Design of a Highly Efficient Bidirectional Isolated LLC Resonant Converter

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Abstract—The isolated unidirectional LLC resonant converter is known for its outstanding efficiency and high power density. Little information has however been published about the possibility of transferring power in the reverse direction. This paper presents modulation schemes for making the LLC converter bidirectional. High efficiencies are predicted for both directions of power flow, though, as the behavior of the resonant tank is substantially different in the reverse direction, some of the inherent benefits of the conventional LLC converter are lost.

Index Terms—Bidirectional, galvanically isolated, LLC series-parallel resonant converter

I. INTRODUCTION

Resonant converters are commonly selected for applications which demand for a high power density and a high energy efficiency. By featuring soft-switching, the switching frequency can in general be chosen much higher than the switching frequency of a comparable hard-switching converter. As a consequence, the volume required for the passive components is drastically reduced, enabling high power densities and high power conversion efficiencies.

In this paper, a highly efficient battery charger is designed, which is capable of bidirectionally charging light electric vehicles (LEVs). The charger will be connected to a dc microgrid at 

\[ V_{dc} = 450 \text{ V} \]

and will feature an output voltage range from 17 V to 56 V. To limit the duty-cycle and/or frequency variation, and to provide galvanic separation from the dc-bus, a transformer is needed.

For this type of application, the LLC resonant converter [1] promises remarkable unidirectional performance [2].

In [3], a bidirectional LLC prototype was built, but no optimized modulation schemes were employed and the converter did not achieve a satisfactory power conversion efficiency. In [4] a symmetric fourth-order resonant converter was built based on an LLC resonant tank, featuring an additional resonant capacitor. However, the proposed CLLC converter operates in boost-mode in both directions and is therefore not very suitable for use as a voltage-regulating converter.

In the following, a bidirectional LLC converter capable of buck-boost operation is designed. Section II summarizes the “classical” unidirectional operating principles of the converter and in Section III, modulation schemes are derived for a highly efficient operation in the reverse direction. In Section IV, the converter is designed for use as a bidirectional LEV charger. The analytic considerations are verified in Section V by time-domain simulations and the performance of the converter is discussed. Finally, the bidirectional LLC converter is compared to the dual active full-bridge converter.

II. OPERATION IN THE FORWARD DIRECTION

Fig. 1 shows the basic circuit diagram of a bidirectional LLC resonant converter. The resonant tank consists of a series capacitor \( C_s \) and a transformer with a turns-ratio of \( n \), into which \( L_s \) and \( L_p \) are integrated. The forward operation has already been analyzed comprehensively for the classical unidirectional LLC converter (e.g. in [1] and [5]). The primary side full-bridge is used to apply a square-wave voltage \( v_1(t) \) to the resonant tank. A near sinusoidal alternating current will flow. The secondary side switches perform synchronous rectification to increase the power conversion efficiency [6].

As the resonant tank acts as a bandpass filter, it is common to assume that only the fundamental components of the currents and voltages in the resonant tank are responsible for the power transfer. This is called the first harmonic approximation (FHA) [1]. In that sense, the square-wave \( v_1(t) \) applied to the resonant tank is represented by its fundamental component

\[ v_{1,(1)} = \hat{v}_{1,(1)} \cdot \sin(\omega t) = V_{dc} \cdot \frac{4}{\pi} \cdot \sin(\omega t), \tag{1} \]

and the rectification stage is replaced by its equivalent ac-resistance

\[ R_{2,ac}' = n^2 \cdot \frac{8 \cdot V_{bat}^2}{\pi^2 \cdot I_{bat}} \tag{2} \]

The ratio of output to input voltage of the converter can be calculated as a function of the switching frequency [5]:

\[ M = \frac{V_{bat}'}{V_{dc}} = \frac{1}{\sqrt{(Q(k - 1/k))^2 + (1 + 1/h - 1/(hk^2))^2}}. \tag{3} \]
The switching frequency is defined relative to the higher resonance frequency of the resonant tank. The switching frequency is defined relative to the higher resonance frequency with switching frequencies above $f_1$. When operated between $(f_1, f_2)$, the range for voltage control is larger because the gain curve is well defined, which is necessary for the PWM to achieve the fundamental component of $v_1$. Fig. 2 illustrates the voltage gain for different designs of the LLC converter operating in the forward direction. (a) The asynchronous clamped mode allows for lower duty-cycles than the clamped mode.

