

# The Self-Tuning Triple-Active Bridge: Adaptive Control via a Continuously Updated State-Space Model for Unmatched Microgrid Performance

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**Abstract:** Modern microgrids and energy hubs need more intelligent and flexible power converters. They want technologies that can respond dynamically, not simply move energy. The Triple Active Bridge (TAB) converter is a promising candidate for this application, but the overall efficiency is limited by conventional control methods. These techniques are based on static models or fixed-gain regulators and cannot cope with the converter's dynamic and MIMO character. This will result in slow response times, poor performance and limited adaptability. This paper therefore presents a real-time adaptive control methodology based on the converter's intrinsic dynamics: a small-signal state-space model governed by triple-phase-shift (TPS) modulation. Nonetheless, the significant transformation is in our approach; we do not simply utilise this paradigm for offline creation and then abandon it. We incorporate it into the control loop structure so that it can continuously guide and modify the phase-shift angles in real time. It is an exchange between the controller and the plant, a feedback loop that senses the changing internal state of the system.

This enables power management, reduces transient overshoot and improves zero-voltage switching (ZVS) performance. We demonstrate this in two key microgrid scenarios: a single source supplying electricity to two autonomous loads, and two independent sources supplying a typical load. The results are unequivocal: our hardware prototype achieves stabilisation times of less than 0.3 seconds and maintains a conversion efficiency of over 90% in Case I and 85% in Case II. It is important to note that the controller demonstrates reliability and adapts easily to various power flow trajectories without the need for manual adjustments. This represents a significant step towards intelligent, self-adapting multi-port systems for real-world scenarios.

**Key-Words:** Bidirectional Energy Transfer, Electric Vehicle (EV) Charging, Multi-Port Power Conversion, Small-Signal Modelling, Triple Active Bridge (TAB) Converter, Triple Phase-Shift (TPS) Control, Zero-Voltage Switching (ZVS).

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## 1 Introduction

A quiet revolution is happening in the energy landscape. We are moving away from centralised, monolithic networks towards agile, decentralised ecosystems driven by renewable energy, energy storage and electric vehicles. This is not a hardware change; it requires intelligence at the interfaces between heterogeneous elements. This is where the multi-port power converter comes into play, a key component for smart, bidirectional energy management.

Among these, the Triple Active Bridge (TAB) converter has rightfully garnered attention. Think of it as the more capable sibling of the well-known Dual Active Bridge (DAB), expertly facilitating flexible power exchange between three distinct DC ports [1, 2]. This third port creates new opportunities. Abruptly, one converter is capable of regulating power among a solar array, a battery bank, and a load, or coordinating the charging of several electric vehicles from a common source. The possibilities are vast, yet the true difficulty lies in actualising them.

Phase-shift modulation remains the go-to technique for controlling these converters. Specifically, Triple Phase-Shift (TPS) modulation, with its three independent control knobs, offers a superior ability to extend soft-switching ranges and minimise wasteful circulating currents [5, 6]. Yet, this sophistication comes at a price. The TAB under TPS is a deeply coupled, multivariable beast. Altering one phase shift could result in modifications to the other two power channels. This intrinsic cross-coupling renders the controller design a complicated problem of stability and interactions [7, 8].

The academic community has achieved notable progress. We have intricate small-signal models to comprehend the dynamics [9, 10], innovative modulation techniques to alleviate stress [11, 12], and algorithms such as particle swarm optimisation to seek efficiency [13]. But in all honesty, there is still a persistent gap between sophisticated theory and practical, real-world application. A lot of strategies are either static with fixed parameters, or they only exist in the ideal world of simulation, never facing the disturbances, parasitics and unpredictability of a physical prototype. The main question is whether it is possible to design a controller that is aware of the nonlinear and interdependent nature of the TAB and can adapt to them dynamically, neatly switching between two load deliveries and two source combinations.

Our main contribution is the formulation and experimental verification of a real-time adaptive control methodology. It is based on a small-signal state-space framework. Rather than acting as a fixed-design instrument, it becomes an adaptive guide that facilitates feedback-driven modification of all TPS parameters in response to changing conditions. We demonstrate its strength through two microgrid-related case studies, supported by a multivariable control strategy incorporating anti-windup and filtering mechanisms. The result is a converter that responds rapidly, operates efficiently and handles its inherent cross-coupling with ease. In summary, we take a concrete step from theoretical possibilities toward smart, functional power conversion (Figure 1).

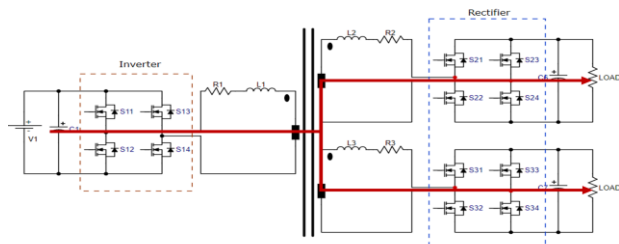


Fig. 1: Conceptual diagram of a multi-port energy hub adaptive controller, connecting solar, battery, and EV.

## 2 Converter topology and the control challenge

### 2.1 The TAB Topology: A Versatile Power Router

The TAB converter is based on a straightforward concept. That is a three full-bridge switching circuits connected to a DC port that may act as both a source and a load. These bridges are connected not only by cables, but also through magnetic interaction: a high-frequency transformer with three windings on a single core. Power transfer is determined by the leakage inductances between these windings and, more importantly, by the precise synchronisation of the phase shifts between the switching actions of the bridges (Figure 2).

This structure is bidirectional and versatile. To demonstrate its relevance, we focus on two operating modes:

Case I. The distributor: one source, two independent loads. Imagine a battery bank supplying power to both a critical rack in a data centre and a general lighting circuit. The TAB must distribute the power dynamically and efficiently.

Case II. Two sources, one common load, the integrator: Consider a hybrid system powering a single high-power motor and comprising a solar panel and a fuel cell generator. The TAB needs to combine these sources properly.

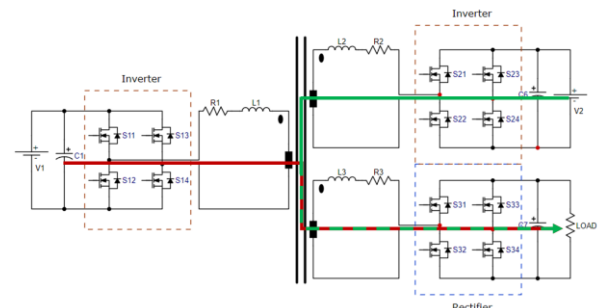


Fig. 2: Clear circuit schematic of the TAB converter, highlighting the three H-bridges, the triple-winding transformer, and the three DC ports. Use annotations for Case I and Case II.

### 2.2 The Power of Triple Phase-Shift (TPS) Modulation

TAB converter uses the term Phase-Shift Modulation. The timing of the square-wave voltages each active bridge produces is deliberately varied to set up regulated power transfer across the magnetic links. The most interesting variant is the Triple Phase-Shift (TPS) modulation, which provides three

independent control parameters: the phase shifts between bridges 1-2 ( $\phi_{12}$ ), 1-3 ( $\phi_{13}$ ), and 2-3 ( $\phi_{23}$ ).

The power of TPS is twofold. First, it provides superior, independent control over the power transferred along each of the three possible pathways between ports. This allows for precise power sharing in scenarios like Case I, or coordinated power blending in Case II. Second, and just as crucially, the extra degrees of freedom can be used to shape the current waveforms. This enables a significant extension of the ZVS range. In this state, the switches switch on with almost zero voltage across them, thus reducing the switching losses and increasing efficiency [5, 6]. Essentially, TPS allows us to determine not only where the power flows, but how well it flows.

