A Natural Bidirectional Input-Series-Output-Parallel LLC-DCX Converter with Automatic Power Sharing and Power Limitation Capability for Li-Ion Battery Formation and Grading System

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Abstract—For the high-power Li-Ion battery (LIB) Formation and Grading System using multi-stage conversion structure, bidirectional isolated DC-DC converter with high input voltage and large output current is needed to achieve isolation and relatively fixed voltage conversion gain. This paper proposes a modulation strategy and modifies the topology to achieve naturally bidirectional power transfer, automatic power sharing and power limitation for an Input-series-output-parallel (ISOP) LLC-DCX converter. Firstly, an active Pulse Width Modulation (PWM) strategy is proposed to avoid synchronous rectifier (SR) sensing and to achieve the nature bidirectional power flow for each sub-module. The resonant current always keeps discontinuous, and the conversion gain can keep constant if ignoring the conduction resistance. Secondly, clamping diodes are added in the conventional LLC converter to limit the inrush current during start-up, and to achieve constant power limitation under over load condition. Thirdly, with the proposed modulation, the power sharing of two-module ISOP LLC converter is analyzed and compared with the conventional ISOP LLC converter. Finally, experiments based on an 8kW (output:15V/533A) prototype composed of two-module ISOP LLC converter are conducted to verify the effectiveness of the proposed solution.

Index Terms—bidirectional isolated DC-DC converter, nature bidirectional power flow, constant power limitation, constant voltage gain

I. INTRODUCTION

DURING recent years, Li-Ion battery (LIB) is growing in popularity for battery electric vehicles, portable electronics, aerospace applications, etc.[1]. The last process for LIB production is to grade battery capacity and performance [2] by a LIB Formation and Grading System with bidirectional power capability[3][4]. In view of topology structure, the system generally consists of three conversion stages including the AC-DC rectifier, isolated bidirectional DC-DC converter with relatively fixed conversion gain (DCX) [5][6] and a bidirectional synchronous buck converter to control charging/discharging power as illustrate in Fig.1. The AC rectifier voltage is firstly step-down to a relatively fixed voltage (around 15V), and then followed with a synchronous bidirectional buck converter. This two-stages structure is used to achieve high conversion efficiency and flexible output power control under wide conversion gain range, as also studied in [7]-[10].

For isolated bidirectional DC-DC conversion stage shown in Fig.1, LLC converter is a suitable topology because of its soft-switching characteristics for all power devices from zero to full load range [11][12]. To achieve high conversion efficiency, Synchronous rectifier (SR) is generally employed for based on detecting the current zero crossing point or sensing the device voltage drop [13][14]. These SR technologies are widely used for LLC converter in unidirectional power applications. While, if it is used in bidirectional power applications, due to the asymmetry of both the topology and gain characteristics, SR must exchange between the two side bridges according to the power direction. As a result, detecting the power direction and shifting control logic will both make the control strategy more complex [15].

In bidirectional power applications, two issues need to be studied for LLC converter. One is the asymmetry issue for both control strategy and topology. The other one is the soft-start and current protection. A symmetrical bidirectional CLLC resonant converter with two resonant tanks was proposed in [16]-[19]. The voltage gain in forward operation mode is the same as that...
in backward mode. However, it needs to change the control logic to achieve bidirectional power flow. In [20], researchers proposed an LLC-LC type bidirectional control strategy for an LLC-LC type resonant converter. It can achieve bidirectional power flow automatically without power flow detection. However, the circulating current of the resonant tank is increased which decreases the efficiency of LLC converter in backward mode. In another point of view, adding an auxiliary inductor can also construct a symmetrical structure which was proposed in [21]. It can automatically change the power flow direction by regulating the switching frequency. However, it is a challenge to transfer power smoothly between different operating modes. In [22], a bidirectional CLTC resonant converter was proposed which adds an auxiliary transformer and an extra resonant capacitor. For these resonant converters, the efficiency and reliability may be reduced as auxiliary devices are added.

To achieve soft start and current protection, some methods were proposed in [23]-[25]. An over-current protection circuit was proposed that it used the output voltage of the LLC converter to clamp the voltage of resonant capacitor through an extra transformer in [23]. An offline calculation method to limit start-up current was proposed in [24]. However, when the parameters of resonant tank are changed because of aging or manufactory mismatching, the theoretical analysis may not be that accurate. In [25], optimal trajectory control (OTC) was used to solve the soft start-up and the short-circuit protection, but this method needs many calculations. The clamping diodes are added to the secondary side switching network to deal with the current protecting issue in [26]. When overload condition is triggered, the resonant converter will operate in DAB-MIXED mode, and an extra controller is needed to employ to decrease the duty cycle and to limit the peak value of the resonant current.

