

Article

Design and Analysis of a 75 kW Five-Phase Two-Switch Buck–Boost Converter for Photovoltaic Systems

Marcin Zygmanski ^{1,*}, Dawid Mańka ² and Jan Strossa ³

¹ Power Electronics, Electrical Drives and Robotics Department (KENER), Silesian University of Technology, 44-100 Gliwice, Poland

² ENEL-PC Sp. z o.o., 44-178 Przyszowice, Poland; d.manka@enel-pc.pl

³ Department of Applied Electronics, Faculty of Electrical Engineering and Computer Science, VSB-Technical University of Ostrava, 708 33 Ostrava, Czech Republic; jan.strossa@vsb.cz

* Correspondence: marcin.zygmanski@polsl.pl

Abstract

This paper presents a five-phase, two-switch buck-boost (5P-TSBB) converter rated at 75 kW, intended for photovoltaic and energy storage applications that require a wide operating voltage range. The proposed system operates with photovoltaic input voltages ranging from 250 V to 1000 V and regulates the DC-link voltage between 600 V and 950 V. The converter supports two distinct operating modes: an independent multi-input mode for multiple independent input sources and an interleaved mode for a single high-power input. The feasible operating area of the converter is determined in the V_{in} – V_{out} plane, taking into account voltage, current, and power limitations. Simulation results and experimental investigations on a laboratory prototype, including measurements of efficiency and power losses, support the theoretical considerations.

Keywords: two-switch buck–boost converter; multi-phase DC–DC converter; photovoltaic systems; efficiency measurements

1. Introduction

Photovoltaic systems and energy storage interfaces increasingly require DC–DC converters capable of operating over a wide input voltage range while maintaining high efficiency and scalability [1–3]. This is particularly relevant for large-scale PV installations, where string voltages may vary significantly due to irradiance, temperature, partial shading, module mismatch, and system configuration [3–5].

For such applications, a step-up/step-down DC–DC converter is required to regulate the DC-link voltage when the PV input voltage is either lower or higher than the required output voltage. Among non-isolated solutions, the Two-Switch Buck–Boost (TSBB) converter is an attractive topology because it provides non-inverting operation and can operate in separate buck and boost modes using a relatively simple power stage [6,7]. Compared with classical single-switch buck–boost-derived converters, the TSBB structure reduces semiconductor voltage stress and avoids continuous operation in the combined buck–boost mode, thereby reducing unnecessary switching losses and improving efficiency over a wide voltage range [6–8].

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Several extensions of TSBB and non-inverting buck–boost converters have also been reported, including multiphase interleaved structures, magnetically coupled variants, and modified high-efficiency step-up/down topologies [9–12].

Most reported multiphase or interleaved buck–boost solutions assume symmetrical operating conditions and shared input sources, under which interleaving yields significant reductions in current ripple and RMS currents [9]. However, in practical photovoltaic applications, individual input sources often operate under asymmetrical voltage and power conditions, which limit the effectiveness of interleaving and increase control complexity [13,14].

For this reason, a converter intended for practical PV systems should not rely exclusively on symmetrical interleaved operation. In multi-string operation, independent processing of each input source is advantageous because it avoids current-sharing requirements between strings operating at different voltage and power levels [13–15]. On the other hand, when a single high-power source is connected, interleaving remains beneficial because it reduces current ripple and distributes thermal stress among several phases. A modular converter structure that supports both operating scenarios is therefore attractive for PV and energy storage interfaces [15,16].

Power losses in TSBB converters have also been addressed in previous studies, mainly in the context of topology modifications, reductions in the number of conducting components, coupled-inductor operation, or control strategies to improve efficiency [8,12]. However, these works usually focus on a specific converter variant or on a selected loss-reduction mechanism. In contrast, the present paper provides a broader loss-oriented analysis of a modular five-phase TSBB converter, including mode-dependent semiconductor conduction and switching losses, inductor winding losses, experimentally identified core losses, and experimental verification at both single-module and complete-converter levels.

To better position the proposed converter with respect to other non-isolated step-up/step-down solutions, Table 1 presents a qualitative comparison of selected converter topologies that can be considered for similar voltage-conversion tasks, including SEPIC, four-switch buck–boost, three-level buck–boost, and TSBB-based converters [17–19]. The comparison is based on the typical structure of each topology and focuses on component count, semiconductor voltage stress, main advantages, and practical limitations. SEPIC converters offer simple single-switch control, but they are associated with higher semiconductor voltage stress and a larger magnetic component count. Four-switch buck–boost converters provide flexible buck/boost operation and enable synchronous control, at the expense of a higher number of active switches and gate drivers. Three-level buck–boost converters can reduce semiconductor voltage stress but require more complex modulation and capacitor voltage balancing. In contrast, the proposed 5P-TSBB converter uses a relatively simple power stage with fewer active switches per module and supports both independent-input and interleaved operation. A fully quantitative comparison of efficiency and power density would require a dedicated optimized design of each converter under the same voltage, power, thermal, magnetic, and filtering constraints, which is beyond the scope of this work.

In this work, a five-phase TSBB converter is developed with a focus on practical operating conditions encountered in photovoltaic and energy-storage systems. The proposed modular structure enables two possible hardware configurations: independent input ports for multiple PV sources and a single-input interleaved configuration for high-power operation. The paper presents the design and analysis of the 75 kW 5P-TSBB converter, including its feasible operating region, mode-dependent current stresses, and power-loss distribution. The experimental verification includes single-module buck and boost tests and high-power operation of the complete five-module prototype in the interleaved single-input configuration.

Table 1. Qualitative comparison of selected non-isolated step-up/step-down topologies.

Topology	Number of Transistors per Module	Number of Diodes per Module	Number of Inductors	Semiconductor Voltage Stress	Advantages	Limitations
SEPIC	1	1	2 or 1 coupled	High ($V_{in} + V_{out}$)	Single switch control	High semiconductor voltage stress
Interleaved Four-Switch Buck-Boost Converter	4	0	1	Moderate (V_{in} or V_{out})	Flexible buck/boost transition	Higher number of active switches
Three-Level Buck-Boost Converter	8	0	1 + balancing capacitors	Low ($V_{in}/2$ or $V_{out}/2$)	Low voltage stress	Higher complexity
Proposed 5P-TSBB Converter	2	2	1	Moderate (V_{in} or V_{out})	Low number of active switches	Pulsating input/output currents

The main contributions of this paper are as follows:

- Development of a 75 kW five-phase Two-Switch Buck–Boost converter laboratory prototype composed of five 15 kW modules;
- Discussion of two possible hardware configurations adapted to multiple-input photovoltaic operation and single-input high-power interleaved operation;
- Determination of the feasible operating area in the V_{in} – V_{out} plane considering voltage, current, and power limitations;
- Formulation of a mode-dependent current and power-loss analysis for buck and boost operation;
- Experimental validation of the loss model using single-module tests and high-power verification of the complete five-phase converter in the interleaved single-input configuration.

2. Converter Description

2.1. Overall Topology

The proposed converter is part of a larger power conversion system that interfaces photovoltaic (PV) sources, battery energy storage (BES), and the utility grid. The overall system architecture, including the location of the analyzed five-phase Two-Switch Buck–Boost (5P-TSBB) converter, is shown in Figure 1.

The 5P-TSBB converter consists of five identical Two-Switch Buck–Boost (TSBB) phases connected to a common DC-link. Each TSBB phase is supplied from an independent input port and operates with a dedicated inductor, enabling a modular structure and independent power processing for multiple input sources.

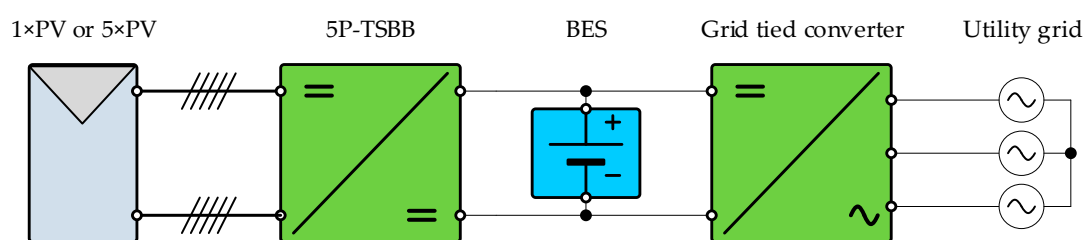


Figure 1. Overall architecture of the PV and BES power conversion system, showing the location of the five-phase Two-Switch Buck–Boost (5P-TSBB) converter under analysis.

2.2. Operating Voltage Ranges

The converter is designed to operate with photovoltaic input voltages ranging from 250 V to 1000 V. The output DC-link voltage is regulated between 600 V and 950 V, making the converter suitable for photovoltaic and energy-storage systems interfaced with grid-connected inverter stages.

2.3. Operating Modes

2.3.1. Independent Multi-Input Mode

In the multi-input operating mode, each TSBB phase processes power from an independent source without phase shifting. As illustrated in Figure 2, the input ports are electrically decoupled, making this operating mode particularly suitable for photovoltaic strings under asymmetric conditions. The proposed topology prevents undesired power coupling between input ports, ensuring independent power processing even under mismatched irradiance or temperature conditions.

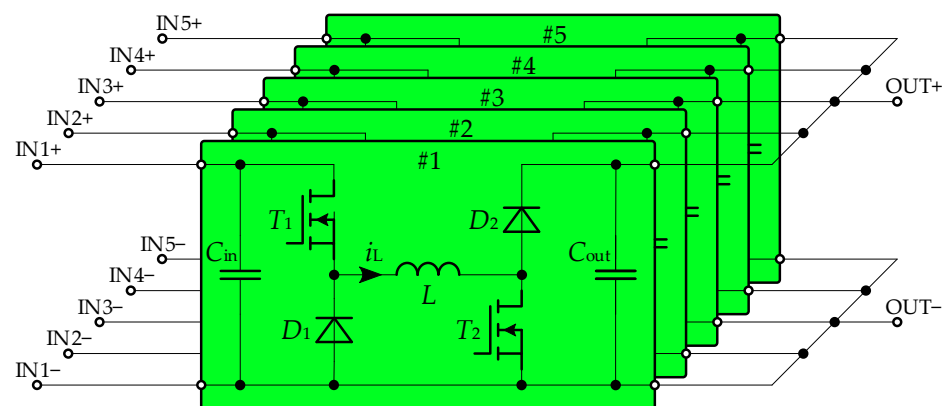


Figure 2. Connection of the five-phase Two-Switch Buck-Boost (5P-TSBB) converter operating in the multi-input mode with electrically decoupled input ports.

2.3.2. Interleaved Single-Input Mode

When a single high-power input source is connected, the converter can operate in an interleaved mode, in which phase shifting between the TSBB phases is applied to reduce current ripple and RMS currents in the power stage. The connection of the 5P-TSBB converter in this operating mode is shown in Figure 3.

