Design Method of 6-Element Boundary Gain for LLC Resonant Converter of Electric Vehicle

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ABSTRACT In the application of on-board chargers, LLC resonant converters need to work in multiple operating modes to achieve wide-range output requirements. For the most commonly used single-phase and three-phase LLC topologies, the time domain expressions of all possible operating modes are introduced to predict voltage and current behaviors. Based on the mode analysis, the 6-element boundary gain design method is proposed to design a wide gain range LLC resonant converter. The resonant current is used as the basis for ZVS judgment and the transformation trend of ZVS in the full operating range is given. Through reasonable distribution of the boost region and the buck region, the resonant devices can be designed more appropriately to achieve lower conduction loss and higher efficiency. Finally, the prototypes of 3.3kW and 6.6 kW are built to verify the theoretical analysis of the method in single-phase LLC and three-phase LLC.

INDEX TERMS LLC resonant converter, zero-voltage switching, wide range output, battery charger, electric vehicle.

I. INTRODUCTION

In the past ten years, electric vehicles (EVs) have been commercialized to reduce demand for fossil fuels [1]–[5]. On electric vehicles, on-board charger (OBC) is installed to charge the high-voltage battery. The performance of OBC is significant for charging time. The typical structure of OBC is illustrated in Fig.1.

The most common OBC charger structure consists of an EMI filter and a two-stage converter. In this two-stage converter, a boost-type AC-DC converter is used as power factor correction and an isolated DC/DC converter is used to directly deal with complex load conditions [6], [7]. The high-voltage battery of EVs have the non-linear load i-v characteristics and a wide adjustable output voltage range. Therefore, the characteristic of this two-stage converter is mainly affected by the DC/DC stage since the voltage and power are regulated in this stage.

Adjustable \( V_{out} \), variable \( I_{out} \), high transmission efficiency, high power density and low electromagnetic interference (EMI) are the requirements for DC/DC converter. Considering these cases, inductor-inductor-capacitor (LLC) resonant converter is one of the most suitable topologies [8]–[10]. According to output power level and input bus voltage, it mainly includes single-phase LLC and three-phase LLC [11], [12]. The output current ripple rate of single-phase LLC resonant converter is 156.7%. Relying on the output capacitor to reduce the ripple is a great challenge for the capacitor. With the development of power devices, the pressure of high output ripple rate on output capacitor can be reduced by increasing the bus voltage. Besides, three-phase LLC resonant converter with three-phase interleaved operation also has a good benefit on ripple suppression [13]. At the same time, more operating modes also exist in the three-phase LLC, which poses a challenge to design the
resonant parameters accurately. As for the design methodologies, it is very difficult to design the LLC resonant parameters adapted to wide voltage gain range with frequency modulation. The fundamental harmonic approximation (FHA) and time-domain model method are commonly used to guide the design.

The conventional FHA method is adopted in [14]–[16]. The design result is accurate and effective when the operating frequency is near the resonant frequency. When the converter operates in a wide range of output voltage, the resonant current is not completely sinusoidal, which will cause a large errors. In order to overcome the shortcomings of the FHA method, the time-domain model is provided in [17]. With the help of calculation software, the expected resonant state of LLC converter can be calculated. The high precision of the calculation results in [17] is carried out under sophisticated calculation software. In [18], a more streamlined method is used to calculate the resonant parameters. The mode analysis provides time domain expressions of voltage, current and DC gain at different resonant states. Unfortunately, the effect of resonant current on ZVS is ignored. In [19], frequency domain analysis and time domain analysis are used as a comparison to design LLC. The realization of ZVS is based on the fact that the resonant current lags behind the resonant network input voltage. Although the impedance of the resonant tank is inductive, there may still be insufficient resonant current that may cause ZVS to fail. The method in [10] also analyzes ZVS based on the resonant current. But it can only guarantee soft switching at the resonant frequency and above the resonant frequency. In [20], the resonant converter is designed to work under an expected operating mode. Although the safety margin of capacitive-mode operation is considered, the optimized design method does not involve efficiency optimization. In [21], the operation mode boundaries and detailed design procedure are provided. The method in [21] can achieve both-side soft-switching in the full range of output. And the importance of resonant current for soft switching is mentioned when designing resonant parameters. Due to insufficient consideration of the transformer, the design method does not reasonably arrange the distribution of the boost region and buck region. At the resonant frequency, the magnetizing current and the resonant current are equal at the switching time. Therefore, in [22], the maximum value of the magnetizing inductor can be calculated to ensure the realization of ZVS. Under wide-range output applications, LLC resonant converter will definitely operate in the boost region under certain conditions. But in boost region, the magnetizing inductor current no longer changes linearly. Thus, the calculation of the critical current in [18], [19], [20] and [22] can not be used to judge the soft switching situation.

