Dual-mode magnetically integrated photovoltaic microconverter with adaptive mode change and global maximum power point tracking

Hamed Mashinchi Maheri1 | Dmitri Vinnikov1,2 | Andrii Chub1,2 | Oleksandr Korkh1 | Argo Rosin1,2 | Ebrahim Babaei3,4

1 Department of Electrical Power Engineering and Mechatronics, Tallinn University of Technology, Tallinn 19086, Estonia
2 Smart City Center of Excellence (Finest Twins), Tallinn University of Technology, Tallinn 19086, Estonia
3 Faculty of Electrical and Computer Engineering, University of Tabriz, Tabriz 5166616471, Iran
4 Engineering Faculty, Near East University, Nicosia 99138, North Cyprus

Abstract
This study proposes a high step-up galvanically isolated dc-dc converter based on the quasi-Z-source (qZS) network. The voltage gain of the converter can change in a wide range. This feature makes the converter suitable for applications with a wide input voltage range. The range of the dc gain is increased by the implementation of the combined energy transfer principle. A reconfigurable structure is used to combine the energy transfer by the isolation transformer and the coupled inductor of the impedance-source network. The adaptive mode change activates the energy transfer via an additional winding in the qZS network at lower voltages, which results in considerable efficiency improvement. Consequently, the input voltage range of the converter is extended to the values useful in photovoltaic applications impacted by partial shading. The proposed approach is verified in a photovoltaic (PV) microconverter, where the energy harvest from a PV module is possible during shading by enabling the global maximum power point tracking (GMPPT). It is demonstrated that the dual-mode operation of the converter helps to achieve a flat efficiency curve along with outstanding maximum power point tracking (MPPT) performance.

1 INTRODUCTION

Renewable energy resources can be a viable alternative for fossil fuels due to their clean and cost-effective features [1]. Power electronic converters are the key components of renewable energy systems. High step-up dc-dc converters are becoming popular for energy harvesting from these sources in low power applications like residential power systems [2–5]. Galvanically isolated dc-dc converters are often used for high voltage step-up in photovoltaic (PV) module-level systems. For renewable energy applications where high voltage step-up and wide input voltage range are required, impedance-source galvanically isolated dc-dc converters can be a suitable choice [6–9]. In these converters, a high-frequency transformer is used to achieve high voltage step-up.

Over the last decade, various topologies of impedance-source galvanically isolated dc-dc converters have been proposed [8]. Among them, galvanically isolated single-switch topologies offer simple switching stage implementation [10]. On the other hand, they suffer from reduced dc voltage gain, which could be increased by combining transformer-based and coupled inductor-based single-switch converter topologies. This approach allows for much higher gain as both an isolation transformer and a coupled inductor of an impedance-source network can have windings connected to the secondary side rectifier, thus providing increased dc voltage gain. The quasi-Z-source (qZS) based topologies are most widely used among the impedance-source converters [8, 11]. The qZS network provides continuous input current and allows for simple integration of its magnetic components on a single core.

Various structures based on the qZS network have been presented. In [12] a cascaded qZS network was used to achieve the high voltage gain. Continuous input current and reduction in the shoot-through duty cycle are the main advantages. A
family of hybrid Z-source converters have been presented in [13]. A qZS network is combined with a traditional Z-source network to improve the boost ability of the converter. These converters provide a wide range of variation in voltage gain, but there is no isolation between the output and the input. Their voltage boosting ability can only be achieved by increasing the duty cycle. The qZS and Y-source networks were also applied to the isolated push-pull converter in [14] and [15], respectively. These converters can achieve high voltage gain, while their range is limited due to high losses at low input voltage. A qZS series resonant dc–dc converter has been proposed in [16] to extend the input voltage regulation range to 1:6 via multimode control. High efficiency was obtained by using a resonant tank in the output side without adding extra components. A qZS network with a three-winding coupled inductor allowing 1:10 input voltage regulation range has been presented in [17]. The proposed converter is suitable for wide input voltage range applications at sub-kW power levels. Its applicability is limited by the non-optimal operation of magnetic networks in a wide range of voltages, which results in their bulky design. To optimise the utilisation of magnetic components, a single-switch galvanically isolated qZS converter was proposed for cost-sensitive applications such as PV module-level power electronics in [18]. Both the transformer and the coupled inductor of the qZS network were used to meet the high step-up conditions along with reasonable power density. High efficiency, very high voltage gain, and a low number of semiconductor devices are the main advantages of that converter. Its input voltage range is still compromised due to limited control freedom of the shoot-through duty cycle, as in most of the converters reviewed above.

As mentioned previously, the galvanically isolated dc–dc converters are commonly used in renewable energy applications, especially in PV module-level power electronics. One of the most important issues around the PV system is the influence of environmental conditions. PV module operation is mostly defined by the weather conditions, that is, ambient temperature and solar irradiance. Partial shading effects are most prominent in small PV systems with distributed maximum power point tracking (MPPT) [19] and could be caused by various obstructions like trees, rooftop structures, neighbouring buildings, dust, or snow, influencing the operation of the PV system significantly [20, 21]. Opaque shading is a severe type of partial shading that could be caused by fallen leaves, snow, or bird droppings, fully blocking sunlight to PV cell(s). Even one opaque shaded PV cell in a substring can block this substring entirely, causing the corresponding bypass diode to conduct. When one or two substrings are blocked, the PV module can potentially lose one- or two-thirds of its output voltage, respectively. To harvest the power under these conditions, an interface converter should typically have an ultra-wide input voltage regulation range with a lower boundary of the MPPT window below 10 V. A high step-up converter with high efficiency has been presented in [18]. In order to improve the operation of the converter and increase the efficiency in the range of low input voltages, it can be realised as a dual-mode converter. In low voltage conditions, such as partial shading, the high efficiency at high voltage gain can be achieved by applying adaptive mode change to the dual-mode operation like in [22, 23].

