A Hybrid PWM-Resonant DC-DC Converter for Electric Vehicle Battery Charger Applications

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Abstract

In this paper, a new hybrid DC-DC converter is proposed for electric vehicle 3.3 kW on-board battery charger applications, which can be modulated in a phase-shift manner under a fixed frequency or frequency variation. By integrating a half-bridge (HB) LLC series resonant converter (SRC) into the conventional phase-shift full-bridge (PSFB) converter with a full-bridge rectifier, the proposed converter has many advantages such as a full soft-switching range without duty-cycle loss, zero-current-switching operation of the rectifier diodes, minimized circulating current, reduced filter inductor size, and better utilization of transformers than other hybrid dc-dc converters. The feasibility of the proposed converter has been verified by experimental results under an output voltage range of 250-420 V dc at 3.3 kW.

Key words: Battery charger, Electric vehicle, Phase-shift full-bridge converter

I. INTRODUCTION

For electric vehicle (EV) applications, an AC-DC converter is required for use as a battery charger. In general, the battery charger for EVs consists of a power factor corrector and an isolated DC-DC converter that directly recharges the battery in EVs [1]-[5]. During the battery recharging process, the output of the DC-DC converter varies greatly within a specific voltage range and its operating range becomes very wide. This makes it difficult to optimally design a converter in the sense of power conversion efficiency.

A conventional phase-shift full-bridge (PSFB) converter with a full-bridge rectifier is the preferred DC-DC topology for EV on-board battery charger applications, because of its natural zero-voltage-switching (ZVS) operation, low current ripple in the battery charging current, and simple structure and control [6]-[8]. However, for wide-output-voltage-range applications like battery chargers, the conventional PSFB converter cannot obtain an optimal power conversion efficiency due to its unique drawbacks such as a narrow ZVS range, large circulating current, and high voltage stress in the rectifier diodes [9]-[11]. In order to improve the performance of the conventional PSFB converter for EV battery charger applications, hybrid dc-dc converters with an LLC series resonant converter (SRC) integrated into the PSFB converter have recently been researched [12]-[14]. By integrating an LLC SRC into a PSFB converter, the resulting hybrid dc-dc converters have many advantages such as a wide ZVS range, reduced size of the filter inductor, and reduced voltage stress. Furthermore, the hybrid converter can work with a reduced operating range during the battery recharging process. Thus, it is possible to obtain a more optimal power conversion efficiency.

This paper proposes a novel hybrid PWM-resonant converter with a half-bridge (HB) LLC SRC integrated into a conventional PSFB converter with a full-bridge rectifier (FBR). While the proposed converter has all of the benefit of the hybrid converters in [12]-[14], it has fewer components and better utilization of the transformers when compared with previous hybrid converters. The circuit configuration, operating principle, and analysis of the proposed converter are presented in this paper. Its feasibility is verified with experimental results under an output voltage range of 250-420 V dc at 3.3 kW.

II. DESCRIPTION OF THE PROPOSED CONVERTER

A circuit diagram of the proposed converter is shown in Fig. 1. As can be seen, it is based on a conventional PSFB converter with a FBR. In addition, a half-bridge (HB) LLC
SRC with a center-tap rectifier (CTR) is integrated into the conventional converter by sharing the lagging-leg switches Q3 and Q4, and stacking two rectifiers in series before the output filter stage consisting of L0 and C01. The lagging-leg switches in the conventional PSFB converter fail easily in terms of ZVS operation as the output load decreases [6]-[8]. However, the lagging-leg switches in the proposed converter can achieve ZVS operation without the effect of the output load condition due to the integrated LLC SRC as documented in [12]-[14]. Moreover, the circulating current flowing only in the primary side of the conventional PSFB converter is removed because the primary current is reset to zero level with the voltage ripple of the resonant capacitor CR1 during the freewheeling phase, as shown in the key operating waveforms of the proposed converter described in Fig. 2. In the process, the rectifier diodes D1 to D4 can be turned off with the zero-current-switching (ZCS) operation. The other output diodes Da1 and Da2 are also turned off with the ZCS due to the resonant operation of the LLC SRC. With these improvements, the proposed converter can achieve a higher power conversion efficiency when compared to the conventional PSFB converter. In addition, it can maintain a high efficiency during the battery recharging process.