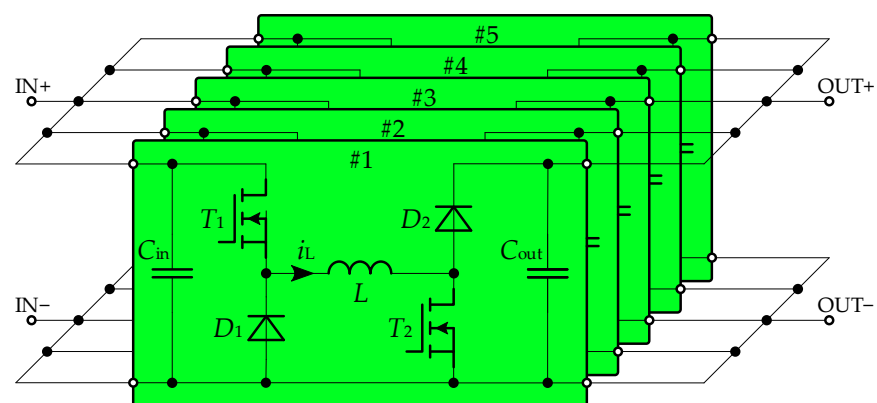


Figure 3. Connection of the five-phase Two-Switch Buck-Boost (5P-TSBB) converter operating in the interleaved single-input mode.

The two operating modes correspond to different hardware configurations of the input connections. The converter is not automatically reconfigured between the multi-input and interleaved single-input modes during operation. The required configuration is selected for a given application during manufacturing, commissioning, or external setup.

In this work, the complete five-module prototype was experimentally tested in the interleaved single-input configuration. The independent multi-input configuration is discussed as a possible hardware configuration enabled by the modular structure of the converter and the separated input ports of the TSBB modules. The experimental single-module tests characterize the buck and boost operations of a TSBB module, the basic building block of the five-module system. A dedicated experimental study of the simultaneous operation of several modules supplied from different input voltages is left for future work.

2.4. Converter Parameters

The developed five-phase Two-Switch Buck–Boost (5P-TSBB) converter consists of five identical TSBB power modules operating in parallel. Due to the modular and symmetrical structure of the converter, most electrical and control parameters are defined at the single-module level. These parameters are summarized in Table 2 and apply equally to each TSBB phase.

The overall ratings of the 5P-TSBB converter, such as the total output power, input current capability, and thermal loading, result from the combined operation of the five modules. Consequently, the parameters listed in Table 2 should be interpreted as per-module values. In contrast, the corresponding system-level values can be obtained by appropriate scaling according to the number of active phases.

Table 2. Electrical parameters of the TSBB module and the 5P-TSBB converter.

Parameter	Symbol	Value
TSBB maximum output power	$P_{\max}$	15 kW
Number of TSBB modules	n	5
5P-TSBB maximum output power	$P_{\max 5P}$	75 kW
Inductor inductance	L	1.2 mH
Input capacitance	C_{in}	20 μF
Output capacitance	C_{out}	20 μF
Maximum input current	$I_{\text{in,max}}$	25 A
Maximum output current	$I_{\text{out,max}}$	25 A
Input voltage range	$V_{\text{in,min}}, V_{\text{in,max}}$	250 V, 1000 V
Output voltage range	$V_{\text{out,min}}, V_{\text{out,max}}$	600 V, 950 V
Switching frequency	f_s	30 kHz

The main components used in the laboratory prototype of a single TSBB module are listed in Table 3.

Table 3. Main components used in the laboratory prototype of a single TSBB module.

Component	Part Type	Comments
Transistors T_1, T_2	SCT3040KL	$V_{\text{DSS}} = 1200 \text{ V}; R_{\text{DS(on)}} = 40 \text{ m}\Omega$
Diodes D_1, D_2	NDSH50120C	$V_{\text{RRM}} = 1200 \text{ V}; I_{\text{F}} = 50 \text{ A}$
Inductors	Custom-made	1.2 mH, 43 m Ω , 6 $\times$ E80 pairs, ferrite F887
Input capacitors C_{in}	B32776G0206K000	20 μF , 1100 V, 6 m Ω
Output capacitors C_{out}	B32776G0206K000	20 μF , 1100 V, 6 m Ω
Gate drivers	ACPL-352J	$v_{\text{GS}} = 18 \text{ V}/0 \text{ V}; R_{\text{Gon}} = 10 \Omega, R_{\text{Goff}} = 10 \Omega$
Microcontroller	TMS320F28027	for PWM generation and converter control

2.5. Control Implementation

The laboratory prototype was controlled using a TMS320F28027 microcontroller from the Texas Instruments C2000 family. The controller generated PWM signals for the TSBB modules and applied the duty-cycle references corresponding to the selected operating point.

The buck or boost operating mode was selected based on the voltage relationship between the input and output sides of the converter. For $V_{in} > V_{out}$, the module operated in buck mode, with transistor T_1 switched and transistor T_2 off. For $V_{in} < V_{out}$, the module operated in boost mode, transistor T_1 remained on, and transistor T_2 was switched. The duty-cycle references were calculated from the analytical relations presented in Section 3.

The experimental tests were performed under steady-state operating conditions for selected input and output voltage levels. For a given voltage operating point, the transferred power was adjusted by changing the duty-cycle command. The duty-cycle command was set with a resolution of 0.001, corresponding to 0.1 percentage points, and was limited to the range from 0.01 to 0.99 in order to avoid extremely short switching pulses.

A dynamic online transition between buck and boost modes was not investigated in this work. A combined buck–boost modulation strategy, in which both transistors are actively switched in the transition region, was not implemented in the present laboratory prototype and is considered as a future development.

In the interleaved single-input configuration, the five TSBB modules were controlled with phase-shifted PWM signals. The phase shift between adjacent modules was equal to $360^\circ/5 = 72^\circ$. In the multi-input configuration, each TSBB module is controlled independently, without interleaving, according to its own input and output conditions.

2.6. Operating Limitations and Feasible Region

The feasible operating region of the converter is constrained by the allowable ranges of input and output voltages, as well as by the current and power limitations of the power stage components. These constraints define the permissible combinations of input voltage, output voltage, and transferred power.

Figure 4 illustrates the operating limits of a single 15 kW TSBB module in the V_{in} – V_{out} plane. The feasible operating area is first defined by the minimum and maximum values of the input voltage ($V_{in,min}$, $V_{in,max}$) and the output voltage ($V_{out,min}$, $V_{out,max}$) listed in Table 1.

Within the admissible voltage operating window, the converter operation is primarily constrained by the maximum allowable input current $I_{in,max} = 25$ A, and the maximum transferable power $P_{max} = 15$ kW. Although the maximum output current is specified as $I_{out,max} = 25$ A, this limitation is practically not independently utilized, since the minimum output voltage is limited to $V_{out,min} = 600$ V. Under these conditions, the maximum power limitation directly results in:

$$I_{out} = \frac{P_{max}}{V_{out,min}} = \frac{15 \text{ kW}}{600 \text{ V}} = 25 \text{ A} \quad (1)$$

Consequently, the output current limit is inherently satisfied by the power limitation and does not introduce an additional independent operating constraint.

As a result, the admissible operating area is divided into two distinct regions. For lower input voltages ($V_{in} < 600$ V), the converter operation is limited by the maximum input current (Figure 4a). In contrast, for higher input voltages ($V_{in} > 600$ V), the operating region becomes limited by the maximum transferable power (Figure 4c).

For completeness, Figure 4b presents the maximum output current values corresponding to the analyzed operating conditions. Figure 4d illustrates the overall feasible operating region of the TSBB converter, including both buck and boost operating modes.

The figure also shows all operating constraints imposed on the converter, along with the 15 experimentally investigated operating points labeled A–O.

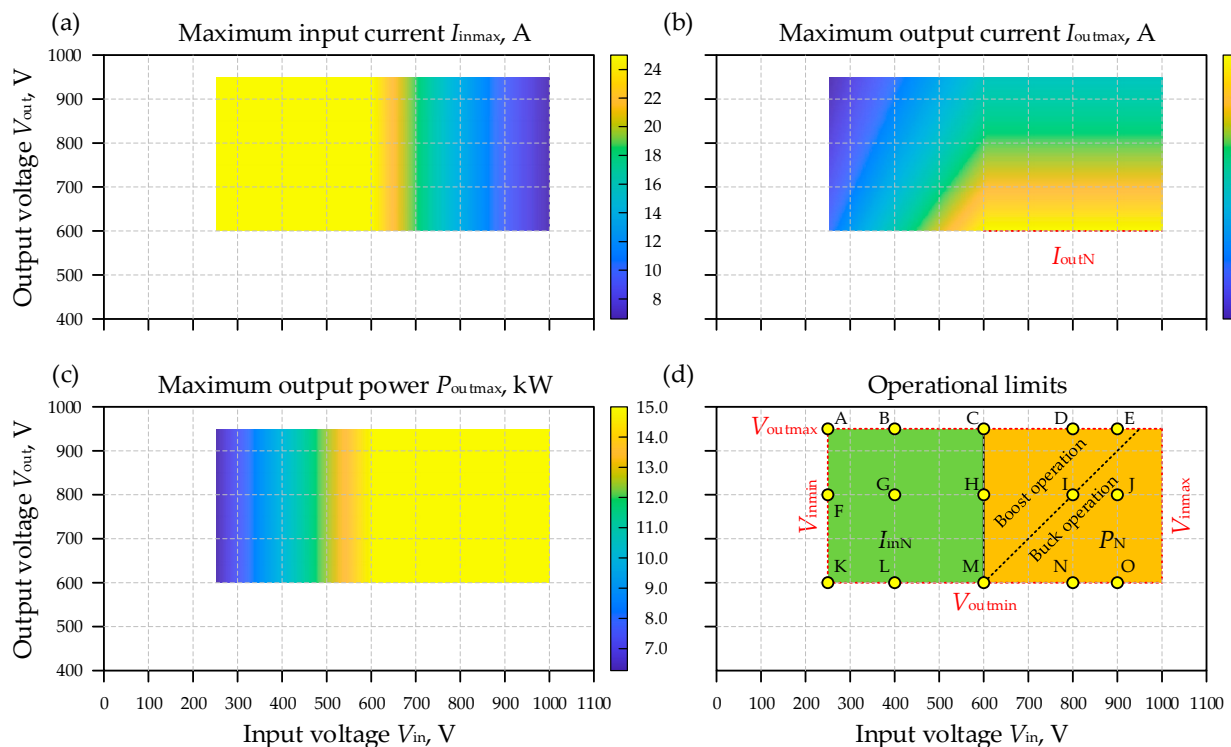


Figure 4. Operating limits and feasible operating regions of a single 15 kW TSBB module represented in the V_{in} - V_{out} plane. The plots show: (a) the maximum allowable input current, (b) the corresponding maximum output current values, (c) the maximum transferable output power assuming an ideal converter efficiency of $\eta = 100\%$, and (d) the feasible operating regions including the experimentally investigated operating points A–O. In (d), the green region denotes the operating area limited by the maximum input current, whereas the orange region denotes the operating area limited by the maximum output power. The dashed line indicates the boundary between boost and buck operation.

