Isolated Single-Input Dual-Output LLC Converter for a Wide Range of Voltage Gain using Mode Transition

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In this study, an LLC converter that can achieve a wide range of voltage gain and high efficiency for two loads is proposed. The efficiency of the proposed LLC converter is found to be greater than that of the conventional full-bridge LLC converter. Depending on the ON/OFF state of the five primary switches, the proposed converter offers five modes of operation. These modes enable a squeezed switching frequency span that is close to the resonance frequency. In all operation modes, one load is controlled by the switching frequency, while the other is controlled by the switching frequency and phase-shift angle. The proposed control method provides a wide operating range and seamless transition between different modes. Furthermore, the mode transitions are performed based on the relationship between the switching frequency and phase-shift angle. First, the theoretical characteristics of the proposed converter are analyzed. Next, the gain characteristics of each operation mode and the mode transition method are described. The prototype is designed to convert a constant input voltage of 360 V to output voltages ranging from 80–420 V and 36–48 V. The experimental results confirmed that steady-state waveforms can be achieved in the multiple proposed modes of operation and voltage gains ranging from less than 0.5 to more than 1.0. Moreover, seamless mode transitions can be achieved without large deviations between the outputs. Finally, the efficiency of the proposed converter is evaluated in comparison to that of a conventional converter. The proposed converter achieved high-efficiency operations in the CC2 region of the battery compared to the conventional converter.

Keywords: LLC converter, single output dual output, wide range voltage gain, on-board charger

1. Introduction

In recent years, plug-in hybrid electric vehicles (PHEVs) have become more popular as a method to reduce fuel consumption and emissions. Generally, PHEVs are equipped with a high-voltage (HV) battery that provides propulsion, a low-voltage (LV) battery for the auxiliary equipment, and an onboard charger to charge them(1)–(3). The onboard charger uses an AC/DC converter in the front stage and an isolated DC/DC converter in the rear stage. In addition, Li-ion cells are used in the HV battery. Figure 1 shows the typical charging profile for a Li-ion cell(4). The Li-ion cells have three charge states, and the battery voltage varies considerably with the charge state. The over-discharging region (CC1) occurs when the battery is over-depleted. In order to recover from the over-discharging region, it is necessary to charge the battery with low voltage and low current. In the CC2 region, the battery is charged with a constant current and is charged until the battery voltage reaches the voltage in the CV region. In general, the voltage in the over-discharging region of the battery is 250 V or less, and the voltage in the CV region is 420 V(5). Therefore, the onboard charger must have a wide output range that is adapted to the battery voltage.

The output characteristics of the onboard charger significantly depend on the DC/DC converter in the rear stage. The LLC resonant converter is one of the most attractive topologies as it has an excellent soft switching performance and wide output range functions and is being researched as a DC/DC converter for battery charging(6–8). Generally, the LLC converter controls the output of the onboard charger by changing the switching frequency, \(f_s\), from the resonance frequency, \(f_r\). However, to adapt to a wide voltage fluctuation range, the \(f_s\) of the LLC converter must change over a wide range and deviate from \(f_r\). When \(f_s\) deviates from \(f_r\), the circulating current in the resonance tank increases. This increases the loss of the converter and reduces its efficiency. Therefore, in a conventional LLC converter, there

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is a tradeoff between the operating range and efficiency. To achieve a wide voltage regulation range and a narrow range simultaneously, several control strategies and modifications to the circuit structure have been studied (9)–(13). In (9), a converter that uses two transformers and five switches was proposed. By controlling the ON/OFF state of the five switches, the converter has four operation modes. The four operation modes provide different voltage gain characteristics, and the converter accommodates a wide output voltage range. However, \( f_s \) needs to swing over a wide range. In (10), a converter with two transformers and an additional active switch was proposed. The converter has two different resonance frequencies owing to the additional switch. As a result, the converter switching frequency always operates near the resonance frequency in the constant current (CC) and constant voltage (CV) modes of the battery. In (11), an LLC converter with three operation modes was proposed to improve the efficiency at low voltage and light load. The proposed converter has two transformers and operates as a full-bridge (FB), a two-phase half-bridge (HB), and a single-phase HB. The seamless transition between the three operation modes enables the converter to achieve a wide voltage range and high-efficiency operation. Thus, a circuit structure with two transformers is effective for expanding the output range of the converter.

On the other hand, onboard chargers are required to have charging functions for LV batteries in addition to HV batteries. The aforementioned method uses two transformers for one load, which reduces the power density. To solve this problem, circuit configurations that use two transformers for two loads have been proposed (14)–(16). In (14), a multiport LLC converter that uses phase-shift control and an asymmetric duty ratio was proposed. The output of the converter can be held constant even when the two input voltage sources are unbalanced. This is useful for renewable or clean energy sources with different output voltages. However, it is not suitable for applications whose output widely fluctuates. In (15), a multiport converter that uses phase-shift control and pulse frequency modulation (PFM) was proposed. Because the two loads are controlled through different control methods, there is no cross regulation between the loads. However, the output range of the converter ranges from 200 V to 400 V, which is not suitable for charging the over-discharge region of the battery. In (16), an LLC converter that uses a combination of HB and FB was proposed. The primary bridge has a conventional FB structure, and two transformers are integrated into a single bridge. This topology is more effective than other multiport converters in terms of efficiency because it reduces the number of switches.

