Single-Phase Single-Stage Bidirectional Isolated
ZVS AC-DC Converter with PFC

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Single-Phase Single-Stage Bidirectional Isolated ZVS AC-DC Converter with PFC

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Abstract—This paper presents a single-phase bidirectional isolated AC-DC converter with Power Factor Correction (PFC) consisting of a half-bridge on the AC side and a full-bridge on the DC side to accomplish single-stage power conversion. The converter applies a new control scheme combining phase-shift and frequency modulation to achieve Zero-Voltage-Switching (ZVS) over the full range of the AC line voltage. Compared to the conventional boost PFC approach, the proposed converter eliminates high frequency harmonic distortions on the mains due to the inherently integrated LC input filter stage. The operating principle in AC-to-DC and DC-to-AC under ZVS conditions by means of analytical considerations are provided. Simulation results and a detailed loss model of a 3.3 kW electric vehicle battery charger to connect to the 230 Vrms / 50 Hz mains considering a battery voltage range of 280 V to 430 V validate the theoretical analysis. The converter can also be used as a submodule in a Cascaded H-Bridge Converter (CHB) for medium or high voltage applications.

Keywords—AC-DC Converter, Power Factor Correction (PFC), Bidirectional, Isolated, Zero-Voltage-Switching (ZVS)

I. INTRODUCTION

Single-phase isolated AC-DC converters are widely used for applications like charging (hybrid-)electric vehicles, interfacing storage batteries (e.g. for uninterruptible power supplies) or supplying energy from photovoltaic systems to the grid. Some of these applications demand bidirectional power flow capability, e.g. for implementing Vehicle-2-Grid (V2G) concepts or grid battery storage systems.

Besides the basic two-stage approach with a boost Power Factor Correction (PFC) rectifier and a subsequent high-frequency isolated DC-DC converter such as Dual Half-Bridge (DHB), Dual Active (Full-)Bridge (DAB) or resonant DC-DC converters [1], several single-stage isolated AC-DC PFC converter topologies have been proposed. A review of state-of-the-art single-phase power quality AC-DC converters is given in [2] whereas [3] summarizes single-phase non-isolated PFC topologies based on the boost converter approach.

Unidirectional single-stage AC-DC PFC converters operating in Discontinuous Conduction Mode (DCM) are proposed in [4], [5] and [6]. However, high power applications ask for PFC operation in Continuous Conduction Mode (CCM) due to high current stresses of the switching devices and conducted emissions in DCM [7]. Furthermore, the output capacitances of the switching devices resonate with the boost inductor during the non-conducting interval.

In [8], [9] a true bridgeless high-efficient AC-DC PFC converter operating in CCM based on a resonant Cuk converter with high power factor is presented. There, integrating isolation into the topology is also possible [10]. Though, the converter system only allows unidirectional power flow.

Bidirectional power flow and single-stage AC-DC operation can be achieved using a Cycloconverter on the primary and a DAB on the secondary side of a medium-frequency transformer [11]. Due to the bidirectional switching devices on the primary side, a sequential multi-step commutation scheme must be used which in turn limits the switching frequency.

Other approaches use phase-shift control as shown in [12], where a half-bridge on the primary and secondary side of the transformer and an input diode rectifier is utilized. The diode rectifier can be replaced by a synchronous rectifier applying MOSFETs to further reduce conduction losses. Furthermore, for shaping the transformer leakage inductance current in order to achieve soft-switching over the full range of the line voltage, a DAB can be used on the transformer secondary side [13].

This paper proposes a single-phase bidirectional isolated AC-DC converter with PFC as shown in Fig. 1 employing combined phase-shift and frequency control to achieve Zero-Voltage-Switching (ZVS) over the full range of the AC line voltage and additionally eliminates the input rectifier stage by use of bidirectional switching devices.

First, in section II the proposed converter topology is explained. The operating modes in AC-to-DC and DC-to-AC case including the converter control to achieve ZVS are described and analyzed in section III. Then, section IV shows a prototype system of the proposed converter including a detailed loss model. Finally, simulation results verify the
II. CONVERTER TOPOLOGY

The topology of the AC-DC converter is shown in Fig. 1. Due to the alternating line voltage, bidirectional switching devices on the AC side have to be employed. These can be realized by an anti-serial connection of two MOSFETs as shown in Fig. 1 or by reverse blocking IGBTs. The AC side of the proposed converter has been presented in [14], [15] considering single-stage AC-AC power conversion. Moreover, a similar primary circuit is used in [16] to supply high pressure sodium lamps.