In addition, for high power (>6.6kW) LIB Formation and Grading System using LLC converter as the isolation stage, the input voltage is relatively high (650V-750V DC bus voltage with 380V three phase AC input), and the output current is large (above 440A, 15V output). To decrease the stress of power devices, modular structure can be adopted. In ideal case, when all the parameters are the same, each sub-module can work at the resonant frequency to achieve stable DCX operation. While, in non-ideal case, to achieve accurate input voltage sharing (IVS) and output current sharing (OCS), frequency modulation control [27]-[29] for the Input-Series-Output-Parallel (ISOP) LLC resonant converter will be needed. These frequency modulation strategies are variable frequency control, and it will cause the operating point to go away from the optimum resonating point. In addition, because the gain changes nonlinearly with the frequency, it is hard to analyze the control mechanism during the transient.

In order to achieve natural bidirectional power flow for LLC-DCX converter used in LIB Formation and Grading System, an ISOP LLC converter topology with automatic power sharing and power limitation capability is studied, which has the following advantages:

1. An active Pulse Width Modulation (PWM) modulation strategy is proposed to avoid SR sensing and to achieve the nature bidirectional power flow. The resonant current always keeps discontinuous, which could achieve constant output voltage gain regardless of the load.
2. Clamping diodes are added in the LLC converter to limit inrush current when the converter starts up, and to achieve constant power limitation during over load operation, which can improve the system reliability.
3. To decrease the power rating of single power switch, ISOP configuration is employed and multi-transformer structure is adopted for each sub-module. Because the proposed modulation strategy can achieve constant voltage gain of each sub-module, the power sharing control can be automatically achieved.

In this paper, a modulation strategy is proposed to achieve natural bidirectional power flow for ISOP LLC converters with automatic power sharing and power limitation capability. The operation principle analysis of the multi-transformer LLC converter under the proposed modulation is given in Section II. In Section III, the analysis about zero-voltage-switching (ZVS) condition, the constant voltage gain and the power limitation of the modified LLC converter under the proposed modulation are presented. With the proposed modulation strategy, the automatic power sharing for ISOP structure is analyzed in Section IV. The design considerations are given in Section V. The experimental results are shown in Section VI to verify the effectiveness of the proposed solution.
II. THE MODIFIED LLC RESONANT CONVERTER WITH ISOP STRUCTURE

The structure of ISOP LLC resonant converter is illustrated in Fig.2(a). There are two sub-modules, and each module is a multi-transformer LLC converter with two clamping diodes which is shown in Fig.2(b). \( V_{in} \) is the total input voltage for the ISOP system, and \( V_o \) is the output voltage.

As Fig.2(b) shows, two split-capacitors \( C_1, C_2 \), two clamping diodes \( D_1, D_2 \) and the resonant inductor \( L_r \) build up the resonant tank. \( L_m \) is the magnetizing inductor of the isolated transformer with turns ratio \( n \).

The frequency of the resonant tank \( f_r \) is shown as

\[
f_r = \frac{1}{2\pi\sqrt{L_r (C_1 + C_2)}} \tag{1}
\]

To achieve symmetrical operation of the multi-transformer rectifier in the secondary side, as shown in Fig.2(b), \( S_{ba} \) is turned on/off synchronously with \( S_{db} \) which is the same for \( S_{ca} \) and \( S_{eb} \). \( S_{ab} \) is turned on/off synchronously with \( S_{db} \) which is the same for \( S_{cb} \) and \( S_{eb} \). In practice, the parameters of each transformer are designed to be the same. Hence, when analyzing the working modes of the converter under proposed PWM modulation strategy, the topology shown in Fig.2(b) can be equivalently represented by the topology shown in Fig.3.

![Figure 3. The equivalent half-bridge LLC resonant converter.](image)

Unlike the conventional LLC converter that operating at a fixed frequency to get a fixed voltage gain, the modified modulation strategy can achieve fixed voltage gain as long as the current is discontinuous under \( f_s > f_r \) where \( f_s \) is the switching frequency. The reason of this phenomenon will be further explained in Section III. In the proposed modulation, the driving signals of \( S_1, S_3 \) and \( S_6 \) are the same which is different with that in conventional LLC converter. Ideally, the driving signals of \( S_2, S_4 \) and \( S_5 \) are the same. And the turn-on time of the driving signals is controlled to be half of the resonant period. In addition, there is no SR issue because all PWM signals are generated actively.

There are four working modes under the proposed modulation. When LIB Formation and Grading Equipment works normally, it operates in Mode 1 and Mode 3 which charges/discharges LIB respectively. When overload occurs, Mode 1 will change to Mode 2 and Mode 3 will change to Mode 4, which can automatically achieve power limitation.

A. Mode 1: Forward Power Transmission Mode

During the forward power transmission mode, the energy is transferred from the primary side to the secondary side of the transformer. Fig.4 shows the working stages for this mode and the key waveforms of the converter under this mode are shown in Fig.5.

![Figure 4. The operation of the converter in forward power transmission mode.](image)

- (a)Stage 1(\( t_1-t_3 \)). (b) Stage 2(\( t_3-t_1 \)).
- (c)Stage 3(\( t_3-t_2 \)). (d)Stage 4(\( t_2-t_3 \)).

