An Isolated/Bidirectional PWM Resonant Converter for V2G(H) EV On-Board Charger

Byung-Kwon Lee, Jong-Pil Kim, Sam-Gyun Kim, and Jun-Young Lee, Member, IEEE

Abstract—This paper suggests another candidate for isolated/bidirectional DC/DC converter in electric vehicle (EV) on-board charger (OBC) based on PWM resonant converter (RC). The PWM-RC has good switching characteristics but it is not adequate for bidirectional applications because it is always operated under ‘buck’ type operation regardless of power flow directions. This problem can be solved by structure change method, which increases the converter gain into double. Also, additional technique to increase the converter gain during discharging operation is suggested by analysis of the gain characteristics. The feasibility of bidirectional PWM-RC is verified with a 6.6kW prototype charger.

Index Terms—Battery charger, Resonant converter, Bidirectional converter

I. INTRODUCTION

INCREASE of transportation energy usage and worries of air pollution have expedited vehicle electrification. Since electrical vehicles (EVs) have a high potential to reduce emission of greenhouse gasses and gasoline usage, they are expected to become commonly used all around the world in the near future as green transportation system. EVs use a high capacity battery pack that can be recharged through power grid so that they provide an opportunity to employ vehicle-to-grid (house) (V2G(H)) technology. V2G(H) technology makes EV an energy storage device by providing the battery power to the grid or the local loads. A large number of EVs interconnected with the utility will be an alternative solution to stabilize the intermittent renewable energy sources and emergency power supply [1–4]. In order for EVs to be operated with V2G(H) function, bidirectional on-board charger (OBC) that regulates bidirectional power flow is essential equipment. Conventional EV bidirectional OBCs are comprised of bidirectional AC/DC converter for harmonic regulation and grid-tied operation followed by isolated/bidirectional DC/DC converter for battery charging and discharging operations. For AC/DC converter, H-bridge-based converters such as single-phase half-bridge, single-phase full-bridge, and three-phase full-bridge converters are generally used for grid-connected converters and multilevel topologies are also considered in high power chargers [5–7].

Recently, many efforts have been focused on improvement of the efficiency in conventional AC/DC converter structures and good efficiencies over 97–99% have been reported based on new devices such as silicon carbide (SiC) devices [8–14]. It means that there are many solutions available for improvement of power density and efficiency in grid-tied converter of bidirectional OBCs. Accordingly, overall efficiency and size are strongly affected by bidirectional DC/DC stage as a result. To be able to be used for battery charger and improve OBC efficiency that seriously affects mileage, DC/DC stage should meet several requirements such as wide output voltage regulation, low electrical stresses, no snubber circuitry, low circulating current, and good switching condition. Above all, the critical requirement for bidirectional application is whether buck/boost operation is available or not [15, 16]. Conventional isolated/bidirectional PWM converters have common structure with voltage-fed side comprised of voltage source and high frequency (HF) inverter and current-fed side comprised of inductor and HF inverter. This structure meets the critical requirement of buck/boost operation but HF inverter in current-fed side suffers from heavy voltage stress caused by transformer leakage inductance [16, 18]. Moreover, this problem becomes worse in high voltage applications such as OBC, which makes it impossible to use high frequency switches of MOSFET even snubber circuitry is adopted. It is a major obstacle for size reduction and efficiency improvement of OBC [17–20]. For high voltage isolated/bidirectional converters, symmetrical structure that has voltage-fed input and voltage-fed output is very advantageous. This structure includes resonant converters and dual-active-bridge (DAB) converter. Among them, DAB is more adequate for bidirectional applications because it has the ‘buck/boost type’ operation for bidirectional power transfer, the ability to accommodate a wide range of voltage levels, and a high power capability [21].

However, the conventional DAB converter has large reactive current flows, which provides electrical stress on its switching elements and

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increase of power losses [22]. To solve these problems and improve the efficiency, various control methods and applications of new semiconductor devices have been suggested [24-26].