Compared to the conventional boost PFC approach, the proposed converter eliminates high frequency harmonic distortions on the mains due to the inherently integrated input filter stage. Capacitors $C_1$ and $C_2$ fully absorb the high frequency switched currents. Nevertheless, the remaining AC voltage ripple on the capacitors leads to a certain degree of high frequency input current distortions.

In either power flow direction, the converter can operate in buck or boost mode. The transformer turns ratio $n$ is chosen such that the primary referred DC voltage $V_{2s}$ is higher than half of the maximum peak voltage of the AC line voltage in every operating point. This guarantees that at the lowest battery voltage the primary and secondary applied half-cycle voltage-second products with respect to $n$ can be kept equal.

III. OPERATING PRINCIPLE

The converter is operated with a combined phase-shift and frequency control as shown later. A simple equivalent circuit is drawn in Fig. 2a which consists of the LC input filter stage and a parallel connected variable impedance load $Z_l$. The converter is operated in such a way, that the reactive power consumed by the filter capacitors is fully compensated by the variable impedance load in order to achieve PFC.

Fig. 2b shows a simplified representation of the converter for phase-shift control. The magnetizing inductance $L_m$ of the transformer is assumed to be much larger than the leakage inductance $L_o$ and is therefore neglected in the following. The primary half-bridge is switched with a constant duty cycle of 50%, which leads to equally distributed capacitor voltages $v_{C1}$ and $v_{C2}$. Neglecting the input inductor $L_1$, the average value over one switching cycle of these voltages follows the AC line voltage according to

$$v_{C1}(t) = v_{C2}(t) = \frac{V_1}{2} \sin(\omega t).$$

Assuming further a negligible small capacitor voltage ripple, the secondary DC-link voltage is $v_{C3} = V_2$ for the following mathematical considerations.

During one half-cycle of the AC line voltage, only two of the primary switching devices are switched at high frequency. These are $S_{1a}$ and $S_{2a}$ for the positive and $S_{1b}$ and $S_{2b}$ for the negative half-wave. The devices switched at low frequency are turned on/off at nearly zero voltage and zero current.

As indicated in Fig. 2b, on the primary side of the transformer, a square-wave voltage $v_p$ with the time-dependent amplitude $|v_1(t)|/2$ is applied. The voltage $v_s$ on the transformer secondary side consists of positive and negative voltage pulses with an amplitude of $V_2$, which allows shaping the transformer leakage inductance current $i_{L_o}$ to transfer the desired instantaneous power and to achieve ZVS. The voltage amplitudes of $v_p$ and $v_s$ are assumed to be constant over one switching cycle because the switching frequency is chosen to be well above the mains frequency.

A. AC-TO-DC Operation

In AC-TO-DC operation, power flows from the mains to the DC side. To describe the phase-shift operation mathematically, two control variables $g(t)$ and $w(t)$ relative to the switching period $T_s$ are introduced. Fig. 3 shows the applied voltages on the transformer primary and secondary side $v_p$ and $v_s$ over one switching period $T_s$ and the definition of the phase-shift control variables $g$ and $w$ during the positive and negative half-cycle of the mains in AC-TO-DC operation (Modulation 1). Moreover, the resulting transformer leakage inductance current $i_{L_o}$ and the states of the switching devices are given.

The transformer leakage inductance current $i_{L_o}$ in AC-TO-DC operation is given by

