High-Gain High-Efficiency IPOS LLC Converter With Coupled Transformer and Current Sharing Capability

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Abstract—In this paper, an input-parallel output-series (IPOS) LLC resonant converter with a coupled transformer and current sharing capability is proposed for high-gain high-efficiency applications such as the renewable energy grid connection. The coupled transformer is utilized to reduce the core size and core loss by implementing the magnetic flux cancellation method, which achieves a natural balance of output voltage simultaneously. The current sharing between the converter modules is achieved by a bridge wire that interconnects the resonant tanks of different modules. Furthermore, the current sharing inside a module is also achieved by the proposed compensation current based on the coordination of compensation inductor and delay control. A 200 kHz 100 V/1.2 kV 3.6 kW two-module IPOS LLC resonant converter prototype is built to verify the proposed method. The peak efficiency can reach as high as 97.5%, and the current sharing between modules and inside a module are well presented.

Index Terms—Current sharing, coupled transformer, LLC converter, input-parallel output-series (IPOS).

I. INTRODUCTION

RENEWABLES accounted for almost half of the growth in power generation (49%) in 2017, and led by wind and solar, the share of renewables in global power generation increased from 7.4% to 8.4% [1], [2]. With the increasing penetration renewables, dc distribution systems are becoming the preferred choice in the utility grid as they provide more natural interface with renewables including solar, fuel cells and wind turbine based on permanent-magnet generators, as well as energy storage including battery and super capacitors. In addition, dc distribution system has better current carrying capacity and faster dynamic response when compared to conventional ac systems.

To interconnect the renewables such as solar panel to the

Fig. 1. IPOS configuration for renewables grid connection.

Fig. 2. IPOS LLC resonant converter.

medium voltage grid, high-gain DC-DC converters are usually required. The input-parallel output-series (IPOS) configuration (Fig. 1) is very attractive in high input current and high output voltage applications [3]–[7]. Due to the parallel and series connection of inputs and outputs, the current and voltage stress on each module is reduced, which allows the use of low-voltage power devices with low on-resistance and enables a higher system efficiency.

In the grid connected applications, the voltage is usually maintained stable for both the low-voltage (LV) and medium-voltage (MV) side. Therefore, an unregulated DC-DC transformer (DCX) can be used to serve as the DC-DC module in IPOS circuit, which helps to improve the efficiency and simplify the control of the isolated stage. The full-bridge converters with phase-shift control have been widely employed in high-power conversion [8]–[12], which is however hard to achieve zero-voltage-switching (ZVS) at light load condition and usually has a high turn-off current [13], [14]. Thus, the efficiency is hard to further improve.

As shown in Fig. 2, the LLC resonant converter is becoming one of the most attractive choices because it can achieve ZVS for primary switches and zero-current switching (ZCS)

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for secondary rectifiers simultaneously [15], [16], resulting in a high efficiency and low electromagnetic interference. The excellent soft switching characteristics also enable the LLC resonant converter operating at a higher frequency and achieving a higher power density. Besides, the LLC converter is particularly suitable for DCX applications because it can run at the fixed resonant frequency and therefore the efficiency benefits can be maximized [16].

For the IPOS configuration, balancing of input currents and output voltages is vital for a reliable and high efficient operation. The relationship between output voltage and input current sharing was revealed in [17] and an output voltage sharing loop was proposed. A current sharing method for interleaved LLC converter with partial energy processing is presented in [18], where a transformer and auxiliary dc-dc converters are added for current sharing loops. However, the additional voltage current sharing loop requires a complex control strategy, which becomes an obstacle when the LLC converter operates as a DCX.

To achieve the automatic current sharing, a common passive impedance method was proposed in [19]–[21], which exhibits a good performance in multi-phase half-bridge LLC converters. However, the Δ/Y equivalent circuit method proposed in [19] and [20] is not applicable to full-bridge LLC converters when current sharing both between modules and inside a module needs consideration and therefore, a different manifest model is required. Besides, the current flows through the interconnection inductor is not under precise control for the passive compensation.

