A modified efficiency-oriented optimal method for three-phase interleaved LLC resonant converter in plug-in hybrid electric vehicle battery chargers

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Abstract
Three-phase interleaved LLC resonant converters are preferably employed to realize high efficiency and high power density in the application of On-board Charger. The parameter design of LLC resonant converter with wide range output has always been rigorous and difficult. Compared with the common time-domain model, a high-accuracy time domain analysis is adopted to analyze the interaction between three phases in detail. This method improves the accuracy of the design results by adding the interaction analysis. According to the operation mode analysis, transformer’s turn ratio, resonant inductor, magnetic inductor and resonant capacitor are optimized based on the design principle of minimizing the conduction loss. Finally, the modified efficiency-oriented optimal method of wide output range LLC resonant converter is proposed and its design results is validated by a 6.6 kW prototype. The efficiency of the design using proposed method is significantly improved than the results of the common time-domain model.

1 | INTRODUCTION

With the development of plug-in hybrid electric vehicles (PHEVs) and pure electric vehicles (EVs), high-efficiency and safe charging of batteries is one of the hotspots of current research [1–4]. According to the complex charging profile shown in Figure 1, there are three stages in the whole charging process, recovery region, bulk charge and constant voltage (CV) charge [5–8]. The initial charging power is about 2800 W. The recovery region will exist if the battery experiences a deep discharge before charging. In this stage, the charging current is not too large and the power is less than rated power. The output power continues to increase as the battery voltage rises. When charging to the threshold of the recovery region, it will enter the bulk charge. The charging equipment has the ability to output the maximum power 6.6 kW when the voltage is between the threshold and the maximum value in bulk charge. Subsequently, the CV charge is applied when the battery rises to the highest voltage or reaches 70%–90% of its capacity. In CV stage, the voltage remains constant and the charging current is much smaller than in the other two stages. The output power also drops sharply at this stage. The charging process ends when SOC (State of Charge) reaches 1. According to the battery charging characteristics mentioned above, an on-board charger (OBC) is installed on the electric vehicle to meet the charging requirements.

The typical chargers consist of an ac–dc converter for active power factor correction and dc–dc converter connected to the battery. The complex charging function of OBC is mainly adjusted by dc–dc stage, so people put forward higher requirements for a series of indicators such as power density, efficiency and electromagnetic interference (EMI) of dc–dc converter. Since the LLC resonant converter has the soft switching characteristics in the full frequency range, it is preferably employed as dc–dc stage to reduce switching losses and EMI [9–11]. Unfortunately, the large output ripple current of single-phase LLC resonant
Multi-phase LLC resonant converter can effectively reduce current ripple and current stress on the rectifier diodes. As the number of phases in the interleaved parallel of the converter increases, the topology and drive circuit will be more complicated. Therefore, three-phase interleaved LLC as shown in Figure 2 is considered the most suitable multi-phase resonant converter in OBC applications [14–16]. Due to the inclusion of multiple resonant elements and complex resonant states, it is difficult to analysis and design, especially for three-phase interleaved LLC. The interleaved topology is achieved either by means of adding auxiliary circuits or with the cost of losing soft switching. Moreover, the interleaved structure between the three phases have low input and output ripple. This is why the size of filter capacitor is much less from the device point of view. As for the design method of resonant converter, the first harmonic approximation (FHA) analysis can provide concise mathematical formulas to describe the relationship between frequency and voltage gain in Ref. [17]. For a charger with an output voltage of 240 to 420 V, the frequency range needs to be very large to meet a wide voltage range. Due to the neglect of the high harmonic components, the design results of this method are not applicable when the operating frequency is far away from the resonant frequency. In order to improve the accuracy of design results, many scholars have proposed a time-domain analysis method. In Refs. [18,19], the operation mode of LLC resonant converter is analysed in detail and the peak gain placement method is introduced. Although this method is accurate and effective, the selection of its limiting conditions is not suitable for the application of OBC. The method in Ref. [20] may result in insufficient resonant current to realize soft switching. In Ref. [21], the analysis of mode boundaries and distribution can guide the design of resonant tank well. Unfortunately, the transformer’s turn ratio is not optimized. In addition, the interaction between three phases is also very important for the design of three-phase LLC, which is usually ignored in previous studies. In order to improve the accuracy of the design results, the modified efficiency-oriented optimal method (MEOM) proposed in this study analyses the operation of three-phase LLC in detail and optimizes the resonant parameters.