This study proposes a dual-mode magnetically integrated PV microconverter with adaptive mode change. Depending on the panel voltage level, the new galvanically isolated qZS converter can change its operation modes adaptively to improve the converter efficiency. The operation principle of the proposed converter is explained in two operation modes along with the derivation of voltage gain expressions. The experimental results are provided to examine the practicality of the converter. The control system is synthesised and explained. Based on that, the global MPPT (GMPPT) is implemented. The efficiency is analysed to validate the performance of the proposed converter in a dual-mode operation. Finally, the conclusions are drawn.

2 | DUAL-MODE MAGNETICALLY INTEGRATED DC-DC CONVERTER

The proposed topology (Figure 1) is originated from the single-switch qZS dc–dc converter [18]. It consists of a coupled inductor TX2, two capacitors (CqZS1 and CqZS2), switches SqZS and S1, an isolation transformer TX1, a blocking capacitor Cb, voltage doubler rectifier (VDR) implemented with diodes D1, D2 and capacitors C1, C2 and a half-wave rectifier (HWR) formed by D3, S2, D4, and C3. The input inductor L1 ensures continuous current from a PV module.

The main switch S1 short circuits the output port of the qZS network to perform the voltage step-up. The isolation transformer TX1 is also connected to this port through the dc blocking capacitor Cb. The diode D3 and the switch S2 constitute a unidirectional switch. The switch S2 performs adaptive mode change by enabling or disabling the HWR connected to the secondary winding of TX2. If the HWR is disabled, the converter operates as the baseline single-switch qZS dc–dc converter. When it is enabled, the output voltage will be equal to the summed outputs of the HWR and VDR.

The dc gain of the converter depends on a duty cycle of S2 as well as on a state of the switch S2, which can be set either on or off for a whole switching period. Switches S1 and SqZS are controlled complementarily with short dead-time. Hence, in the ideal case, the switching period $T$ could be broken into two time intervals, that is, when the switch $S_1$ is conducting or is in the off-state.

3 | GENERALISED OPERATION PRINCIPLE

As it was mentioned above, the switch $S_2$ operates as a static (on/off) switch, thus enabling or disabling the HWR. Therefore, the operation of the converter is analysed in two modes: With disabled HWR when $S_2$ is turned off and enabled HWR with $S_2$ turned on.
3.1 Converter operation with disabled HWR

3.1.1 Time interval $t_{on}$

Within this time interval, the switch $S_1$ is turned on, and $S_{qZS}$ is turned off. The equivalent circuit is shown in Figure 2(a). Waveforms of the control signals, voltages and currents are presented in Figure 3. The energy transmission process in this time interval is such that the capacitors $C_{qZS1}$ and $C_{qZS2}$ are discharged while the inductors $L_{qZS1}$ and $L_{qZS2}$ are charged. Diodes $D_1$ and $D_2$ are reverse biased and conducting, respectively. The secondary winding of $TX1$ charges the capacitor $C_1$ through the diode $D_2$, $C_2$ is discharged by the output current.

The relationship between mutual inductance and leakage inductance is described as follows:

$$L_{lk1} = 1 - k_1^2 L_{lep},$$  \hspace{1cm} (1)$$
$$L_{lk2} = 1 - k_2^2 L_{qZS2},$$  \hspace{1cm} (2)

where $k_1$ and $k_2$ are the coupling coefficients and $k_1 = k_2 = 1$ at the perfect coupling.

The relationship between the primary and the secondary windings of the transformer and the coupled inductor is as follows:

$$V_{N1} = N_p V_{Np} = n_1 V_{Np},$$  \hspace{1cm} (3)$$
$$V_{Nsec} = N_{qZS} V_{NqZS} = n_2 V_{NqZS},$$  \hspace{1cm} (4)

where $V_{N1}$ and $V_{Nsec}$ are the voltages of the primary and secondary windings of $TX1$, respectively, $V_{NqZS}$ is the voltage of the secondary winding of $TX2$, and $V_{NqZS}$ is the voltage of $I_{qZS2}$. The voltage across the inductor $L_{qZS1}$ is as follows:

$$v_{LqZS1} = V_{in} + V_{CqZS2}$$  \hspace{1cm} (5)$$
The current through the inductor $L_{qZS1}$ is expressed as

$$i_{L_{qZS1}} = \frac{V_{in} + V_{CqZS2}}{L_{qZS1}} t + I_{L_{qZS1}}. \quad (6)$$

where $I_{L_{qZS1}}$ is the minimum current through the $L_{qZS1}$. Considering Equation (6), during this time interval, the inductor $L_{qZS1}$ is charged, and current through it has been increased, so that at $t = D \cdot T$ it reaches the maximum value. Applying $t = D \cdot T$ to Equation (6), the maximum current through the inductor is obtained as

$$I_{L_{qZS1}} = \frac{(V_{in} + V_{CqZS2}) D}{L_{qZS1} f} + I_{L_{qZS1}}. \quad (7)$$

