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Gain Modeling and Parameter Optimization of Symmetric Bidirectional CLLC Resonant Converters in LVDC Distribution Systems

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Abstract—Low-voltage DC (LVDC) distribution systems are increasingly adopted in data centers, electric vehicle charging, and industrial power supply due to their ability to reduce multi-stage energy conversion, enhance efficiency, and integrate distributed energy sources. At their core, high-performance DC-DC converters enable efficient power transfer and provide electrical isolation across different voltage levels. Beyond these fundamental roles, they regulate voltage, improve power quality, and enhance system reliability under varying load and source conditions. Their versatility makes them indispensable in applications ranging from renewable energy integration to low-voltage DC distribution networks. Among various topologies, the CLLC resonant converter, with its high efficiency, wide operating range, and inherent bidirectional power flow capability, is particularly suitable for isolated energy conversion and system-level energy management in LVDC applications. However, in the process of analyzing and designing the CLLC, the conventional fundamental harmonic approximation (FHA) method exhibits significant discrepancies when analyzing the actual operating behavior of the CLLC converter. This paper first introduces the basic topology of the CLLC resonant converter and subsequently demonstrates the limitations of the FHA-based analysis. A time-domain analytical (TDA) model investigates the high-efficiency operating modes and derives their gain expressions from intrinsic time-domain characteristics. Boundary conditions further constrain the operation to avoid transition into loss-dominant modes. Furthermore, we designed and implemented a 135W laboratory prototype, which ensures reliable soft-switching operation across the intended load range and achieves a peak efficiency of 98.1%.

Index Terms—CLLC Resonant Converter, TDA, Gain Formula, Boundary Analysis, Optimization Design

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I. INTRODUCTION

Low-Voltage DC (LVDC) distribution systems demonstrate substantial advantages in energy efficiency, structural simplicity, and the integration of distributed resources, making them highly attractive for applications such as data centers, residential buildings, and electric transportation. Compared with conventional AC-based architectures, LVDC networks minimize conversion stages and reduce transmission losses, which leads to improved overall efficiency. Within such systems, DC-DC converters operate as the primary interface to provide voltage matching, galvanic isolation, and energy management. As renewable generation and energy storage continue to expand, power transfer increasingly follows a bidirectional pattern rather than a single direction. To accommodate dynamic load requirements and variable grid conditions, bidirectional DC-DC converters emerge as indispensable units in LVDC networks, enabling two-way power transfer, achieving voltage regulation, and optimizing energy utilization to strengthen system reliability and operational flexibility [1]-[2].

Conventional DC-DC converter topologies exhibit distinct limitations. For instance, the Buck-Boost converter supports only unidirectional energy flow. Connecting two converters in parallel enables bidirectional operation; however, this approach results in increased cost, a larger volume, and reduced integration feasibility. The bidirectional half-bridge converter can realize zero-voltage switching (ZVS) through phase-shift control, yet its narrow phase-shift range significantly constrains its application scenarios [3]. In the case of the LC resonant bidirectional full-bridge converter, the voltage gain remains below unity due to the series configuration of resonant inductance, resonant capacitance, and load, thereby limiting its operation to step-down mode. The LLC resonant bidirectional full-bridge converter exhibits favorable light-load performance and achieves ZVS across the entire power range [4]-[7]. However, its structural asymmetry results in inconsistent behavior during forward and reverse operations. Specifically, during reverse operation, the

converter behaves as an LC resonant circuit, which prevents it from operating in boost mode [8].

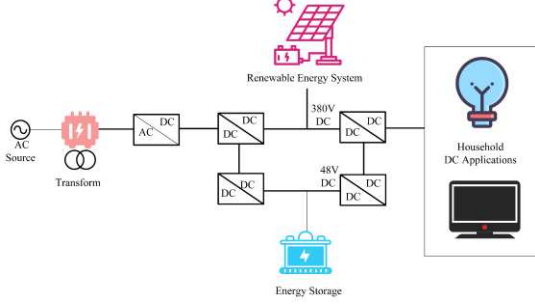


Fig.1. Typical Applications of LVDC Distribution Systems.

As shown in Fig. 1, LVDC networks are typically employed in applications where reliable bidirectional power conversion plays a critical role. Within this context, the full-bridge isolated bidirectional CLLC resonant converter enables bidirectional power transfer and supports a wide range of output voltages and currents. By adjusting the resonant frequency or duty cycle, the output can be flexibly regulated over a broad range while achieving ZVS or zero-current switching (ZCS) for the power devices. Such control significantly reduces switching losses and electromagnetic interference (EMI), thereby enhancing overall efficiency and reliability. Many researchers have focused on this topology due to its advantages, leading to extensive investigations [9]–[13].

The parameter design of the resonant components in symmetric CLLC bidirectional DC–DC converters is critical to ensure proper operation. Such a design strongly depends on the accurate modelling of the converter. The conventional modelling method is the fundamental harmonic approximation (FHA), which simplifies the design process by reducing the complexity of the system. Previous work introduced a parameter design method using the voltage gain characteristics derived from FHA[14]. Further investigated the gain characteristics and resistive boundary conditions using the FHA-based model[15]. However, these methods only consider the fundamental harmonic, neglecting higher-order harmonics, which results in inaccurate gain predictions and an inability to distinguish between different operating modes.

In contrast to FHA, the time-domain analytical (TDA) models the converter by solving the differential equations under various operating modes. The derived modal definitions not only provide a theoretical basis for understanding the diversity of operating states but also enable accurate prediction of converter behavior and key performance characteristics. [16]–[18] provide a systematic characterization of the time-domain operating modes of the LLC resonant converter, owing to the close topological resemblance and resonant energy-transfer commonality between LLC and CLLC converters. these time-domain modes extend directly to the operating-state analysis of the CLLC DC–DC converter, thereby establishing a theoretical foundation for subsequent mechanism studies and control design[19]–[23]. [24] assumed an infinite magnetizing inductance, which inevitably neglects the magnetic field establishment process required for energy transfer. [25]

presented an analysis of the CLLC gain and load operating regions, laying the groundwork for achieving a wide voltage gain. However, its gain formula is derived using the FHA, which exhibits significant errors as the switching frequency gradually deviates from the resonant frequency.

Given the above limitations, this paper first verifies the limited accuracy of FHA. Then, based on a time-domain model, a more precise gain expression is derived by leveraging the symmetry and continuity of the waveform. An optimization method for the resonant parameters of the CLLC converter is proposed based on this gain formulation. The proposed approach is validated through simulation, followed by hardware prototyping to verify the effectiveness of the optimized resonant parameters.