![Figure 5. The key waveforms of the proposed converter in forward power transmission mode.](image)

Stage 1 (\( t_1-t_3 \)) [Fig.4(a)]: Before \( t_1, S_1 \) and \( S_6 \) are turned off. And the voltages of the junction capacitors of \( S_1 \) and \( S_6 \) have been decreased from \( V_o \) to zero. The parasitic anti-parallel diodes of \( S_3 \) and \( S_5 \) then conduct, which provides ZVS condition for \( S_3 \) and \( S_5 \). During this mode, \( S_2, S_4 \) and \( S_6 \) are turned on, and resonance occurs among \( C_1, C_2 \) and \( L_r \). The resonant current \( i_r \) increases sinusoidally in positive direction from zero which achieves zero-current-switching (ZCS) for \( S_1 \). The voltage of \( L_m \) is clamped to \( nV_o \), and the magnetizing...
current $i_{Lm}$ decreases linearly in negative direction until it is larger than zero. After that, the magnetizing current $i_{Lm}$ increases linearly in positive direction. The secondary current of transformer $i_s$ resonates with $L_m$ and the voltage across $L_m$ is clamped to $nV_o$. The current of magnetizing inductor decreases in positive direction, the anti-parallel diodes of $S_4$ and $S_5$ are conducted. Afterwards, the voltage of $L_m$ is inverted, and the magnetizing current $i_{Lm}$ is decreased linearly in positive direction. The secondary current of transformer $i_s$ is changed to be negative.

Stage 3 ($t_2-t_3$) [Fig.4(c)]: After $t = t_2$, the resonant inductor $L_r$ and the junction capacitors of $S_1$ and $S_2$ keep on resonating, because the voltage across $L_m$ is clamped to $nV_o$. The current of magnetizing inductor decreases in positive direction, but its value is always larger than zero which guarantees the anti-parallel diodes of $S_4$ and $S_5$ to keep on conducting.

Stage 4 ($t_3-t_4$) [Fig.4(d)]: Before $t = t_3$, the anti-parallel diodes of $S_1$ and $S_2$ are conducting which ensure ZVS turn-on for $S_4$ and $S_5$. At $t = t_3$, $S_2$, $S_4$ and $S_5$ turn on, the current of resonant inductor increases sinusoidally in negative direction which is similar to Stage 1.

B. Mode 2: Forward Power Limitation Mode

When charging current of LIB is greater than that in rated condition, Mode 1 will change to this mode. In overload situation, both the currents of $L_r$ and the voltage across $C_1$ or $C_2$ increase. Once $V_{C1}$ or $V_{C2}$ reaches $V_{o1}$, $D_2$ or $D_1$ starts conducting. Then $C_1$ and $C_2$ will be bypassed. Therefore, the resonance among $C_1$, $C_2$ and $L_r$ will be stopped. When $D_1$ or $D_2$ conducts, the energy transfer will be interrupted from primary side to the secondary side of the transformer which means that the power limitation occurs. Moreover, the resonant current $i_n$ will be limited which protects the switching networks and the resonant capacitors. Fig.6 demonstrates the working stages and the key waveforms of the converter under this mode are shown in Fig.7.

Stage 1 ($t_0-t_1$) [Fig.4(a)]: This stage is the same as that in Mode 1.

Stage 2 ($t_1-t_2$) [Fig.6(a)]: Before $t = t_1$, $L_r$ resonates with $C_1$ and $C_2$, and the voltage across $C_2$ increases sinusoidally. After that, $V_{C2}$ reaches $V_{o1}$, and $D_1$ are conducted which causes the voltage across resonant inductor to change to $-nV_o$. After $t = t_1$, the current of $L_r$ decreases linearly in positive direction. The resonance tank stops work, and there is no energy transmitting to the secondary side of the transformer.

Stage 3 ($t_2-t_3$) [Fig.6(b)]: At $t = t_2$, $S_1$, $S_3$ and $S_6$ are turned off. And the anti-parallel diodes of $S_2$, $S_4$ and $S_5$ conduct after the voltages of their junction capacitors decreases to zero. The voltage across resonant inductor $L_r$ then changes to $-(V_{o1}+nV_o)$ which speeds up the reduction for the current of $L_r$ till it decreases to zero.

Stage 4 ($t_3-t_4$) [Fig.6(c)]: After the current of $L_r$ decreases to zero, the junction capacitors of $S_1$ and $S_3$ begin to resonate with $L_r$. The anti-parallel diodes of $S_4$ and $S_5$ keep on conducting because $i_s$ is larger than zero.

Stage 5 ($t_4-t_5$) [Fig.4(d)]: This stage is the same as that in Mode 1.

C. Mode 3: Reverse Power Transmission Mode

For LIB Formation and Grading System, when LIB is discharged, the energy of LLC converter will be transmitted from the secondary side to the primary side of transformer, which is in contrary to the forward transmission mode. Fig. 8 shows the working stages, and the key waveforms of the converter under this mode are shown in Fig. 9.
and $S_6$ are turned on with ZCS. And the magnetizing current $i_{Lm}$ decreases linearly in negative direction until it changes to zero for the voltage across the magnetizing inductor $L_m$ is $nV_o$. The magnetizing current will increase in positive direction after it changes to zero. This mode ends when the secondary current $i_r$ is decreased to zero.