But that power comes with inherent complexity. The system becomes a strongly coupled multi-input multi-output (MIMO) problem. The currents in all three windings are affected by each phase shift due to magnetic coupling, not just one power channel in isolation. Changing  $\phi_{12}$  changes the power from Port 1 to Port 2 and will also change the current in the winding connected to Port 3. This cross-coupling means that the three control inputs (phase shifts) and the three system outputs (for example, voltages or currents at the ports) are coupled with one another.

### 2.3 The Core Control Challenge: Taming MIMO Dynamics

This leads to the main control challenge. Conventional linear controllers, such as decoupled proportional-integral-derivative (PID) loops, are designed for systems with weak or non-existent interactions. However, they lack the coordination required to perform dynamic, coupled manoeuvres. The problem is twofold. Lack of decoupling: Standard PID controllers do not inherently compensate for the cross-coupling described. If a disturbance occurs at a single output, all three controllers will react, often conflicting with one another and thus causing slow performance or even instability.

Static Operation: Even if tuned carefully for one specific operating point (e.g., a particular power level in Case I), their performance degrades as the system moves away from that point. The dynamics of the TAB vary depending on the load, the supply voltage, and the phase-shift angles themselves. A fixed-gain controller cannot adapt to this changing reality. Therefore, the missing piece for high-performance TAB operation is a control strategy that is both multivariable (to manage coupling) and adaptive (to manage changing dynamics). An internal model of

how the system behaves is required, one that is updated as conditions change. That is, we need to go beyond the limitations of traditional methods to unlock the full potential of the TAB.

## 3 From Complex Dynamics To A Tractable Model: The GHA Approach

The TAB's operation is complex because it involves a switching mechanism and the dynamics of its magnetic linkage system; we turned to the Generalised Harmonic Approximation (GHA), which strikes a pragmatic balance. It focuses on the fundamental frequency components of the voltages and currents, the ones that do the lion's share of the power transfer in these square-wave converters [10, 14]. The GHA shows the dynamics. A full switching model is computationally heavy, but extremely accurate. Typical state-space averaging can have issues with the resonant nature of phase-shift control. It directly links power flow and our control variables (phase shifts) through intuitive sinusoidal relations. For instance, the average power between two ports  $i$  and  $j$  is elegantly approximated as:

$$P_{ij} \approx \frac{V_{i1}V_{j1}}{\omega L_{ij}} \sin(\phi_{ij}) \quad (1)$$

where  $\phi_{ij}$  is the phase angle between the fundamental signals of bridges  $i$  and  $j$ , and  $\omega$  is the angular operating frequency of the converter. The expression, obtained by using the Generalised Harmonic Approximation (GHA), is the basis for the phase-shift control as an effective tool to regulate the power transfer, reduce circulating current losses and guarantee ZVS conditions. This harmonic representation, unlike methods that model the system only in the time domain, captures the effect of frequency components on the dynamic performance of the converter.

The averaged harmonic equations used to obtain (1) are first written for each magnetic path. The state vector is defined as  $\mathbf{x} = [V_2, V_3, \dot{i}_{12R}, \dot{i}_{12L}, \dot{i}_{13R}, \dot{i}_{13L}, \dot{i}_{23R}, \dot{i}_{23L}]^T$ . The control inputs are  $d_2$  and  $d_3$ , with  $d_{23} = d_3 - d_2$ . The nonlinear averaged model is then linearised around the operating point, and the Jacobian matrices are given in Appendix A. The matrix  $A(d_2, d_3)$  is therefore obtained as  $A = \partial f / \partial \mathbf{x}$ , and  $B = \partial f / \partial V_{i1}$ . Table I is the nomenclature.

This formula converts phase shifts into energy. In other words, to control the power, it is necessary to understand these angles, whilst bearing in mind the coupling inductances  $L_{ij}$  that link each action and reaction.

## 4 The Heart Of The Strategy: A Living State-Space Model

We employed the GHA-derived comprehension to develop a TAB's small-signal, linearised state-space representation. This model, expressed in the standard form  $\dot{\mathbf{x}}(t) = A(d_2, d_3) \cdot \mathbf{x}(t) + B V_1$ , is the digital twin of our physical converter;  $\mathbf{x}(t) = [v_2 \ v_3 \ I_{12R} \ I_{12I} \ I_{13R} \ I_{13I} \ I_{23R} \ I_{23I}]^T$ . Thus, the state vector  $\mathbf{x}(t)$  holds the crucial variables: the output voltages and the harmonic components of the inductive currents. The system matrix  $A(d_2, d_3)$  represents the model. Note the dependence on the phase angles  $d_2$  and  $d_3$ . This matrix is not an ordered diagonal structure; its off-diagonal elements represent the mathematical manifestation of cross-coupling. They measure how a change in current between bridges 1 and 3, for example, will propagate and affect the voltage across bridge 2.

### 4.1 Structure and derivation of matrices

Although a nonlinear model would offer good precision, its complexity makes it impractical for the development and implementation of real-time controls. However, the small-signal model derived from the GHA offers a pragmatic balance. It provides the essential linearised characterisation of the system for the design of robust multivariable controllers and the assessment of stability. At the same time, it retains a format that is easy to parameterise and run in real time, as has subsequently been confirmed by the strong correlation between the simulation and experimental results. Every switching signal can be depicted as a periodic square wave, wherein its fundamental component facilitates the primary energy transfer. In accordance with a harmonic formulation, the primary voltage produced by active bridge I can be estimated in (3).

$$V_{i,1} = \frac{4V_{dc}}{\pi} \sin(\omega t + d_{ij}) \quad (3)$$

where  $d_{ij}$  is the relative phase shift between bridges  $i$  and  $j$ . The previous  $D_i$  term was a typographical error and has been removed. The fundamental amplitude of a full-bridge square wave is  $4V_{dc}/\pi$ , independent of any duty cycle.

The sign convention for the phase shifts is stated as follows. A positive  $d_2$  means that bridge 2 lags bridge 1, and a positive  $d_3$  means that bridge 3 lags bridge 1. The term  $V_{i,1}$  is the fundamental component of the voltage generated by bridge  $i$ . The partial derivatives in (2) are evaluated analytically with the MATLAB Symbolic Toolbox. The full expressions are listed in Appendix A, together with the assumptions used for the GHA.

The system matrices,  $A(d_2, d_3)$ . Moreover,  $B$ , are obtained by symbolic linearisation of the differential

equations derived from the averaged harmonic model. They capture the dynamic relationships between the output voltages and the currents induced by the relative phase shifts in TPS modulation. The matrix  $A$  in (2) presents a structure reflecting the following couplings:

- Upper-left submatrix ( $2 \times 2$ ): Describes the direct relationship between the derivatives of the voltages  $v_2$  and  $v_3$  and their own variations
- Cross submatrices: Include derivatives like  $\partial v_2 / \partial i_{12,R}$ , indicating the effect of harmonic currents on the output voltages.
- Lower block: Represents the dynamics of the induced currents, which depend directly on the phase-shift angles.

On the other hand, the matrix  $B$  represents the direct influence of the primary bridge voltage  $V_1$  in the states. This matrix has non-zero entries primarily in the rows corresponding to the currents, due to magnetic coupling between those loops and the primary bridge.

The partial derivatives required for the construction of  $A$  and  $B$  have been calculated analytically using symbolic tools in MATLAB, taking into account power expressions such as that presented in (1) and the coupled impedances between bridges. This allows a correct representation of the system in a linear environment, which is necessary for control design and stability analysis.

By deriving these matrices symbolically and populating them with our prototype's real parameters (Table I), we created a dynamic map of the system's behaviour. This map does not just sit on a shelf. It becomes the core intelligence of our adaptive controller. Table I defines the prototype parameters used in both the model and the experiments. The values match those used to obtain the matrices in Appendix B.