This paper proposes a high-gain high-efficiency IPOS LLC converter with coupled transformer and current sharing capability. To improve efficiency, the magnetic flux cancellation concept is utilized to construct a coupled transformer, which can not only reduce the core size and the core loss, but also balances the output voltage between modules as well. To achieve resonant current sharing between modules, a structure that utilizes a bridge wire to interconnect the resonant tanks between modules is proposed and the equivalent model is given to present the cause and solution for the current unbalance. In order to further achieve the current sharing inside a module, an active control method by coordinating the compensation inductor and delay control is proposed to generate compensation current. At last, to verify the proposed IPOS LLC converter, a laboratory prototype was built with rated parameters of 200 kHz 100 V/1.2 kV 3.6 kW. The experimental results show the proposed converter has a good performance on current sharing and a peak efficiency of 97.5%.

II. COUPLED TRANSFORMER WITH FLUX CANCELLATION

An IPOS LLC resonant converters contains multiple DC-DC modules, and each module requires a transformer, and these transformers can account for a large part of the whole converter loss, thus the transformer design is critical. Magnetic flux cancellation can reduce the core losses and size, and a matrix transformer configuration with magnetic cancellation is proposed in [22], and this concept is applied to the IPOS LLC

Fig. 3. Diagram of transformer. (a) With separate cores. (b) Coupled by a common core.

Fig. 4. Equivalent magnetic circuit of transformer.

converter to reduce the core losses.

The concept of the magnetic flux cancellation is shown in Fig. 3. When two separate magnetic cores are adopted (Fig. 3(a)), properly arranging primary windings of the two modules can let Φ_1 and Φ_2 , the fluxes in core 1 and core 2, flow in the same direction. When the winding current i_1 and i_2 of two modules are the same, the fluxes in the neighbor legs of core 1 and core 2 are of the same magnitude but in reverse direction. Fig. 4 shows the equivalent magnetic circuit model of Fig. 3, and we can derive:

$$\Phi_1 = \Phi_2 = \Phi, \quad (1)$$

when

$$\begin{cases} F_1 = F_2 \\ R_{m1} = R_{m2} \end{cases}. \quad (2)$$

Therefore, the center cylinder of two cores can be reaped and the two transformers can be implemented using a single U-shape core, as shown in Fig. 3(b). Despite the elimination of center cylinder, the magnetic flux flows in the remaining part of core is unchanged. Hence, the core size and the core loss are

Fig. 5. IPOS LLC converter with coupled transformer.

well reduced.

With the coupled transformer in Fig. 3(b), the magnetic flux flows through winding of two modules is forced to be identical. The voltage on the secondary side of transformer can be derived as follows:

$$v_{s1} = v_{s2} = n \frac{d\Phi}{dt}. \quad (3)$$

According to (1), the secondary side voltage of two modules v_{s1} and v_{s2} can be naturally balanced, therefore output voltage V_{o1} and V_{o2} in Fig. 2 are naturally balanced.

III. CURRENT SHARING BETWEEN MODULES

For a converter with multiple modules, it is essential to distribute current evenly among different modules to allow the thermal balance and to ensure stable operation. When the transformers of the IPOS modules are coupled, as shown in Fig. 5, current sharing between modules becomes critical. To simplify the analysis, gate signals of the power devices of two modules in Fig. 5 are considered the same and without delays, we obtain:

$$\begin{cases} v_1 = v_2 \\ v_3 = v_4 \end{cases}. \quad (4)$$

When the two modules operate nears the resonant frequency, based on fundamental harmonic approximation (FHA) [23], [24], an equivalent circuit is shown in Fig. 6. The voltage relationship between input and output is expressed as follows:

$$\begin{cases} v_{13} = i_1 \cdot Z_1 + \frac{v_{s1}}{n} \\ v_{24} = i_2 \cdot Z_2 + \frac{v_{s2}}{n} \end{cases}, \quad (5)$$

where

$$\begin{cases} Z_1 = j\omega L_{r1} + \frac{1}{j\omega C_{r1}} \\ Z_2 = j\omega L_{r2} + \frac{1}{j\omega C_{r2}} \end{cases}. \quad (6)$$

Fig. 6. Equivalent circuit of IPOS LLC converter.

