Mode Analysis and Optimum Design of Bidirectional CLLC Resonant Converter for High-frequency Isolation of DC Distribution Systems

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Abstract—This paper points out the low accuracy of the traditional fundamental harmonic approximation (FHA) in predicting the gain characteristics and resonant behaviors of the full-bridge isolated bidirectional CLLC resonant converter, which results in great impediments in optimally designing the resonant tank parameters. Given these drawbacks of FHA for CLLC converter, this paper proposes a novel operation mode analysis for CLLC resonant converter. Compared with traditional FHA, the operation mode analysis not only explains the appearance of various operation modes, but also provides the accurate description of the gain characteristics and resonant behaviors. Based on the operation mode analysis and the typical application in dc distribution systems for CLLC resonant converter, the influence of multiple factors for CLLC resonant converter is clarified and an optimal design methodology is developed further. Applying this design method, the converter can minimize the conduction loss while the required gain of a fixed switching frequency range is maintained. At last, the experiment results verify the correctness of the operation mode analysis and the excellent performance of CLLC resonant converter constructed by the proposed design methodology.

Keywords—bidirectional dc-dc converter; distribution system; CLLC resonant tank; mode analysis; optimal design.

I. INTRODUCTION

With the rapid development of the power electronic technology, modern control strategy and dc output renewable energy sources, the dc distribution systems are superior to the traditional ac distribution system due to its advantages of small device cost, high power conversion efficiency and high reliability[1-2].

Fig. 1 illustrates a typical configuration of a dc distribution system [3]. It is apparent that the bidirectional dc-dc converters (BDCs) are the everlasting key components for dc distribution systems to interface between different dc buses [4-5]. Especially, isolated BDCs are valuable to replace the low-frequency transformers for the high-frequency galvanic isolation between dc buses and ac grids with significantly reduced weight, volume and expense. Various isolated BDCs have been developed for higher power density, lower power conversion losses and more stable operation ability. They generally adopt half-bridge or full-bridge type switches for high frequency converting and target control while employing the high-frequency transformers for galvanic isolation and voltage transformation.

Among various isolated BDC types, the full-bridge isolated bidirectional CLLC resonant converter is newly presented [6-7]. It has zero-voltage switching capability for primary power switches and soft commutation capability for rectifiers. In addition, the turn-off power loss is low due to the small turn-off current of primary power switches. Also, it dispenses with snubber circuits to reduce the voltage stress of switching devices and uses simple variable-frequency control method as traditional unidirectional LLC resonant converters. Due to these advantages, the bidirectional CLLC resonant converter has attracted a lot of research interests and has been practically applied for the dc distribution systems [8-9].

However, the inherent characteristic and resonant behaviors of the bidirectional CLLC resonant converter are hard to obtain and analyze for its complex resonant tank structure and various operation modes. That brings about a great obstacle in guiding the design of the resonant elements and control loop for CLLC resonant converters. As the widely applied theoretical research method for the resonant type converters, the traditional fundamental harmonic approximation (FHA) method has been used.

Fig. 1. Typical configuration of a low voltage dc distribution system.
to obtain the gain characteristics of LLC resonant converter [6-7]. But this dc gain is not accurate for FHA only takes the fundamental component of the circuit current and voltage into account while neglecting the considerable high-order components. In recent years, a generalized operation mode analysis has been presented to predict and study the gain characteristics and steady operation behaviors for LLC resonant converter [10-11]. These predictions are consistent in the experiment results. The operation mode analysis stems from the basic circuit time domain theory and advanced numerical calculation method. Hence, it has widely applicable ability for several resonant type converters including the bidirectional LLC resonant converter.

In this paper, the discussion concentrates on the full-bridge isolated bidirectional LLC resonant converter for a typical dc distribution system shown in Fig. 1. Low accuracy for the FHA in predicting the gain characteristics and resonant behaviors for LLC converter is pointed out. Owing to these drawbacks, the analysis for the operation mode of LLC converter is proposed. By this analysis, the exact gain curves and resonant waveforms according to the switching frequency, load conditions and resonant parameters are obtained. Then the influences of the dead time, load conditions and resonant elements for LLC converter are clarified respectively. On the basis of the analysis above, an optimal design methodology is presented to minimize the conduction loss while the required gain of a fixed switching frequency range is maintained. A 1-kW prototype with 380V input and 380V output is constructed to verify the validity of the operation mode analysis and the optimal design methodology.