Stage 2 ($t_1$-$t_2$) [Fig.8(b)]: At $t = t_1$, $S_1$, $S_3$ and $S_6$ are turned off. After that, the junction capacitors of $S_1$, $S_3$, $S_2$ and $S_6$ resonate with $L_r$ and the voltage across the junction capacitor of $S_2$ decreases to zero. Then, the anti-parallel diode is conducted. Afterwards, the voltage across $L_m$ is inverted, and the magnetizing current $i_{Lm}$ is decreased linearly in positive direction.

Stage 3 ($t_2$-$t_3$) [Fig.8(c)]: After $t = t_2$, the resonant inductor $L_r$ and the junction capacitors of $S_1$, $S_3$, $S_2$ and $S_6$ keep on resonating for the voltage across $L_m$ is clamped to $nV_o$. The current of magnetizing inductor decreases in positive direction, but its value is always larger than zero which guarantees the anti-parallel diodes of $S_2$ to keep on conducting.

Stage 4 ($t_3$-$t_4$) [Fig.8(d)]: Before $t = t_3$, the anti-parallel diode of $S_2$ is conducting which ensures ZVS turn-on for $S_1$. At $t = t_3$, $S_2$, $S_1$ and $S_6$ are turned on, the current of resonant inductor increases sinusoidally in positive direction which is similar to Stage 1.

D. Mode 4: Reverse Power Limitation Mode

When discharging current is higher enough, Mode 3 will change to this mode. The voltage of $C_1$ or $C_2$ will reach $V_{in1}$ if the current of $L_r$ is high enough. This will cause the conduction of $D_1$ or $D_2$, and stop the resonance in the resonant tank.
The current waveshapes are illustrated in Fig. 7. During Stage 2, the resonant tank stops working, whose key waveforms are illustrated in Fig. 11. Then resonance occurs among $L_1$, $C_1$, and $V_{in}$, and the current of $L_1$ decreases linearly in positive direction. The energy keeps on transmitting from the secondary side of the transformer to the primary side.

Stage 3 ($t_3$-$t_4$) [Fig. 10(b)]: This stage begins when $S_1$, $S_3$, and $S_6$ are turned off. After that, the junction capacitors of $S_1$ and $S_6$ are discharged by $i_t$ until the voltages across $S_1$ and $S_6$ decrease to zero. The anti-parallel diodes of $S_1$ and $S_6$ then conduct which changes the voltage across the magnetizing inductor. Then the current of the magnetizing inductor $L_m$ decreases linearly in negative direction. The voltage across the resonant inductor $L_r$ changes to $V_{in}$, and the current of $L_r$ decreases quickly to zero.

Stage 4 ($t_4$-$t_5$) [Fig. 10(c)]: At $t = t_4$, $i_t$ changes to zero, and the anti-parallel diode of $S_2$ then conducts. After that, $L_r$ resonates with the junction capacitors of $S_2$, $S_1$, $S_3$, $S_5$, and $S_6$.

Stage 5 ($t_5$-$t_6$) [Fig. 8(d)]: This stage is the same as that in Mode 3.

### III. ANALYSIS OF THE MODIFIED CONVERTER UNDER THE PROPOSED MODULATION

#### A. Constant Voltage Gain under the Proposed Modulation

As Fig. 12 shows, $i_{out}(t)$ is equal to $[2n(i(t) - i_{in}(t))]$, the average value of $i_{out}(t)$ is written by

$$\bar{i}_{out} = \frac{2}{T} \int_{t_1}^{t_2} 2n[i(t) - i_{in}(t)] \, dt$$

(2)

Where $i_{in}(t)$ is the input current of the primary side switching network, and $i_{out}(t)$ is the output current of the secondary side switching network which is shown in Fig. 3. The average value of $i_{in}$ is zero. Hence, the average value of $i_{out}(t)$ can be simplified as

$$\bar{i}_{out} = \frac{4n}{T} \int_{t_1}^{t_2} i(t) \, dt = 4n \cdot \bar{i}_{in}$$

(3)

The input power and output power should keep balance if ignoring the losses, so the voltage gain can be expressed by

$$M = \frac{V_{out}}{V_{in}} = \frac{\bar{i}_{out}}{\bar{i}_{in}} = \frac{1}{4n}$$

(4)

From the analysis above, it is concluded that the voltage gain keeps a constant value as long as power limitation condition is not triggered. In practice, the conduction resistance and the line impedance could not be neglected, which causes voltage drop of the converter, affecting the voltage gain. Thus, the output voltage considering the conduction resistances can be calculated as

$$V_o = \frac{V_{\text{out}}}{8n} \sqrt{\frac{V_{\text{in}}^2 - 16P_c}{R_p + R_{p,\text{ds}} + n^2(R_L + R_{\text{copper}} + 2R_{\text{clp}})}}$$

(5)

where $R_p$ is the copper resistance of the primary side, $R_{p,\text{ds}}$ is the conduction resistance of the primary side switches. $R_{\text{copper}}$ is the copper resistance of the transformer secondary side. $R_L$ is the line impedance in output side of the converter. $R_{\text{clp}}$ is the conduction resistance of the secondary side switches.