The originality of this paper is the control strategy of the load with voltage fluctuation over a wide range and the load with low voltage level, and the management method of operation mode transition. The primary side bridge of the proposed circuit adopts the variable topology proposed in (9). However in (9), there is only one load, and power transmission to multiple loads is not considered. Furthermore, a clear control method for operation mode transition is not shown. In this study, the control method extended to two loads and the design guideline and control strategy for realizing the operation mode transition are shown. These are the decisive differences from (9). The proposed circuit has five operation modes that have different voltage gain characteristics. In addition, the switching frequency always operates at \( f_s \leq f_r \). The proposed converter provides output voltages in the ranging from 80 V to 420 V and 36 V to 48 V under constant input voltage conditions. One of the separated loads is assumed to be the Li-ion battery with wide voltage fluctuations, and the other low voltage level load is assumed to be the auxiliary battery. The five operation modes and the mode transitions allow the proposed converter to be designed with a higher magnetizing inductance or lower leakage inductance than conventional converters. In this paper, the circuit configuration and principle of the proposed converter are discussed, and its effectiveness is confirmed through examination in prototype. The experimental results reveal that the proposed circuit achieves higher efficiency over a wider operating range than conventional circuits.

2. Conventional Converter and Proposed Circuit

2.1 Circuit Configuration of the Conventional Converter

Figure 2 shows the configuration of a conventional converter for PHEVs. According to (17), the circuit in Fig. 2 is a standard onboard charger. The onboard charger has a grid-to-HV (G2H) battery charge function and an HV-to-LV (H2L) battery charge function. Furthermore, the LLC converters in the front and rear stages control the voltage/current of the HV and LV batteries, respectively, and the two batteries are controlled by \( f_r \). Both the front and rear bridges of the circuit in Fig. 2 operate with FB and HB as shown in Fig. 3 and control the two batteries. In FB mode, switches \( S_1, S_4 \) and switches \( S_2, S_3 \) operate complementarily. In HB mode, \( S_1 \) is always off and \( S_4 \) is always on. \( S_1 \) and \( S_4 \) operate complementarily. These are the same for the converter in the rear stage. The voltage gain obtained in HB mode is half the gain obtained in FB mode. Because the voltages/currents of the two battery fluctuate, \( f_r \) must also swing over a wide range. Therefore, the loss of the converter may increase.

2.2 Circuit Configuration of the Proposed LLC Converter

Figure 4 shows the proposed single-input,
1. In this mode, Eqs. (1) and (2), respectively, as follows:

\[
R_{e1} = \frac{8N^2}{\pi^2} R_{L1} \tag{1}
\]

In mode 1, the voltage applied to the magnetizing inductor changes depending on the ratio of the magnitudes of the two loads. Therefore, when the voltage gain obtained by \( f_s = f_r \) is determined, the derivation is performed using the relationship of the equivalent circuit in Fig. 6.

Two inductors are considered as one inductor, as shown in the equivalent circuit in Fig. 6. Therefore, \( i_{L1} \) and \( i_{L2} \) are equal. The change in \( \varphi_1 \) changes \( i_{L1} \) and at the same time \( i_{L2} \). Due to the change in the current \( i_{L2} \) changes the output voltage \( V_{o2} \), the change in \( \varphi_1 \) affects \( V_{o2} \). Similarly, the change in \( \varphi_2 \) affects not only \( V_{o2} \) but also \( V_{o1} \). This is because the primary sides of the two resonant tanks are connected in series and are equivalently regarded as one resonant tank. Therefore, in the proposed circuit, one of the two-phase shift angles is 0 in order to reduce the interference of the phase shift angles.

2.4 Operation mode 2 of the Proposed LLC Converter

Figure 5(b) shows the equivalent circuit of operation mode 2. \( S_{p2} \) and \( S_{p5} \) are driven complementarily with a certain dead time, \( S_{p1} \) and \( S_{p4} \) are constantly ON, and \( S_{p3} \) is constantly OFF. In this mode, the primary bridge operates as the HB, and the transformers on the primary side are connected in parallel to the primary bridge. Figure 7 shows the steady-state waveform at \( f_s = f_r \) in operation mode 2. As the parameters of the two transformers are different, resonant currents \( i_{L1} \) and \( i_{L2} \), flowing through leakage inductors \( L_{V1} \) and \( L_{V2} \) are different. When \( f_s = f_r \), output voltage \( V_{o1} \) is given by Eq. (3) as follows:

\[
V_{o1} = \frac{V_{in}}{2N_1} \tag{3}
\]

Similarly, the output voltage \( V_{o2} \) is obtained by Eq. (4).

\[
V_{o2} = \frac{V_{in}}{2N_2} \tag{4}
\]

Eqs. (3) and (4) are the same as the LLC converters that operate on conventional HB.