II. LOW ACCURACY OF TRADITIONAL FUNDAMENTAL HARMONIC APPROXIMATION

The circuit topology of the isolated bidirectional LLC resonant converter shown in Fig. 2 has the symmetrical structure. The high-frequency transformer is modeled with the ideal transformer whose turn ratio is 1 and the magnetizing inductance. Besides the resonant network is composed of the series inductance and series capacitors. Full-bridge H1 and H2 act as the inverting stage and rectifying stage respectively in forward operation mode, which is reverse in backward operation mode. Via the symmetrical structure and identical control method, the operation behaviors in both power flow directions are same and this paper focuses on the forward operation mode. Here the power switches of H1 adopt variable frequency control with 50% duty cycle while that of H2 is out of control and the parasitic diodes of the power switches, M1—M8, serve as the rectifier.

![Fig. 2. Circuit topology of isolated bidirectional LLC resonant converter](image)

Fig. 2 illustrates the equivalent circuit of the full-bridge isolated bidirectional LLC resonant converter using FHA [6-7], where \( V_{\text{ro,FHA}} \) and \( V_{\text{rin,FHA}} \) are the fundamental components of the input and output voltage of the resonant network respectively and change with switching frequency \( f_s \). \( R_{eq} \) is the ac output equivalent resistance of the resonant network and is calculated in (1) with the resistive load, \( R_o \)

\[
R_{eq} = 8R_o / \pi^2 \tag{1}
\]

Analyzing the equivalent circuit of LLC resonant converter in Fig. 3, the voltage gain \( M \) varied with the normalized switching frequency \( f_s \) could be derived as follows:

\[
M = \frac{V_o}{V_{in}} = \left| \frac{V_{\text{ro,FHA}}(f_s)}{V_{\text{rin,FHA}}(f_s)} \right| = \frac{f_s^2}{\sqrt{f_s^2 E_1^2(f_s) + Q^2 E_2^2(f_s)}} \tag{2}
\]

where \( E_1(f_s) \) and \( E_2(f_s) \) are function components as given in (3) and (4), respectively

\[
E_1(f_s) = f_s^2(1+1/k) - 1/k \tag{3}
\]

\[
E_2(f_s) = f_s^4(2+1/k) - f_s^2(2+2/k) + 1/k \tag{4}
\]

Moreover, \( f_s = f_1 / f_2, f_s = 1 / (2\pi\sqrt{L_r/C_r}) \) is the series resonant frequency, \( Q = Z_o / R_{eq} \) is the quality factor, \( Z_o = \sqrt{L_r/C_r} \) is the characteristic impedance, \( k = L_o / L_r \) is the inductance ratio.

The specifications for the LLC resonant converter in [6-7] are applied to (1)—(4). Thus, the curves for the voltage gain \( M \) versus the normalized switching frequency \( f_s \) under different \( k \) can be obtained in Fig. 4. In addition, the corresponding Psipce simulation results are provided for comparison.

![Fig. 3. Equivalent circuit of isolated bidirectional LLC resonant converter with FHA](image)
When $S_{11}, S_{44}$ turn on and $S_{22}, S_{33}$ turn off.

to merely analyze the converter behaviors of the first half cycle and voltage waveforms of one half cycle is symmetrical to that of the other in the resonant network. Therefore, it is reasonable to merely analyze the converter behaviors of the first half cycle when $S_{12}, S_{43}$ turn on and $S_{21}, S_{34}$ turn off.

**CLLC resonant converter.**

### III. RESEARCH OF ISOLATED BIDIRECTIONAL CLLC RESONANT CONVERTER WITH OPERATION MODE ANALYSIS

Due to the fact that FHA has the drawbacks of low accuracy and low applicability for the study and design of CLLC resonant converter, this section details the operation modes and gain characteristics of CLLC resonant converter.

Because **CLLC resonant converter** adopts the full-bridge topologies to set up the inverting and rectifier networks under the variable frequency control with 50% duty cycle, the current and voltage waveforms of one half cycle are symmetrical to that of the other in the resonant network. Therefore, it is reasonable to merely analyze the converter behaviors of the first half cycle when $S_{11}, S_{44}$ turn on and $S_{22}, S_{33}$ turn off.