3. Analysis of Converter Operation

3.1. Buck Converter Mode

The converter operation is first analyzed in buck mode, where $V_{in} > V_{out}$. In this operating mode, transistor T_1 operates as the actively switched device, while transistor T_2 remains permanently turned off. The corresponding duty cycle, D_{BUCK} , is defined by (2).

$$D_{BUCK} = \begin{cases} \frac{V_{out}}{V_{in}} & \text{for } I_{out} \geq I_{out,crit} \\ \sqrt{\frac{I_{out}}{I_{outref}} \frac{V_{out}^2}{V_{in}(V_{in} - V_{out})}} & \text{for } I_{out} < I_{out,crit} \end{cases} \quad (2)$$

where the applied expression depends on the output current. For $I_{out} \geq I_{out,crit}$, the converter operates in continuous conduction mode (CCM), whereas for $I_{out} < I_{out,crit}$, it operates in discontinuous conduction mode (DCM). The critical current $I_{out,crit}$ is given by

$$I_{out,crit} = I_{outref} \left(1 - \frac{V_{out}}{V_{in}} \right) \quad (3)$$

where the reference output current $I_{out,ref}$ is defined as

$$I_{out,ref} = \frac{V_{out} T_S}{2L} \quad (4)$$

Based on the inductor voltage in the switching intervals, the peak-to-peak inductor current ripple is given by

$$\Delta I_{LBUCK} = \frac{(V_{in} - V_{out})}{L} D_{BUCK} T_S \quad (5)$$

This expression holds for both CCM and DCM, as it depends only on the inductor voltage during the on-state interval. The obtained ripple is further used to determine the current waveforms and their rms values, which are required to calculate conduction losses in semiconductor devices and passive components. In addition, the average diode currents are determined, as required by the adopted diode power-loss model. The average currents of diodes D_1 and D_2 are expressed as

$$I_{D1,AV} = \begin{cases} I_{out} D_{DBUCK} & \text{for } I_{out} \geq I_{out,crit} \\ \frac{\Delta I_{LBUCK}}{2} D_{DBUCK} & \text{for } I_{out} < I_{out,crit} \end{cases} \quad (6)$$

$$I_{D2,AV} = I_{out} \quad (7)$$

The conduction duty ratio of diode D_1 , denoted as D_{DBUCK} , depends on the operating mode. In DCM, it is given by

$$D_{DBUCK} = \frac{(V_{in} - V_{out})}{V_{out}} D_{BUCK} \quad (8)$$

In CCM, it reduces to

$$D_{DBUCK} = 1 - D_{BUCK} \quad (9)$$

The rms currents of the semiconductor devices and the inductor can be expressed as

$$I_{T1,rms} = \begin{cases} \sqrt{\left(I_{out}^2 + \frac{\Delta I_{LBUCK}^2}{12} \right) D_{BUCK}} & \text{for } I_{out} \geq I_{out,crit} \\ \Delta I_{LBUCK} \sqrt{\frac{D_{BUCK}}{3}} & \text{for } I_{out} < I_{out,crit} \end{cases} \quad (10)$$

$$I_{D1,rms} = \begin{cases} \sqrt{\left(I_{out}^2 + \frac{\Delta I_{LBUCK}^2}{12} \right) D_{DBUCK}} & \text{for } I_{out} \geq I_{out,crit} \\ \Delta I_{LBUCK} \sqrt{\frac{D_{DBUCK}}{3}} & \text{for } I_{out} < I_{out,crit} \end{cases} \quad (11)$$

$$I_{L,rms} = \begin{cases} \sqrt{\left(I_{out}^2 + \frac{\Delta I_{LBUCK}^2}{12} \right)} & \text{for } I_{out} \geq I_{out,crit} \\ \Delta I_{LBUCK} \sqrt{\frac{D_{BUCK} + D_{DBUCK}}{3}} & \text{for } I_{out} < I_{out,crit} \end{cases} \quad (12)$$

Since diode D_2 conducts the inductor current over the entire conduction interval, its rms current is equal to the rms current of the inductor, i.e., $I_{D2,rms} = I_{L,rms}$.

To illustrate the converter operation in buck mode, the characteristic current waveforms within one switching period are shown in Figure 5 for both CCM and DCM conditions.

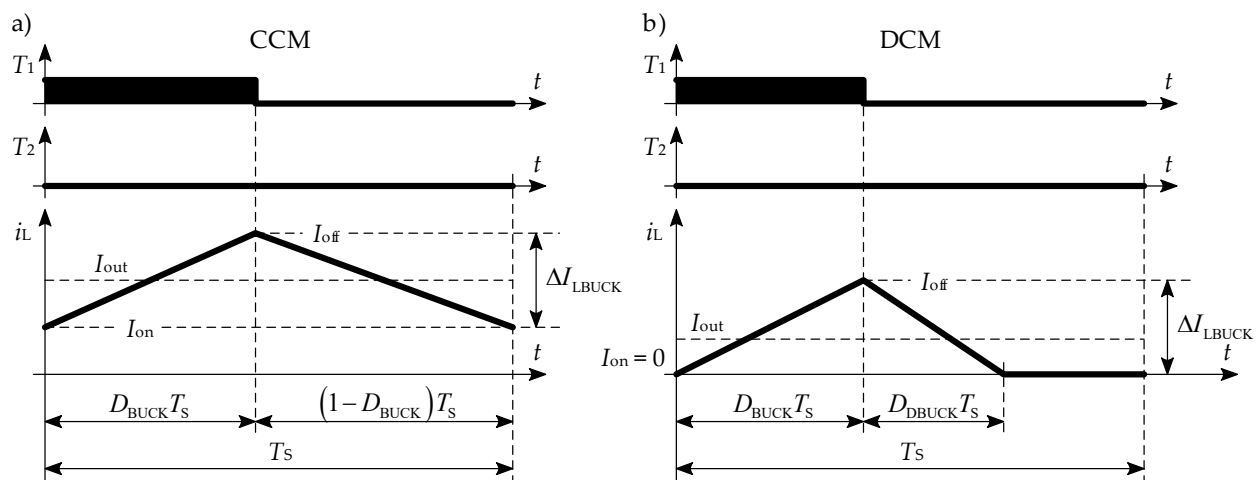


Figure 5. Characteristic current waveforms of the converter operating in buck mode: (a) continuous conduction mode (CCM); (b) discontinuous conduction mode (DCM).

As shown in Figure 5, in CCM, the inductor current remains continuous and varies around its average value, which equals the output current, I_{out} . In contrast, in DCM, the current starts from zero, increases during the on-state interval, and returns to zero during the freewheeling interval.

The conduction intervals of transistor T_1 and diode D_1 directly follow from the inductor current waveform and correspond to the duty ratios D_{BUCK} and D_{DBUCK} , respectively. These waveforms form the basis for the derivation of the average and rms currents presented in the previous section.

The current values I_{on} and I_{off} marked in Figure 5 correspond to the transistor current at turn-on and turn-off, respectively. These values are obtained directly from the inductor current waveform and used to calculate switching losses using the datasheet energy characteristics. In CCM, I_{on} corresponds to the minimum value of the inductor current, whereas I_{off} corresponds to its maximum value. In DCM, the transistor turns on at zero current and turns off at the peak inductor current. Therefore, these currents are expressed as:

$$I_{on} = \begin{cases} I_{out} - \frac{1}{2} \Delta I_{LBUCK} & \text{for } I_{out} \geq I_{out,crit} \\ 0 & \text{for } I_{out} < I_{out,crit} \end{cases} \quad (13)$$

$$I_{off} = \begin{cases} I_{out} + \frac{1}{2} \Delta I_{LBUCK} & \text{for } I_{out} \geq I_{out,crit} \\ \Delta I_{LBUCK} & \text{for } I_{out} < I_{out,crit} \end{cases} \quad (14)$$

These relations complete the description of the current waveforms in buck operation and provide the quantities required for the subsequent analysis of semiconductor currents and switching losses.

3.2. Boost Converter Mode

Next, the converter operation is analyzed for the boost mode, corresponding to $V_{in} < V_{out}$. In this operating mode, transistor T_2 is actively switched, while transistor T_1 remains turned on. The switching of transistor T_2 is performed with a duty cycle D_{BOOST} , defined as

$$D_{BOOST} = \begin{cases} 1 - \frac{V_{in}}{V_{out}} & \text{for } I_{out} \geq I_{out,crit} \\ \sqrt{\frac{I_{out}}{I_{out,ref}} \frac{V_{out}}{V_{in}} (V_{out} - V_{in})} & \text{for } I_{out} < I_{out,crit} \end{cases} \quad (15)$$

where the applied expression depends on the output current. For $I_{out} \geq I_{out,crit}$, the converter operates in continuous conduction mode (CCM), whereas for $I_{out} < I_{out,crit}$, it operates in discontinuous conduction mode (DCM). The critical current $I_{out,crit}$ is given by

$$I_{out,crit} = I_{out,ref} \left(1 - \frac{V_{in}}{V_{out}} \right) \left(\frac{V_{in}}{V_{out}} \right)^2 \quad (16)$$

where $I_{out,ref}$ is defined in (4). Based on the inductor voltage in the switching intervals, the peak-to-peak inductor current ripple is given by

$$\Delta I_{LBOOST} = \frac{V_{in}}{L} D_{BOOST} T_S \quad (17)$$

This expression holds for both CCM and DCM, as it depends only on the inductor voltage during the on-state interval. The average currents of diodes D_1 and D_2 are expressed as

$$I_{D1,AV} = 0 \quad (18)$$

$$I_{D2,AV} = I_{out} \quad (19)$$

The conduction duty ratio of diode D_2 , denoted as D_{DBOOST} , depends on the operating mode. In DCM, it is given by

$$D_{DBOOST} = \frac{V_{in}}{V_{out} - V_{in}} D_{BOOST} \quad (20)$$

In CCM, it reduces to

$$D_{DBOOST} = 1 - D_{BOOST} \quad (21)$$

The following analytical current expressions are derived using the ideal lossless power-balance relation between the input and output sides of the converter, which allows the input-side inductor current to be expressed with respect to the output current.

The rms currents of the semiconductor devices and the inductor can be expressed as (22)–(24).

$$I_{T2,rms} = \begin{cases} \sqrt{\left(\left(I_{out} \frac{V_{out}}{V_{in}} \right)^2 + \frac{\Delta I_{LBOOST}^2}{12} \right) D_{BOOST}} & \text{for } I_{out} \geq I_{out,crit} \\ \Delta I_{LBOOST} \sqrt{\frac{D_{BOOST}}{3}} & \text{for } I_{out} < I_{out,crit} \end{cases} \quad (22)$$

$$I_{D2,rms} = \begin{cases} \sqrt{\left(I_{out} \frac{V_{out}}{V_{in}} \right)^2 + \frac{\Delta I_{LBOOST}^2}{12}} D_{DBoost} & \text{for } I_{out} \geq I_{out,crit} \\ \Delta I_{LBOOST} \sqrt{\frac{D_{DBoost}}{3}} & \text{for } I_{out} < I_{out,crit} \end{cases} \quad (23)$$

$$I_{L,rms} = \begin{cases} \sqrt{\left(I_{out} \frac{V_{out}}{V_{in}} \right)^2 + \frac{\Delta I_{LBOOST}^2}{12}} & \text{for } I_{out} \geq I_{out,crit} \\ \Delta I_{LBOOST} \sqrt{\frac{D_{DBoost} + D_{DBoost}}{3}} & \text{for } I_{out} < I_{out,crit} \end{cases} \quad (24)$$

Since transistor T_1 conducts the inductor current over the entire conduction interval, its rms current is equal to the rms current of the inductor, i.e., $I_{T1,rms} = I_{L,rms}$.

To illustrate the converter operation in boost mode, the inductor current waveforms within one switching period are shown in Figure 6 for both CCM and DCM conditions. In this mode, transistor T_1 remains continuously turned on, whereas transistor T_2 is the actively switched device.