II. FHA MODELING AND LIMITATION ANALYSIS

Fig.2 illustrates the topology of a symmetric bidirectional CLLC resonant DC–DC converter. In this circuit, S_1 – S_4 are the primary-side switches, operated in complementary pairs with a fixed duty cycle of 50%. S_5 – S_8 are the secondary-side switches, which are not actively driven but operate under a synchronous rectification strategy. L_{r1} and C_{r1} represent the primary resonant inductor and capacitor, while L_{r2} and C_{r2} denote the corresponding secondary-side resonant elements. The magnetizing inductance of the transformer is denoted by L_m , and n represents the transformer turns ratio. Each switch incorporates a parasitic body diode (D_1 – D_8) and output capacitance (C_{oss1} – C_{oss8}).

To ensure symmetric operation during both forward and reverse power flow, the circuit must satisfy the condition in (1).

$$\begin{cases} L_{r1} = n^2 L_{r2} \\ C_{r1} = \frac{C_{r2}}{n^2} \end{cases} \quad (1)$$

Fig.3 shows the equivalent circuit of the CLLC resonant converter derived using the FHA. The variables u_{AB} and u_{CD} represent the fundamental components of the voltages at ports AB and CD in the network shown in Fig.2. R_{eq} denotes the equivalent output resistance, which can be obtained via Fourier transform.

$$R_{eq} = \frac{8}{\pi^2} R_o. \quad (2)$$

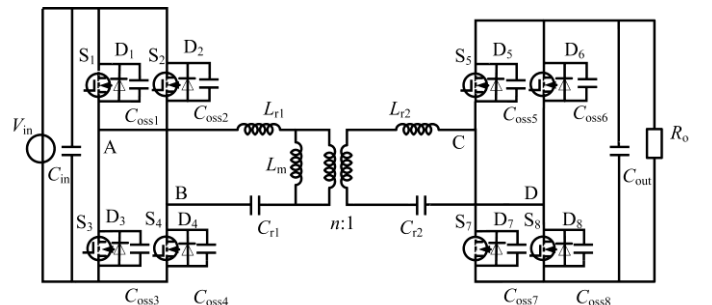


Fig.2. Symmetrical CLLC Bidirectional DC–DC Converter.

Based on the equivalent circuit shown in Fig.3 and the voltage division principle, the First Harmonic Approximation yields the voltage gain of the CLLC resonant converter as:

$$M = \frac{u_{CD}}{u_{AB}} = \frac{1}{\sqrt{\left(1 + \frac{1}{k} - \frac{1}{kf_n^2}\right)^2 + \frac{Q^2}{k^2} \left[(2k+1)f_n - \frac{2k+2}{f_n} + \frac{1}{f_n^3}\right]^2}} \quad (3)$$

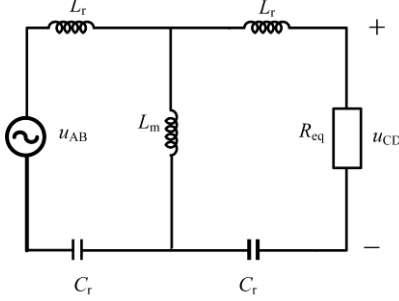


Fig.3. The Equivalent Circuit of the CLLC Resonant Converter Based on the FHA.

Where k denotes the ratio of the magnetizing inductance to the primary-side resonant inductance, Q is the quality factor, and f_n represents the normalized switching frequency. These parameters are defined as follows:

$$\begin{cases} f_n = \frac{f_s}{f_r} \\ f_r = \frac{1}{2\pi\sqrt{L_{r1}C_{r1}}} \\ k = \frac{L_m}{L_{r1}} \\ Q = \frac{Z_r}{R_{eq}} \\ Z_r = \sqrt{\frac{L_{r1}}{C_{r1}}} \end{cases} \quad (4)$$

Where f_s denotes the switching frequency, f_r is the resonant frequency, and Z_r represents the characteristic impedance.

As illustrated in Fig.4, a noticeable discrepancy exists between the conclusions derived from the FHA analysis and the actual simulation results, suggesting that the accuracy of FHA is difficult to guarantee. This deviation arises because FHA neglects the impact of higher-order harmonics on circuit performance. Therefore, FHA fails to accurately describe the full-frequency dynamic response and actual characteristics.

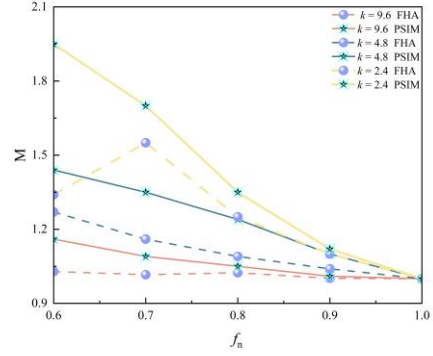


Fig.4. Comparison of the gain characteristics of the CLLC resonant converter obtained by FHA and circuit simulation.

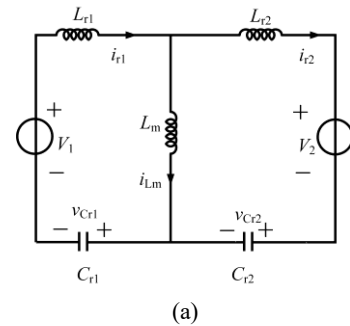
III. MODAL ANALYSIS OF CLLC RESONANT CONVERTERS IN TDA

When performing modal analysis of CLLC resonant converters, TDA can provide a more accurate characterization of the complex dynamic behaviors exhibited by the converter. Compared to the FHA, TDA directly captures the time-domain response of the system. Particularly in high-frequency or fast-transient scenarios, TDA provides significantly more precise representation of dynamic performance.

The CLLC resonant converter employs a 50% duty cycle pulse frequency modulation scheme with complementary switching signals for the primary-side full-bridge. To simplify the analysis, the study examines the first half-cycle, with S_1 and S_4 conducting and S_2 and S_3 turned off. The operational modes of the CLLC resonant converter are classified based on the conduction states of the rectifier-side diodes:

- 1) **P mode**
During positive conduction, diodes D_5 and D_8 conduct while D_6 and D_7 remain off.
- 2) **N mode**
During negative conduction, diodes D_6 and D_7 conduct while D_5 and D_8 remain off.
- 3) **O mode**
During open circuit, all diodes D_5 and D_8 are off.

Fig.5 illustrates the equivalent circuits for the three fundamental operating modes. Both theoretical analysis and simulation verification demonstrate that eight distinct operating modes can achieve ZVS during the first half-cycle through different sequential combinations of the modes: P, PO, NP, PON, PN, NOP, OPO, and O.



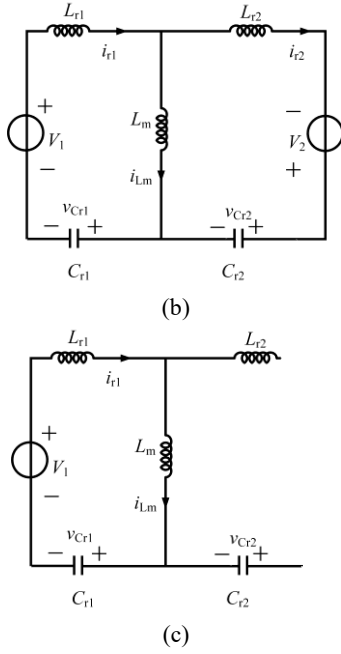


Fig.5. Time-domain equivalent circuit model of the CLLC resonant converter. (a) P mode. (b) N mode. (c) O mode.