In this paper, another candidate for isolated/bidirectional OBC is proposed using PWM resonant converter (RC). The PWM-RC has constant frequency PWM control and good switching characteristics. By proper design of the resonant tank, upper switches can be operated with zero-voltage turn-on, and other switching devices including bottom switches and output rectifier can have zero-voltage turn-on and zero-current turn-off [27]. This PWM-RC can be modified for bidirectional power flow by replacing rectifier with switches. However, the bidirectional PWM-RC is always operated with ‘back type’ operation regardless of power flow directions so that it is difficult to have discharging operation over entire battery voltage range. This defect can be overcome by structure change method that changes the rectifier structure to increase the voltage gain into double. Also, an additional technique to increase the converter gain during discharging operation is suggested by analysis of the gain characteristics.

II. ANALYSIS OF THE PROPOSED CONVERTER

A. Description of the proposed converter

Fig. 1 is the schematic diagram of the proposed bidirectional DC/DC converter. The converter has symmetrical structure like DAB converter. It has 8 switches for charging or discharging operations. Resonant tank comprised of resonant capacitor $C_r$ and resonant inductor $L_r$ is located on the secondary side for resonant PWM operation and the capacitor $C_r$ located on the primary side is for voltage-doubling operation. Since the converter is controlled only by PWM different from conventional resonant converters, efficiency degradation due to excessive high switching frequency, and audible noise or no-load regulation problems due to excessive low switching...
frequency can be avoided [27]. The proposed bidirectional charger keeps structural advantages similar to DAB converter such as low voltage stress of switches and absorption of transformer leakage inductance by circuit parameter. Different from DAB converter, the proposed converter operates with ‘buck type’ regardless of power flow direction, which makes it difficult for the converter to form bidirectional power flow. To solve this problem, the proposed converter adopts voltage-doubler rectification structure in case of discharging operation, which increases the voltage gain into double [28]. This function can be implemented by keeping $M_3$ on conducting-state during discharging operation. The operation will be explained in the next section.

B. Mode analysis of charging operation

In charging operation, the power flow is controlled by $M_1$-$M_4$ and body diodes of $M_5$-$M_8$ are used for full-bridge rectifier. In this case, $C_s$ is used for DC-coupling capacitor so that it does not affect the operation if $C_s$ has a sufficiently large value. Fig. 2 is key waveforms for mode analysis of charging and discharging operations. In this figure, symbols in parentheses are for discharging operation. Mode diagrams and their equivalent circuits in charging operation are in Fig. 3. Before analysis, it is assumed that all switching devices are ideal without drain-source capacitance $C_{ds}$ and the resonant capacitor voltage $v_c$ does not exceed the battery voltage $V_{batt}$. Also, $M_1$ is on conducting-state before mode 1.

Mode 1 ($t_0 \leq t < t_1$): When $M_1$ is turn on at $t_0$, the primary current $i_p$ flows through $M_1$, $M_s$, and the transformer primary side. The secondary current $i_s$ increases from zero through $L_s$ and $C_r$, body diodes of $M_3$ and $M_7$, and transformer secondary side. From the equivalent circuit of mode 1, $v_{cr}$ and $i_s$ can be derived as

$$v_{cr}(t) = \left(\frac{V_{in}}{n} - V_{batt}\right) - \left(\frac{V_{in}}{n} + V_{crf} - V_{batt}\right) \cos \omega_r(t - t_0)$$  \hspace{2cm} (1)

$$i_s(t) = \frac{1}{Z_{cr}} \left(\frac{V_{in}}{n} + V_{crf} - V_{batt}\right) \sin \omega_r(t - t_0)$$  \hspace{2cm} (2)