2.5 Operation mode 3 of the Proposed LLC Converter

Figure 5(c) shows the equivalent circuit of operation mode...
In the proposed circuit, one load is controlled as the FB. Moreover, connected in parallel, and the primary side bridge operates 5. In this mode, the primary sides of the transformers are because output voltage 4ever, because output voltage 3. Output 1-2 of the primary bridge is a two-level square wave of 1. On the other hand, due to the 360 IEEJ Journal IA, Vol.10, No.3, 2021

3. 1-2-Sp3 and Sp2-Sp5 are driven complementarily with a certain dead time, and Sp4 is constantly ON. In this mode, the bridge on the primary side operates as the combination of FB and HB. Figure 8 shows the steady-state waveform at 5 in mode 3. Output vab of the primary bridge is a two-level square wave of +Vin and −Vin. Moreover, vcb is a two-level square wave of 0 and −Vin. Therefore, vab operates on FB, on the other hand, vcb operates on HB. When 5 = fr, output voltage Vo1 can be obtained using Eq. (5). However, because output voltage Vo2 operates on the HB, it can obtained using Eq. (4).

\[ V_{o1} = \frac{V_{m}}{N_1} \] (5)

2.6 Operation mode 4 of the Proposed LLC Converter

Figure 5(d) shows the equivalent circuit of operation mode 4. In this mode, the bridge on the primary side operates as a combination of FB and HB, similar to mode 3. Sp2-Sp4 and Sp3-Sp5 are driven complementarily with a certain dead time, and Sp1 is constantly ON. Figure 9 shows the steady-state waveform at 5 = fr in mode 4. Output vab of the primary bridge is a two-level square wave of +Vin and 0. In addition, vcb is a two-level square wave of Vin and −Vin. When 5 = fr, output voltage Vo1 can be obtained using Eq. (3). However, because output voltage Vo2 operates on the FB, it can be obtained using Eq. (6).

\[ V_{o2} = \frac{V_{m}}{N_2} \] (6)

2.7 Operation mode 5 of the Proposed LLC Converter

Figure 5(e) shows the equivalent circuit of operation mode 5. In this mode, the primary sides of the transformers are connected in parallel, and the primary side bridge operates as the FB. Moreover, Sp1-Sp5 and Sp2-Sp4 are driven complementarily with a certain dead time, and Sp3 is constantly ON. Figure 10 shows the steady-state waveform at 5 = fr in mode 5. vab and vcb are both two-level square waves of Vin and −Vin. When 5 = fr, output voltages Vo1 and Vo2 can be obtained using Eqs. (5) and (6), respectively.

2.8 Operational Principle with Phase Shift Control Applied

In the proposed circuit, one load is controlled by the switching frequency and the other load is controlled by the phase shift in all operating modes. The phase shift angle \( \varphi_1 \) or \( \varphi_2 \) is inserted between the rise of the primary side switch and the rise of the load side switch. Figure 11 shows the steady-state waveform at \( \varphi_1 > 0 \) and \( \varphi_2 = 0 \) in mode 5. The phase shift angle \( \varphi_1 \) is inserted between the rising edge of the primary side switches Sp1 and Sp5 and the rising edge of the load side switch Sp2. On the other hand, due to the \( \varphi_2 \) is 0, the rise of Sp1 is synchronized with Sp1 and Sp5, and the rise of Sp2 is synchronized with Sp2 and Sp4. Similarly, in other operation modes, the phase shift angle is provided on one of the load-side bridges.

There are six switching stages in the half cycle of the proposed converter. Figure 12 shows the equivalent circuit for each switching stage.

Stage 1 \([t_0 \sim t_1][\text{Fig. 12(a)}]): Before \( t_0 \), the switches Sp1 and Sp5 on the primary side are ON, and at \( t = t_0 \), Sp1 and Sp5 are OFF. At \( t = t_1 \), switches Sp2 and Sp4 turn ON. During this period, the two transformers do not transmit power to their respective loads.

Stage 2 \([t_1 \sim t_2][\text{Fig. 12(b)}]): Figure 12(b) shows the equivalent circuit of \( t = t_1 \) to \( t_2 \) in Fig. 11. \( L_r \) and \( C_r \) resonate through input voltage \( V_m \), and resonant current \( i_{Lr} \) changes linearly. Moreover, transformer \( T_{r1} \) is short-circuited by \( S_{r1} \) and \( S_{r2} \), and power is stored in \( L_r \) rather than being supplied to load \( R_{L1} \). On the other hand, the resonance current \( i_{Lr} \) is a sinusoidal current because it is not affected by the phase shift angle \( \varphi_1 \).

Stage 3 \([t_2 \sim t_3][\text{Fig. 12(c)}]): Sp2 turns off at \( t = t_2 \), and \( S_{r1} \) turns on after the dead time period. During period \( t_2 \) and \( t_3 \), \( D_{r2} \) and \( S_{r1} \) are turned ON, and \( T_{r1} \) transmits power to the load. The energy stored in the leakage inductor on the
primary side during the period of stage 2 is transmitted to the load through $T_{r1}$. Therefore, the voltage of $V_{o1}$ is boosted.