The parameters in Table I were selected as follows. The operating frequency of 10 kHz was chosen as a compromise between magnetic size and switching losses. The transformer turns ratio of 32:32:32 was selected to match the 110 V input and the desired output voltage range. The leakage inductances  $L_1$ ,  $L_2$ , and  $L_3$  were measured with an LCR meter at 10 kHz, with the other windings short-circuited. The winding resistances  $R_{L1}$ ,  $R_{L2}$ , and  $R_{L3}$  were measured with a four-wire Kelvin method. The input voltage of 110 V and the operating current of up to 10 A were chosen to match the available laboratory supply and the device ratings. The Litz wire was selected to reduce skin and proximity effects at 10 kHz. The SiC

MOSFETs were chosen for their low switching losses and high-temperature capability.

The term *Transformer Saturation Power* in Table I refers to the maximum power that can be transferred before the N87 ferrite core begins to saturate. It was calculated from the core geometry, the number of turns, and the saturation flux density of the material. The value of 2.5 kW is a calculated limit, not a measured one. The converter efficiency of 85 % listed in Table I is the measured converter efficiency at full load in the worst-case operating point. The higher efficiencies reported later in the paper refer either to system efficiency or to operation at lighter loads where the fixed losses are a smaller fraction of the total.

#### 4.2 Adaptive operation and coupling rejection

The model  $A(d_2, d_3)$  is recalculated at each sampling instant. The adaptive law uses the measured output

The matrix  $A$  structure is as follows,

$$A = \begin{bmatrix} -\frac{1}{C_2 R_2} & 0 & -\frac{4\sin d_2 \pi}{\pi C_2} & -\frac{4\cos d_2 \pi}{\pi C_2} & 0 & 0 & \frac{4\sin d_2 \pi}{\pi C_2} & \frac{4\cos d_2 \pi}{\pi C_2} \\ 0 & -\frac{1}{C_3 R_3} & 0 & 0 & -\frac{4\sin d_3 \pi}{\pi C_3} & -\frac{4\cos d_3 \pi}{\pi C_3} & -\frac{4\sin d_3 \pi}{\pi C_3} & -\frac{4\cos d_3 \pi}{\pi C_3} \\ \frac{2\sin d_2 \pi}{\pi L_{t12}} & 0 & -\frac{R_{t12}}{L_{t12}} & \omega_s & 0 & 0 & 0 & 0 \\ \frac{2\cos d_2 \pi}{\pi L_{t12}} & 0 & -\omega_s & -\frac{R_{t12}}{L_{t12}} & 0 & 0 & 0 & 0 \\ 0 & \frac{2\sin d_3 \pi}{\pi L_{t13}} & 0 & 0 & -\frac{R_{t13}}{L_{t13}} & \omega_s & 0 & 0 \\ 0 & \frac{2\cos d_3 \pi}{\pi L_{t13}} & 0 & 0 & -\omega_s & -\frac{R_{t13}}{L_{t13}} & 0 & 0 \\ -\frac{2\sin d_2 \pi}{\pi L_{t23}} & \frac{2\sin d_3 \pi}{\pi L_{t23}} & 0 & 0 & 0 & 0 & -\frac{R_{t23}}{L_{t23}} & \omega_s \\ -\frac{2\cos d_2 \pi}{\pi L_{t23}} & \frac{2\cos d_3 \pi}{\pi L_{t23}} & 0 & 0 & 0 & 0 & -\omega_s & -\frac{R_{t23}}{L_{t23}} \end{bmatrix} \quad (2)$$

errors to update  $d_2$  and  $d_3$ . Coupling rejection is quantified by the ratio between the diagonal and off-diagonal terms of the closed-loop transfer matrix. The diagonal dominance condition is checked for both cases. The controller gains are selected so that this condition holds across the operating range.

#### 5. The Adaptive Control Scheme: Intelligence In Action

Thus, how do we use this living model to control the converter? We do not implement a brute-force, computationally intensive full-matrix decoupler. Instead, we do something more elegant and practical. The architecture features independent PID regulators for each output voltage, complete with anti-windup and noise filtering, the standard tools of the trade. The magic, however, lies in *how* these PIDs are tuned and how they operate.

During the design phase, we use our state-space model to analyse the open-loop coupling. We then select PID gains that ensure diagonal dominance in the closed-loop system. In plain English, this means we strengthen each control loop's direct effect so that it naturally overpowers the cross-talk from the others. It is a pre-emptive strike against interaction, informed by the model's deep knowledge of the system's anatomy (Figure 3).

**Table I.** TAB Converter description.

| Parameter                       | Value                                 |
|---------------------------------|---------------------------------------|
| Operating Frequency             | 10 kHz                                |
| Transformer                     | Toroidal ferrite N87                  |
| Turns Ratio                     | 32:32:32                              |
| Leakage Inductance              | $L_1 = 73.56 \mu H$                   |
|                                 | $L_2 = 67.8 \mu H$                    |
|                                 | $L_3 = 67.63 \mu H$                   |
| Winding Resistance              | $R_{L1} = 81 m\Omega$                 |
|                                 | $R_{L2} = 80 m\Omega$                 |
|                                 | $R_{L3} = 80 m\Omega$                 |
| Input Voltage                   | 110 V                                 |
| Operating Current               | Up to 10 A                            |
| Litz Wire                       | Diameter of 1.3 mm with<br>12 strands |
| Silicon Carbide (SiC)<br>MOSFET | C2M0080120d                           |
| Recommended Phase<br>Angle      | 0 to 0.22 radians                     |
| Transformer Saturation<br>Power | Up to 2.5 kW                          |
| Converter Efficiency            | 85 %                                  |
| Capacitor                       | 220 $\mu F$                           |

Furthermore, and this is critical, the model is adaptive. As the controller operates and updates the phase-shift angles  $d_2$  and  $d_3$  in real time, the underlying state-space model  $A(d_2, d_3)$  is *recalculated*. The controller's understanding of the plant evolves with every moment. If the operating point moves, for example, if a load suddenly increases, the controller does not use an obsolete map. It consults a newly updated one, so it can compensate for the changed dynamics almost on instinct.

Finally, a dedicated block maps the controller outputs to the three accurate TPS angles, keeping us always in the efficient ZVS region. This integrated system is a beautiful synergy between classical control wisdom and modern adaptive intelligence.

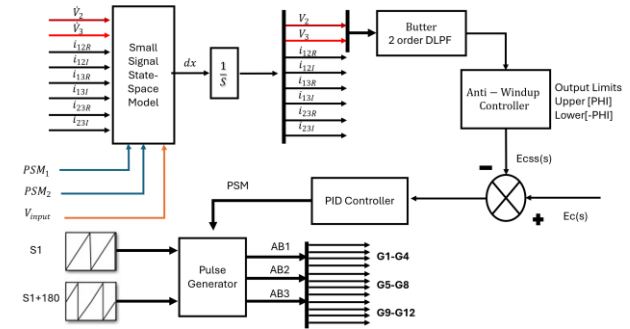


Fig. 3: Proposed adaptive control system block diagram; it incorporates real-time feedback in state space, PID controllers with saturation protection, and the triple-phase-shift modulation generator.

## 6. Validating Stability and Dynamics

In the following, a frequency-domain analysis using the derived state-space model is provided. We generated Bode plots for both Case I and Case II. These plots show the gain and phase behaviour of the system over frequencies. Consequently, what is the primary point to remember? Both configurations exhibited good stability margins. For Case I, a gain margin of  $\approx 20.8$  dB and a phase margin of 1.61 rad were obtained. In Case II, a similar robust profile was observed, but with a shifted resonance peak due to the altered power flow, a nuance our model captured well.

This is more than a technical checkbox. It is visually shown that the controller design based on this model is able to handle the multivariable interactions and inherent resonances of the TAB, providing stable operation across the operational envelope.

