

## Article

# Non-Linear Control of a Triple Active Bridge Converter Used in a DC Microgrid with Multiples Buses

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## Abstract

This article presents a novel nonlinear control strategy for an active triple-bridge converter used in a direct current (DC) microgrid with multiple buses to interconnect its three constituent buses, which are comprised of generation systems and generalized loads. The proposed controller aims to regulate the voltage on two of the DC buses and manage the power exchanged between them in response to load and voltage changes. Unlike the existing literature, the control strategy is designed based on the generalized state-space averaged model of the converter, obtained from the delta equivalent circuit of the high-frequency transformer. The performance of the designed controller is validated through simulation results, demonstrating good performance under changes in generated power, load variations, and voltage fluctuations on both DC buses of the microgrid.

**Keywords:** triple active bridge; DC microgrid; generalized load; generalized space state averaged model; nonlinear control

## 1. Introduction

In recent decades, the increasing penetration of renewable energy sources, storage systems, and power electronics-based loads has driven the development of DC microgrids as an efficient alternative to traditional alternating current (AC) distribution systems [1–3]. This is primarily because DC microgrids allow for more direct integration of photovoltaic generators, batteries, and supercapacitors, reducing conversion stages and improving the overall efficiency and reliability of the system [4–7].

However, a critical challenge in DC microgrids arises from the widespread deployment of power electronic loads, such as tightly regulated converters and electric vehicle (EV) battery chargers, which typically exhibit constant power load (CPL) or constant current load (CCL) characteristics [8,9]. Specifically, CPLs possess a negative incremental impedance that can affect DC bus stability and induce oscillations throughout the microgrid [10,11]. These adverse effects are further exacerbated in multi-bus DC microgrids, where strong electrical coupling exists between subsystems [12].

In this context, electronic power converters play a fundamental role, as they are responsible for interconnecting the different subsystems, providing voltage regulation and power management [13,14]. Among the different topologies available, the Triple Active Bridge (TAB) DC–DC converter has aroused increasing interest in DC microgrid



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applications with multiple buses, thanks to its high power density, bidirectional power transfer capability, galvanic isolation, and the possibility of interconnecting three DC buses using a single high-frequency transformer (HFT) [15–17].

Despite these advantages, the Triple Active Bridge (TAB) converter exhibits highly nonlinear and coupled dynamics, as the power exchanged between its ports simultaneously depends on multiple control variables, typically the phase-shift angles between the active bridges. This cross-coupling hinders the independent regulation of each DC bus and poses a significant challenge for controller design, particularly under large-signal disturbances and varying load conditions. Consequently, the development of control strategies remains a research topic of great interest [16,18]. For this reason, accurate modeling of the TAB converter is essential for the synthesis of advanced control strategies. In particular, averaged modeling techniques based on the Generalized State Space Averaging (GSSA) method allow for capturing both DC and AC dynamics, as well as the coupling effects of the TAB converter, resulting in a reliable and highly precise model [19,20].

Regarding the design of controllers for the TAB DC-DC converter, several approaches based on linear control techniques have been proposed in the specialized literature. In [21], a Proportional–Integral (PI) controller designed from its linearized GSSA model is proposed and used to regulate the voltage at two of the converter’s ports, to which only resistive loads are connected. Linear decoupling control techniques have also been proposed, such as in [22], where a controller is designed to regulate the voltage at two of the converter’s ports using its simplified model, thus reducing the controller’s complexity. However, this strategy has only been validated with purely resistive loads.

Due to the inherent limitations of linear control strategies, such as performance degradation and stability issues during large-scale load excursions or set-point changes, as they are synthesized using linearized models, the literature has explored nonlinear control designs. These approaches directly address dynamic coupling and system nonlinearities, ensuring more precise regulation and enhanced robustness against load variations. In [23], a Sliding Mode Control (SMC) is designed to regulate voltage and balance total power, achieving satisfactory performance under voltage and load variations; however, the HFT current dynamics are neglected, which may compromise system stability. A voltage controller for a TAB DC-DC converter is presented in [24], employing an effective diagonal matrix decoupling strategy based on differential flatness control. While this method offers favorable dynamic performance under disturbances, it remains highly dependent on model accuracy and is validated solely with resistive loads. Recently, advanced strategies such as model-free prediction and fixed-time control have been proposed for AC microgrids [25,26], showing significant robustness. Furthermore, although AI-based controllers for DC-DC converters have been explored to handle complex nonlinearities [27], they often require extensive training data and high computational power. Similarly, while advanced mathematical approaches for high-gain bidirectional converters exist [28], they frequently focus on specific topologies. Finally, a Model Predictive Control (MPC) is proposed in [29], providing significant improvements in dynamic response and port decoupling; nevertheless, it entails a high computational burden, exhibits sensitivity to parameter uncertainty, and is evaluated only under resistive load conditions. Unlike the aforementioned AI-based, SMC, and MPC methods, the proposed Feedback Linearization (FL) controller achieves exact decoupling with lower mathematical complexity and no training overhead. This ensures exponential error reduction and a simple control strategy to implement in standard microcontrollers, avoiding SMC chattering problems and the high computational cost of MPC or AI approaches.

This paper proposes the design of a new nonlinear control strategy for a TAB DC-DC converter used to interconnect three buses in a DC MG, each consisting of different types of loads. The main contributions of this work are listed below:

- ◆ Unlike the approaches described in the literature, which mostly use the equivalent star model of the HFT, the proposed controller has been designed based on a delta ( $\Delta$ ) representation of the coupling inductances. This choice allows the formulation of the GSSA model, which captures the fundamental AC component of the currents and the DC component of the voltages, avoiding the need to calculate the neutral point voltage present in the HFT star equivalent. Thanks to this approach, the model is defined by the phase shifts of the first bridge to the other two, omitting the direct interaction between the second and third bridges. This significantly simplifies the control laws and reduces the computational load.
- ◆ The controller is designed using a nonlinear control technique, employing feedback linearization (FL), which allows the system to be controlled using linear controllers through a coordinate transformation and nonlinear feedback. To ensure system stability and incorporate the dynamics of the HFT currents, an internal control loop is designed for the real and imaginary components of the HFT's secondary currents, and an external control loop is designed to manage the power at the ports and thus regulate their voltages. Compared to conventional linear strategies, such as the one proposed in [21], and nonlinear control strategies, such as the SMC controller proposed in [30], this approach provides improved dynamic response and superior rejection of disturbances during load transients and setpoint variations.
- ◆ Unlike most proposals in the literature, the strategy proposed in this work is validated with a generalized load model, which consists of components of constant impedance, constant current, and constant power (ZIP load). This ensures the robustness of the control under complex and nonlinear load dynamics.

## 2. System Description

The system under study in this work is illustrated in Figure 1. It consists of a DC microgrid (MG) featuring three DC buses: one connected to DC energy sources and the remaining two connected to generalized ZIP loads. These MG buses are interconnected via a Triple Active Bridge (TAB) DC-DC converter, which provides galvanic isolation between the DC stages to ensure system safety. Furthermore, since the converter facilitates bidirectional power flow among the MG buses, the overall operational flexibility of the system is maximized.

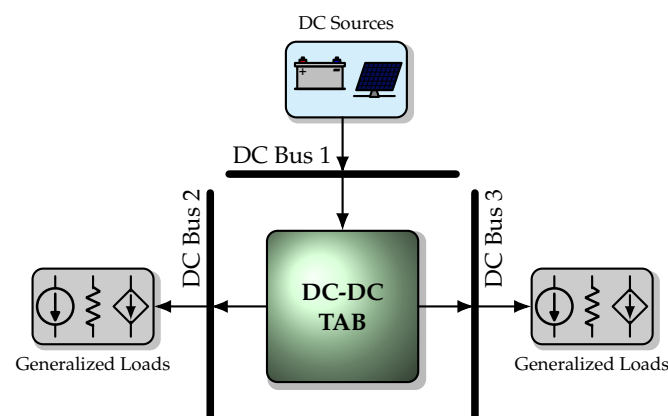


**Figure 1.** DC microgrid scheme with multiple buses interconnected by a TAB DC-DC converter.

In Figure 2, a simplified diagram of the DC-DC TAB converter considered in this work is shown. It is composed of three active bridges (AB-1, AB-2, and AB-3), (all in full-bridge configuration), whose activation signals are  $u_1$ ,  $u_2$  and  $u_3$ . On the other hand,  $C_1$ ,  $C_2$  and  $C_3$  are the DC filter capacitors for buses 1, 2 and 3 respectively,  $v_1$ ,  $v_2$  and  $v_3$  correspond to the voltages at each bus of the DC MG,  $i_{01}$  is the current of bus 1,  $i_{02}$  is the current in the second bus, and  $i_{03}$  corresponds to the current in the third DC bus of the MG. Finally,  $i_{p1}$ ,  $i_{p2}$  and  $i_{p3}$  are the input currents to each of the ports (AB-1, AB-2 and AB-3, respectively).