During this time interval, the voltage of the inductor $L_{qZS2}$ is given by

$$v_{L_{qZS2}} = V_{CqZS1}. \quad (8)$$

The current through the inductor $L_{qZS2}$ is as follows:

$$i_{L_{qZS2}} = \frac{V_{CqZS1}}{L_{qZS2}} t + I_{L_{qZS1}}. \quad (9)$$

Similar to $L_{qZS1}$, the maximum current through the inductor $L_{qZS2}$ is obtained as

$$I_{L_{qZS2}} = \frac{V_{CqZS1} D}{L_{qZS2} f} + I_{L_{qZS1}}. \quad (10)$$

As the switch $S_1$ is turned on, the voltage across it equals zero. By considering Figure 2(a), the voltage of the primary winding of $TX1$ is given by

$$v_{Np} = -V_{Cb}. \quad (11)$$

Considering Equations (3) and (11), the voltage of the secondary winding of $TX1$ can be calculated as

$$v_{N1} = -n_1 V_{Cb}. \quad (12)$$

Applying Kirchhoff’s voltage law (KVL) to the secondary side of $TX1$, the following relation is obtained:

$$V_{c1} = v_{L_{lk1}} - v_{N1}, \quad (13)$$

where

$$v_{L_{lk1}} = L_{lk1} \frac{di_{L_{lk1}}}{dt} = n_1 \frac{1 - k_1^2}{k_1^2} I_{L_{lk}} \frac{di_{L_{lk}}}{dt} = n_1 \frac{1 - k_1^2}{k_1^2} i_{L_{lk}}. \quad (14)$$

Considering Equations (12), (13), (14), and Figure 2(a), the voltage across the capacitor $C_1$ could be found by

$$V_{C1} = \left( \frac{2k_2^2 - 1}{k_1^2} \right) n_1 V_{Cb}. \quad (15)$$

During this time interval, the current through the capacitors is defined by the currents of the $qZS$ inductors:

$$i_{CqZS1} = -i_{L_{qZS1}}. \quad (16)$$

$$i_{CqZS2} = -i_{L_{qZS2}}. \quad (17)$$

### 3.1.2 Time interval $t_{off}$

During this time interval, $S_1$ and $S_{qZS}$ are turned off and on, correspondingly. The equivalent circuit of the proposed converter is shown in Figure 2(b). Inductors $L_{qZS1}$ and $L_{qZS2}$ release part of their energy to charge up the capacitors $C_{qZS1}$ and $C_{qZS2}$. The VDR diodes $D_1$ and $D_2$ are conducting and reverse biased, respectively. Through the conducting diode $D_1$, the capacitor $C_1$ is discharged by the output current and the capacitor $C_2$. The voltage of the inductor $L_{qZS2}$ is given by

$$v_{L_{qZS2}} = V_{in} - V_{CqZS1}. \quad (18)$$

The current through the inductor is as follows:

$$i_{L_{qZS2}} = \frac{V_{in} - V_{CqZS1}}{L_{qZS2} f} t + I_{L_{qZS2}}. \quad (19)$$

During $T_{off}$, the inductor is discharging; so its current is decreasing to $I_{L_{qZS1}}$ at the end of $T_{off}$. Applying $t = (1-D)T$ to Equation (19), the minimum current through the inductor $L_{qZS1}$ could be found:

$$I_{L_{qZS1}} = \frac{(V_{in} - V_{CqZS1})(1-D)}{L_{qZS1} f} + I_{L_{qZS2}}. \quad (20)$$

According to Figure 2(b), the voltage of the inductor $L_{qZS2}$ is obtained as follows:

$$v_{L_{qZS2}} = -V_{CqZS2}. \quad (21)$$

Considering Equation (22), the current through the inductor $L_{qZS2}$ is given by

$$i_{L_{qZS2}} = \frac{-V_{CqZS1}}{L_{qZS2}} t + I_{L_{qZS2}}. \quad (22)$$
The voltage across the switch $S_1$ is defined by the voltage of the qZS capacitors:

$$V_{S1} = V_{CqZY1} + V_{CqZY2}$$  \hspace{3cm} (23)

The voltage of the primary winding of $TX1$ is obtained as follows:

$$v_{Np} = V_{CqZY1} + V_{CqZY2} - V_{C3}$$  \hspace{3cm} (24)

The voltage of the secondary winding of $TX1$ is as follows:

$$v_{Ns} = n_1(V_{CqZY1} + V_{CqZY2} - V_{Cb})$$  \hspace{3cm} (25)

By applying KVL to the secondary side of $TX1$, the following equation is obtained:

$$V_C1 - V_C2 = v_{Llk1} - v_{Ns}.$$  \hspace{3cm} (26)

The following expression is derived from Equation (26) taking into account Equations (14), (24), (25), and Figure 2(b) as follows:

$$V_C1 - V_C2 = n_1 \left( \frac{1 - 2k_1^2}{k_1^2} \right) (V_{CqZY1} + V_{CqZY2} - V_{C3}).$$  \hspace{3cm} (27)

