \left(\beta \tilde{x}_2 + \frac{x_1}{C} - \frac{x_2}{CR_o} \right) \right] + x_c, \quad (8)$$

with the current reference x_1^* computed by the RCG (21), a given voltage reference x_2^* and the integral action x_c in (19), with $x_c(0) \in \mathbb{R}$ and sufficiently large tuning gains α , β and κ_I , is well-defined and asymptotically stabilizes the system trajectories towards the reference values.

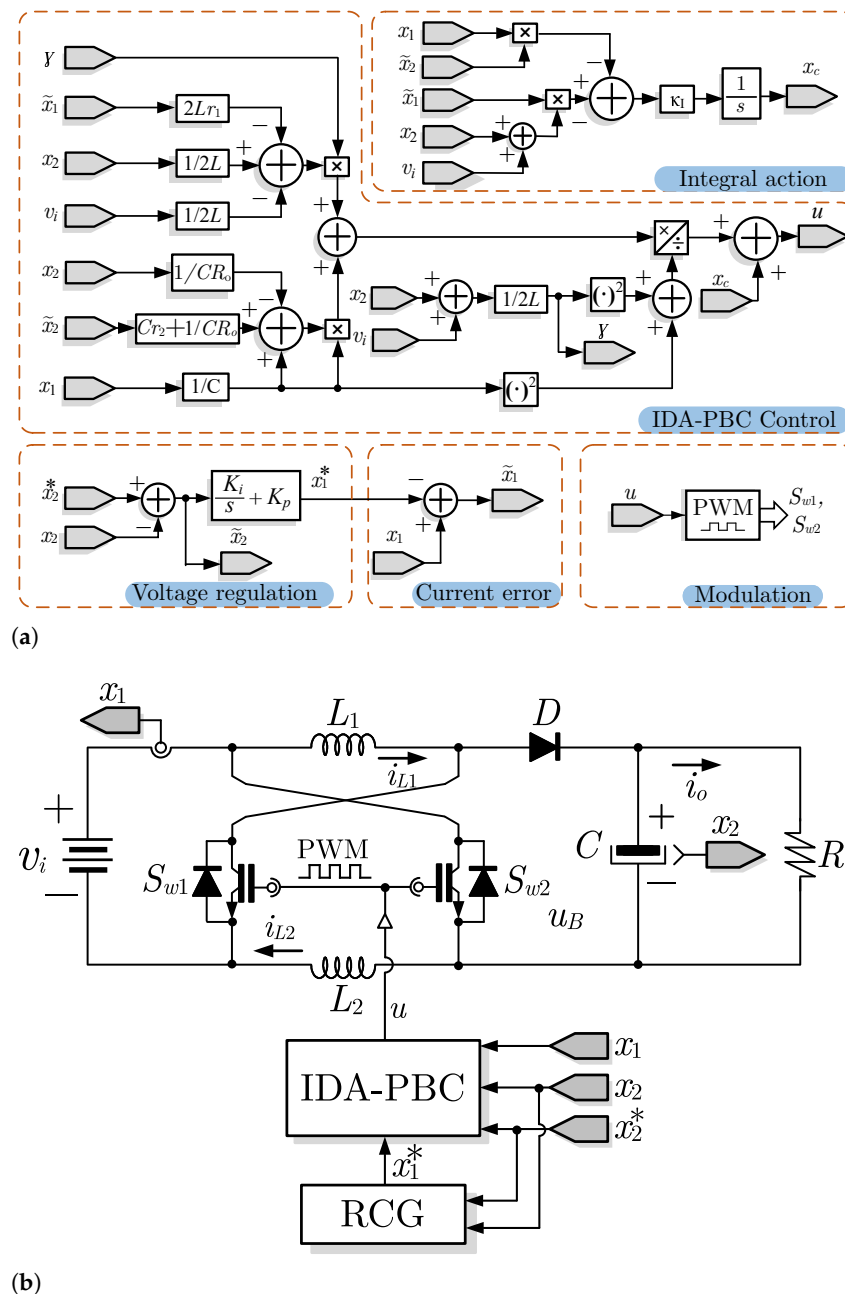


Figure 2. IDA-PBC control scheme: (a) detailed structure, (b) overall implementation.

Proof. To fulfill the control objectives stated above, the following desired closed-loop dynamics are prescribed as

$$\dot{x} = [\mathcal{J}_d - \mathcal{R}_d] \nabla_x W + \mathcal{G}(x) \kappa_I \nabla_{x_c} W \quad (9)$$

$$\dot{x}_c = -\kappa_I \mathcal{G}^\top(x) \nabla_x W, \quad (10)$$

where (9) represents the desired port-Hamiltonian dynamics of the ASL-BC, in which $\mathcal{J}_d$ and $\mathcal{R}_d$ are the desired interconnection and damping matrices, respectively, and $W(\tilde{x}, x_c)$ is the total energy function that serves as a Lyapunov function for the closed-loop system. The second Equation (10) defines the dynamics of the integral action x_c , which is driven by the port characteristic matrix $\mathcal{G}(x)$ and the gradient $\nabla_x W$, and is introduced to compensate for the steady-state error caused by the unmodeled parasitic resistances in the system. In particular, $\mathcal{J}_d = \mathcal{J} + \mathcal{J}_a = -\mathcal{J}_d^\top$ is the desired interconnection matrix and $\mathcal{R}_d = \mathcal{R} + \mathcal{R}_a$ is the desired damping matrix, with both auxiliary matrices selected as

$$\mathcal{J}_a = \begin{bmatrix} 0 & -j \\ j & 0 \end{bmatrix}, \quad \mathcal{R}_a = \begin{bmatrix} r_1 & 0 \\ 0 & r_2 \end{bmatrix},$$

where the positive control gains r_1 and r_2 inject virtual damping into the system. Notice that setting $j = -1/2LC$, the desired interconnection matrix results in

$$\mathcal{J}_d = \mathcal{J} + \mathcal{J}_a = \mathcal{J} = \begin{bmatrix} 0 & -1/2LC \\ 1/2LC & 0 \end{bmatrix} + \begin{bmatrix} 0 & 1/2LC \\ -1/2LC & 0 \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix} \quad (11)$$

Hence, interconnections are canceled, and the control scheme design is simplified. In addition, the total energy storage function $W(\tilde{x}, x_c)$ is the sum of the desired Hamiltonian function $\mathcal{H}_d(\tilde{x})$ and the dynamic extension energy function, and both are defined as

$$W(\tilde{x}, x_c) = \mathcal{H}_d(\tilde{x}) + \frac{1}{2}\kappa_I^{-1}x_c^2, \quad (12)$$

where

$$\mathcal{H}_d(\tilde{x}) = \frac{1}{2}(2L\tilde{x}_1^2 + C\tilde{x}_2^2). \quad (13)$$

