BIDIRECTIONAL ISOLATED ZVS DC-DC CONVERTER WITH NONPULSATING INPUT & OUTPUT CURRENT

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Abstract - A new bidirectional isolated Zero-Voltage-Switching (ZVS) DC-DC converter with nonpulsating input and output current is proposed. This converter is ideally suited for battery charging application and is applied as a submodule in a battery storage system based on a modular multilevel converter (M²C). The current output of the proposed DC-DC converter enables small current ripple for directly interfacing a battery without the need of additional filter components. The derivation of topology, the operating principle, the analysis of the ZVS area as well as simulation results of a 12.4 kW DC-DC converter and a M²C submodule are provided.

Keywords – DC-DC Converter, Nonpulsating Input/Output Current, Zero-Voltage-Switching, Modular Multilevel Converter, Battery Storage

I. INTRODUCTION

With increasing use of renewable energy sources like wind power and solar energy, energy storage systems become an important part of future energy distribution systems due to the inherently fluctuating and stochastic nature of renewable energy sources. The storage systems are able to compensate for energy shortages during low wind conditions or at nighttime. In order to store considerable amounts of energy and ensure balancing supply and demand, suitable high power electronic equipment interfacing the storage media play an important role.

Besides the well known energy storage systems using the potential energy of water, in recent years additional concepts based on compressed air, flywheels, thermal energy storage in molten salt [1] as well as chemical storage in batteries have been investigated. The battery based systems offer the advantage of high energy and power density, high cycle efficiency as well as being location-independent and easily scalable. Such systems also become more important in grid applications for load leveling providing fast frequency regulation [2].

At high power levels the battery energy storage system is usually connected to the medium voltage distribution grid. Because of the relatively high voltages, multilevel converter systems are advantageous to use due to lower harmonics, robust operation and reduced switching losses. The multilevel converters [3], and especially the modular multilevel converters (M²C) [4], additionally offer the benefit to easily split the energy storage into smaller modules as proposed in [5, 6], and have the benefit to be highly modular and fault tolerant sys-

Fig. 1. M²C based battery storage system consisting of 2n submodules per phase leg and subsequent DC-DC converters to charge electric vehicles (a), the conventional half-bridge submodule topology with subsequent isolated DC-DC converter (b) and the proposed bidirectional isolated ZVS DC-DC converter with nonpulsating input and output current for use as a submodule (c).
tems. With the different submodules a balancing of the batteries is also possible [7, 8].

Fig. 1(a) shows a M²C where the submodules additionally include an isolated DC-DC converter as shown in Fig. 1(b). Due to the galvanic isolation, the submodule outputs of the three phases \( a, b \) and \( c \) can be connected in parallel in order to obtain a continuous power flow from the grid to each battery. Additionally, with this parallel connection an interleaving of the three converters is possible, which reduces the current ripple and enables a significant size reduction of the passive components.

As shown by the dashed lines in Fig. 1(a), non-isolated DC-DC converters can be connected to the batteries of the submodules and the outputs of these converters can be paralleled in order to perform a high-power charging of electric vehicles. Due to the high charging power an ultra-fast charging of electric vehicles in less than 5 minutes is possible. This charging concept is investigated in the project “Ultra-Fast Charging of Electric Vehicles” [6].

Usually, half-bridge input stages as shown in Fig. 1(b) are used in the submodule of the M²C for interfacing the grid. A subsequent isolated DC-DC converter is utilized to perform galvanic isolation, voltage adaptation and charging control. Using for example a dual-active-bridge converter [9] as an isolated DC-DC converter results in a total number of 10 switching devices per submodule. Furthermore, for charging the batteries a low current ripple is required [10] resulting in additional effort for filtering the output current or increasing the output capacitors of the DC-DC converter.

Therefore, in this paper a new bidirectional isolated DC-DC converter with nonpulsating input and output current as shown in Fig. 1(c) is proposed. The converter is based on an isolated Cuk converter but features soft-switching for all four switching devices. It can be used as a submodule in a M²C based battery storage system but also as single DC-DC converter for charging batteries. A typical application can be in interconnecting a high voltage DC bus with the 12 V battery in a (hybrid-)electric vehicle.