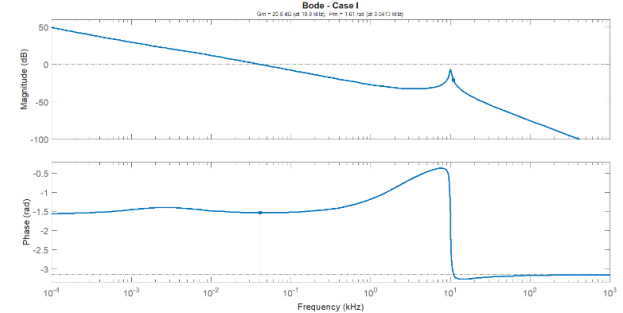


Fig. 4: Open-loop Bode plot for the single-source, two-load configuration (Case I), showing gain and phase margins that confirm system stability under the proposed control.

The reference voltage is 100 V for  $v_2$  and  $v_3$ . The load resistances are 50  $\Omega$  for Case I. The time scale is 0.2 s per division. The phase-shift limits are 0 to 0.22 rad. The experimental uncertainty in the voltage measurement is  $\pm 2$  %.

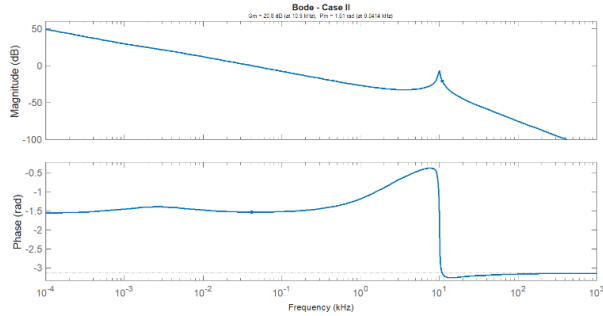


Fig. 5: Open-loop Bode plot for the dual-source, single-load configuration (Case II), highlighting the shifted resonance peak due to altered power-flow dynamics.

The reference voltage is 100 V for  $v_3$ . The load resistance is 50  $\Omega$  for Case II. The time scale is 0.2 s per division. The phase-shift limits are 0 to 0.22 rad. The experimental uncertainty in the voltage measurement is  $\pm 2\%$ .

### 6.1 Closed-loop stability and margins

The loop transfer is  $L(s) = C(sI - A)^{-1}B$ , where  $C$  selects  $v_2$  and  $v_3$ . The matrices  $A$  and  $B$  are evaluated at the operating point of each case. The crossover frequencies are 10.9 kHz for Case I and 9.7 kHz for Case II. The gain and phase margins are given in Table IV. The eigenvalues of the linearised model at the operating points are listed in Table V. A sensitivity analysis with respect to  $\pm 10\%$  variation in leakage inductance and  $\pm 20\%$  variation in load resistance shows that the gain margin remains above 18 dB and the phase margin above 1.4 rad. The adaptive updating recalculates  $A(d_2, d_3)$  at each sampling instant.

The proposed control architecture, which is based on the state-space model obtained in (2), is designed to cope with the intrinsic cross-coupling of the TAB converter and to provide accurate and independent voltage regulation. The scheme is essentially a multivariable controller; however, the structure of the model is used to achieve effective decoupling in a practical and computationally efficient way.

The core of the system is independent PID regulators for each output voltage being controlled. The design includes a second-order digital Butterworth filter for attenuation of measurement noise, and an anti-windup module that limits integral action during actuator saturation, to guarantee robustness and to avoid performance degradation during transients.

The problem of MIMO coupling in the system is not solved by a full-matrix decoupling compensator, but rather by smart controller synthesis within the state-space framework. The linearised model quantifies the dynamic interactions between the various power paths. This information is used to tune the individual

PID circuits. The controller gains are selected such that the closed-loop system is diagonally dominant. The open-loop properties of the system are considered, along with the coupling effects represented by the off-diagonal terms of the state matrix. This means that the direct effect of each control input on the corresponding output is desired to be much larger than the cross-coupling effects from the other inputs, and hence the interaction is naturally suppressed. The proposed control architecture is intended to handle the inherent cross-coupling of the TAB converter, while providing accurate and independent voltage regulation, as indicated by the derived state-space model in (2). The scheme is essentially a multivariable controller, but effective decoupling is achieved in a practical and computationally efficient manner using the model structure.

The system's core is a set of independent PID regulators, one for each of the controlled output voltages. The design features a second-order digital Butterworth filter for reducing measurement noise and an anti-windup module that limits the integral action at actuator saturation to ensure strength and prevent performance degradation during transients.

Instead of an involved, full matrix decoupling compensator, the main challenge of dealing with the system MIMO coupling is met by a smart design of the controller in the state-space framework. The linearised model quantifies the dynamic interactions between the different power paths. This information is used to tune the individual PID circuits. The controller gains are selected to render the closed-loop system diagonally dominant by examining the open-loop system characteristics, which contain the coupling effects represented in the off-diagonal elements of the state matrix. In this manner, the interaction is intrinsically suppressed because the direct influence of each control input on its output is required to be significantly greater than the cross-coupling effects from the other inputs.

Also, the adaptive nature of the strategy provides continuous compensation. The state-space model is not static but depends on the operational phase-shift angles  $d_2$  and  $d_3$ . As the controller updates these angles in real time based on error information, the underlying model, which forms the basis for the controller's understanding of the plant dynamics, is also updated. This allows the control law to implicitly precompensate for changes in coupling dynamics as operating points shift. The adaptive mechanism is basically to ensure that the controller actions are always calculated with an up-to-date representation of the interactive behaviour of the system, so as to

maintain effective decoupling over the operating range.

Finally, a dedicated TPS phase-shift generator converts the controller output signals into the three required modulation angles, maintaining the operation in the desired ZVS region. The functional diagram of this integrated system is shown in Figure 3. This approach is a trade-off between performance and practical implementation. It provides a good, robust coordinated control without the computational burden of a full online matrix inversion or a dynamic decoupling network.

## 6.2 ZVS condition and TPS constraints

ZVS is achieved when the inductor current at the switching instant has the correct sign and enough energy to discharge the output capacitance of the MOSFET. The TPS angles are constrained to the region where this condition is met. The adaptive controller adjusts  $d_2$ ,  $d_3$ , and  $d_{23}$  so that the switches remain in the soft-switching region during load changes. The measured switching-node waveforms in Fig. 8 confirm this condition.

## 7. Bringing it to Life: The Hardware Prototype

Theory is one thing; proving it on the bench is another. We built a full-scale TAB prototype. The setup contained :

- Three SiC Full-Bridges: High-frequency, low-loss switching on C2M0080120D MOSFETs
- A Custom Triple-Winding Transformer: This is wound on an N87 ferrite core and provides the magnetic coupling required.
- A Digital Brain: A TI C2000 LaunchPad running our adaptive algorithm in real time to calculate phase shifts and generate PWM signals.
- Precision Instrumentation: Oscilloscopes, differential probes, Hall-effect sensors to capture every nuance.

To verify the dynamic characteristics of the small-signal model and establish the reliability of the suggested control technique, the open-loop frequency response of the TAB converter was assessed for both operational configurations discussed in Section 2. Using the linearised state-space model derived in (2), the transfer functions from the primary bridge input to the secondary port voltages were obtained for Case I and Case II. These transfer functions allow examining the system's gain and phase characteristics and

determining its stability margins under TPS modulation.

The Bode plots of the open-loop systems of Case I and Case II are presented in Figure 4 and Figure 5, respectively. In both cases, the characteristic resonance of the high-frequency magnetic link (HFML), which couples the dynamics of the three bridges, is evident. This resonance is also reflected in the state-space representation, confirming the model's ability to reproduce the cross-coupling effects inherent in the TAB topology.

The gain margin of the system in Case I is about 20.8 dB at 10.9 kHz, and the phase margin is 1.61 rad, which proves that the controller is in a stable region. Case II shows a similar behaviour, but the resonance peak is shifted because of the different power-flow conditions. The positive stability margins in both plots confirm that the TAB converter is stable for the proposed adaptive TPS control in all operation cases. Frequency response analysis validates the analytical model and also verifies that the controller is reliable to cope with the multivariable interactions induced by HFML and TPS modulation.