The study is organised as follows: Section 2 introduces the three-phase time domain model and analyses above-resonance region and below-resonance region respectively. Section 3 gives the optimization methods of inductance ratio, characteristic impedance and provides the design process. According to the design results of the proposed method, experimental test results are placed in Section 4.

2 | THREE PHASE LLC ANALYSIS AND PROPOSED TIME-DOMAIN MODEL

Figure 2 shows the topology of three-phase LLC resonant converter proposed. The circuit structure on the primary side consists of three switching legs, three single-phase resonant tanks with Y-connection structure. The transformer is regarded as the combination of the magnetizing inductance and the ideal transformer. The three switching legs switches at the same frequency with 120° phase-shift. The phase difference of two MOSFETs in each phase is half a resonance period. And there is also a dead time of about 100 ns at the switch alternate moment to realize the soft switching process. The phase difference of two MOSFETs in each phase is half a resonance period. And there is also a dead time of about 100 ns at the switch alternate moment to realize the soft switching process. The phase difference of two MOSFETs in each phase is half a resonance period. And there is also a dead time of about 100 ns at the switch alternate moment to realize the soft switching process. The phase difference of two MOSFETs in each phase is half a resonance period. And there is also a dead time of about 100 ns at the switch alternate moment to realize the soft switching process. The phase difference of two MOSFETs in each phase is half a resonance period. And there is also a dead time of about 100 ns at the switch alternate moment to realize the soft switching process. The phase difference of two MOSFETs in each phase is half a resonance period. And there is also a dead time of about 100 ns at the switch alternate moment to realize the soft switching process.
inductor $L_r$, magnetic inductor $L_m$ and high-frequency capacitor $C_r$. The single-phase equivalent circuit of LLC resonant converter is shown in Figure 3. The way to get $R_{eq,ac}$ actual load is converted to the input of the uncontrolled rectifier. According to the transformer’s turns ratio, the load is reflected to the primary side of the transformer.

The voltage state of $L_m$ is the core of time domain model. Three typical states P(Positive), N(Negative) and O(Off) exist in LLC half-bridge resonant converter (Figure 4). The states of P and N happen when $L_m$ is always clamped by the output voltage in the forward or reverse direction. The state of O appears when D1, D2 turn OFF. Magnetic inductor $L_m$ participates in the series resonance with $L_r$ and $C_r$. These three states basically include all the resonant states of the half-bridge LLC. In half switching cycle, there are six major operation modes (PN, PON, PO, OPO, NP, and NOP), which are used to named according to different combinations of up to three subintervals.

In LLC half-bridge resonant converter, there are two levels of input voltage: 0 V and $V_{in}$. As shown in Figure 5, four levels exist in three-phase LLC resonant converter. Compared with LLC half-bridge resonant converter, three-phase LLC resonant converter has five similar states ($\frac{2}{3}P, \frac{2}{3}N, \frac{1}{3}P, \frac{1}{3}N$ and O) and two additional states (O-P and O-N). The stages and corresponding the voltage of magnetic inductor is shown in Table 1, where $n$ is transformer turns ratio.

Since the voltage and current waveforms change periodically, the resonant states only needs to be analyzed in half a period. There are up to three subintervals in Half-bridge LLC resonant converter, which can be arranged by the appearance order. In comparison, three-phase LLC resonant converter has more intervals, where intervals P and N in half-bridge LLC are divided into five subintervals as shown in Table 2. In steady state, some intervals may or may not occur and there are at most 17 subintervals.

Three-phase LLC resonant converter handle a wide range output voltage and load by adjusting the frequency and exist a resonant frequency $f_r$ defined as:

$$f_r = \frac{1}{2\pi\sqrt{L_rC_r}} = \frac{1}{2\pi\omega_o}$$

(1)

The normalized time variable $\theta$ and the value of half period $\xi$ are given by

$$\begin{cases} \theta = \omega_o t \cr \xi = \omega_o T_s \end{cases}$$

(2)

Characteristic impedance is given by

$$Z_o = \sqrt{\frac{L_r}{C_r}}$$

(3)

We would like to normalize the voltage, current and power to make them easier in description. Lowercase letters $m$ and $j$ are used to denote normalized voltage and current. In the design of a wide output voltage range, the range of input voltage only comes from the output ripple of PFC. Therefore, $(2/3) V_{in}$ should be chosen as the base voltage and the current variables are normalized by $(2/3) V_{in}/Z_o$.