A. Operation Principle

In order to analyze the operation of the proposed converter, several assumptions are made as follows.

1) The switch devices are ideal MOSFETs except for the parasitic capacitors and the internal body diodes;
2) The parasitic capacitors of all the MOSFETs have the same capacitance as Coss;
3) The rectifier diodes are ideal;
4) The output inductor L0 is large enough to be treated as a constant current source during a switching period;
5) The magnetizing inductance of T1 is very large so that the magnetizing current can be ignored;
6) The leakage inductor of T1 is included into the resonant inductor Lg.

The driving signals are the same as those for the conventional PSFB converter and the output is regulated in the phase-shift manner at a fixed frequency, which is also the same as the conventional converter. Referring to Fig. 2, there are 14 operating modes in a switching period which can be divided into two half cycles: t0-t1 (from mode 1 to mode 7) and t1-t14 (from mode 8 to mode 14). The operation principles of two half cycles are symmetric. Therefore, only the first half cycle is described and operating circuits during the cycle are shown in Fig. 3.

Mode 1 [t0-t1]: Mode 1 begins when the switches Q1 and Q2 are in the ON-state, and it ends when the switch Q1 is turned off. The output current IBattery flows through the diodes D1 to D4 and the transformers of T1 and T2. The capacitor C02 is charged with the resonant current generated by the resonance between Lg and C02 in the LLC SRC via the diode Da2. During this mode, the voltage across the resonant capacitor CR1 is linearly charged by the load current (or n1IBattery) reflected to the primary side and it can be expressed as:

\[ \Delta V_{CR1} = n_1V_{IN} + V_{O1} - \frac{V_{R}}{2} \]
converters transfer the power required in the output port.

Since the output current
\[ I_{O1} \]
through the zero level. Since the output current
\[ I_{O1} \]

7. Fig. 3. Operating circuits of the proposed converter. (a) Mode 1. (b) Mode 2. (c) Mode 3. (d) Mode 4. (e) Mode 5. (f) Mode 6. (g) Mode 7.

\[ V_{CV1}(t) = \frac{n I_{battery}}{C_{R1}} (t - t_s) - 0.5 \Delta V_s, \]

where \( \Delta V_s \approx \frac{n I_{battery} T_s}{C_{R1}} \).

The rectifier output voltage \( V_{rec}(t) \) is the sum of the voltages of the transformers and it is expressed as:
\[ V_{rec}(t) = V_{P1}(t) + V_{P2}(t) = n (V_{PSFB} - V_{CV1}(t)) + V_{O1}. \]  

From eq. (2), it is known that both the PSFB and LLC converters transfer the power required in the output port.

**Mode 2** \([t_1-t_2]\): Mode 2 begins when the switch \( Q_1 \) is turned off and it ends when the primary voltage \( V_{P1}(t) \) reaches the zero level. Since the output current \( I_{battery} \) still flows through \( D_{1\&2} \), the voltages across \( Q_1 \) and \( Q_2 \) are linearly charged or discharged by the energy stored in the output inductor \( L_0 \) as follows:

\[ V_{Q1}(t) = \frac{n I_{battery}}{2C_{OSS}} (t - t_s) \quad \text{and} \quad V_{Q2}(t) = V_{IN} - V_{Q1}(t) = V_{P1}(t). \]  

\( V_{P1}(t) \) decreases linearly to the zero level as \( V_{Q1}(t) \) and \( V_{Q2}(t) \) also ramps down to the output of the LLC SRC, \( V_{O1} \). The equation for \( V_{Q2}(t) \) is the same as Equation (2). The resonant current of the LLC SRC continuously charges \( C_{O1} \).

At the end of this mode, the switch \( Q_1 \) is turned on with ZVS.