where $\omega_r = 1/(L_sC_r)^{0.5}$ and $Z_{cr} = (L_sC_r)^{0.5}$. $V_{crf}$ is the peak voltage of $v_{cr}$ in charging operation and its expression is as shown in eq. (8). The primary current $i_p$ is the sum of $i_s$ referred to the primary side and the magnetizing current $i_{gm}$. When $M_1$ is turned off, the primary current is used to charge and discharge $C_{ds}$ of $M_1$ and $M_2$. If $C_{ds}$ of $M_2$ is completely discharged before the end of mode 3, ZVS of $M_1$ can be accomplished. Different from ZVS condition of $M_2$, ZVS of $M_4$ is not easy because only the peak value of the magnetizing current is used for ZVS and it is affected by load condition. The

$$i_p(t) = -\frac{1}{Z_{cr}} (v_{cr}(t) + V_{batt}) \sin \omega_r(t - t_0) + i_{gm}(t_0) \cos \omega_r(t - t_0)$$  \hspace{2cm} (4)

With the assumption that dead-times between gate signals of upper and lower switches are short enough to be neglected, $i_p(t)$ can be found from eqs. (1) and (2) as follows:

$$v_{cr}(t) = v_{cr}(D/T_s) = \left(\frac{V_{in}}{n} - V_{batt}\right) - \left(\frac{V_{in}}{n} + V_{crf} - V_{batt}\right) \cos \omega_r D/T_s$$  \hspace{2cm} (5)

$$i_s(t) = i_s(D/T_s) = \frac{1}{Z_{cr}} \left(\frac{V_{in}}{n} + V_{crf} - V_{batt}\right) \sin \omega_r D/T_s$$  \hspace{2cm} (6)

where $D$ is the duty-ratio of $M_1$ (or $M_2$) in charging operation. From eqs. (4), (5), and (6), the duration of mode 2, $T_{M2f}$, can be derived as

$$T_{M2f} = \frac{1}{\omega_r} \tan^{-1} \left( -\frac{V_{in} - V_{batt} - V_{crf}}{V_{in} + V_{crf} - V_{batt}} \sin \omega_r D/T_s \right)$$  \hspace{2cm} (7)

Mode 3 ($t_2 \leq t < t_3$): After mode 2, only the magnetizing current $i_{gm}$ circulates through $M_1$ and the body diode of $M_2$. During mode 3, the resonant capacitor voltage $v_{cr}$ is kept as $v_{cr}(t_2)$ which is the peak value of the resonant capacitor defined by $V_{crf}$. From eqs. (3) and (7), $V_{crf}$ can be derived as follows:

$$v_{cr}(t_2) = \frac{V_{in} \left(\frac{V_{in}}{n} - V_{batt}\right)}{2V_{batt}} \left(1 - \cos \omega_r D/T_s\right) = V_{crf}$$  \hspace{2cm} (8)

$V_{crf}$ should not exceed $V_{batt}$ to guarantee the normal operation by preventing body diodes of $M_3$ and $M_7$ from conducting abnormally. Referring to key waveform in Fig. 2 and mode diagram in Fig. 3, only the peak magnetizing current is used to charge and discharge $C_{ds}$ of $M_1$ and $M_2$. If $C_{ds}$ of $M_2$ is completely discharged before the end of mode 3, ZVS of $M_1$ can be accomplished. Different from ZVS condition of $M_2$, ZVS of $M_4$ is not easy because only the peak value of the magnetizing current is used for ZVS and it is affected by load condition. The

![Fig. 5 Equivalent circuits of modes 1–3(a) and their modifications (b)](image-url)
operation of the next half-cycle comprised of modes 4–6 is same as that of the first half-cycle.

C. Mode analysis of discharging operation

The power flow is controlled by $M_5$–$M_8$ in discharging operation. To increase the voltage gain, the rectifier using body diodes of $M_1$ and $M_2$ is operated as voltage-doubler rectifier by keeping $M_3$ on conducting-state. The capacitor $C_r$ is used for charge-pump capacitor and the capacitor voltage $v_{cr}$ becomes $V_{batt}/2$. Mode diagrams and their equivalent circuits are as shown in Figs. 4 and 5(a). Assuming that the impedance of the magnetizing inductance refered to the transformer secondary side, $\omega_LM/n^2$ is sufficiently large compared with the resonant tank impedance of $\omega_LL_1+1/(\omega_Cr)$, the equivalent circuit can be redrawn as Fig. 5(b) by exchanging locations of $L_m/n^2$ and series network of $L_r$ and $C_r$. Where, $\omega_s$ is switching frequency expressed in rad/sec.