There are some assumptions and explanations that need to be provided first.

1. All rectifier diodes ignore their turn-on voltage drop
2. The input and output voltages are constant during operation

FIGURE 3 AC equivalent circuit of LLC resonant converter

FIGURE 4 LLC half-bridge resonant converter

FIGURE 5 Input voltage comparison of single-phase LLC and three-phase LLC

\[
\begin{align*}
\theta &= \omega_o t \\
\xi &= \omega_o T_s / 2
\end{align*}
\]
3. Power transformers are seen as a parallel combination of magnetic inductance and ideal transformer.
4. This study does not consider the effects of tolerance in device parameter values between three phases.

During a wide range frequency variation, wide range output voltage can be obtained from three-phase LLC resonant converter due to the ability of variable voltage gain. The dc voltage gain of three-phase LLC is given by

\[ M = \frac{nV_{\text{out}}}{V_{\text{in}}} \]  \hspace{1cm} (4)

When the resonant converter operates above the resonant frequency, the phase of magnetic inductor is ahead of the input voltage of resonant tank as shown in Figure 6. In a half switching period, \( 0 \leq \theta \leq \zeta \), four subintervals \( \frac{1}{3}N \), \( \frac{2}{3}P \), \( \frac{2}{3}P \) and \( \frac{1}{3}P \) appear in sequence.

For the \( \frac{2}{3}P \) and \( \frac{2}{3}N \), the input voltage of resonant tank is \( \pm \frac{2}{3}V_{\text{in}} \) and the voltage across magnetic inductor is clamped by \( \pm \frac{2}{3}V_{\text{out}} \). The normalized capacitor voltage \( m_{C,\frac{2}{3}P/N} \) and resonant tank current \( j_{L,r} \) can be obtained by writing two second-order circuit equations and the output current is the absolute value of the difference between magnetizing inductance current \( j_{L,m} \) and resonant tank current \( j_{L,r} \). The expressions of resonant state are given by

\[ m_{C,\frac{2}{3}P/N}(\theta) = [m_{C,\frac{2}{3}P/N}(0) - 1 \pm M]\cos(\theta) + j_{L,\frac{2}{3}P/N}(0)\sin(\theta) + 1 \mp M \]

\[ j_{L,\frac{2}{3}P/N}(\theta) = [1 \mp M - m_{C,\frac{2}{3}P/N}(0)]\sin(\theta) + j_{L,\frac{2}{3}P/N}(0)\cos(\theta) \]

\[ m_{L,m,\frac{2}{3}P/N}(\theta) = \pm M \]

\[ j_{L,m,\frac{2}{3}P/N}(\theta) = j_{L,m,\frac{2}{3}P/N}(0) \pm M * I * \theta \]

\[ j_{\text{out},\frac{2}{3}P/N}(\theta) = \left| j_{L,\frac{2}{3}P/N}(\theta) - j_{L,m,\frac{2}{3}P/N}(\theta) \right| \]  \hspace{1cm} (5)
where, \( m_{C_{1}P/N}(0) \), \( j_{L_{1}P/N}(0) \), and \( j_{L_{m1}P/N}(0) \) represent the initial normalized variables of \( \frac{1}{2}P \) and \( \frac{1}{2}N \) subintervals.