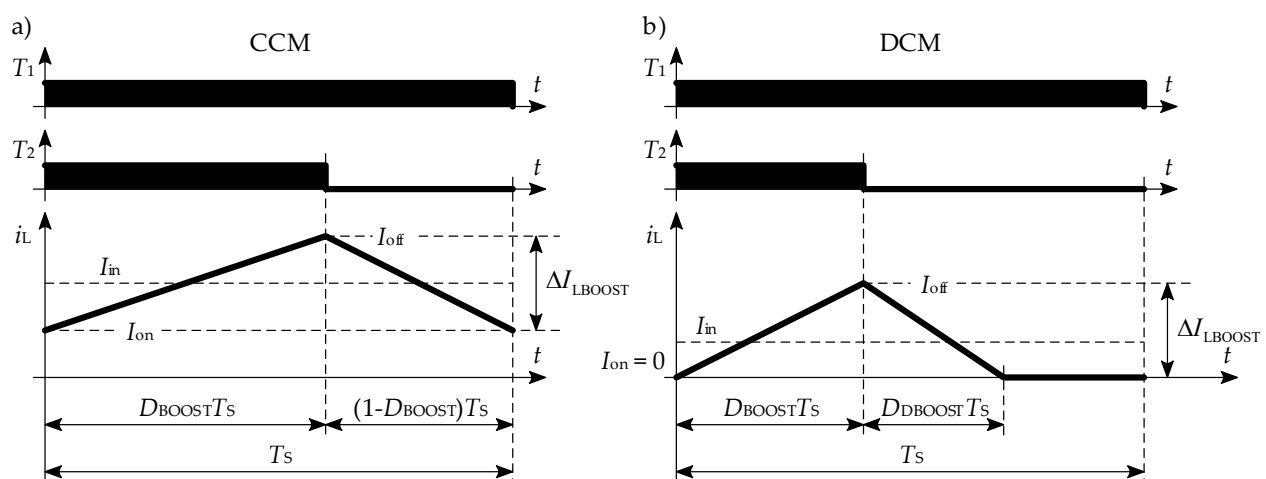


Figure 6. Characteristic current waveforms of the converter operating in boost mode: (a) continuous conduction mode (CCM); (b) discontinuous conduction mode (DCM).

As shown in Figure 6, turn-on and turn-off currents of T_2 are determined from the inductor current waveform in the same manner as in buck operation. In CCM, I_{on} corresponds to the minimum inductor current, and I_{off} to its maximum. In contrast, in DCM, the turn-on current is zero, and the turn-off current corresponds to the peak inductor current. These relations summarize the current waveforms in boost operation and will be used in the subsequent analysis of the mode-dependent current stresses.

$$I_{on} = \begin{cases} I_{out} \frac{V_{out}}{V_{in}} - \frac{1}{2} \Delta I_{LBOOST} & \text{for } I_{out} \geq I_{out,crit} \\ 0 & \text{for } I_{out} < I_{out,crit} \end{cases} \quad (25)$$

$$I_{\text{off}} = \begin{cases} I_{\text{out}} \frac{V_{\text{out}}}{V_{\text{in}}} + \frac{1}{2} \Delta I_{\text{LBOOST}} & \text{for } I_{\text{out}} \geq I_{\text{out,crit}} \\ \Delta I_{\text{LBOOST}} & \text{for } I_{\text{out}} < I_{\text{out,crit}} \end{cases} \quad (26)$$

The relations derived in Sections 3.1 and 3.2 are therefore used in Section 4 to identify the conducting and actively switched semiconductor devices in buck and boost operation.

4. Power Loss Analysis

4.1. Assumptions of the Loss Model

The power loss analysis was performed to determine the losses of the investigated TSBB converter over the considered operating range. The model focuses on power losses in individual semiconductor devices and in the inductor. This is important because the semiconductor devices responsible for current conduction and high-frequency switching depend on the converter operating mode.

As shown in Section 3, the TSBB converter can operate in buck or boost mode. In these modes, different semiconductor devices conduct current, and different devices are switched at the switching frequency. Therefore, the loss calculation was carried out separately for buck and boost operation, using the current relations derived in the previous section. The analysis was performed for steady-state operation. For a given voltage operating point, the transferred power may vary within the admissible range; therefore, the losses were evaluated as functions of the corresponding current values rather than as a single value assigned to the voltage point. The rms and average currents required for the loss calculation were determined using the analytical current relations of the converter. The numerical calculations and loss-map generation were performed using MATLAB R2023b (MathWorks, Natick, MA, USA).

Voltage sources were used to represent the input and output sides of the converter in the analytical loss model. Consequently, losses in the input and output capacitors were not included in the analysis. This simplification was adopted because, under these conditions, the ripple current is shared between the local capacitor banks and the external voltage sources, and its distribution depends on the source impedance, cable impedance, and the internal impedance of the capacitor banks. A detailed capacitor-loss model would therefore require an additional model of the complete external test setup, which was outside the scope of this work. Preliminary ESR-based estimations indicated that these losses were small compared with the semiconductor and inductor losses. Therefore, the adopted loss model includes semiconductor conduction and switching losses, as well as inductor losses. Semiconductor conduction losses were calculated using the device current values and datasheet-based on-state characteristics. Switching losses were calculated only for the devices commutated at the switching frequency in the corresponding operating mode, using the datasheet switching-energy characteristics [20,21].

Inductor losses were divided into winding losses and core losses. Winding losses were calculated from the rms inductor current and the assumed constant winding resistance. The inductor core losses were determined experimentally as the difference between the measured total inductor losses and the calculated winding losses.

A complete electrothermal model was not developed in this work. This simplification is acceptable because, during the experimental tests, the temperature rise of the heatsink and the inductor did not exceed approximately 50 °C. Therefore, the model was formulated as a practical loss model to evaluate converter power losses under different operating conditions.

4.2. Semiconductor Losses

Semiconductor losses were analyzed separately for buck and boost operation, because different devices participate in current conduction and high-frequency switching in each operating mode. In both modes, the semiconductor losses include conduction and switching components; however, their distribution among the devices is mode-dependent.

During boost operation, power losses occur in three semiconductor devices: transistors T_1 and T_2 and diode D_2 . Transistor T_1 remains continuously turned on and conducts the inductor current, whereas transistor T_2 is switched at a duty cycle D_{BOOST} . Diode D_2 conducts the inductor current during the off-state interval of T_2 , while diode D_1 does not conduct in this mode. During buck operation, power losses are also generated in three semiconductor devices: transistor T_1 , diode D_1 , and diode D_2 . In this mode, transistor T_1 is switched with the duty cycle D_{BUCK} , while transistor T_2 remains off and does not contribute to power losses. Diode D_1 conducts the switched current, whereas diode D_2 conducts the inductor current.

For this reason, conduction and switching losses were calculated using separate current relations for buck and boost operation. Conduction losses were determined from the average and rms currents of the conducting devices, whereas switching losses were calculated only for the device switched at the switching frequency in each operating mode.

4.2.1. Conduction Losses

Conduction losses were calculated from the average and rms currents of the semiconductor devices in buck and boost operation. The current values were obtained from the analytical relations presented in Section 3.

For MOSFETs, the conduction losses were calculated using the on-state resistance:

$$P_{\text{cond,TX}} = R_{\text{DS(on)}} I_{\text{TX,rms}}^2 \quad (27)$$

where $R_{\text{DS(on)}}$ is the drain-source on-state resistance of transistor T_X and $I_{\text{TX,rms}}$ is its rms current.

For diodes, the conduction characteristic was approximated by a threshold voltage, V_{F0} , and a differential resistance, r_D . Therefore, the conduction losses were calculated as:

$$P_{\text{cond,DX}} = V_{F0} I_{\text{DX,AV}} + r_D I_{\text{DX,rms}}^2 \quad (28)$$

where $I_{\text{DX,AV}}$ is the average diode current, and $I_{\text{DX,rms}}$ is the rms diode current.

4.2.2. Switching Losses

Switching losses were calculated only for the semiconductor devices involved in high-frequency commutation. In each operating mode, switching losses occur in one transistor and one diode. In boost operation, they are associated with transistor T_2 and diode D_2 , whereas in buck operation, they are associated with transistor T_1 and D_1 .

The switching losses of transistor T_X were calculated from the turn-on and turn-off energies obtained from the datasheet characteristics:

$$P_{\text{sw,TX}} = f_s k_V [E_{\text{on}}(I_{\text{on}}) + E_{\text{off}}(I_{\text{off}})] \quad (29)$$

where f_s is the switching frequency, E_{on} and E_{off} are the turn-on and turn-off energies, and I_{on} and I_{off} are the transistor currents at turn-on and turn-off, respectively. These current values were determined from the analytical current relations derived in Section 3. The coefficient k_V accounts for the difference between the actual switching voltage and the reference voltage $V_{\text{ref}} = 600$ V used in the datasheet [20]. For transistor T_1 , switched in buck operation, $k_V = V_{\text{in}}/V_{\text{ref}}$. For transistor T_2 , in boost operation, $k_V = V_{\text{out}}/V_{\text{ref}}$.

In this work, the transistor switching energies were not identified by a separate double-pulse test. The laboratory prototype was designed for converter-level operation, whereas direct measurement of the transistor switching energy would require modifications of the power PCB and the current-measurement path to measure the transistor current during the switching transient with sufficient bandwidth and accuracy. Therefore, the switching losses were calculated using the datasheet energy characteristics. This approach was considered acceptable because the gate-driver voltage, gate resistance, switching voltage, and switching current ranges in the experimental setup were selected to closely match the device manufacturer's reference test conditions. Consequently, the datasheet-based switching-energy characteristics are expected to provide a reasonable approximation of the actual transistor switching losses. However, deviations may still occur due to layout parasitics, thermal conditions, and differences between the converter operation and the datasheet test setup.

The switching losses of the SiC diodes were calculated from the capacitive charge characteristic $Q_C(V_D)$ provided in the diode datasheet [21]. Since SiC diodes exhibit negligible reverse-recovery effects, the diode switching losses were associated with charging and discharging of the diode junction capacitance during high-frequency commutation.

The energy related to the capacitive charge of diode D_x was approximated as:

$$E_{C,Dx} = \frac{1}{2} V_D Q_C(V_D) \quad (30)$$

The corresponding diode switching losses were then calculated as:

$$P_{sw,Dx} = f_s E_{C,Dx} \quad (31)$$

In the buck operation, this calculation was applied to diode D_1 , with $V_D = V_{in}$. In the boost operation, it was applied to D_2 , with $V_D = V_{out}$. The remaining diodes were not included in the switching-loss calculation because they were not involved in high-frequency commutation in the corresponding operating mode.

4.3. Inductor Losses

The total inductor losses were experimentally determined under different operating conditions. Since the measured losses include both winding losses and core losses, the winding contribution was first calculated and subtracted from the measured value. The winding losses $P_{L,w}$ were calculated from the rms value of the inductor current $I_{L,rms}$, and the measured dc winding resistance R_L :

$$P_{L,w} = R_L I_{L,rms}^2 \quad (32)$$

In the calculations, the winding resistance was assumed constant at $R_L = 43 \text{ m}\Omega$.

The remaining part of the measured losses was attributed to the inductor core losses:

$$P_{L,core} = P_{L,meas} - P_{L,w} \quad (33)$$

where $P_{L,meas}$ is the measured total inductor loss.