Furthermore, as observed in Figure 2, the three active bridges are interconnected through an AC link formed by an HFT, which features a primary winding with leakage inductance  $L_1$  and internal resistance  $R_1$ , and two secondary windings with leakage inductances  $L_2$  and  $L_3$  and internal resistances  $R_2$  and  $R_3$ , respectively. The currents  $i_{t1}$ ,  $i_{t2}$  and  $i_{t3}$  correspond to the primary current and the currents of the two secondary windings;  $v_{p1}$  is the voltage on the AC side of AB-1, while  $v_{p2}$  and  $v_{p3}$  are the voltages on the AC side of AB-2 and AB-3, respectively. Finally,  $n_{12}$ ,  $n_{13}$  and  $n_{23}$  represent the HFT turns ratios between ports 1, 2, and 3.

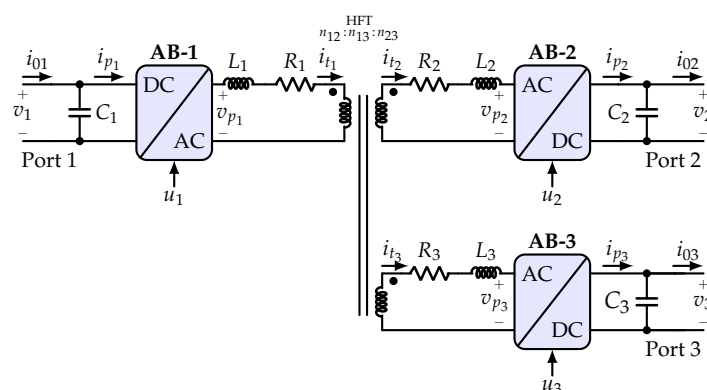


Figure 2. DC-DC TAB converter.

Figure 3 illustrates the switching signals  $u_1$ ,  $u_2$ , and  $u_3$  corresponding to the three active bridges of the TAB DC-DC converter (AB-1, AB-2, and AB-3, respectively). These signals take values in the set  $\{-1; 1\}$  and are out of phase with each other. Specifically, a phase shift  $\varphi_{12}$  exists between signals  $u_1$  and  $u_2$ , and a phase shift  $\varphi_{13}$  is present between  $u_1$  and  $u_3$ , whereas the phase shift between  $u_2$  and  $u_3$  is defined as  $\varphi_{23} = \varphi_{12} - \varphi_{13}$ .

Therefore, by shifting the switching signals of the three active bridges, it is possible to control the power transfer between the three converter ports and, thereby, regulate the voltage on DC buses 2 and 3.

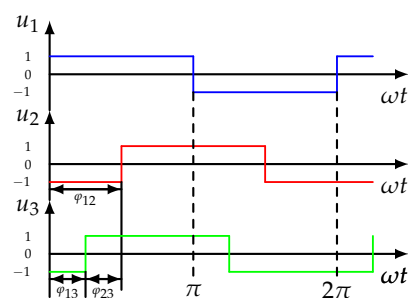


Figure 3. Switching signals for active bridges AB-1, AB-2 and AB-3, respectively.

To perform the mathematical modeling of the system considering both the DC bus voltage dynamics and the AC current dynamics in the HFT, it is first necessary to obtain an equivalent circuit of the TAB DC-DC converter that represents the cross-coupling between

ports produced by the HFT. To this end, this work utilizes the delta equivalent circuit of the TAB DC-DC converter shown in Figure 4, where the active bridges have been replaced by square wave voltage sources with voltages  $v_{p1} = u_1 v_1$ ,  $v_{p2} = n_{12} u_2 v_2$ , and  $v_{p3} = n_{13} u_3 v_3$ . On the other hand,  $i_{12}$ ,  $i_{13}$  and  $i_{23}$  are the currents between each of the ports, while  $L_{12}$ ,  $L_{13}$ ,  $L_{23}$ ,  $R_{12}$ ,  $R_{13}$  and  $R_{23}$  are the equivalent HFT inductances and resistances, which are given by the following relationships:

$$L_{ij} = \frac{L_i L_j + L_i L_k + L_j L_k}{L_k},$$

$$R_{ij} = \frac{R_i R_j + R_i R_k + R_j R_k}{R_k}, \quad (i, j, k = 1, 2, 3, \text{ with } i \neq j).$$
(1)

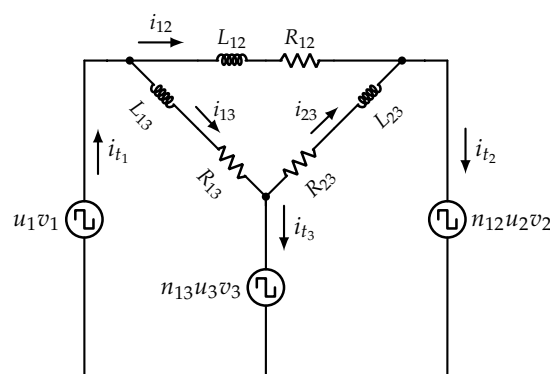


**Figure 4.** Delta equivalent circuit of DC-DC TAB converter.

On the other hand, the power delivered or absorbed by each port can be defined as:

$$p_1 = p_{12} + p_{13},$$

$$p_2 = p_{23} + p_{12},$$

$$p_3 = p_{13} + p_{23},$$
(2)

where  $p_{12}$  is the power transferred between port 1 and 2,  $p_{23}$  is the power between port 2 and 3, and  $p_{13}$  is the power transferred between port 1 and 3.

Next, the power transferred between ports is defined as:

$$p_{12} = v_2 i_{p1} = \frac{n_{12} v_1 v_2}{2 f_s L_{12}} \varphi_{12} (1 - \varphi_{12}),$$
(3)

$$p_{13} = v_3 i_{p3} = \frac{n_{13} v_1 v_3}{2 f_s L_{13}} \varphi_{13} (1 - \varphi_{13}),$$
(4)

$$p_{23} = v_3 i_{p3} = \frac{n_{23} v_2 v_3}{2 f_s L_{23}} \varphi_{23} (1 - \varphi_{23}),$$
(5)

where  $f_s$  is the switching frequency, while  $L_{12}$ ,  $L_{13}$  and  $L_{23}$  correspond to the equivalent HFT leakage inductances between such ports, which are given by (1).

### 3. Mathematical Modeling of the DC-DC Converter TAB

To design a control strategy that regulates the voltage on buses 2 and 3 of the MG to a constant value, a mathematical model representing the converter dynamic behavior is required. Initially, the switched-mode model will be obtained, and finally, it will be derived into the GSSA model of the TAB DC-DC converter, which will be used in the design of the control strategy proposed.

### 3.1. Switched Model

Using the schematic diagram of the TAB DC-DC converter in Figure 2, the switching signals in Figure 3 and the delta equivalent circuit in Figure 4, and under the consideration that the power transistors of the three active bridges are ideal, the dynamics of the currents  $i_{12}$ ,  $i_{13}$  and  $i_{23}$  and the voltages  $v_2$  and  $v_3$  can be obtained as follows:

$$\begin{aligned} \dot{i}_{12} &= \frac{1}{L_{12}}(-R_{12}i_{12} + u_1v_1 - n_{12}u_2v_2), \\ \dot{i}_{13} &= \frac{1}{L_{13}}(-R_{13}i_{13} + u_1v_1 - n_{13}u_3v_3), \\ \dot{i}_{23} &= \frac{1}{L_{23}}(-R_{23}i_{23} + u_3v_3 - n_{23}u_2v_2), \\ \dot{v}_2 &= \frac{1}{C_2}(i_{12} + i_{23} - i_{02}), \\ \dot{v}_3 &= \frac{1}{C_3}(i_{13} - i_{23} - i_{03}). \end{aligned} \quad (6)$$

Expressions (6) represent the switched model of the TAB DC-DC converter for any HFT turns ratio, formulated in terms of the dynamics of the currents  $i_{12}$ ,  $i_{13}$  and  $i_{23}$  arising from the HFT delta-equivalent circuit. Therefore, to obtain the currents in the three HFT windings, the HFT equivalent circuit shown in Figure 4 can be utilized:

$$i_{t1} = i_{12} + i_{13}, \quad i_{t2} = i_{12} + i_{23}, \quad i_{t3} = i_{13} - i_{23}. \quad (7)$$

Finally, assuming that the HFT has a unity turns ratio ( $n_{12} = n_{13} = n_{23} = 1$ ), and substituting the relationships from (7) into (6), the switched model in terms of the currents of the three HFT windings and the voltages at ports 2 and 3 results in:

$$\dot{i}_{t1} = -\frac{R_t}{L_t}i_{t1} + \frac{2}{L_t}u_1v_1 - \frac{1}{L_t}u_2v_2 - \frac{1}{L_t}u_3v_3, \quad (8)$$

$$\dot{i}_{t2} = -\frac{R_t}{L_t}i_{t2} + \frac{1}{L_t}u_1v_1 - \frac{2}{L_t}u_2v_2 + \frac{1}{L_t}u_3v_3, \quad (9)$$

$$\dot{i}_{t3} = -\frac{R_t}{L_t}i_{t3} + \frac{1}{L_t}u_1v_1 - \frac{1}{L_t}u_2v_2 - \frac{2}{L_t}u_3v_3, \quad (10)$$

$$\dot{v}_2 = \frac{1}{C_2}(u_2i_{t2} - i_{02}), \quad (11)$$

$$\dot{v}_3 = \frac{1}{C_3}(u_3i_{t3} - i_{03}), \quad (12)$$

where  $R_t = R_{12} = R_{13} = R_{23}$  and  $L_t = L_{12} = L_{13} = L_{23}$  result from considering that the HFT has a unity turns ratio.