For the \( \frac{1}{2}P \) and \( \frac{1}{2}N \), the input voltage of resonant tank is \( \pm \frac{1}{2}V_{\text{in}} \) and the voltage across magnetic inductor is clamped by \( \pm \frac{1}{2}nV_{\text{out}} \). The expressions of resonant state are given by

\[
\begin{align*}
    m_{C_{1}P/N}(\theta) &= [m_{C_{1}P/N}(0) - \frac{1}{2} \pm \frac{1}{2}M \cos(\theta)] + j_{L_{1}P/N}(0) \sin(\theta) + \frac{1}{2} \mp \frac{1}{2}M
\\
    j_{L_{1}P/N}(\theta) &= \frac{1}{2} \mp \frac{1}{2}M - m_{C_{1}P/N}(0) \sin(\theta) + j_{L_{1}P/N}(0) \cos(\theta)
\\
    m_{L_{m1}P/N}(\theta) &= \pm \frac{1}{2}M
\\
    j_{L_{m1}P/N}(\theta) &= j_{L_{m1}P/N}(0) \pm \frac{1}{2}M \ast l \ast \theta
\\
    j_{\text{out}1P/N}(\theta) &= \left| j_{L_{1}P/N}(\theta) - j_{L_{m1}P/N}(\theta) \right|
\end{align*}
\]

where \( m_{C_{1}P/N}(0) \), \( j_{L_{1}P/N}(0) \), and \( j_{L_{m1}P/N}(0) \) represent the initial normalized variables of \( \frac{1}{2}P \) and \( \frac{1}{2}N \) subintervals.

In order to obtain the corresponding resonant state, there are some other conditions to be observed. First, the inductor current and capacitor voltage cannot be changed suddenly. Therefore, the value of variables should be same at boundary moment of adjacent subintervals. Second, the resonant tank current and magnetizing current must be equal when the polarity of magnetizing inductor voltage is reversed. Third, according to the symmetry of the waveform, the initial value of normalized voltage and current should be opposite to the value of the half period. Besides, the energy is only transmitted to the load when magnetizing inductor voltage is clamped by output voltage. The normalized output power is given by

\[
P_{\text{oa}} = \frac{P_{o}}{(\frac{2}{3}V_{\text{in}})^{2}}Z_{o} = \frac{nV_{\text{in}}T_{o}}{(\frac{2}{3}V_{\text{in}})^{2}}Z_{o} = \frac{M}{\zeta j_{\text{out}}}
\]
where \( j_{\text{out}} \) is the sum of the effective values of the output currents of each subinterval. After adding these conditions, sufficient number of equations can be obtained to solve the initial value of normalized voltage and current.

When the operating frequency moves to below-resonance region, the operation mode is more complex and discontinuous conduction mode as shown in Figure 7. Similar to single-phase LLC, three-phase LLC also has stage O in below-resonance region. The resonant tank current and magnetizing current remain equal in this subinterval. Because the magnetizing inductor participates in the series tank’s resonance, there is sufficient current in the resonant tank to ensure ZVS in this region. The secondary side current of this phase is zero and no energy flows to the load. The normalized equations of subinterval O describing the circuit states are

\[
\begin{aligned}
\begin{cases}
 m_{\text{C}, O}(\theta) &= m_{\text{C}, O}(0) - \frac{1}{2} \cos(k\theta) + \frac{j_{\text{L}, o}(0)}{k} \sin(k\theta) + \frac{1}{2} \frac{3}{4} M \\
 j_{\text{L}, O}(\theta) &= \frac{j_{\text{L}, o}(0)}{k} \sin(k\theta) + j_{\text{L}, o}(0) \cos(k\theta) \\
 m_{\text{L}, o}(\theta) &= \frac{1}{l} \left[ 1 - \frac{3}{4} - m_{\text{L}, o}(0) \right] \\
 j_{\text{L}, o}(\theta) &= j_{\text{L}, o}(0) \\
 j_{\text{out}, O}(\theta) &= 0
\end{cases}
\end{aligned}
\]

(8)

From the waveform diagram of Figure 7, there are two special subintervals that exist before the input voltage level change. The current situation on the secondary side is shown in Figure 8. In the interval \([0, \beta]\), the magnetizing inductor is first clamped by \( \frac{1}{2} n V_{\text{out}} \). Since the phase difference between the three phases is 120°, the phase C circuit just enters subinterval O at the moment \( \beta \). The existence of interval \((\alpha, \beta)\) is caused by the action of phase C, so it can be defined as O-P stage. The O-N stage occurs when the magnetizing inductor is clamped reversely. In the subinterval O-P, the rectifier diode \( D_{o5} \) and \( D_{o6} \) are turned off and the secondary side current flows only in phase A and phase B. The magnetizing inductor voltage of phase A is clamped by \( \frac{1}{2} n V_{\text{out}} \). The normalized equations of subinterval O-P and O-N can be described by