**CLLC resonant converter** can operate in diverse modes depending on switching frequency and load condition when the circuit parameters are determined. Each mode is composed of multiple stages. In the first half cycle, three different stages may exist, which are differentiated by the output voltage of the resonant network, $V_{out}$: the positive clamped stage $V_{o} = V_{2}$, the negative clamped stage $V_{o} = -V_{2}$ and the freewheeling stage $|V_{o}| < V_{2}$. They are named $P_{b}$ stage, $N_{b}$ stage and $O_{b}$ stage, and their equivalent circuits are depicted in Fig. 5(a), (b) and (c) respectively. Subscript $b$ stands for bidirection, which is used to distinguish the stages of the bidirectional **CLLC converter** from the $P$ stage, $N$ stage and $O$ stage of traditional **LLC converters** [10].

Through the theoretical research and simulation verification, eight operation modes are searched out and named after the sequence of the stages in the first half cycle: $O_{b}, P_{b}O_{b}P_{b}, N_{b}P_{b}, P_{b}P_{b}N_{b}, P_{b}O_{b}$, and $P_{b}O_{b}N_{b}$. They all can realize the zero-voltage switching for the power switches of the inverting network. The following discussion focuses on three common modes: $P_{b}, P_{b}O_{b}$ and $P_{b}O_{b}N_{b}$.

When the switching frequency $f_{s}$ equals the series resonant frequency $f_{r}$, **CLLC resonant converter** operates in $P_{b}$ mode and the operation waveforms are depicted in Fig. 6(a). In $P_{b}$ mode, the primary resonant current $i_{L_{1}}$ and the magnetizing current $i_{L_{m}}$ have the same initial value. Then $i_{L_{1}}$ begins to outweigh $i_{L_{m}}$, $V_{o}$ turn on and $M_{2}, M_{3}$ turn off. That makes $V_{o}$ clamped by $V_{2}$ and all the passive components of the **CLLC resonant network** involved in the operation. $P_{b}$ mode ends until the $i_{L_{1}}$ and $i_{L_{m}}$ are equal again. The equivalent circuit of $P_{b}$ mode is shown in Fig. 5(a). $i_{L_{1}}, i_{L_{2}}, V_{C_{1}}$ and $V_{C_{2}}$ are the selected state variables and their state equations could be established. After solving the state equations and normalizing with the base values in (4), the normalized time domain expressions of the state variables are obtained as (5).

\[
V_{b} = V_{2} \quad Z_{b} = Z_{r} \quad I_{b} = V_{b} / Z_{b}
\]

\[
\begin{align*}
\left\{ \begin{array}{l}
 i_{L_{1}p_{b}} (\theta) = & I_{1p_{b}} \sin(\theta + \theta_{p_{b}}) + I_{2p_{b}} \sin(k_{1} \theta + \theta_{2p_{b}}) \\
 i_{L_{2}p_{b}} (\theta) = & I_{1p_{b}} \sin(\theta + \theta_{p_{b}}) - I_{2p_{b}} \sin(k_{1} \theta + \theta_{2p_{b}}) \\
 v_{c_{1}p_{b}} (\theta) = & -I_{1p_{b}} \cos(\theta + \theta_{p_{b}}) - I_{2p_{b}} / k_{1} \cos(k_{1} \theta + \theta_{2p_{b}}) + 1 / M \\
 v_{c_{2}p_{b}} (\theta) = & -I_{1p_{b}} \cos(\theta + \theta_{p_{b}}) + I_{2p_{b}} / k_{1} \cos(k_{1} \theta + \theta_{2p_{b}}) - 1
\end{array} \right.
\]

where $\theta = \omega_{o} t$ is the operational angle, $\omega_{o} = 2\pi f_{r}$ is the resonant angular frequency, $k_{1} = 1/(1 + 2k)$, $I_{1p_{b}}, \theta_{1p_{b}}, I_{2p_{b}}$ and $\theta_{2p_{b}}$ are the unknowns.

From the symmetry of the operation waveforms in the first half cycle, the initial value of each state variable is opposite to its end, which is described as follows:

\[
\begin{align*}
 i_{L_{1}p_{b}} (0) = & -i_{L_{1}p_{b}} (\pi) \\
 i_{L_{2}p_{b}} (0) = & -i_{L_{2}p_{b}} (\pi) \\
 v_{c_{1}p_{b}} (0) = & -v_{c_{1}p_{b}} (\pi) \\
 v_{c_{2}p_{b}} (0) = & -v_{c_{2}p_{b}} (\pi)
\end{align*}
\]

Moreover, the normalized output current of **CLLC resonant converter** could be calculated and simplified as
The equation set for Pb mode is established with (5)—(7) and can be solved for the analytic solutions of the unknowns $M, i_{1P}, \theta_{1P}, i_{2P}$ and $\theta_{2P}$. So the resonant behaviors of CLLC resonant converter under Pb mode can be exactly described. It should be noted that the obtained voltage gain $G$ equals 1.