### 3.2. Generalized State-Space Averaged Model

The model given by (8)–(12) obtained in the previous subsection accurately describes the switched behavior of the AC stage currents ( $i_{t1}$ ,  $i_{t2}$  and  $i_{t3}$ ) and the DC voltages ( $v_2$  and  $v_3$ ). However, the control strategy proposed in this work requires an averaged model of the converter that captures the system dynamics without explicitly relying on the switching variables.

Furthermore, because this topology includes an AC stage, classical averaging techniques are unsuitable, as they only consider the DC component of the system variables, ignoring their AC components. Therefore, the GSSA modeling strategy is adopted, which allows the AC variables of the system to be represented using a Fourier series expansion.

sion, resulting in a dynamic averaged model that preserves the AC information of the HFT currents [31].

The GSSA modeling strategy is based on the complex representation of Fourier series of state variables in the interval  $t - T \leq \tau \leq t$ , as described below:

$$x(t) = \sum_{k=-\infty}^{\infty} x_k(t) e^{jk\omega t}, \quad (13)$$

where  $\omega$  is the angular switching frequency, while  $x_k(t)$  represents the Fourier coefficient associated with the  $k$ -th harmonic of the AC variables. This coefficient can be interpreted as a moving average over a switching period, making it useful for obtaining the average value of the AC variables present in the DC-DC TAB converter, and it is defined as follows:

$$x_k(t) = \frac{1}{T} \int_{t-T}^t x(\tau) e^{-jk\omega\tau} d\tau. \quad (14)$$

The GSSA modeling strategy has two fundamental properties. The first establishes the derivative of the moving average given by (14), and the second defines the average of the  $k$ -th harmonic of a product between two variables [31]. These are given by

$$\begin{aligned} \frac{d}{dt} \langle x \rangle_k &= \left\langle \frac{d}{dt} x \right\rangle_k(t) - jk\omega \langle x \rangle_k, \\ \langle xy \rangle_k(t) &= \sum_i \langle x \rangle_{k-i} \langle y \rangle_i. \end{aligned} \quad (15)$$

To obtain the GSSA model of the TAB DC-DC converter, in this work, only the component ( $k = 0$ ) is considered for the DC variables, and the component ( $k = \pm 1$ ) for the AC variables.

Therefore, after applying properties (13), (14), and (15) to the commutated model given by (8)–(12), the GSSA model of the TAB DC-DC converter is obtained as follows:

$$\dot{x}_1 = -\frac{R_t}{L_t} x_1 + \omega x_2 - \frac{2}{L_t} \mu_1 x_5 + \frac{1}{L_t} \mu_3 x_6, \quad (16)$$

$$\dot{x}_2 = -\frac{R_t}{L_t} x_2 - \omega x_1 - \frac{2}{\pi L_t} v_1 - \frac{2}{L_t} \mu_2 x_5 + \frac{1}{L_t} \mu_4 x_6, \quad (17)$$

$$\dot{x}_3 = -\frac{R_t}{L_t} x_3 + \omega x_4 - \frac{2}{L_t} \mu_3 x_6 + \frac{1}{L_t} \mu_1 x_5, \quad (18)$$

$$\dot{x}_4 = -\frac{R_t}{L_t} x_4 - \omega x_3 - \frac{2}{\pi L_t} v_1 + \frac{1}{L_t} \mu_2 x_5 - \frac{2}{L_t} \mu_4 x_6, \quad (19)$$

$$\dot{x}_5 = \frac{1}{C_2} (2(\mu_1 x_1 + \mu_2 x_2) - i_{02}), \quad (20)$$

$$\dot{x}_6 = \frac{1}{C_3} (2(\mu_3 x_3 + \mu_4 x_4) - i_{03}), \quad (21)$$

where

$$\begin{aligned} x_1 &= \langle i_{t1} \rangle_{1R}, \quad x_2 = \langle i_{t1} \rangle_{1I}, \quad x_3 = \langle i_{t2} \rangle_{1R}, \quad x_4 = \langle i_{t2} \rangle_{1I}, \quad x_5 = \langle v_2 \rangle_0, \\ x_6 &= \langle v_3 \rangle_0, \quad v_1 = \langle v_1 \rangle_0, \quad i_{02} = \langle i_{02} \rangle_0, \quad i_{03} = \langle i_{03} \rangle_0, \quad \langle u_1 \rangle_{1R} = 0, \\ \langle u_1 \rangle_{1I} &= -\frac{2}{\pi}, \quad \mu_1 = \langle u_2 \rangle_{1R} = -\frac{2}{\pi} \sin(\pi\varphi_{12}), \quad \mu_2 = \langle u_2 \rangle_{1I} = -\frac{2}{\pi} \cos(\pi\varphi_{12}), \\ \mu_3 &= \langle u_3 \rangle_{1R} = -\frac{2}{\pi} \sin(\pi\varphi_{13}), \quad \mu_4 = \langle u_3 \rangle_{1I} = -\frac{2}{\pi} \cos(\pi\varphi_{13}). \end{aligned} \quad (22)$$

At the same time, the subscript “0” denotes the DC component, and the subscripts ‘1R’ and “1I” represent the real and imaginary parts of the fundamental AC component.

It is necessary to clarify that, from a physical perspective, the state variables  $x_1$  to  $x_4$  represent the evolution of the high-frequency current components within the transformer’s leakage inductances. Meanwhile,  $x_5$  and  $x_6$  describe the average dynamics of the DC link voltages. The coupling terms in the model reflect the physical power flow between ports, where the phase angles  $\varphi_{12}$  and  $\varphi_{13}$  allow for regulating the power transferred between the converter’s ports through magnetic coupling.

Next, the reference values of the system variables are obtained. To do this, expressions (16)–(21) are evaluated at their equilibrium point, ignoring the effect of  $R_t$  since, in practice, it results in a value close to zero, thus obtaining the following:

$$\begin{aligned} x_1^* &= \frac{1}{\omega L_t} \left( -2\mu_2^* x_5^* + \mu_4^* x_6^* - \frac{2v_1}{\pi} \right), \quad x_3^* = \frac{1}{\omega L_t} \left( \mu_2^* x_5^* - \mu_4^* x_6^* - \frac{2v_1}{\pi} \right), \\ \mu_1^* &= -\frac{2}{\pi} \sin(\pi\varphi_{12}^*), \quad \mu_2^* = -\frac{2}{\pi} \cos(\pi\varphi_{12}^*), \quad \mu_3^* = -\frac{2}{\pi} \sin(\pi\varphi_{13}^*), \quad \mu_4^* = -\frac{2}{\pi} \cos(\pi\varphi_{13}^*), \end{aligned} \quad (23)$$

where  $x_5^*$  and  $x_6^*$  are the desired values of the voltages at ports 2 and 3, respectively. On the other hand, the phase shifts at the operating point can be obtained from (3) and (4), assuming a unity transformation ratio and that the converter is in steady state, resulting in:

$$\varphi_{12}^* = \frac{1}{2} \left( 1 - \sqrt{1 - \frac{8f_s L_t i_{02}^e}{v_1}} \right), \quad \varphi_{13}^* = \frac{1}{2} \left( 1 - \sqrt{1 - \frac{8f_s L_t i_{03}^e}{v_1}} \right), \quad (24)$$

where  $i_{02}^e$  and  $i_{03}^e$  are the output currents at the equilibrium point for ports 2 and 3, respectively.

It should be clarified that the reference values  $x_2^*$  and  $x_4^*$  will be obtained from the external control loop, which will be designed in the next section.

## 4. Control Strategy

The following section presents the design of the control strategy proposed in this work. The objective of the designed controller is to regulate the voltage of ports 2 and 3 to their respective reference values, to which generalized ZIP loads are connected.