The obtained values of $P_{L,core}$ were first analyzed as a function of the peak-to-peak ripple current ΔI_L with the average inductor current $I_{L,av}$ used as a parameter, as is shown in Figure 7a. Since the ripple current is proportional to the flux-density variation in the inductor core, the measured losses were approximated using a Steinmetz-type power function, following the common approach used for magnetic core-loss estimation under nonsinusoidal excitation and dc-bias conditions [22,23].

$$P_{L,core} = k_c (I_{L,av}) \Delta I_L^\beta \quad (34)$$

The exponent β was determined from experimental data and found to be 1.77. For each value of the average inductor current, the proportionality coefficient $k_c(I_{L,av})$ was identified from the measured loss data. The obtained values of $k_c(I_{L,av})$ were then approximated by a quadratic function of the average inductor current, as expressed in (35). The resulting approximation is shown in Figure 7b.

$$k_c(I_{L,av}) = aI_{L,av}^2 + bI_{L,av} + c \quad (35)$$

Finally, the inductor core losses used in the converter loss model were calculated as:

$$P_{L,core} = \left(\left(0.0012 \frac{W}{A^2} \right) I_{L,av}^2 - \left(0.0245 \frac{W}{A} \right) I_{L,av} + 0.55 W \right) \left(\frac{\Delta I_L}{A} \right)^{1.77} \quad (36)$$

This formulation allows estimation of the inductor core losses at any converter operating point, given the average inductor current $I_{L,av}$, and the peak-to-peak inductor current ripple ΔI_L .

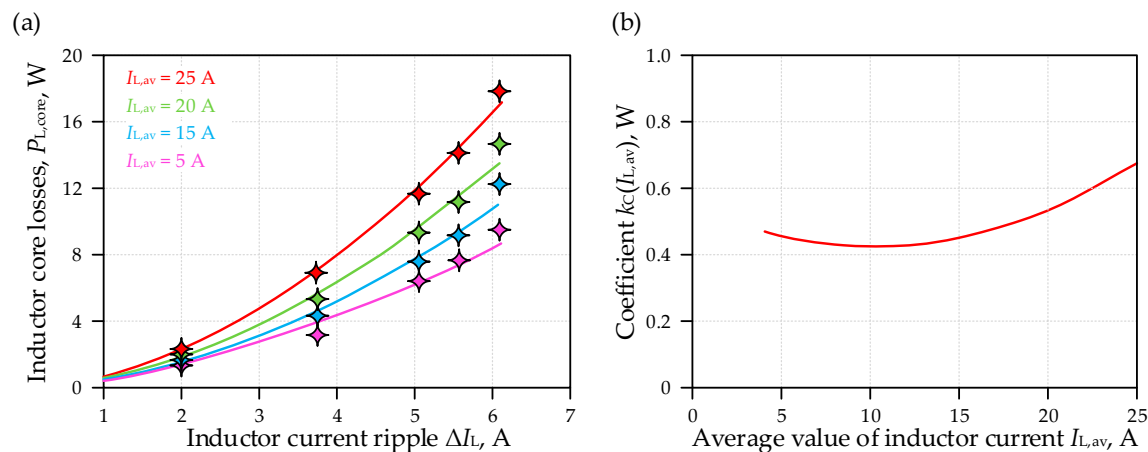


Figure 7. Identification of the inductor L core-loss model; (a) measured and fitted core losses versus inductor current ripple for selected average current values; (b) quadratic approximation of the coefficient $k_c(I_{L,av})$. In (a), the star markers indicate measured data points, while the solid lines represent the fitted curves

5. Experimental Verification and Efficiency Analysis

The experimental verification was performed in two stages. First, a single 15 kW TSBB module was tested at the operating points A–O defined in Section 2.5 to validate the proposed power-loss model, including semiconductor and inductor losses. Then, the complete five-phase converter was tested under high-power conditions to confirm the scalability of the modular structure and the applicability of the single-module loss model to the 75 kW system. Detailed loss comparisons are presented for selected representative operating points, while the full set of measurements is used to evaluate the efficiency trend over the analyzed operating range.

In all efficiency measurements presented in this section, the laboratory DC sources were connected to the converter with the PV-side Pi-type EMI filters bypassed. This allowed the losses of the main TSBB power stage to be evaluated without including the additional losses and parasitic effects of the EMI filters.

5.1. Single-Module 15 kW Experimental Test

The first experimental stage was performed using a single 15 kW TSBB module. The measurements were carried out at the operating points A–O defined in Section 2.5,

covering both buck and boost operation. For each point, the input and output voltages, currents, powers, converter efficiency, and total power losses were determined.

The measured losses were compared with the losses calculated using the model described in Section 4, including semiconductor conduction and switching losses, inductor winding losses, and experimentally identified inductor core losses.

Although measurements were performed at all operating points A–O, four representative cases were selected for detailed presentation: K, G, C, and O. Points K, G, and C correspond to boost operation and represent low-input-voltage, typical, and high-output-voltage operating conditions, respectively. Point O represents a clearly defined buck operating condition. These points were selected away from the immediate buck–boost transition boundary to ensure a clear interpretation of the loss mechanisms.

The experimental waveforms recorded for the selected operating points are shown in Figure 8. Points K, G, and C correspond to boost operation, whereas point O represents buck operation. The waveforms illustrate the inductor current and the corresponding voltage under the same operating conditions used for the power-loss measurements.

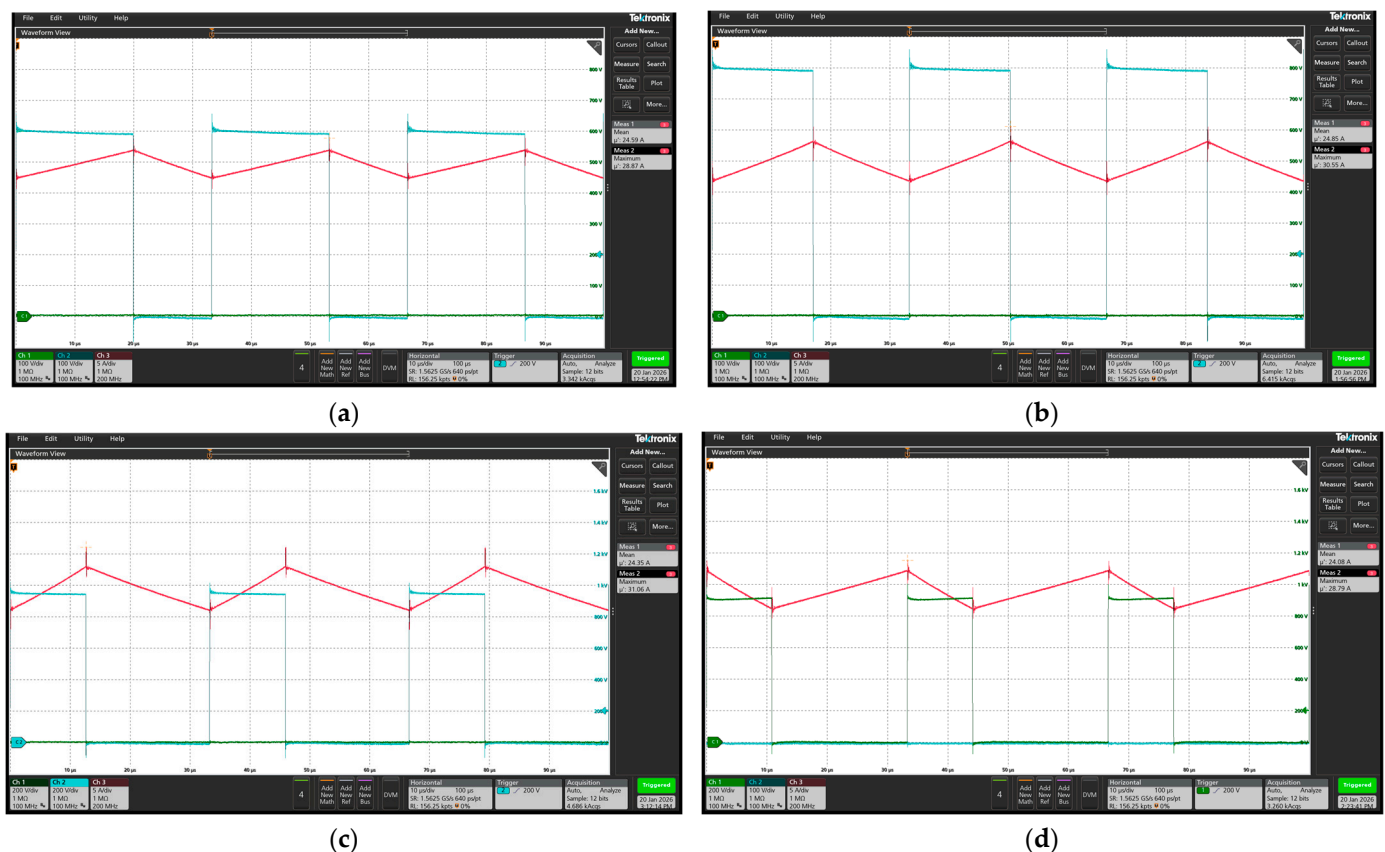


Figure 8. Measured waveforms illustrating the operation of the 5P-TSBB converter in different operating points: (a) point K, $V_{in} = 250$ V, $V_{out} = 600$ V; (b) point G, $V_{in} = 400$ V, $V_{out} = 800$ V; (c) point C, $V_{in} = 600$ V, $V_{out} = 950$ V; and (d) point O, $V_{in} = 900$ V, $V_{out} = 600$ V. Ch1 (green)—voltage across transistor T_1 ; Ch2 (blue)—voltage across diode D_2 ; and Ch3 (red)—inductor current i_L .

The measured waveforms shown in Figure 8 confirm the expected operation of the converter in both operating modes. In the boost mode, transistor T_1 remains continuously turned on; therefore, the voltage across T_1 is close to zero. In buck mode, diode D_2 conducts continuously, which is indicated by the voltage across D_2 remaining close to zero. Thus, the presented waveforms enable verification of the conduction states of the selected semiconductor devices and their relationship to the inductor current waveform.

Figure 9 presents the measured and calculated power-loss characteristics for the selected operating points. In all cases, the calculated losses follow the measured characteristics with satisfactory accuracy. The remaining discrepancies can be attributed mainly to the simplified thermal treatment of the semiconductor parameters, the assumption of constant winding resistance, and the use of datasheet-based switching-energy characteristics. Although the switching conditions in the prototype were selected to be close to the datasheet reference conditions, the actual switching losses may still differ due to layout parasitics, device temperature, and the fact that the switching energies were not measured directly in the converter prototype.