Stage 4 [$t_3 \sim t_4$] [Fig. 12(d)]: At $t = t_3$, the polarity of the resonant current $i_{Lr2}$ is reversed. The resonant current $i_{Lr1}$ changes in a sinusoidal shape.

Stage 5 [$t_4 \sim t_5$] [Fig. 12(e)]: At $t = t_4$, the polarities of the magnetizing currents flowing in $L_{m1}$ and $L_{m2}$ are reversed. During this period, a resonance current flows on the primary side as in stage 3 and stage 4.

Stage 6 [$t_5 \sim t_6$] [Fig. 12(f)]: At $t = t_5$, the values of $i_{Lr1}$ and $i_{Lm1}$ become equal. Hence, the secondary current in $T_{r1}$ becomes 0. $T_{r2}$ transmits power to the load through $D_{s1}$ and $S_{p2}$ through resonance on the primary side. As a result, $D_{s2}$ is turned OFF, and the secondary current of transformer $T_{r1}$ becomes 0.

The remaining half cycle also operates symmetrically with the first half cycle. When $\varphi_1 > 0$ and $\varphi_2 = 0$, the resonance tank constituting $T_{r2}$ operates in the same operation as the conventional LLC converter. The phase shift angle is inserted into the bridge of the load that requires higher boost operation.

Figure 13 shows the voltage and current waveforms of the primary and secondary bridges in secondary side phase shift control. When the secondary side phase shift control is used, unlike the primary side phase shift control, the large circulating current does not occur in the primary side bridge. However, as shown in Fig. 13, the circulating current is occurred on the secondary side. The large phase shift angle provides the converter with high boost capability, however it causes a large circulating current.

### 2.9 Gain Characteristics of the Proposed LLC Converter

Figure 14 shows a schematic diagram of the gain characteristics of the PFM and phase-shift control in the proposed LLC converter. Figure 14(a) shows the gain...
characteristics of modes 1, 2, and 5 of \( R_{L1} \) in the PFM control. The output voltage gain, \( G_{o1} \), is normalized by the output voltage obtained by \( f_s = f_r \), and \( f_s = f_r \) in mode 5. Output voltage gains \( G_{o1} \) and \( G_{o2} \) are given by Eqs. (7) and (8), respectively, as follows:

\[
G_{o1} = \frac{N_1 V_{o1}}{V_{in}} \quad (7)
\]
\[
G_{o2} = \frac{N_2 V_{o2}}{V_{in}} \quad (8)
\]

Operation modes 2 and 5 exhibit the same gain characteristics as the LLC converters that operate with a conventional HB and FB, respectively. In mode 1, the primary side of the two transformers is connected in series; hence, a gain lower than the gains of modes 2 and 5 is achieved. The proposed circuit has a wide range of voltage gain owing to the use of two transformers connected in series; hence, a gain lower than the gains of modes 2 and 5 is achieved. The gain relationship shown in Fig. 15, \( G_{o1} \) and \( G_{o2} \) are controlled by switching frequency \( f_s \) and phase-shift angle \( \phi_2 \). After the transition, \( G_{o1} \) is controlled by switching frequency \( f_s \) and phase-shift angle \( \phi_2 \). This also applies to the operation modes in which the transformers are connected in series, \( \phi_1 \) and \( \phi_2 \) interfere with each other. In the proposed LLC converter, the two loads are controlled by the PFM and phase shift; one load is controlled by the frequency, and the other load is controlled by the phase shift.

3. Operation Mode Transition Sequence

3.1 Mode Transition Method

Figure 15 shows a schematic diagram of the gain characteristics before and after the mode transition. Gain \( G_{o1} \) of load \( R_{L1} \) changes, and gain \( G_{o2} \) of load \( R_{L2} \) is constant. Before the transition, the converter operates in mode 2, and after the transition, it operates in mode 3. In the proposed converter, one load is controlled by using the switching frequency, and the other load is controlled by the switching frequency and the phase-shift angle.

3.2 Phase Shift Angle Compensation Method

To achieve a seamless mode transition, a transition that compensates for the phase-shift angle is proposed. When \( f_s \) and \( f_r \) are equal, the relationship between the gain and the phase-shift angle can be expressed, as shown in Eq. (9)(18)(19), as follows:

\[
G_{ox} = \frac{\sqrt{\pi Q_x} + \sqrt{-2(\cos \varphi_x)^2 + \pi Q_x + 2}}{\sqrt{Q_x} (1 + \cos \varphi_x)} \quad (9)
\]

where \( x \) is 1 or 2, and \( f_r \) and \( Q_x \) are defined, respectively, as follows:

\[
f_r = \frac{1}{2 \pi \sqrt{L_r C_r}} \quad (10)
\]
\[
Q_x = \frac{\sqrt{L_x / C_r}}{N_x R_{Lx}} \quad (11)
\]

Immediately after the transition, the switching frequency transitions to the resonance frequency regardless of the load conditions. The phase-shift angle is then compensated by Eq. (9) according to the \( Q \) value of the load so that the gain becomes equal before and after the transition. By compensating for the frequency and the phase-shift angle, the proposed converter achieves dynamic transitions between modes.