By replacing Equations (15) in (27), the capacitor voltage $V_{C2}$ can be obtained:

$$V_{C2} = n_1 \left( \frac{2k_1^2 - 1}{k_1^2} \right) (V_{CqZY1} + V_{CqZY2}).$$  \hspace{3cm} (28)

By applying the volt-second balance to the inductors $L_{qZY1}$ and $L_{qZY2}$, and considering Equations (15) and (27), the voltages across the capacitors $C_{qZY1}$ and $C_{qZY2}$ are obtained as follows:

$$V_{CqZY1} = \frac{1 - D}{1 - 2D} V_{in}$$  \hspace{3cm} (29)

$$V_{CqZY2} = \frac{D}{1 - 2D} V_{in}$$  \hspace{3cm} (30)

Considering Equation (23), the boost factor of the proposed converter with disabled HWR could be found as:

$$B = \frac{V_{out}}{V_{in}} = \frac{V_{CqZY1} + V_{CqZY2}}{V_{in}} = \frac{1}{1 - 2D}.$$  \hspace{3cm} (31)

Taking into account Equations (28) to (30) and Figure 2(b), the voltage gain of the proposed converter is calculated as follows:

$$V_{out} = V_{C2}.$$  \hspace{3cm} (32)

3.2 Converter operation with enabled HWR

For enabling the HWR, the switch $S_2$ is turned on and the diode $D_4$ is reverse biased for a whole switching period. It should be mentioned that depending on the polarity of the secondary winding of $TX2$ (Figure 4), the HWR could rectify either the positive or negative half-wave of $TX2$ voltage. The procedure of the analysis is similar for both polarities. Here, the converter is analysed for the rectification of the positive half-wave. Similar to the previous section, the switching period is divided into two time intervals corresponding to the turn-on ($T_{on}$) and turn-off ($T_{off}$) states of the switch $S_1$.

3.2.1 Time interval $t_{on}$

The equivalent circuit of the proposed converter is shown in Figure 5(a). During this time interval, the diode $D_3$ is conducting, the capacitor $C_3$ is charging. Waveforms of the control signals, voltages, and currents are shown in Figure 6. The relations for the current and the voltage of the inductors and the capacitor are the same as those at the turned-off switch $S_2$. According to Figure 5(a), the voltage of the capacitor $C_3$ is obtained as follows:

$$v_{NqZY} = v_{LqZY} = V_{CqZY1}.$$  \hspace{3cm} (36)

$$v_{N\text{sec}} = n_2 V_{CqZY1}.$$  \hspace{3cm} (37)

FIGURE 4 Polarity of $TX2$: (a) positive; (b) negative
Applying KVL to the secondary side of $TX2$, the following relation is obtained:

$$V_{C3} = v_{NSec} - v_{Llk2}, \tag{38}$$

where

$$v_{Llk2} = L_{Llk2} \frac{di_{Llk2}}{dt} = n_2 \frac{1 - k_2^2}{k_2^2} I_{qZS} \frac{di_{qZS}}{dt} = n_2 \frac{1 - k_2^2}{k_2^2} v_{LqZS}, \tag{39}$$

Considering Equations (37), (38), (39), and Figure 5(a), the voltage across the capacitor $C_3$ is as follows:

$$V_{C3} = \left( \frac{2k_2^2 - 1}{k_2^2} \right) n_2 V_{CqZS}. \tag{40}$$

Replacing Equations (29) in (40) yields:

$$V_{C3} = n_2 \left( \frac{2k_2^2 - 1}{k_2^2} \right) \left( 1 - D \right) V_{in}. \tag{41}$$

### 3.2.2 Time interval $t_{off}$

During this time interval, the diode $D_3$ will be reverse biased when the switch $S_3$ is turned off. The capacitor $C_3$ is discharging while supplying the output current. The equivalent circuit of the proposed converter is shown in Figure 5(b). The relations of the current and the voltage of the capacitor and inductors are the same as the converter operates as a single switch $qZS$ converter. According to Figure 5(b), the output voltage is equal to the summed outputs of the HWR and VDR.

Using Equations (29) to (31) and (42), the voltage gain of the proposed converter is calculated as follows:

$$V_{out} = V_{C2} + V_{C3}. \tag{42}$$

$$V_{out} = n_1 \left( \frac{2k_1^2 - 1}{k_1^2} \right) \left( V_{CqZS1} + V_{CqZS2} \right) + n_2 \left( \frac{2k_2^2 - 1}{k_2^2} \right) V_{CqZS1}. \tag{43}$$

Assuming $k_1 = k_2 = 1$, for the rectification of the positive half-wave of $TX2$, the voltage gain could be obtained as follows:

$$G_{pos} = \frac{V_{out}}{V_{in}} = \frac{n_1 + n_2(1 - D)}{1 - 2D}. \tag{44}$$

The dc input-to-output current conversion factor for the rectification of the positive voltage half-wave is given by

$$\left( \frac{I_{out}}{I_{in}} \right)_{pos} = \frac{1 - 2D}{n_1 + n_2(1 - D)}. \tag{45}$$

As it was mentioned, the procedure of the steady-state analysis when the negative half-wave is rectified is the same as for the positive half-wave. Therefore, only the final equation of the voltage gain and the dc input-to-output current conversion factor are given:

$$G_{neg} = \frac{V_{out}}{V_{in}} = \frac{n_1 + n_2D}{1 - 2D}. \tag{46}$$

$$\left( \frac{I_{out}}{I_{in}} \right)_{neg} = \frac{1 - 2D}{n_1 + n_2D}. \tag{47}$$