**Mode 3** \([t_1-t_3]\): Mode 3 begins when the switch \( Q_2 \) is turned on with ZVS, and it ends when the primary current \( i_{P1}(t) \) reaches the zero level. During this mode, \( V_{P1}(t) \) is zero, \( i_{P1}(t) \) is reset to the zero level with the voltage ripple of \( C_{R2} \) as in eq. (4), and commutation occurs between the output diodes \( D_{1\&3} \) and \( D_{2\&3} \).
\[
i_{p}(t) = n_{\text{battery}} \cdot \frac{0.5 \Delta V}{L_{m1}} (t - t_2).
\]  

\( V_{\text{rec}}(t) \) is the output of the LLC SRC, \( V_{O1} \). This means that only the LLC SRC transfers the power required in the output port.

In this mode, the circulating current existing in the conventional PSFB converter is removed, and the output diodes \( D_{163} \) are turned off with ZCS. Consequently, the conduction power loss in the active switches and the switching power loss in the rectifier diodes are reduced during the battery recharging process.

**Mode 4** \([t_2-t_3]\): Mode 4 begins when the commutation between \( D_{163} \) and \( D_{9} \) is completed, and it ends when the power transfer by the LLC SRC (from \( V_{IN} \) to \( C_{O1} \)) is completed. In this mode, the output power is supplied from the output of the LLC SRC as in Mode 3. At the end of this, the diode \( D_{9} \) is turned off with ZCS.

**Mode 5** \([t_3-t_4]\): Mode 5 begins when the power transfer by the LLC SRC is completed, and it ends when the switch \( Q_{4} \) is turned off. The output power is supplied from the output of the LLC SRC as always. The primary current \( i_{P2}(t) \) becomes the magnetizing current of the transformer \( T_{2} \) in this mode.

**Mode 6** \([t_4-t_5]\): Mode 6 begins when \( Q_{1} \) is turned off, and it ends when \( V_{P2}(t) \) reaches \(-V_{IN} \). During this mode, the voltages across \( Q_{1} \) and \( Q_{2} \) are charged or discharged by the energy stored in the magnetizing inductor \( L_{m2} \) of \( T_{2} \) as follows:

\[
V_{O1}(t) = \frac{i_{P2}(t)}{2C_{OSS}} (t - t_3) = \frac{i_{P2}(t)}{2C_{OSS}} (t - t_6) \quad \text{and} \quad V_{op}(t) = V_{IN} - V_{O1}(t).
\]  

At the end of this mode, the switch \( Q_{4} \) is turned on with ZVS.

**Mode 7** \([t_5-t_6]\): Mode 7 begins when the switch \( Q_{5} \) is turned on with ZVS. Since \( V_{P2}(t) \) is the negative input voltage, \( i_{P2}(t) \) is decreased as in eq. (6) and the commutation between the output diodes \( D_{164} \) and \( D_{9} \) occurs. The resonance between \( L_{R} \) and \( C_{O2} \) is progressed in the LLC SRC and the resonant current generated by the resonance starts to flow to \( C_{O2} \) via \( D_{9} \).

\[
i_{p}(t) = \frac{V_{IN} - 0.5 \Delta V}{L_{m2}} (t - t_4).
\]  

The rectifier output voltage \( V_{rec}(t) \) is equal to eq. (2), and both the PSFB and LLC converters start to transfer the power required in the output port. This mode ends when the commutation between \( D_{164} \) and \( D_{9} \) is completed.

**B. Voltage Conversion Ratio, M**

The \( V_{rec}(t) \) in Fig. 2 can be simplified as in Fig. 4(a) after ignoring the durations of modes 2 and 6 because their durations are very narrow in practice. Fig. 4(b) shows \( V_{rec}(t) \) in the conventional PSFB converter with a FBR. As shown in the figure, the conventional PSFB converter processes only the power transfer within the powering interval defined as the effective duty-cycle time, \( DT_{S} \). On the other hand, the proposed converter continuously transfers the power required in the output port with the help of the LLC SRC. Due to the difference between these powering concepts, the proposed converter can have a lower turns-ratio of \( T_{i} \) when compared with the conventional PSFB converter as in eq. (7).

\[
n_{\text{1\_proposed\_converter}} < n_{\text{1\_conventional\_converter}}. \quad (7)
\]

Since the lower turns-ratio reduces both the conduction power loss generated in the primary side of the transformers and the voltage stress in the rectifier diodes, it is possible that the proposed converter gets a power conversion efficiency that is better than the conventional PSFB converter for charging the battery.