4. Design Guidelines and Control Methods

4.1 HV Battery Voltage/Current Characteristics

As shown in Fig. 1, the Li-ion cell assumed as the HV
battery changes its voltage/current characteristics according to the charge state of the battery. The CC1 region is the over-discharged region of the battery and is charged at a CC rate until the battery voltage returns to the nominal voltage. CC2 is a CC charge state of the battery, which is charged with a CC until the battery voltage to the CV mode. The current rate of CC2 is higher than that of CC1. The CV region is the CV state of the battery and is charged at a CV. Table 1 shows the voltage/current values from points A to E in Fig. 1. The battery voltage fluctuates from 80 V to 420 V, and the nominal state of the battery is point C.

4.2 Design Guidelines Generally, the LLC converter has high efficiency when the switching frequency operates near the resonance frequency. To achieve high efficiency near the nominal voltage, $V_n$, of the battery, turn ratio $N_1$ of $T_{r1}$ is given by Eq. (12) as follows:

$$ N_1 = \frac{V_m}{V_n} \tag{12} $$

Turn ratio $N_2$ of $T_{r2}$ is determined using the voltage values of the assumed LV battery and the operating mode to be used, and it is given by

$$ N_2 = \frac{V_m}{2 \times V_{02,\text{min}}} \tag{13} $$

In this study, modes 1, 2, and 3 are mainly used positively because the voltage fluctuation of the LV battery is small. When the voltage fluctuation of the LV battery is large, operation modes 4 and 5 are used. The parameters of the two resonant tanks are designed using the FHA method. The gain characteristics of the LLC converter is given by Eq. (14) using the FHA method.

$$ G_{ox} = \frac{1}{\sqrt{1 + \frac{1}{k_x} \left(1 - \frac{1}{f_r} \right)^2 + Q_x^2 \left(\frac{f_n - 1}{f_n} \right)^2}} \tag{14} $$

where, each parameter can be expressed as follows.

$$ f_n = \frac{f_x}{f_r} \tag{15} $$

$$ k_x = \frac{L_{mx}}{L_{r}} \tag{16} $$

As the fluctuation of $V_{02}$ is smaller than that of $V_{01}$, it is possible to use a large magnetizing inductor for $L_{m2}$. The parameters of each resonant tank must be designed to achieve mode transitions. To perform a mode transition, the gain obtained at the minimum switching frequency must be at least twice the minimum gain for that mode.

Additionally, to determine parameters $L_{m1}$ and $L_{m2}$, it is necessary to find the upper limits of $L_{m1}$ and $L_{m2}$ such that ZVS is maintained within the given dead time, $t_d$. In order to realize ZVS of the primary MOSFET, it is necessary to charge and discharge the output capacitance of the MOSFET during the dead time. In conventional LLC converter, the dead time required to achieve ZVS is given below.

$$ t_d > \frac{8 \times L_{mx} \times C_{oss} \times V_{m} \times f_s}{N_1 \times V_n} \tag{17} $$

When the primary bridge operates as HB, its output voltage is obtained at $f_s = f_r$ by Eqs. (3) and (4). Therefore, by substituting these into Eq. (17), the magnetizing inductor for satisfying the ZVS condition is expressed by the following equation.

$$ L_{mx} < \frac{t_d}{8 \times C_{oss} \times f_s} \tag{18} $$

When the primary bridge operates as FB, the maximum value of the magnetizing inductor is expressed by the following from Eqs. (5) and (6).

$$ L_{mx} < \frac{t_d}{16 \times C_{oss} \times f_s} \tag{19} $$

In mode 2, the two bridges operate as HB, therefore the constraint condition for the two magnetizing inductors is Eq. (18). In mode 5, the two bridges operate as FB, therefore the constraint condition for the two magnetizing inductors is Eq. (19). Eq. (19) is the strictest constraint on the magnetized inductor. Therefore, the magnetizing inductor of the proposed circuit is determined based on the constraint condition of Eq. (19).

4.3 Control Method Figure 16 shows the proposed control block diagram. The two loads are controlled using a PI controller. The smaller value of the PI controller outputs, $v_1$ or $v_2$, is used to modulate the switching frequency, and the difference between $v_1$ and $v_2$ is used to modulate the phase-shift angle. The minimum phase-shift angle is 0°. The operation mode is determined using the values of output voltages $V_{01}$ and $V_{02}$, and Eqs. (3)–(6) can be used to calculate the threshold values. For example, when $V_{01}$ and $V_{02}$ are less than or equal to the values calculated using Eqs. (3) and (4), the proposed LLC operates in mode 1. Furthermore, mode transitions are performed using flag and feed-forward compensation. The flag is detected at the time of operation mode transition, and compensation is performed through the feed-forward process when the flag rises. Flags 1 and 2 determine the phase-shift angles, $\varphi_1$ and $\varphi_2$, to be compensated. When $\varphi_1$ is 0 immediately before the transition, $\varphi_2$ is compensated, and when $\varphi_2$ is 0, $\varphi_1$ is compensated.