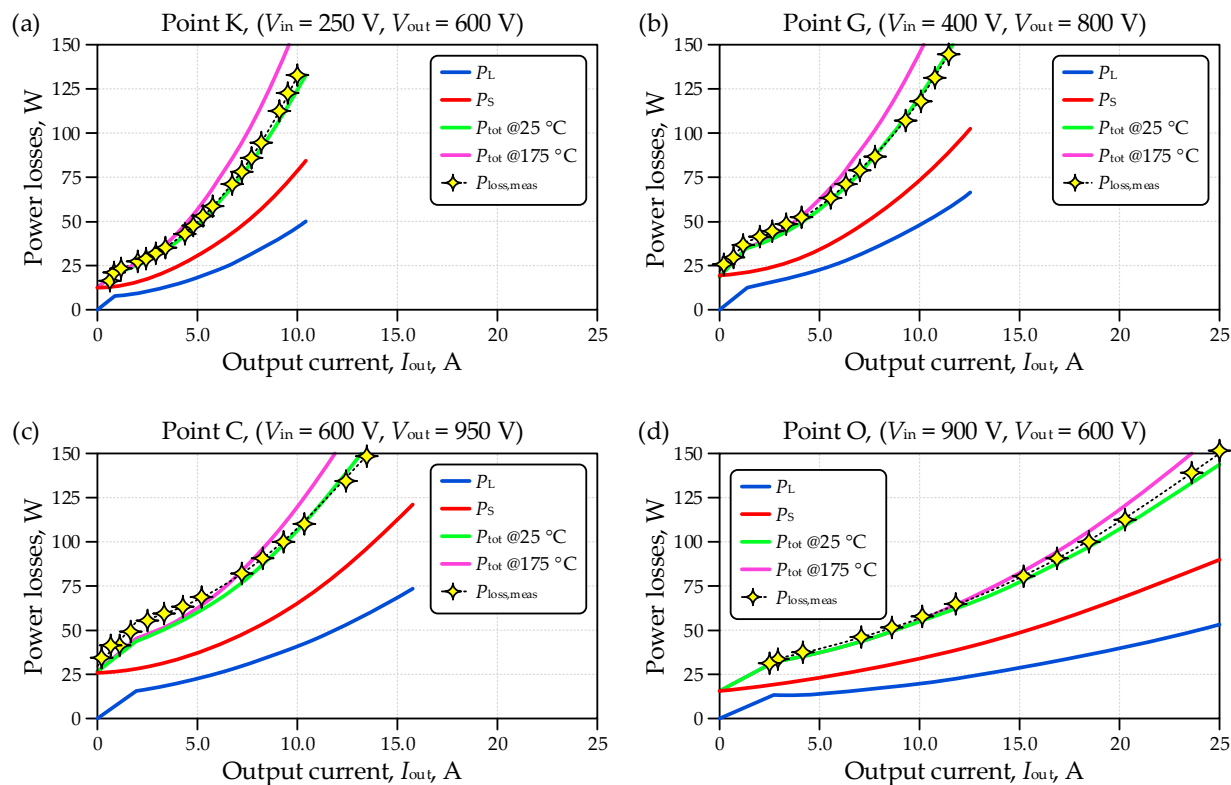


Figure 9. Experimental verification of the single-module loss model for four representative operating points: (a) point K, $V_{in} = 250$ V, $V_{out} = 600$ V; (b) point G, $V_{in} = 400$ V, $V_{out} = 800$ V; (c) point C, $V_{in} = 600$ V, $V_{out} = 950$ V; and (d) point O, $V_{in} = 900$ V, $V_{out} = 600$ V. The calculated inductor losses P_L , semiconductor losses P_S , and total converter losses P_{tot} are compared with the measured total losses $P_{loss, meas}$.

As shown in Figure 9, good agreement between the measured and calculated losses is observed across all selected operating points. In most cases, the measured total losses are higher than the calculated total losses at $T = 25$ °C but lower than those at $T = 175$ °C. This is consistent with the expected temperature dependence of semiconductor conduction losses and confirms that the adopted loss model provides a realistic estimate of the converter losses over the analyzed current range.

The inductor losses are significant, reaching approximately 75 W in the analyzed cases. However, the inductor losses remain lower than the total semiconductor losses over the entire current range presented. This confirms that both loss components must be included in the model, although the semiconductor devices remain the dominant source of losses in the investigated operating points.

It should also be noted that the maximum output current differs between the selected operating points. Only in point O does the converter reach $I_{out} = I_{out, max} = 25$ A. In the

remaining points, the maximum output current is limited by either the maximum input current of 25 A or the maximum output power of 15 kW. Consequently, the available output current range depends on the specific operating voltage.

For very low output currents, below approximately 1–2 A, the influence of discontinuous inductor current operation is visible. This effect is particularly noticeable in the inductor losses, because the core-loss component depends strongly on the peak-to-peak inductor current ripple ΔI_L . In discontinuous conduction mode, changes in the peak-to-peak current ripple do not significantly affect the rms current value. Therefore, a similar effect is not clearly observed in the semiconductor losses, which are mainly related to the rms and average current values.

A slightly different tendency can be observed for point C. At lower output currents, the measured losses exceed the calculated values, whereas at higher currents, they become slightly lower than the calculated losses. This discrepancy may be related to simplified thermal assumptions, datasheet-based switching-energy characteristics, and the greater influence of voltage-dependent switching and parasitic effects at high output voltages.

The total losses in the presented cases do not exceed approximately 150 W, which corresponds to about 1% of the rated 15 kW output power of a single TSBB module. Therefore, the efficiency of the converter in this operating range is expected to be close to 99%.

The verified loss model was then used to evaluate the performance of a single 15 kW TSBB module over the entire analyzed voltage range. The calculations were performed in the V_{in} – V_{out} plane for the maximum allowable operating point at each voltage combination. Therefore, the output power was limited by $I_{in,max} = 25$ A, $I_{out,max} = 25$ A, or $P_{out,max} = 15$ kW, depending on the considered operating point. This approach is consistent with the feasible operating region defined in Section 2.6 and directly relates the loss distribution to the actual converter limitations.

Figure 10 shows the resulting efficiency map, total converter losses, inductor losses, and semiconductor losses calculated at $T = 25$ °C. The dashed line indicates the boundary between boost and buck operation, corresponding to $V_{in} = V_{out}$.

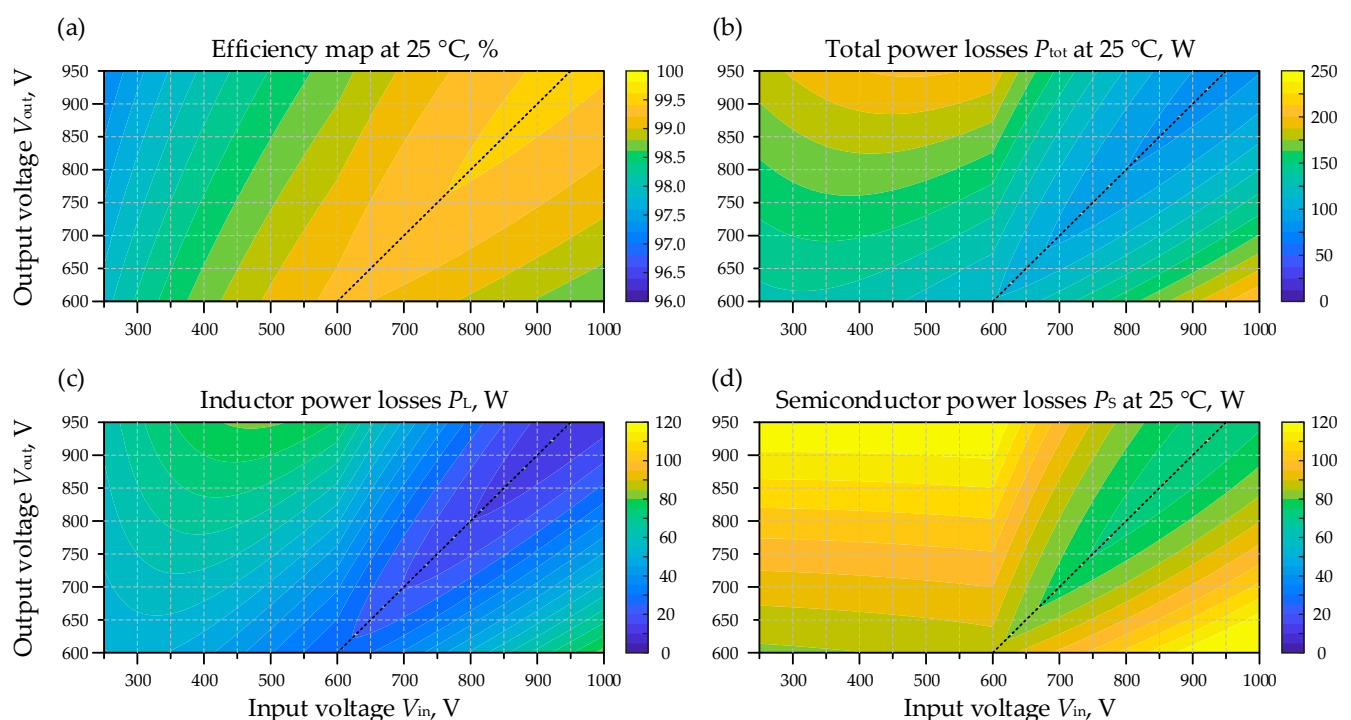


Figure 10. Calculated efficiency and power-loss maps of a single 15 kW TSBB module in the V_{in} – V_{out} plane for $T = 25$ °C: (a) efficiency map, (b) total converter losses P_{tot} , (c) inductor losses P_L , and (d)

semiconductor losses P_s . The calculations were performed at the maximum allowable operating point for each voltage pair, considering $I_{in,max} = 25$ A, $I_{out,max} = 25$ A, and $P_{out,max} = 15$ kW.

As shown in Figure 10, the converter efficiency remains high across most of the analyzed operating range, generally close to 99%. The highest efficiency is obtained in the region where both the current stress and the voltage conversion ratio remain moderate. In contrast, efficiency decreases at low input voltages, where the maximum input current and the inductor current increase significantly, thereby limiting the transferred power.

The total loss map in Figure 10b shows that the highest losses occur mainly in the low-input-voltage boost region. In this region, the converter must process relatively high input current, which increases both conduction losses in the semiconductor devices and winding losses in the inductor. The inductor-loss map in Figure 10c confirms that P_L is strongly affected by both the rms value of the inductor current and the peak-to-peak current ripple. Therefore, the inductor losses become particularly relevant in regions with high current stress and increased ripple.

The semiconductor-loss map in Figure 10d shows a different distribution. In the boost region, semiconductor losses are influenced by the current conducted by T_1 , the switching losses of T_2 , and the conduction and capacitive switching losses of D_2 . In the buck region, the loss distribution changes because T_1 becomes the actively switched transistor, while D_1 and D_2 conduct current. This confirms that the power-loss model must distinguish between buck and boost operation rather than applying a single averaged loss formulation over the entire voltage range.

To further analyze the semiconductor-loss mechanisms, the total semiconductor losses were separated into the losses generated in individual devices. This decomposition is important because the devices involved in conduction and high-frequency commutation depend directly on the operating mode of the TSBB converter. In the boost operation, losses are generated mainly in T_1 , T_2 , and D_2 , whereas D_1 remains inactive. In the buck operation, T_1 , D_1 , and D_2 contribute to the semiconductor losses, while T_2 remains turned off.

Figure 11 presents the calculated loss distribution among the semiconductor devices for $T = 25$ °C.

The obtained loss distribution clearly reflects the mode-dependent operation of the TSBB converter. Transistor T_1 generates losses over the entire operating range, but the physical origin of these losses differs between the two modes. In boost operation, T_1 remains continuously turned on, and its losses are conduction losses. In buck operation, T_1 becomes the actively switched device; therefore, its losses include both conduction and switching components.