$$i_{L_o}(\tau) = \begin{cases} \frac{|v_1|}{2} + nV_2 & (\tau - \tau_0) + i_{L_o}(\tau_0) \quad \tau_0 \leq \tau \leq \tau_1 \\ \frac{|v_1|}{2} & (\tau - \tau_1) + i_{L_o}(\tau_1) \quad \tau_1 \leq \tau \leq \tau_2 \\ \frac{|v_1|}{2} - nV_2 & (\tau - \tau_2) + i_{L_o}(\tau_2) \quad \tau_2 \leq \tau \leq \tau_3 \\ \frac{|v_1|}{2} - nV_2 & (\tau - \tau_2) + i_{L_o}(\tau_2) \quad \tau_2 \leq \tau \leq \tau_3 \\ \frac{|v_1|}{2} + nV_2 & (\tau - \tau_3) + i_{L_o}(\tau_3) \quad \tau_3 \leq \tau \leq \tau_4 \\ \frac{|v_1|}{2} + nV_2 & (\tau - \tau_3) + i_{L_o}(\tau_3) \quad \tau_3 \leq \tau \leq \tau_4 \\ \frac{|v_1|}{2} + nV_2 & (\tau - \tau_3) + i_{L_o}(\tau_3) \quad \tau_3 \leq \tau \leq \tau_4 \\ \frac{|v_1|}{2} - nV_2 & (\tau - \tau_4) + i_{L_o}(\tau_4) \quad \tau_4 \leq \tau \leq \tau_5 \\ \frac{|v_1|}{2} - nV_2 & (\tau - \tau_4) + i_{L_o}(\tau_4) \quad \tau_4 \leq \tau \leq \tau_5 \\ \frac{|v_1|}{2} + nV_2 & (\tau - \tau_5) + i_{L_o}(\tau_5) \quad \tau_5 \leq \tau \leq \tau_6 \end{cases}$$

with the values at the switching instants

$$i_{L_o}(\tau_0) = -\frac{|v_1|}{2} + (g - w)nV_2,$$

$$i_{L_o}(\tau_1) = \frac{(g - \frac{1}{2}) |v_1| + (g + w)nV_2}{2f_s L_o},$$

$$i_{L_o}(\tau_2) = \frac{(\frac{1}{2} - w) |v_1| + (g + w)nV_2}{2f_s L_o},$$

$$i_{L_o}(\tau_3) = \frac{|v_1|}{2} + (g - w)nV_2,$$

$$i_{L_o}(\tau_4) = \frac{|v_1|}{2} - nV_2,$$

$$i_{L_o}(\tau_5) = \frac{|v_1|}{2} - nV_2,$$

$$i_{L_o}(\tau_6) = \frac{|v_1|}{2} + nV_2.$$
devices differ during the positive and negative half-wave of the AC line voltage.

\[ i_{L_s}(\tau_1) > I_s, \quad i_{L_s}(\tau_3) > I_s. \]  

The ZVS conditions considering a minimum commutation current \( I_s \) for the resonant transition are given by

\[ i_{L_s}(\tau_1) > I_s, \quad i_{L_s}(\tau_3) > I_s. \]  

In order to guarantee a trapezoidal current shape as shown in Fig. 3, \( i_{L_s}(\tau_1) = i_{L_s}(\tau_3) \) is demanded, which leads to one single ZVS condition for the whole switching cycle inducing

\[ w(t) = \left| v_1(t) \right| \left( 1 - 2g(t) \right) \quad \left( 11 \right) \]

The instantaneous power over one switching cycle \( T_s \) transferred from the AC to the DC side can then be calculated by

\[ p_t(t) = \frac{\left| v_1(t) \right| nV_2 \left( g(t) - 2g^2(t) + w(t) - 2w^2(t) \right)}{4f_sL_\sigma}. \quad \left( 12 \right) \]

taking (11) into account. Inserting (11) into (12) and setting the derivative with respect to \( g \) to zero, leads to the maximum transferrable instantaneous power

\[ p_{t,\text{max}}(t) = \frac{|v_1(t)| nV_2 \left( v_1^2(t) + 4|v_1(t)| nV_2 + 4n^2V_2^2 \right)}{32f_sL_\sigma \left( v_1^2(t) + 4n^2V_2^2 \right)} \]  

dependent on the AC line voltage \( v_1(t) \).

The transfer of active power takes place in the intervals \([\tau_1, \tau_3]\) and \([\tau_4, \tau_6]\) which is also shown by the shaded area in Fig. 3. During the intervals \([\tau_3, \tau_4]\) and \([\tau_5, \tau_6]\), only reactive power occurs that is required to achieve ZVS. The control function \( g(t) \) will be given in section III-C.