In this article, section 2 introduces two representative topologies in on-board charges. According to the different operating frequencies, we use time domain model to analysis to study the resonant states in single-phase LLC and three-phase LLC. All typical intervals of resonant states are proposed. In section 3, the selection method of the transformer turns ratio makes the distribution of the resonant converter in the boost region and the buck region more reasonable. According to the mode boundary gain, the maximum value of the inductance ratio can be obtained. The change trend of the switching current is provided for design considerations to achieve ZVS. According to the relationship of 6 elements related to the design, we provides a set of OBC design procedure. The design parameters of the proposed method are tested in the experimental platforms and the test results are in Section 4.

II. RESONANT STATE ANALYSIS AND TIME DOMAIN MODEL

The LLC resonant converter consists of square wave generator, resonant tank, rectifier diodes and filter capacitor. The topological structure of single-phase full-bridge LLC and three-phase interleaved LLC converter is illustrated in Fig.2. As shown in Fig.2, the resonant tank contains three components: resonant capacitor $C_r$, magnetizing inductor $L_m$, resonant inductor $L_r$.

The converter uses $f_r = 1/2\pi \sqrt{L_m/C_r}$ as resonant frequency. As shown in Fig.3 and Fig.4, both single-phase and three-phase LLC have similar characteristics. All resonant states, can be distinguished by the state of the magnetizing inductor. In single-phase LLC, stage P and N occur when the magnetizing inductance is positive or negative clamped. Since the input voltage of the three-phase LLC has 4 levels, there are 4 stages to correspond: (2/3) P, (1/3) P, (2/3) N, (1/3) N. These 4 stages reflect the positive and negative clamping states of the magnetizing inductor in the three-phase LLC.
When the operating frequency is greater than the resonant frequency, the operating waveform is shown in Fig.3. The most obvious feature in the below-resonance region is that the phase of the magnetizing inductor voltage lags behind the tank input voltage. As shown in fig.3(a), in interval \([t_0-t_1]\), the single-phase LLC operates in stage P. Magnetizing inductor current increases linearly to \(t_2\). In a half cycle, the stage P in single-phase LLC corresponds to \((1/3)P\), \((2/3)P\) and \((1/3)P\) of three-phase LLC. Stage N in single-phase LLC corresponds to \((1/3)N\), \((2/3)N\) and \((1/3)N\) of three-phase LLC. As shown in fig.3(b), in interval \([t_2-t_3]\), resonant state \((2/3)P\) occurs when the magnetizing inductor is clamped by two-thirds of the output voltage \((2/3)V_{\text{out}}\).

When the corresponding output diodes turn off, the energy will not be transferred to the secondary side, and single-phase and three-phase LLC will have stage O in below-resonance region. The interval \([t_1-t_2]\) in the single-phase and the interval \([t_3-t_6]\) in the three-phase are the resonant converter working in the stage O. In particular, when one phase of the three-phase LLC operates in stage O, it will affect the resonant state of other two phases. State O-P and O-N are defined as the special stage in three-phase LLC. In fig.3(b), interval \([t_1-t_2]\) and interval \([t_3-t_4]\) correspond to resonant state O-P. Interval \([t_5-t_6]\) and interval \([t_6-t_7]\) correspond to resonant state O-N.