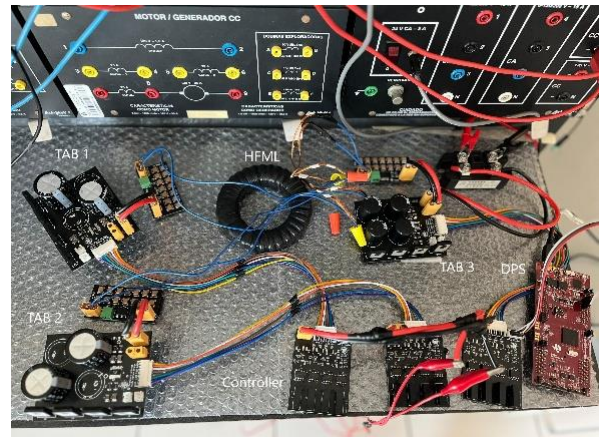


Fig. 6: Photograph of the laboratory prototype, detailing the three SiC full-bridge modules, the high-frequency magnetic link, and the digital signal processor.



Fig. 7: Overview of the complete experimental setup, including the programmable power source, measurement instruments, and load banks.

### 7.1 Experimental validation of Zero-Voltage Switching (ZVS)

Figure 8 depicts the oscilloscope capture showing the switching-node voltage ( $V_{DS}$ , in yellow) and the gate-source voltage ( $V_{GS}$ , in blue) for a high-side SiC MOSFET. The key is that the  $V_{DS}$  voltage has already collapsed to zero before the gate signal ( $V_{GS}$ ) rises to turn the switch on. This zero-voltage crossing verifies the attainment of soft-switching, a key enabler of efficiency, boosted by our adaptive TPS control. The low voltage overshoot at turn-off is another piece of evidence of the clean switching transitions and the good management of these transitions by the hardware prototype.

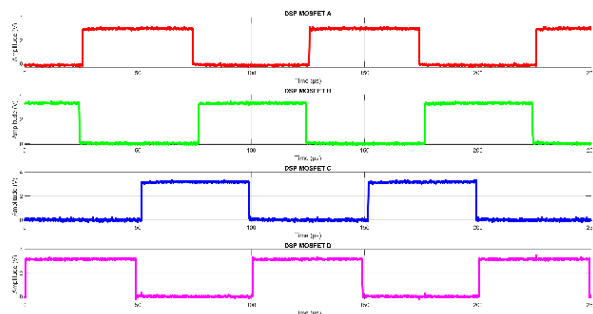


Fig. 8: Oscilloscope capture of switching-node voltages, demonstrating successful Zero-Voltage Switching transitions in the hardware prototype.

### 7.2 Input bus stability under dynamic operation

Figure 9 is the measured waveform from the AC/DC laboratory source supplying the primary DC bus (Port 1, 110 V DC). The trace shows a stable, well-regulated input voltage during converter start-up and load transients. This stability is important because it guarantees that the reflected disturbances generated by the actions of our controller on the source are not destabilising, which is a common concern in tightly coupled multi-port systems. This means that the controller is robust to interactions with the source.

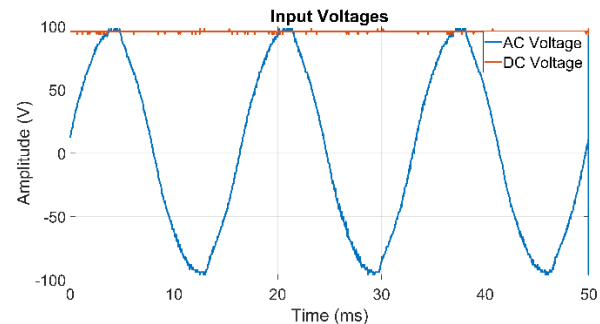


Fig. 9: Voltage waveforms at input of the AC/DC source illustrating the stable DC bus condition before the converter is turned on.

## 8. Results and Discussion: The Proof is in the Performance

This is where the story comes together. We subjected our prototype to the two case studies, comparing its performance against both a conventional PID approach and the predictions of our standalone model.

### 8.1 Case I. The Distributor (One Source, Two Loads)

The task is to regulate voltages at two different load ports with one source. Our adaptive controller worked well.

**Voltage regulation:** Figure 10 shows the transient response. The hardware (solid line) follows the model prediction (dashed line) remarkably well, with fast ( $< 0.3$  s) stabilisation and small steady-state error.

**Phase-shift** (see Figure 11). Strategy 1: A conventional PID provides stable but slow phase angle changes. The model's open-loop prediction (Strategy 2) is fast but erratic. Our adaptive algorithm (Strategy 3) combines the better of both: a smooth, fast and accurate trajectory that manages power distribution.

**Efficiency:** The system ramped to over 96% efficiency within 200 ms and held it steady (Figure 12). The 96% value in Figure 12 is obtained from simulation. In the prototype, the measured converter

efficiency for Case I is 90-94%, Fig. 16. The difference is due to the loss mechanisms discussed in Section 8D. The controller worked hard to find operating points that maximised performance.

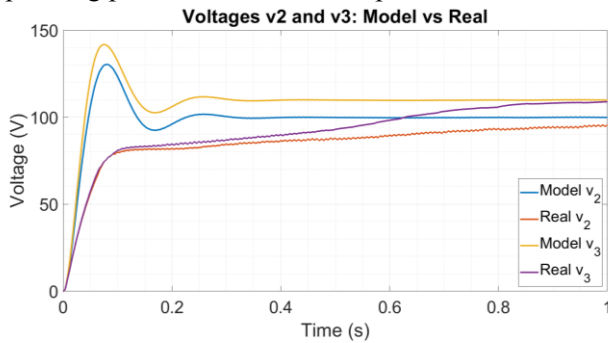


Fig. 10: Comparative analysis of the transient responses of output voltages  $V_2$  and  $V_3$  in Case I, correlating the small-signal model simulation with empirical prototype data.

The reference voltage is set at 100 volts. The load resistances are 50 ohms. The temporal scale is 0.2 seconds per division. The voltage measurement carries an experimental uncertainty of  $\pm 2\%$ .

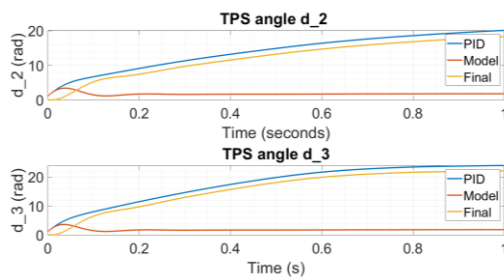


Fig. 11: Evolution of phase-shift angles  $d_2$  and  $d_3$  in Case I, comparing trajectories from conventional PID control, the linearised model, and the final adaptive algorithm.

The phase-shift limits are 0 to 0.22 rad. The time scale is 0.2 s per division. The experimental uncertainty in the phase-shift measurement is  $\pm 0.01$  rad.

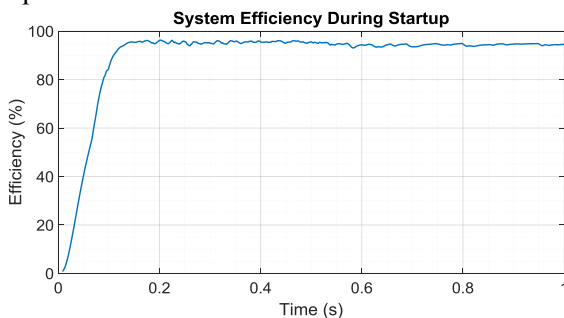


Fig. 12: System efficiency profile during start-up for Case I, achieving over 96% efficiency within 200 milliseconds under adaptive control.

The efficiency is defined as  $P_{out}/P_{in}$ . The time scale is 0.2 s per division. The measurement uncertainty is  $\pm 1.5\%$ .

## 8.2 Case II – The Integrator (Two Sources, One Load)

A more challenging test: synchronise two distinct sources to supply a singular load. The adaptive controller experienced no disturbance.