The losses in T_2 are concentrated in boost operation, where this transistor is switched at the switching frequency. This confirms that T_2 is the main switched transistor during boost operation. In buck operation, T_2 remains inactive and does not contribute significantly to the semiconductor losses.

The losses in diode D_1 are associated mainly with buck operation. In this mode, D_1 conducts during the freewheeling interval and participates in the high-frequency commutation with transistor T_1 ; therefore, both its conduction losses and capacitive switching losses are included. In boost operation, D_1 does not conduct, and its losses are negligible.

Diode D_2 contributes to losses in both operating modes, but its role differs in each case. In the boost operation, D_2 conducts during the energy-transfer interval, when transistor T_2 is turned off. In buck operation, D_2 remains in the main current path and conducts the inductor/output current over the whole switching period, T_s . Therefore, the losses in D_2 increase with higher inductor and output currents.

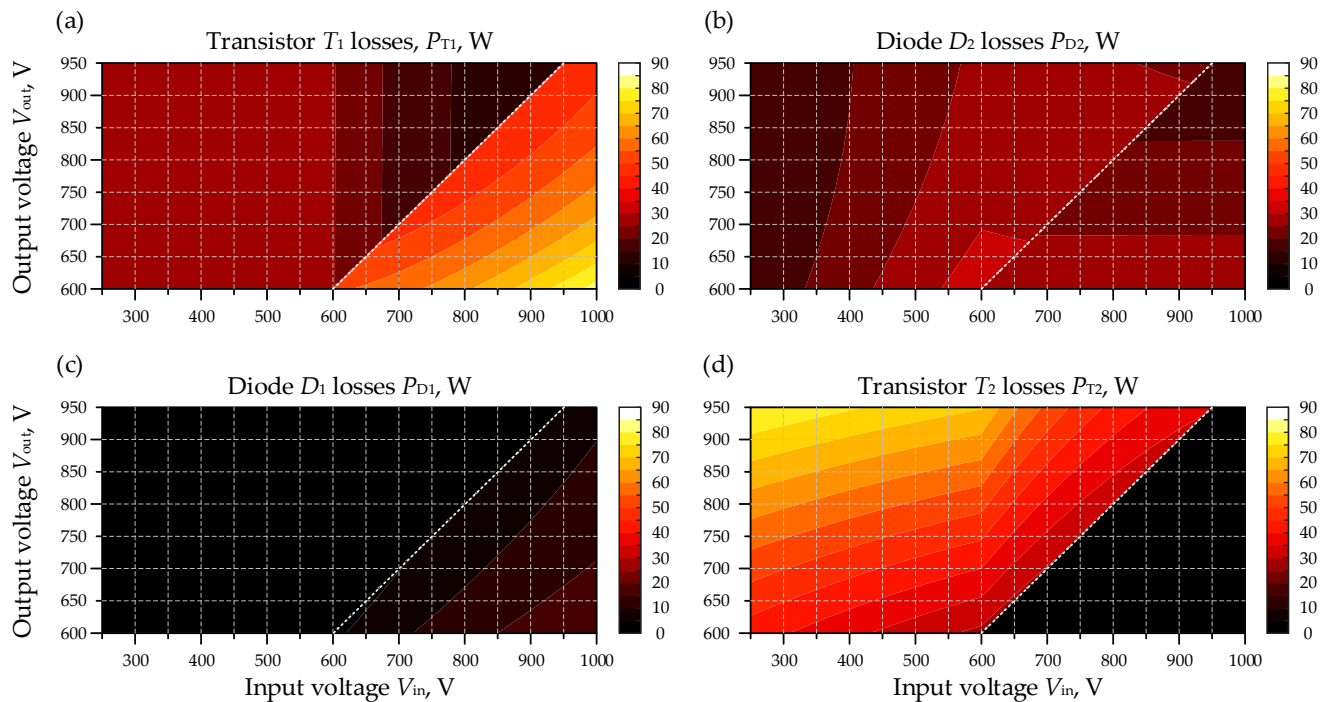


Figure 11. Calculated distribution of semiconductor losses in a single 15 kW TSBB module at $T = 25$ °C: (a) transistor T_1 losses P_{T1} , (b) diode D_2 losses P_{D2} , (c) diode D_1 losses P_{D1} , and (d) transistor T_2 losses P_{T2} . The dashed line indicates the boundary between boost and buck operation.

These results confirm that the semiconductor loss distribution cannot be described only by the total semiconductor losses, P_s . The contribution of individual devices varies substantially between buck and boost operation, which justifies separately calculating semiconductor losses for each operating mode.

The agreement obtained confirms that the proposed loss model can be used to estimate losses for a single TSBB module over the analyzed operating range. Therefore, the model is further used as the basis for evaluating the complete five-phase converter.

5.2. Extension of the Loss Model to the Five-Phase Converter

The second experimental stage was performed using the complete five-phase TSBB converter. The purpose of this test was to verify the operation of the proposed 5P-TSBB converter at the system level and to confirm that the modular architecture can be scaled from a single 15 kW TSBB module to a 75 kW five-phase converter. This converter is shown in Figure 12.

The tested converter was operated in the interleaved single-input configuration. In this mode, all five TSBB modules are supplied from the same DC source and operate with phase-shifted control signals. Such an operation reduces ripple components in the input and output currents compared with operating a single TSBB module. As a consequence, the RMS value of the ripple current flowing through the input and output capacitors can also be reduced, thereby limiting the power losses in these capacitors.

Although the converter operates as a five-phase system, its hardware structure remains modular. In the tested prototype, a single common capacitor bank was not used for the entire 5P-TSBB converter. Instead, the five-module structure was retained, using the same TSBB modules as those discussed in the previous section. Therefore, the five-phase converter can be treated as a parallel connection of five identical TSBB modules operating with interleaved control.



Figure 12. Laboratory prototype of the developed 75 kW 5P-TSBB converter.

The experimental setup used to verify the complete 5P-TSBB converter is shown in Figure 13. A circulating-power test configuration was applied to reduce the required power rating of the external DC power supply. In this setup, the output side of the converter was connected back to the input side via the auxiliary resistor R_b . As a result, most of the processed power circulated between the output and input terminals of the converter, whereas the external DC supply delivered only the converter losses and the power dissipated in the auxiliary resistor.

Two power quantities were measured during the test. The power P_1 was measured at the output of the external DC power supply, while the power P_2 was measured at the output side of the 5P-TSBB converter. The measured power P_2 represents the power processed by the converter, whereas P_1 represents the additional power supplied to the circulating-power system. The measured powers can be expressed as

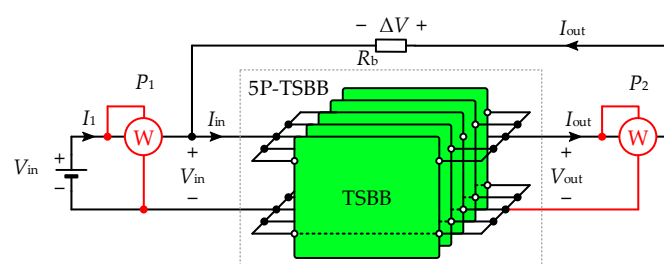


Figure 13. Experimental setup for the circulating-power test of the 5P-TSBB converter with the auxiliary resistor R_b connected between the output and input sides.

$$P_1 = I_1 \times V_{in} \quad (37)$$

$$P_2 = P_{out,5P} = I_{out} \times V_{out} \quad (38)$$

It should be noted that the power P_1 is not the total input power of the converter. In the circulating-power configuration, part of the output power is returned to the input side through the auxiliary resistor R_b . Therefore, the actual input power of the 5P-TSBB converter exceeds the power delivered by the external DC supply.

According to the current directions shown in Figure 13, the input current of the converter I_{in} can be expressed as

$$I_{in} = I_1 + I_{out} \quad (39)$$

Thus, the input power of the 5P-TSBB converter is

$$P_{in,5P} = V_{in} (I_1 + I_{out}) = P_1 + V_{in} I_{out} \quad (40)$$

The total losses of the five-phase converter can therefore be calculated from the difference between the input and output powers:

$$P_{loss,5P} = P_{in,5P} - P_{out,5P} = P_1 - P_2 + V_{in} I_{out} \quad (41)$$

The efficiency of the converter is calculated as:

$$\eta = \frac{P_{out}}{P_{in}} = \frac{I_{out} \cdot V_{out}}{(I_1 + I_{out}) \cdot V_{in}} = \frac{P_2}{P_1 + I_{out} \cdot V_{in}} \quad (42)$$

A limitation of this method is that the output voltage of the converter must be higher than the input voltage; therefore, the converter must operate in boost mode. The output voltage can be expressed as

$$V_{out} = V_{out} + \Delta V \quad (43)$$

where ΔV is the voltage drop across the auxiliary resistor R_b . The power measured at the DC power supply used in the test is the sum of the power losses in the 5P-TSBB converter and the power dissipated in the resistor R_b , i.e.,

$$P_1 = P_{loss,5P} + P_{Rb} = P_{loss,5P} + I_{out}^2 R_b \quad (44)$$

Assuming that the maximum output current $I_{out,5P,max}$ is equal to five times the maximum output current of a single module, i.e.,

$$I_{out,5P,max} = 5 \times 25 \text{ A} = 125 \text{ A} \quad (45)$$

and substituting the maximum power of the laboratory DC power supply, $P_{1max} = 15 \text{ kW}$, the maximum value of the auxiliary resistance R_b can be calculated. In this estimation, the converter losses are assumed to be negligible compared with the power dissipated in the auxiliary resistor. Since

$$P_{Rb} = R_b (I_{out,5P,max})^2 \approx P_{1max} \quad (46)$$

The auxiliary resistance R_b should fulfill the following condition:

$$R_b \leq \frac{P_{1max}}{(I_{out,5P,max})^2} = 0.96 \Omega \quad (47)$$

In the experimental setup, a lower R_b value of 0.32Ω was selected to ensure that the power drawn from the laboratory supply remained below its maximum rating.

The selected auxiliary resistance R_b limits the maximum admissible voltage difference, ΔV , between the output and input because

$$\Delta V = R_b I_{out,5P} \quad (48)$$

Thus, the maximum value of ΔV corresponds to the maximum output current. For the 5P-TSBB converter, the current and power ratings are 5 times those of a single TSBB module. Therefore, the maximum input current is 125 A, and the maximum processed power is 75 kW. The boundary between the current-limited and power-limited operating

regions is obtained as $V_{in} = 75 \text{ kW}/125 \text{ A} = 600 \text{ V}$. At $V_{in} = 600 \text{ V}$, the difference voltage can be increased to $\Delta V = 0.32 \Omega \times 125 \text{ A} = 40 \text{ V}$.

It should be noted that the TSBB converter requires a minimum output voltage of $V_{out,min} = 600 \text{ V}$. Since $V_{out} = V_{in} + \Delta V$ and the maximum available voltage difference in the test setup is 62.5 V , operation at input voltages significantly lower than 600 V is not considered in this test. For $V_{in} < 600 \text{ V}$, the operating range is limited by the maximum input current, and the processed power is lower than the rated power. Consequently, the output current and the corresponding voltage drop ΔV across R_b are also lower.

For $V_{in} > 600 \text{ V}$, operation is limited by the maximum power that can be processed. In this region, the output current decreases as the output voltage increases, and therefore, ΔV also decreases. As a result, the highest value of ΔV occurs at the boundary between these two operating regions, i.e., at $V_{in} = 600 \text{ V}$.