The following substitutions are introduced:

$$V_{\text{bat}} = r V_{\text{bat}}$$
$$f_1 = \frac{1}{2 \pi \sqrt{L_s C_s}}$$

The switching frequency is defined relative to the higher resonance frequency of the resonant tank $f_1$ by the parameter $k$. $Z_0$ is the characteristic impedance of the resonant tank. To make (3) independent of the actual values of $L_p$, $L_s$ and $C_s$, the variables $h$ and $Q$ are introduced. $P_{\text{bat}}$ is the power that is transferred to the battery. The lower resonance frequency $f_2$ is the frequency at which the voltage gain $M$ is maximal.

Fig. 2 illustrates the voltage gain for different designs (variations of $h$) and different loading conditions (variations of $Q$) at different relative switching frequencies. At lighter loads (lower values of $Q$), the maximum achievable voltage gain is high. When operated between $f_2$ and $f_1$, the LLC converter is able to boost $(M > 1)$. Fig. 2 implies, that a limited capability for voltage conversion ratios lower than one exists with switching frequencies above $f_1$. However, the necessary switching frequency variation is large because the gain curve is very flat in this region. In these modes, the resonant tank behaves inductively and zero voltage switching (ZVS) can be performed on the primary side. Zero current switching (ZCS) is achieved on the secondary side, because the secondary side switches are performing synchronous rectification. If the converter was operated below the lower resonance frequency, ZVS would be lost as the resonant tank behaves capacitive in this region. Thus, this mode is typically avoided.

A. Clamped Mode

By making use of the clamped switching state available with the primary side full-bridge, the range for voltage control in buck mode can be expanded. This is necessary for the later introduced bidirectional design. In terms of the FHA, the PWM reduces the fundamental component of $v_1$ [8]:

$$\hat{v}_{1,1} = \frac{4V_{dc}}{\pi} \cdot \cos \left( (1 - d) \frac{\pi}{2} \right)$$

The duty-cycle is defined as $d = \frac{2\tau}{T}$. A duty-cycle of $d = 1$ corresponds to the block mode in which the LLC converter is typically operated. The lower limit of $d$ is defined by the current which is available for ZVS when the converter is leaving the clamped voltage state. The lower $d$ gets, the more will the point in time, at which the clamped state is left, move towards the zero-crossing of the current until the parasitic drain-source capacitances of the switches can no longer be fully discharged. In Fig. 3, the case where ZVS fails is illustrated by a small zigzag arrow.

B. Asynchronous Clamped Mode Modulation

To further increase the possibility of voltage regulation, the asynchronous clamped mode (ACM) [8] can be used:

$$\hat{v}_{1,1} = \frac{V_{dc}}{\pi} \sqrt{10 + 6 \cos \left( (1 - d) \pi \right)}$$

This time, the clamped state is only entered during the first conduction half-period. The voltage waveform now has a dc-offset which is blocked from the transformer by the resonant capacitor $C_s$. Fig. 3 shows the currents and voltages when using the asynchronous clamped mode. This time, soft-switching is achieved. The limiting case for this modulation scheme is reached where $v_1$ is only toggled between negative (resp. positive) and zero (clamped state). However, a smooth transition to this mode is not possible because soft-switching would be lost with very low duty cycles.

III. OPERATION IN THE REVERSE DIRECTION

To transfer power in the reverse direction, the secondary side switches can apply a square wave voltage $v_2(t)$ to the resonant tank. Similar to the analysis of the forward direction, the ratio of input voltage to output voltage can be expressed with the help of the FHA. Because the resonant-tank is not symmetric, the gain equation for the reverse direction differs from the gain equation for the forward direction:

$$M_{\text{rev}} = \frac{V_{dc}}{V_{\text{bat}}} \frac{1}{\sqrt{(Qk - Q)^2 + 1}}$$

$Q$ is defined in an analogue manner, $f_1$ and $k$ are still the same as for the forward direction:

$$Q = \frac{R_{1,ac}}{Z_0} = \frac{Z_0 \pi^2 P_{\text{dc}}}{8 V_{dc}}.$$
$P_d$ denotes the power that is fed to the dc-link.