B. DC-to-AC Operation

In DC-to-AC operation, power flows from the secondary to the primary side of the transformer which demands a different sign of the applied transformer voltages \( v_p, v_s \) and the transformer leakage inductance current \( i_{L_s} \). Fig. 4 shows the transformer voltages \( v_p, v_s \), the leakage inductance current \( i_{L_s} \) and the gating signals of the switching devices for the positive and negative half-wave of the mains (Modulation 1).

The analytical expression of the transformer leakage inductance current \( i_{L_s} \) in DC-to-AC operation can be determined the same way as in AC-to-DC operation considering the switching instants from Fig. 4. Setting \( i_{L_s}(\tau_3) = i_{L_s}(\tau_5) \) leads to (11) for the control variable \( w(t) \).

The instantaneous power transferred from DC to AC side is given by (12) through \(-p_t(t)\). The same holds for the

\[ V_2' \]

Modulation 1

\[ V_2' \]

Fig. 4. Primary and secondary voltages \( v_p, v_s' \) applied to the transformer and resulting transformer leakage inductance current \( i_{L_s} \) over one switching period \( T_s \) in DC-to-AC operation. The gate signals of the primary switching devices differ during the positive and negative half-wave of the AC line voltage.
maximum transferrable power, which is defined by (13) by means of $-\Delta p_{\text{transf}}(t)$.

Active power is transferred in the intervals $[\tau_0, \tau_2]$ and $[\tau_1, \tau_3]$ which is also shown by the shaded area in Fig. 4. Only reactive power occurs during the intervals $[\tau_2, \tau_3]$ and $[\tau_0, \tau_1]$ that is required to achieve ZVS.

C. Converter Control under ZVS Conditions (Modulation 1)

To achieve ZVS for all switching devices during one half-cycle of the mains, the turn-off currents have to be large enough to charge/discharge the drain-source capacitances of the switching devices in a bridge-leg. Looking at the resonant transition and introducing energy equivalent capacitances $C_{\text{eq.p}}, C_{\text{eq.s}}$ for the parallel connection of the drain-source capacitances in a bridge-leg both for AC and DC side, the conditions for ZVS in both AC-to-DC and DC-to-AC operation are

$$i_{\text{LS}}(\tau_1) > \frac{V_1}{\sqrt{\frac{L}{C_{\text{eq.p}}}}} \quad \text{and} \quad i_{\text{LS}}(\tau_1) > \frac{V_2}{\sqrt{\frac{L}{C_{\text{eq.s}}}}},$$

$$i_{\text{LS}}(\tau_3) > \frac{V_1}{\sqrt{\frac{L}{C_{\text{eq.p}}}}} \quad \text{and} \quad i_{\text{LS}}(\tau_3) > \frac{V_2}{\sqrt{\frac{L}{C_{\text{eq.s}}}}}.$$  \hspace{1cm} (14) (15)

Conditions (14) and (15) result in a minimum for the control variable $g(t)$ given by

$$g_{\min}(t) = \frac{4f_s L_s I_s}{|v_1(t)| + 2nV_2} I_s = \max \left\{ \frac{V_1}{\sqrt{\frac{L}{C_{\text{eq.p}}}}}, \frac{V_2}{\sqrt{\frac{L}{C_{\text{eq.s}}}}} \right\}. \hspace{1cm} (16)$$

Furthermore, the maximum transferrable instantaneous power according to (13) occurs at

$$g_{\max}(t) = \frac{v_1^2(t) - |v_1(t)| n V_2 + 2n^2 V_2^2}{2 (v_1^2(t) + 4n^2 V_2^2)}, \hspace{1cm} (17)$$

which is set to be the upper limit for the control variable $g(t)$. To keep a certain power control reserve without losing ZVS, the control variable $g_{\text{ref}}(t)$ is introduced, which provides the average power between controlling with (16) and (17) as shown in Fig. 5a. The analytical expression is found to be

$$g_{\text{ref}}(t) = \frac{\alpha |v_1(t)|^3 + (\alpha n V_2 + \beta) v_1^2(t) + 4\alpha n^2 V_2^3 + 4\beta n^2 V_2^2}{4 (v_1^2(t) + 4n^2 V_2^2) (|v_1(t)| + 2nV_2)}, \hspace{1cm} (18)$$

with $\alpha, \beta$ given by

$$\alpha = 2 - \sqrt{2}, \quad \beta = 8\sqrt{2} f_s L_s I_s. \hspace{1cm} (19)$$

In Fig. 5a the instantaneous power transferred from the AC to the DC side during a half-cycle of the mains for controlling with $g_{\min}(t)$, $g_{\text{ref}}(t)$ and $g_{\max}(t)$ is drawn. At every point in time, a power control reserve of $\pm \Delta p(t)$ is obtained.