Notice that the control error is defined as the difference between the system state and the reference value, i.e.,

$$\tilde{x} = x - x^* = \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} - \begin{bmatrix} x_1^* \\ x_2^* \end{bmatrix}, \quad (14)$$

where $\tilde{x}_1$ is the current reference computed by the outer loop below in (21) and $\tilde{x}_2$ is the constant output voltage reference. Furthermore, the partial derivative of W with respect to the state x and the controller dynamic extension x_c are

$$\begin{aligned} \nabla_x W &= \frac{\partial W}{\partial x} = \begin{bmatrix} 2L\tilde{x}_1 \\ C\tilde{x}_2 \end{bmatrix} \\ \nabla_{x_c} W &= \frac{\partial W}{\partial x_c} = \kappa_I^{-1}x_c, \end{aligned} \quad (15)$$

where $x_c \in \mathbb{R}$ is the dynamic extension (integral action) and $\kappa_I > 0$ is the integral gain. Recall that the dynamic extension is added to the control scheme to deal with non-modeled parasitic resistances in the system. Afterward, in order to compute the explicit controller u , the PCH system model (3) is matched with the desired closed-loop dynamics (9) as follows:

$$[\mathcal{J}_d - \mathcal{R}_d]\nabla_x W + \mathcal{G}(x)\kappa_I\nabla_{x_c} W = [\mathcal{J} - \mathcal{R}]\nabla_x \mathcal{H} + \mathcal{G}(x)u + \zeta. \quad (16)$$

This matching equation equates the desired closed-loop dynamics with the open-loop ASL-BC model; its purpose is to bridge the physical plant description and the target stable closed-loop behavior, such that solving it for the control input u directly yields the IDA-PBC control. In this way, solving (16) for the control signal u results in

$$u = \mathcal{G}^\dagger(x) \left([\mathcal{J}_d - \mathcal{R}_d]\nabla_x W - [\mathcal{J} - \mathcal{R}]\nabla_x \mathcal{H} - \zeta \right) + x_c \quad (17)$$

where $\mathcal{G}^+(x)$ is a pseudo inverse, and is obtained from [36]

$$\mathcal{G}^+(x) = \left(\mathcal{G}^\top(x) \mathcal{G}(x) \right)^{-1} \mathcal{G}^\top(x) = \frac{1}{\left(\frac{x_2 + v_i}{2L} \right)^2 + \left(\frac{x_1}{C} \right)^2} \begin{bmatrix} \frac{x_2 + v_i}{2L} & -\frac{x_1}{C} \end{bmatrix}. \quad (18)$$

Hence, based on (17), the control signal u results in (8), with $\alpha = 2Lr_1$ and $\beta = Cr_2 + 1/CR_o$ and the dynamic extension (integral action)

$$\dot{x}_c = -\kappa_I \mathcal{G}^\top(x) \nabla_x W = -\kappa_I \left((x_2 + v_i) \tilde{x}_1 - x_1 \tilde{x}_2 \right), \quad (19)$$

where $x_c(0) \in \mathbb{R}$. On the other hand, the stability of the current loop with the dynamic extension is proven straightforward. Consider the closed-loop desired energy function (12) as a Lyapunov function. Note that $W(\tilde{x}, x_c)$ is a positive definite function and $W(0, 0) = 0$. In this way, the time derivative of (12) along the trajectories of (9) and (10) and considering the skew-symmetry property of $\mathcal{J}_d$, is computed as

$$\begin{aligned} \dot{W}(\tilde{x}, x_c) &= \nabla_x W^\top \dot{\tilde{x}} + \nabla_{x_c} W^\top \dot{x}_c \\ &= \nabla_x W^\top [\mathcal{J}_d - \mathcal{R}_d] \nabla_x W + \mathcal{G}(x) \kappa_I \nabla_{x_c} W \\ &\quad - \nabla_{x_c} W^\top \kappa_I \mathcal{G}^\top(x) \nabla_x W \\ &= \nabla_x W^\top [\mathcal{J}_d - \mathcal{R}_d] \nabla_x W \\ &\quad + \underbrace{\nabla_x W^\top \mathcal{G}(x) \kappa_I \nabla_{x_c} W - \nabla_{x_c} W^\top \kappa_I \mathcal{G}^\top(x) \nabla_x W}_{\text{(cancellation by transposition)}} \\ &= \underbrace{\nabla_x W^\top \mathcal{J}_d \nabla_x W}_{(\mathcal{J}_d \text{ selected as zero})} - \nabla_x W^\top \mathcal{R}_d \nabla_x W \\ &= -\nabla_x W^\top \mathcal{R}_d \nabla_x W < 0 \end{aligned} \quad (20)$$

where $\dot{W}(\tilde{x}, x_c)$ is negative definite and consequently the origin of (9) is stable in the sense of Lyapunov, and asymptotic stability is achieved by invoking La Salle's invariance principle. $\square$

Remark 3. The control law (8) involves the pseudo-inverse $\mathcal{G}^+(x)$ defined in (18), whose denominator is given by $D_{en} = \left(\frac{x_2 + v_i}{2L} \right)^2 + \left(\frac{x_1}{C} \right)^2$. Since $v_i > 0$ is a strictly positive constant, evaluating D_{en} at the worst-case condition $x_1 = 0$ and $x_2 = 0$ yields $D_{en} = \left(\frac{v_i}{2L} \right)^2 > 0$; therefore, $\mathcal{G}^+(x)$ is well-defined and bounded away from singularity under all feasible physical operating conditions.