Because $v_d'(t) = v_cL_2(t)$ is directly applied to $L_p$, $I_p$ neither participates in the resonance of $L_s$ and $C_s$ nor in the power transfer. Equation (6) is thus exactly the same voltage gain relationship already known from the half-bridge series resonant converter (SRC), revealing a familiar drawback [1]: The inability to control the voltage in the no-load case.

This problem can be overcome by making use of the clamped voltage state available with the employed full-bridges. In the following, switching patterns are derived for both a reverse buck mode and a reverse boost mode that allow for full voltage control while transferring power in the reverse direction.

### A. Reverse Operation in Buck Mode

When $V_{dc} < V_{bat}'$, the LLC converter can operate in backward buck mode similar to the series resonant converter is operated in mode V in [9]. Fig. 4 shows the characteristic voltages and currents for this mode. The switching period is defined to begin at $t_0$, when the primary side resonant tank current $\dot{i}_{L_s}$ is zero. The primary side switches perform synchronous rectifications, so ZVS/ZCS is achieved on the primary side. The switching works as follows:

$t \in [t_0, t_1]$: Before the conduction cycle starts, $v_{d1}'$ is still clamped to zero by $S_0$ and $S_d$, $v_1$ is negative ($S_2$ and $S_3$ were in the on-state to reduce the conduction losses but are now turned off to allow for the current to commutate on the primary side). At $t = t_0$, $v_{d1}'$ is switched to $v_{d1}' = V_{bat}'$ by turning $S_0$ off and $S_d$ on. $i_{L_s}$ then begins to rise and $D_1$ and $D_2$ become conducting; $S_1$ and $S_2$ are afterwards turned on again at zero voltage. Power is transferred to the dc-link.

$t \in [t_1, T_2]$: At $t = t_1$, $v_{d1}'$ gets actively clamped to zero by turning off $S_d$ (and subsequently turning on $S_2$). No power is obtained from the battery anymore. The energy stored in the resonant inductor is transferred into the resonant capacitor and to the dc-link. Hence, $i_{L_s}$ begins to decrease. When $i_{L_s}$ has reached zero, the first half-period of the conduction-cycle is over. At this time, $v_{C_s}$ has reached its peak value.

A steady-state trajectory is illustrated in Fig. 5 for a given power demand, a fixed dc-link voltage $V_{dc}$ and a fixed battery voltage $V_{bat}'$. During $T_1 = t_1 - t_0$, the trajectory lies on a circle around the excitation voltage $V_e = V_{bat}' - V_{dc}$. For the time period $T_2 = T - t_1$, $v_{d1}'$ is clamped to zero and the system trajectory lies on a circle around $-V_{dc}$. The state of $L_p$ is not considered in the diagram, because $L_p$ does not participate in the resonance.

1) **Control scheme**: In order to implement the outlined modulation scheme, the switching times $T_1$ and $T_2$ can be expressed as a function of the resonant capacitors peak voltage $v_{C_s,0} = |v_{C_s}(0)|$:

$$T_1 = \frac{\varphi_1}{\omega_0} = \frac{1}{\omega_0} \cdot \cos^{-1} \left( \frac{r_{1}^2 + V_{bat}'^2 - r_{2}^2}{2r_1V_{bat}'} \right)$$  \quad (8)$$

$$T_2 = \frac{\varphi_2}{\omega_0} = \frac{1}{\omega_0} \cdot \cos^{-1} \left( \frac{r_{2}^2 + V_{bat}'^2 - r_{1}^2}{2r_2V_{bat}'} \right)$$  \quad (9)$$

$V_{bat}'$ and $V_{dc}$ are assumed constant, the resonance frequency $\omega_0 = 2\pi f_{L_2}$ is given and $r_1$ and $r_2$ are marked in Fig. 5 and can be calculated with the law of cosine:

$$r_1 = v_{C_s,0} + V_{bat}' - V_{dc}$$  \quad (10)$$

$$r_2 = v_{C_s,0} + V_{dc}$$  \quad (11)$$

An analytic expression for the average power transfer is obtained from calculating the energy transferred each half-period:

$$W_{T/2} = \int_{0}^{t_1} V_{bat}' i_{L_s}(t) \, dt \cdot 2 \cdot f_s = V_{bat}' \int_{0}^{t_1} \frac{r_1}{Z_0} \sin(\omega_0 t) \, dt$$

$$= \frac{V_{bat}' r_1}{\omega_0 Z_0} \left[ 1 - \cos(\varphi_1) \right] \frac{1}{\omega_0} = \frac{2v_{C_s,0} V_{dc}}{Z_0 \omega_0}$$