To keep the power control reserve and guarantee the transfer of the required instantaneous power, an additional degree of freedom, the switching frequency $f_s$ is considered as time-dependent control variable. To maintain ZVS at low powers around the zero-crossing of the mains voltage, $g$ is kept greater than zero. Therefore, the switching frequency has to rise, in order to lower the transferred power. The instantaneous power which has to be transferred to the transformer secondary side for a given input current amplitude $I_1^\ast$ can be expressed as

$$p_1^\ast(t) = V_1 I_1^\ast \sin^2(\omega t) - \frac{\omega (C_1 + C_2)}{4} V_1^2 \sin(\omega t) \cos(\omega t), \hspace{1cm} (20)$$

which takes the reactive power consumed by the capacitors $C_1, C_2$ into account to achieve PFC. Inserting (18) into (12) while considering (11) and setting

$$p_1(t) = p_1^\ast(t) \hspace{1cm} (21)$$

gives the solution of the additional control variable $f_s(t)$. The upper limit of the switching frequency is set to 120 kHz, which is reached near the zero-crossing of the AC line voltage. There, the reference value of the transfer power $p_1^\ast(t)$ according to (20) falls below the transferrable power obtained by controlling with $g_{\min}(t)$ as can be seen in Fig. 5b. In order to avoid high switching losses, ZVS is maintained by controlling with $g_{\min}(t)$ what in turn leads to a small input current distortion.

To keep the power control reserve and guarantee the transfer of the required instantaneous power, an additional degree of freedom, the switching frequency $f_s$ is considered as time-dependent control variable. To maintain ZVS at low powers around the zero-crossing of the mains voltage, $g$ is kept greater than zero. Therefore, the switching frequency has to rise, in order to lower the transferred power. The instantaneous power which has to be transferred to the transformer secondary side for a given input current amplitude $I_1^\ast$ can be expressed as $p_1^\ast(t) = V_1 I_1^\ast \sin^2(\omega t) - \frac{\omega (C_1 + C_2)}{4} V_1^2 \sin(\omega t) \cos(\omega t)$, which takes the reactive power consumed by the capacitors $C_1, C_2$ into account to achieve PFC. Inserting (18) into (12) while considering (11) and setting $p_1(t) = p_1^\ast(t)$ gives the solution of the additional control variable $f_s(t)$. The upper limit of the switching frequency is set to 120 kHz, which is reached near the zero-crossing of the AC line voltage. There, the reference value of the transfer power $p_1^\ast(t)$ according to (20) falls below the transferrable power obtained by controlling with $g_{\min}(t)$ as can be seen in Fig. 5b. In order to avoid high switching losses, ZVS is maintained by controlling with $g_{\min}(t)$ what in turn leads to a small input current distortion.

D. Partial Load Operation (Modulation 2)

Decreasing the load, especially below 50% of the maximum output power, the saturation region of the switching frequency around the zero-crossing of the AC line voltage is extended.
with the proposed modulation strategy which results in a remarkable input current distortion. Fig. 7 shows the Total Harmonic Distortion (THD) of the input current and the Power Factor (PF) of the converter over the input power range up to the maximum input power of 3.68 kW. From there, it can be concluded that the proposed control strategy is applicable almost down to 40% of the maximum input power, keeping the THD below 5% and the PF above 0.985. The limit is set to 45% of the input power where the vertical line is drawn in Fig. 7.