3.2. Reference Current Generator

The control law (8) requires both references x_1^* and x_2^* to implement the control error $\tilde{x}$ defined in (14) and to fulfill the control objectives (6) and (7). Meanwhile, the voltage error $\tilde{x}_2$ is obtained from the direct measurement of the output voltage x_2 and the explicit definition of the constant output voltage reference x_2^* ; to implement the current error, besides the current measurement, a current reference x_1^* needs to be defined. Different ways have been implemented in the literature for this end; for instance, the equilibrium point $\bar{x}_1$ in (2) might act as the current reference if the load R_o is known but it varies during operation and consequently an adaptation law to approximate the load is required if $\bar{x}_1$ is used to construct x_1^* [33,34]. On the other hand, based on traditional CMC, an outer voltage loop based on a simple PI action over the output voltage x_2 is traditionally proposed to generate the current reference x_1^* ; for instance, see [12,15,35]. In this work, to avoid using

adaptive schemes, and based on traditional CMC, an outer voltage loop based on a PI action over the output voltage x_2 is proposed to generate the current reference x_1^* as

$$\begin{aligned}x_1^* &= -K_p \tilde{x}_2 - K_i \tilde{\zeta}, \\ \dot{\tilde{\zeta}} &= \tilde{x}_2\end{aligned}\quad (21)$$

where K_p and K_i are the positive control gains to be selected. Remark that x_1^* is used as input to the current loop in (14). In order to prove the stability of the outer loop, the time-scale separation principle is invoked, in accordance with the CMC framework [12,15,35]. It is considered that the current dynamics are faster than the voltage dynamics; therefore, it is possible to decouple these two dynamics. This decoupling is consistent with the singular perturbation theory provided in [37]. In this way, the inner loop ensures that $x_1 \rightarrow x_1^*$; consequently, the voltage dynamics in (1) are rewritten as

$$\dot{x}_2 = \frac{1}{C} \left((1 - \bar{u}) x_1^* - \frac{x_2}{R_o} \right), \quad (22)$$

where $\bar{u}$ denotes the steady-state value reached by the control signal u , and x_1^* is now considered the new control input to dominate the output voltage dynamics. Recalling that $\tilde{x}_2 = x_2 - x_2^*$; then, Equation (22) can be rewritten in terms of the voltage error as

$$\dot{\tilde{x}}_2 = \frac{1}{C} \left((1 - \bar{u}) x_1^* - \frac{\tilde{x}_2 + x_2^*}{R_o} \right). \quad (23)$$

Afterwards, consider (23) in closed loop with (21), which result in

$$\begin{aligned}\dot{\tilde{x}}_2 &= - \left(\frac{\bar{u}' K_p}{C} + \frac{1}{C R_o} \right) \tilde{x}_2 - \frac{\bar{u}' K_i}{C} \tilde{\zeta} - \frac{x_2^*}{C R_o} \\ \dot{\tilde{\zeta}} &= \tilde{x}_2\end{aligned}\quad (24)$$

where $\bar{u}' = (1 - \bar{u})$. The equilibrium point of the last expression is $(\tilde{x}_2, \tilde{\zeta}) = (0, -\frac{x_2^*}{\bar{u}' K_i R_o})$ and considering a new coordinate $\tilde{\tilde{\zeta}} = \tilde{\zeta} - \bar{\zeta}$, the slow subsystem (24) can be expressed in the compact matrix form $\dot{\tilde{z}} = A \tilde{z}$, as

$$\underbrace{\begin{bmatrix} \dot{\tilde{x}}_2 \\ \dot{\tilde{\tilde{\zeta}}} \end{bmatrix}}_{\dot{\tilde{z}}} = \underbrace{\begin{bmatrix} - \left(\frac{\bar{u}' K_p}{C} + \frac{1}{C R_o} \right) & - \frac{\bar{u}' K_i}{C} \\ 1 & 0 \end{bmatrix}}_A \underbrace{\begin{bmatrix} \tilde{x}_2 \\ \tilde{\tilde{\zeta}} \end{bmatrix}}_{\tilde{z}}, \quad (25)$$

where the stability of the equilibrium $\tilde{z}^* = 0$ is determined by the eigenvalues of A . The characteristic polynomial of A is given by

$$p(\lambda) = \det(\lambda I - A) = \lambda^2 + \underbrace{\left(\frac{\bar{u}' K_p}{C} + \frac{1}{C R_o} \right)}_{a_1} \lambda + \underbrace{\frac{\bar{u}' K_i}{C}}_{a_0} = 0. \quad (26)$$

By the Routh–Hurwitz stability criterion, both roots of (26) have strictly negative real parts if and only if $a_1 > 0$ and $a_0 > 0$. Since $\bar{u}' > 0$, $K_p > 0$, $K_i > 0$, $C > 0$, and $R_o > 0$ hold by construction, both conditions are satisfied for any admissible operating point. Therefore, the matrix A is Hurwitz, and the equilibrium $\tilde{z}^* = 0$ of the slow subsystem is asymptotically stable, establishing the convergence $(\tilde{x}_2, \tilde{\tilde{\zeta}}) \rightarrow (0, 0)$ as $t \rightarrow \infty$. The overall block diagram of the control scheme is presented in Figure 2.

4. Numerical and Experimental Results and Discussion

The proposed multi-loop nonlinear control strategy based on IDA-PBC given in (8)–(19) and (21), and shown in Figure 2b for the output voltage regulation of the ASL-BC is validated through both numerical and experimental results. In Table 1, the system parameters used for both cases are shown. To validate the performance, as well as the effectiveness of the proposed controller, the following real-life scenarios are considered for both numerical and experimental results:

- i. Changes in the output load: The output resistance is varied between 100 Ω and 126 Ω , and a voltage reference x_2^* equals to 120 V is set for the output voltage.
- ii. Changes in the input voltage: Stepwise variations in the input voltage from 33 V to 43 V are implemented. The voltage reference is established at 120 V as well.
- iii. Voltage reference changes: The voltage reference is configured to change in steps from 110 V to 120 V. In this case, the input voltage and the load are at their nominal constant values.

Table 1. System parameters.

System Parameters	Value
Input voltage v_i	33 V
Nominal load R_o	126 Ω
Inductance L	1 mH
Capacitor C	380 μ F
Voltage reference x_2^*	120 V
Switching frequency f_s	75 kHz

4.1. Simulation Results and Comparative Analysis

For simulation results, the ASL-BC is implemented using the SimPowerSystem toolbox from MATLAB/Simulink, version number R2025a. Consequently, for both inductors, the parasitic resistance is set to 0.5 Ω . Additionally, for the three scenarios, the changes are introduced at a rate of 4 Hz. The comparative analysis considers the proposed IDA-PBC controller with the dynamic extension, a nested PI controller that substitutes the proposed inner current loop by a standard PI action over the current error $\tilde{x}_1$, and finally the PI-PI controller designed in [13]. The gains for each controller are gathered in Table 2. In addition, Table 3 presents the performance error indices computed for the output voltage error $\tilde{x}_2$ of each controller under the three operating scenarios. Four indices are considered: the Integral of Squared Error (ISE), the Integral of Absolute Error (IAE), the Integral of Time-Weighted Absolute Error (ITAE), and the Integral of Time-Weighted Squared Error (ITSE). Overall, the proposed IDA-PBC controller achieves the smallest error values across all scenarios, confirming the superior tracking performance observed in Figures 3–5.