The admissible operating region of the 5P-TSBB converter during the circulating-power test is presented in Figure 14 in the V_{in} – V_{out} plane. The shaded area indicates the voltage combinations that satisfy the converter limitations and the constraints imposed by the test setup. The marked points correspond to the experimental operating conditions selected for verification at $V_{in} = 600 \text{ V}$, 700 V , and 800 V . These points were chosen to cover the practically relevant part of the boost-mode operating range.

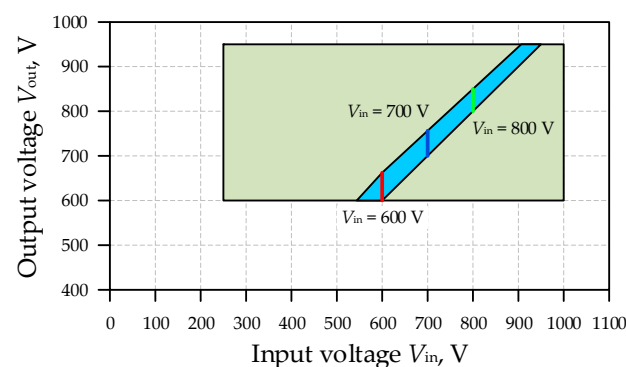


Figure 14. Admissible operating region of the 5P-TSBB converter in the V_{in} – V_{out} plane during the circulating-power test, with marked experimental test points.

The analysis presented above describes the circulating-power test configuration in a general form and defines the method used to determine the losses and efficiency of the complete 5P-TSBB converter. However, the comparison between the measured losses of the five-phase converter and the loss model developed for a single TSBB phase is valid only if the five phases are loaded in a sufficiently balanced way. In the interleaved single-input mode, this condition requires the inductor currents of the individual TSBB modules to be distributed nearly uniformly.

To verify this condition, the inductor currents of all five phases were measured during the experimental test. Figure 15 shows the measured current waveforms at the operating point with $V_{in} = 600 \text{ V}$. This point was selected because it represents operation near the rated power of the 5P-TSBB converter and corresponds to a phase current of approximately 25 A . For this operating point, the measured average inductor currents were 23.90 A , 24.18 A , 23.13 A , 23.71 A , and 23.24 A for phases 1–5, respectively. The mean value was 23.63 A . The maximum phase-to-phase difference was 1.05 A , which corresponds to approximately 4.4% of the mean value. The standard deviation was approximately 0.40 A , giving a coefficient of variation of about 1.7% . These results confirm that the current sharing between the five TSBB modules was sufficiently balanced. Therefore, the measured losses of the complete 5P-TSBB converter can be compared with the loss model previously developed for a single TSBB phase, which is extended to the five-phase system by

summing the losses of the individual phases. The observed current distribution justifies the use of this modular loss-modeling approach for the operating point under analysis.

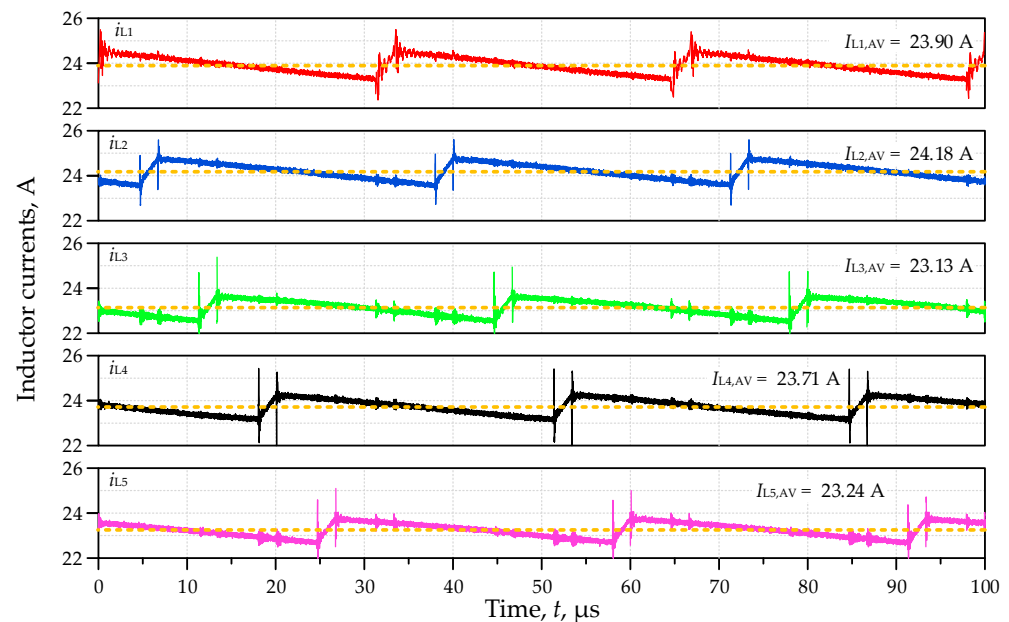


Figure 15. Measured inductor currents of the five TSBB modules during interleaved operation. The waveforms are shifted by one fifth of the switching period, confirming proper phase-shifted control and balanced current sharing. The horizontal dashed lines indicate the average values of the corresponding inductor currents, with the numerical values given in the figure.

Figure 16 shows the measurement results for the 5P-TSBB converter obtained during the circulating-power test at input voltages of $V_{in} = 600, 700$, and 800 V. The measured power losses are presented as a function of the output power and compared with the analytical loss model. The model results were obtained for a single TSBB module and then scaled by the number of operating modules ($n = 5$), with the temperature set to 25 °C.

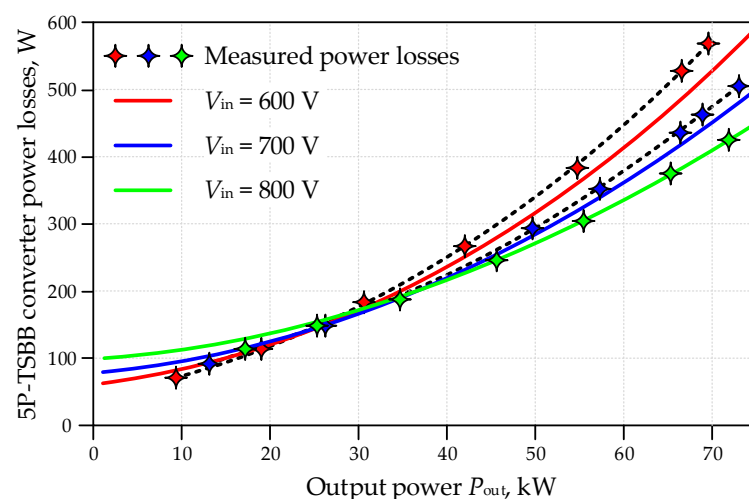


Figure 16. Measured and analytical power losses of the 5P-TSBB converter obtained in the circulating-power test for $V_{in} = 600, 700$, and 800 V. Solid lines represent analytical results, whereas dashed lines represent fitted curves based on the measured data points.

Figure 16 compares the measured power losses of the complete 5P-TSBB converter with the results obtained from the analytical loss model. A satisfactory agreement

between the experimental results and the scaled analytical model is achieved, confirming that the modular loss-modeling approach can be applied to the complete 75 kW converter. The remaining differences may result from unequal thermal conditions of the individual modules, current-sharing deviations, parasitic resistances in the experimental setup, and measurement uncertainty in the circulating-power test.

In boost operation, a higher V_{in} reduces the inductor current required for the same output power. Since this current flows through the conducting semiconductor devices and the inductor, the current stress and conduction losses decrease.

Figure 17 shows an example measurement result displayed by the Yokogawa WT5000 power analyzer during the circulating-power test of the 5P-TSBB converter. The presented result corresponds to one selected operating point from the characteristics shown in Figure 16, obtained for $V_{in} = 600$ V. In this test configuration, the power values displayed by the analyzer represent the quantities measured in the circulating-power setup.

Udc1	Udc2	----	----
598.395 V	633.204 V		
Idc1	Idc2	----	----
7.3311 A	109.670 A		
P1	P2	PLoss	η
4.38692 kW	69.4436 kW	569.383 W	99.1867 %
----	----	----	----

Figure 17. Measurement result displayed by the Yokogawa WT5000 power analyzer during the circulating-power test of the 5P-TSBB converter for a selected operating point at $V_{in} = 600$ V and $I_{out} = 109.67$ A. The displayed efficiency value corresponds to the converter efficiency calculated according to (42).

For the presented operating point (Figure 17), the output voltage V_{out} was higher than the input voltage V_{in} by the voltage drop across the auxiliary resistance R_b , i.e., $\Delta V = 109.67$ A $\cdot 0.32 \Omega = 35.1$ V.

It should be noted that the circulating-power configuration used in this work enables system-level high-power validation of the complete 5P-TSBB converter only in boost operation. In a buck operation, where $V_{out} < V_{in}$, this passive test configuration cannot return the converter output power to the input side in the same manner; therefore, an active power sink/source or a different regenerative setup would be required. Buck-mode operation was verified at the single-module level. Since the five-phase converter consists of identical TSBB modules, the buck-mode switching states, current paths, semiconductor stresses, and inductor-current relations of each phase remain unchanged. Thus, the single-module buck-mode results support the expected five-phase buck-mode behavior, although the lack of system-level buck-mode measurements is acknowledged as a limitation.

6. Conclusions

This paper presented the design, analysis, and experimental verification of a 75 kW five-phase Two-Switch Buck–Boost converter intended for photovoltaic and energy-storage applications. The converter consists of five identical 15 kW TSBB modules and can be configured either for independent multi-input operation or for interleaved single-input operation. In this work, the experimental verification included single-module buck and boost tests and complete five-module operation in the interleaved single-input

configuration. The independent multi-input configuration was discussed as a possible hardware configuration of the proposed modular converter, while its experimental validation under different input-voltage and input-power conditions is left for future work.

The feasible operating region of the converter was analyzed in the V_{in} – V_{out} plane, accounting for voltage, current, and power limitations. The analysis showed that the operation is mainly limited by the maximum input current at low input voltages and by the maximum output power at higher input voltages. The maximum output current does not introduce an additional independent limitation, because it is already satisfied by the rated power and the minimum output voltage.

A power-loss model of a single TSBB module was developed separately for buck and boost operation. The model includes semiconductor conduction and switching losses, winding losses, and experimentally identified inductor core losses. The results confirmed that the loss distribution strongly depends on the operating mode and that separate loss calculations are required for buck and boost operation.

The experimental verification of the single 15 kW module showed good agreement between calculated and measured losses for representative operating points. The obtained efficiency was generally close to 99%, with lower values mainly in the low-input-voltage boost region, where the inductor and semiconductor currents are higher.

The complete five-phase converter was also tested using a circulating-power setup. The measured losses confirmed that the results for the 75 kW converter are consistent with predictions from the single-module loss model. The measurements also showed that increasing the input voltage reduces converter losses at the same transferred power, primarily due to reduced inductor current.

The results confirm that the proposed 5P-TSBB converter is a suitable solution for high-power photovoltaic and energy-storage interfaces that require a wide operating voltage range. Future work will focus on electrothermal modeling, optimization of the magnetic components, and verification under dynamic operating conditions.

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