#### B. Power Limitation of the Modified Converter

When the converter works under forward transmission mode, the transmitted power will be limited if the voltage across $C_1$ or $C_2$ reaches $V_{in}$. Simultaneously, the clamping diode of $C_1$ or $C_2$ conducts, which cuts off the path for the energy to transmit.

At $t = t_1$, the resonant tank stops working, whose key waveshapes are illustrated in Fig. 7. During $t_0$ to $t_1$, the differential equations of the resonant tank can be expressed as

$$L \frac{di_{in}(t)}{dt} = V_{\text{in}} - 2nV_o - u_{c_2}(t)$$

(6)

$$2C \frac{du_{c_2}(t)}{dt} = i_c(t)$$

Where

$$u_{c_2}(t) + u_{c_2}(t) = V_{\text{in}}$$

(7)

Based on (6) and (7), $i_c(t)$, $u_{c_2}(t)$ and $u_{c_2}(t)$ can be expressed as

$$i_c(t) = i_c(0)\cos(\omega t) + \frac{V_{\text{out}} - 2nV_o - u_{c_2}(0)}{\omega L_c} \sin(\omega t)$$

(8)
guarantee ZVS condition for $S_1$ and $S_5$ when they are turned on. In order to discharge/charge the junction capacitors completely, the energy stored in magnetizing inductor should be higher than that in the four junction capacitors. 

$$\frac{1}{2} L_{m_{peak}} i_{m_{peak}}^2 \geq \frac{1}{2} (C_{S_5} + C_{S_4} + C_{S_3} + C_{S_6}) V_o^2$$

(15)

Where $C_{S_3}$, $C_{S_4}$, $C_{S_5}$ and $C_{S_6}$ are the junction capacitors of $S_3$, $S_4$, $S_5$ and $S_6$ respectively. $i_{m_{peak}}$ is the peak value of the magnetizing current which can be expressed as

$$i_{m_{peak}} = \frac{n V_o}{4 L_m f_s}$$

(16)

$i_i$ will decrease after $S_1$, $S_3$ and $S_6$ are turned off, but it is always positive which guarantees ZVS for $S_2$, $S_4$ and $S_5$. So the current of magnetizing inductor should be larger than zero during the deadtime.

$$i_{m_{peak}} - T_{dead} \frac{n V_o}{L_m} \geq 0$$

(17)

Where $T_{dead}$ is the deadtime which can be expresses as

$$T_{dead} = \frac{1}{2 f_s} - \frac{1}{2 f_i}$$

(18)

In practice, the same type devices are used for a full bridge, so the values of $C_{S_3}$, $C_{S_4}$, $C_{S_5}$ and $C_{S_6}$ are assumed the same. Based on (15) and (16), the range of the magnetizing inductance can be calculated, which is expressed as

$$L_m \leq \frac{n^2}{64 f_s C_{S_3}}$$

(19)

Meanwhile, magnetizing current should be larger than zero during the deadtime, so the duration of $T_{dead}$ is limited. According to (17) and (18), it can be deduced that

$$2 f_i \geq f_s$$

(20)

In the reverse power transmission mode, the switch network of primary side can achieve ZVS, the resonant inductor current equals to $i_{m_{peak}}$ when $S_1$, $S_3$ and $S_6$ are turned off. The junction capacitors of $S_1$ and $S_2$ are charged/discharged by $i_i$, and then the anti-parallel diode of $S_2$ conducts. In order to guarantee ZVS condition of $S_2$, the value of $i_i$ should be large enough to charge/discharge the junction capacitors completely.

$$\frac{1}{2} L_i i_{m_{peak}}^2 \geq \frac{1}{2} (C_{S_1} + C_{S_2}) V_{int}^2$$

(21)

It is assumed that the values of $C_{S_1}$ and $C_{S_2}$ are the same. Based on (16) and (21), the range of the magnetizing inductance is

$$L_i \leq \frac{n^2 V_i^2 L_f}{32 C_{S_1} f_s V_{int}}$$

(22)

In this mode, the deadtime should also be properly designed to guarantee ZVS condition, which is the same with that in the forward power transmission mode.

From analysis above, the range of the magnetizing inductance is shown as (19) and (22). From (20), it can be seen that the resonant frequency needs to be lower than twice of the switching frequency, otherwise, the switches will lose ZVS.
IV. ANALYSIS OF POWER SHARING FOR ISOP LLC SYSTEM UNDER THE PROPOSED MODULATION

For high power LIB Formation and Grading System, the AC/DC stage is a three-phase rectify whose output voltage is usually controlled between 600V and 750V (380VAC input), and the output current is large (such as 500A). In order to reduce the voltage stress for the primary side switching network and the current stress of the secondary side switching network, an ISOP structure is used. The schematic block of two-module ISOP LLC system is shown in Fig.14, where \( V_{in1} \) and \( V_{in2} \) are the input voltage of each LLC module. \( I_{out1} \) and \( I_{out2} \) are the output current of each LLC module. It is assumed that the efficiency of each LLC converter is 100%, so the input power of the LLC converter is equal to output power which is shown as

\[
\begin{align*}
V_{in1} \cdot I_{in1} &= I_{out1} \cdot V_{o1} \\
V_{in2} \cdot I_{in2} &= I_{out2} \cdot V_{o2}
\end{align*}
\]