When the operating frequency is less than the resonant frequency, the operating waveform is shown in Fig.4. In above-resonance region, the phase of the magnetizing inductor voltage leads the tank input voltage. Compared with below-resonance region, there will be no stage O before the end of the half cycle. The magnetizing inductor is still clamped for a short period of time after the switching is over. The interval \([t_2-t_3]\) in the single-phase and the interval \([t_6-t_7]\) in the three-phase are the characteristic intervals corresponding to the above-resonance region. The magnetic current will continue to increase linearly until the next resonant state occurs.

According to resonant state analysis, time-domain method can be used to describe the status of each subinterval in a half cycle. To generalize the analysis, normalized variables are defined as follows:

- Circuit Variable | SYMBOL | Normalized Variable
---|---|---
Switching frequency | \(f_s\) | \(f_s = f_s/f_r\)
Single-phase voltage | \(V_{\text{in}}(t)\) | \(V_{\text{in}}(t) = V_{\text{in}}(t)/V_n\)
Three-phase voltage | \(V_{\text{in}}(t)\) | \(V_{\text{in}}(t) = \frac{3V_{\text{in}}(t)}{V_n}\)
Single-phase current | \(I_{\text{in}}(t)\) | \(I_{\text{in}}(t) = I_{\text{in}}(t)Z_{\text{in}}/V_n\)
Three-phase current | \(I_{\text{in}}(t)\) | \(I_{\text{in}}(t) = 3I_{\text{in}}(t)Z_{\text{in}}/2V_n\)
Single-phase output power | \(P_{\text{out}}\) | \(P_{\text{out}} = P_{\text{out}}Z_{\text{out}}/V_{\text{in}}^2\)
Three-phase output power | \(P_{\text{out}}\) | \(P_{\text{out}} = 3P_{\text{out}}Z_{\text{out}}/4V_{\text{in}}^2\)
Voltage gain | \(n\) | \(n = n_{\text{out}}/V_n\)
Inductance ratio | \[L_1/L_2\] | 
Characteristic impedance | \(Z_{\text{char}} = \sqrt{L_1/C_2}\) |
equations. The state of Stage P and N can be expressed as

\[
\begin{align*}
\{ m_{P,N}(\theta) = [m_{C,P,N}(0) - 1 \pm M] & \cos(\theta) \\
+ jLr_{P,N}(0) & \sin(\theta) + 1 \mp M \\
jLr_{P,N}(\theta) = [1 \mp M - m_{C,P,N}(0)] & \sin(\theta) \\
+ jLr_{P,N}(0) & \cos(\theta) \\
m_{Lm,r,P,N}(\theta) = & \pm M \\
jLm_{l,r,P,N}(0) = & jLm_{l,r,P,N}(0) \pm Ml \theta \\
j_{out,p,N}(\theta) = & \left| jLr_{P,N}(\theta) - jLm_{l,r,P,N}(\theta) \right|
\end{align*}
\]

The state of stage O can be expressed as:

\[
\begin{align*}
\{ m_{C,O}(\theta) = [m_{C,O}(0) - 1] & \cos(k\theta) \\
+ jLr_{O}(0) & \sin(\theta) + 1 \\
jLr_{O}(\theta) = & k[1 - m_{C,O}(0)] \sin(\theta) \\
+ jLr_{O}(0) & \cos(\theta) \\
m_{Lm,o,O}(\theta) = & \frac{1}{1 + l} [1 - m_{Lm,o,O}(\theta)] \\
jLm_{o,O}(\theta) = & jLm_{o,O}(\theta) \\
j_{out,o}(\theta) = & 0
\end{align*}
\]

According to the equations of stage P, N, and O, \((2/3)V_{in}\) is used to normalize variables to acquire the stage of \((2/3)P\), \((1/3)P\), \((2/3)N\), \((1/3)N\), \((2/3)P\), \((1/3)P\), \((2/3)N\), and \((1/3)N\) in three-phase LLC. Besides, the stage of O-P and O-N can be expressed as:

\[
\begin{align*}
\{ m_{C,O-P,N}(\theta) = [m_{C,P,N}(0) - 1 - \frac{3}{4}M] & \cos(\theta) \\
+ jLr_{P-N}(0) & \sin(\theta) + 1 + \frac{3}{4}M \\
jLr_{O-P,N}(\theta) = [\frac{1}{2} - \frac{3}{4}M - m_{C,P,N}(0)] & \sin(\theta) \\
+ jLr_{P,N}(0) & \cos(\theta) \\
m_{Lm,o-P,N}(\theta) = & \frac{3}{4}M \\
jLm_{o-P,N}(\theta) = & jLm_{o-P,N}(0) \pm \frac{3}{4}Ml \theta \\
j_{out,o-P,N}(\theta) = & \left| jLr_{P,N}(\theta) - jLm_{l,r,P,N}(\theta) \right|
\end{align*}
\]