$P_{av}$ is calculated as follows:

$$P_{av} = W_{T/2} \cdot 2f_s = \frac{2v_{C_s,0} V_{dc}}{Z_0 \omega_0} \cdot 2 \cdot \frac{1}{2(T_1 + T_2)}$$

To control the converter, a reference power demand $P_{av}$ is set and (12) is solved for $v_{C_s,0}$. The result is plugged into (8) and (9) to obtain the switching times. A PI-controller can be used.
to adjust $P_{av}$ to control the output voltage. Once $T_1$ and $T_2$ are calculated, the switching frequency is given by:

$$f_s = \frac{1}{2(T_1 + T_2)} \quad (13)$$

The switching frequency can thus not be chosen independently of the power-transfer and is in general above $f_{f1}$. In the presence of the non-idealities in an actual implementation, it may be necessary to measure the zero crossing of $i_L$, to accurately determine the switching times.

2) Switching: As the primary side performs synchronous rectification, the primary side switches are turned on and off at quasi zero current and zero voltage. On the secondary side however, a superposition of the resonant tank current $i_{Ls}$ and the magnetizing current $i_{Lp}$ has to be turned off when entering the clamped state at $t = t_1$:

$$i_{2off} = \frac{r_1}{Z_0} \sin(\varphi_1) + \frac{1}{2} \frac{n^2 V_{bat}}{T_p} V_{bat} \cdot T_1. \quad (14)$$

When leaving the clamped state at $t = T/2$, $i_{Ls}$ is zero and only the magnetizing current has to be turned off.

One could think of a resonant pulse mode, where one would wait for the resonance to complete (and thus for the $i_{Ls}$, to come down to zero by itself) to not switch the secondary side current at its peak value, but a considerable magnetizing current would have built up by then, leading to high turn-off losses on the secondary side, eliminating the potential benefits from this approach.

3) Quasi Burst Mode: Calculations show, that at light load, the switching frequency is relatively high when using the modulation scheme discussed above. To avoid this, the converter can be forced into a quasi inactive state with $v_p^s = 0$ when $i_{Ls}$ has reached zero at $t = t_2$. As long as $|V_{C0}| < V_{dc}$, the primary side diodes will not conduct. Power is transferred in small bursts during $T_1$ and $T_2$. During the newly introduced idle time $T_3 = T - t_2$, the dc-link stays energized, but no power is transferred. The two idle states are marked in Fig. 5 by two stars. Fig. 6 shows the simulation results of the LLC converter making use of the quasi burst mode. At $t = t_2$, $v_p^s$ gets clamped to zero. Notice the ringing on the diodes, as the voltage is not clamped on the primary side. When $V_{C0}$ is not very close to $V_{dc}$, additional power losses will occur due to the ringing. These have been neglected in a first approximation.

The idle-time can be expressed as an angle ($\varphi_3 = T_3 \cdot \omega_0$) for calculating the power-transfer:

$$P_{av} = \frac{W}{2} \cdot 2f_s = \frac{2V_{C0}V_{dc}}{Z_0} \cdot \frac{1}{(\varphi_1 + \varphi_2 + \varphi_3)} \quad (15)$$

B. Reverse Operation in Boost Mode

In case of $V_{dc} > V_{bat}$, the bidirectional LLC converter can operate in a backward boost mode similar to the mode mentioned in [10] for the SRC. Fig. 7 shows the characteristic currents and voltages during this mode of operation. The switching period starts at $t = t_0$ where $i_{Ls}$ is zero. In the following, the switching is explained:

$t \in [t_0, t_1]$: Before the switching cycle starts, $S_3$ and $S_2$ have been conducting on the primary side (synchronous rectification), and $S_6$ and $S_7$ have been conducting on the secondary side. At $t = t_0$, $v_p^s$ is switched from $v_p^s = -V_{bat}$ to $v_p^s = V_{bat}$ by turning $S_6$ and $S_7$ off and afterwards turning $S_5$ and $S_4$ on. The load voltage gets clamped at $V_{bat}$ (see Section III-B1 for how to achieve ZVS). The resonant tank current $i_{Ls}$ begins to rise.