Figure 6 shows the comparison of the theoretical voltage gains of the proposed converter for the operating modes with
the disabled and enabled HWR. Moreover, the dc gain with enabled HWR is visualised for both connection possibilities of the HWR. As can be observed, the best gain extension could be achieved by the rectification of the positive half-wave of the secondary voltage of \( TX2 \). Figure 7 shows the variations of the normalised voltage of the switch \( S_1 \) versus voltage gain for the unity turns ratio of \( TX1 \) and \( TX2 \). The curves are plotted based on the equation given by [24]. As can be seen, the voltage stress of the switch \( S_1 \) is lower for the rectification of the positive half-wave. High voltage gain with low voltage stress of the switch shows that the proposed converter has the best configuration when the HWR rectifies the positive half-wave of \( TX2 \).

4 | EXPERIMENTAL VERIFICATION

This section provides the experimental results to validate the performance of the proposed converter. The experimental prototype with the peak power of 300 W was assembled following the schematics shown in Figure 1. The main specifications and types of semiconductor components used in the prototype are listed in Table 1. The converter was designed for MPPT and interfacing of the 60- and 72-cell PV modules into a dc bus with the nominal operating voltage of 400 V.

4.1 | Steady-state waveforms

The steady-state waveforms of the experimental prototype are shown in Figures 8 and 9. In Figure 8, the converter is operating with disabled HWR, and to ensure the demanded dc gain, the duty cycle of \( S_1 \) was set to 0.27. Evidently, the current through the \( I_{lqZS1} \) equals zero when the HWR is disabled. Therefore, the current \( I_{lqZS1} \) is changing linearly. The output voltage is equal to the average voltage of the capacitor \( C_2 \). For the turn-on state of the switch \( S_1 \), the primary voltage of the \( TX1 \) is equal to the voltage of the capacitor \( C_1 \).

By turning on the switch \( S_2 \), the HWR is activated to increase the voltage gain (Figure 9). As a result, the duty cycle is decreased from 0.27 to 0.227 for the same operating point. The output voltage is equal to the sum of the voltages of \( C_2 \) and \( C_3 \), which are 330 and 70 V, respectively. The voltage stress of the switch \( S_1 \) is 55 V, which is the voltage of \( C_2 \) divided by the turns ratio of \( TX1 \). The decrease in the duty cycle value is positively mirrored on the input current ripple, which is 25% smaller with an activated HWR. Also, the primary winding voltage of \( TX1 \) is lower than that for the operation with disabled HWR.

Figures 8 and 9 show that due to the presence of the input inductor, the converter features the continuous input current, irrespective of the operating mode. To reduce the number of magnetic components in the circuit, the input inductor \( L_{qZS1} \) could be integrated into the magnetic core of \( TX2 \), as reported in [25]. However, due to the increased ripple of the core flux, such a three-winding coupled inductor would have a considerably increased volume of the magnetic core, which will negatively impact the power density of the converter.

### Table 1 Specifications and components of the converter

| Test conditions | Symbol | Range/value |
|-----------------|--------|-------------|
| Input voltage range | \( V_{in} \) | 5–65 V |
| Input current range | \( I_{in} \) | 0–10 A |
| Input power range | \( P \) | 0–300 W |
| Switch duty cycle | \( D \) | 0.05–0.45 |
| Switching frequency | \( f_s \) | 80 kHz |
| Output voltage | \( V_{out} \) | 400 V |

| Passive components | Designator | Value |
|--------------------|------------|-------|
| Input inductor | \( L_{qZS1} \) | 24 \( \mu \)H |
| Magnetising inductance of \( TX1 \) | \( L_m \) | 29 \( \mu \)H |
| Turns ratio of \( TX1 \) | \( n_1 \) | 6 |
| Leakage inductance of \( TX1 \) | \( L_{lk1} \) | 25 \( \mu \)H |
| Magnetising inductance of \( TX2 \) | \( L_{qZS2} \) | 28 \( \mu \)H |
| Turns ratio of \( TX2 \) | \( n_2 \) | 6 |
| Leakage inductance of \( TX2 \) | \( L_{lk2} \) | 14 \( \mu \)H |
| Quasi-Z-source (qZS) and blocking capacitor | \( C_{qZS1}, C_{qZS2}, C_b \) | 52 \( \mu \)F |
| Resonant capacitor | \( C_1 \) | 220 nF |
| Output capacitors | \( C_2, C_3 \) | 1 \( \mu \)F |

| Semiconductors components | Part number |
|-----------------------------|------------|
| \( S_1 \) and \( S_{qZS} \) | Infineon BSC035N10SS5 |
| Rectifier diodes \( D_1–D_4 \) | CREE C3D02060E |
| Static switch \( S_2 \) | Infineon IPD60R280P7 |
4.2 Efficiency analysis

Figure 10 shows the efficiency variations of the experimental prototype versus the input voltage. First, the efficiency was acquired within the input voltage range from 8 to 65 V, and at the input current of 2 A. As is seen from Figure 10(a), the activation of the HWR allows for extending the lower bound of the input voltage range from 15 to 5 V without any remarkable impact on the efficiency.