Since the voltage across \( C_{Ri} \) in the proposed converter is small in practice, consider the \( V_{rec}(t) \) without the effect of \( C_{Ri} \) in Fig. 4(a) for getting the voltage conversion ratio of the proposed converter. The voltage conversion ratio, \( M \) can be obtained by averaging \( V_{rec}(t) \) as:

\[
M(D) = \frac{V_{O1}}{V_{IN}} = 2n_D + 0.5n_L M_{\text{LLC\_SRC}}, \quad (8)
\]

where \( D \) is the duty-cycle that is a function of the phase-shifted time, \( T_{ph} \) as \( D = T_{ph} / T_{S} \).

In eq. (8), \( M_{\text{LLC\_SRC}} \) is the gain of the LLC SRC and is defined as:
more that the turns-ratio of conventional PSFB converter. From this, it is known once of the proposed converter is always larger than that of the magnitude as the switching frequency, \( C \).

This scheme lowers the utilization of the powering sections of the PSFB converter and the LLC SRC are separated over a switching period. This scheme lowers the utilization of the PSFB transformer as:

\[
M_{LLC\_SRC} = \frac{V_o}{n_2(V_i/2)} = \frac{k}{\sqrt{1+Q^2(k^2(f_s - 1)/f_R)}},
\]

where:

\[
Q = \frac{1}{R_w} \sqrt{\frac{L_R}{L_{R2}}}, \quad k = \frac{n_{o2}}{L_R}, \quad f_R = \frac{1}{2\pi \sqrt{L_k C_{R2}}}, \quad f_s = \frac{f_s}{f_R},
\]

\[
R_w = \frac{8R_L}{(n_2 \pi)^2}, \quad R_L = V_o/I_{battery}, \text{ and } f_s \text{ is the switching frequency.}
\]

From eq. (8), it is known that the output of the proposed converter can be regulated with three methods: phase-shift modulation at a fixed frequency, frequency modulation at a fixed duty-cycle, and hybrid modulation of the duty-cycle and frequency. In this paper, the control is focused on only the phase-shift modulation at a fixed frequency, which is the same as the conventional PSFB converter.

Fig. 5 shows the curve of \( M_{LLC\_SRC} \). If the resonant frequency of the LLC SRC, \( f_R \) is designed with the same magnitude as the switching frequency, \( M_{LLC\_SRC} \) has a gain value of 1.0 as shown in the figure. Then, eq. (8) can be further simplified and \( M \) can be expressed by normalizing the voltage conversion ratio based on the turns-ratio \( n_1 \) of the PSFB transformer as:

\[
M_{\text{norm, proposed}}(D) = \frac{V_o}{n_2 V_{IN}} = 2D + 0.5\alpha,
\]

where \( 0 < \alpha = n_2/n_1 \leq 1 \).

Fig. 6 shows the curve of eq. (10). As can be seen, the gain of the proposed converter is always larger than that of the conventional PSFB converter. From this, it is known once more that the turns-ratio of \( T_1 \) in the proposed converter can be designed with a smaller value when compared with the conventional converter as in eq. (7).

**C. Transformer Utilization, T.U**

Fig. 4(c) shows the \( V_{rec}(t) \) of the hybrid converters in [12] and [13]. As can be seen in the figure, the powering sections of the PSFB converter and the LLC SRC are separated over a switching period. This scheme lowers the utilization of the transformers for the battery charger applications as mentioned in [15]. For example, if the converters charge a battery with the CC-CV charging strategy, the \( T.U \) in the converters can be calculated as follows:

\[
T.U = \frac{P_{\text{out(max)}}}{P_{\text{PSFB(max)}} + P_{\text{LLC\_SRC(max)}}} = \frac{V_{O\_\text{max}} \cdot I_{\text{Battery}}}{\left( V_{O\_\text{max}} - V_{O\_\text{min}} \right) \cdot I_{\text{Battery}} + 200(V_o \cdot I_{\text{Battery}})} = 0.67.
\]

On the other hand, the PSFB converter in the proposed converter always receives the help of the LLC SRC in processing the power transfer. In this structure, the \( T.U \) is calculated as follows:

\[
T.U = \frac{V_{O\_\text{max}} \cdot I_{\text{Battery}}}{(V_{O\_\text{max}} - V_{O\_\text{min}}) \cdot I_{\text{Battery}} + 200(V_o \cdot I_{\text{Battery}})} = 1.0
\]

Consequently, the proposed converter has better utilization of the transformers when compared with the hybrid converters in [12] and [13].