### 4.1. System Assumptions and Practical Impact

To facilitate the controller design, the following assumptions are considered:

- **Component Ideality:** Losses in switches and the high-frequency transformer (HFT) are neglected in the control model. In practice, these act as small disturbances that the proposed FL controller compensates for through its active feedforward and integral action.
- **Constant Parameters:** Inductances and capacitances are assumed to be at their nominal values. Real-world variations (e.g., due to temperature) are handled by the integral term in the control laws, ensuring zero steady-state error.
- **Ideal Sensing:** Measurements of voltages and currents are assumed to be noise-free. Real implementations require low-pass filters in the Analog-to-Digital Converter (ADC) stages; the stability margins of the proposed control are robust enough to handle the minor delays introduced by these filters.

### 4.2. Proposed Controller Design

Because the system described by (16)–(21) is nonlinear in nature, given that it has products between state, control, and measurement variables, this paper proposes the design of a nonlinear control strategy based on FL. This choice is due to the fact that, thanks to the

coordinate transformation and nonlinear feedback provided by this control technique, it is possible to control a nonlinear system with a linear controller [25,32].

To ensure system stability and avoid unstable zero dynamics, an indirect control scheme is designed. It consists of an external control loop, which regulates the voltage at ports 2 and 3, and an internal control loop, which regulates the real and imaginary parts of the currents of both HFT secondaries, as shown in Figure 5.

To do this, the following output vector is defined:

$$\mathbf{y} = \begin{bmatrix} z_1 & z_2 & z_3 & z_4 \end{bmatrix}^T = \begin{bmatrix} x_1 & x_2 & x_3 & x_4 \end{bmatrix}^T. \quad (25)$$

By deriving the output vector (25) until the inputs appear explicitly, we obtain a system of equations described in the new coordinates,

$$\begin{bmatrix} \dot{z}_1 \\ \dot{z}_2 \\ \dot{z}_3 \\ \dot{z}_4 \end{bmatrix} = \underbrace{\begin{bmatrix} -\frac{R_t}{L_t}x_1 + \omega x_2 \\ -\frac{R_t}{L_t}x_2 - \omega x_1 - \frac{2}{\pi L_t}v_1 \\ -\frac{R_t}{L_t}x_3 + \omega x_4 \\ -\frac{R_t}{L_t}x_4 - \omega x_3 - \frac{2}{\pi L_t}v_1 \end{bmatrix}}_{\mathbf{A}(\mathbf{x})} + \underbrace{\begin{bmatrix} -\frac{2}{L_t}x_5 & 0 & \frac{1}{L_t}x_6 & 0 \\ 0 & -\frac{2}{L_t}x_5 & 0 & \frac{1}{L_t}x_6 \\ \frac{1}{L_t}x_5 & 0 & -\frac{2}{L_t}x_6 & 0 \\ 0 & \frac{1}{L_t}x_5 & 0 & -\frac{2}{L_t}x_6 \end{bmatrix}}_{\mathbf{E}(\mathbf{x})} \begin{bmatrix} \mu_1 \\ \mu_2 \\ \mu_3 \\ \mu_4 \end{bmatrix}. \quad (26)$$

In this way, the dynamics described in the new coordinates can be represented by a linear system in terms of the auxiliary control inputs ( $\gamma_1$ ,  $\gamma_2$ ,  $\gamma_3$ , and  $\gamma_4$ ), which can be designed using linear design tools. These are defined as:

$$\begin{bmatrix} \dot{z}_1 & \dot{z}_2 & \dot{z}_3 & \dot{z}_4 \end{bmatrix}^T = \begin{bmatrix} \gamma_1 & \gamma_2 & \gamma_3 & \gamma_4 \end{bmatrix}^T = \boldsymbol{\gamma}. \quad (27)$$

Next, if the matrix  $\mathbf{E}(\mathbf{x})$  in (26) is invertible, the control inputs can be obtained by

$$\begin{bmatrix} \mu_1 & \mu_2 & \mu_3 & \mu_4 \end{bmatrix}^T = \mathbf{E}(\mathbf{x})^{-1}(\boldsymbol{\gamma} - \mathbf{A}(\mathbf{x})). \quad (28)$$

Therefore, since  $|\mathbf{E}(\mathbf{x})| = \frac{9x_5^2x_6^2}{L_t^4} \neq 0$ , because  $x_5 > 0$  and  $x_6 > 0$ , it is possible to obtain the control inputs using (26) since the matrix  $\mathbf{E}(\mathbf{x})$  has an inverse. So, from (26) and (28), the control inputs  $\mu_1$ ,  $\mu_2$ ,  $\mu_3$ , and  $\mu_4$  are obtained as follows:

$$\mu_1 = -\frac{1}{3x_5} \left( 2\gamma_1 + \frac{R_t}{L_t}(2x_1 - x_3) - \omega(2x_2 + x_4) + \gamma_3 \right), \quad (29)$$

$$\mu_2 = -\frac{1}{3x_5} \left( 2\gamma_2 + \frac{R_t}{L_t}(2x_2 + x_4) + \omega(2x_1 + x_3) + \gamma_4 + \frac{6}{\pi L_t}v_1 \right), \quad (30)$$

$$\mu_3 = -\frac{1}{3x_6} \left( \gamma_1 + \frac{R_t}{L_t}(x_1 + 2x_3) - \omega(x_2 + 2x_4) + 2\gamma_3 \right), \quad (31)$$

$$\mu_4 = -\frac{1}{3x_6} \left( \gamma_2 + \frac{R_t}{L_t}(x_2 + 2x_4) + \omega(x_1 + 2x_3) + 2\gamma_4 + \frac{6}{\pi L_t}v_1 \right). \quad (32)$$

Expressions (29) and (30) correspond to the real and imaginary parts of the phase shift between the first and second active bridges ( $\varphi_{12}$ ), while (31) and (32) represent the real and imaginary parts of the phase shift between the first and third active bridges ( $\varphi_{13}$ ).

#### 4.3. External Control Loop Design

Since the purpose of the control strategy designed is to regulate the voltages at the second and third ports of the converter, an external control loop is designed to control the voltage  $v_2$  ( $x_5$ ) and  $v_3$  ( $x_6$ ).

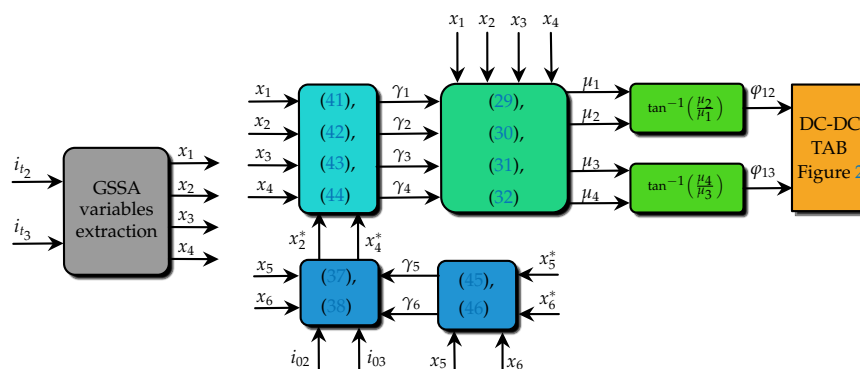


Figure 5. Control Scheme proposed.

Initially, it is considered to have a settling time ten times slower than the internal control loop. Therefore, it can be assumed that  $x_1 = x_1^*$ ,  $x_2 = x_2^*$ ,  $x_3 = x_3^*$ ,  $x_4 = x_4^*$ ,  $\mu_1 = \mu_1^*$ ,  $\mu_2 = \mu_2^*$ ,  $\mu_3 = \mu_3^*$ , and  $\mu_4 = \mu_4^*$ .