Moreover, by using the HWR, the power conversion efficiency in the operating points with 15 and 25 V was improved by 5.8% and 2.2%, respectively. At the input current of 4 A, the activation of the HWR results in the efficiency rise from 1% to 3%, depending on the operating point (Figure 10(b)).

Figure 11 shows the comparison of the efficiency between two possible connection possibilities of the TX2 and HWR (positive and negative, see Figure 4), within the input voltage range from 5 to 30 V and at the input current of 10 A. It is evident that the operation of the converter with the rectification of the positive half-wave is more beneficial for the high step-up applications as it can give up to 3% of efficiency rise at the lower values of the input voltage.

Figure 12 shows the efficiency profile of the converter operating with continuous power of 100 W in the range of input voltages from 8 to 65 V. For the input voltages below 25 V, the enabling of the HWR results in higher efficiency, especially in the case of the positive half-wave rectification by the HWR.

In conclusion, the proposed approach with the possibility of mode change of the combined energy transfer significantly enhances the step-up performance of the baseline single-switch qZS dc-dc converter, allowing either for a wider input voltage regulation range or higher power conversion efficiency at large dc gain values.

4.3 Adaptive mode change implementation

For the proposed approach of the combined energy transfer, the optimal tradeoff between the dc gain range and the power conversion efficiency could be achieved by the proper selection of the turns ratio of TX1 and TX2. Another possibility of the dc gain extension without serious penalties to the efficiency is the implementation of the adaptive mode change (AMC) principle derived from the topology-morphing control [26, 27]. In that case, the operating mode of the converter is automatically selected by the control system to change the dc voltage gain instead of regulating the duty cycle outside the favourable range. As a result, the duty cycle is constrained to a region of high efficiency despite variations of the input voltage in a wide range.
Figure 13 shows the AMC realisation principle in the experimental converter operating at the constant power of 100 W. The transition between modes is realised at $V_{in} = 24$ V by turning on or off the switch $S_2$. This resulted in an efficiency rise of up to 4% at the lowest values of the input voltages (Figure 14). Moreover, owing to AMC implementation, the converter now features a relatively flat efficiency of over 92% within a dc gain range from 8 to 26.

5.1 EXPERIMENTAL EVALUATION OF MPPT PERFORMANCE

In this section, we will demonstrate how the proposed approach could be used in a low-cost PV microconverter for improving its MPPT performance. More specifically, an extension in a dc gain variation range using AMC enables the implementation of a GMPPT based on a periodic P-V curve sweep [22], which was technically impossible with a baseline single-switch qZS microconverter due to the dc gain limitation issues [10].
5.1 Implementation of GMPPT

The GMPPT method is examined under several operation conditions to show the performance of the converter. For a typical Si-based PV module with three sub-strings, the shading conditions on the sub-strings can make three MPPs: two local MPPs (LMPPs) and the global MPP (GMPP). As mentioned previously, in the partial shading or opaque shading conditions, the range of the input voltage is low. In these conditions, to achieve the GMPP, activation of the HWR at low voltages is needed to achieve acceptable efficiency and cover the demanded input voltage range.

Applying the method presented in [28] and using the reconfiguration ability of the proposed converter, GMPPT is enabled along with improved efficiency even at low irradiations. The global MPPT routine is implemented in two stages. First, the converter has to preset the operating point to the maximum achievable voltage of the PV module, which corresponds to the minimum allowed duty cycle of the switch $S_1$. Second, the reference voltage is gradually decreasing down to the minimum allowed input voltage. During the voltage sweep, the control system measures the input voltage and current in tabular form with a certain voltage step. Implementation of these stages results in the identification of the actual GMPP position. The control system defines the position of the GMPP and gradually presets the converter towards the vicinity of this point. After approaching the actual position of the GMPP close enough, the local MPPT takes over and performs accurate tracking of the GMPP.

To achieve optimal power conversion efficiency during the GMPPT, the converter has to be switched between the operating modes employing AMC. Considering the experimental results in the previous section, there is an optimum transition point for the AMC where the efficiency curves of the converter with enabled and disabled HWR intersect. Theoretically, the AMC threshold voltage can be approximated using a polynomial expression as demonstrated in [28]. However, this imposes increased requirements for the measurement sensors as well as the main microcontroller which is an important issue in such cost-sensitive applications like PV module-level power electronic converters. Therefore, to simplify AMC implementation and, consequently, constrain the cost of the proposed technology, this study embraces low-cost realisation using a simple hysteresis transition around the input voltage of 20 V. This threshold voltage was identified as the most probable transition point for the AMC at high input currents, that is, highest losses and thermal loading of semiconductors, where the optimisation of the efficiency has the highest influence on the converter reliability. To improve converter dynamics, especially during transients, feedforward control is implemented.