$t \in [t_1, T/2]$: At $t = t_1$, $V_{bat}$ is switched to $V_{bat}$ by turning $S_3$ and afterwards turning $S_1$ on at zero voltage. Power is transferred to the dc-link. When $i_{Ls}$ has reached zero, the first half-period of the switching cycle is over.

For reasons of symmetry, a discussion of the second half-period of the conduction cycle is omitted. The steady-state trajectory is very similar to the steady-state trajectory for the boost mode and is not shown for the sake of brevity. The switching times $T_1$ and $T_2$ are calculated similar to the way shown in Section III-A:

$$T_1 = \frac{\varphi_1}{\omega_0} = \frac{1}{\omega_0} \cos^{-1} \left( \frac{r_1^2 + V_{bat}^2 - r_2^2}{2r_2 V_{bat}} \right), \quad (16)$$

$$T_2 = \frac{\varphi_2}{\omega_0} = \frac{1}{\omega_0} \cos^{-1} \left( \frac{r_1^2 + V_{bat}^2 - r_2^2}{2r_2 V_{bat}} \right). \quad (17)$$

$r_1$ and $r_2$ describe the amplitude of the oscillation:

$$r_1 = V_{C0} + V_{bat}, \quad (18)$$

$$r_2 = V_{C0} + V_{bat} - V_{bat}. \quad (19)$$

The switching frequency is again given by (13) and the power transfer is derived as shown for the buck mode:

$$P_{av} = \frac{2V_{bat}V_{C0}}{Z_0} \cdot \frac{1}{\varphi_1 + \varphi_2} \quad (20)$$
1) **Switching:** In backward boost mode, the secondary switches only have to turn off the magnetizing current, making this mode of operation far more attractive than the buck mode. Before entering the clamped switching state on the primary side at \( t = t_1 \), \( S_2 \) and \( S_3 \) are kept in the on-state for a short additional period of time. When the direction of the current has changed, \( S_2 \) is turned off and the current is used to discharge the parasitic drain-source capacitances of \( S_1 \) and \( S_2 \), until \( S_1 \) can be turned on at zero voltage. This way, the primary side is switched at zero voltage. For the sake of simplicity, the delayed switching of the primary side is not considered in the calculation of the switching times.

2) **Quasi Burst Mode:** In case of \( V_{dc} > V_{bat}' \), a quasi burst mode can be derived similar to the quasi burst mode discussed for the backward buck mode. Fig. 8 shows the characteristic currents and voltages in this mode of operation. At \( t_0 \), the converter enters the quasi-inactive state by switching \( v'_{2} \) to zero. Before leaving the inactive state, the switches \( S_2 \) and \( S_3 \) are turned on at \( t = t_3 \) to allow for quasi ZVS on the primary-side: a small current will build up in the resonant-tank which is used to discharge the parasitic drain-source capacitances of \( S_1 \) and \( S_4 \) when entering the clamped state at \( t = t_4 \). For this current to be high enough, the difference between \( v_{cc,0} \) has to be sufficiently large, which presents a trade-off. With \( V_{dc} \neq V_{0} \), the primary side will, strictly speaking, no longer switch at zero voltage at \( t = t_3 \). Simulations show, that the required voltage difference to enable ZVS is however small. Since the switching frequency is also well below the nominal operating frequency in burst mode, the partial loss of ZVS has been neglected in a first approximation.

Because of the intermediate phase between \( t_3 \) and \( t_4 \), there is no smooth transition between the burst mode and the regular backward boost mode. Consequently, the implementation of the control algorithm becomes more complex.

### IV. DESIGN

With the modulation schemes discussed, a bidirectional LLC converter can be designed. To reduce the turn-off currents on the secondary side, the turns-ratio is chosen as low as possible. Consequently, the LLC converter will have to work in buck mode in the forward direction as well. Simulations have shown, that with \( n = 12.5 \), ZVS can still be achieved on the primary side for the whole operating area when the asynchronous clamped mode is used. With the regular clamped mode, ZVS is no longer possible. For \( n = 12.5 \), the LLC converter has to operate in buck mode in the forward direction for battery voltages of \( V_{bat} < \frac{450}{1.25} = 36 \) V.

In order to cope with the high turn-off currents during reverse operation, four Infineon IPB020R045CP MOSFETs are paralleled for each switch on the secondary side. Infineon IPW60R045CP MOSFETs are used on the primary side. Both are chosen for their exceptional figure of merit [11].