\[
\begin{aligned}
\begin{cases}
 m_{\text{C}, O}(\theta) &= \left[ m_{\text{C}, P}(0) - \frac{1}{2} + \frac{3}{4} M \right] \cos(\theta) + j_{\text{L}, P}(0) \sin(\theta) + \frac{1}{2} \frac{3}{4} M \\
 j_{\text{L}, O}(\theta) &= \left[ \frac{1}{2} + \frac{3}{4} M - m_{\text{C}, P}(0) \right] \sin(\theta) + j_{\text{L}, P}(0) \cos(\theta) \\
 m_{\text{L}, o}(\theta) &= \frac{3}{4} M \\
 j_{\text{L}, o}(\theta) &= j_{\text{L}, o}(0) + \frac{3}{4} M \ast l \ast \theta \\
 j_{\text{out}, O}(\theta) &= |j_{\text{L}, P}(\theta) - j_{\text{L}, o}(\theta)|
\end{cases}
\end{aligned}
\]

(9)

In addition to the four conditions mentioned before, there are extra conditions that to be met. First, the time points appearing \((0, \alpha, \beta, \gamma, \delta, \epsilon, \zeta)\) in below-resonance region can be expressed as

\[
\begin{aligned}
\begin{cases}
 \alpha &= \frac{1}{3} \zeta - (\zeta - \epsilon) = \epsilon - 2 \frac{1}{3} \zeta \\
 \beta &= \frac{1}{3} \zeta \\
 \gamma &= \frac{2}{3} \zeta - (\zeta - \epsilon) = \epsilon - \frac{1}{3} \zeta \\
 \delta &= \frac{2}{3} \zeta 
\end{cases}
\end{aligned}
\]

(10)

where \((\zeta - \epsilon)\) is the length of the subinterval O.

In this operation mode, the magnetizing current and resonant current are the same at the end of the subinterval when entering the subinterval O. Therefore, at the moment \( \epsilon \), the condition is given by

\[
\begin{aligned}
 j_{\text{L}, o}(\epsilon) &= j_{\text{L}, o}(\epsilon)
\end{aligned}
\]

(11)

These steady-state solutions in this operation mode \([m_{\text{C}, o}(0), j_{\text{L}, o}(0), j_{\text{L}, o}(0), \epsilon, \zeta, M]\) can be solved by the state equations and conditions. Unfortunately, these equations will lead to transcendental equations and have no closed-form solutions.
switching instants, the primary resonant tank current will be decrease rapidly, resulting in insufficient current to achieve ZVS. At the same time, the secondary side has output current. The rectifier diodes have reverse recovery loss. Therefore, the operation mode shown in Figure 6 is preferable state in below-resonance region. The boundary gain curve can be regarded as reasonable design boundary.

According to Equation (7), the relationship between the normalized output power and the characteristic impedance is proportional when the load power and input voltage are constants. The gain curve corresponding to different characteristic impedances is show in Figure 10, and it is obvious that the gain curve is getting closer to the boundary gain curve as the characteristic impedance increases. The increase of characteristic impedance also means the decrease of circulating current. Therefore, in the case of meeting the gain requirement, the characteristic impedance value should be increased as much as possible. The boundary conditions can be given as

\[ m_{\text{loss}}(\theta_{\text{end}}) = \frac{1}{3} M \]  

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

By adding boundary conditions, the boundary gain can be directly derived, which is determined only by the inductance ratio. The boundary gain curves for different inductance ratios are shown in Figure 11. It is visible that the higher value of \( I \) results in a larger gain capability within a given frequency range. Therefore, the boundary gain at the minimum operating frequency also increases as the value of \( I \) increases. When solving equations with numerical software, we can use \( I \) as a loop variable to obtain boundary gain at the minimum operating frequency. \( M_{Bf_{\text{min}}} \) denotes the boundary gain at the minimum operating frequency and \( M_{Bf_{\text{min}}} - I \) relationship curve is instructive for parameter design. \( M_{\text{max,full}} \) denotes the max gain at full load. Considering the boundary gain may cause the resonant current to be insufficient at the switching instants, the full load gain-frequency curve should be kept at a little

**FIGURE 9** Gain-frequency mode boundaries of single-phase LLC

Numerical calculation tool such as MATLAB or 1stopt can be used to obtain numerical solutions.