In the following, first the novel topology is derived in section II and then, the operating principle in grid-to-battery and battery-to-grid operation as well as the ZVS operating area are analyzed in section III. Finally, section IV shows a concept of a 250 kW / 2 MWh battery storage system and simulation results of a 12.4 kW DC-DC converter submodule, which is part of the energy storage system.

II. DERIVATION OF TOPOLOGY

The commonly used half-bridge submodule in a M²C as presented in [4] can be connected to a DC-DC converter to achieve galvanic isolation and voltage adaptation as shown in Fig. 1(b). Nevertheless, for interfacing a battery additional filter components are required to ensure a small current ripple. A possible implementation of an isolated DC-DC converter is a dual-half-bridge [11] or a dual-active-bridge [9] with a high frequency isolation transformer and a subsequent output filter to limit the current ripple. However, the main disadvantage is the high number of switching devices (6 or 10), an increased control complexity, higher losses and reduced reliability.

The bidirectional isolated Cuk converter shown in Fig. 2 basically offers nonpulsating input and output currents and achieves galvanic isolation with only two switching devices [12]. In both power flow directions only one switching device is actively turned on/off during operation. However, the isolated Cuk converter is a hard-switching converter, which results in switching losses, and has the disadvantage that the energy stored in the transformer leakage inductance must be absorbed by the output capacitance of the switching device, resulting in voltage ringing and higher voltage stress [13].

To avoid the voltage ringing across the main switching device and achieve soft-switching at the same time, an auxiliary switch and a clamp capacitor can be connected in parallel to the switching device [14] or the transformer primary side [15]. There, the clamp capacitor absorbs the energy stored in the transformer leakage inductance and the main as well as the auxiliary switching devices are then turned on/off under ZVS conditions. However, these topologies only allow unidirectional power flow.

By inserting the active clamp circuit consisting of an auxiliary switch and a clamping capacitor on both the transformer primary (connection \( a_1 - a_2 \) in Fig. 2) and secondary side (connection \( b_1 - b_2 \) in Fig. 2) the new topology shown in Fig. 1(c) results, whose primary circuit has been presented in [16, 17]. The proposed DC-DC converter features nonpulsating input and output current, bidirectional power flow, galvanic isolation and ZVS of all four switching devices over a wide load range.

In the following section, the operating principle of the converter is analyzed in detail.

III. OPERATING PRINCIPLE

For simplification of circuit analysis, the magnetizing inductance \( L_m \) of the transformer is neglected in the following. Thus, the primary and secondary transformer leakage inductance can be summed up to \( L_s = L_{s1} + L_{s2} \), where \( L_{s2} \) is referred to the primary side. The switching devices \( S_I/S_2 \) and \( S_3/S_4 \) in the two half-bridge legs are inversely controlled. Hence, there are 4 switching states.

One of the capacitors \( C_1/C_2 \) as well as one of the capacitors \( C_3/C_4 \) form a resonant circuit together with the transformer leakage inductance \( L_s \) which depends on the switching state of the converter. The switching frequency is chosen to be well above the resonant frequency of this resonant circuit.

This paper focuses on controlling the two half-bridge legs with a duty cycle of 50%. However, also duty cycles different than 50% are possible. The phase-shift between the square-wave voltages applied on primary and secondary side of the transformer determines the output power.
\[ P = \frac{N_1}{N_2} \cdot \frac{V_1 V_2 d (1 - 2d)}{f_s L_\sigma} \] (1)

where \( V_1 \) is the input voltage, \( V_2 \) the output voltage, \( N_1/N_2 \) the transformer turns ratio, \( d \) the relative phase-shift between the two square-wave voltages on primary and secondary side of the transformer and \( f_s \) the switching frequency. At a fixed output voltage, the converter behaves like a phase-shift controlled current source with average output current

\[ I_2 = \frac{N_1}{N_2} \cdot \frac{V_1 d (1 - 2d)}{f_s L_\sigma}. \] (2)

For a relative phase-shift of \( d = 0.25 \) the output power reaches its maximum at

\[ P_{\text{max}} = \frac{N_1}{N_2} \cdot \frac{V_1 V_2}{8 f_s L_\sigma}. \] (3)

The control strategy leads to capacitor voltages \( v_{C1}/v_{C2} \) equal to the input voltage \( V_1 \) and capacitor voltages \( v_{C3}/v_{C4} \) equal to the output voltage \( V_2 \). Therefore, the voltage stress on the primary switching devices is double the input voltage and double the output voltage on the secondary switching devices. The voltage stress can be reduced by controlling the primary half-bridge with a duty cycle above 50% and the secondary half-bridge with a duty cycle below 50% as will be discussed in a future paper.