Next, for the design of the external loop, the energy at ports 2 and 3 are selected as outputs, respectively:

$$\begin{aligned} z_5 &= \frac{1}{2} C_2 x_5^5, \\ z_6 &= \frac{1}{2} C_3 x_6^2. \end{aligned} \quad (33)$$

After deriving expressions (33), we obtain

$$\dot{z}_5 = C_2 \dot{x}_5 x_5 = 2x_5(\mu_1^* x_1^* + \mu_2^* x_2^*) - i_{02} x_5, \quad (34)$$

$$\dot{z}_6 = C_3 \dot{x}_6 x_6 = 2x_6(\mu_3^* x_3^* + \mu_4^* x_4^*) - i_{03} x_6. \quad (35)$$

Equating (34) and (35) with two new auxiliary control inputs

$$\begin{bmatrix} \dot{z}_5 & \dot{z}_6 \end{bmatrix}^T = \begin{bmatrix} \gamma_5 & \gamma_6 \end{bmatrix}^T, \quad (36)$$

and after replacing (36) in (34) and (35), the references for the internal control loop  $x_2^*$  and  $x_4^*$  are obtained

$$x_2^* = \frac{1}{\mu_2^*} \left( \frac{1}{2} \left( \frac{1}{x_5} \gamma_5 + i_{02} \right) - \mu_1^* x_1^* \right), \quad (37)$$

$$x_4^* = \frac{1}{\mu_4^*} \left( \frac{1}{2} \left( \frac{1}{x_6} \gamma_6 + i_{03} \right) - \mu_3^* x_3^* \right). \quad (38)$$

#### 4.4. Design of Control Parameters

In the following subsection, the auxiliary control inputs given by (27) and (36) are designed. To do this, the errors of the transformed variables are initially defined as

$$\begin{aligned} e_1 &= z_1 - z_1^* = x_1 - x_1^*, \quad e_2 = z_2 - z_2^* = x_2 - x_2^*, \\ e_3 &= z_3 - z_3^* = x_3 - x_3^*, \quad e_4 = z_4 - z_4^* = x_4 - x_4^*, \\ e_5 &= z_5 - z_5^* = \frac{1}{2} C_2 (x_5^2 - x_5^{*2}), \quad e_6 = z_6 - z_6^* = \frac{1}{2} C_3 (x_6^2 - x_6^{*2}), \end{aligned} \quad (39)$$

and its dynamics are assigned as follows:

$$\begin{aligned} \dot{e}_1 + k_{p1} e_1 &= 0, \quad \dot{e}_2 + k_{p2} e_2 = 0, \quad \dot{e}_3 + k_{p3} e_3 = 0, \\ \dot{e}_4 + k_{p4} e_4 &= 0, \quad \dot{e}_5 + k_{p5} e_5 = 0, \quad \dot{e}_6 + k_{p6} e_6 = 0. \end{aligned} \quad (40)$$

After replacing (27) and (36) in (40), the auxiliary control inputs are obtained as follows:

$$\gamma_1 = -k_{p_1}e_1, \quad (41)$$

$$\gamma_2 = -k_{p_2}e_2, \quad (42)$$

$$\gamma_3 = -k_{p_3}e_3, \quad (43)$$

$$\gamma_4 = -k_{p_4}e_4, \quad (44)$$

$$\gamma_5 = -k_{p_5}e_5 - k_{i_1} \int e_5 dt, \quad (45)$$

$$\gamma_6 = -k_{p_6}e_6 - k_{i_2} \int e_6 dt, \quad (46)$$

where  $k_{p_j}$  (with  $j = 1, 2, 3, 4, 5, 6$ ) are proportional gains that allow the desired dynamic response to be achieved. On the other hand,  $k_{i_1}$  and  $k_{i_2}$  are integral gains added to eliminate the steady-state error in the voltages of ports 2 and 3 that may occur due to parameter variations.

From a physical perspective, the control variables  $\gamma_5$  and  $\gamma_6$  represent linear controllers, the proportional term provides an immediate reaction to voltage deviations caused by power mismatches between ports, while the integral term accounts for the physical losses of the converter and the average power demand of the loads, ensuring zero steady-state error even under the negative incremental impedance effect of CPLs.

The proportional gains of the internal control loop ( $k_{p_1}$ ,  $k_{p_2}$ ,  $k_{p_3}$ , and  $k_{p_4}$ ) are obtained after solving the first four differential equations of (40) and considering a criterion of 2% in the response, resulting in:

$$k_{p_1} = k_{p_2} = k_{p_3} = k_{p_4} = \frac{4}{t_{s_1}}, \quad (47)$$

where  $t_{s_1}$  is the settling time for the real and imaginary parts of the HFT secondary currents (inner loop).

On the other hand, the proportional and integral gains of the external controller ( $x_5$  and  $x_6$ ) are obtained after adding control laws (45) and (46) to (36) and applying the Laplace transform, resulting in,

$$k_{p_5} = \frac{8C_2}{t_{s_2}}, \quad k_{p_6} = \frac{8C_3}{t_{s_2}}, \quad k_{i_5} = \omega_n^2 C_2, \quad k_{i_6} = \omega_n^2 C_3, \quad (48)$$

where  $t_{s_2}$  is the settling time for stresses  $x_5$  and  $x_6$ , while  $\omega_n$  is the natural frequency whose value is set at  $\omega_n = 2.54 \times 10^3$  rad/s, to achieve a response in stresses with a maximum overshoot of 2%.

Considering the parameters in Table 1, and that the internal loop settling time is 0.2 ms, while that of the external loop is 2 ms, the values for the gains shown in Table 2 are obtained.

**Table 1.** System parameters.

| Converter Parameters    |       |
|-------------------------|-------|
| Description             | Value |
| Input voltage, $v_1$    | 250 V |
| Port 2 voltage, $x_5^*$ | 120 V |
| Port 3 voltage, $x_6^*$ | 120 V |

Table 1. Cont.

| Converter Parameters                        |              |
|---------------------------------------------|--------------|
| Description                                 | Value        |
| HFT transformation ratio, $1:n_{12}:n_{13}$ | 1:1:1        |
| Leakage inductance, $L_1 = L_2 = L_3$       | 14 $\mu$ H   |
| Internal resistance, $R_1 = R_2 = R_3$      | 0.2 $\Omega$ |
| Port 1 capacitance, $C_1$                   | 470 $\mu$ F  |
| Port 2 capacitance, $C_2$                   | 470 $\mu$ F  |
| Port 3 capacitance, $C_3$                   | 470 $\mu$ F  |
| Switching frequency, $f_s$                  | 20 kHz       |
| Load parameters at port 2                   |              |
| Description                                 | Value        |
| Resistive load, $R_2$                       | 15 $\Omega$  |
| Constant power load, $P_{CPL_2}$            | 1.25 kW      |
| Constant current load $I_2$                 | 4 A          |
| Load parameters at port 3                   |              |
| Description                                 | Value        |
| Resistive load, $R_3$                       | 15 $\Omega$  |
| Constant power load, $P_{CPL_3}$            | 1.4 kW       |
| Constant current load $I_3$                 | 4 A          |

Table 2. Controller parameters.

| Gain                                    | Value            |
|-----------------------------------------|------------------|
| $k_{p_1} = k_{p_2} = k_{p_3} = k_{p_4}$ | $20 \times 10^3$ |
| $k_{p_5} = k_{p_6}$                     | 1.88             |
| $k_{i_5} = k_{i_6}$                     | 1.19             |

Finally, Figure 5 shows the control scheme of the proposed controller. Initially, the real and imaginary components of the currents of both HFT secondaries are extracted and used to obtain the auxiliary control actions ( $\gamma_1$ ,  $\gamma_2$ ,  $\gamma_3$  and  $\gamma_4$ ). On the other hand, from the measurement of the voltages  $x_5$  and  $x_6$  and their references, the auxiliary control actions  $\gamma_5$  and  $\gamma_6$  used by the external control loop to obtain the references  $x_2^*$  and  $x_4^*$  are obtained. Then, the control actions  $\mu_1$ ,  $\mu_2$ ,  $\mu_3$ , and  $\mu_4$  are calculated, from which the phase shifts  $\varphi_{12}$  and  $\varphi_{13}$  are obtained, which are applied to the switching signals of the active bridges of the TAB DC-DC converter in Figure 2.

Regarding the practical implementation of the control scheme shown in Figure 5, the GSSA variable extraction block is implemented using the filtering scheme described in [33], which allows for real-time estimation of the fundamental components with low computational overhead. The overall control law is executed at a fixed sampling frequency, where the gains were designed with a bandwidth separated from the HFT switching frequency to avoid instability. This implementation ensures that the complex GSSA mathematical framework is reduced to a set of direct arithmetic operations suitable for standard digital signal processors (DSPs).

#### 4.5. Stability Analysis for Zero Dynamics

Since the relative vector degree of the output vector (27) is  $r = 4$  and the order of the system given by (16)–(21) is  $n = 6$ , the resulting dynamic has zeros given by (20) and (21), whose stability must be verified [32]. To do this, we consider that the system is in a stable

state, so we assume that  $x_1 = x_1^*$ ,  $x_2 = x_2^*$ ,  $x_3 = x_3^*$ ,  $x_4 = x_4^*$ ,  $\mu_1 = \mu_1^*$ ,  $\mu_2 = \mu_2^*$ ,  $\mu_3 = \mu_3^*$ , and  $\mu_4 = \mu_4^*$ . Therefore, after replacing the references (37) and (38) in (20) and (21), we obtain

$$\dot{x}_5 = \frac{1}{C_2 x_5} \gamma_5, \quad (49)$$

$$\dot{x}_6 = \frac{1}{C_3 x_6} \gamma_6. \quad (50)$$

Using the controller and converter parameters given by Tables 1 and 2, the phase portraits in Figure 6 corresponding to the dynamics of the zeros of (49) and (50) respectively are obtained.