3 | THE MEOM AND DESIGN PROCEDURE

Under the given normalized power, normalized gain curves for various loads are obtained by solving the equations based on the aforementioned analysis. In the OBC applications, designers usually focus on voltage gain or power, which makes it difficult for designers to map specific design requirement to the time-domain model. In the single-phase LLC resonant converter, the major operation modes in below-resonance region can be divided by adding boundary conditions as shown in Figure 9. For instance, the boundary of PO and PON mode is that the normalized magnetizing inductor voltage of the PO mode is \(-M\) at the end of the O subinterval. In three-phase LLC, subinterval \( \frac{1}{3} N \) will appear after subinterval O when the load continues to increase or the operation frequency decreases. If the resonant converter enters into \( \frac{1}{3} N \) in advance at

**FIGURE 8** Current in the secondary side of the converter. (a) Secondary current waveform. (b) Current variation in two stage
distance from the boundary gain curve. We can use the condition \( M_{\text{max, full}} = 0.02 \) as the method of selecting the inductance ratio \( l \). After determining \( l \), the power boundary of the corresponding mode boundary is also determined. As long as the maximum output power curve is limited to the expected operating mode area, then all the operating states of the converter will not exceed this mode. The relationship between normalized resonant current, gain and frequency under full load is shown in Figure 12. Obviously, the normalized resonant current is sufficient to achieve ZVS.

The inductance ratio \( l \) and characteristic impedance \( Z_0 \) are the two most important indicators affecting the electrical characteristics of the LLC resonant converter. It is worth mentioning that greater gain capability is at the expense of increasing circulating current by reducing input impedance. In short, the means of optimizing resonant parameters can be achieved by reducing the value of \( l \) or increasing the value of \( Z_0 \).

According to Equation (4), the transformer turns ratio determined the output voltage at the resonant frequency when the input voltage is fixed. \( V_{\text{out,f=f}} \) denotes the output voltage.
at the resonant frequency. The higher $V_{\text{out},f=fr}$ means that the minimum voltage gain is very small, which is contrary to the poor buck capability of LLC resonant converter. If $V_{\text{out},f=fr}$ is lower, a larger inductor ratio is required to satisfy the larger maximum voltage gain. However, larger inductance ratio leads to lower overall efficiency. According to the charging curve in Figure 1, the battery voltage is above 330 V for full power charging. So it is suitable to choose 330 V as $V_{\text{out},f=fr}$. It is worth noting that the transformer turns ratio usually needs to be rounded, so the voltage selection can be slightly higher than 330 V.

Based on the above analysis, the design procedure can be outlined in seven steps.

Step 1: The initial design value of the transformer turns ratio is calculated by

$$n = \frac{V_{\text{in,min}}}{V_{\text{out},f=fr}} \quad (13)$$

where $V_{\text{out},f=fr}$ is 330 V.

In this case, the converter can ensure that it operates in below-resonance region when the output voltage is above 330 V. Since the primary and secondary sides of the transformer must be integers, the initial calculated turns ratio value needs to be decrease to meet the design requirements.

Step 2: Calculate the maximum gain according to Equation (13)

Step 3: According to the relationship between the inductance ratio and minimum frequency gain given in Figure 11, the value of inductance ratio can be obtained. The curve of Figure 11 needs to be obtained with the aid of numerical calculation software. Boundary gain is 0.02 less than maximum gain.

Step 4: Determine the boundary power value according to the obtained $I$ value with the help of numerical calculation software.

Step 5: The relationship between the normalized output power and the characteristic impedance is proportional. Reduce the $Z_o$ value based on the boundary power.

Step 6: When the maximum gain at full load meets the design requirements, the $Z_o$ value is finally obtained.

Step 7: The turns ratio, inductance ratio and characteristic impedance are the final results.