A. Grid-to-Battery Operation

In grid-to-battery operation, the square-wave voltage applied to the transformer primary side leads the square-wave voltage applied to the transformer secondary side – similar to a phase-shift controlled dual-active-bridge. The operation over a whole switching period \( T_c \) can be described using 12 modes which are shown in Fig. 4 with the corresponding key waveforms in Fig. 3.

At time \( t_0 \), diodes \( D_1 \) and \( D_3 \) are conducting. Switches \( S_1 \) and \( S_3 \) are turned on, \( S_2 \) as well as \( S_4 \) are turned off. The voltage \( v_{C1}/v_{C2} \) across the capacitors \( C_1/C_2 \) equals \( V_1 \), \( v_{C3}/v_{C4} \) across the capacitors \( C_3/C_4 \) equals \( V_2 \).

1) Mode 1 (\( t_0 < t < t_1 \)): The absolute value of the leakage inductance current \( i_{L\sigma} \) is larger than \( i_{L2} \), so that diode \( D_3 \) is conducting. Furthermore, \( i_{L\sigma} \) and the output current \( i_{L2} \) are linearly increasing while the input current \( i_{L1} \) is linearly decreasing.

2) Mode 2 (\( t_1 < t < t_2 \)): At \( t_1 \), \( |i_{L\sigma}| = i_{L2} \), so that the current through diode \( D_3 \) reverses its polarity and therefore commutates to switch \( S_3 \).

3) Mode 3 (\( t_2 < t < t_3 \)): Since \( i_{L\sigma} > i_{L1} \), the current through diode \( D_1 \) reaches zero at time \( t_2 \), so that switch \( S_1 \) starts to conduct.

4) Mode 4 (\( t_3 < t < t_4 \)): At \( t_3 \), switch \( S_3 \) is turned off and capacitances \( C_{r3}/C_{r4} \) provide ZVS conditions. \( C_{r3}/C_{r4} \) is charged/discharged by the current driven by leakage inductance \( L_\sigma \) and the output inductor \( L_2 \). At \( (t_4 - t_3)/2 \) the output current \( i_{L2} \) reaches its maximum and starts to decrease linearly. As soon as \( C_{r4} \) is completely discharged, diode \( D_4 \) turns on at \( t_4 \). Then, switch \( S_4 \) can be turned on at nearly zero voltage.

5) Mode 5 (\( t_4 < t < t_5 \)): The input current \( i_{L1} \) and output current \( i_{L2} \) are linearly decreasing. Depending on the output voltage \( V_2 \) referred to the primary side in comparison to the input voltage \( V_1 \), current \( i_{L\sigma} \) increases (\( V'_{2} > V_1 \)), remains constant (\( V'_{2} = V_1 \)) or decreases (\( V'_{2} < V_1 \)).

6) Mode 6 (\( t_5 < t < t_6 \)): Switch \( S_1 \) is turned off at \( t_5 \) and capacitances \( C_{r1}/C_{r2} \) provide ZVS conditions. \( C_{r1}/C_{r2} \) is charged/discharged by the current driven by \( L_\sigma \) and inductor \( L_1 \) until \( C_{r2} \) is completely discharged and diode \( D_2 \) starts to conduct at \( t_6 \). Now, switch \( S_2 \) can be turned on at nearly zero voltage. The input current \( i_{L1} \) reaches its minimum at \((t_6 - t_5)/2\) and starts to increase linearly again.

7) Mode 7 (\( t_6 < t < t_7 \)): The current \( i_{L\sigma} \) decreases linearly and at \( t_7 \), \( i_{L\sigma} = i_{L1} \), so that diode \( D_2 \) stops conducting and the current commutates to \( S_2 \) after \( t_7 \).