The conduction losses are considered for a junction temperature of \( T_j = 125 \) °C. The current is assumed to be evenly split up among paralleled MOSFETs. The energy dissipated at turn-off in one of the secondary side switches is calculated according to [12]:

\[
E_{S,off} = (L_D + L_S) \cdot i_{S,off}^2 \cdot \frac{V_{2,pk}}{V_{3,pk} - V_2} \tag{21}
\]

\( i_{S,off} \) denotes the current flowing through a switch \( S_x \) prior to the turn-off. For a prototype design, it is presumed that a combined circuit and packaging inductance of \( L_D + L_S \approx 12 \) nH is achieved per switch\(^1\). During turn-off, the voltage across the switch will overshoot to \( V_{2,pl} \). This voltage is presumed to be no higher than the rated avalanche voltage \( V_{bat} \) of the device itself. Hence, \( V_{pl} \) of the selected switches (\( V_{bat} = 75 \) V) has been taken for the calculations. The switching losses for zero voltage switching and the switching losses for zero current switching have in a first approximation been neglected.

#### A. Design of the Resonant Tank

According to the design method discussed in [14], \( L_p \) is matched to the load when operating at resonance.

\[
L_p = \frac{n^2 R_0}{2 \pi f_{11}}. \tag{22}
\]

In a first approximation, this minimizes the conduction losses at the nominal operating point (36 V, 30 A) when operating in the forward direction [14]. \( R_{vcc,2} \) denotes the equivalent ac-resistance in (2).

To satisfy the maximum voltage demand at maximum power draw, \( L_p = 1 \cdot L_p \) is sufficient. No additional degree of freedom exists for the resonance capacitance since the nominal switching frequency is specified.

\[
C_t = \frac{1}{(2 \pi f_{11})^2 \cdot L_t}. \tag{23}
\]

A summary of the component values obtained is given in Table II.

---

\(^1\)The inductance of the To-236 package is around 5 nH [13] and an additional 7 nH have been assumed for the PCB stray inductance.

**Table I:** Specification for a light electric vehicle charger. Lithium-based batteries with voltages of 25 V, 36 V resp. 48 V can be charged and discharged.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Battery voltage range</td>
<td>( V_{bat} ) 17 V ... 56 V</td>
</tr>
<tr>
<td>DC-link voltage</td>
<td>( V_{dc} ) 450 V</td>
</tr>
<tr>
<td>Maximum output current</td>
<td>( I_{bat,max} ) ±30 A</td>
</tr>
<tr>
<td>Maximum output power</td>
<td>( P_{bat,max} ) ±1.65 kW</td>
</tr>
<tr>
<td>Nominal switching frequency</td>
<td>( f_s ) 140 kHz</td>
</tr>
</tbody>
</table>
TABLE II: Component values for the LLC converter designed for bidirectional power transfer

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transformer core material</td>
<td>EPCOS N87</td>
</tr>
<tr>
<td>Transformer turns-ratio</td>
<td>n</td>
</tr>
<tr>
<td>Series capacitance</td>
<td>C_s</td>
</tr>
<tr>
<td>Transformer leakage inductance</td>
<td>L_s</td>
</tr>
<tr>
<td>Leakage layer thickness</td>
<td>s</td>
</tr>
<tr>
<td>Transformer magnetizing inductance</td>
<td>L_p</td>
</tr>
<tr>
<td>Air gap length</td>
<td>g</td>
</tr>
<tr>
<td>Primary side MOSFETs</td>
<td>Infineon IPW60R045CP</td>
</tr>
<tr>
<td>Secondary side MOSFETs</td>
<td>Infineon IPB020NE7N3</td>
</tr>
<tr>
<td>Resonant capacitor</td>
<td>EPCOS 2 x B32652A2332J</td>
</tr>
<tr>
<td>Primary side filter capacitance</td>
<td>EPCOS B32522N6154J</td>
</tr>
<tr>
<td>Secondary side filter capacitance</td>
<td>EPCOS 6 x B32522C106K</td>
</tr>
</tbody>
</table>