Table 2. Numerical test control gains for each controller.

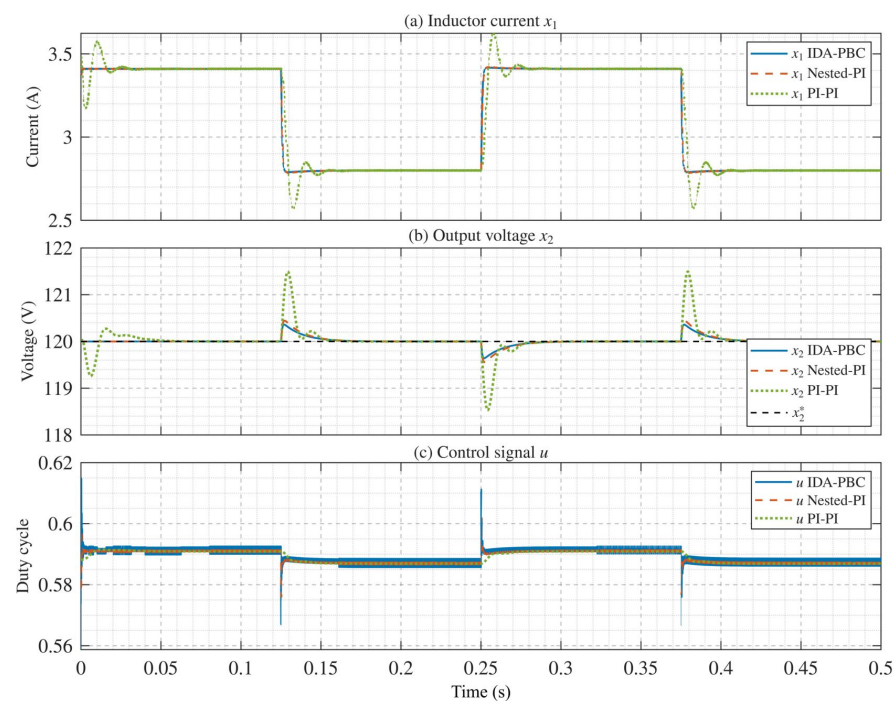
Control Gain	Value
IDA-PBC + Dynamic Extension	
Damping injection r_1	100×10^3
Damping injection r_2	100
Gain Dyn. Ext. κ_I	1
Proportional gain K_p	1.5
Integral gain K_i	120

Table 2. *Cont.*

Control Gain	Value
Nested PI	
Outer proportional gain K_p	1.5
Outer integral gain K_i	120
Inner proportional gain K_p	0.3
Inner integral gain K_i	5
PI-PI [13]	
Outer proportional gain K_p	1
Outer integral gain K_i	50
Inner proportional gain K_p	60
Inner integral gain K_i	40

Table 3. Performance error indices for the comparative analysis of the three controllers under three operating scenarios. The smallest error value in each category is highlighted in **bold**.

Scenario	Controller	ISE	IAE	ITAE	ITSE
Load step change	IDA-PBC	0.003065	0.0153	0.004016	0.0007879
	Nested-PI	0.00314	0.01547	0.004055	0.0008072
	PI-PI [13]	0.03734	0.04575	0.009638	0.008734
Input voltage change	IDA-PBC	0.002597	0.01374	0.003604	0.0006688
	Nested-PI	0.01309	0.02864	0.007629	0.003366
	PI-PI [13]	0.05527	0.04992	0.01294	0.01421
Voltage reference change	IDA-PBC	0.141	0.04866	0.01238	0.0355
	Nested-PI	0.4037	0.09201	0.02354	0.1015
	PI-PI [13]	0.471	0.143	0.03696	0.1199

**Figure 3.** Simulation comparison of the proposed IDA-PBC controller against the Nested-PI and the controller in [13] for load step change scenario.

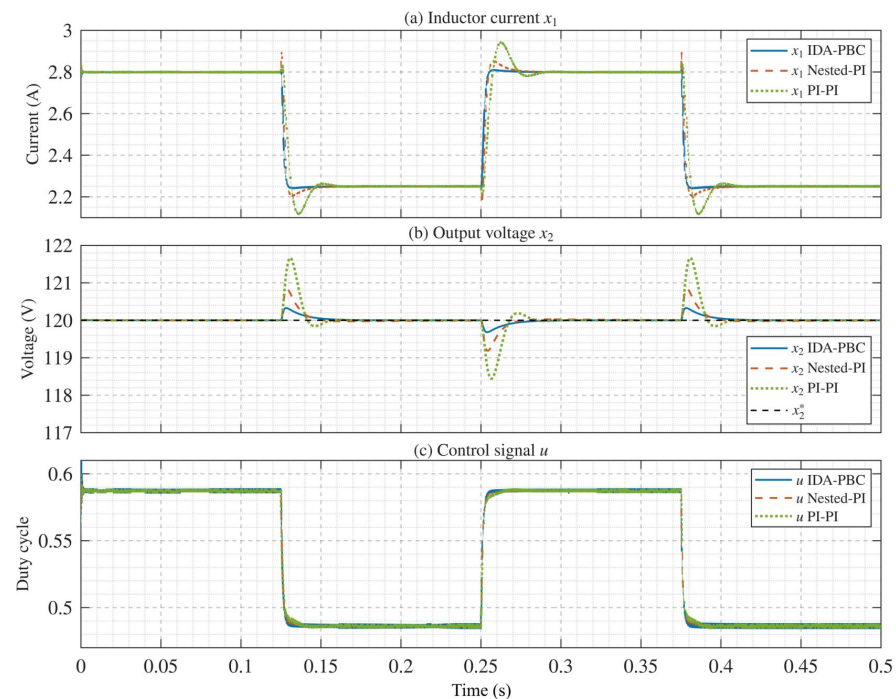


Figure 4. Simulation comparison of the proposed IDA-PBC controller against the Nested-PI and the controller in [13] for input voltage variation scenario.