As shown in Figure 13, this article provides a design procedure for three-phase LLC resonant converter with wide range of output voltage. The optimal parameters of $n$, $I$ and $Z_o$ can be obtained at the end of the design procedure. Finally, the resonant inductor, magnetic inductor and resonant capacitor can be solved by

$$\begin{align*}
L_r &= \frac{Z_o}{2\pi f_r} \\
C_r &= \frac{1}{2\pi f_r Z_o} \\
L_m &= \frac{L_r}{T}
\end{align*} \quad (14)$$

Among all the losses of LLC, conduction loss often accounts for about 80%. The resonant current directly determines the conduction loss. For example, the resonant current flows through the on-state resistance of the MOSFET to cause conduction loss. The on-state resistance $R_{ds}$ can be acquired from datasheet. The normalized root mean square (RMS) resonant current $I_{\text{RMS,n}}$ is given by Equation (15)

$$I_{\text{RMS,n}} = \frac{\pi}{2\sqrt{2}} \sqrt{P_{\text{on}} + \frac{l}{(1-l)^2}} \quad (15)$$

The relationship of $I_{\text{RMS,n}}$ and $I$ is show in Figure 14.
It appears that for a given load condition $P_{on}$, the RMS currents will be lower if $I$ decrease, and so does the conduction loss. For a certain $P_{on}$ and $I$, the conduction loss will be lower with a greater characteristic impedance $Z_\circ$.

The design method proposed in this article is to reduce $I$ and increase $Z_\circ$ as much as possible, which is the method to minimize conduction loss.

4 | EXPERIMENTAL RESULTS

An experimental prototype with the output power 6.6 kW has been built to verify the proposed method. The DC output voltage is for 240 to 420 V and the maximum output current is 20 A. As shown in Figure 15, electronic load is used to simulate the battery and the high voltage DC source is used as the OBC voltage detection.

In the debugging of the resonant power supply, the resonant parameters often need to be debugged several times to get the optimal parameters. Therefore, as shown in Figure 16, the independent printed circuit board is used to place the resonant device, especially in the high-power applications.

The parameter design results of FHA, the single-phase time domain model and the proposed method are shown in Table 3. Three sets of design results were tested experimentally. $M$ represent the voltage gain of the converter, and its range directly reflects the voltage regulation capability of the output voltage. We found that non-optimal parameter values such FHA design results will lead to failure to meet the design requirements. Although the gain requirement can be met within a predetermined frequency range, a large value of $I$ will result in low efficiency. As shown in Figure 17, because the single-phase time domain model ignores the influence between the three phases, its design results ultimately fail to meet the gain requirements. Following the proposed method, the parameters of resonant tank are designed and chosen as follow: $n = 16:14$, $L_r = 10 \mu$H, $L_m = 60$ mH, $C_r = 272$ nF. The circuit components used in the prototype are provided in Table 4. In Figure 17, the black dots represent the experimentally measured voltage gain. The theoretical value and the experimental data value are basically the same, especially at low frequency. The influence of the parasitic parameters of the transformers and rectifier diodes on the circuit makes the experimentally measured data slightly higher than the theoretical value at high frequency.

![Figure 16](image1.png)

**FIGURE 16** Printed Circuit board with Independent resonant tank

| Parameter | Symbol | Value |
|-----------|--------|-------|
| Input voltage range | $V_{in,min} - V_{in,max}$ | 360–420 V |
| Output voltage range | $V_{out,min} - V_{out,max}$ | 240–420 V |
| Maximum output power | $P_{out,max}$ | 6.6 kW |
| Frequency range | $f_{min} - f_{max}$ | 65–250 kHz |
| Resonant frequency | $fr$ | 96.5 kHz |
| Transformer turns ratio | $n$ | 16:14 |
| Resonant capacitance | $C_r$ | 68 nF × 4 |
| Resonant inductance | $L_r$ | 10 mH |
| Magnetic inductance | $L_m$ | 60 mH |
| Filter capacitor | $C_m$ | 16 μF |

![Figure 17](image2.png)

**FIGURE 17** Experimental testing of different design results

**TABLE 3** Parameter design results for different methods

| Resonant inductance ($\mu$H) | FHA | Single-phase time domain model | Proposed method |
|-----------------------------|-----|--------------------------------|----------------|
| Magnetizing inductance ($\mu$H) | 70  | 60                            | 60             |
| Resonant capacitance ($nF$) | 180 | 340                           | 272            |

**TABLE 4** Design specification and some parameter for LLC.
Partial test waveforms of minimum output voltage 240 V and maximum output voltage 420 V under full load are shown in Figures 18 and 19. It should be noted that the output ripple of voltage and current are the waveforms after removing the DC component. When the output voltage is the lowest voltage, the resonant converter operates in above-resonance region,

**FIGURE 18** Experimental test waveform at $V_{in} = 365$ V, $V_{out} = 240$ V, and $P_o = 2.9$ kW. (a) GS-DS waveform, (b) output voltage ripple, and (c) output current ripple.