(23)

If the output current balance between each LLC converter are realized, so

\[
I_{out1} = I_{out2}
\]

(24)

According to (23), in order to achieve the output current balance, \( V_{in1} \) and \( V_{in2} \) should be controlled to keep balance, namely

\[
V_{in1} = V_{in2}
\]

(25)

It can be known from (24) and (25) that IVS control or OCS control should be adopted to keep two sub-modules sharing the same output power. However, for the high output current application, it does require more effort to sample output current than input voltage. So, achieving IVS is preferred, which controls the voltage gain of each LLC converter.

As for the convention LLC converter, a variable-frequency control is adopted to realize input voltage balance in facing with parameters mismatching. First Harmonic Approximation (FHA) is employed to analyze the voltage gain for the LLC converters. For a single LLC converter, the equivalent model is shown in Fig.15.

\[
R_{ac} = \frac{8\pi^2 R_c}{\pi^2}
\]

(26)

where \( R_c \) is the load resistance. The voltage gain of the LLC converter is equal to the gain of the resonant tank. The gain of the LLC converter can be expressed as

\[
M = \frac{f^2(m-1)}{\sqrt{(m/f)^2-1} + f^2((f^2-1)^2 \cdot (m-1)^2 \cdot Q^2}}
\]

(27)

Where

\[
\begin{align*}
f &= \frac{f_s}{f_r} \\
Q &= \frac{\sqrt{L_c/C_r}}{R_w} \\
m &= \frac{L_m + L_r}{L_r}, f_r = \frac{1}{2\pi\sqrt{L_rC_r}}
\end{align*}
\]

(28)

The output gain curves with different loads are illustrated in Fig.16. It can be seen from Fig.16 that the variable-frequency control can be adopted to keep input voltage balance through adjusting the operating frequency of each LLC converter when the load is changed or the parameters of two LLC converter are different.

![Figure 14. Schematic block of two-module ISOP LLC system.](image)

If the LLC converter operates under the proposed modulation, the voltage gain of the LLC converter is different with that in variable-frequency control. Fig.17 shows the voltage gain curves of the LLC converter under the proposed modulation. It can be seen from Fig.17 that the voltage gain will be kept to a constant value if the operating frequency of each converter is below the resonant frequency. If the parameters of each sub-module are the same, the voltage gain will be constant for each sub-module, which can achieve automatic power sharing. When the parameters are different between two sub-modules, the voltage gain of two modules will be different if the duty cycles of the driving signals are the same. In order to get a fixed voltage gain of two modules, it can change the duty cycle of the driving signals to deal with parameters mismatching while the operating frequency keeps a stable.
constant value and the working modes for two sub-modules are the same. As for the convention LLC converter, in order to achieve IVS and OCS control, the operating frequency of two sub-modules may be different, which will cause difference in working condition and ZVS condition between two sub-modules.

V. DESIGN CONSIDERATION

A. The Turns Ratio of the Transformer

As analyzed in Section III, with proposed modulation, the multi-transformer LLC converter can achieve constant voltage gain as long as the output power is not higher than the power limitation value.

The relationship between the voltage gain and the turns ratio of the transformer was deduced in (4). And the turns ratio of the transformer is expressed as

\[ n = \frac{V_m}{8V_o} \]  

B. Resonant Tank Design

When LLC converter works in power limitation mode, the voltages across the resonant capacitors are clamped by the clamping diodes which limits the power. So, the resonant capacitance determines the maximum output power and it can be calculated according to the maximum power, which is shown as

\[ C_i = \frac{P_{\text{max}}}{2f_iV_{m1}^2} \]  

\[ P_{\text{max}} \] is the maximum output power of each LLC converter. And the resonant inductance can be calculated by

\[ L_r = \frac{1}{8\pi C_i f_r^2} \]  

C. Current Stress of The Clamping Diodes

When overload occurs, \( D_1 \) and \( D_2 \) will be conducted. As Fig.7 shows, at \( t = t_1 \), the voltage of \( C_1 \) decreases to zero and \( D_1 \) then conducts. After that, the current of \( D_1 \) decreases linearly, because the voltage across \( L_r \) changes to \(-2nV_o\). After \( S_1 \) is turned off, the voltage across \( L_r \) changes to \(<(2nV_o+V_{m1})\). So, the current of \( D_1 \) can be shown as

\[ i_{D1}(t) = \begin{cases} 
  i_{D1_{max}} - \frac{2nV_o}{L_r}(t-t_1), & t_1 < t \leq t_2 \\
  i_{D1_{max}} - \frac{V_{m1} + 2nV_o}{L_r}(t-t_2), & t_2 < t \leq t_3 
\end{cases} \]  

where \( I_{D_{\text{max}}} \) is the current of \( D_1 \) when \( t = t_1 \) which is shown as

\[ I_{D_{\text{max}}} = i_i(\Delta t) \]  

where \( \Delta t \) is shown in (11). And \( t_2 \) can be solved by

\[ t_2 - t_1 = \frac{1}{2f_r} \Delta t \]  