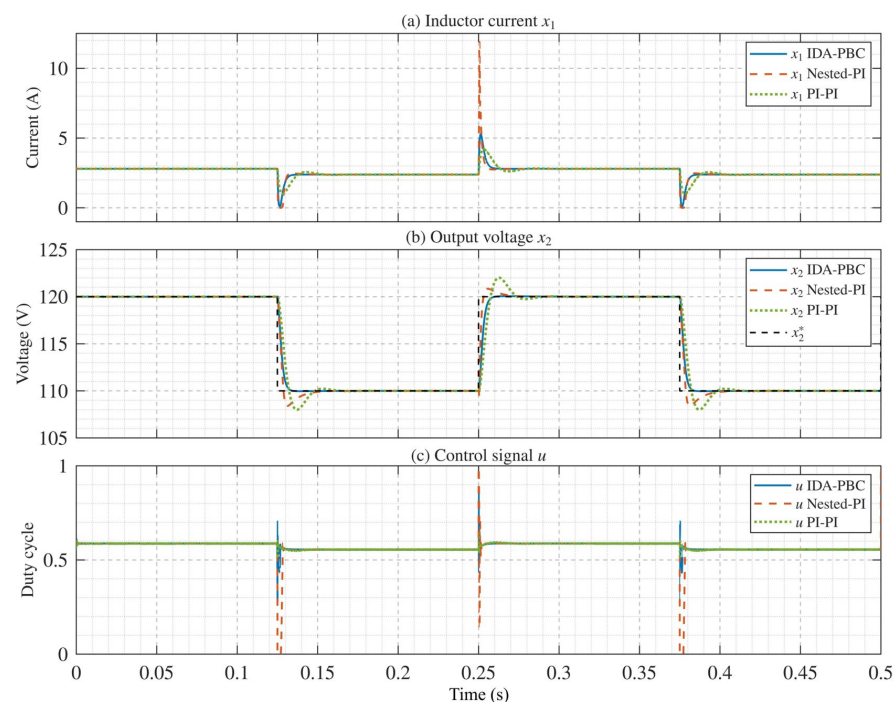


Figure 5. Simulation comparison of the proposed IDA-PBC controller against the Nested-PI and the controller in [13] for output voltage reference transition scenario.

4.2. Numerical Results: Changes in the Output Load

The performance of the three controllers under load variations is depicted in Figure 3. First, Figure 3a shows the inductor current x_1 for each controller, which varies from an initial condition of 3.4 A to 2.8 A at each load transition. Notice that the IDA-PBC and Nested-PI responses are close to each other, whereas the PI-PI controller [13] exhibits an oscillatory behavior, reaching ± 0.3 A at each transition. The output voltage x_2 performance for the three controllers is shown in Figure 3b. In this case, all controllers achieve a similar

settling time of approximately 0.04 s; however, the proposed IDA-PBC presents smaller over- and undershoots with respect to the other controllers, with a deviation of ± 0.38 V. Finally, the control signal u is shown in Figure 3c, where it is observed that the IDA-PBC controller exhibits a more energetic response than the others, with larger transient over- and undershoots that remain within feasible operating bounds.

4.3. Numerical Results: Changes in the Input Voltage

The performance of the three controllers under input voltage variations is depicted in Figure 4. First, Figure 4a shows the inductor current x_1 for each controller, which varies from 2.8 A to 2.25 A at each input voltage transition. Notice that the IDA-PBC response is less oscillatory and more stepwise than the other two controllers, which display a more oscillatory behavior. The output voltage x_2 performance for the all controllers is shown in Figure 4b. In this case, the IDA-PBC and Nested-PI controllers reach the reference in approximately 0.04 s, while the PI-PI controller [13] takes approximately 0.05 s with a more oscillatory response. The proposed IDA-PBC presents smaller over- and undershoots with respect to the other controllers, with a deviation of ± 0.38 V. Finally, the control signal u is shown in Figure 4c, where the three controllers remain within feasible operating bounds, experiencing duty cycle variations from 0.487 to 0.587 with smooth transitions.

4.4. Numerical Results: Reference Voltage Changes

The performance of the three controllers under output voltage reference variations is depicted in Figure 5. First, Figure 5a shows the inductor current x_1 for each controller, which varies from 2.8 A to 2.3 A at each transition. Notice that the IDA-PBC and Nested-PI controllers reach steady state in approximately 0.01 s, while the PI-PI controller [13] takes approximately 0.03 s with a more oscillatory response, yet with smaller over- and undershoots than the former. The output voltage x_2 performance for the three controllers is shown in Figure 5b. In this case, the IDA-PBC controller reaches the reference in approximately 0.01 s, while the Nested-PI and PI-PI [13] controllers take approximately 0.03 s and 0.05 s, respectively, exhibiting a more oscillatory response. The proposed IDA-PBC presents no over- or undershoots in contrast to the other controllers. Finally, the control signal u is shown in Figure 5c, where the IDA-PBC and PI-PI [13] controllers remain within feasible operating bounds, while the Nested-PI exhibits a large transient excursion of the control signal, reaching near-zero values and approaching saturation at unity. All control signals present duty cycle variations from 0.55 to 0.58.

4.5. Experimental Results and Dynamic Extension Comparative Analysis

In Table 1, the control parameters are shown for the experimental case. The experimental test bench depicted in Figure 6 is built, where a dSPACE 1104 control board is used for the digital implementation of the control scheme. Furthermore, in each aforementioned scenario, the obtained results are shown with and without the dynamic extension to emphasize the importance of non-modeled parasitic resistance effect mitigation and steady-state error elimination. Note that changes are at a rate of 0.5 Hz. The controller gains are listed in Table 4.

Table 4. Experimental control gains for the proposed controller.

Control Gain	Value
Damping injection r_1	2×10^3
Damping injection r_2	2×10^3
Gain Dyn. Ext. κ_I	1
Proportional gain K_p	2.5
Integral gain K_i	0.2

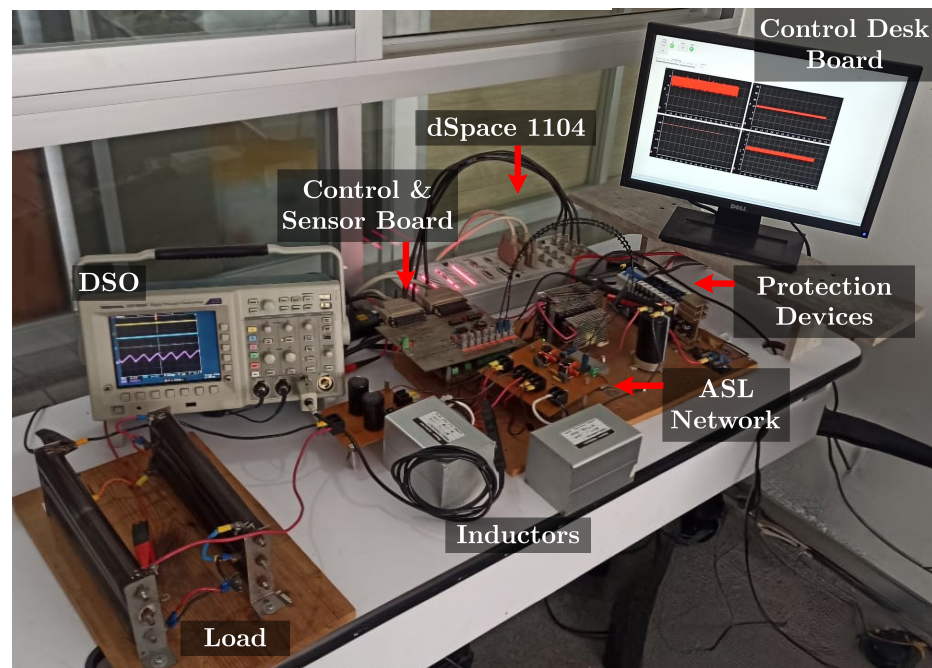


Figure 6. Experimental test bench of the ASL-BC.