Mode 1 ($t_0 \leq t < t_1$): When $M_5$ is turned on at $t_0$, the secondary current $i_{cr}$ flows through $L_r$, $C_r$, $M_6$, $M_7$, and transformer secondary side. The primary current $i_p$ increases from zero through the body diode of $M_2$, the channel of $M_3$, $C_r$ and the transformer primary side. From the equivalent circuit in Fig. 5(b), $v_{cr}$ and the primary current referred to the secondary side $i_{pp}$ can be approximately derived as

$$v_{cr}(t) = \left( V_{batt} - \frac{V_{batt} + V_{err} - V_m}{2n} \right) \sin \omega_t (t - t_0)$$

$$i_{pp}(t) = -\left( V_{cr}(t) + \frac{V_m}{2n} \right) \cos \omega_t (t - t_0)$$

$V_{err}$ is the peak voltage of $v_{cr}$ in discharging operation and its expression is as shown in eq. (16). The secondary current $i_s$ is equal to the sum of $i_{pp}$ and $i_{in}$. When $M_6$ is turned off, $i_s$ is used to make ZVS condition of $M_5$.

Mode 2 ($t_1 \leq t < t_2$): When $M_7$ is turned on, mode 2 begins. The power flow is formed as shown in Fig. 4 and $i_p$ begins to decrease. From the equivalent circuit of mode 2, $v_{cr}$ and $i_{cr}$ can be derived as

$$v_{cr}(t) = \left( V_{batt} - \frac{V_{batt} + V_{err} - V_m}{2n} \right) \sin \omega_t (t - t_1)$$

$$i_{cr}(t) = \left( V_{cr}(t) + \frac{V_m}{2n} \right) \cos \omega_t (t - t_1)$$

Eq. (16) is the peak value of $v_{cr}$ during discharging operation defined by $V_{cr}$. When $M_6$ is turned off, the peak magnetizing current referred to secondary side is used to charge and discharge $C_{ab}$’s of $M_5$ and $M_7$. The first half-cycle comprised of modes 1–3 is charging period of the charge-pump capacitor $C_r$ through the body diode of $M_2$ and the channel of $M_3$, and the next half-cycle comprised of modes 4–6 is powering period through the charge-pump capacitor $C_r$ and the body diodes of $M_1$ and $M_3$. The two operations are same except for rectifier operation.

D. Input-output relationship of the proposed converter

< Charging operation >

The average input power $P_{gy}$ can be calculated by averaging the product of switch current of $M_1$–$M_4$ and the input voltage $V_m$ over half of the switching cycle. Since the magnetizing current is circulating current, it does not contribute to power transfer. Accordingly, $P_{gy}$ can be derived with eqs. (2) and (8), which is as follows:

$$P_{gy} = \frac{2}{T_s} \int_{0}^{T_s} i_m(t) dt = \frac{4C_rV_{batt}K_rV_{batt} - V_m}{nT_s(2V_{batt} - K_rV_{batt}/n)}$$

Where, $K_r = 1 - \cos \omega_0D_{T_s}$. Since the output power $P_{of}$ is expressed as

$$P_{of} = V_{batt}I_{batt}$$

battery voltage expression according to charging current can be derived as

$$V_{batt} = \frac{4(V_m/n)^2C_rK_r + (V_m/n)K_rI_{batt}}{2I_{batt}T_s + 4(V_m/n)C_rK_r}$$
by equating eq. (17) with eq. (18).