These operating modes and feedforward compensation are performed by “Selection” in the control block diagram. “Selection” determines the operation mode based on the relationship between the voltage thresholds shown in Table 2. Feedforward compensation is performed based on the chart shown in Fig. 17 mode $x_{(k)}$ is the detected operation mode, and mode $x_{(k-1)}$ is the previously detected operation mode. $x$ is 1 to 5.

5. Experimental Results

5.1 Experimental Conditions Table 3 shows the experimental parameters of the proposed LLC converter. The values of leakage inductor $L_r$ and resonant capacitor $C_r$, which comprise the resonance tank, are the same, whereas
the values of magnetizing inductors \( L_{m1} \) and \( L_{m2} \) are different. The leakage inductor \( L_r \) is adjusted by using an external inductor so that the two leakage inductors have the same value. Each parameter is designed according to the design guidelines given in Section 4. Resonance frequency \( f_r \) of the resonance tank is 100 kHz. As the fluctuation of \( V_{o1} \) is smaller than that of \( V_{o2} \), the transition from mode 1 to mode 2 occurs at \( V_{o1} = 125 \text{ V} \) according to the parameters in Table 2 and Eq. (3).

Similarly, the transition from mode 2 to mode 3 occurs at \( V_{o1} = 250 \text{ V} \), \( I_{o1} = 1.0 \text{ A} \) or \( V_{o1} = 250 \text{ V} \), \( I_{o2} = 2.38 \text{ A} \) according to the parameters in Table 2 and Eq. (5). The gain of the LLC converter is low at high loads. Therefore, each parameter of the proposed circuit is designed to satisfy the following relationship under the load conditions of \( V_{o1} = 250 \text{ V} \) and \( I_{o1} = 2.38 \text{ A} \).

\[
V_{o1,\text{mode2 maxi}} \geq V_{o1,\text{mode3 mini}} \tag{20}
\]

5.2 Steady State Waveforms Figure 18 shows the steady-state waveforms in the CC1 and CC2 regions of the HV battery. Figure 18(a) shows the steady-state waveforms of mode 1, in which \( V_{o1} \) and \( V_{o2} \) are 80 V and 36 V, respectively. In mode 1, the two transformers operate equivalently as a single transformer; hence, the difference between the bridge outputs is a square wave with two values, 0 and \( V_{in} \). \( V_{o1} \) is controlled by the switching frequency and the phase-shift angle, whereas \( V_{o2} \) is controlled by the switching frequency. Therefore, \( \varphi_2 \) is 0. Figure 18(b) shows the steady-state waveform of mode 2. Because the primary bridge operates as HB, \( v_{ab} \) is a binary square wave with \( V_{in} \) and 0, and \( v_{cb} \) is the same with 0 and \( -V_{in} \). As \( V_{o1} \) is controlled by the switching frequency and \( V_{o2} \) is controlled by the switching frequency and the phase-shift angle, \( \varphi_1 \) is 0, and \( V_{o1} \) and \( V_{o2} \) are 200 V and 48 V, respectively. Figure 18(c) shows the steady-state waveform of mode 3. The primary bridge operates as a combination of FB and HB. \( v_{ab} \) is a two-level square wave of \( V_{in} \) and 0, and \( v_{cb} \) is a binary square wave of 0 and \( -V_{in} \). Moreover, \( V_{o1} \) and \( V_{o2} \) are 250 V and 36 V, respectively. These voltages are obtained using Eqs. (5) and (4), respectively, and the two loads are controlled by the switching frequency.

Figure 19 shows the switching waveforms of mode 2 and mode 3. In mode 2, the primary bridge operates as HB. In
mode 3, one of the two bridges operates as FB and the other operates as HB. In Fig. 19(b), the phase shift angle \( \phi_2 \) is provided to control the two loads. From the measured \( V_{GS} \) and \( V_{DS} \) waveforms, it can be seen that ZVS of the primary MOSFETs are achieved.

Figure 20 shows the steady-state waveforms of the proposed operation modes 4 and 5. In mode 4, the primary bridge operates as a combination of FB and HB, similar to mode 3. \( v_{ab} \) operates on HB, and \( v_{cb} \) operates on FB. In mode 5, the primary bridge operates as FB. Therefore, \( v_{ab} \) and \( v_{cb} \) are binary square waves of \( V_{in} \) and \( V_{in} \). In these operation modes, as in the other operation modes, one load is controlled by the switching frequency, and the other load is controlled by the switching frequency and the phase-shift angle.