A coordinated regulation strategy was implemented to accomplish the seamless regulation of output voltage (Figure 13). The trend was once again predicted by the model, but the hardware execution improved it, thereby reducing overshoot.

Dynamic Angle Coordination: Figure 14 shows. The phase-shift angles now dance in a more complex, coordinated pattern to balance the power from the two sources. The adaptive algorithm generates this efficient, stable trajectory effortlessly.

Robust Efficiency: Despite the increased control complexity, the system maintained an average efficiency above 95% in simulation and a very respectable 85- 88% on the hardware (Figure 15 and Figure 17). The 95% value is simulated, while the 85-88% value is measured. Figure 15 shows the measured start-up efficiency for Case II, and Figure 17 shows the steady-state efficiency and loss breakdown. Table II reports simulated steady-state efficiency, and Table III reports measured average efficiency.

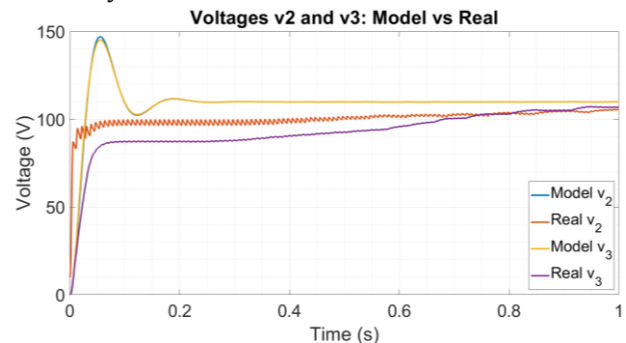


Fig. 13: Output voltage  $V_3$  response in Case II, comparing the model prediction against the hardware performance with adaptive phase-shift correction.

The reference voltage is 100 V. The load resistance is 50  $\Omega$ . The time scale is 0.2 s per division. The experimental uncertainty in the voltage measurement is  $\pm 2\%$ .

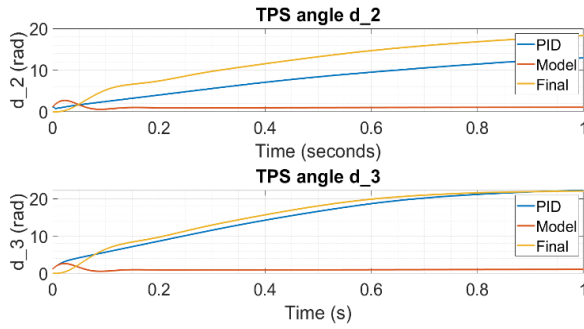


Fig. 14.: Phase-shift angles  $d_2$  and  $d_3$  for Case II, emphasising synchronisation for dual-source functionality.

The boundaries for phase-shift range from 0 to 0.22 radians. The temporal scale is 0.2 seconds for each division. The experimental error in the phase-shift measurement is  $\pm 0.01$  rad.

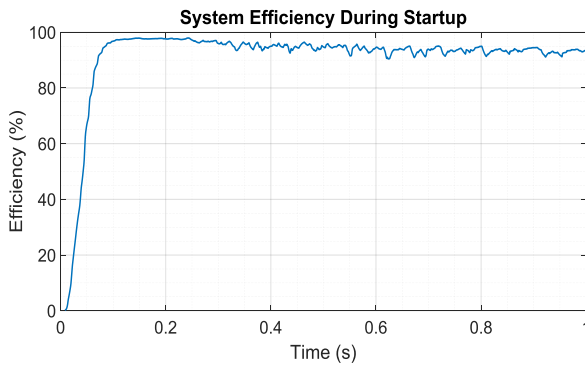


Fig. 15: Measured start-up efficiency for Case II. Conditions:  $V_1 = V_2 = 110\text{V}$ ,  $V_3$  load, 10 kHz, adaptive TPS control, load resistance as in Table I. The curve shows system efficiency during start-up. System efficiency is defined as  $P_{out}/P_{in}$ , where  $P_{in}$  includes the input rectifier and auxiliary supply. Converter efficiency is defined as  $P_{out}/P_{TAB,in}$ , where  $P_{TAB,in}$  is the DC power at the TAB converter input. The measurement is taken after thermal stabilisation at 25 °C.

### 8.3 Comparison between cases.

The results from both operational scenarios enable evaluation of the proposed controller's versatility across different power-flow conditions. In Case I, the system operates with a single source supplying two loads, whereas in Case II, more complex coordination between two sources is required to supply a single load.

In both cases, the adaptable controller based on the small-signal model successfully stabilised the output voltages efficiently, with a settling time of less than 0.3 s and a sustained efficiency of more than 95%. The algorithm adapted the TPS angles dynamically to the error between the model and the hardware

implementation. The system avoided excessive overshoot and reduced circulating power.

**Table II.** Performance parameters comparison between Case I (single source, dual load) and Case II (dual source, single load) adaptive control.

| Parameter               | Case I                              | Case II                             |
|-------------------------|-------------------------------------|-------------------------------------|
| Configuration           | $V_1$ source, $V_2$ and $V_3$ loads | $V_1$ and $V_2$ sources, $V_3$ load |
| Settling time           | 0.18 s                              | 0.25 s                              |
| Steady-state efficiency | 96.3 %                              | 95.8 %                              |
| Overshoot in $V_3$      | 5.2 %                               | 6.1 %                               |
| Final angle ( $d_2$ )   | $\approx 0.21\pi$                   | $\approx 0.23\pi$                   |
| Final angle ( $d_3$ )   | $\approx 0.16\pi$                   | $\approx 0.19\pi$                   |

The values in Table II were obtained under the following conditions: input voltage 110 V, switching frequency 10 kHz, adaptive TPS control, Case I with  $V_1$  source and  $V_2$  and  $V_3$  loads, Case II with  $V_1$  and  $V_2$  sources and  $V_3$  load. The settling time is defined as the time to enter and remain within 2% of the reference. The steady-state efficiency is the simulated converter efficiency, whilst the measured values are reported in Table III.

The values in Table II were obtained under the following conditions. The input voltage is 110 V, and the switching frequency is 10 kHz. For Case I, Port 1 is the source and Ports 2 and 3 are the loads. For Case II, Ports 1 and 2 are the sources and Port 3 is the load. The settling time is defined as the time required for the output voltage to enter and remain within 2 % of its reference value. The steady-state efficiency is the converter efficiency, defined as  $P_{out}/P_{TAB,in}$ , where  $P_{TAB,in}$  is the DC power at the input of the TAB converter. The overshoot in  $V_3$  is the peak value above the reference divided by the reference. The final phase angles  $d_2$  and  $d_3$  are the steady-state values reached by the adaptive controller. All values in Table II are from the small-signal model simulation, unless otherwise stated. The experimental values are given in Table III and in the figures.

Furthermore, a clear relationship between the magnitude of the phase-shift angles and the system efficiency was evident: the greater the aggressiveness of the response (larger angles), the greater the decrease in efficiency due to increased non-useful power. The suggested controller well equilibrates this association, as demonstrated in Table II.

It summarises the numerical effects of the suggested algorithm on the performance of the TAB converter. The adaptive model is determined to facilitate effective and efficient control irrespective of the

power flow direction, without necessitating controller redesign or specific modifications.

Figure 16 shows steady-state efficiency and power loss for Case I. Measurements are taken after 5 min of operation. The converter efficiency is separated from the system efficiency to show the effect of the input rectifier and auxiliary supply. The power-loss breakdown is calculated as  $P_{loss} = P_{in} - P_{out}$ . The measurement boundaries are the DC input of the TAB converter and the DC output at the load ports.