< Discharging operation>

Fig. 6 is the equivalent circuit expressed with full-bridge rectifier instead of voltage-doubler rectifier. The equivalent circuit has exactly same structure with that in charging operation so that the derivation of input-output relationship in discharging operation is same to the case of charging operation. Thus, the results is

$$V_{in} = 2 \times \frac{4(nV_{batt})^2(C_i/n^2)K_i + 2(nV_{batt})K_iT_i}{4I_{in}T_i + 4(nV_{batt})(C_i/n^2)K_i}$$

(20)

where, $K_i = 1 - \cos \omega DT_i$, and $I_{in}$ is the average input current. Since the discharging control is performed not by $I_{in}$ but by $I_{batt}$ and $V_{in}$ is the constant output voltage regulated by a harmonic regulator, eq. (20) can be rewritten as the following equation:

$$V_{batt} = \frac{V_{in}I_{batt}T_i + (C_i/n)K_iV_{in}^2}{2C_iV_{in}K_i + nI_{in}T_iK_i}$$

(21)

with the condition of $V_{batt}I_{batt} = V_{in}I_{in}$. Eq. (21) gives the information of the minimum battery voltage available at a discharge current command.

III. DESIGN AND CONTROL STRATEGY

The main object of OBC is to charge battery and it has a long operation time of 4–8 hours under full load condition. On the contrary, discharging operation may not occur frequently like charging operation. Thus, main focus of design is to optimize charging operation. To guarantee the charging operation explained in the mode analysis and the soft-switching conditions of power devices, the peak voltage of resonant capacitor should not exceed the battery voltage and the duration from mode 1 to mode 2 should be shorter than half of the switching period. The design procedure has been well explained in reference [26] and it is summarized in section A.

A. Design for charging operation

Since $V_{cfr}$ cannot be a negative value, the transformer turns-ratio $n$ should be smaller than $V_{o} / V_{batt, max}$. Also, the maximum operational duty-ratio $D_{f,max}$ should be chosen to secure the duration of mode 3. Using eqs. (7) and (8), and the selected values of $D_{f,max}$ and $n$, a resonant frequency $\omega_r$ can be chosen among the candidates satisfying normal operation conditions of $V_{cfr} < V_{batt}$ and $DT_i + T_{M2}/2 < T_i$. After then, the resonant capacitor $C_i$ is selected to admit the maximum battery voltage at $D_{f,max}$ and the maximum charging power using eq. (19). The resonant inductor $L_i$ is calculated with the selected $C_i$. The magnetizing inductance $L_m$ is designed to meet ZVS of upper switches and its guideline is as follows:

$$L_m = \frac{T_{dead}D_iT_i}{4C_i}$$

(22)

It is derived based on the condition that $\Delta T$ of the transition time of drain-to-source voltages in switch-bridge is shorter than $T_{dead}$. To design $L_m$, it is necessary to select $D_{f, min}$ that is the worst case to accomplish complete ZVS. The designed parameters are verified by checking whether the conditions of $V_{cfr} < V_{batt}$ and $DT_i + T_{M2}/2 < T_i/2$ are satisfied under entire battery voltage ranges at maximum charging powers.

B. Control strategy of discharging operation

In discharging operation, the normal operation conditions of $V_{cfr} < V_{o} / n$ and $DT_i + T_{M2}/2 < T_i$ should be met like charging operation. Using eq. (15), (16), and (21), the minimum discharge voltage satisfying the normal operation conditions can be investigated according to various discharge currents. The result of partial derivatives of eq. (21) according to $I_{batt}$ is

$$\frac{\partial V_{batt}}{\partial I_{batt}} = \frac{C_iT_iV_{in}^2(1 - \cos \omega DT_i)}{2C_iV_{in}K_i + nI_{in}T_iK_i}$$

(23)

and it is always positive value, which says that the minimum discharge voltage is increased as discharge current is increased. Therefore, the minimum discharge voltage that can output entire discharge current can be found from the maximum discharge current condition. Unfortunately, the minimum discharge voltage may not meet design specifications because the design parameters are optimized for charging operation. To overcome this problem, another control variable is required to increase the voltage gain for discharging operation. Rearranging eq. (21), it can be written as follows:

$$V_{batt} = \frac{V_{in}I_{batt}T_i + (C_i/n)\omega_iV_{in}^2}{2C_iV_{in} + nI_{in}T_i}$$