In both cases, it can be observed that a single minimum phase equilibrium point is obtained. Therefore, the controller proposed in this work ensures that the closed-loop system is stable.

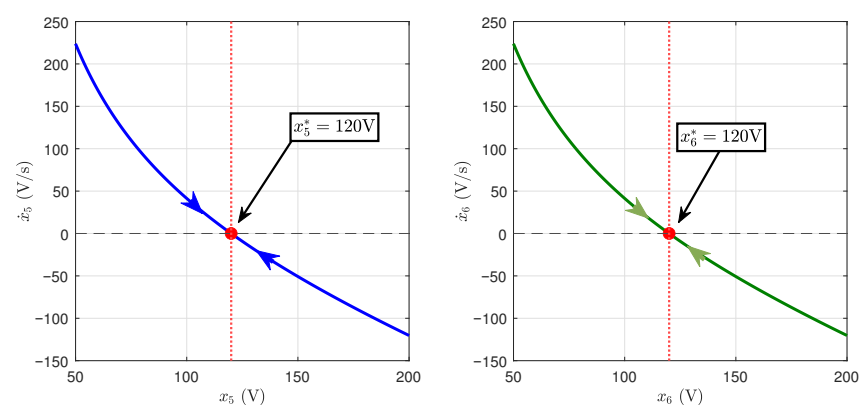


**Figure 6.** Phase portrait for the dynamics of the zeros. For (49) (left) and for (50) (right).

It is necessary to clarify that the proposed FL-based controller is inherently robust against the destabilizing effects of constant power loads (CPLs) since, as shown in the control laws (45) and (46), the terms  $i_{02}$  and  $i_{03}$  act as an active feedforward compensation. This algebraic cancellation of the load dynamics ensures that the closed-loop error remains governed by (45) and (46). Consequently, the stability of the converter is independent of the load power level, maintaining a constant and robust response even under severe CPL conditions.

## 5. Simulation Results

To validate the performance of the designed controller, simulation tests were performed with *Matlab R2022a Simulink* using the *SimPowerSystems* library and the *Runge–Kutta* simulation method with a fixed step size of  $T = 0.16 \mu\text{s}$ . MOSFET transistors with an internal resistance  $R_{DS} = 25 \text{ m}\Omega$  were used for the three active bridges, and a real transformer model was used for the HFT in order to achieve a realistic simulation scenario.

The controller parameters are shown in Table 2, while the system parameters are shown in Table 1.

The simulation tests were performed on the system shown in Figure 2 using the control scheme shown in Figure 5.

To validate the performance and robustness of the proposed controller in the face of reference changes and variations in input voltage and load disturbances, four simulation tests were carried out. In the first, reference changes and variations in input voltage are made; in the second, variations in connected loads occur; and in the third, the performance of the proposed strategy is compared with a PI and SMC controller in response to reference

and load changes. Finally, a fourth test was performed, where variations were made to the converter parameters, with the aim of validating the robustness of the proposed controller.

### 5.1. Reference Changes and Variations in Input Voltage

In this 110 ms test, a reference change was made from 120 V to 130 V at  $t = 10$  ms on voltage  $v_2$ , which was returned to 120 V at  $t = 30$  ms. Then, at  $t = 50$  ms, the voltage  $v_3$  was increased from 120 V to 130 V and returned to 120 V at  $t = 70$  ms. Finally, at  $t = 90$  ms, the input voltage ( $v_1$ ) was increased from 250 V to 260 V and returned to its value of 250 V at 100 ms.

Figure 7a shows the voltages at ports 2 and 3, where good tracking of the references can be observed with a slightly underdamped response that settles within 5% of the final value in the required settling time. Given the variations in the input voltage  $v_1$  ( $t = 90$  ms to 100 ms), both output voltages show small deviations ( $\pm 0.75$  V) from their reference values. Furthermore, it can be observed that when one of the output voltages changes, the voltage of the remaining port does not vary significantly (around  $\pm 1$  V), demonstrating the decoupling between ports.

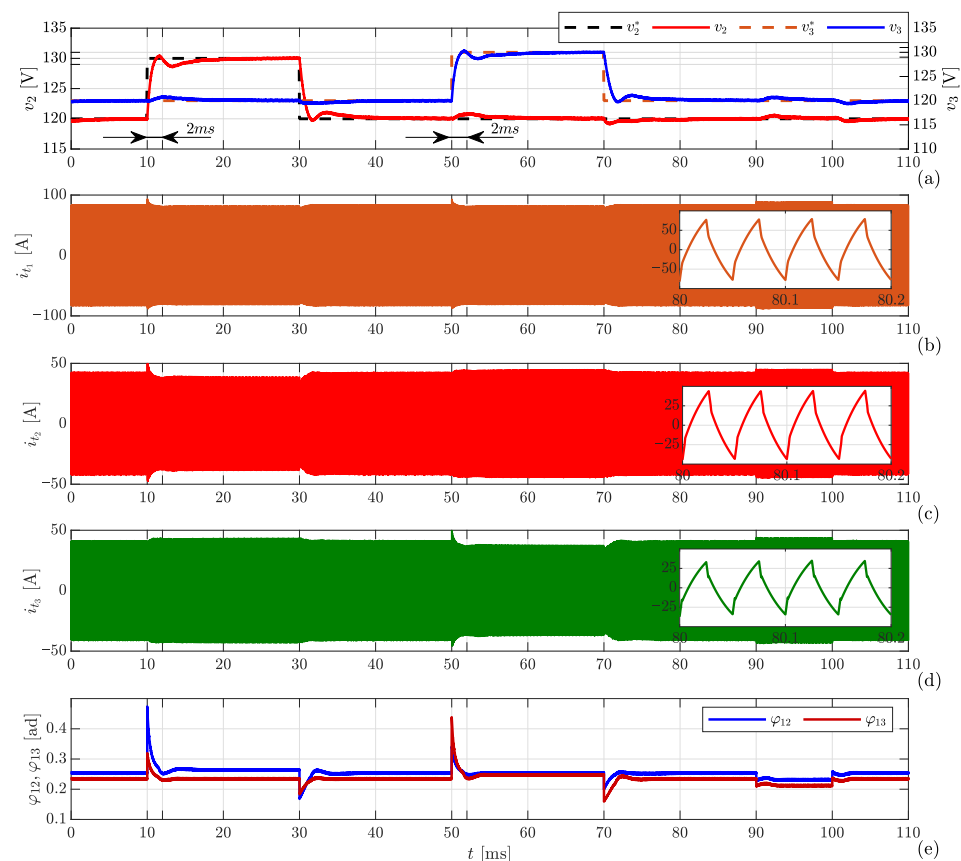


**Figure 7.** Converter performance in response to reference changes and variations in port 1 voltage. (a) Voltages at ports 2 and 3. (b) Primary current of the HFT. (c) Secondary current ( $i_{i2}$ ) of the HFT. (d) Secondary current ( $i_{i3}$ ) of the HFT. (e) Control actions.

Figure 7b–d shows the three HFT currents, which do not exhibit significant overshoots or oscillations in response to the changes made and have a zero average value. The inset boxes show the characteristic trapezoidal waveform, with a switching frequency of 20 kHz.

Finally, Figure 7e shows the control actions  $\varphi_{12}$  and  $\varphi_{13}$ . In both cases, transients with small oscillations and overshoots are observed after reference changes; however, the signals remain within their operating limits at all times, without reaching saturation.

## 5.2. Load Variations

To evaluate the robustness and regulation capability of the proposed control strategy, a 130 ms transient simulation was performed. During the test, a reference value of 120 V was set for the voltages at ports 2 ( $v_2$ ) and 3 ( $v_3$ ). The test started with a 15  $\Omega$  resistive load and a 1 kW constant power load (CPL) at both ports. At  $t = 10$  ms, the resistive load at port 2 was disconnected, maintaining the CPL. Subsequently, at  $t = 30$  ms, the port 2 CPL was increased to 1.25 kW, and at  $t = 50$  ms, a 4 A constant current source was connected. Next, at  $t = 70$  ms, the resistive load at port 3 was disconnected, maintaining the CPL, which was then increased to 1.4 kW at  $t = 90$  ms. Finally, at  $t = 110$  ms, a 4 A constant current source was connected to port 3.

The system response to load variations is shown in Figure 8. It can be seen that the controller manages to maintain both voltages ( $v_2$  and  $v_3$ ) at their reference values (see Figure 8a), showing decreases of no more than 6% with respect to their reference value (120 V) in response to load increases, and when the resistive load is disconnected, in both cases an increase of around 10% is observed with respect to their nominal value, although in all load disturbances the controller manages to reach the reference value again.