When the switching frequency equals the resonant frequency, the converter operates in P mode. At the beginning of P mode, the primary resonant current i_{r1} equals the magnetizing current i_{Lm} . The two currents then diverge and converge again, enabling ZCS of the converter.

When the switching frequency falls below the resonant frequency, the converter enters PO mode. Under this condition, the current evolution during the P stage matches that in P mode, after which the O stage starts, where i_{r2} persistently zero. This condition creates a temporary decoupling between primary and secondary sides of the transformer, establishing the necessary conditions for ZCS realization.

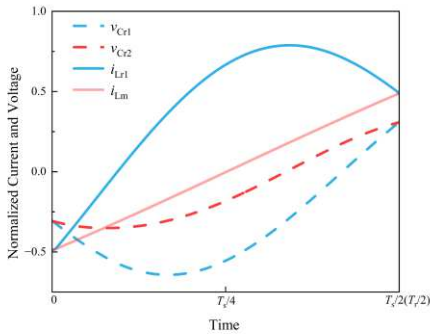


Fig.6. Operating Waveforms under P mode.

Applying Kirchhoff's laws yields the state equations for the equivalent circuits under different operating modes. (5) presents the loop equation corresponding to the circuit in Fig.5-(a).

$$\begin{cases} V_1 - L_{r1} \frac{di_{Lr1}}{dt} - L_m \frac{di_{Lm}}{dt} - v_{Cr1} = 0 \\ V_2 + L_{r2} \frac{di_{Lr2}}{dt} - L_m \frac{di_{Lm}}{dt} + v_{Cr2} = 0 \\ i_{r1}(t) = C_{r1} \frac{dv_{Cr1}}{dt} \\ i_{r2}(t) = C_{r2} \frac{dv_{Cr2}}{dt} \\ i_{Lm}(t) = i_{r1}(t) - i_{r2}(t) \end{cases} \quad (5)$$

$i_{Lr1}, i_{Lr2}, v_{Cr1},$ and v_{Cr2} are chosen as the state variables. (5) is normalized using base reference values.

$$\begin{cases} V_{base} = V_2 \\ Z_{base} = Z_r \\ I_{base} = V_2 / Z_r \end{cases} \quad (6)$$

Substituting (6) into (5) yields the state-space equations for P mode operation:

$$\begin{cases} i_{Lr1P}(t) = I_{1P} \sin(2\pi f_r t + \phi_{1P}) + I_{2P} \sin(2k_1 \pi f_r t + \phi_{2P}) \\ i_{Lr2P}(t) = I_{1P} \sin(2\pi f_r t + \phi_{1P}) - I_{2P} \sin(2k_1 \pi f_r t + \phi_{2P}) \\ v_{Cr1P}(t) = -Z_r I_{1P} \cos(2\pi f_r t + \phi_{1P}) - \\ \quad I_{2P} \frac{Z_r}{k_1} \cos(2k_1 \pi f_r t + \phi_{2P}) + V_1 \\ v_{Cr2P}(t) = -Z_r I_{1P} \cos(2\pi f_r t + \phi_{1P}) + \\ \quad I_{2P} \frac{Z_r}{k_1} \cos(2k_1 \pi f_r t + \phi_{2P}) + V_2 \end{cases} \quad (7)$$

Where $k_1 = \sqrt{1/(1+2k)}$, M represents the voltage gain, $I_{1P}, I_{2P}, \phi_{1P}, \phi_{2P}$ are unknown variables.

The NP mode and PON mode are the two operating modes most susceptible to power losses, which transition from P mode and PO mode, respectively.

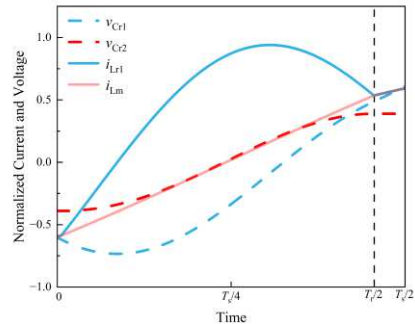


Fig.7. Operating Waveforms under PO mode.

As the switching frequency gradually increases beyond the resonant frequency, the resonant converter operates in the NP mode. At the beginning of this mode, $i_{r1} < i_{Lm}$, and the N stage

ends when the two currents become equal. Since the switching period is shorter than the resonant period, i_{r1} and i_{Lm} remain unequal at the end of the P stage, thereby preventing the realization of ZCS.

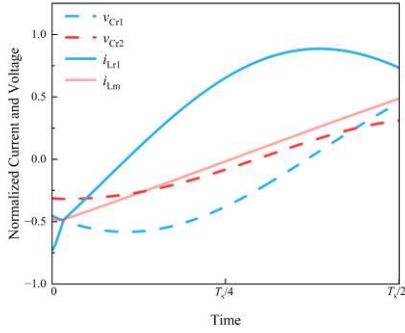


Fig.8. Operating Waveforms under NP mode.

One can derive the state-space equations for N mode through analysis similar to (5) and (6) using the equivalent circuit shown in Fig.5-(b).

Under PO mode operation, continuous reduction of the switching frequency will cause the converter to transition into PON mode. As evidenced by Fig.9, the first two stages of PON mode remain essentially identical to those in PO mode. The emergence of N mode reduces the resonant current. Prior to the switching state transition, premature conduction of the secondary-side diode causes the primary-side current to deviate, which hinders the realization of ZVS.

$$\begin{cases} i_{Lr1N}(t) = I_{1N}\sin(2\pi f_r t + \varphi_{1N}) + I_{2N}\sin(2k_1\pi f_r t + \varphi_{2N}) \\ i_{Lr2N}(t) = I_{1N}\sin(2\pi f_r t + \varphi_{1N}) - I_{2N}\sin(2k_1\pi f_r t + \varphi_{2N}) \\ v_{Cr1N}(t) = -Z_r I_{1N}\cos(2\pi f_r t + \varphi_{1N}) - \\ \quad I_{2N} \frac{Z_r}{k_1} \cos(2k_1\pi f_r t + \varphi_{2N}) + V_1 \\ v_{Cr2N}(t) = -Z_r I_{1N}\cos(2\pi f_r t + \varphi_{1N}) + \\ \quad I_{2N} \frac{Z_r}{k_1} \cos(2k_1\pi f_r t + \varphi_{2N}) + V_2 \end{cases} \quad (8)$$

Where I_{1N} , I_{2N} , φ_{1N} , φ_{2N} represent the unknown variables to be solved.