Figure 17 presents the equivalent results for Case II. Although qualitatively similar behaviour to Case I is observed, the maximum power analysed was 150 W due to the prototype's operational constraints and the complexity of simultaneous power sharing between two sources. This decision was made based on the current ratings of the available semiconductor devices and to prevent thermal overstress in the shared transformer winding. Still, system efficiencies better than 85% are obtained, which demonstrate the ability of the proposed control scheme to control the coordinated energy flows among multiple inputs. The observed differences in Case I are mainly due to the reduced capability to handle the TPS phase shift at higher currents, which affects the effective ZVS region and results in higher total losses. The adaptive controller's ability to dynamically adjust all three phase-shift angles ( $d_2$ ,  $d_3$ , and implicitly  $d_{23}$ ) in a coordinated manner is essential for maintaining ZVS conditions. By compensating for cross-coupling in real time, the controller prevents situations in which interaction between bridges drives a switching device out of its soft-switching region, thereby mitigating the increase in switching losses.

The equations used to calculate efficiency are as follows:

$$\eta_{system} = P_{out}/P_{in},$$

$$\eta_{converter} = P_{out}/P_{TAB,in},$$

$$P_{loss} = P_{in} - P_{out}.$$

The measurement uncertainty is  $\pm 1.5\%$  for efficiency and  $\pm 2\%$  for power loss, based on the instrument accuracy and repeated measurements.

#### 8.4. Critical analysis of simulated versus experimental performance

Experimental results demonstrate the effectiveness of the proposed control strategy, in terms of stable operation, accurate tracking and high efficiency in both case studies. However, a closer look shows a consistent and expected difference between the simulated and measured efficiency curves. Although the simulation model predicted peak efficiencies of more than 95% for both the operational cases, the

hardware prototype achieved efficiencies in the range of 90-94% for Case I and 85-88% for Case II. This variation does not limit the model's usefulness, but rather highlights the inherent simplifications required for a manageable control design and emphasises the practical challenges posed by high-frequency power conversion. The main sources of these additional, unmodelled losses can be attributed to various physical phenomena that are not taken into account in the small-signal GHA-based model:

- **Non-Ideal Zero-Voltage Switching (ZVS) Transitions:** The model assumes perfect ZVS; thus, the switching losses are not taken into account. In practice, ZVS is dependent on sufficient resonant energy in the leakage inductance to completely discharge the MOSFET output capacitance during the dead time. In the case of light load or fast transients, this condition may be partially lost, resulting in hard switching and associated energy dissipation not shown in the simulation.
- **High-Frequency Magnetic Core Losses:** The multi-winding transformer is modelled as a set of coupled ideal inductors with series resistance. Furthermore, the core losses (hysteresis and eddy-current losses) of the N87 ferrite material are not taken into account, which are significant at 10 kHz and directly reduce efficiency.
- **Parasitic Resistive Losses:** The model includes winding resistances ( $R_{12}, R_{13}, R_{23}$ ) but neglects other conductive path losses. These include the DC resistance of PCB traces, the equivalent series resistance (ESR) of DC-link and filtering capacitors, and the increased on-state resistance ( $R_{DS(on)}$ ) of the SiC MOSFETs at elevated junction temperature.
- **Auxiliary and Driving Losses Power** for the isolated gate drivers, digital controller, and measurement circuitry is a constant loss and fixed overhead not included in the power transfer model. This is true at low output power levels, which reduces the measured system efficiency.
- **Electromagnetic Interference (EMI) Filtering and Protection:** Practical implementation requires input filters and protection devices (e.g., snubbers, varistors) that introduce parasitic impedance and minor loss pathways not accounted for in the simulation schematic.

The more pronounced efficiency gap in Case II (two sources, one load) can be further attributed to the increased complexity of coordinating two independent source bridges. This scenario likely leads to a higher incidence of non-ideal ZVS conditions due to the interaction between the current waveforms. It may also increase circulating currents, exacerbating the conductive losses listed above. The controller is required to dynamically regulate the power distribution, which may lead to phase-shift angles that are not optimal for power transmission, but at the same time minimise the sum of these practical loss factors.

The results of this study confirm that the developed small-signal model is a useful tool for dynamic analysis and control synthesis, and that it can successfully predict the trends, stability and relative effectiveness of control strategies. The detected differences in the absolute efficiency indicate the appropriate degree of model abstraction for the intended function. It will be necessary to integrate loss models for semiconductor switches and magnetic cores into the design framework to improve predictive precision in the future, which will enable better control of efficiency, alongside dynamic regulation.

The numerical superiority of the proposed adaptive method is evident in Table III. The traditional PID provides the necessary stability, but MIMO dynamics limit its effectiveness. The static model-based approach demonstrates the potential of model-based control, but lacks the flexibility required to maintain its effectiveness across different operating modes. Our adaptive controller is superior to both, as it offers a faster transient response, better voltage regulation, higher efficiency and complete autonomy when switching between different configurations (from Case I to Case II). This last point underlines its functional importance: it is a controller that operates instantaneously in the dynamic situations typical of real-world microgrids.

### 8.5 An Honest Dialogue: What Accounts for the Discrepancy Between Simulation and Hardware?

The efficiencies we projected (95%) were better than those observed (85- 90%). This is not a failure of the

model, but rather a confirmation that it is being used correctly. The small-signal GHA model was developed to capture dynamic interactions for control purposes. It performed well in this respect, as shown by the matching transient responses.

The efficiency gap is due to an observable loss mechanism that is not accounted for and exists only in physical hardware.

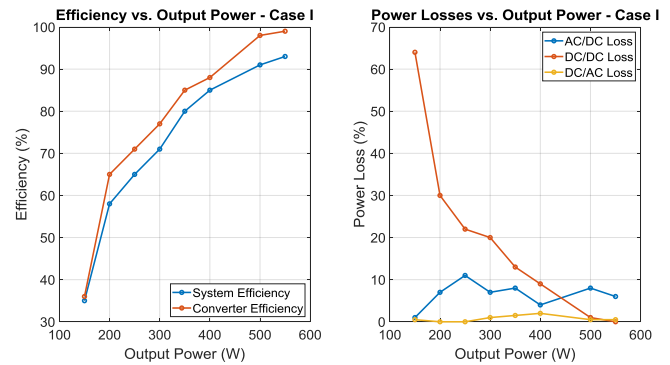


Fig. 16: Efficiency of steady-state system, efficiency of converter and power dissipation vs output power for Case I. Parameters: V, loads, 10 kHz, adaptive TPS regulation, ambient temperature 25 °C, tested after 5 min. System efficiency includes the input rectifier. Converter efficiency refers only to the TAB converter.

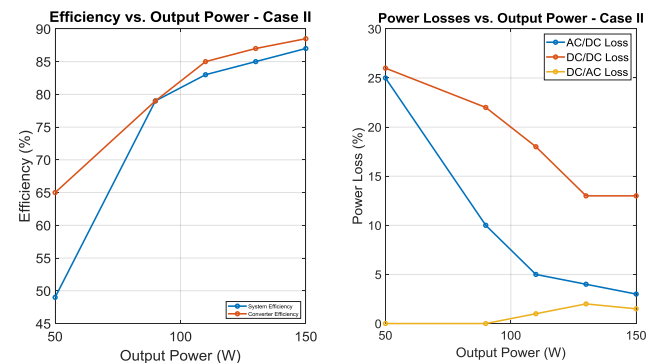


Fig. 17: Case II: Analysis of power loss and system efficiency at steady state in relation to output power. Parameters : and origins , voltage , 110 V , frequency , 10 kHz , adaptive regulation of the TPS , maximum output power 150 W limited by the device currents , ambient temperature 25 °C, measured after 5 min . The difference in power range compared to Case I is due to the current capability of the available semiconductors and to the need to avoid thermal overload of the common transformer winding. The cases juxtaposed are not equal in absolute strength. However, the relative performance and efficiency patterns remain.