### 5.3 Gain Characteristics and mode Transition Control

Figure 21 shows the measurement results of the gain characteristics during PFM and phase-shift control in the proposed converter. Figure 21(a) shows the gain characteristics of each operation mode in the PFM control. In modes 2 and 3, \( Q_1 = 0.35 \) (\( R_1 = 250 \text{ V}/2.38 \text{ A} \)), and in mode 1 the measurement is performed using a load with a \( Q \) value lower than that of modes 2 and 3. Operation mode 2 is the gain characteristic of the HB operation, whereas mode 3 is the gain characteristic of the FB operation. Operation mode 1 provides a lower voltage gain than modes 2 and 3. Figure 21(b) shows the gain characteristics of mode 2 in the phase-shift control. \( V_{oc} \) is controlled to be constant at 36 V, and \( V_{gs} \) is controlled based on the voltage/current characteristics of the HV battery. There is \( Q_1 = 0.15 \), and the measurement is performed using a load of \( Q_2 = 0.25 \). Moreover, \( V_{oc} \) is controlled by the switching frequency, and \( V_{gs} \) is controlled by the switching frequency and the phase shift. Furthermore, \( V_{oc} \) is boosted by increasing phase-shift angle \( \phi_1 \), and \( V_{gs} \) is controlled to be constant. In mode 2, the operation mode transitions to
mode 3 when $V_{o1}$ reaches 250 V. It is confirmed that these measured gain characteristics are in agreement with the simulation results.

Figure 22 shows the transition waveform from mode 2 to mode 3. In Fig. 22(a), $V_{o1}$ is controlled to be 250 V, and $V_{o2}$ is controlled to be 36 V before and after the transition. Before the transition, $V_{o1}$ is controlled by the switching frequency and phase shift angle, and $V_{o2}$ is controlled by the switching frequency. After the transition, $V_{o1}$ is controlled by the switching frequency, and $V_{o2}$ is controlled by the switching frequency and the phase shift angle. Therefore, phase-shift angle $\varphi_2$ is compensated by Eq. (9) so that the gain of $V_{o2}$ becomes equal before and after the transition. The switching frequency after the transition is equal to the resonance frequency, and phase-shift angle $\varphi_1$ is 0. Thus, the $Q$ value of the load at the mode transition point can be known from the charging characteristics of the battery. Figure 22(b) shows the transition waveform when $V_{o1}$ is controlled to be 250 V and $V_{o2}$ is controlled to be 48 V. Before the transition, $V_{o1}$ is controlled by the switching frequency and phase-shift angle, and $V_{o2}$ is controlled by the switching frequency. After the transition, $V_{o1}$ is controlled by the switching frequency, and $V_{o2}$ is controlled by the switching frequency and phase-shift angle. Therefore, phase-shift angle $\varphi_2$ is compensated by Eq. (9) so that the gains are equal before and after the transition. It is confirmed that there is no magnitude deviation between the outputs before and after the transition, and that a seamless mode transition is achieved. However, because Eq. (9) is an approximate equation that expresses the relationship between the gain of the converter and the phase-shift angle, it is not possible to completely suppress the deviation immediately after the transition.

### 5.4 Efficiency Characteristics

To evaluate the performance of the converter, the losses of the proposed LLC converter and the conventional LLC converter shown in Fig. 2 are compared. The input voltage and voltage/current characteristics of each load are the same in the conventional converter and the proposed converter. However, the parameters of the resonance tank of the conventional converter are different from those of the proposed converter. The turn ratios, $n_1$ and $n_2$ of the conventional converter are given by Eqs. (21) and (22), respectively.

\[
\begin{align*}
n_1 &= \frac{V_{in}}{2 \times V_{o1\text{ min}}} = \frac{360}{2 \times 80} = 2.25 \quad \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdOTS
\end{align*}
\]

\[
\begin{align*}
n_2 &= \frac{V_{o2\text{ max}}}{2 \times V_{o2\text{ min}}} = \frac{420}{2 \times 36} = 5.83 \quad \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdOTS
\end{align*}
\]

The conventional converter operates in two modes: FB and HB. Therefore, the turn ratio of the transformer is designed such that the minimum voltage of each load is achieved at $f_s = f_r$. The minimum output voltage is obtained in the HB mode shown in Fig. 3(b). The transformers of conventional converters are designed to minimize loss with respect to the magnetic flux density, $\Delta B$. Thus, the copper and iron losses of the transformer are given by Eqs. (23) and (24), respectively, and $\Delta B$ is determined such that the derivative of the sum of these equations is close to zero.

\[
P_{\text{copper}} = \frac{\rho_{\text{c}}}{A_w} \frac{f_s}{f_{\text{rms}}} + \frac{\rho_{\text{c}}}{A_w} \frac{f_s}{f_{\text{rms}}} \quad \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdOTS
\]

\[
P_{\text{iron}} = k \cdot f_s \cdot \Delta B \cdot V_c \quad \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdOTS
\]
Single-Input Dual-Output Converter (Yuki Kinoshita et al.)

where $A_w$ is the cross-section of the winding; $\rho$ is the resistivity; $l$ is the winding length; subscripts $p$ and $s$ are the primary and secondary, respectively; $k$, $\alpha$, and $\beta$ are the Steinmetz parameters; and $V_c$ is the volume of the transformer. The magnetizing inductors, $L_m1$ and $L_m2$, of the conventional converter are $220\mu H$ and $280\mu H$, respectively. The values of $L_r$ and $C_r$ in the proposed and conventional LLC converters are equal. In addition, the switching frequency ranged from $60\ kHz$ to $100\ kHz$.