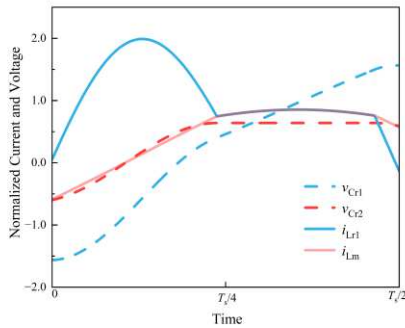


Fig.9. Operating Waveforms under PON mode.

One can derive the state-space equations for O mode from the equivalent circuit in Fig.5-(c) by following a similar analytical procedure to that used for N mode.

$$\begin{cases} i_{Lr1O}(t) = I_{1O}\sin(2k_2\pi f_r t + \varphi_{1O}) \\ v_{Cr1O}(t) = -I_{1O} \frac{Z_r}{k_2} \cos(2k_2\pi f_r t + \varphi_{1O}) + V_1 \end{cases} \quad (9)$$

Where $k_2 = \sqrt{1/(1+k)}$, I_{1O} , φ_{1O} are the unknown variables to be determined.

The following briefly introduces other operating modes:

1) **PN mode**

as the transferred power increases, the resonant converter may enter the PN mode, which, similar to the NP mode, suffers from relatively high-power loss.

2) **NOP mode**

In the NP mode, as the transferred power decreases, the converter may shift to the NOP mode. This mode operates at a frequency consistently higher than the resonant frequency, which results in reduced power transmission efficiency.

3) **OPO mode and O mode**

There are correspond to the light-load and no-load operating conditions of the resonant converter, respectively.

The solution of the mode equations reveals that the experimental results obtained via TDA are in close agreement with the simulation outcomes, which demonstrates the high fidelity of TDA.

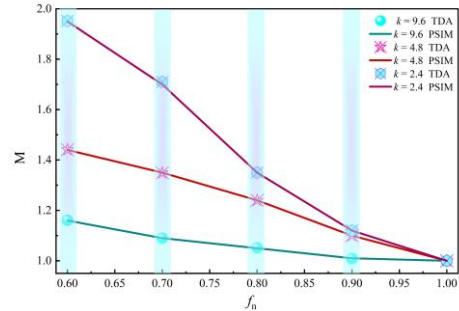


Fig.10. Comparison of the gain characteristics of the CLLC resonant converter obtained by TDA and circuit simulation.

IV. OPTIMIZATION DESIGN CONSIDERATIONS FOR CLLC RESONANT CONVERTERS

In CLLC resonant converter design, parameter selection is critical for ensuring proper operation. The ultimate optimization objective is to achieve soft switching and enhance overall efficiency. Precise realization of ZVS in the designed operating mode shifts the primary power loss mechanism toward conduction losses. In CLLC converters, the characteristics of the primary-side current play a key role in determining the losses. The design of resonant parameters directly influences the Current waveform and peak magnitude and the Current distribution. Therefore, precise optimization

of these three key parameters (L_{r1} , C_{r1} , and L_m) enables the CLLC converter to achieve maximum energy conversion efficiency during target operating modes.

A. ZVS Operation

Fig.11 illustrates the ZVS operational waveforms of the CLLC resonant converter, displaying from top to bottom: Gate drive voltages u_{GS} of switches S_1 - S_4 , Primary-side resonant current and magnetizing current, Switch voltage u_s waveforms.

At the first dead time, t_{d1} , when S_2 and S_3 turn off while S_1 and S_4 remain off, all primary-side switches are in the off state. During this interval, i_{r1} equals i_{Lm} with the current direction unchanged, and the converter enters freewheeling mode. The resonant current charges the parasitic capacitors of S_2/S_3 while discharging those of S_1/S_4 until their voltages reach zero. This enables near-zero-voltage switching and effectively reduces switching losses. To ensure stable ZVS achievement, sufficient i_{Lm} must be maintained during t_d to fully discharge the MOSFETs' parasitic capacitance, requiring satisfaction of [26]:

$$t_d \geq 8f_{smax} L_m C_{oss(max)}. \quad (10)$$

Where f_{smax} denotes the maximum operating frequency, $C_{oss(max)}$ represents the maximum value of the MOSFET's output capacitance. In typical design approaches, the dead time t_d for CLLC resonant converters typically requires to set an initial value. Consequently, we can reformulate (10) as:

$$L_m \leq \frac{t_d}{8f_{smax} C_{oss(max)}}. \quad (11)$$

Fig.11. Operating waveforms of the CLLC resonant converter illustrating ZVS.

B. Conduction Losses in CLLC Resonant Converters

Because the CLLC resonant converter inherently employs soft-switching techniques that significantly reduce switching losses, the conduction loss becomes the dominant factor. This section develops a precise loss model by incorporating equivalent loop resistances on both primary and secondary sides through time-domain analysis, investigating their formation mechanisms and influencing factors to guide parameter optimization for enhanced system performance.

The conduction losses originate from the resistive losses in the power switches and the resonant circuit of the converter.

The primary-side conduction losses are mainly caused by the MOSFET on-state resistance, $R_{ds(on)}$, and the primary winding resistance of the transformer, R_{T1} . The secondary-side conduction losses arise from the equivalent load resistance, R_o , and the secondary winding resistance, R_{T2} , of the transformer. By employing the equivalent circuit analysis method, we derive the equivalent resistance of the primary side:

$$R_1 = R_{ds(on)} + R_{T1} + \frac{1}{n^2} (R_o + R_{T2}). \quad (12)$$

The values of $R_{ds(on)}$, R_{T1} and R_{T2} are closely related to the selected components. The following equation defines the conduction loss:

$$P_{cond} = I_{rms}^2 R_1. \quad (13)$$

Where P_{cond} represents the conduction loss, I_{rms} denotes the root-mean-square value of the primary current.

Under ideal conditions—assuming a mutual coupling coefficient of 1, a constant load that is significantly larger than the resonant impedance, the total current can be expressed as:

$$\begin{cases} I_{r1}^2 = \frac{I_{Lm}^2}{k^2} \\ I_{rms}^2 = I_{Lm}^2 + I_{r1}^2 = I_{Lm}^2 (1 + 1/k^2) \end{cases}. \quad (14)$$

As shown in the above expression, when the inductance ratio k increases, I_{rms} decreases significantly, thereby reducing conduction losses. However, when the ideal assumptions do not hold, the relationship between the primary-side current and the magnetizing current cannot be derived simply using k . Based on the time-domain model, the relationship between k and I_{rms} is analyzed. As clearly illustrated in Fig.12, under various normalized frequency conditions, I_{rms} consistently decreases with increasing k . Accordingly, the design process favors a relatively large value of k .