Considering the conditions explained above, the control system shown in Figure 15 was synthesised. The MPPT block determines the reference input voltage. The reference input voltage is applied to the feedforward block. This feedforward control calculates the theoretical duty cycle, while in practice, the real value should be a little larger. This difference between the theoretical and the experimental duty cycle values results in the input voltage setting error applied to the PI controller. The PI controller compensates for this difference by the signal $\Delta D$. Moreover, the regulation speed is improved as the PI controller needs to handle only a small portion of the control signal. The feedforward duty cycle $D_f$ is calculated by the equations below, depending on the operation mode:

$$D = \frac{V_{in} - n_1V_{in}}{2V_{out}} \tag{49},$$

$$D = \frac{V_{in} - (n_1 + n_2)V_{in}}{2(V_{out} - n_2V_{in})} \tag{50}.$$  

Clearly, Equation (49) corresponds to the disabled HWR, while Equation (50) should be used when the HWR is enabled. The AMC threshold voltage $V_{TH}$ defines selection between these equations. Hence, the AMC is seamlessly integrated into the closed-loop control system. The resulting duty cycle that is applied to the modulator is obtained by adding the signal produced by the PI controller ($\Delta D$) to the feedforward term $D_f$ and taking into account the relative duration of the dead-time $D_{DT}$. The hysteresis block is used for stable AMC implementation to avoid frequent changes between the modes when the GMPP is close to the threshold voltage $V_{TH} = 20$ V. In Figure 15, variables in red colour are either acquired from sensors or were predefined in the software.

5.2 Experimental validation of GMPPT

To evaluate the GMPPT performance of the proposed microconverter, different operating scenarios have been tested. This study considers flexible CIGS PV module MiPV-165W as this type of PV modules is rapidly gaining popularity in building-integrated PV, recreational vehicles, industrial uses, and so forth. These PV modules contain tens of bypass diodes as there is one in every cell, which results in countless shading scenarios possible.
Closed-loop control system comprising maximum power point tracking (MPPT) and adaptive mode change of the proposed converter intended for PV module integration into a stable dc bus

Test P-V profiles used in the evaluation of MPPT performance: $PV_1$ (red line) describes the standard test conditions, and $PV_2$ (green line) is used to model hard shading scenarios.

Figure 16 shows two P-V profiles that were implemented employing a solar array simulator Agilent E4360A controlled by the LabView software. Due to the limitations of the testbench, the PV module was represented by its equivalent sectioned into four substrings with four bypass diodes. The profile $PV_1$ with one MPP is for the PV module operation under nominal operating cell temperature (NOCT) of 45°C and uniform irradiance of 800 W/m² without shading. The profile $PV_2$ corresponds to the operating conditions derived from the profile $PV_1$ where cells in three substrings are shaded down to 200 W/m². This is a reason for having two MPPs.

Figure 17 shows the experimental waveforms acquired from the implementation of MPPT. The waveforms show the trend of voltage and current changes from the GMPPT scanning start until the converter operates around the GMPP. To show the dual-mode operation, the signal $S_2$ measured at the gate driver output is shown. The converter was controlled by microcontroller STM32F334 using Cortex-M4 core and high-resolution PWM periphery.

For profile $PV_1$, the AMC performed the switching between the modes twice. First, during the GMPPT scanning where the input voltage was lower than the threshold voltage of 20 V. Second, the converter switches from operation with the HWR to the baseline single-switch topology near GMPP. For the profile $PV_2$, the AMC results in the switch $S_2$ turn-on during the GMPPT scanning. The GMPP was set for low input voltage because of shading conditions. The GMPP was reached at the best efficiency due to the implementation of the dual-mode AMC. This means that the HWR was active during the GMPPT scanning at the input voltage of below 20 V. Near the GMPP, it was disabled to increase the efficiency.

Conclusions

In this study, a dual-mode galvanically isolated dc-dc converter has been proposed. It can utilise voltage step-up by means of both an isolation transformer and a coupled inductor utilised with half-wave and a VDR, respectively. The proposed topology uses adaptive mode change to define whether the HWR is enabled. As a result, both magnetic elements provide voltage step-up to the output at low input voltages, while only a highly efficient transformer is used at higher voltages. Implementation of the adaptive mode change has resulted in the extension of the input voltage range. Also, the efficiency has been improved for the low range of the input voltage.

The efficiency of the converter in different modes was investigated experimentally. According to the results, the converter...
ACKNOWLEDGMENTS

This research was supported by the Estonian Research Council (grant PUT1443), by the Estonian Centre of Excellence in Zero Energy and Resource Efficient Smart Buildings and Districts (ZEBE), grant 2014-2020.4.01.15-0016 funded by the European Regional Development Fund, and by the European Commission through the H2020 project Finest Twins (grant No. 856602).