**D. Reduction of the Circulating Current**

The conventional PSFB converter working in the phase-shift manner has freewheeling phases. During the freewheeling phases, the non-powering current circulates in the primary side of the transformer. This current is officially called the circulating current, and it is pinpointed as a significant drawback of the conventional PSFB converter operating under wide output voltage range applications like battery chargers. This is because the freewheeling phases are maximized when the battery voltage is at its minimum value, and then the conduction power loss coming from the circulating current is maximized.

On the other hand, the circulating current is minimized in the proposed converter due to the voltage ripple of \( C_{R1} \) as analyzed in detail in mode 3. As a result, the RMS value of the primary current can be reduced and the conduction power...
loss can also be reduced.

E. Soft-Switching Characteristics

In the conventional PSFB converter, the ZVS operation of the leading-leg switches is less sensitive to output load conditions due to the large ZVS energy coming from the large output filter inductor. On the other hand, the lagging-leg switches fail very easily in terms of the ZVS operation as the output load current decreases. This is because the source of the ZVS energy is the energy stored in the small leakage inductor of $T_1$ and this energy is very small.

The ZVS operation of the leading-leg switches in the proposed converter is obtained with the same principle as the conventional PSFB converter. Hence, it is less sensitive to the output load condition. Meanwhile, the ZVS of the lagging-leg switches ($Q_3$ and $Q_4$) is achieved with the energy stored in the magnetizing inductor of $T_2$ as explained in eq. (5). This is the same ZVS principle as LLC SRC. Therefore, the lagging-leg switches are always able to operate with ZVS without the effect of the output load.

All of the rectifier diodes in the conventional PSFB converter are under hard switching status. Meanwhile, for the proposed converter, the diodes $D_1$ and $D_2$ are turned off with ZCS due to the voltage ripple of $C_{R1}$ as explained in mode 3. The diodes $D_{a1}$ and $D_{a2}$ are also turned off with ZCS due to the resonant process of the LLC SRC.

F. Duty-Cycle Loss

For a wide ZVS range, the conventional PSFB converter requires the addition of an external inductor. However, this method greatly increases the commutation time of the primary current and a very large duty-cycle loss occurs. To compensate for the increased duty-cycle loss, the turns-ratio of the transformer $(n)$ should be designed with a larger value. As a result, both the current stress in the primary side of the transformer and the voltage stress of the rectifier diodes increase significantly.

Meanwhile, the ZVS range in the proposed converter is extended without adding an external inductor. As a result, the commutation time of the primary current is minimized, and the effective duty-cycle $D_{eff}$ comes close to the real duty-cycle $D$, which is defined as $D = \frac{T_p}{T_s}$. Consequently, the turns-ratio of the PSFB transformer $T_1$ can be designed to be more beneficial when compared with the conventional PSFB converter.

G. Filter Requirements

In the proposed converter, the output power is supplied continuously from the input stage over a switching period. Meanwhile, the conventional PSFB converter processes only the required power within the effective duty-cycle time due to the freewheeling phases. As a result of this difference, the proposed converter can have a much lower $V_{level}$ and a smaller current ripple of the output inductor $\Delta I_{Lo}$ than the conventional converter, as shown in Fig. 7. In other words, the output inductor in the proposed converter can be designed with a much smaller inductance when compared with the conventional converter under the same specifications for the current ripple. Eq. (13) shows how to select a magnetic core for the design of the output inductor. As in the equation, the proposed converter can employ an output inductor with a smaller size due to the smaller inductance.

$$A_p > \frac{L \cdot I_{Lo}^2}{k_w \cdot \Delta B \cdot J}.$$  \hspace{1cm} (13)

where $A_p$ is the window-area product of the magnetic core, $L$ is the required inductance, $I_{Lo}$ is the RMS or DC value of the current flowing through the inductor, $\Delta B$ is the working flux density, $J$ is the current density, and $k_w$ is the winding factor.