In case of an input power below 45% of the maximum input power, the modulation is slightly changed (Modulation 2). Phase-shift control as shown in Fig. 8 is applied, where the positive pulse of the secondary voltage waveform lies totally within the positive pulse of the primary voltage waveform. Demanding \( i_{Ls}(t_1) = 0 \) (again in order to get one single ZVS condition) leads in this case also to (11) for the solution of \( w(t) \). To maintain ZVS also at low input powers, \( g(t) > 0 \) has to be guaranteed for a minimum commutation current of zero. Practically, ZVS can not be maintained because the resonant transition after turn-off to charge/discharge the output capacitances of the switching devices requires a certain turn-off current.

![Fig. 6. Control variables \( q_p(t) \), \( w_p(t) \) and \( f_s(t) \) over a half-cycle of the mains period in AC-to-DC operation for a mains voltage of 230 V\(_\text{rms} \), an input current of 16 A\(_\text{rms} \) and an output voltage of \( V_2 = 355 \text{ V} \). The controller adjusts \( g_p(t) \) in the bounds \([g_{\text{min}}(t); g_{\text{max}}(t)]\), the solution of \( g_p(t) \) is used as feed-forward variable.](image1)

![Fig. 7. THD of the input current and PF over the input power range from 10% to 100% of maximum input power for a battery voltage of 355 V applying the proposed modulation scheme (Modulation 1) according to Fig. 3 and Fig. 6 (solid lines). The dashed lines result from the change of the modulation scheme (Modulation 2) according to Fig. 8 at 45% of the input power (vertical line).](image2)

![Fig. 8. Primary and secondary voltages \( v_p, v_s \) applied to the transformer and resulting transformer leakage inductance current \( i_{Ls} \) over one switching period \( T_s \) in partial load AC-to-DC operation.](image3)

### IV. Prototype System

As a prototype system, a 3.3 kW electric vehicle battery charger to connect to the single-phase 230 V\(_\text{rms} / 50 \text{ Hz} \) mains with an output voltage range of 280 V\(_\text{dc} \) to 430 V\(_\text{dc} \) of a Lithium-ion battery is considered. The primary and secondary switching devices are chosen to be 650 V MOSFETs of type STY139N65MS with an on-resistance of 14 mΩ at a junction temperature of 25°C [17]. The system parameters are listed in detail in Table I. A 3D drawing of the prototype system is shown in Fig. 13.

#### A. Converter Components and Loss Model

In the following, the converter components are described with their loss models to calculate the efficiency of the converter applying the proposed control scheme. Table II summarizes the components of the prototype system whereas Fig. 9 shows the calculated efficiencies over the input power range from 10% to 100% for the battery voltages 280 V, 355 V and 430 V. For the proposed modulation strategy during high load conditions (Modulation 1), a peak efficiency of 97.8% is reached at an input power of 2.58 kW and a battery voltage of 280 V. The estimated power density is around 3.9 kW/L. Fig. 10 shows the calculated distribution of the power losses among the converter components at the

### TABLE I

PARAMETERS OF THE PROTOTYPE SYSTEM.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Mains voltage ( V_1 )</td>
<td>230 V(_\text{rms} ) ± 10%</td>
</tr>
<tr>
<td>Mains frequency ( f_1 )</td>
<td>50 Hz</td>
</tr>
<tr>
<td>Battery voltage ( V_2 )</td>
<td>280 V(<em>\text{dc} ) … 430 V(</em>\text{dc} )</td>
</tr>
<tr>
<td>Output power ( P_2 )</td>
<td>3.3 kW</td>
</tr>
<tr>
<td>Switching frequency ( f_s )</td>
<td>20 kHz…120 kHz</td>
</tr>
<tr>
<td>Transformer turns ratio ( n )</td>
<td>10/1.3</td>
</tr>
<tr>
<td>Transformer leakage inductance ( L_{Le} )</td>
<td>11 mH</td>
</tr>
<tr>
<td>Inductors ( L_1, L_2 )</td>
<td>100 μH</td>
</tr>
<tr>
<td>Capacitors ( C_1, C_2 )</td>
<td>10 μF</td>
</tr>
<tr>
<td>Capacitor ( C_3 )</td>
<td>20 μF</td>
</tr>
</tbody>
</table>
TABLE II
COMPONENTS OF THE PROTOTYPE SYSTEM.