Based on (1), (2) and (3), the relative relationship between the resonant currents of the two modules can be derived:

$$\frac{i_1}{i_2} = \frac{Z_2}{Z_1} = \frac{j\omega L_{r2} + \frac{1}{j\omega C_{r2}}}{j\omega L_{r1} + \frac{1}{j\omega C_{r1}}}. \quad (7)$$

Current unbalance between modules is mainly attributed to the component tolerance in the resonant inductor or capacitor. Considering 5% tolerance in L_{r1} and C_{r1} respectively, and we use $|(i_1 - i_2)/i_2|$ to evaluate the relative error of currents between two modules. Fig. 7 shows the relative error of the currents changing with the resonant components L_{r2} and C_{r2} respectively, where L_r and C_r represent the nominal resonant inductance and capacitance. The current-sharing error increase dramatically with light different between resonant components and can reach 100% when $L_{r1} = 1.05L_{r2}$ or $C_{r1} = 1.05C_{r2}$.

To balance the current between modules, a wire bridge is inserted to short-circuit the resonant capacitors of two modules, as shown in Fig. 8, whose equivalent circuit model is shown in Fig. 9. With this short wire, we obtain:

$$v_5 = v_6. \quad (8)$$

Furthermore, the voltage constraints can be described as:

$$\begin{cases} v_{15} = i_1 \cdot j\omega L_{r1} + \frac{v_{s1}}{n} \\ v_{26} = i_2 \cdot j\omega L_{r2} + \frac{v_{s2}}{n} \end{cases}, \quad (9)$$

$$\begin{cases} v_{53} = i_3 \cdot \frac{1}{j\omega C_{r1}} \\ v_{64} = i_4 \cdot \frac{1}{j\omega C_{r2}} \end{cases}. \quad (10)$$

Combining (3), (4), (8), (9) and (10) the relationship between the resonant currents of the two modules can be derived:

(a)

(b)

Fig. 7. Current sharing error versus. (a) Inductance tolerance. (b) Capacitance tolerance.

Fig. 8. IPOS LLC converter with short-circuit resonant capacitors.

$$\begin{cases} \frac{i_1}{i_2} = \frac{L_{r2}}{L_{r1}} \\ \frac{i_3}{i_4} = \frac{C_{r1}}{C_{r2}} \end{cases} \quad (11)$$

Still consider 5% tolerance in L_{r1} and C_{r1} respectively, and Fig. 10 shows the current sharing error IPOS modules using the proposed balancing technique. The current-sharing is considerably improved, and even in the worst case, no more than 5% error is observed, which is acceptable in the practice

Fig. 9. Equivalent circuit with short-circuit resonant capacitors.

(a)

(b)

Fig. 10. Current sharing error versus. (a) Inductance tolerance. (b) Capacitance tolerance.

since the primary purpose of current sharing is to achieve thermal balance. The current sharing can be further improved by choosing resonant components with smaller tolerance.

IV. CURRENT SHARING INSIDE A MODULE

The interconnection of resonant tanks can balance the current between modules (i_1 and i_2 , i_3 and i_4), but current inside a module can still be different (i_1 and i_3 , i_2 and i_4) because the current i_5 in Fig. 9 that flows through the interconnection wire:

Fig. 11. IPPOS LLC converter with compensation inductor.

Fig. 12. (a) Gate signals and H-bridge output voltages. (b) Resonant currents.

$$i_5 = i_1 - i_3 = i_4 - i_2. \quad (12)$$

The current i_5 reflects the difference between i_1 and i_3 , i_2 and i_4 . To balance the current inside a module, additional method is proposed to suppress the current i_5 . A compensation inductor L_s is utilized to replace the interconnection wire, as shown in Fig. 11. With proper gate signal delay control, a compensation current i_{comp} can be generated to offset i_5 and balance the currents inside a module.

To analyze the principle for generating compensation current i_{comp} , consider the gate signals of module 1 lag behind that of module 2 by a tiny period of time Δt , as shown in Fig. 12(a).

Fig. 13. Equivalent circuit under narrow pulse voltages.

Fig. 14. Simplified equivalent circuit under narrow pulse voltages.