Energy is transmitted to load in P, N, \((2/3)P\), \((1/3)P\), \((2/3)N\), and \((1/3)N\). Thus, the normalized output power is given by

\[
\begin{align*}
P_{on,p} = & \frac{M}{V} \int_{0}^{\phi_{p,0}} \left[ jLr_{P}(\theta) - jLm_{p}(\theta) \right] d\theta \\
P_{on,n} = & \frac{M}{V} \int_{0}^{\phi_{n,0}} \left[ jLm_{n}(\theta) - jLr_{N}(\theta) \right] d\theta
\end{align*}
\]

According to the corresponding operation mode, the time domain equations are listed in the order of appearance. If the corresponding voltage and current are solved, additional constraints need to be added. Waveform symmetry can be used as a limiting condition for solving equations. The initial value and half-period value of voltage and current in one cycle should be opposite to each other. The value of the capacitor voltage and inductor current cannot be changed suddenly. This restriction always occurs at the critical moment when the subinterval changes. At the end of the interval P (including

\((2/3)P\) and \((1/3)P\), \((2/3)N\) and \((1/3)N\)) and N\((2/3)N\) and \((1/3)N\), the resonant current and magnetizing current must be equal to enter the next interval. Finally, if there is another state after the O interval, the magnetizing inductor voltage must resonate to the clamped state. The corresponding time-domain analytical formulas can complete the solution of LLC resonant by adding additional constraints.

### III. 6-ELEMENT BOUNDARY GAIN DESIGN METHOD

It is difficult to design and debug the LLC resonant converter with a wide gain range. There are two parameters that reflect the output characteristics of LLC: voltage gain \(M\) and output power \(P_o\). These two characteristics are determined by transformer turns ratio \(n\), inductance ratio \(l\), characteristic impedance \(Z_o\) and resonant current at switching moment \(jLr(\xi)\).

#### A. TRANSFORMER TURNS RATIO \(n\)

When the converter has a wide output voltage, the turns ratio \(n\) directly affects the distribution of the below-resonance region and above-resonance region. The voltage gain is 1 and corresponding output voltage \(V_o\) is only determined by \(n\) when the operating frequency is the resonant frequency. When output voltage \(V_o > V_r\), the converter operates in below-resonance region. When \(V_o < V_r\), the converter operates in above-resonance region. In the entire output voltage range, maximum gain \(M_{max}\) and minimum gain \(M_{min}\) is

\[
\begin{align*}
M_{max} &= \frac{nV_{out,max}}{V_{in,min}} \\
M_{min} &= \frac{nV_{out,min}}{V_{in,max}}
\end{align*}
\]

According to the resonant state and time domain model, load-gain curves of the single-phase LLC and three-phase LLC in below-resonance region are respectively shown in Fig.5(a) and (b). \(P_{on}\) is the normalized output power under different output power. The boundary between the most commonly used PON mode and PO mode is selected for reference.

Compared to three-phase LLC, as shown in figure.5, single-phase LLC has wider operating frequency range and stronger voltage gain capability. The three-phase LLC has a higher minimum frequency \(f_{min}\), making the operating frequency range narrow and the gain capability poor. In the meanwhile, if the slope of the load curves in the low-frequency region is too large, it may cause loop instability. Therefore, in order to make full use of the gain capability of single-phase LLC, the lowest output voltage is set as the starting point for design. High-voltage power batteries usually start charging at full power when the voltage is 330V.

So, \(V_o = 330V\) is set as the starting point for design. The maximum turns ratio of single-phase and three-phase
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FIGURE 5. Operation range of the below-resonance region (a) Single-phase LLC, (b) Three-phase LLC.