<table>
<thead>
<tr>
<th>Component</th>
<th>Material/Details</th>
</tr>
</thead>
<tbody>
<tr>
<td>MOSFETs</td>
<td>2x STYY139N65M5, 650 V, 14 mT</td>
</tr>
<tr>
<td>Capacitors</td>
<td>2x AMCC-4 VITROPERM 500 10 primary turns, 120μm copper foil 13 secondary turns, 120μm copper foil</td>
</tr>
<tr>
<td>Inductors</td>
<td>L₁: 2x Kool Mu E 4317 26u 27 turns, round wire, diameter 2 mm</td>
</tr>
<tr>
<td></td>
<td>L₂: 2x Kool Mu E 4317 26u 28 turns, round wire, diameter 2 mm</td>
</tr>
<tr>
<td>Capacitors C₁</td>
<td>18x Syfer 1825500564KX, 560 nF</td>
</tr>
<tr>
<td>Capacitors C₃</td>
<td>36x Syfer 1825500564KX, 560 nF</td>
</tr>
</tbody>
</table>

**Fig. 9.** Calculated efficiencies over the input power range from 10 % to 100 % of maximum input power for battery voltages 280 V, 355 V and 430 V. The vertical line defines 45 % of the input power where the modulation scheme is changed to limit input current distortions.

**Fig. 10.** Calculated distribution of the power losses among the converter components at the maximum input power of 3.68 kW and a battery voltage of 355 V.

maximum input power of 3.68 kW and a battery voltage of 355 V.

1) **Power MOSFETs:** The losses of the power MOSFETs are mainly determined by conduction losses. Switching losses $P_{Si,swe}$ per MOSFET are approximated by measurement data. To reduce conduction losses, $N_s$ number of MOSFETs are paralleled, so that the power loss per switching device is then approximated by

$$P_{Si} = \frac{R_{ds,on}}{N_s} i_{Si,rs}^2 + N_s P_{Si,swe}$$

with $i = \{1a, 1b, 2a, 2b, 3, 4, 5, 6\}$. The prototype system applies two MOSFETs in parallel for all switching devices ($N_s = 2$). For conduction loss calculations, a junction temperature of 55 °C is assumed where the on-resistance $R_{ds,on}$ of the chosen MOSFET is found to be 18.2 mΩ.

2) **Transformer:** The turns ratio of the transformer $n$ is chosen such that $v_{C_{23}} > V_1/2$ (with $v_{C_{23}}$ referred to the primary side of the transformer) is always satisfied, also at the lowest battery voltage of 280 V. The maximum transformer leakage inductance $L_g$ is determined in such a way, that the peak of the instantaneous power $\tilde{P}_1$ (neglecting the reactive power term in (20)) at full input power of 3.68 kW can be transferred at the lowest switching frequency of 20 kHz and the lowest battery voltage of 280 V by controlling the converter with $g_p(t)$. $L_g$ can be obtained by inserting (18) into (12) while considering (11) and solving

$$\tilde{P}_1 = \tilde{V}_1 \tilde{I}_1 = p(t = T_g/4).$$

The calculated $L_g$ guarantees the above mentioned power control reserve $\pm \Delta p(t)$ also at the peak of the instantaneous power.

For the transformer, two C-cores of size AMCC-4 [18] with VITROPERM 500 [19] material form an E-core where 2 of them are stacked. The primary winding is wound around the center leg and the secondary winding around the center and an outer stray leg as shown in [20]. In the stray leg, an air gap is inserted with a gap length of 0.85 mm to get the desired leakage inductance. In the loss model, the core losses per volume are calculated by applying the improved Generalized Steinmetz Equation (iGSE) [21].

For the primary and secondary windings, the optimal foil thickness is determined as 132 μm and 116 μm according to [22], which gives a minimum value of effective AC resistance. 120μm copper foil is used for the prototype system, with 10 primary and 13 secondary turns. The skin and proximity effect losses per unit length in foil conductors for each current harmonic are then calculated according to [23]. The external magnetic field strength for calculating proximity losses is derived by a 1D approximation using the Dowell method [24].

3) **Inductors:** The primary and secondary inductors are built with two stacked E-cores of type Kool Mu 4317 with material 26u from Magnetics [25]. Powder cores are ideally suited for the prototype system because they offer a distributed air gap and a high saturation flux density of around 1 T and are therefore advantageous over a ferrite core with a large air gap exhibiting considerable fringing magnetic field. Both inductors are wound with round wire of 2 mm diameter since the high frequency components in the input/output current are relatively small. The number of turns for primary inductor L₁ are 27, for secondary inductor L₂ 28, so that a minimum inductance value of 100 μH is guaranteed at the highest peak current.