As a result, the square voltage v_{13} (module 1) will fall behind v_{24} (module 2) by Δt , which will produce narrow pulse voltages v_{12} and v_{34} . From the waveform, the phase of v_{12} is 90° behind the square voltage v_{13} and v_{24} generated by H-bridges. Besides, v_{12} and v_{34} are of the same magnitude but in reverse phase:

$$v_{12} = -v_{34}. \quad (13)$$

The pulse v_{12} and v_{34} can be employed to balance the current inside a module. Fig. 13 shows the equivalent circuit. Because the coupled transformer shares the same core and $v_{s1} = v_{s2}$, as described in (1), the equivalent circuit can be further simplified as Fig. 14. Ignore the tolerance of resonant components, we obtain:

$$\begin{cases} L_{r1} = L_{r2} = L_r \\ C_{r1} = C_{r2} = C_r \\ \omega_r = \frac{1}{\sqrt{L_r C_r}} \end{cases} \quad (14)$$

Based on Fig. 14, the transfer function from v_{12} to i_{comp} can be derived, by combining with (11) and (12):

$$\frac{i_{comp}(\omega)}{v_{12}(\omega)} = H\left(\frac{\omega}{\omega_r}, \frac{L_s}{L_r}\right) \cdot \frac{1}{\omega_r L_r} \quad (15)$$

where:

$$H\left(\frac{\omega}{\omega_r}, \frac{L_s}{L_r}\right) = \frac{\left(\frac{\omega}{\omega_r}\right)^2 + 1}{j\left(\frac{\omega}{\omega_r}\right) \left[\frac{L_s}{L_r} - \frac{L_s}{L_r} \left(\frac{\omega}{\omega_r}\right)^2 + 2 \right]}. \quad (16)$$

To analyze the characteristics of i_{comp} , the bode diagram

Fig. 15. The bode diagram of H . (a) Amplitude-frequency. (b) Phase-frequency.

of $H(\omega/\omega_r L_s/L_r)$ is depicted in Fig. 15, where different compensation inductance L_s are considered. As shown in Fig. 15(b), the fundamental harmonic component of i_{comp} always lags behind v_{12} by 90° , while v_{12} lags behind the unbalanced current i_5 by 90° , as illustrated in Fig. 12. As a result, the fundamental harmonic component of i_{comp} lags behind i_5 by 180° , which can be utilized to offset the unbalanced current i_5 .

It is worth noting that in practice, the direction of current i_5 is determined by the relative magnitude of the resonant component values. Based on Fig. 9, (11) and (12), we can derive:

$$i_5 = \frac{\frac{C_{r2}}{C_{r1}} - \frac{L_{r1}}{L_{r2}}}{1 + \frac{C_{r2}}{C_{r1}}} \cdot i_1. \quad (17)$$

When $C_{r2}/C_{r1} > L_{r1}/L_{r2}$, the current i_5 flows in the same phase to i_1 , as depicted in 13(b), which requires the gate signals of module 1 lagging behind that of module 2 as discussed above. When $C_{r2}/C_{r1} < L_{r1}/L_{r2}$, the current i_5 flows in the opposite phase to i_1 . Therefore, the gate signals of module 1 should be adjusted to lead ahead of that of module 2 to generate the compensation current in opposite phase.

Meanwhile, the higher order harmonics of i_{comp} needs to be suppressed. As shown in Fig. 15, if L_s is too small, higher order harmonics i_{comp} of may be larger than its fundamental component, leading to the distortion of resonant current. Since i_{comp} only contains odd harmonics, the gain of third and higher harmonics should be restricted to be smaller than fundamental harmonic, which can be expressed as:

$$\begin{cases} \left| H\left(3, \frac{L_s}{L_r}\right) \right| \leq \left| H\left(1, \frac{L_s}{L_r}\right) \right| \\ \angle H\left(3, \frac{L_s}{L_r}\right) = 90^\circ \end{cases}. \quad (18)$$

By solving (18), we obtain:

$$L_s \geq \frac{2}{3} L_r. \quad (19)$$

Although larger L_s value seems to have a better suppression on higher order harmonics according to and Fig. 15 and (16), too large inductance will degrade the automatic current sharing between modules as discussed in Section III, in which case v_5 and v_6 cannot be deemed as equal for the large impedance introduced by L_s . Therefore, the design of L_s needs a compromise and L_s is chosen as $2/3 L_r$ in this work.