(24)

The solution of partial derivative of eq. (24) according to $T_i$ is

$$\frac{\partial V_{batt}}{\partial T_i} = \frac{1 - \cos \omega DT_i}{(1 - \cos \omega DT_i)^2} \left(1 - \cos \omega DT_i\right) \frac{V_{in}I_{batt}}{2C_iV_{in} + nI_{in}T_i}$$

(25)

As shown in this equation, the slope of battery voltage is influenced by $T_i$ but its sign should be kept constant in order that $T_i$ is used as a control variable. Investigating eq. (25), the slope has negative sign if the following condition is met

$$M = \frac{1 - \cos \omega DT_i}{(1 - \cos \omega DT_i)^2} \left(1 - \cos \omega DT_i\right) \frac{V_{in}I_{batt}}{2C_iV_{in} + nI_{in}T_i} \leq 0$$

To the extent that $\omega DT_i$ meets eq. (26), the voltage gain of eq. (20) can be additionally increased by increasing $T_i$, regardless of battery current. Therefore, a lower battery voltage can be used for discharging operation.

The minimum battery voltage $V_{batt, min}$ can be calculated using eqs. (15), (16), (21), and $I_{batt, max}$ at a nominal switching frequency. The target minimum battery voltage can be selected with a slightly higher voltage than the calculated minimum battery voltage and the switching frequency is changed to an appropriate value selected among the candidates that satisfy a target minimum battery voltage and eq. (26).

IV. EXPERIMENTAL RESULTS

The prototype charger has been designed with the specifications described in Table 1. The selected devices are listed in Table 2 together with parameters designed with the procedure described in section III. The capacitor $C_i$ is used to prevent the transformer saturation during charging operation and it is also used as charge-pump capacitor having half of the
TABLE 1

<table>
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<tr>
<td>Maximum charging/discharging power $P_{\text{max}}$</td>
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<tr>
<td>Maximum battery current $I_{\text{batt, max}}$</td>
<td>20A</td>
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<td>Battery voltage $V_{\text{batt}}$</td>
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<td>Battery voltage $V_{\text{batt}}$</td>
<td>400V</td>
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<td>Nominal switching frequency $f_s$</td>
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TABLE 2

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<tr>
<td>Resonant inductor $L_r$</td>
<td>35μH</td>
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<td>Transformer turns ratio $n$</td>
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<td>Charge pump capacitor $C_i$</td>
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<td>Output filter inductor $L_m$</td>
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<td>Transformer Core</td>
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Fig. 7 Plot of $V_{\text{crf}}$, $V_{\text{batt}}$, and $D_fT_s+T_{\text{chg}}$ according to duty-ratio variations at $I_{\text{batt}}=20A$

Fig. 8 Maximum charge voltage according duty-ratio and charging input voltage $V_{\text{in}}$. In addition, resonant frequency formed by $C_r$ and $L_r$ should not be affected by $C_i$. Considering these roles of $C_r$, it is better that $C_r$ has a large value, but too large values produce bulky design because high frequency capacitors such as film capacitor are utilized. In this design, a value over 20 times larger than that of resonant capacitor has been chosen for $C_r$, in order that the resonant frequency is determined only by $C_r$ and $L_r$. As shown in Figs. 7 and 8, this design satisfies the conditions

Fig. 9 Minimum discharge voltages according to duty-ratio and discharging current changes under satisfying normal operation conditions

Fig. 10 Index $M$ according to frequency changes at maximum battery current

Fig. 11 Minimum discharge voltages according to duty-ratio and switching frequency changes under satisfying normal operation conditions