REFERENCES

1. Carriere, T., et al.: Strategies for combined operation of PV/storage systems integrated into electricity markets. IET Renewable Power Gener. 14(1), 71–79 (2020)
2. Zhang, Y., et al.: A wide input-voltage range quasi-z-source boost dc–dc converter with high-voltage gain for fuel cell vehicles. IEEE Trans. Ind. Electron. 65(6), 5201–5212 (2018)
3. Forouzesh, M., et al.: High-efficiency high step-up dc–dc converter with dual coupled inductors for grid-connected photovoltaic systems. IEEE Trans. Power Electron. 33(7), 5967–5982 (2018)
4. Forouzesh, M., et al.: Step-up DC–DC converters: A comprehensive review of voltage-boosting techniques, topologies, and applications. IEEE Trans. Power Electron. 32(12), 9143–9178 (2017)
5. Tseng, K.C., Huang, C.C.: High step-up high-efficiency interleaved converter with voltage multiplier module for renewable energy system. IEEE Trans. Ind. Electron. 61(3), pp. 1311–1319 (2014)
6. Gorji, S.A., et al.: Galvanically isolated switched-boost-based DC-DC converter. In Proceedings 2016 IEEE Energy Conversion Congress and Exposition (ECCE), pp. 1–6, Milwaukee, WI (2016), https://doi.org/10.1109/ECCE.2016.7855129.
7. Forouzesh, M., Baghramian, A.: Galvanically isolated high gain Y-source dc–dc converters for dispersed power generation. IET Power Electron. 9(6), 1192–1203 (2016)
8. Chub, A., Vinnikov, D., Blaabjerg, F., Peng, F. Z.: 'A review of galvanically isolated impedance-source dc–dc converters. IEEE Trans. Power Electron. 31(4), pp. 2808–2828 (2016)
9. Vinnikov, D., et al.: New high-gain step-up dc/dc converter with high-frequency isolation. In Proceedings of APEC, pp. 1204–1209, Orlando, FL, USA (2012), https://doi.org/10.1109/APEC.2012.6165972
10. Vinnikov, D., et al.: Single-switch galvanically isolated step-up DC-DC converter for residential photovoltaic applications. In: IECON, pp. 6578–6582, Florence, Italy (2016), https://doi.org/10.1109/IECON.2016.7793776
11. Liang, W., et al.: Night operation, analysis, and control of single-phase quasi-Z-source photovoltaic power system. IET Renewable Power Gener. 13(15), 2817–2829 (2019)
12. Vinnikov, D., et al.: Step-up DC/DC converters with cascaded quasi-Z-source network. IEEE Trans. Ind. Electron. 59(10), 3727–3736 (2012)
13. Shen, H., et al.: Hybrid Z-source boost DC–DC converters. IEEE Trans. Ind. Electron. 64(1), 310–319 (2017)
14. Swiakoti, Y.P., et al.: High-voltage boost quasi-Z-source isolated DC/DC converter. IET Power Electron. 7(9), 2387–2395 (2014)
15. Swiakoti, Y.P., et al.: Y-source impedance-network-based isolated boost DC/DC converter. In: IPEC Hiroshima, pp. 1801–1805, Hiroshima, Japan (2014), https://doi.org/10.1109/IPEC.2014.6869828
16. Vinnikov, D., et al.: High-performance quasi-Z-source series resonant DC-DC converter for photovoltaic module-level power electronics applications. IEEE Trans. Power Electron. 32(5), 3634–3650 (2017)
17. Chub, A., et al.: Novel isolated power conditioning unit for micro wind turbine applications. IEEE Trans. Ind. Electron. 64(7), 5984–5993 (2017)
18. Chub, A., et al.: Single-switch impedance-source galvanically isolated dc–dc converter with combined energy transfer. In: Proceedings of RTUCON, pp. 1–6, Riga, Latvia (2016), https://doi.org/10.1109/RTUCON.2018.8659851
19. Sinapis, K., et al.: A comprehensive study on partial shading response of c-Si modules and yield modeling of string inverter and module level power electronics. Sol. Energy 135, 731–741 (2016)
20. Bulbaram, A., et al.: Control and circuit techniques to mitigate partial shading effects in photovoltaic arrays. IEEE J. Photovoltaics 2(4), 532–546 (2012)
21. Lappalainen, K., Valkealaki, S.: Fluctuation of PV array global maximum power point voltage during irradiance transitions caused by clouds. IET Renewable Power Gener. 13(15), 2864–2870 (2019)
22. Liang, Z., et al.: A high-efficiency PV module-integrated dc–dc converter for PV energy harvest in FREEDM systems. IEEE Trans. Power Electron. 26(3), 897–909 (2011)
23. Liang, Z., et al.: A high efficiency DC MIC for PV energy harvest in FREEDM systems. In: Proceedings of APEC, pp. 1–6, Fort Worth, TX, USA (2011), https://doi.org/10.1109/APEC.2011.5744612

24. Vinnikov, D., et al.: Voltage gain extension techniques for high step-up galvanically isolated DC-DC converters. In: Proceedings of IEEE ICIT 2020, pp. 1021–1027, Buenos Aires, Argentina (2020), https://doi.org/10.1109/ICIT45562.2020.9067115

25. Vinnikov, D., et al.: Magnetically integrated high step-up resonant DC-DC converter for distributed photovoltaic systems. In: Proceedings of IEEE IEC 2017, pp. 7691–7697, Beijing, China (2017), https://doi.org/10.1109/IECON.2017.8217348

26. Jovanović, M.M., Irving, B.T.: On-the-fly topology-morphing control—efficiency optimization method for LLC resonant converters operating in wide input- and/or output-voltage range. IEEE Trans. Power Electron. 31(3), 2596–2608 (2016), https://doi.org/10.1109/ECCE.2019.8912292

27. Vinnikov, D., et al.: Fault-tolerant bidirectional series resonant DC-DC converter with minimum number of components. In: Proceedings of IEEE ECCE 2019, pp. 1359—1363, Baltimore, MD, USA (2019)

28. Vinnikov, D., et al.: MPPT performance enhancement of low-cost PV microconverters. Sol. Energy 187, 156–166 (2019)

How to cite this article: Maheri HM, Vinnikov D, Chub A, Korkh O, Rosin A, Babaei E. Dual-mode magnetically integrated photovoltaic microconverter with adaptive mode change and global maximum power point tracking. *IET Renew Power Gener.* 2021;15:86–98. https://doi.org/10.1049/tpg2.12007