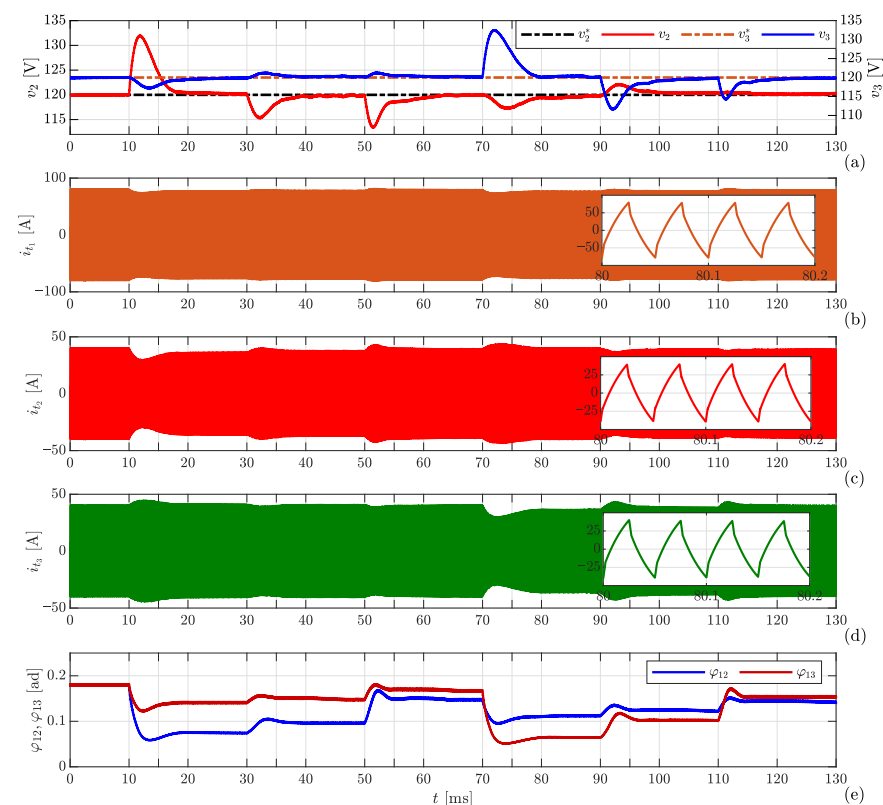


**Figure 8.** Converter performance under load changes at ports 2 and 3, respectively. (a) Voltages at ports 2 and 3. (b) Primary current of the HFT. (c) Secondary current ( $i_{t2}$ ) of the HFT. (d) Secondary current ( $i_{t3}$ ) of the HFT. (e) Control actions.

The currents of the three HFT windings for this test are shown in Figure 8b–d, which exhibit transient behavior without significant oscillations or overshoot, and maintain their average value at zero, as can be verified in the details shown in each figure.

Figure 8e shows the phase angles  $\varphi_{12}$  and  $\varphi_{13}$  applied to the switching signals of ports 2 and 3, respectively. It can be observed that both phase shifts exhibit a transient response with small overshoots in response to disturbances, although at no point do these control actions reach saturation.

Figure 9 (above) shows the output currents  $i_{02}$  and  $i_{03}$ , which validate the system response to the load variations imposed at the different times when the load changes occur. Finally, Figure 9 (below) shows the voltages at the terminals of the HFT. It can be observed that all the voltages have a square waveform, with a mean value of zero. In the first detail, it can be seen that the phase shifts between the active bridges ( $\varphi_{12}$  and  $\varphi_{13}$ ) are different from each other since the connected loads are not equal, while in the second detail, the phase shifts are similar because the power of the loads connected to both ports has a small difference between them.

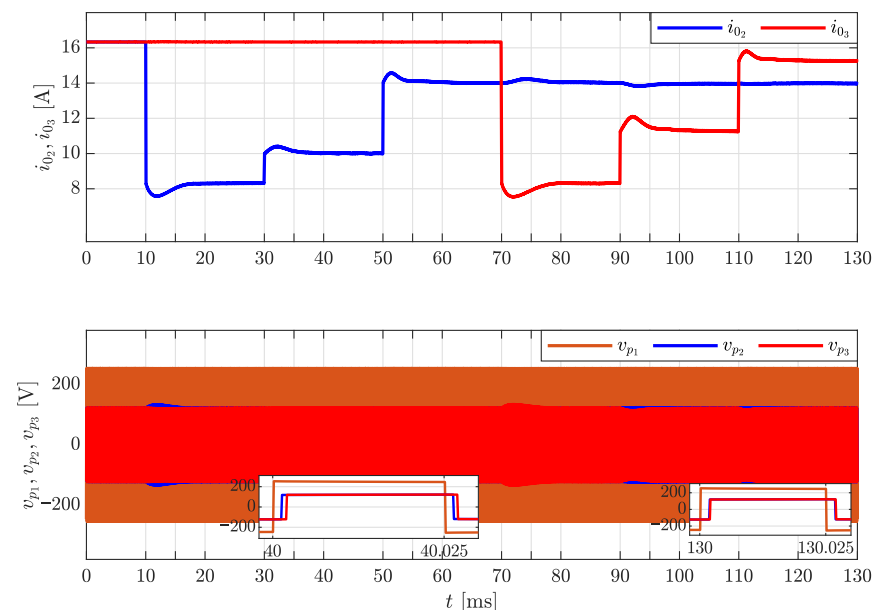


**Figure 9.** Converter performance under load changes at ports 2 and 3, respectively. **(Top):** outputs ports currents. **(Bottom):** voltages across the HFT terminals.

### 5.3. Comparison of the Proposed Control Strategy with a PI and SMC Controllers

Finally, the control strategy proposed in this work (FL) was compared with a PI controller proposed in [21] and a Sliding Mode Control (SMC) strategy proposed in [30], which were both adjusted with the same performance requirements established for the FL controller.

The test performed has a duration of 130 ms and begins with a resistive load of  $15 \Omega$  and CPL of 1 kW at each port. The voltage reference changes, from 120 V to 130 V, were applied at  $t = 10$  ms for  $v_2$  and at  $t = 30$  ms for  $v_3$ . Subsequently, at  $t = 50$  ms, the CPL of port 2 was increased to 1.2 kW, followed by the connection of a constant current source of 4 A at  $t = 70$  ms. Next, the CPL of port 3 is increased to 1.4 kW at  $t = 90$  ms, and finally at  $t = 110$  ms, a constant current source of 4 A is connected.

The dynamic response of the voltages of both ports ( $v_2$  and  $v_3$ ) is shown in Figure 10a. The proposed FL controller exhibits the best performance, with the fastest recovery time and the lowest overshoot among the three techniques. The SMC shows a robust response but with slightly higher oscillations during transients compared to FL. On the other hand, due to the nonlinear nature of the system and the connected loads, the PI exhibits the more pronounced oscillations and the longest settling time after each disturbance. Specifically, during the current source connection at  $t = 70$  ms and  $t = 110$  ms, the FL strategy maintains the voltages  $v_2$  and  $v_3$  with significantly greater stability than the SMC and PI.

Finally, Figure 10b–d shows the HFT currents  $i_{t1}$ ,  $i_{t2}$ , and  $i_{t3}$ , respectively. It can be seen that, in steady state, the HFT currents for the three controllers have a similar waveform (see details for each one), although the FL strategy achieves a response with fewer oscillations

in response to changes in reference and load. The zoom-in views confirm that while the basic waveform is preserved, the FL controller provides a smoother transition between operating points.

To summarize the performance comparison, Table 3 presents a qualitative evaluation of the three control strategies based on the obtained results.