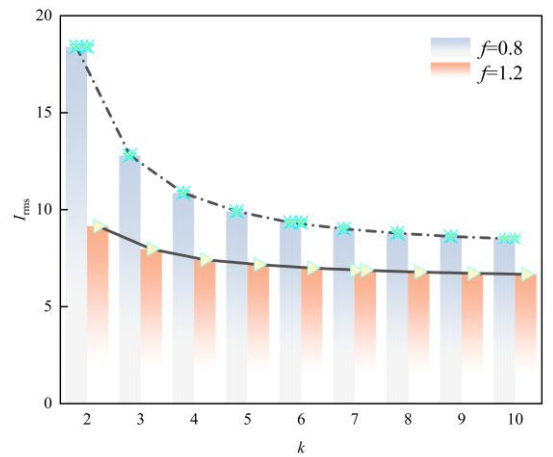


Fig.12. Variation of I_{rms} with k at Different Normalized Switching Frequencies f_n .

C. Voltage Gain Analysis of CLLC Resonant Converters

Voltage gain analysis is of paramount importance in the CLLC resonant converter, as it defines the feasible operating

frequency range under different input and load conditions, guarantees stable output regulation, and establishes the basis for achieving soft switching, which in turn impacts overall efficiency and switching losses.

From the above analysis, it can be concluded that the PO mode enables precise realization of soft switching. In the PO operating mode, the gain expression must be derived based on current continuity and half-cycle symmetry, in conjunction with (7) and (9). Subsequently, the continuity and half-cycle symmetry conditions associated with the PO mode are presented.

$$v_{Cr1P}\left(\frac{T_r}{2}\right) = v_{Cr1O}(0). \quad (15)$$

$$v_{Cr2P}\left(\frac{T_r}{2}\right) = v_{Cr2O}(0). \quad (16)$$

$$i_{Lr1P}\left(\frac{T_r}{2}\right) = i_{Lr1O}(0). \quad (17)$$

$$i_{Lr2P}\left(\frac{T_r}{2}\right) = i_{Lr2O}(0). \quad (18)$$

$$v_{Cr1P}(0) = -v_{Cr1O}\left(\frac{T_s}{2}\right). \quad (19)$$

$$i_{Lr1P}(0) = -i_{Lr1O}\left(\frac{T_s}{2}\right). \quad (20)$$

Equation (15)-(18) characterizes the continuity condition, whereas (19)-(20) illustrates the half-cycle symmetry inherent in PO-mode operation. Where T_r and T_s denote the resonant and switching periods, respectively, in the O phase, the secondary side is equivalent to an open circuit, during which the resonant current i_{r2} and resonant capacitor voltage v_{Cr2} remain unchanged and exhibit symmetry only within the P phase. It is particularly noteworthy that, in the PO mode, the time point $T_r/2$ marks both the end of the P phase and the beginning of the O phase.

Initially, solve the simultaneous equations formed by (15) and (19), (17) and (20) together with (7) and (9), results in:

$$Z_r I_{1P} \cos \varphi_{1P} - I_{2P} \frac{Z_r}{k_1} \cos(k_1 \pi + \varphi_{2P}) = -I_{1O} \frac{Z_r}{k_2} \cos(k_2 \pi + \varphi_{1O}). \quad (21)$$

$$I_{1O} \sin(k_2 \pi + \varphi_{1O}) = -I_{1P} \sin \varphi_{1P} + I_{2P} \sin(k_1 \pi + \varphi_{2P}). \quad (22)$$

$$Z_r I_{1P} \cos \varphi_{1P} = 2V_1 - I_{1O} \frac{Z_r}{k_2} \cos\left(\frac{k_2 \pi}{f_n} + \varphi_{1O}\right) - I_{2P} \frac{Z_r}{k_1} \cos \varphi_{2P}. \quad (23)$$

$$I_{2P} = \frac{I_{1O} \sin(k_2 \pi + \varphi_{1O})}{2 \sin\left(\frac{k_1 \pi}{2}\right)}. \quad (24)$$

$$I_{1P} \sin \varphi_{1P} + I_{2P} \sin \varphi_{2P} = -I_{1O} \sin\left(\frac{k_2 \pi}{f_n} + \varphi_{1O}\right). \quad (25)$$

Solving (21) and (23) simultaneously gives:

$$\begin{aligned} 2V_1 - I_{1O} \frac{Z_r}{k_2} \cos\left(\frac{k_2 \pi}{f_n} + \varphi_{1O}\right) - I_{2P} \frac{Z_r}{k_1} \cos\left(\frac{k_1 \pi}{2}\right) = \\ I_{2P} \frac{Z_r}{k_1} \cos \frac{k_1 \pi}{2} - I_{1O} \frac{Z_r}{k_2} \cos(k_2 \pi + \varphi_{1O}). \end{aligned} \quad (26)$$

During the P mode the current of the magnetizing inductance can be expressed as:

$$i_{LmP}(t) = i_{r1P}(t) - i_{r2P}(t) = 2I_{2P} \sin(2k_1 \pi f_r t + \varphi_{2P}). \quad (27)$$

During the O mode, since the secondary side is open-circuited, the magnetizing inductance current equals the primary-side resonant inductance current:

$$i_{LmO}(t) = i_{Lr1O}(t). \quad (28)$$

The continuity condition of the magnetizing current still holds, leading to:

$$i_{LmP}\left(\frac{T_r}{2}\right) = i_{Lr1O}(0). \quad (29)$$

Substituting (9) and (23) into (25) yields:

$$2I_{2P} \sin(k_1 \pi + \varphi_{2P}) = I_{1O} \sin(k_2 \pi + \varphi_{1O}). \quad (30)$$

(22) and (26) can be solved simultaneously to yield:

$$I_{1P} \sin \varphi_{1P} = -I_{2P} \sin(k_1 \pi + \varphi_{2P}). \quad (31)$$

Since the secondary side is open-circuited during the O phase, the symmetry of the secondary resonant current is only exhibited during the P phase:

$$i_{Lr2P}(0) = -i_{Lr2P}\left(\frac{T_r}{2}\right) = 0. \quad (32)$$

Substituting (28) into (7) yields:

$$I_{1P} \sin \varphi_{1P} - I_{2P} \sin \varphi_{2P} = -I_{1P} \sin(\pi + \varphi_{1P}) + I_{2P} \sin(k_1 \pi + \varphi_{2P}) = 0. \quad (33)$$

By analyzing (29), it can be concluded that:

$$\begin{cases} I_{1P} \sin \varphi_{1P} = I_{2P} \sin \varphi_{2P} \\ I_{1P} \sin \varphi_{1P} = -I_{2P} \sin(k_1 \pi + \varphi_{2P}) \\ \varphi_{2P} = -\frac{k_1 \pi}{2} \end{cases} \quad (34)$$