When the switching frequency $f_s$ gradually reduces from the series resonant frequency $f_r$, CLLC resonant converter will operate in PbOb mode as illustrated in Fig. 6(b). In PbOb mode, the resonant behaviors of Pb stage are identical to that of Pb mode. When Pb stage ends, the amplitude of $v_{o1}$ is smaller than $V_2$, $M_1$—$M_4$ turn off and Ob stage begins to appear. In Ob stage, $L_{r2}$ and $C_{r2}$ no longer join the operation of CLLC converter. Hence, $i_{L2}$ equals zero, $u_{C2}$ stays constant and $L_{r1}, C_{r1}, L_{m}$ constitute the resonant tank. The equivalent circuit of Ob stage is given in Fig. 5(c) and the normalized time domain expressions of the state variables $i_{L1}$ and $u_{C1}$ during Ob stage are derived as follows:

$$\begin{align*}
i_{L1Ob}(\theta) &= I_{o1} \sin(k_2 \theta + \theta_{o1}) \\
u_{C1Ob}(\theta) &= -I_{o1} / k_2 \cos(k_2 \theta + \theta_{o1}) + 1 / M
\end{align*}$$

where $k_2 = \sqrt{1/(1 + k)}$, $I_{o1}$ and $\theta_{o1}$ are the amplitude and initial angle of $i_{L1Ob}$ and they are the unknowns.

Because (5) and (8) are the transcendental equation sets, the equation set for PbOb mode which is composed of (5), (8) and the constraints has no analytic solutions. However, its numerical solutions can be fast obtained using the advanced numerical calculation method with suitable initial values. Then, the resonant behaviors of PbOb mode could be described. In PbOb mode, the initial value of $i_{L1}$ is negative and $i_{L2}$ equals zero during Ob stage. Hence, the power switches of the inverting network achieve zero-voltage switching and the diodes of the rectifier network realize soft commutation, which brings about the high operation efficiency for the CLLC resonant converter. Meanwhile, the voltage gain $M$ monotonically increases from 1 according to the decrease of the switching frequency $f_s$ from $f_r$. It suggests PbOb mode is a boost mode and its closed-loop control stability could be ensured with a simple linear feedback controller. Thus, PbOb mode is an excellent operation mode for the CLLC resonant converter.

With the switching frequency $f_s$ decreasing continually, $N_0$ stage will emerge following Ob stage and CLLC converter will operate in PbObN0 mode. When Ob stage ends, $v_{o1}$ drops to $-V_2$, which forces the conduction of $M_3$ and $M_4$. Therefore, CLLC converter moves into $N_0$ stage and all the resonant components join the operation. Throughout the $N_0$ stage, $i_{L1}$ is smaller than $i_{Lm}$ and $i_{L2}$ is negative. Fig. 5(b) depicts the equivalent circuit of $N_0$ stage. The normalized time domain expressions of the state variables $i_{L1}, i_{L2}, v_{C1}$ and $v_{C2}$ during $N_0$ stage are given in (9).

$$\begin{align*}
i_{L1N0}(\theta) &= I_{1N0} \sin(\theta + \theta_{1N0}) + I_{2N0} \sin(k_2 \theta + \theta_{2N0}) \\
i_{L2N0}(\theta) &= I_{1N0} \sin(\theta + \theta_{1N0}) - I_{2N0} \sin(k_2 \theta + \theta_{2N0}) \\
u_{C1N0}(\theta) &= -I_{1N0} \cos(\theta + \theta_{1N0}) \\
u_{C2N0}(\theta) &= -I_{1N0} \cos(\theta + \theta_{1N0}) + I_{2N0} / k_1 \cos(k_2 \theta + \theta_{2N0}) + 1 / M \tag{9}
\end{align*}$$

where $I_{1N0}, \theta_{1N0}, I_{2N0}$ and $\theta_{2N0}$ are the unknowns.