LLC is

\[
\begin{align*}
\eta_{\text{max, single}} &= \frac{V_{\text{in, max}}}{V_{r, \text{single}}} \\
\eta_{\text{max, three}} &= \frac{V_{\text{in, max}}}{V_{r, \text{three}}}
\end{align*}
\]  

(6)

where \(V_{r, \text{single}} = V_{\text{in, min}}, V_{r, \text{three}} = 330\)

**B. INDUCTANCE RATIO \(l\) AND CHARACTERISTIC IMPEDANCE \(Z_o\)**

Inductance ratio \(l\) and characteristic impedance \(Z_o\) have a significant impact on converter performance. PON and PO are the two most suitable operating modes for maximum load. However, PON mode ends in N stage and the primary resonant current will be decrease rapidly, which is unacceptable in high current converter applications. We can use boundary curve of PO and PON as the design boundary of PO mode. The boundary conditions of single-phase and three-phase LLC can be given as

\[
\begin{align*}
M_{B, fn} &= f_{\text{min}} \\
M_{\text{max, full}} &= \frac{1}{3} M
\end{align*}
\]  

(7)

where \(\theta_{o, \text{end}}\) is the end of the O subinterval.

\(Z_o\) is proportional to \(P_{\text{on}}\). Therefore, as long as the maximum load curve falls above the boundary curve, it is guaranteed that all resonant states will not operate in PON mode.

**C. RESONANT CURRENT WHEN \(\theta = \zeta j_{Lr}(\zeta)\)**

Zero voltage switching (ZVS) is the most significant advantage of LLC resonant converter. The converter may cause ZVS failure due to improperly resonant tank parameters. The effective value of the resonant current determines the efficiency of the converter. The resonant current at the moment of switching \(j_{Lr}(\zeta)\) must be large enough to charge

FIGURE 6. Boundary gain at different \(l\) in single-phase and three-phase LLC.

\(l\) directly determined the boundaries of the operating modes. \(M_{B, fn=f_{\text{min}}}\) is defined as the boundary gain at the lowest frequency. With the help of numerical calculation software, we can get the boundary gain at different value of \(l\) as shown in Fig.6.

According to Fig.6, we can determine the maximum inductance ratio \(l_{\text{max}}\) in the design of LLC resonant. \(M_{\text{max, full}}\) is defined as the peak gain at full load when the output voltage is maximum. The maximum value of \(l\) occurs when \(M_{B, fn=f_{\text{min}}} = M_{\text{max, full}}\). Different \(l\) values will have corresponding \(Z_o\) that meets the gain requirements. From the comparison of the curves in Fig.6, the single-phase LLC provides better gain capability than the three-phase LLC. The three-phase LLC resonant converter will often choose a smaller transformer turns ratio to improve the previously poor gain capability.

Lower inductance ratio and higher characteristic impedance values leads to a bigger input impedance, which will effectively reduce the circulating current and conduction loss. In the case of satisfying the output voltage and output power, the inductance ratio \(l\) should be as small as possible to improve the transmission efficiency. The smaller value of \(l\) also means that the slope of the load curve in the above-resonance region is smaller. As the output voltage or the load continues to decrease, the converter cannot output the specified voltage by increasing the frequency. In that condition, the converter will adopt the burst mode to achieve specified voltage. Using burst mode in advance will cause excessive output ripple. Therefore, the choose of LLC resonant tank parameters is a comprehensive design between minimizing the conduction loss and maintaining the desired gain range.
and discharge the corresponding parasitic capacitors. When the current is insufficient, high-frequency oscillation occurs because soft switching is not implemented. Therefore, the value of \( j_{Lr}(\zeta) \) is very critical when designing the resonant parameters of LLC.

The most excellent operating area is PO and the \( j_{Lr}(\zeta)-f_n \) curves of different normalized output power are shown in Fig.7.

As shown in Fig.7, the current in PON mode is insufficient to meet the boundary current (minimum current) required by ZVS. Therefore, the curves of the PON and the adjacent PON area are regarded as unsuitable operating conditions.