By combining (33) with (22) and (24), we obtain:

$$2I_{2P} \sin(k_1 \pi + \varphi_{2P}) = I_{1O} \sin(k_2 \pi + \varphi_{1O}). \quad (35)$$

$$2I_{2P} \sin \varphi_{2P} = -I_{1O} \sin\left(\frac{k_2 \pi}{f_n} + \varphi_{1O}\right). \quad (36)$$

$$I_{1O} \sin\left(\frac{k_2 \pi}{f_n} + \varphi_{1O}\right) = I_{1O} \sin(k_2 \pi + \varphi_{1O}). \quad (37)$$

Based on the inherent properties of trigonometric functions:

$$\frac{k_2 \pi}{f_n} + \varphi_{1O} + k_2 \pi + \varphi_{1O} = \pi. \quad (38)$$

$$\varphi_{1O} = \frac{1}{2}(\pi - k_2 \pi - \frac{k_2 \pi}{f_n}). \quad (39)$$

The voltage across the secondary resonant capacitor varies during the P phase and remains constant during the O phase. Its half-cycle symmetry can be expressed as:

$$v_{Cr2}(0) = -v_{Cr2}\left(\frac{T_r}{2}\right). \quad (40)$$

By combining (7) with (40), we obtain:

$$\begin{aligned} -Z_r I_{1P} \cos \varphi_{1P} + I_{2P} \frac{Z_r}{k_1} \cos \varphi_{2P} - V_2 = \\ Z_r I_{1P} \cos(\pi + \varphi_{1P}) - I_{2P} \frac{Z_r}{k_1} \cos(k_1 \pi + \varphi_{2P}) + V_2. \end{aligned} \quad (41)$$

By applying trigonometric identities along with (33), we

obtain the expression for V_2 as follows:

$$V_2 = I_{2P} \frac{Z_r}{k_1} \cos \frac{k_1 \pi}{2}. \quad (42)$$

Combining (26), (39) and (42) yields the following result:

$$V_1 - V_2 = I_{10} \frac{Z_r}{k_2} \sin \left(\frac{k_2 \pi}{2} - \frac{k_2 \pi}{2f_n} \right). \quad (43)$$

Solving (34) and (35) simultaneously yields the expression for I_{2P} as follows:

$$I_{2P} = \frac{I_{10} \sin(k_2 \pi + \phi_{10})}{2 \sin(\frac{k_1 \pi}{2})}. \quad (44)$$

Substituting (44) into (42) yields:

$$I_{10} = \frac{2V_2 k_1 \tan(\frac{k_1 \pi}{2})}{Z_r \cos(\frac{k_2 \pi}{2} - \frac{k_2 \pi}{2f_n})}. \quad (45)$$

Substituting (45) into (43) yields:

$$V_1 = [1 + \frac{2k_1 \tan(\frac{k_1 \pi}{2}) \tan(\frac{k_2 \pi}{2} - \frac{k_2 \pi}{2f_n})}{k_2}] V_2. \quad (46)$$

We reorganize the expressions of k_1 and k_2 in terms of k and substitute them into the gain expression to derive the voltage gain for the PO mode:

$$M_{PO} = \frac{1}{1 + \frac{2\sqrt{k+1}}{\sqrt{2k+1}} \tan(\frac{\pi}{2\sqrt{2k+1}}) \tan[\frac{\pi}{2\sqrt{k+1}}(1 - \frac{1}{f_n})]}. \quad (47)$$

In designing a CLLC resonant converter, researchers typically begin by specifying the input and output voltages, thereby establishing the foundation for resonant parameter matching, ensuring soft-switching conditions, and further improving system efficiency. With the voltage gain predetermined, they can then identify the k_{max} from equation (47) by selecting an appropriate range of normalized frequencies.

D. Boundary Condition Analysis

As illustrated in Fig.9, the voltage gain in PON mode initially increases and subsequently decreases, which, as previously discussed, adversely affects closed-loop stability. Consequently, the converter should maintain PO mode operation while avoiding PON mode transitions. The boundary between these operating modes arises from the PO mode. A continuous decrease in switching frequency or load current leads to premature diode conduction and the loss of O mode operation, causing a decoupling between the primary current and the magnetizing current, and ultimately resulting in current roll-off. Consequently, the magnetizing voltage progressively decreases and eventually stabilizes at a negative value, V_2 . To maintain the PO mode boundary condition, the following must be satisfied:

$$v_{LmO}(\frac{T_s}{2}) - v_{Cr2P}(\frac{T_r}{2}) \geq -V_2. \quad (48)$$

In the equivalent circuit, $v_{LmO}(\frac{T_s}{2})$ can be expressed as:

$$v_{LmO}(\frac{T_s}{2}) = V_1 - v_{Lr1O}(\frac{T_s}{2}) - v_{Cr1O}(\frac{T_s}{2}). \quad (49)$$

Substituting (49) into (48) yields:

$$V_1 + V_2 > v_{Cr1O}(\frac{T_s}{2}) + v_{Cr2O}(\frac{T_s}{2}) = v_{Cr1P}(\frac{T_r}{2}) + v_{Cr2P}(\frac{T_r}{2}) + \frac{i_{Lmp}(\frac{T_r}{2})}{C_r} (\frac{T_s}{2} - \frac{T_r}{2}). \quad (50)$$

Based on charge conservation and half-cycle symmetry, we obtain:

$$v_{Cr1O}(\frac{T_s}{2}) = -v_{Cr1P}(0) = v_{Cr1P}(0) + \frac{P_{out} T_s}{2C_r V_2} + \frac{i_{Lmp}(\frac{T_r}{2})}{C_r} (\frac{T_s}{2} - \frac{T_r}{2}). \quad (51)$$

$$v_{Cr2O}(\frac{T_s}{2}) = -v_{Cr2P}(0) = v_{Cr2P}(0) + \frac{P_{out} T_s}{2C_r V_2}. \quad (52)$$

By combining (50), (51), and (52), we derive:

$$Z_r < \frac{2V_1 V_2 f_n}{\pi P_{out}}. \quad (53)$$

where P_{out} represents the output power, elimination of V_2 from the equation allows the forward and reverse operating states to be expressed as:

$$\begin{cases} Z_{r1} < \frac{2V_1^2 M_{PO} f_n}{n\pi P_{out}} \\ Z_{r2} < \frac{2V_1^2 f_n}{n\pi P_{out} M_{PO}} \end{cases}. \quad (54)$$

Z_{r1} and Z_{r2} represent the impedance constraint ranges for forward and reverse operation, respectively.

V. NETWORK PARAMETER DESIGN FOR CLLC RESONANT CONVERTERS DESIGN SPECIFICATIONS

A. Converter Design Specifications

To verify the feasibility of the proposed approach, a prototype converter was developed with specifications as shown in Table I.

TABLE I
CLLC RESONANT CONVERTER DESIGN SPECIFICATIONS

Parameter	Value
Rated Input Voltage	72V
Rated Output Voltage	48V
Output Voltage Range	42-54V
Resonant Frequency	106kHz
Rated Output Power	135W

The transformer turns ratio (n) is critically important as it directly affects voltage gain, resonant characteristics, and efficiency. Designing the rated operating point at the resonant frequency maximizes efficiency, reduces switching losses, and simplifies the matching of resonant tank parameters, ensuring optimal full-load performance. At the resonant frequency, the voltage gains equal 1, yielding a turns ratio $n = 1.5$.