**FIGURE 19** Experimental test waveform at $V_{in} = 390$ V, $V_{out} = 420$ V, and $P_o = 6.6$ kW. (a) GS-DS waveform, (b) output voltage ripple, and (c) output current ripple.
which has natural ZVS capability. The GS and DS waveforms of the MOSFET corresponding to the lowest voltage are shown in Figure 18a. It can be found that the region of ZVS is very obvious. Due to the poor buck capability of the LLC resonant converter, the minimum voltage requirement cannot be met by increasing the frequency. The burst mode is usually used to adjust the output voltage. The introduction of this method will make the ripple voltage larger. Due to the three-phase parallel interleaving structure, the ripple current can also be kept low even when the filter capacitor is small as shown in Figure 18c. In general, the design requirements are basically met at the lowest output voltage. Improper resonant parameters may lead to insufficient gain or efficiency. As shown in Figure 19a, ZVS can still meet to achieve higher transmission efficiency when the converter is maximum output voltage. In addition, the output ripple voltage is only 12.5 V and the ripple current is only 1.18 A.

Figure 20 shows the operating waveforms at maximum and minimum output voltage. The resonant components are all patch components. The ds current of Q1 is the half-cycle waveform of the resonant current. Therefore, we use the ds current of Q1 to describe the resonant current. The resonant current and resonant capacitor voltage basically reached the expected waveforms. The input voltage waveform of the resonant tank is similar to the principle waveform at the lowest output voltage. However, the input voltage waveform of the resonant tank oscillated due to the presence of parasitic parameters and limited test conditions. The resonant current in the three phases differ by 120° in phase. The ds currents of Q1, Q3, and Q5 are measured to verify the resonant current state. As shown in Figure 21, the resonant current state meets the design expectations.

The efficiency test results of the converter under the highest output voltage and the lowest output voltage are shown in Figure 22. Because the design results of the FHA analysis method do not meet the gain requirements, they are not added to the efficiency test comparison. According to the structural design, the resonant components are all patch components. The current of resonant tank cannot be measured, and thus the efficiency cannot be directly measured. In order to ensure the rigour of the efficiency test method, the efficiency comparison test of the whole OBC machine is provided under the premise.
of only changing the parameters of the resonant tank. It is obvious that the proposed method improves efficiency performance. It can be seen that the converter efficiency maintains above 89% from 10% load to full load. By compared with the above-resonance region, the below-resonance region not only has the ability of the primary ZVS, but also the secondary-side rectifier diode has the ability of ZCS. In addition, the corresponding operating current is small at high output voltages, and the main loss is the conduction loss under the premise of implementing soft switching. As shown in Figure 22b, efficiency at output voltage of 420 V is slightly higher than the efficiency at output voltage of 240 V and the peak efficiency is 95.88%. The experimental results show that the calculated parameters meet the requirements of the converter’s gain and efficiency at the same time.

5 | CONCLUSION

The three-phase interleaved LLC resonant converter applied PHEV battery charger has been analysed in detail, and a MEOM is proposed. When the operating frequency is farther away from the resonance frequency, the greater proportion of the O stage in the entire period, the greater impact on the accuracy of the time domain model. By comparing with traditional design methods, the MEOM proposed comprehensively considers the interaction between the three phases. Considering minimizing conduction loss, we give the optimal design method of transformer’s turn ratio $n$, inductance ratio $l$ and characteristic impedance $Z_w$. According to the analysis of the optimized results, the design method is accurate and effective. Debugging and development cycle time can be reduced, which is significant in three-phase LLC design. A 6.6 kW output prototype is built using the proposed method, which obviously improves transmission efficiency.

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