8) Mode 8 (\( t_7 < t < t_8 \)): The current \( i_{L1} \) is increasing linearly and \( i_{L2} \) as well as \( i_{L\sigma} \) are decreasing linearly. Furthermore, \( i_{L\sigma} \) changes its direction.

9) Mode 9 (\( t_8 < t < t_9 \)): At \( t_8 \), the current through diode \( D_3 \) reaches zero, so that \( D_4 \) turns off and switch \( S_4 \) starts conducting.

10) Mode 10 (\( t_9 < t < t_{10} \)): At \( t_9 \), switch \( S_4 \) is turned off and capacitances \( C_{r3}/C_{r4} \) provide ZVS conditions. \( C_{r3}/C_{r4} \) is discharged/charged by the current driven by \( L_\sigma \) and \( L_2 \). Capacitor \( C_{r3} \) is discharged until diode \( D_3 \) turns on, so that switch \( S_1 \) can be turned on at nearly zero voltage. The output current \( i_{L2} \) reaches its minimum at \((t_{10} - t_9)/2\) and starts increasing linearly.

Fig. 3. Key waveforms of the proposed DC-DC converter in grid-to-battery operation.
Fig. 4. Operating modes of the proposed DC-DC converter over one switching period $T_s$ in grid-to-battery operation.
11) Mode 11 (t_{10} < t < t_{11}): The currents \( i_{L1} \) and \( i_{L2} \) are linearly increasing and \( i_{L1} \) increases \((V'_2 > V_1)\), remains constant \((V'_2 = V_1)\) or decreases \((V'_2 < V_1)\) depending on the output voltage \( V'_2 \) in comparison to \( V_1 \).

12) Mode 12 (t_{11} < t < t_{12} = T_s): At \( t_{11} \), switch \( S_2 \) is turned off and capacitances \( C_{eq1}/C_{eq2} \) provide ZVS conditions. \( C_{eq1}/C_{eq2} \) is discharged/charged by the current driven by \( L_\sigma \) and \( L_1 \). \( C_{eq1} \) is discharged until diode \( D_1 \) turns on, so that switch \( S_1 \) can be turned on at nearly zero voltage. Current \( i_{L1} \) reaches its maximum at \((t_{12} - t_{11})/2\) and starts decreasing linearly.

At the end of mode 12, i.e. after time \( t_{12} = T_s \), mode 1 starts again.

B. Battery-to-Grid Operation

In battery-to-grid operation the control signals of \( S_1 \) and \( S_2 \) have to be exchanged with the ones of \( S_3 \) and \( S_4 \) compared to grid-to-battery operation, i.e. the primary and secondary side circuit in Fig. 1(c) are exchanged. The square-wave voltage applied to the transformer primary side lags the square-wave voltage applied to the transformer secondary side.

C. Zero-Voltage-Switching Area

To ensure ZVS conditions, the current flowing through the switching device at turn-off has to be large enough to charge/discharge the two capacitors parallel to the switches in a bridge leg. The conditions in grid-to-battery operation are given by

\[
\dot{V}_{r,1} = Z_{0,2} \cdot (N_1 N_2 \cdot i_{L\sigma}(t_3) + i_{L2}(t_3)) > 2V_2 \quad (4)
\]

\[
\dot{V}_{r,2} = Z_{0,1} \cdot (i_{L\sigma}(t_3) - i_{L1}(t_3)) > 2V_1 \quad (5)
\]

\[
\dot{V}_{r,3} = Z_{0,2} \cdot (-N_1 N_2 \cdot i_{L\sigma}(t_9) - i_{L2}(t_9)) > 2V_2 \quad (6)
\]

\[
\dot{V}_{r,4} = Z_{0,1} \cdot (-i_{L\sigma}(t_11) + i_{L1}(t_11)) > 2V_1 \quad (7)
\]

with

\[
Z_{0,1} = \sqrt{\frac{L_\sigma L_1}{(L_\sigma + L_1)C_{eq1}}} \quad (8)
\]

\[
Z_{0,2} = \sqrt{\frac{L_\sigma N_2^2 L_2}{(N_2^2 L_1^2 + L_2)C_{eq2}}} \quad (9)
\]