B. Selection of Inductance Ratio k

The maximum and minimum voltage gain range for forward operation can be determined based on the transformer's input-output voltage requirements:

$$\begin{cases} M_{\max 1} = \frac{nU_{0\max}}{U_{\text{in}}} = 1.125 \\ M_{\min 1} = \frac{nU_{0\min}}{U_{\text{in}}} = 0.875 \end{cases} \quad (55)$$

Similarly, the maximum and minimum voltage gain ranges for reverse operation are derived as follows:

$$\begin{cases} M_{\max 2} = \frac{U_{\text{in}}}{nU_{0\min}} = 1.14 \\ M_{\min 2} = \frac{U_{\text{in}}}{nU_{0\max}} = 0.89 \end{cases} \quad (56)$$

The upper and lower bounds of the voltage gain can be obtained by considering the extreme values of both forward and reverse operation:

$$\begin{cases} M_{\max} = \{M_{\max 1}, M_{\max 2}\} = 1.14 \\ M_{\min} = \{M_{\min 1}, M_{\min 2}\} = 0.875 \end{cases} \quad (57)$$

As evident from (47), the voltage gain depends on both the inductance ratio, k , and the normalized frequency, f_n . Once the allowable voltage gain range and switching frequency range are predetermined, the maximum value of k can be determined.

By setting the minimum normalized frequency to $f_{n,\min} = 0.75$ to achieve a relatively wide switching frequency range, and substituting this into the PO mode gain formula, we obtain:

$$k < 6.67. \quad (58)$$

C. Determination of the characteristic impedance Z_r

The characteristic impedance determines the converter's ability to maintain voltage regulation under variations in input and load. If Z_r is huge, according to (4) and (11), it would lead to an impractically large L_m , making it difficult to maintain reasonable operational efficiency. Therefore, Z_r must be constrained by an upper limit. The gain in PO mode decreases with increasing normalized frequency. Under full-load operation, when the normalized frequency reaches 0.75, (54) combined with Table 1 yields:

$$\begin{cases} Z_{r1} < 13.93 \\ Z_{r2} < 10.72 \end{cases} \quad (59)$$

The characteristic impedance must satisfy the operational constraints for both forward and reverse modes. The final

selected value is $Z_r = 10\Omega$.

D. Resonant Parameter Design

Based on the specifications in Table 1 and the completed design of parameters k and Z_r , the resonant parameters can be calculated using the following (1) and (4). The final optimized resonant parameters are summarized in Table II.

TABLE II
PARAMETERS OF CLLC RESONANT CONVERTER

Parameter	Value
L_{r1}	15.01 μH
C_{r1}	150.17 nF
L_m	100.07 μH
L_{r2}	6.68 μH
C_{r2}	337.88 nF

To achieve soft switching for reducing switching loss, the parameters must satisfy (11).

This design employs MOSFET model STP35N60DM2. The datasheet specifies an output capacitance C_{oss} of 110 pF. With t_d set to 200 ns, the maximum allowable L_m calculated from (11) is 1.52 mH. Since this value exceeds the designed L_m of 100.57 μH in this work, the implemented parameters satisfy the soft-switching requirements.

V. EXPERIMENTAL VALIDATION

Based on the resonant parameters derived from the proposed design method, a prototype is constructed. Given the symmetry between forward and reverse power flow characteristics, the analysis focuses on the forward power flow results.

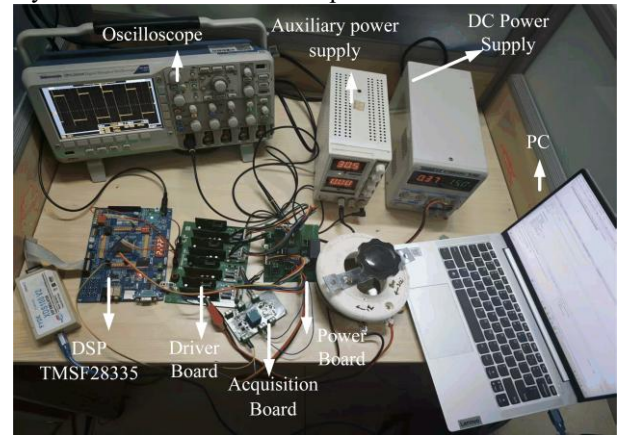
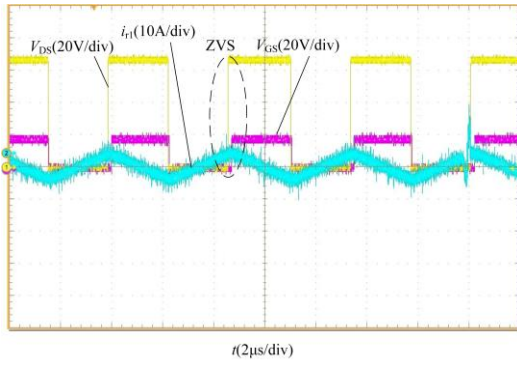
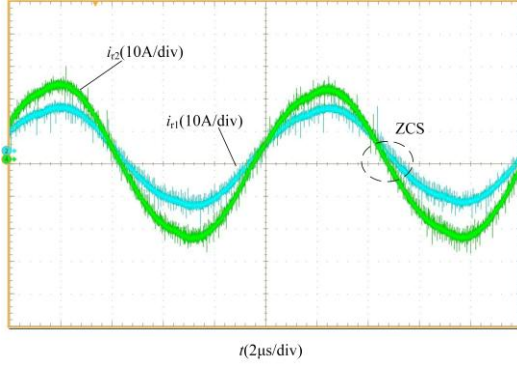


Fig.13. Hardware setup for the experimental test.

The experimental waveforms in Fig.13 show the results measured on the hardware prototype of the CLLC resonant converter operating in the forward direction under both P mode and PO mode. In Figs.14-(a) and 15-(a), the primary-side switches achieve accurate ZVS. In addition, Figs.14-(b) and 15-(b) illustrate that currents i_{r1} and i_{r2} achieve precise ZCS in both operating modes.