4.6. Experimental Results: Changes in the Output Load

In Figure 7, the behavior of the system states and the control signal following step changes in the output resistance is shown. To begin with, Figure 7a depicts the tracking of the current x_1 to the desired value x_1^* . It shows tight tracking with no steady-state error within a range of 2.8 A and 3.44 A. In addition, the regulation of the output voltage x_2 at 120 V is depicted in Figure 7b. The voltage over- and undershoots are due to stepwise variations in the output load. However, these voltage transients are approximately 2% of 120 V with a settling time to reference value x_2^* of 0.4 s. Remark that load variation causes output power variations between 144 W (100 Ω) and 114.28 W (126 Ω). Thus, Figure 7a,b reflect the fulfillment of the control objectives mentioned in Section 3. The time history of u is displayed in Figure 7c, where variations are from 55.83% to 56.71%. On the other hand, the behavior of the control scheme without the integral action (19) is shown in Figure 8. The tracking lack of x_1 to x_1^* is evident in Figure 8a, where a steady-state error is present. Additionally, the error magnitude decreases if the control gain r_1 increases; however, increasing the value of r_1 causes undesirable noise in the transient response of the current x_1 . The regulation of voltage x_2 to the desired voltage x_2^* at 120 V is illustrated in Figure 8b. The magnitude of the voltage peaks caused by variations in the output load is higher compared to Figure 7b. Simultaneously, the control signal shown in Figure 8c differs from the values of its counterpart in Figure 7c. Notice that the controller performance without the dynamic extension does not achieve the current tracking objective. However, voltage regulation is assured and stability is exhibited.

4.7. Experimental Results: Changes in the Input Voltage

The performance of the control scheme following changes in input voltage is illustrated in Figure 9. In addition, in Figure 9a, a tight current tracking to x_1^* is shown with variations from 2.22 A to 2.85 A, and no steady-state error is exhibited. Moreover, Figure 9b shows the output voltage regulation x_2 to the reference value of 120 V, where transients of approximately ± 1 V are exhibited. In addition, the settling time to the reference value is 0.2 s. Furthermore, the control signal is observed in Figure 9c, and displays adjustments from 44.8% to 55.6% to achieve the current tracking and voltage regulation. On the other hand, Figure 10 shows the closed-loop performance when the integral action (19) is not

considered. The steady-state error of current x_1 is shown in Figure 10a, where the maximum error is almost 0.3 A. It is possible to reduce the difference between current x_1 and its reference value x_1^* by increasing the control gains. However, it causes voltage and current overshoots that can result in component damage. Notice that although the current tracking target is not reached, the voltage regulation x_2 to the desired reference x_2^* is achieved in Figure 10b. Similarly, the control signal in Figure 10c, adjusts itself to cope with the stepwise variations in the input voltage.

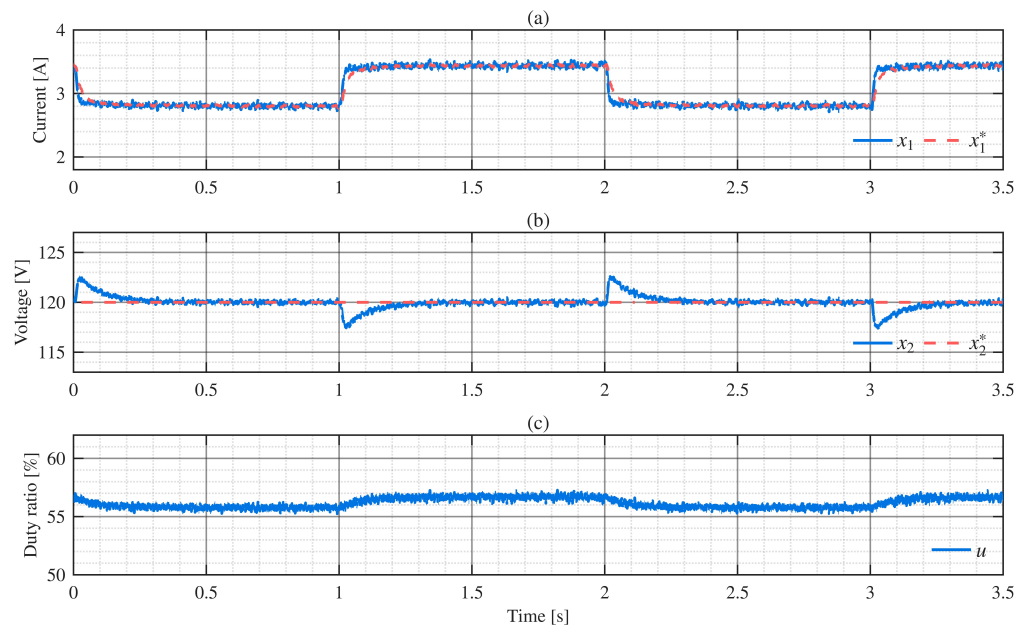


Figure 7. Experimental results for the case of changes in output load with controller dynamic extension. (a) Inductor current x_1 and reference current x_1^* , (b) output voltage x_2 and reference voltage x_2^* , (c) control signal u .