The equation set for PbObN0 mode could be built and solved as PbOb mode. When the operation of CLLC converter moves into PbObN0 mode from PbOb mode, the gain $M$ firstly increases and then decreases with switching frequency $f_s$ decreasing. It indicates that the gain curve reaches its peak in this mode. The switching frequency ought to be higher than that of the peak, $f_{p'}$, to guarantee the control stability. Moreover, the switching frequency of the boundary between PbOb and PbObN0 modes is near to $f_r$. Hence, PbOb is the main operation mode in the switching frequency range of $f_r$ to $f_{p'}$.

With the equation sets for the other five modes established and solved, their operation behaviors can also be described. Thereinto, $O_0$ mode exists under no load conditions for CLLC converter; $O_0P_1O_0$ and $N_0P_0P_0P_0$ modes appear when the load of the CLLC converter is light; $P_0N_0$ mode exists after $P_0$ mode, the gain $M$ first increases and then decreases with the load increasing. $N_0P_0$ emerges when $f_s$ is lower than $f_r$ and is a buck operation mode.

Based on the establishment and solutions for the equation sets of the operation modes, the resonant behaviors and gain characteristic can be obtained exactly. Fig. 7 depicts that the obtained gain curves are in conformity with the simulation data and verify the high accuracy of the operation mode analysis.
IV. OPTIMAL DESIGN CONSIDERATION AND METHOD FOR CLLC RESONANT CONVERTER

In view of the exact operation mode analysis in Section III, the gain curves and operation behaviors of the isolated bidirectional CLLC resonant converter can be accurately predicted. On this basis, the influences of the dead time, load conditions and resonant components on the performances of CLLC converter are discussed in this section for design considerations. And then an optimal design methodology is proposed for CLLC resonant converter.

A. Design Consideration of Dead Time

The ZVS operation of the primary power switches is a vital factor for the optimal design of CLLC converter and has a close relation with the dead time $t_d$ [6]. During the dead time, the output capacitance of primary power switches must be charged or discharged completely even under the worst-case conditions, i.e., when the switching frequency reaches its maximum under output voltage regulation and no load conditions. And then the correlation between $t_d$ and the magnetizing inductance $L_m$ for the achievement of the ZVS operation is derived as follows:

$$t_d \geq 8C_p f_{s,\text{max}} L_m$$

(10)

where $C_p$ is the effective output capacitance of the primary power MOSFETs, which can be obtained from the specializations, $f_{s,\text{max}}$ is the requested upper limit of the switching frequency.

Considering $t_d$ is usually given in advance for the design of the resonant converter [6-7], (10) is able to be transformed into the following inequality:

$$L_m \leq t_d / 8C_p f_{s,\text{max}}$$

(11)

Inequality (11) indicates that there is a maximum value for $L_m$ and it could be used as a constraint condition for the optimal design of CLLC converter.

B. Design Consideration of Load Conditions

With the mode equation sets solved, the gain characteristics of CLLC resonant converter can be exactly obtained. Fig. 8 shows a typical dc gain characteristic plot with the load current $I_2$ varying from zero to rated value. The enlarged view of the dc gain characteristic around the resonance ($f_n = 1$) is illustrated as well.

It can be seen from Fig. 8 that a majority of the gain curves cross unity at point A ($f_n = 1$) while the rest in a small range of light-load to no-load cross unity along the line AB. In the paper, CLLC converter is used for the high-frequency isolation as a dc transformer and its voltage gain ought to be equal to unity in the range of rated-load to no-load. Therefore, the line AB is the most suitable operation region for CLLC converter without regard to the other factors, such as the power loss. Nevertheless, due to the voltage drop from the power loss of semiconductor switches, resonant components and high-frequency transformer, the gain above unity of CLLC converter has to be sufficient to regulate the output voltage.

In Fig. 8, each gain curve has a peak value larger than unity below the resonance ($f_n < 1$). For a certain operation frequency range above the frequency of the peak gain, the gain increases monotonically with the decrease of the frequency. It implies that the maximum gain variation exists at the minimum operation frequency $f_{n, \text{min}}$ ($f_{n, \text{min}} = f_{\text{min}} / f_n$). Meanwhile, compared to other lighter load, the gain curve under rated load has the lowest gain value at the same frequency. Thus, if the gain at $f_{n, \text{min}}$ under rated load is larger than the required maximum gain $M_{\text{max}}$, CLLC converter has ability to regulate the output voltage for the whole load range. On the other hand, the frequency of the peak gain under rated load, $f_{n, \text{rated}}$ is larger than that under other lighter load conditions, therefore, $f_{n, \text{rated}}$ should be smaller than $f_{n, \text{min}}$ to ensure the control stability. On the basis of the analysis above, the following two constraints can be obtained:

$$M_{f_{n, \text{min}}, \text{rated}} \geq M_{\text{max}}$$

(12)

$$f_{p, \text{rated}} \leq f_{n, \text{min}}$$

(13)

where $M_{f_{n, \text{min}}, \text{rated}}$ is the gain under rated load at $f_{n, \text{min}}$.