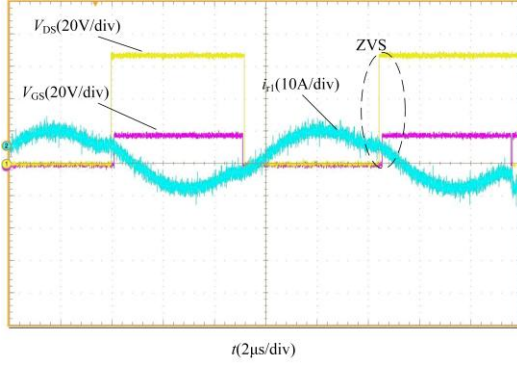


(a)

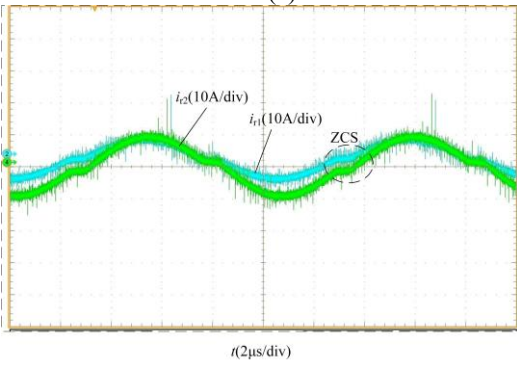


(b)

Fig.14. Experimental Waveforms in P Mode. (a) Primary-Side Current Achieving ZVS. (b) Secondary-Side Current Achieving ZCS.

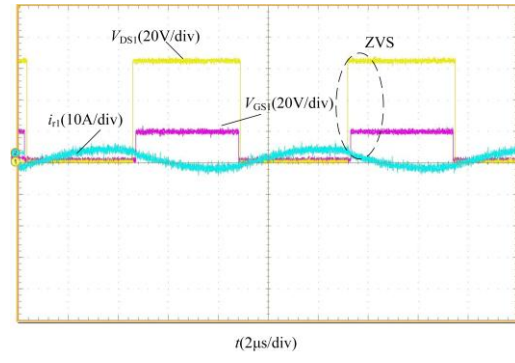


(a)



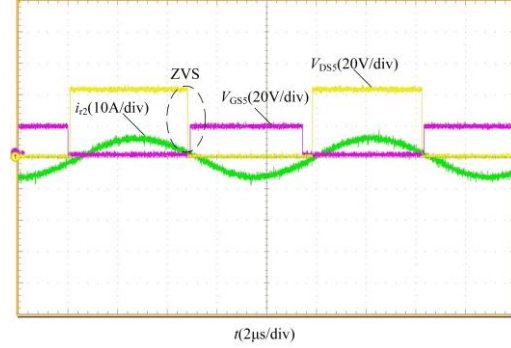
(b)

Fig.15. Experimental Waveforms in PO Mode. (a) Primary-Side Current Achieving ZVS. (b) Secondary-Side Current Achieving ZCS.

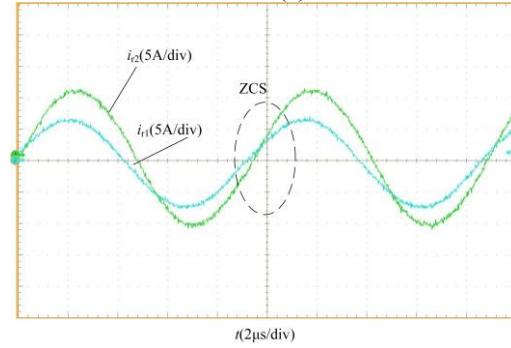


(a)

Fig.16. Experimental Waveforms in NP Mode.

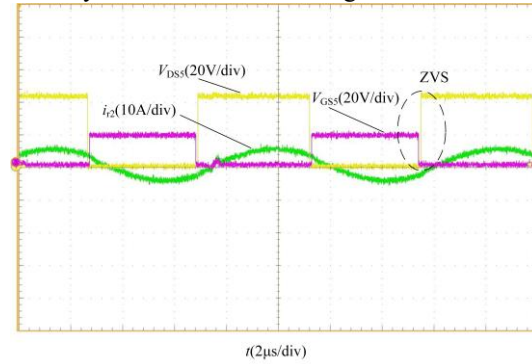


(a)



(b)

Fig.17. Experimental Waveforms of P Mode under Reverse Operation. (a) Primary-Side Current Achieving ZVS. (b) Secondary-Side Current Achieving ZCS.



(a)

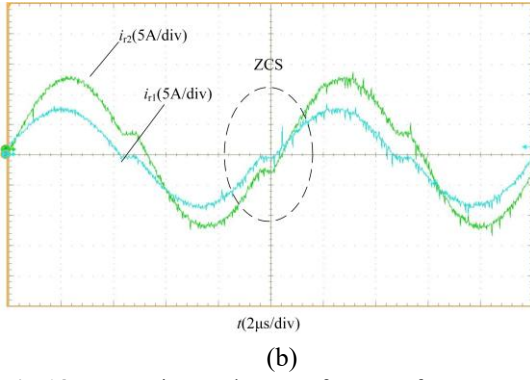


Fig.18. Experimental Waveforms of PO Mode under Reverse Operation. (a) Primary-Side Current Achieving ZVS. (b) Secondary-Side Current Achieving ZCS.

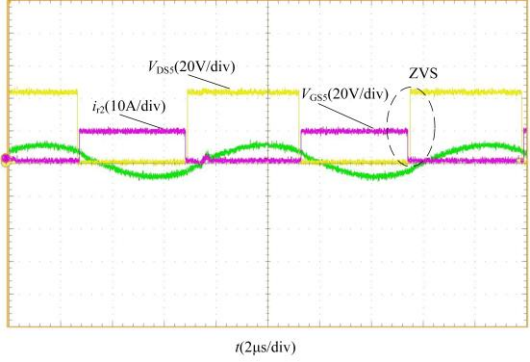


Fig.19. Experimental Waveforms of NP Mode under Reverse Operation.

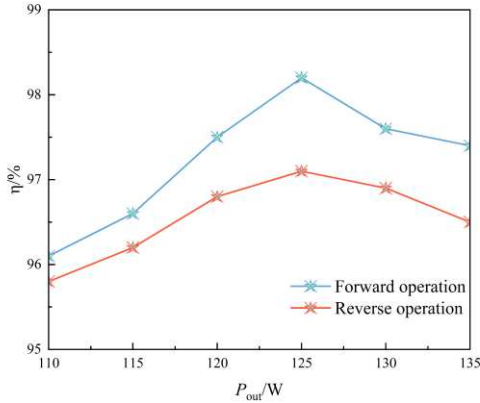


Fig.20. Efficiency of the prototype at various load levels.

Fig. 20 presents the efficiency curves of the prototype operating in both forward and reverse directions under different load conditions. In each case, we fixed the input voltage according to Table I. The results show that the proposed parameter design methodology enables the converter to achieve reliable soft-switching operation while maintaining high efficiency across varying load levels.

VI. CONCLUSION

This paper addresses the accuracy limitations of the FHA in analyzing the CLLC resonant converter. A detailed

investigation is carried out in the time domain under specific operating modes, with a simplified time-domain model proposed for the most efficient mode. The influence of dead time, voltage gain, and mode boundaries on converter performance is discussed. A systematic design methodology is summarized, and a prototype is developed to validate the proposed approach through experimental results.

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