At \( t = t_3 \), the current of \( D_1 \) decreases to zero, so \( t_3 \) can be solved as

\[ t_3 = \frac{V_{m1} + 2f_rL_rI_{D_{\text{max}}} + 4f_rV_{m1}n}{2f_rV_{m1} + 4f_rV_{m1}n} \]  

Then the RMS of \( i_{D2}(t) \) can be expressed as (36).

\[ \text{Fig.18. The comparison of the resonant current and the clamping diode current when over load condition.} \]

VI. COMPARISON AND EXPERIMENTAL VERIFICATION

A. Modulation Strategy Comparison

With the proposed modulation, LLC converter achieves naturally bidirectional power transfer and theoretically keeps a constant voltage gain. In practice, for the primary side and secondary side switches, all driving signals whose duty cycle always keeps constant are the same regardless of the load. Besides, when the power flow direction is changed, the energy of resonant tank will be automatically and smoothly transferred. Under the proposed modulation, PWM modulation of LLC converter is actively generated, no needing of considering the power transfer direction and current zero crossing point. In addition, the output voltage gain can keep constant in ideal theoretical analysis and changes slightly in practice due to the effect of conduction resistances.

In [21], a modulation was proposed to achieve automatic forward and backward mode transition for LLC converter.

\[ i_{D1_{RMS}} = \sqrt{2f_i\left(\frac{1}{2}v_{D1_{RMS}}^2+\frac{1}{2}i_{D1_{RMS}}^2\right)} \]  

To simplify the calculation, magnetizing current is neglected.

\[ \text{Fig.18. The comparison of the resonant current and the clamping diode current when over load condition.} \]
There are six working modes of the converter with different voltage gain and output current state. In the modulation, all the MOSFETs in the converter are operated with same frequency, but the pulse width of gate driving signals for primary side MOSFETs and secondary side MOSFETs are different according to the working modes.

In [20], a control strategy was proposed to achieve bidirectional power flow automatically without power flow detection. Under the modulation, the key waveforms of the converter which are simulated by PSIM are illustrated in Fig.19.

As seen, the pulse width of the driving signals of primary side switches and secondary side switches always keep constant whatever the power flow direction is. When the converter works in backward mode, for the asymmetric driving signal of primary side switches and secondary side switches, circulating current will be generated, which leads to high RMS value of transformer secondary current.

The comparisons for resonant current and transformer secondary current are show in Fig.20. As seen, when LLC converter works in forward mode, both the RMS value of the resonant current and RMS value of transformer secondary current are almost the same for two kinds of control strategies. However, when the converter works in reverse mode under the control strategy proposed in [20], the RMS value of transformer secondary current is higher than that when the converter works under the proposed modulation. Consequently, with the proposed modulation, the circulating current of LLC converter will be reduced.

B. Experimental results

An 8kW experimental prototype composed of two DC-DC modules using ISOP configuration is built to verify the effectiveness of the proposed solution. The specifications of each DC-DC module are given in TABLE I. The input voltage of the converter is 720V, and the output voltage is 15V. The switching frequency of the converter is 75kHz. The primary side switches are IPW65R041CFD produced by Infineon Technologies and the secondary side switches are PSMN1R5–40PS produced by Nexperia. The clamping diodes are ES8GC produced by Jingdao Microelectronics. Each switch on the secondary side is connected in parallel by two MOSFETs. The experimental prototype is shown in Fig.21.

The steady-state waveforms under the proposed modulation are shown in Fig.22. Fig.22(a) is the forward transmission waveforms for the resonant current and the driving signals, and Fig.22(b) shows the waveforms under the reverse power transmission mode. It can be seen that the driving signal is the same for the primary side and the secondary side. And the resonant current is discontinuous as the resonance starts at when the driving signals are turned on.

Fig.23. shows the soft-switching waveforms. In the forward power transmission mode, the switches of the secondary side can realize ZVS whose waveforms are illustrated in Fig.23(a),
and the switches of the primary side can achieve ZVS in the reverse power transmission mode whose waveforms are shown in Fig.23(b).