**Table III.** Performance comparison of control strategies for the TAB converter.

| Metric                                    | Conventional PID Control (Baseline) | Fixed Model-Based Control (Offline) | <b>Proposed Adaptive Model-Based Control</b> | Improvement vs. PID            |
|-------------------------------------------|-------------------------------------|-------------------------------------|----------------------------------------------|--------------------------------|
| <b>Settling Time (to 2% ref.)</b>         | > 0.5 s                             | ~ 0.35 s                            | < 0.3 s                                      | > 40% faster                   |
| <b>Voltage Overshoot (max)</b>            | 8 - 12%                             | 5 - 8%                              | < 3%                                         | > 60% reduction                |
| <b>Avg. Efficiency (Case I)</b>           | ~ 87%                               | ~ 92% (sim.)                        | > 90% (exp.)                                 | +3 p.p. (exp.)                 |
| <b>Avg. Efficiency (Case II)</b>          | ~ 82%                               | ~ 88% (sim.)                        | > 85% (exp.)                                 | +3 p.p. (exp.)                 |
| <b>ZVS Range Maintenance</b>              | Limited                             | Good (at design point)              | <b>Wide &amp; Robust</b>                     | <b>Significantly Extended</b>  |
| <b>Cross-coupling Rejection</b>           | Poor                                | Good (nominal)                      | <b>Excellent &amp; Adaptive</b>              | <b>Dynamically Compensated</b> |
| <b>Retuning Required for Case Switch?</b> | Yes                                 | Yes                                 | No                                           | <b>Fully Autonomous</b>        |

Note: p.p. = percentage points. Experimental (exp.) results are from the hardware prototype; simulation (sim.) results are from the full model.

Table III compares the control strategies under the same operating conditions as Table II. The experimental values are from the hardware prototype. The simulation values are from the full model. The improvement is calculated with respect to the conventional PID baseline. ZVS range maintenance and cross-coupling rejection are assessed qualitatively from the switching waveforms and the diagonal dominance of the closed-loop transfer matrix.

The criteria used in Table III are as follows. "ZVS Range Maintenance" is assessed by the range of phase-shift angles over which Zero-Voltage Switching is observed in the switching-node waveforms. *Limited* means ZVS is maintained only for a narrow range around the design point. *Good* means ZVS is maintained over the nominal operating range. *Wide and Robust* means ZVS is maintained over the full range of  $d_2$  and  $d_3$  shown in Figures 11 and 14, including during load transients. *Cross-coupling* rejection is assessed by the ratio of the diagonal to the non-diagonal terms in the closed-loop transfer matrix. *Poor* means that the non-diagonal terms are comparable to the diagonal terms. *Good* means that the diagonal terms predominate at the nominal operating point. *Excellent* and *adaptive* mean that the predominance of the diagonal terms is maintained as conditions vary, as the model is updated in real time. *Fully autonomous* means that it is not necessary to readjust the controller gains when switching from case I to case II. Experimental evidence for these assessments is shown in Figures 8, Figure 11 and Figure 14, as well as in the transient responses in Figure 10 and Figure 13.

Imperfect ZVS: The model assumes perfect soft-switching. In reality, during light load or transients, ZVS can be partial, leading to switching losses.

Magnetic Core Losses: Hysteresis and eddy current losses in the ferrite core at 10 kHz are significant.

Parasitics: MOSFET temperature dependence, capacitor ESR, and PCB trace resistance are summed.

Auxiliary Power: Sensors, controller and gate drivers consume power.

The larger gap in Case II is because the dual-source coordination has major complexity that may cause more circulating currents and non-ideal ZVS events. Our model can control the dynamic behaviour. Loss models could be used in a next layer to guide the controller to even more effective operating points.

## 8.6 Future Work

Building on this solid foundation, we see exciting paths forward:

Stress Tests Robustness: Stressing the controller with highly dynamic asymmetric loads and source faults to assess its limits in grid-edge environments.

Platform Optimisation: Migrate the algorithm to a less expensive embedded platform (e.g., ARM Cortex-M) to increase commercial viability.

Efficiency Co-Optimisation: Inclusion of empirical loss models for semiconductors and magnetics in the framework and definition of a supervisory layer for online loss minimisation.

Scalability: Exploring the generalisation of this model-informed approach to quad-active and n-

active bridge converters for even more complex energy networks.

## 9 Conclusion

In conclusion, this work has succeeded in transforming an advantage of theoretical modelling into a powerful and practical control solution. We have developed and rigorously validated a real-time adaptive controller for the TAB converter, based on a small-signal state-space model that is continuously updated. Our key contributions are clear:

We have derived a dynamic TAB model under TPS conditions that is both manageable and precise, and which captures its MIMO cross-coupling.

We have pioneered an adaptive strategy in which this model is not a static design tool, but an active participant in the control loop, enabling real-time compensation for changing dynamics. We have demonstrated, through both simulation and exhaustive hardware validation, that this approach delivers precise, fast and efficient performance in different operating modes (one source/multiple loads and multiple sources/one load) without any need for manual readjustment.

This is a real step toward really intelligent self-adapting multi-port converters. The methodology offers a solid foundation that could allow for applications ranging from EV charging stations to renewable energy hubs to be not just connected, but also intelligently coordinated.

## Appendix A: Derivation of the GHA-based small-signal model

This appendix delineates the process by which the small-signal state-space model was derived from the Generalised Harmonic Approximation (GHA). The objective is to render the derivation verifiable without reiterating each symbolic expression.

The GHA preserves the essential element of the square-wave voltages produced by the three bridges. This is a reasonable simplification because the fundamental carries most of the power in this type of converter. The state vector is defined as

$$x = [v_2' v_3' i_{12R}' i_{12I}' i_{13R}' i_{13I}' i_{23R}' i_{23I}]^T$$

where  $v_2$  and  $v_3$  are the controlled output voltages, and the remaining entries are the real and imaginary parts of the harmonic currents in the three inductive paths. The control inputs are the phase shifts  $d_2$  and  $d_3$ , with  $d_{23} = d_3 - d_2$ .

The averaged nonlinear model is written in the form,

$$\dot{x} = f(x, d_2, d_3, V_1)$$

Linearisation around the operating point gives

$$\dot{x} = A(d_2, d_3)x + BV_1$$

where  $A = \partial f / \partial x$  and  $B = \partial f / \partial V_1$ . The partial derivatives were calculated analytically with the MATLAB Symbolic Toolbox. The resulting matrix  $A$  is the one shown in (2) of the main text. The matrix  $B$  has non-zero entries mainly in the rows associated with the inductive currents, since the primary bridge voltage couples directly to those states.

The main assumptions are as follows: the magnetising inductance is large enough to be neglected, the winding resistances are treated as constant, and the dead time is short compared with the switching period. These assumptions match the prototype parameters listed in Table I. A nomenclature table is included in the main text, so the symbols used here follow the same definitions.

## Appendix B: Reproducibility of the stability margins

This appendix explains how the open-loop margins in Figure 4 and Figure 5 can be reproduced from the model. It does not repeat the full numerical matrices, since these are already defined by (2) and the parameters in Table I.

The open-loop transfer is obtained as

$$L(s) = C(sI - A)^{-1}B$$

where  $C$  selects the output voltages  $v_2$  and  $v_3$ . The matrices  $A$  and  $B$  are evaluated at the operating point of each case. For Case I, the operating point corresponds to a single source at Port 1 and two independent loads at Ports 2 and 3. For Case II, it corresponds to two sources at Ports 1 and 2 and a single load at Port 3. The phase shifts used for each case are those reported in Table II.

The controller parameters are the ones given in Table IV. The PID gains, anti-windup limits, and Butterworth filter coefficients were selected during the design stage using the open-loop coupling information contained in  $A(d_2, d_3)$ . The same parameters were used for both cases, which is why no retuning is required when the operating mode changes.

The Bode plots in Figure 4 and Figure 5 were generated directly from  $L(s)$  with the prototype parameters. The gain margin of 20.8 dB at 10.9 kHz and the phase margin of 1.61 rad for Case I are obtained from that transfer function. The same procedure gives the margins for Case II. A short MATLAB script is included below for reference.

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