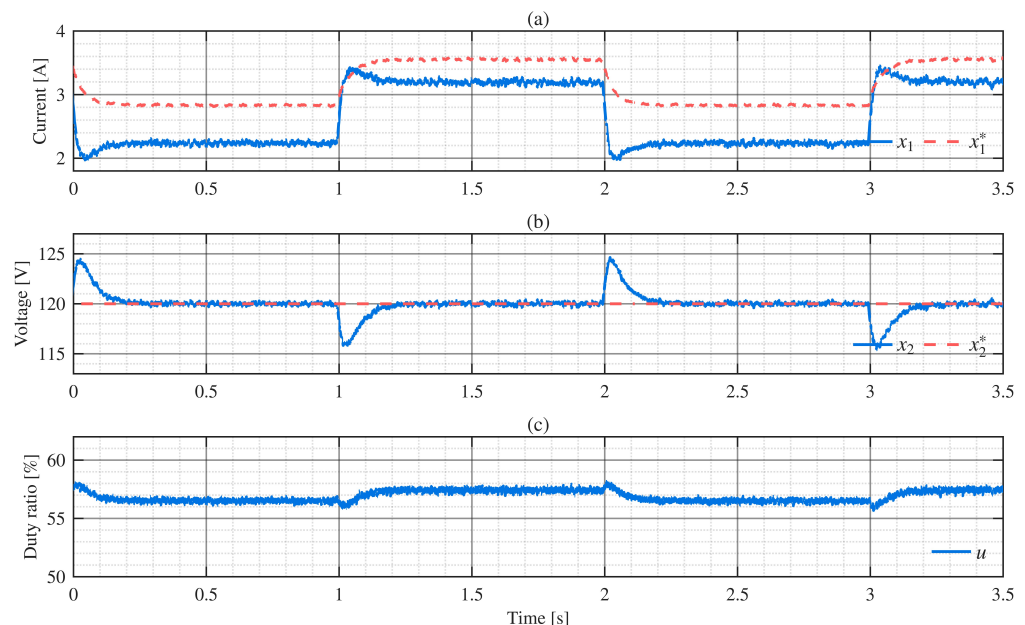


Figure 8. Experimental results for changes in output load without controller dynamic extension. (a) Inductor current x_1 and reference current x_1^* , (b) output voltage x_2 and reference voltage x_2^* , (c) control signal u .

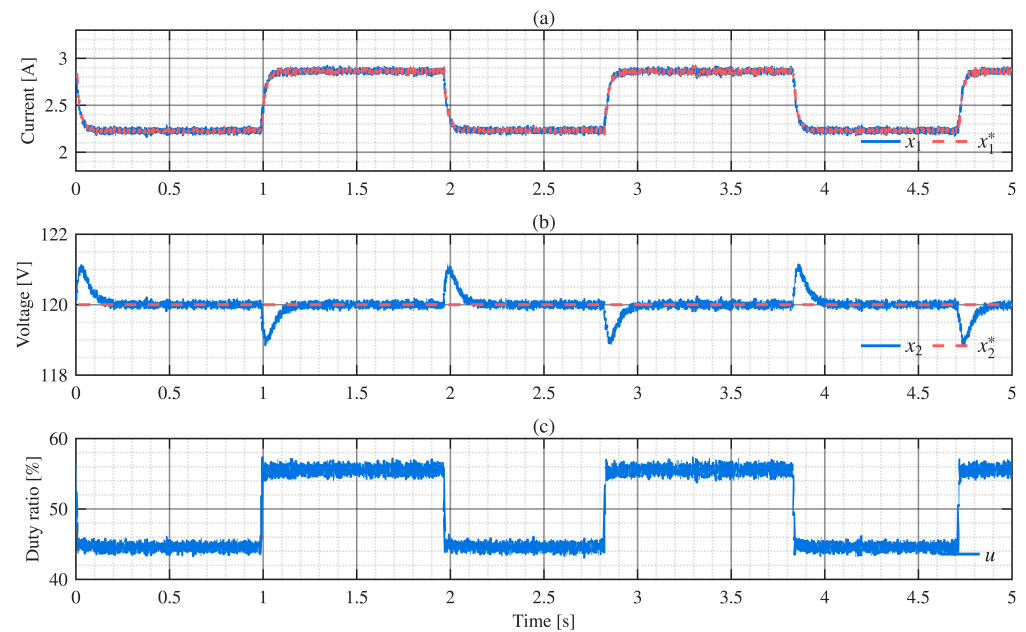


Figure 9. Experimental results for step changes in the input voltage with controller dynamic extension. (a) Inductor current x_1 and reference current x_1^* , (b) output voltage x_2 and reference voltage x_2^* , (c) control signal u .

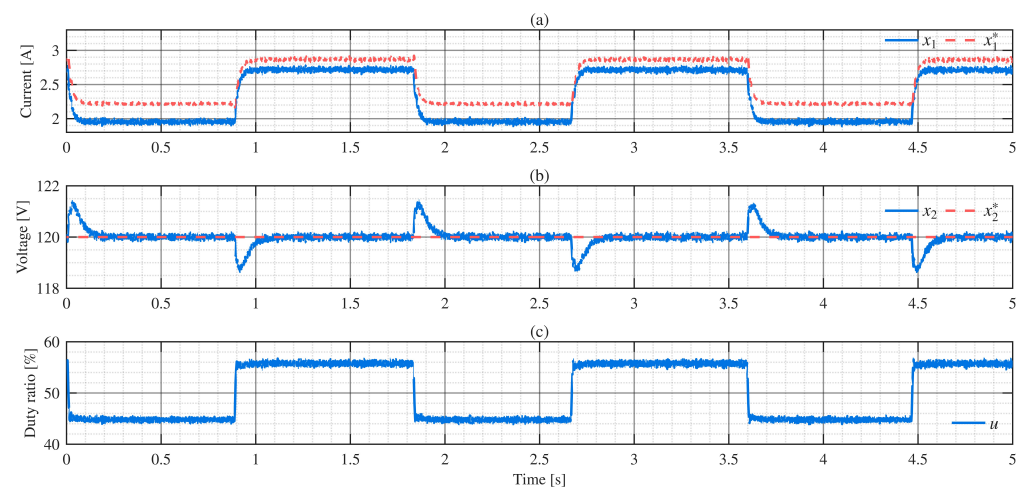


Figure 10. Experimental results for step changes in the input voltage without controller dynamic extension. (a) Inductor current x_1 and reference current x_1^* , (b) output voltage x_2 and reference voltage x_2^* , (c) control signal u .