III. EXPERIMENTAL RESULTS

To verify the effectiveness of the proposed converter, a 3.3 kW prototype circuit has been built with the components listed in Table I and tested with the battery charging strategy.
Fig. 9. Operating waveforms ($i_{p1}(t)$, $i_{p2}(t)$, $V_{p1}(t)$ and $V_{rec}(t)$) of the proposed converter during constant-current mode with the charging current of 7.85A: (a) when $V_o=250V$, (b) when $V_o=300V$, (c) when $V_o=350V$, and (d) when $V_o=420V$ (100% load condition).

Fig. 10. ZVS waveforms of lagging-leg switches during constant-current mode with the charging current of 7.85A: (a) when $V_o=250V$, (b) when $V_o=300V$, (c) when $V_o=350V$, and (d) when $V_o=420V$ (100% load condition).
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Fig. 11. ZVS waveforms of lagging-leg switches during constant-voltage mode at 420V: (a) when $I_{O}=0.78A$ (10% load), (b) when $I_{O}=1.57A$ (20% load), and (c) when $I_{O}=3.93A$ (50% load).

Consisting of the constant-current and constant voltage modes.

Fig. 8 shows the prototype converter. The prototype circuit has the following specifications:
- Input voltage: $V_{IN}=385$ V
- Output voltage: $V_{O}=250-420$ V
- Charging current during CC mode: $I_{Battery}=7.85$ A
- Switching frequency: $f_{S}=100$ kHz

The resonant tank of LLC SRC was designed with the following procedure. Firstly, in order to guarantee the ZVS operation of the lagging-leg switches, the following equation should be satisfied:

$$\frac{1}{2}L_{m2}i_{p2}^2(t_i) \geq \frac{1}{2}C_{oss}V_{IN}^2$$

where $i_{p2}(t_i) = \frac{V_{IN}}{4n_2f_{S}L_{m2}}$.  

Eq. (13) can be represented as follows:

$$L_{m} \leq \frac{1}{C_{oss}}\left(\frac{V_{IN}}{4n_2f_{S}L_{m2}}\right)^2.$$  

Considering the power handling capacity of the half-bridge LLC SRC in EV battery charger applications, the output voltage of the LLC SRC, $V_{O}$, was set as 200V. In order to minimize the circulating current generated by the LLC SRC, the resonant frequency ($f_{R}$) was designed with the same magnitude as the switching frequency ($f_{S}$). For the LLC SRC, the resonant frequency is the load-independent frequency to guarantee the ZVS turn on of the lagging-leg switches from zero to full load. In addition, the gain of the LLC SRC becomes 1.0 at this frequency. Then, in turn, from eq. (9), it can be noted that $n_{2}$ is equal to 1.0. According to equation (14), with the design parameters of $n_{2}=1.0$, $V_{IN}=385$V, $V_{O}=200$V, $C_{oss}$ of IPP60R074C6, and $f_{S}=100$ kHz, the magnetizing inductance of $T_{2}$, $L_{m2}$ was designed as 120μH. In designing the LLC SRC, $k$ is commonly designed in the range from 4.0 to 6.0 considering the input voltage variation and the conduction power loss. In this paper, $k$ was selected as 6.0. Then, the resonant inductance for $L_{R}$ is given as 20μH. Finally, the resonant capacitor $C_{R2}$ can be designed with the following equation:

$$C_{R2} = \frac{1}{(2\pi f_{R})^2 L_{R}} = 100nF.$$  

Fig. 10 and 11 show the ZVS waveforms of the lagging-leg switches during the battery charging process. As can be seen, the lagging-leg switches are successfully turned on with ZVS under the overall load conditions due to the combination with an LLC resonant converter.

Fig. 12 shows the efficiency measured during the battery charging process:

$$\frac{1}{2}L_{m2}i_{p2}^2(t_i) \geq \frac{1}{2}C_{oss}V_{IN}^2$$

where $i_{p2}(t_i) = \frac{V_{IN}}{4n_2f_{S}L_{m2}}$.  