The core losses are calculated with the improved generalized approach and the flux in the air-gap is assumed homogenous. The proximity effect in the windings are considered according to [15]. The transformer has been designed for minimum power losses with the help of MATLAB’s global optimization toolbox. An E-core is used to allow for simple analytic calculations of the magnetic field. The computer performs a free variation of the geometric dimensions of the core and in each step picks the best number of turns to realize the turns-ratio. The performance of the transformer is assessed at the three characteristic operating points (25 V; 30 A), (36 V; 30 A) and (48 V; 30 A). The target size of the transformer has been set to a box volume of \( V_{\text{box}} = 0.05 \text{dm}^3 \). Both the primary side windings and the secondary side windings are realized with litz wire. The internal skin effect and the internal and external proximity effect in the windings are considered according to [15]. The \( H \)-Field is calculated with a one-dimensional approach and the flux in the air-gap is assumed homogenous. The core losses are calculated with the improved generalized steinmetz equation (IGSE) [16].

Fig. 9 shows the optimized transformer design. A spacer with a thickness of \( s = 4.2 \text{ mm} \) is inserted between the windings to integrate the series inductance \( L_s \). The core has an air gap to integrate the parallel inductance \( L_p \). To keep the calculations simple, both have been regarded as decoupled from each other.

C. Filter Capacitors

The limit for the secondary side voltage ripple has been set to 5% of the output voltage and the filter capacitors are chosen accordingly. The filter capacitances have been calculated as \( C_1 > 0.24 \mu \text{F} \) and \( C_2 > 52.1 \mu \text{F} \). The primary side filter is realized with two EPCOS B32522N6154J film capacitors. On the secondary side, six EPCOS B32522C106K are paralleled. The resonance capacitance of \( C_s = 7.5 \mu \text{F} \) is realized by five EPCOS B32652A2332J film capacitors in parallel.

V. SIMULATION

A time-domain simulation has been performed with GeCoCIRCUITS [17] to verify the analytic considerations. A PI-controller is used to control the voltage by varying the switching frequency in forward mode. In forward buck-mode, the duty-cycle is calculated according to (5). In the reverse direction, the modulation schemes discussed in Section III are applied. The simulation results are discussed in the following.

A. Forward Operation

Very low power losses occur when operating the converter in the forward direction. ZVS is always achieved on the primary side, and ZCS is always achieved on the secondary side.

Fig. 11 shows the power conversion efficiency as a function of the operating point. The efficiency surface is almost flat over the entire output voltage range. At low loads, the reactive currents in the resonant tank account for a large share of the overall power losses.

Fig. 13 shows the distribution of the power losses among the different components. The semiconductor power losses are divided into switching losses \( P_{\text{sw}} \) and conduction losses \( P_{\text{cond}} \). The power losses in the transformer are composed of the winding losses \( P_{\text{wall}} \) and the core losses \( P_{\text{Fe}} \). A small share of power \( P_{\text{filt}} \) is dissipated in the filter capacitors and the resonant capacitor \( P_{\text{C_s}} \). The gate-drive power losses are denoted by \( P_{\text{gate}} \). At full load, the switching losses dominate the overall power losses.

B. Backward Operation

When the LLC converter is operated in the reverse direction, large turn-off currents are observed on the secondary side. In buck mode, the full AC-link current plus the magnetizing current has to be turned off. As a result, relatively high switching losses occur and the efficiency drops towards the upper end of the output voltage range as illustrated in Fig. 11. As the switching frequency is higher than \( f_{\text{gs}} \), the gate drive power losses are higher than in the forward direction. The different operating modes are illustrated as a function of the operating point in Fig. 10.
have to be turned off on the secondary side. This is illustrated in Fig. 10. The power losses are calculated based on the triangular and trapezoidal current mode [19]. With this approximation, the power losses dominate the overall power losses. Fig. 12 shows the current, which has to be turned off by the secondary side switches.

VI. COMPARISON WITH THE DUAL ACTIVE FULL-BRIDGE

While the unidirectional LLC converter can offer a high efficiency in configurations with only two active semiconductors [6], the bidirectional LLC converter has to compete with other full-bridge topologies. In [18], the dual-active full-bridge (DAB) has been found to be the most promising isolated bidirectional converter for a very similar voltage range and power rating. The bidirectional LLC converter is thus compared to the DAB converter.