Although the PO mode is the most suitable operating state, the resonant converter will have insufficient current when the curve falls near the PO/PON boundary. It is easy to find from Fig.7 that all load curves are concentrated at the resonant frequency point, which happens to be the P mode operation. When the normalized output power value is 0.35 higher than the PO/PON boundary, \( j_{Lr}(\zeta) \) increases monotonically with decreasing frequency. In a single-phase LLC resonant converter, ZVS can be realized to avoid high frequency oscillation as long as \( j_{Lr}(\zeta) \) is greater than boundary current.

In the three-phase LLC resonant converter, the PO/PON boundary and each curves are converged at the resonant frequency point, which happens to be the P mode operation. When the resonant converter operates at the resonant frequency, the voltage gain \( M \) is equal to 1 and \( j_{Lr}(\zeta) \) can be solved by

\[
j_{Lr}(\zeta) = j_{Lm}(\zeta) = \frac{M\zeta}{2} = \frac{l\zeta}{2} \quad (8)
\]

When the normalized output power value is 0.35 higher than the PO/PON boundary, \( j_{Lr}(\zeta) \) increases monotonically with decreasing frequency. In a single-phase LLC resonant converter, ZVS can be realized to avoid high frequency oscillation as long as \( j_{Lr}(\zeta) \) is greater than boundary current.

In the three-phase LLC resonant converter, the PO/PON boundary and the resonant current under different load conditions are shown in Fig.8.

Similar to single-phase LLC, the PO/PON boundary and each curves are converged at the resonant frequency. It is worth noting that the mode boundary increases monotonically with decreasing frequency. As long as all operating curves of the resonant converter are above the mode boundary, there can be enough current to achieve ZVS.

D. CONDUCTION LOSS ANALYSIS

Since the conduction loss is the majority of the overall loss in LLC resonant converter, it is particularly important to minimize the conduction loss in the design process. With the equal on-state resistance value, the magnitude of resonant current and magnetizing current directly reflect the magnitude of conduction loss. When the converter operates in above-resonance region and below-resonance region, there are no simple expressions for the RMS resonant and magnetizing current. According to fig.7 and fig.8, the resonant current at the resonant frequency can reflect the overall current trend, which can also reflect the level of conduction loss. When the operating frequency is the resonant frequency, the RMS of the resonant current and magnetizing current are given by

\[
\begin{align*}
  I_{rRMS} &= \frac{\pi}{2\sqrt{2}} \sqrt{\frac{P_{on}}{(nV_o)^2} + \frac{(nV_o)^2}{1 - l^2}Z_r^2} \\
  I_{mRMS} &= \frac{\pi}{2\sqrt{3}} \frac{nV_o}{Z_o} \\
\end{align*}
\]

Among the design variables, the RMS of the resonant current and magnetizing current are only related to the inductance ratio \( l \). For a given load condition, the RMS currents will be lower if \( l \) decreases, and so does the conduction loss. The smaller characteristic impedance \( Z_o \) also means the smaller conduction loss. Inductance ratio \( l \) and characteristic impedance \( Z_o \) always appear correspondingly. Once \( l \) is determined, \( Z_o \) is also determined. Therefore, \( l \) is a variable parameter that directly reflects the conduction loss. When designing resonant parameters, we should reduce \( l \) to reduce conduction loss.

In order to speed up the process of debugging the parameters of the resonant tank, the relationship between the parameters in the LLC resonant converter is shown in Fig.9.

Fig.9 not only serves as a guide for independent LLC design, but can also be used for rapid secondary design for different battery voltage ranges of the same power. The
The parameters reflecting the output characteristics of the resonant converter are: voltage gain \( M \) and normalized output power \( P_{on} \). Therefore, the change of the load curve determined by \( Z_0 \) also determines the voltage gain \( M \) and \( jL_r(\zeta) \) of the converter. When other variables are consistent, the larger \( Z_0 \) means the smaller \( M \) and the smaller \( jL_r(\zeta) \). The inductance ratio \( l \) determines the boundary of operating mode, which means that it affects both the overall level of the voltage gain and the value of \( jL_r(\zeta) \). The transformer turns ratio \( n \) directly determines the maximum and minimum gain of the converter, which also means it affects the distribution of the above-resonance region and below-resonance region.