Eq. (13) can be represented as follows:

$$L_{m} \leq \frac{1}{C_{oss}}\left(\frac{V_{IN}}{4n_2f_{S}L_{m2}}\right)^2.$$  

For the control of the battery charging based on the CC-CV charging strategy, two control loops were used: an inner current control loop and an outer voltage control loop. In addition, two PI-type controllers were used to regulate the battery voltage and the charging current. The entire control diagram was digitally implemented using a TMS320F28069.

Fig. 9 shows the key operating waveforms of the proposed converter, while the battery is being charged from 250V to 420V with a constant-current of 7.85A. As shown in the figure, all of the measured waveforms follow the theoretical waveforms described in Fig. 2 well. Moreover, it is confirmed that the proposed converter has reduced the circulating current in the primary side when the primary voltage $V_{Pr}(i)$ is at the zero level during the battery charging process. It can also be seen that there is a staircase voltage waveform in the output of rectifier, $V_{rec}$ in the figure. From this, it is known that the output power is transferred continuously from the input stage to the output stage, and that the size of the output filter inductor can be smaller than that of the conventional PSFB converter.

Figs. 10 and 11 show the ZVS waveforms of the lagging-leg switches during the battery charging process. As can be seen, the lagging-leg switches are successfully turned on with ZVS under the overall load conditions due to the combination with an LLC resonant converter.

Fig. 12 shows the efficiency measured during the battery charging process:
charging process. As shown in Fig. 12, the proposed converter has better performance than the conventional PSFB converter. This is due to the reduced circulating current, wider ZVS range, ZCS operation of all of the rectifier diodes, and lower turns-ratio. The proposed converter has a maximum efficiency of 97.74% under the full load condition and it maintains high efficiency during the battery charging process.

IV. CONCLUSION

In order to improve the performance of the conventional PSFB converter for EV battery charger applications, hybrid type converters have been recently researched. In this paper, a new hybrid DC-DC converter with a full-bridge rectifier has been presented. While the proposed converter has all of the benefit mentioned for the hybrid converters in [12]-[14], the proposed converter has fewer components and better utilization of the transformers when compared with the previously presented hybrid converters. A summary of the advantages is given as follows:

1) Full ZVS range of all of the active switches without a duty-cycle loss
2) ZCS turn-off operation of the rectifier diodes
3) Minimized circulating current
4) Smaller output filter inductor
5) Better utilization of the transformers for battery charger applications

Experimental results with a 3.3 kW prototype converter show a 97.74% peak efficiency and a high efficiency during the battery charging process. From these results, it can be said that the proposed converter is suitable for EV battery on-board charger applications.

REFERENCES


Il-Oun Lee received his B.S degree in Electrical and Electronic Engineering from Kyungpook National University, Taegu, Korea, in 2000; his M.S. degree in Electrical Engineering from Seoul National University, Seoul, Korea, in 2002; and his Ph.D. degree from the Korea Advanced Institute of Science and Technology (KAIST), Daejeon, Korea, in 2013. From 2003 to 2008, he was a R&D Engineer in the PDP (Plasma Display Panel) Development Group, Samsung SDI, Korea, where he was involved in circuit and product development. From 2008 to 2013, he was a Senior Engineer in the Power Advanced Development Group, Samsung Electro-Mechanics Co. LTD., Korea, where he was involved in the development of EV battery chargers, LED lighting drivers, high efficiency server or network power supplies, and high power density adapters. From 2013 to 2015, he was a Senior Researcher in the Energy Saving Lab., Korea Institute of Energy Research (KIER), Daejeon, Korea, where he carried out research on advanced high reliability microgrids and the development of a simulator system for a 20kW diesel engine generator. In 2015, he joined the School of Electrical and Electronic Engineering, Keimyung University, Daegu, Korea, as an Associate Professor. His current research interests include DC-DC converters, power factor correction (PFC) AC-DC converters, LED drivers, battery chargers for electric vehicles, digital display power systems, and digital control approaches for DC-DC converters.