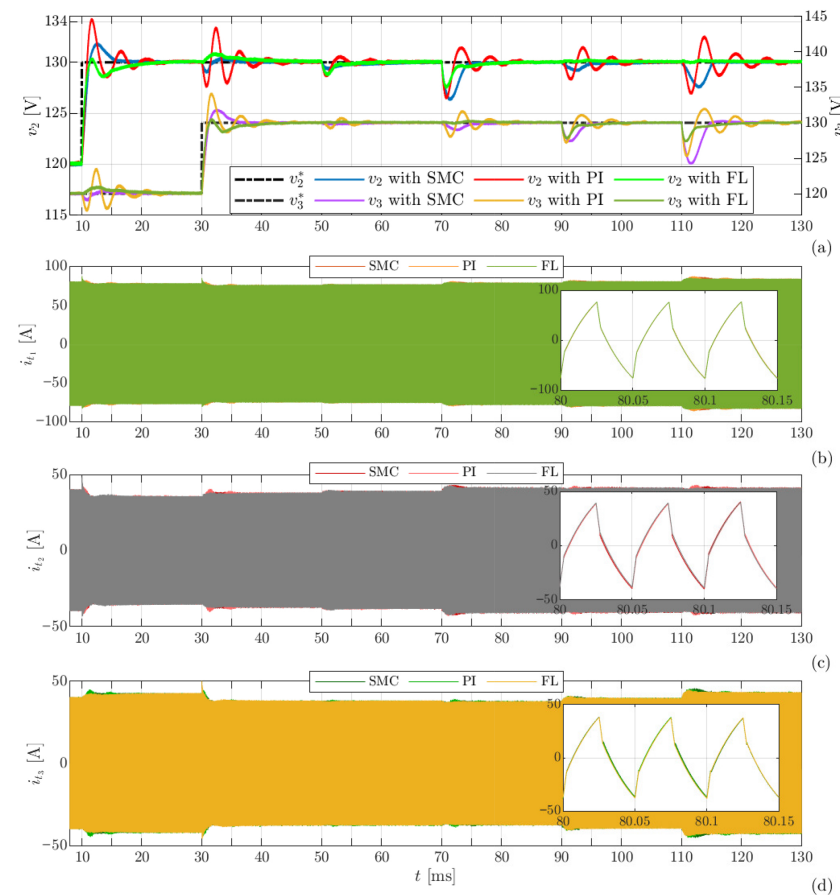


**Figure 10.** Comparison of the proposed control strategy with a conventional PI controller [21], and SMC controller [30], in response to reference and load changes. (a) Voltages at ports 2 and 3. (b) Primary current of the HFT. (c) Secondary current ( $i_{t2}$ ) of the HFT. (d) Secondary current ( $i_{t3}$ ) of the HFT.

**Table 3.** Performance comparison between PI, SMC, and the proposed FL controller.

| Parameter                      | PI [21]     | SMC [30]    | FL (Proposed) |
|--------------------------------|-------------|-------------|---------------|
| Settling time [ms]             | $\approx 6$ | $\approx 2$ | $\approx 2$   |
| Steady-state error [V]         | $< 0.3$     | $< 0.1$     | $< 0.1$       |
| Max. Voltage Overshoot [V]     | $\approx 4$ | $\approx 2$ | $\approx 0.4$ |
| Max. Voltage Drop [V]          | $\approx 4$ | $\approx 4$ | $\approx 2$   |
| Recovery Time (Load Step) [ms] | $\approx 8$ | $\approx 4$ | $\approx 4$   |
| Robustness to Load Changes     | Low         | High        | High          |
| Complexity                     | Low         | Moderate    | High          |

#### 5.4. Robustness Analysis Against Parametric Variations

To evaluate the reliability and robustness of the proposed control strategy, a parametric sensitivity analysis was performed. This test considers a variation of  $\pm 10\%$  in the nominal values of the main passive components of the converter: the DC-bus capacitors ( $C_1$ ,  $C_2$ ,  $C_3$ ), the leakage inductances of the high-frequency transformer (HFT), and the series resistances

of its windings ( $R_t$ ). The objective is to demonstrate that the controller maintains stability and performance even under modeling uncertainties or component aging.

The test conditions are identical to those described in the Section 5.3 regarding reference steps and load disturbances, but focusing on the parametric variation (PV) effects. The dynamic response of the system is shown in Figure 11. It can be observed in Figure 11a that the output voltages  $v_2$  and  $v_3$  follow their respective references with negligible deviations, regardless of whether the parameters are at their nominal value, +10% PV, or −10% PV. The recovery time and the damping of the oscillations remain almost identical in the three cases, confirming the high robustness of the controller proposed in this work.

Furthermore, Figure 11b–d shows the HFT currents  $i_{t1}$ ,  $i_{t2}$ , and  $i_{t3}$ . The zoom-in views provided in each subplot reveal that the current waveforms maintain their symmetry and phase-shift relations. Although the peak values of the currents exhibit minor variations due to the change in the leakage inductance ( $L_t$ ), the controller effectively compensates for these effects without degrading the voltage regulation at the output ports. These results indicate that the proposed strategy is suitable for practical applications where manufacturing tolerances or thermal effects may shift the passive component values, ensuring stable operation across a wide parametric range.

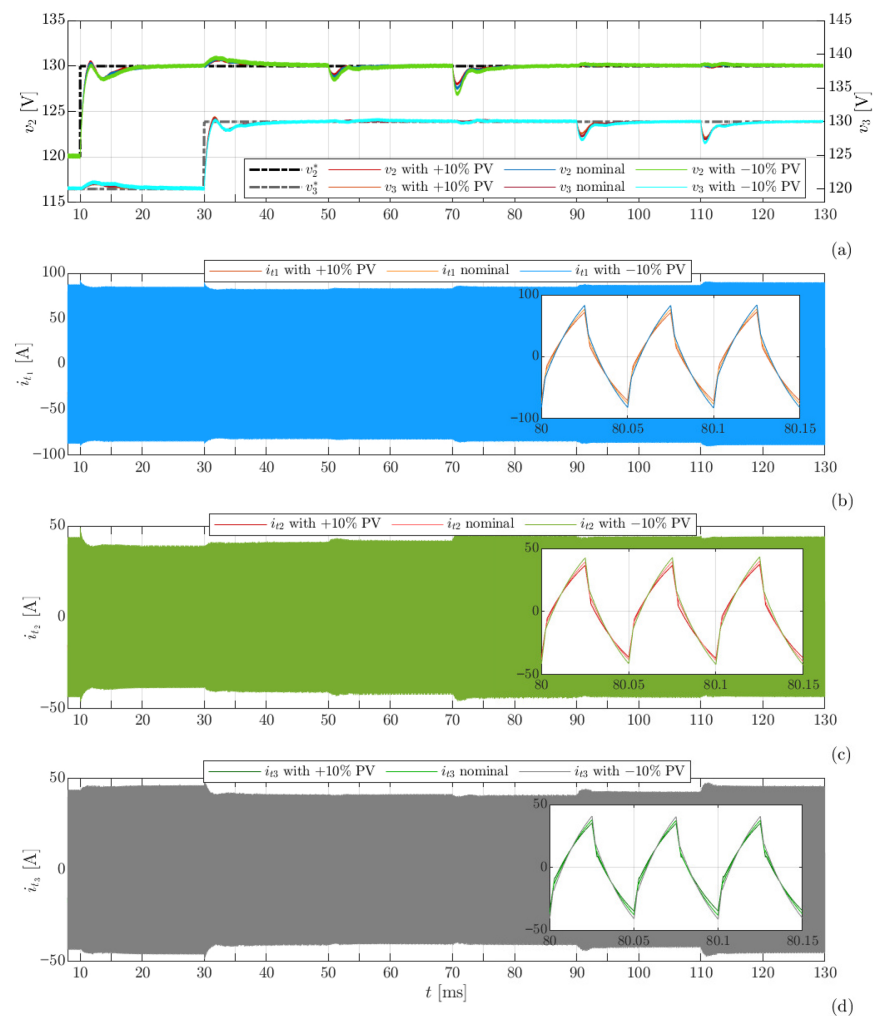


**Figure 11.** Performance of the proposed control strategy under parametric variations of  $\pm 10\%$ . (a) Voltages at ports 2 and 3. (b) Current  $i_{t1}$  of the HFT. (c) Current  $i_{t2}$  of the HFT. (d) Current  $i_{t3}$  of the HFT.

## 6. Conclusions

In this paper, a new control strategy was presented for a TAB DC-DC converter used to interconnect three buses of a DC MG. The designed controller allows the voltage to be regulated in two of the MG buses, to which generalized loads composed of components of constant impedance, constant current, and constant power, are connected. Unlike the works proposed in the literature, the converter was modeled based on the equivalent triangle circuit of the HFT and using GSSA. In this way, the dynamics of the fundamental AC component for the HFT currents and DC for the port voltages are captured, and the calculation of the midpoint voltage obtained with the star equivalent is avoided.

The controller was designed using a nonlinear FL-based technique, and thanks to the construction of a nested control scheme, with an internal loop for controlling the real and imaginary parts of the HFT currents and an external control loop for controlling the energy at ports 2 and 3, and thus regulating the voltage at both ports, a controller with good dynamic performance is obtained that ensures the stability of the system in the face of variations in reference and input voltages and large load excursions.

Simulation results demonstrate the superior dynamic performance of the proposed controller. The system exhibits a robust response to reference changes and input voltage perturbations. Furthermore, it ensures stable operation even under high inherent instability caused by the interaction between constant power (CPL) and constant current (ICC) loads, while maintaining zero steady-state error. Finally, comparative studies involving conventional PI and SMC strategies reveal that the proposed nonlinear FL control achieves enhanced dynamic performance and faster convergence, similar to the SMC. Furthermore, unlike the PI scheme, which exhibits degraded performance due to port cross-coupling, the proposed strategy ensures precise regulation and robustness under large-scale load variations, offering a superior balance between implementation simplicity and transient response for TAB-based DC microgrids.

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