According to the relationship of 6 elements, the proposed design procedure is shown in Fig. 10. Where the line labeled ① is single-phase LLC and the line labeled ③ is three-phase LLC.

The design requirements of the resonant converter need to be provided at the beginning of the design procedure. At the end of the design procedure, the transformer turns ratio \( n \), inductance ratio \( l \) and characteristic impedance \( Z_0 \) will be obtained. Resonant inductor can be calculated by \( L_r = Z_r/2\pi f_r \). Resonant capacitor can be calculated by \( C_r = 1/2\pi f_r Z_0 \). Magnetizing inductor can be calculated by \( L_m = L_r/l \).

IV. EXPERIMENTAL RESULTS

The proposed method is designed for LLC parameters and the results of single-phase and three-phase LLC are verified by prototype. The unoptimized time-domain model design method uses resonant current as the design reference for ZVS. However, the resonant current can only ensure that the current at the resonant frequency is sufficient. As shown in Fig. 7, when the resonant converter works in the below-resonance region, the switching current \( I_{sw} \) may not be able to meet the needs to discharging the parasitic capacitors of MOSFETs. The unoptimized design results are \( L_r = 8mH \), \( C_r = 290nF \) and \( L_m = 55mH \). The design results are tested on the three-phase LLC experimental platform and the corresponding ZVS experimental waveform is shown in Fig. 11.
The three-phase drive waveform in Fig. 11 has a phase difference of 120 degrees. From the measured waveforms of drain-source voltage $V_{ds,Q2}$, there is an oscillation when Q2 is just turned off. This oscillation is precisely caused by insufficient switching current and can cause additional switching losses. The design results of the 6-element boundary gain design method are $L_r$ is 10mH, $C_r$ is 272nF and $L_m$ is 60mH. The increased inductance ratio and the corresponding characteristic impedance can also meet the output requirements. The experimental waveforms at the highest output voltage and the lowest output voltage are shown in Figure 12. Compared with the previous design results, the drain-source voltage waveform has no obvious oscillations under the two extreme conditions.

Before the arrival of each drive signal $V_g$, the drain-source voltage close to zero. Through the test waveforms in Figure 12, the characteristic of zero voltage turn-on can be realized in each cycle.

In addition, we test three resonant states of the resonant converter with the proposed method. The maximum output power of single-phase LLC is 3.3kW and the output voltage range is 230-430V. According to the proposed method, single-phase calculation results: $L_r$ is 40mH, $C_r$ is 70nF and $L_m$ is 300mH. The maximum output power of three-phase LLC is 6.6kW and the output voltage range is 240-420V. Three-phase calculation results: $L_r$ is 10mH, $C_r$ is 272nF and $L_m$ is 60mH.

**A. ABOVE-RESONANCE REGION**

When the operating frequency is greater than the resonant frequency, the LLC resonant converter works in the above-resonance region (buck region). The experimental waveforms of single-phase LLC and three-phase LLC in this area are shown in Fig. 13.

In Fig. 13 (a), the phase of the magnetizing inductor voltage leads the input voltage, which is one of the characteristics of the above-resonance region. LLC resonant converter works in NP mode and the resonant current is sufficient in this mode to achieve ZVS. According to the parameters designed by the proposed method, the drain-source voltage waveform is smooth and there is no obvious resonance, which shows that soft switching can also be better realized in single-phase LLC. All test waveforms of single-phase LLC achieve the expected results. In three-phase LLC, the resonant current cannot be directly measured because the resonant inductor and transformer are of patch structure. The three-phase LLC experimental waveforms selects the drain-source currents of Q2, Q4 and Q6. The actual test current waveform is half of the...
real resonant current and the other half is a symmetrical waveform. The phase relationship of the three test waveforms is consistent with the theory. Due to the three-phase LLC topology, the resonant current is more abundant at the switching moment. When the load continues to lighten, the frequency continues to increase to 250kHz and the converter will enter burst mode.

B. BELOW-RESONANCE REGION

When the operating frequency is less than the resonant frequency, the LLC resonant converter works in the below-resonance region (boost region). The experimental waveforms of single-phase LLC and three-phase LLC in this area are shown in Fig.14.