In order to well cancel the current i_5 , a proper delay time Δt should be chosen. The generated compensation current i_{comp} (Fig. 11) should satisfy:

$$|i_{\text{comp}}(\omega_r)| = |i_5|, \quad (20)$$

where $i_{\text{comp}}(\omega_r)$ can be derived from (15), and i_5 can be represented by $\varepsilon \cdot i_1$, ε is defined as the error coefficient based on (17). Therefore:

$$|i_{\text{comp}}(\omega_r)| = \left| \frac{v_{12}(\omega_r)}{j\omega_r L_r} \right| = |\varepsilon \cdot i_1|, \quad (21)$$

$v_{12}(\omega_r)$ is the fundamental component of narrow pulse voltage v_{12} :

$$v_{12}(\omega_r) = \frac{2\Delta t}{T_s} \frac{4}{\pi} V_i \cdot \sin \omega_r t, \quad (22)$$

i_1 is determined by the equivalent load R_{e1} (FHA), as shown in Fig. 9:

$$i_1 = \frac{1}{R_{e1}} \frac{4}{\pi} V_i \cdot \sin(\omega_r t + 90^\circ). \quad (23)$$

Combining (21), (22) and (23), the delay time Δt is derived as:

$$\Delta t = \frac{\varepsilon \pi L_r}{R_{e1}} = \frac{\varepsilon \pi^3 L_r}{2R_L} \cdot n^2. \quad (24)$$

R_L represents the load of the converter, and n represents turns ratio of the transformer, as shown in Fig. 11. The larger error coefficient ε requires a longer delay time Δt . Normally, the error coefficient is within the range of 0–5%.

V. EXPERIMENTAL RESULTS

A. Prototype

To verify the proposed coupled transformer and current sharing capability of IPOS LLC resonant converter, a 200 kHz 3.6 kW converter prototype is constructed as shown in Fig. 16, and the parameter specifications are shown in Table I. The prototype serves as a dc/dc transformer with a fixed voltage gain, and operate at a fixed switching frequency.

The primary devices are IRFP4568PBF MOSFETs from Infineon. The secondary rectifiers are IDH20G120C5 SiC Schottky diodes from Infineon. Printed circuit board (PCB) planar winding structure are adopted both for the transformer and resonant inductors. The magnetic material of core is ferrite DMR95 from DMEGC. By utilizing interleaving windings, the leakage inductance of the transformer is 50 nH (from primary side) as measured, which is negligible in the experiment. Therefore, the potential soft switching failure of power devices caused by the leakage inductance is suppressed [25]. Due to the implementation of current sharing, the current that flows through the compensation inductor is small, which will not cause large power dissipation. Therefore, an SMT inductor SRP7050TA-1R8M from BOURNS is chosen as the compensation inductor. The power density of the overall prototype is around 33 W/in³.

A PCB planar transformer is designed to demonstrate the coupled transformer concept. The cost of the PCB with more than 4 layers is usually high. To reduce the cost, the windings of the transformer is implemented using three identical 4-layer PCBs and each PCB has 4:4 turns, as shown in Fig. 17. The primary side of three PCBs are connected in parallel, while the secondary side are connected in series, and the resultant turns ratio is 4:12. For each PCB board, primary and secondary windings are perfectly interleaved layer by layer, with two coils on each layer. The magnetic motive force (MMF) between windings is minimized. The PCB winding structure is optimized based on the design method proposed in [26], [27]. Therefore, the leakage inductance as well as ac winding loss of transformer can be minimized.

Fig. 18(a) shows the 2-D finite-element simulation result of the proposed coupled transformer. The result shows that the MMF between windings is fairly small, which induces a small leakage inductance as well. Besides, Fig. 18(b) shows the simulation result of a separate transformer with the same core but a single winding. By comparison, it is obvious that the magnetic field distribution in the core is nearly identical either in the coupled transformer or separate transformer, which verifies that the coupled transformer does not change the normal magnetic flux density in the core, and the loss in two separate transformer cores can be reduced to that in a coupled transformer core.

Fig. 16. Prototype of a 200 kHz 3.6 kW converter.