The capacitance \( C_{eq1}/C_{eq2} \) is the constant, energy equivalent capacitance for the two nonlinear output capacitances of the switching devices plus the auxiliary capacitance in parallel to the switch. To enable ZVS turn-on of the switches, the two capacitances must be completely charged/discharged within the interlocking delay. The resonant transition time is given by

\[
t_{r,\alpha} = \frac{1}{\omega_{0,\alpha}} \cdot \arcsin \left( \frac{2V_2}{\dot{V}_{r,\beta}} \right) \quad (10)
\]

\( \alpha \in \{1, 2\} \) denotes the bridge leg, \( \beta \in \{1, 2, 3, 4\} \) the resonant transition. The angular frequencies of the primary and secondary resonant transitions are

\[
\omega_{0,1} = \sqrt{\frac{L_\sigma + L_1}{L_\sigma L_1 C_{eq1}}} \quad (11)
\]

\[
\omega_{0,2} = \sqrt{\frac{L_\sigma N_2^2 L_2}{N_2^2 L_1^2 + L_2 C_{eq2}}} \quad (12)
\]

In Fig. 5 the ZVS area determined with the equations above is given. With a relative phase-shift in the range of \( d \in [0,0.25] \) a much smaller ZVS area results than with \( d \in [0.25,0.5] \). The reason is the much smaller peak current \( i_{L\sigma} \) which limits the ZVS area significantly.

IV. 250 kW / 2 MWh Battery Storage System

Due to the non-pulsating input and output current, the soft-switching DC-DC converter presented in the previous sections is ideally suited for the battery storage system shown in Fig. 1(a). The system consists of 38+2 submodules per leg/phase and is connected to the 6.6 kV medium voltage AC grid. Each submodule has a maximum output power of 12.4 kW. The total system power is 250 kW. The DC link of the battery storage system is kept at a constant voltage of 13 kV.

The outputs of three submodules each in a different phase are connected in parallel and interleaved in order to further reduce the current ripple and the size of the output inductors. Furthermore, a continuous power flow from the grid to the Lithium-ion battery pack with an average voltage of 300 V and an energy capacity of 50 kWh is obtained with this parallel connection. On the primary side of the submodules, the input inductors are replaced by a common inductor \( L_{a1}/L_{a2} \) reducing the system volume. In the following, detailed results for a submodule are presented.
A. Simulation Results of the Submodule

The proposed circuit topology is simulated as DC-DC converter and as single submodule in the M$^2$C battery storage system shown in Fig. 1(a). The simulation model corresponds with Fig. 1(c) but neglects the magnetizing inductance $L_m$ of the transformer as well as the nonlinearity of the output capacitances of the switching devices.

The DC-DC converter is simulated with a constant input voltage of 600 V, a constant output voltage of 300 V, a switching frequency of 20 kHz and the parameters given in Table I. The simulation was carried out at a relative phase-shift of $d = 0.15$.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{1,nom}$</td>
<td>50 V ... 600 V</td>
</tr>
<tr>
<td>$V_{2,nom}$</td>
<td>240 V ... 360 V</td>
</tr>
<tr>
<td>$f_s$</td>
<td>20 kHz</td>
</tr>
<tr>
<td>$L_1$</td>
<td>10 mH (in M$^2$C: 2 mH)</td>
</tr>
<tr>
<td>$L_2$</td>
<td>6 mH (in M$^2$C: 2 mH)</td>
</tr>
<tr>
<td>$N_1/N_2$</td>
<td>2</td>
</tr>
<tr>
<td>$L_{o1}$</td>
<td>90 $\mu$H</td>
</tr>
<tr>
<td>$L_{o2}$</td>
<td>22.5 $\mu$H</td>
</tr>
<tr>
<td>$L_m$</td>
<td>neglected</td>
</tr>
<tr>
<td>$C_1, C_2$</td>
<td>15 $\mu$F (in M$^2$C: 2.5 $\mu$F)</td>
</tr>
<tr>
<td>$C_3, C_4$</td>
<td>60 $\mu$F</td>
</tr>
<tr>
<td>$C_{eq1}, C_{eq2}$</td>
<td>1 nF</td>
</tr>
<tr>
<td>$C_{r1}, C_{r2}, C_{r3}, C_{r4}$</td>
<td>2 nF</td>
</tr>
</tbody>
</table>

Fig. 6 shows the simulated key waveforms in grid-to-battery operation whereas Fig. 7 shows the key waveforms in battery-to-grid operation. ZVS operation of all switching devices is achieved as shown in Fig. 8.