4.8. Experimental Results: Reference Voltage Changes

Closed-loop experimental results considering changes in the output voltage reference x_2^* are shown in Figure 11. Again, the variations are performed in steps at a rate of 0.5 Hz from 110 V to 120 V. The output load is set at 126 Ω , and the resulting output power ranges from $P_o = 96.03$ W ($x_2^* = 110$ V) to $P_o = 114.28$ W ($x_2^* = 120$ V). In Figure 11a, tight current tracking is shown with no steady-state error. Afterward, tight voltage regulation is depicted in Figure 11b despite sudden variations in x_2^* . Hence, the control objectives are achieved. In addition, the control signal in Figure 11c displays changes between 55.93% ($x_2^* = 120$ V) and 52.46% ($x_2^* = 110$ V). In comparison, in Figure 12, the closed-loop system behavior without dynamic extension is observed for this case scenario. Figure 12a shows the steady-state error of the current, which is almost 0.2 A. Nevertheless, the voltage regulation illustrated in Figure 12b is preserved. Finally, the control signal is observed in Figure 12c, which presents changes between 55.93% and 52.46%.

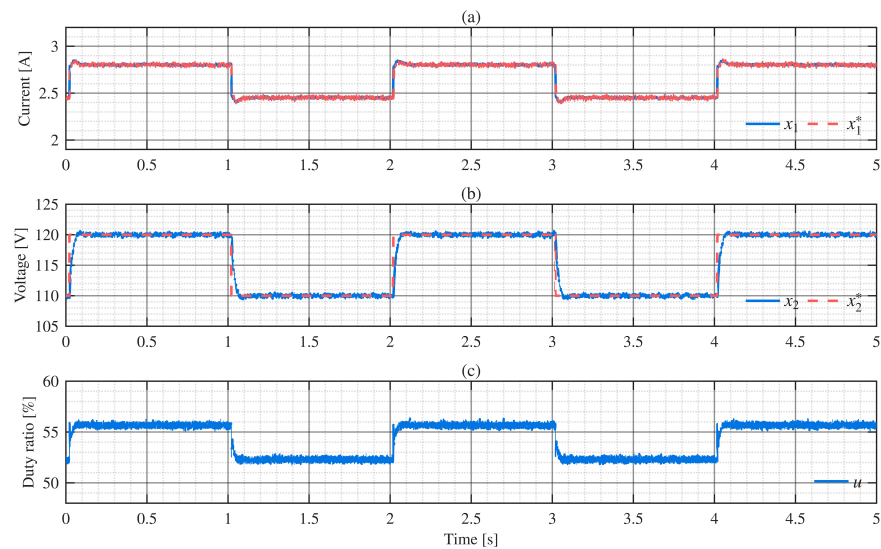


Figure 11. Experimental results for step changes in the voltage reference with controller dynamic extension. (a) Inductor current x_1 and reference current x_1^* , (b) output voltage x_2 and reference voltage x_2^* , (c) control signal u .

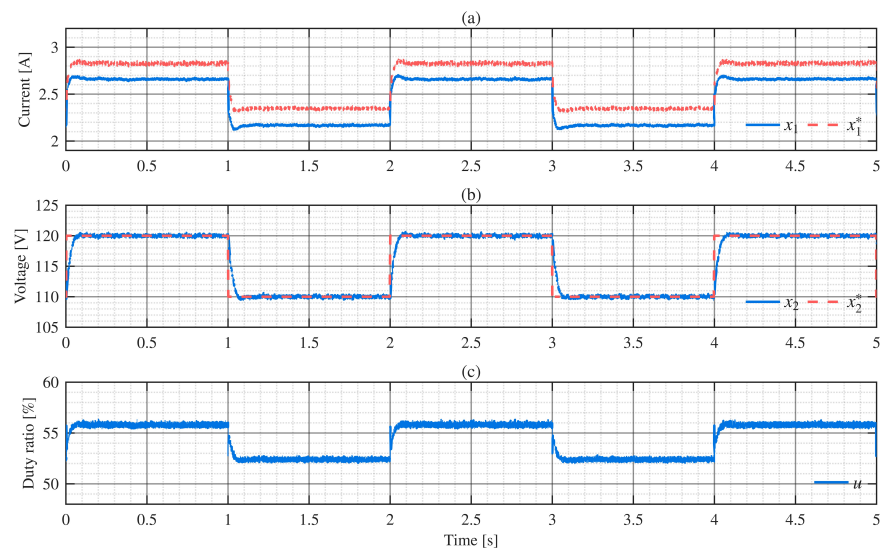


Figure 12. Experimental results for step changes in the voltage reference without controller dynamic extension. (a) Inductor current x_1 and reference current x_1^* , (b) output voltage x_2 and reference voltage x_2^* , (c) control signal u .

5. Conclusions and Future Work

This paper has presented the control strategy design for current tracking and voltage regulation in an ASL-BC. The proposed approach combined interconnection and damping assignment PBC with dynamic extension. The control strategy achieved high-performance tracking of the desired current and output voltage regulation, and no restrictions were imposed over the tuning gains nor singularities were present in the control law. Furthermore, this scheme ensured the asymptotic stability and passivity of the closed-loop system, whereas the dynamic extension improved tracking and regulation performance by eliminating the effect of the non-modeled parasitic resistances present in the system. Numerical and experimental results have shown fast and accurate tracking of the desired current and regulation of the output voltage, even in the presence of load changes, input voltage disturbances, and output voltage reference changes. For future work, the proposed control strategy can be extended to other step-up or step-down high-gain DC-DC converters to exploit renewable energy sources like FC or PV systems.

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Abbreviations

The following abbreviations are used in this manuscript:

IDA-PBC	Interconnection and damping assignment passivity-based control
DC	Direct current
BC	Boost converter
PV	Photovoltaic
FC	Fuel cell
ASL	Active switched inductor
CMC	Current-mode control
PI	Proportional–integral
EL	Euler–Lagrange
RCG	Reference current generator
PCH	Port-controlled Hamiltonian system
CCM	Continuous conduction mode
DCM	Discontinuous conduction mode
v_i	Input voltage
L_1, L_2, L	Converter inductors
S_{w1}, S_{w2}	Active switches
D	Diode
C	Capacitor
R_o	Nominal resistive load
x_1, i_L	Inductor current
x_2, v_C	Output capacitor voltage
u	Duty cycle, control signal
$\bar{u}$	Nominal duty cycle
$\bar{x}_1$	Inductor current eq. point
$\bar{x}_2$	Output capacitor voltage eq. point
$(\mathcal{J}_d), \mathcal{J}$	(Desired) Interconnection matrix
$(\mathcal{R}_d), \mathcal{R}$	(Desired) Dissipation matrix
$(\mathcal{H}_d), \mathcal{H}$	(Desired) Hamiltonian function
$\mathcal{G}$	Port characteristic matrix
ξ	Disturbance vector
x_1^*	Current reference
x_2^*	Voltage reference
α, β	Control gains
x_c	Integral action
κ_I	Integral gain
$(\tilde{\cdot})$	Error variable
W	Energy function

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