The core losses per volume are calculated by using the iGSE, the Steinmetz parameters are obtained from [25]. The skin and proximity effect losses per unit length in round conductors for each current harmonic are calculated according to [23]. Also for the inductors, the external magnetic field strength is derived by a 1D approximation using the Dowell method [24].
4) Capacitors: For primary and secondary capacitors $C_1, C_2, C_3$, paralleled 560 nF ceramic capacitors with dielectric X7R from Syfer [26] are used. Multilayer ceramic capacitors exhibit high energy density and are therefore ideally suited to achieve high power densities. Since the voltage ripple on the capacitors is relatively small at high switching frequencies, dielectric losses are not accounted for in the loss model. Therefore, only the thermal losses according to

\[ P_{C_i} = \frac{R_{esr}}{N_i} i_{C_i, \text{rms}} \quad i = \{1, 2, 3\} \]  

(24)

where $R_{esr}$ denotes the equivalent series resistance obtained from datasheet and $N_i$ the number of capacitors paralleled, are considered.

5) Auxiliary Losses: Besides the load dependent loss shares shown in the previous sections, a constant loss share for pre-charging relay, gate drives, control, sensing and cooling of 8 W is considered. Additional losses caused by an EMI filter are approximated by an equivalent resistance of 4 mΩ.

6) Cooling System: The number of semiconductors basically defines the base plate size of the heat sink as 80 mm $\times$ 65 mm for primary and secondary side switching devices, so that a double-sided heat sink can be used. Two 40 mm $\times$ 40 mm fans of type San Ace 40 are applied for forced convection cooling. After optimizing the cooling system as described in [27], a thermal heat sink to ambient resistance of $R_{h,a} = 0.25 \text{ K/W}$ results which in turn leads to a Cooling System Performance Index (CSPi) of 12.6.

B. Simulation Results

The proposed converter is simulated in GeckoCIRCUITS [28] with a simulation model according to Fig. 1 and the parameters given in Table I for an input current reference $I_1^* = \sqrt{2} \cdot 16 \text{ A}_{\text{rms}}$ and an output voltage $V_2 = 355 \text{ V}$ in AC-to-DC operation for a mains voltage of 230 V$_{\text{rms}}$ as can be seen in Fig. 11.

V. FURTHER APPLICATIONS

The AC-DC converter depicted in Fig. 1 can also be used as a submodule in a medium/high voltage battery energy storage system shown in Fig. 12 where the submodules per phase are connected in series sharing single filter inductances $L_a, L_b, L_c$ and the three phase branches form a star connection like in the Cascaded Multilevel PWM Converter presented in [29] also known as the Cascaded H-Bridge Converter (CHB). The battery energy storage system drawn in Fig. 12 additionally functions as a fast charging station for (hybrid-)electric vehicles [30], [31]. Subsequent DC-DC converters are connected to the storage batteries whose outputs are paralleled to achieve high output powers for DC fast charging. Alternatively, the outputs of the submodules can be directly connected to the battery of an electric vehicle.

VI. CONCLUSION

A single-phase single-stage bidirectional isolated AC-DC converter with PFC is proposed. The converter achieves ZVS for all switching devices over the full range of the AC line voltage due to the combined phase-shift and frequency modulation. The inherently integrated LC input filter stage allows a substantial reduction of harmonic distortions on the mains. For validating the theoretical analysis, a 3.3 kW electric vehicle battery charger to connect to the single-phase 230 V$_{\text{rms}}$ / 50 Hz mains with a battery voltage range of 280 V to 430 V is investigated. During high load conditions, a peak efficiency of 97.8% is reached at a power density of around 3.9 kW/L. The converter can also be used as a submodule
in a CHB in star connection for medium and high voltage applications

ACKNOWLEDGMENT

The authors would like to thank Swisselectric Research and the Competence Center Energy and Mobility (CCEM) very much for their strong financial support of the research work and VACUUMSCHMELZ for providing the core material for the transformer.

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