Additionally, the submodule is simulated with a sinusoidal input voltage from 50 V to 600 V as it is the case in the M$^2$C battery storage system. Simulation results of the input current and the interleaved output power to the battery are shown in Fig. 9. Compared with the simulation parameters for the DC-DC converter, the inductance $L_2$ can be decreased by a factor of 3 for a constant current ripple due to interleaved control of the submodules in the three different phases (see Fig. 1(a)). Furthermore, the capacitances $C_1$ and $C_2$ are reduced to 2.5 $\mu$F to limit the reactive power flow needed to charge and discharge the primary capacitors with the grid frequency. The gate control signals are only mathematically calculated. In a next step, a controller will be implemented.

B. Submodule Prototype

In the M$^2$C based battery storage system, the input voltage of a submodule varies sinusoidal from 50 V to 600 V. Due to the series connection of the submodules in a phase leg, the balancing of the batteries needs to be done by controlling the input voltages of the submodules. Therefore, the submodule prototype is designed for an input voltage range up to 800 V to provide power control reserve. A battery voltage swing of 300 V ± 20% is considered. The submodule is operated at a switching frequency of 20 kHz.

Using a three-level neutral point clamped (NPC) input stage for the submodule prototype as shown in Fig. 10 divides the required blocking voltage of the primary switching devices. The primary and secondary switching devices are chosen to be 1.2 kV silicon carbide MOSFETs from CREE [18].

In Table II the components of the submodule prototype are given while Table III shows the estimated losses of the switching devices and the passive components at an output power of 12.4 kW. Fig. 11 shows a possible hardware realization of the submodule.
Fig. 8. ZVS of all switching devices of the proposed DC-DC converter in simulation (\(d = 0.15\)). \(v_{S1}\) to \(v_{S4}\) depict the voltages across the switching devices.

Fig. 9. Simulated input current of the submodule and the interleaved output power \(p_2\) to the battery in the M²C battery storage system. The voltage \(v_1\) corresponds to the voltage seen by a single submodule. The input current is only pre-controlled.

**TABLE II**

<table>
<thead>
<tr>
<th>Component</th>
<th>Type</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inductor core (L_1)</td>
<td>Metglas AMCC-100 (Alloy 260SA1)</td>
</tr>
<tr>
<td>Inductor core (L_2)</td>
<td>Metglas AMCC-80 (Alloy 260SA1)</td>
</tr>
<tr>
<td>Transformer core (2\times)</td>
<td>Metglas AMCC-320 (Alloy 260SA1)</td>
</tr>
<tr>
<td>Capacitor (C_{1}/C_{2})</td>
<td>Electronicon E53.H59-232T10</td>
</tr>
<tr>
<td>Capacitor (C_{3}/C_{4})</td>
<td>Electronicon E53.M59-603T20</td>
</tr>
<tr>
<td>Switching devices</td>
<td>CREE CMF20120D</td>
</tr>
<tr>
<td>Estimated volume</td>
<td>6.3 dm</td>
</tr>
</tbody>
</table>

**V. CONCLUSION**

In this paper, a new bidirectional isolated DC-DC converter with nonpulsating input and output current and ZVS over a wide load range is presented. The new topology is derived from a Cuk converter and the operating details are discussed. Due to the low input and output current ripple, this converter is ideally suited for battery charging applications, which is demonstrated with a 250 kW / 2 MWh medium voltage battery energy storage system that can also be used for ultra-fast...
charging of electric vehicles. The storage system is based on a M2C structure that utilizes the proposed converter topology as submodules. For validating the proposed concept, a 12.4 kW prototype of a submodule is investigated.

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