Fig. 12 Minimum discharge voltages according to duty-ratio and switching frequency changes
of $V_{cr}<V_{batt}$ and $D_{r}T_s+T_{M2}<T_s/2$ under entire battery voltage ranges at the maximum charging power. Also, the design can deal with entire battery voltage range with normal switching frequency of 50kHz. Fig. 9 is the operational regions satisfying the conditions of $V_{cr}<V_{batt}$ and $D_{r}T_s+T_{M2}<T_s/2$, which are depicted according to discharging current changes. As predicted in eq. (24), it can be seen that the minimum battery discharge voltage becomes high as battery current is increased. The minimum battery voltage, that can operate under all load conditions while satisfying the conditions of $V_{cr}<V_{batt}$ and $D_{r}T_s+T_{M2}<T_s/2$, occurs at 273V. Using eq. (26), the index $M$ can be depicted in Fig. 10 according to frequency changes at the
maximum battery current. For example, $D_i$ less than 0.47 can be used at $f_s=40$kHz. Fig. 11 is the minimum discharge battery voltage satisfying the conditions $V_{ref} < V_{batt}$ and $D_i T_s + T_{M2} < T_s/2$. It is depicted using eqs. (15), (16), and (21) at the maximum battery current according to duty-ratio of discharging operation and switching frequency changes. Duty-ratio ranges according to switching frequency changes in Fig. 11 are included in negative $M$ in Fig. 10. Accordingly, the switching frequencies depicted in Fig. 11 can be used for another control variable. At the nominal switching frequency of 50kHz, the minimum battery voltage capable of flowing entire discharging current is about 273V and frequency changing voltage should be selected above this voltage. In this design, we have selected 290V considering a margin and 40kHz has been selected at which the lowest battery voltage is available. The resulting minimum discharge voltage can be depicted in Fig. 12. Fig. 13 is the implemented prototype bidirectional charger comprised of the proposed converter and single-phase inverter for test use. The controller shown in Fig. 1 was implemented with DSP TMS320F28335. Figs. 14 and 15 are switching waveforms measured according to power flow directions. From these figures and eqs. (2) and (10), it can be seen that the more the battery voltage is decreased, the higher the peak value of transformer current at switching instant occurs during charging operation. Accordingly, a high switching loss happens at a low battery voltage and the switching loss becomes worse at a high output power so that the efficiency degradation is likely to occur at a low battery voltage and a high output power. On the other hand, the peak value of transformer current happens at a high battery voltage during discharging operation so that high switching loss occurs at the high battery voltage. Also, the nominal switching frequency of 50kHz can accommodate the entire battery voltage range during charging operation, but switching frequency is changed to 40kHz at lower battery voltage during discharging operation. Fig. 16 shows the transient between charging and discharging operations. It shows that the transition times are not fast because the proposed method has the selective bidirectional operation dominated by external power flow commands. Fig. 17(a) is the measured efficiency according to battery voltages, which includes driving power consumption, controller power consumption, and output filter loss. In charging operation, 97.2% has been recorded at 330V/6.6kW and the maximum efficiency of 97.8% is measured at 415V/6.6kW. On the other hand, 96.2% has been recorded at 330V/6.6kW and the maximum efficiency of 96.8% is measured at 250V/5kW in discharging operation. Overall efficiency including single-phase inverter is as shown in Fig. 17(b). Table 3 shows the performance comparisons between the proposed charger and recently reported topologies. It shows that it is not inferior to recent works in the aspect of structure and efficiency despite that the proposed converter mainly uses constant frequency control.

V. CONCLUSIONS

An isolated/bidirectional PWM-RC has been suggested for bidirectional OBCs. Using the structure change method, the proposed charger can be operated under bidirectional power flow by overcoming gain characteristic which has always ‘buck type’ operation regardless of power flow directions. Analysis has been performed to derive the voltage gain and the design
equations, and an additional control technique to increase the converter gain has been proposed based on the analysis. To verify the performances, a 6.6kW prototype charger has been implemented with the design guidelines. Experimental results show that 97.2% has been recorded at 330V/6.6kW in charging operation and 96.2% has been recorded at the same condition in discharging operation. Therefore, it may be another candidate for DC/DC stage in V2G (H) EV chargers or backup power supplies such as emergency power system and household energy storage system.

REFERENCES


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