Therefore, only if the gain curve under rated load conditions meets (12) and (13), CLLC converter could regulate the output voltage in the whole load range.
Fig. 9. Gain curves under different switching frequency. (a) $k$ varies from 5 to 20 and $Z_r$ equals 12. (b) $Z_r$ varies from 6 to 24 and $k$ equals 10.

### C. Design Consideration of $k$ and $Z_r$

The resonant components, $L_r$, $L_m$, and $C_r$, strongly affect the converter’s performances and are the ultimate targets of the optimal design. They can be expressed with the inductance ratio $k$ and the characteristic impedance $Z_r$ as shown in (14). Thus, the original design of three resonant components can be simplified to the design of $k$ and $Z_r$. The influences of $k$ and $Z_r$ on the operating range and the power conversion efficiency are discussed below and the optimal design idea for $k$ and $Z_r$ is given further.

\[
L_r = \frac{Z_r}{\omega} \quad L_m = kL_r \quad C_r = \frac{1}{(Z_r, \omega)} \tag{14}
\]

Under the rated load condition, the typical gain characteristic plots with $k$ or $Z_r$ variations are depicted in Fig. 9 (a) and (b) respectively. The required maximum gain $M_{\text{max}}$ and minimum normalized switching frequency $f_{n_{\text{min}}}$ are also given for design considerations: $M_{\text{max}}$ is set as 1.05 for the purpose of compensating for the voltage drop from the power loss while $f_{n_{\text{min}}}$ is set as 0.85 to obtain a relatively narrow switching frequency range.

It is obvious from Fig. 9(a) that with a fixed small $Z_r$, $f_{\text{RMS}}$ is always lower than $f_{n_{\text{min}}}$ for the gain curves under different $k$, which suggests (13) is met. But with $k$ increasing, the $M_{f_{\text{min}}, \text{rated}}$ decreases and will be lower than $M_{\text{max}}$ when $k$ is greater than a certain value. That is to say, there is an upper limit $k_{\text{max}}$ for $k$ to meet (12). From Fig. 9(b), with $k$ fixed and $Z_r$ increasing, $f_{\text{RMS}}$ moves to $f_{n_{\text{min}}}$ and will outweigh $f_{n_{\text{min}}}$ at last. It indicates the existence of an upper limit $Z_{r1}$ for $Z_r$ to guarantee (13), in addition, $M_{f_{\text{min}}, \text{rated}}$ remains unchanged for the gain curves satisfying (13). It can be concluded that the satisfaction of (12) rests with $k$ while that of (13) depends on $Z_r$. Hence, $k$ and $Z_r$ can be confined for meeting (12) and (13) respectively as follows:

\[
k \leq k_{\text{max}} \tag{15}
\]

\[
Z_r \leq Z_{r1} \tag{16}
\]

The determination process of $k_{\text{max}}$ and $Z_{r1}$ is detailed below. From Section III it can be aware that when LLC converter with certain load operates in the frequency range of $f_n$, $P_{\text{OH}}$ is its main operation mode. And the mode distribution of $P_{\text{OH}}$ is illustrated in Fig. 9(a) and (b). It is obvious from Fig. 9(a) that the converter operates in $P_{\text{OH}}$ mode at $f_{n_{\text{min}}}$ as $f_{n_{\text{min}}}$ belongs to the frequency range mentioned above. Meanwhile, a one-to-one correspondence exists between $M$ and $k$ at $f_{n_{\text{min}}}$. Let $M$ equal $M_{\text{max}}$, $f_n$ equal $f_{n_{\text{min}}}$ and solve the equation set for $P_{\text{OH}}$ mode, thus $k_{\text{max}}$ can be obtained. In Fig. 9(b), the peak gain curve and the boundary curve between $P_{\text{OH}}$ and $P_{\text{ON}}$ modes closely coincide. Hence, the information of the peak gain point, such as $f_n$, can be approximated by that of the boundary. On this boundary, another one-to-one correspondence exists between $f_n$ and $Z_r$. Let $k$ equal $k_{\text{max}}$, $f_n$ equal $f_{n_{\text{min}}}$ and solve the equation set for this boundary, thus $Z_{r1}$ can be obtained.