The DAB converter is designed to operate with a combination of the trapezoidal and triangular current mode [19]. With the extensions discussed in [18], ZVS is always achieved on the primary side and ZCS is achieved on the secondary side in a large share of the operating area. To allow for a comparison, the exact same switches are chosen for both converters. Other than the LLC converter, the DAB only needs a series inductance in the ac-link. The series inductance is designed to allow the converter to fulfill the maximum power requirement. A margin is included to have sufficient reservoirs for voltage control. A turns-ratio of \( n = 12.5 \) is chosen at which the average power conversion efficiency at full output current is maximal. The transformer is optimized in the same way as done for the LLC converter. Both transformers share the same target volume of \( V_{\text{target}} = 0.05 \text{dm}^3 \). In order to have sufficient degrees of freedom for the optimization of the transformer, the turns-ratio has been rounded to half-integers. Table III summarizes the design of the DAB converter.

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transformer core material</td>
<td>EPCOS</td>
</tr>
<tr>
<td>Leakage layer thickness ( s )</td>
<td>1.6 mm</td>
</tr>
<tr>
<td>Transformer turns-ratio ( n )</td>
<td>11.5</td>
</tr>
<tr>
<td>Transformer leakage inductance ( L_a )</td>
<td>68.1 ( \mu \text{H} )</td>
</tr>
<tr>
<td>Primary side MOSFETs</td>
<td>Infineon</td>
</tr>
<tr>
<td>Secondary side MOSFETs</td>
<td>Infineon</td>
</tr>
<tr>
<td>Primary side filter capacitance</td>
<td>EPCOS B32522N6154J</td>
</tr>
<tr>
<td>Secondary side filter capacitance</td>
<td>EPCOS 4 ( \times ) B32522C106K</td>
</tr>
</tbody>
</table>

TABLE III: Component values for the designed DAB converter

A. Power Conversion Efficiency

Fig. 11 shows the calculated power conversion efficiency of the LLC converter working in the forward direction. The peak efficiency is predicted to be \( \eta_{\text{max, LLC}} = 98.3\% \). In the model, the currents and voltages have been calculated for the ideal triangular and trapezoidal current mode with an ideal DAB converter model. The power losses are calculated based on the so obtained voltages and currents in the transformer and the switches. The operating point is assumed to be independent of the power losses. With this approximation, the power losses are virtually the same for both directions of power flow. The different operating mode are illustrated as a function of the operating point in Fig. 10.

When the DAB converter is operating in forward boost-mode or backward buck-mode, also relatively high currents have to be turned off on the secondary side. This is illustrated in Fig. 12. At the lower end of the output voltage range, ZCS is achieved. A reduction of the turn-off currents could be achieved by choosing \( n \) even lower, at the cost of reducing the efficiency at the lower end of the output voltage range. Recall, that this is unpractical with the bidirectional LLC converter: to reduce the turn-off currents, a further extension of the forward buck range would be required, which would entail the loss of ZVS on the primary side at lower battery voltages.

It becomes evident that the DAB converter outperforms the bidirectional LLC converter in terms of power conversion efficiency. The turn-off currents in the DAB converter are less on the secondary side and could be reduced even further. The transformer of the DAB converter only needs to integrate the series-inductance. As a consequence, more degrees of freedom are available in the optimization process.

VII. CONCLUSION

Up until now, little information has been available on the capabilities of operating the LLC converter in the reverse direction. In this paper, possible modes of bidirectional power flow are analyzed and a bidirectional LLC converter is presented. The converter is targeted towards a light electric vehicle charger for large-quantity, small-accumulator vehicle-to-grid applications. While the converter is predicted to achieve an outstanding power conversion efficiency of \( \eta_{\text{max, LLC}} > 97.6\% \) for both directions of power-flow, some of the inherent benefits of its unidirectional originator have been lost: ZCS is no longer achieved on the secondary side when operating in the reverse direction. At high output voltages, the turn-off currents are significant and high switching losses occur. To allow for voltage control within the whole operating area, the bidirectional LLC converter needs full-bridges on both sides. Consequently, it has to compete with the DAB converter which exhibits a better overall performance and requires less passive components.

For applications with higher battery voltages and a similar output power, the turn-off currents on the secondary side voltages will be reduced. Consequently, less energy is stored in the parasitic inductances of the switches and PCB connections,
allowing the bidirectional LLC converter to achieve ZVS on both sides in both modes (low turn-on voltages are accepted in the backward boost burst mode), which is subject to further research.

**REFERENCES**