**TABLE I**

**PARAMETERS OF THE EXPERIMENTAL PROTOTYPE**

<table>
<thead>
<tr>
<th>Item</th>
<th>Detail</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Voltage ($V_{in}$)</td>
<td>720V</td>
</tr>
<tr>
<td>Output Voltage ($V_{o}$)</td>
<td>15V</td>
</tr>
<tr>
<td>Rated Power ($P$)</td>
<td>8000W</td>
</tr>
<tr>
<td>Turns Ratio ($n$)</td>
<td>6:1</td>
</tr>
<tr>
<td>Resonant Capacitance ($C_1$ &amp; $C_2$)</td>
<td>250nF</td>
</tr>
<tr>
<td>Resonant Inductance ($L_r$)</td>
<td>5.8μH</td>
</tr>
<tr>
<td>Magnetizing Inductance ($L_m$)</td>
<td>125μH</td>
</tr>
<tr>
<td>Switching Frequency ($f_s$)</td>
<td>75kHz</td>
</tr>
<tr>
<td>Primary side switches</td>
<td>IPW65R041CFD</td>
</tr>
<tr>
<td>Secondary side switches</td>
<td>PSMN1R5-40PS</td>
</tr>
<tr>
<td>Clamping Diodes</td>
<td>ES8G°C</td>
</tr>
</tbody>
</table>

![Figure 22. Steady-state waveforms under the proposed modulation.](image)

(a) Forward power transmission mode.
(b) Reverse power transmission mode.

![Figure 23. Soft-switching waveforms under the proposed modulation.](image)

(a) Forward power transmission mode.
(b) Reverse power transmission mode.

![Figure 24. Dynamic waveforms under the proposed modulation.](image)

(a) Forward power transmission mode to reverse power transmission mode.
(b) Reverse power transmission mode to forward power transmission mode.

![Figure 25. Dynamic response of the load step-change for the converter.](image)

(a) Forward power transmission mode.
(b) Reverse power transmission mode.

Fig.25 shows the dynamic response of the load step-change for the converter. Fig.25(a) illustrates the response time, which is 2ms, for the load changes from 3600W (15V, 240A) to 5400W (15V, 360A). The load changes from 3600W (15V, 240A) to 5400W (15V, 360A) is shown in Fig.25(b) and the response time is 1ms.
The soft-start waveforms are shown in Fig. 26. It can be seen that the current is limited in an acceptable range. The waveforms in power limiting mode are shown as Fig. 27. The waveforms for the voltage of resonant capacitor and the current of resonant inductor in normal operation mode are shown in Fig. 27(a). From Fig. 27(b), it can be seen that the voltage across the resonant capacitor is clamped by the clamping diodes and the output voltage is dropped to achieve constant output power.

![Figure 25](image-url) Dynamic response of the load step-change for the converter. (a) 3600W to 5400W. (b) 5400W to 3600W.

![Figure 26](image-url) Soft-start waveforms.

![Figure 27](image-url) Power limitation waveforms. (a) Normal operation mode. (b) Over load operation mode.

![Figure 28](image-url) IVS and OCS waveforms. (a) Steady waveforms (b) Load changing waveforms.

The waveforms of IVS and OCS are shown in Fig. 28. It can be seen that the resonant currents of two LLC modules are the same except the phase-shifted angle for that interleaved PWM signals are adopted to reduce the output current ripples. And the input voltages are the same which are shown in Fig. 28(a). When the load is changed, IVS and OCS can also be obtained with proposed modulation, which are shown in Fig. 28(b) and Fig. 28(c) where the input voltages of two modules are almost the same.

Fig. 29 shows the waveforms when resonate parameters of the two modules are mismatched. The resonant capacitor of module #1 is 250nF, and the other one is 313nF, which is 25% larger. Fig. 29(a) shows the resonant current and driving signals for two modules, and Fig. 29(b) shows the input voltage sharing waveforms. As seen, even though the mismatching of parameters will cause overshoot of resonant current, the difference of the input voltages ($V_{in1}$-$V_{in2}$) is still small as shown in Fig. 29(b), indicating that the power sharing can be well...
achieved. Therefore, with the proposed modulation, the two modules of the converter will keep balance in power even though the parameters of resonant tank are mismatched in a controlled tolerance (25%).

The difference will be larger as the load increases, mainly caused by the conduction voltage drop of the power devices. Although the voltage gain is changed for the conduction resistance, the load regulation of the converter is acceptable for LIB Formation and Grading System, because there is another voltage regulation stage as shown in Fig.1. It can be seen from the curves that the output voltage is close to that in ideal case when output power is below 2kW. When the power increases, the voltage drop will increase slightly, which is within 0.5V. And the experimental results are close to the results of the theoretical analysis. The efficiency of the LLC converter is shown in Figure 31. The maximum efficiency of the LLC converter is 95% and the full load efficiency is 91%.

![Figure 29. The waveforms when the parameters of two resonant tank are mismatched.](image)

(a) The driving signals and the resonant currents.
(b) The balance in the input voltages of two modules.

![Figure 30. The output voltage curves with different power.](image)

**TABLE II**

<table>
<thead>
<tr>
<th>Item</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_{\text{gs1}}$</td>
<td>0.12 mΩ</td>
</tr>
<tr>
<td>$R_{p_{\text{ds}}}$</td>
<td>20.5 mΩ</td>
</tr>
<tr>
<td>$R_{\text{gs2}}$</td>
<td>0.13 mΩ</td>
</tr>
<tr>
<td>$R_{L}$</td>
<td>0.11 mΩ</td>
</tr>
<tr>
<td>$R_{d_{\text{ds}}}$</td>
<td>1.3 mΩ</td>
</tr>
</tbody>
</table>

![Figure 31. Efficiency of the LLC converter.](image)

**REFERENCES**


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