On the other hand, with accurate current waveforms solved from the mode equation sets, the RMS of these waveforms can be calculated out and the RMS of the normalized primary current $I_{L1n}$ is expressed as follows:

\[
I_{L1n, \text{RMS}} = \sqrt{\frac{1}{f_n} \int_{0}^{\pi/2} i_{L1n}(\theta)^2 d\theta} \tag{17}
\]

With the rated load, the $I_{L1n, \text{RMS}}$ according to the normalized frequency for different $k$ and $Z_r$ are shown in Fig. 10(a) and (b). It can be seen from Fig. 10(a) that the $I_{L1n, \text{RMS}}$ decreases with $k$ increasing. Therefore, high $k$ can reduce the conduction loss of the LLC converter. In Fig. 10(b), the $I_{L1n, \text{RMS}}$ decreases as $Z_r$ increases. But this decrease is no longer clear when $Z_r$ is larger than a certain value $Z_{r2}$. Given high $Z_r$, the $I_{L1n, \text{RMS}}$ decreases as $Z_r$ increases. Therefore, high $k$ and $Z_r$ can reduce the conduction loss of the LLC converter.

Given the discussion above and the constraints (15) and (16), the final constraints for $k$ and $Z_r$ can be obtained as follows:

\[
k = k_{\text{max}} \tag{18}
\]

\[
Z_r = \min\{Z_{r1}, Z_{r2}\} \tag{19}
\]

Substituting $k$ and $Z_r$ into (14), the optimal values for the resonant components, $L_r$, $L_m$ and $C_r$, can be calculated out.

Integrating the design considerations for the dead time, load conditions and resonant components, the optimal design procedure is illustrated in Fig. 11. Then the design results are applied for the construction of a 1-kW prototype with 380V input and 380V output.
Fig. 10. RMS of normalized primary current under different frequency. (a) k variations from 4 to 20. (b) Zr variations from 4 to 16.

Design Specializations

\[ V_{in}, V_{out}, f, f_{min}, f_{max}, M_{max}, t_{dead} \]

Assume \( k = 20 \)

Calculate \( L_{max} \) with \( t_{dead} \)

Calculate \( Z_r \) with \( k, L_{max}, f \)

Solve \( k_{min} \) with \( P_{Ob} \) mode equation set

Solve \( Z_1 \) with \( P_{Ob} / P_{ObNs} \) boundary mode equation set

Search \( Z_2 \) with correlations between \( Z_r \) and \( L_{max}, RMS \)

Calculate \( L_r, C_r \) and \( L_m \) with \( k_{min} \) and \( \min \{ Z_1, Z_2 \} \)

Fig. 11. Optimal design procedure for resonant components \( L_r, L_m \) and \( C_r \).

V. EXPERIMENTS

To verify the validity of the mode analysis and the optimal design methodology. The 1-kW prototype with 380V input and 380V output is built with the optimal resonant parameters. The series resonant frequency is set to 65kHz with 40uH \( L_r \), 150nF \( C_r \) and 380uH \( L_m \). Experimental waveforms of \( P_b \) and \( P_{Ob} \) modes are illustrated in Fig. 12, which is in consistent with the mode analysis results.

Fig. 12. Experiment waveforms. (a) \( P_b \) mode. (a) \( P_{Ob} \) mode.

The power conversion efficiency of the prototype under different load conditions is shown in Fig. 13. It indicates that the optimal design has high operation efficiency.

Fig. 13. Efficiency curve under different load conditions.

VI. CONCLUSION

Due to the low accuracy of the traditional FHA, this paper offers an accurate operation mode analysis for the newly proposed full-bridge isolated bidirectional \( CLLC \) resonant converter. Eight major operation modes formed by different combinations of three stages are found and their mode equation sets are built.
Exact descriptions of gain characteristic and resonant behaviors are obtained with the equation sets solved. On the basis of the accurate mode analysis, the influence of the dead time, load conditions and resonant parameters on the performances of CLLC converters are clarified and an optimum design methodology is proposed. The design method is implemented in the prototype and the experiment results verify the operation mode analysis and optimal design methodology.

REFERENCES