TABLE I
SPECIFICATIONS OF PROTOTYPE

Parameter	Value
Switching frequency	200 kHz
Input voltage	100 V
Output voltage	1200 V
Dead time	150 ns
Transformer turns ratio	1 : 3
Resonant capacitor	216 nF(C_{r1}), 238 nF(C_{r2})
Resonant inductor	2.6 μ H(L_{r1}), 2.8 μ H(L_{r2})
Magnetizing inductance	25 μ H
Compensation inductor	1.8 μ H

B. Experimental Waveforms

Fig. 19 shows the experimental waveforms to validate the current sharing between modules without and with the interconnected resonant tanks respectively, which is tested under 50 V input, 600 V output and 600 W load condition. v_{13} and v_{24} are voltages on the resonant tanks of module 1 and 2 respectively. i_1 and i_2 denote the resonant currents through the resonant inductors L_{r1} and L_{r2} respectively. When there is no compensation inductor interconnecting two modules, approximately 30% current unbalance can be observed between two resonant tanks, as shown in Fig. 19(a). While the compensation inductor L_s is added to enable the current sharing, the two resonant currents that flows in L_{r1} and L_{r2} can reach nearly the same, as shown in Fig. 19(b).

Fig. 20 shows the experimental waveforms to validate the current sharing effect inside a module under different delay time between two modules, which is tested under 100 V input, 1.2 kV output and 2.4 kW load condition. i_3 and i_4 denotes the resonant current flows through the resonant capacitors C_{r1} and C_{r2} respectively. No delay, 20 ns delay and 40 ns delay are tested as shown in Fig. 19. With the increase of delay time, the current difference inside a module is reduced, as shown in Fig. 20(b) and (c). Besides, the waveform of the current in interconnection inductor L_s is shown in Fig. 21. By cancelling i_5 with i_{comp} , the actual current flows through the inductor L_s is fairly small.

Figs. 22 and 23 show the experimental waveform under light

Fig. 17. Winding structure of the PCB planar transformer.

Fig. 18. 2-D finite-element simulation. (a) Coupled. (b) Separate transformer.

(a)

(b)

Fig. 19. Experimental waveforms for current sharing between modules. (a) Without and (b) with interconnected resonant tank.

(a)

(b)

(c)

Fig. 20. Experimental waveforms for current sharing inside a module. (a) No delay time. (b) 20 ns delay. (c) 40 ns delay.

load and full load condition respectively. The oscillation under light load is attributed to the parasitic capacitor introduced by the planar transformer, which, however does not influence the normal operation of the converter and will disappear under heavy load. The results show that ZVS is achieved for primary devices of both two modules simultaneously under both light and heavy load conditions.

The measured efficiency curve of the proposed IPOS

LLC converter is shown in Fig. 24, where the efficiency of a single LLC converter is also plotted. The output power of two converters are normalized for comparison. The proposed IPOS LLC converter has a peak efficiency of 97.5% at 2.4 kW load and a full-load efficiency of 96.5%. It is obvious that the efficiency of the proposed IPOS LLC converter is higher than that of a single LLC converter over the load range, due to the optimized coupled transformer. With rated 100 V input

Fig. 21. Waveform of current in L_s .

Fig. 22. Waveform under 1.2 kW (light load).

Fig. 23. Waveform under rated 3.6 kW (full load).

Fig. 24. Efficiency curves.

and 1.2 kV output voltage, high gain and high efficiency are achieved for the IPOS LLC converter.

Fig. 25. Loss breakdown under 3.6 kW load.

Based on the experimental results, the loss breakdown of the prototype under 3.6 kW full-load condition is analyzed by theoretical calculation, as shown in Fig. 25. It is evident that the core loss of transformer is relatively small and the major loss is distributed at the conduction loss and turn-off loss of the power devices, which may be further improved by choosing devices with lower on-resistance and better performance on high frequency applications such as SiC and GaN.

VI. CONCLUSION

An IPOS LLC converter is proposed in this paper to deal with the high voltage gain and high efficiency applications. A high efficiency is obtained by utilizing a coupled transformer to implement the magnetic flux cancellation, and the size and power losses of the magnetic core are reduced and the output voltage is naturally balanced due to the common magnetic flux that flows through different windings. A bridge wire that interconnects resonant tanks is proposed to balance the current between modules. Furthermore, the current inside one module is balanced by the coordination of compensation inductor and delay control. The proposed structure has been verified on a 200 kHz 100 V/1.2 kV 3.6 kW prototype. Experimental results show that the converter has a good performance on current sharing and can achieve high efficiency up to 97.5%.

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