As shown in Fig.14(a), single-phase LLC operates in PO mode. It is formally ensured that the resonant current can still be sufficient because the magnetic inductor is added to the resonance before the switching moment. It is obvious that the magnetizing inductor voltage lags behind the input voltage in phase, which is also a typical characteristic of the boost region. Also at full load, the three-phase resonant converter works at the maximum output voltage. As shown in Fig.14(b), the resonant current has more resonant states in the cycle. At this time, the operating frequency is also the lowest.

FIGURE 15. Operating waveforms at resonant frequency (a) single-phase LLC \( V_{\text{in}} = 380V, V_{\text{out}} = 280V \) and \( P_0 = 2165W \) (b) three-phase LLC \( V_{\text{in}} = 390V, V_{\text{out}} = 350V \) and \( P_0 = 6600W \).

If the frequency is lower, it may enter the capacitive region, which is extremely dangerous for the resonant converter. Therefore, it is extremely important to accurately design the boost region. From the experimental results in boost region, the design results meet the expected requirements very well.

C. RESONANT FREQUENCY

When LLC resonant converter operates at resonant frequency, the experimental waveforms of single-phase LLC and three-phase LLC at this frequency are shown in Fig.15.

At resonant frequency, the input voltage of the resonant tank and magnetizing inductor voltage are equal in phase and amplitude. Capacitor voltage phase lags behind resonant current by 90 degrees. The resonant current waveforms changes sinusoidally. As shown in Fig.15, the resonant current is ahead of the input voltage in phase. Only in this way, the current will return to the resonant tank to charge and discharge the parasitic capacitors.

The parameters of the resonant tank also greatly affect EMC. The raised frequency point is often related to the switching frequency. Unreasonable parameter design may directly lead to the failure of the test and must use shielding layer or other methods to suppress noise. Therefore, a reasonable design of resonant parameters is very necessary. The test results are shown in Fig.16.
Whether it is single-phase or three-phase LLC, the efficiency range and there is no unnecessary switching oscillation. Ensure that the switching current is sufficient in the whole output range. When the switching current changes to the critical value of the ZVS design, the oscillation caused by insufficient switching current will increase the switching loss. We use the design results of unoptimized design method as a reference to single-phase and three-phase LLC resonant converter is proposed. 6 elements that affect the performance of the converter are analyzed one by one and the relationship between them is provided to guide the design. The 6-element boundary gain design method can reduce the cumbersomeness caused by multiple cycles and improve the practicality of the time domain method. The research and calculation of current $\dot{I}_{L}(t)$ can ensure ZVS in the entire output range. In addition, the proposed method can reduce the operating range of the boost region to obtain a lower inductance ratio. Experimental results are captured on LLC lab prototypes and the overall efficiency is improved with the proposed approach. Especially in the high output voltage (below-resonance region), the efficiency value can even be increased by about 0.5%.

**V. CONCLUSION**

A comprehensive LLC design method that can be applied to single-phase and three-phase LLC resonant converter is proposed. The experimental test uses the standard of CISPR22 CLASSB. From the test results of single-phase and three-phase LLC, both the peak value and the average value have a certain margin. The maximum value of noise basically occurs near the switching frequency of LLC, which also affirms the importance of accurate resonant parameters.

The resonant converter outputs the full range of voltage for efficiency testing and the test results are shown in Fig.17. Compared with other design methods, the time domain model has higher design accuracy. The design results can usually meet the requirements of a wide gain range. The calculation method of the unoptimized time domain model does not notice the change of the switching current over the entire output range. When the switching current changes to the critical value of the ZVS design, the oscillation caused by insufficient switching current will increase the switching loss. We use the design results of unoptimized design method as a comparison efficiency test. The optimized design result can ensure that the switching current is sufficient in the whole range and there is no unnecessary switching oscillation. Whether it is single-phase or three-phase LLC, the efficiency of the proposed method has been obviously improved. When the converter outputs high voltage, low current means low conduction loss, so the efficiency will be improved overall. The overall efficiency of single-phase LLC is above 97%. The overall efficiency of three-phase LLC is above 96%.

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