Figure 23 shows the efficiency characteristics of the proposed and the conventional converters, and Fig. 23(a) shows the efficiency characteristics of the CC1 region of the battery. In the conventional converter, the front converter operates as HB. The proposed circuit operates in mode 1 and mode 2. In this region, the efficiency of the conventional converter exceeds that of the proposed converter in regions where the battery voltage is low. This is because the switching frequency of the conventional converter operates near the resonance frequency. However, in regions where the load is heavy, the efficiency of the proposed converter exceeds that of the conventional converter. This is due to the difference in the parameters of the resonance tanks that constitute both converters. Because the proposed circuit uses at least three operation modes over a wide operating range, a large magnetizing inductor can be used, thereby suppressing the circulating current of the transformer. Figures 23(b) and (c) show the efficiency characteristics of the CC2 region of the battery. In the operating region of (b), in the conventional converter, the converter in the front stage operates as FB and the converter in the rear stage operates as HB. The proposed circuit operates in mode 3. In (c), in the conventional circuit, the converter in the front stage operates as FB, and the converter in the rear stage switches from HB to FB at the load of around 800 W. In (b) and (c) regions, the efficiency of the proposed converter exceeds that of the conventional converter in all regions. This is because, as in the CC1 region, when the load in the conventional converter becomes heavy, the switching frequency significantly deviates from the resonance frequency such that the circulating current of the circuit increases. Therefore, the proposed converter has high efficiency in the region that is above the nominal voltage of 250 V of the battery.

5.5 Loss Analysis of the Proposed Converter

Figure 24 shows the loss analysis results in Fig. 23(b). The error between the loss analysis result and the loss measured in the experiment is within 10%, confirming the validity of the loss analysis. The main losses are copper loss of magnetic parts and switching loss of the load side HV port. The circulating loss on the secondary side is caused by the phase shift angle $\Phi_1$ with the bridge on the primary side. Therefore, circulating loss increases as the output power increases. This is mentioned as a problem of the converter using the phase shift control. Nevertheless, as shown in Figs. 23(b) and 23(c), the efficiency of the proposed circuit is higher than that of the conventional converter configuration under heavy load conditions.

Figure 25 shows the current characteristics at $V_o1=250\ V$ to $420\ V$ and $V_o2=36\ V$ and the loss analysis of the conventional circuit. In the proposed circuit, $i_{Lr1}$ increases as the output power increases. The loss of the converter in the front stage of the conventional circuit is caused by $i_{Lr1}$. According to Fig. 25(b), the loss of the conventional circuit is dominated by the loss of the converter in the front stage. In particular, due to the effective value of the primary current is
large, the conduction loss and copper loss of the switch are large. On the other hand, the $i_{L,t1}$ of the proposed circuit is significantly reduced compared to the $i_{L,t1}$ of the conventional circuit. Therefore, it can be concluded that the proposed circuit reduces the loss caused by the current in the front stage converter of the conventional circuit.

5.6 Comparison of Similar Counterparts Solutions

Table 4 shows a comparative study of the proposed circuit and the counterparts solutions. In Ref. (22), proposed an improved LLC converter to increase the gain of conventional converter. The wide voltage gain of 0.2 to 1.2 is obtained only by simple frequency modulation. Furthermore, the peak efficiency exceeds 95%, confirming that it is one of the effective methods for expanding the voltage range. However, there is only one load. References (15) and (16) are LLC converters with two loads. The output voltage range of Ref. (15) is 200 V to 400 V, and the output voltage range of Ref. (16) is 250 V to 420 V. The proposed circuit achieves a wide voltage range for two loads. The output voltage range of the proposed circuit is 80 V to 420 V. Therefore, the proposed circuit is superior in terms of voltage range for systems with two loads compared to these references.

6. Conclusions

In this paper, a single-input, dual-output LLC resonant converter is proposed for a wide range of voltage gain applications, such as in onboard chargers. Depending on the ON/OFF state of the five switches, the converter offers five different modes of operation. One of the two loads is controlled by the switching frequency, and the other load is controlled by the switching frequency and the phase-shift angle. Therefore, there is no significant control interference between the two loads. A control method that seamlessly transitions between different operation modes is proposed to achieve a wide voltage gain. Mode transition control is executed based on the relationship between the phase-shift angle and the switching frequency. To validate the effectiveness of this topology through experiments and a theoretical analysis, a laboratory prototype is designed with an input of 360 V and outputs ranging from 80 V to 420 V and from 36 V to 48 V. The experimental results confirmed that the steady-state waveforms of the five proposed modes of operation have results that are similar to those of theoretical studies. Furthermore, the proposed mode transition control method achieved a seamless transition between different modes. The efficiency characteristics of the proposed LLC converter are evaluated through comparison with those of conventional LLC converters, and the proposed converter demonstrated good efficiency characteristics for HV battery with wide voltage fluctuations and LV battery. It is confirmed that the efficiency of the proposed converter exceeded that of the conventional converter in the entire region of CC2, which is the main region of the onboard charger. Therefore, the proposed LLC converter is suitable for applications with wide output voltage ranges, including in onboard chargers for the batteries of electric vehicles.

References

Single-Input Dual-Output Converter (Yuki Kinoshita et al.)


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