A Novel Design Method of LCC-S Compensated Inductive Power Transfer System Combining Constant Current and Constant Voltage Mode via Frequency Switching

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ABSTRACT
Inductive power transfer (IPT) is an attractive wireless power charging option in many applications such as electric vehicle biomedical devices and consumer electronics, etc. Constant current and constant voltage (CC-CV) charging profile is widely used for charging applications due to affecting the life and the reliability of the Li-ion battery. However, the IPT systems that can achieve both CC and CV modes with soft switching are not well studied. This paper presents an inductive power transfer system for wireless charging without a back-end converter, which achieves the required two-stage charging profile with high efficiency by keeping inverters’ soft switching operation. The characteristics of an LCC-S compensation topology have been investigated thoroughly. Two fixed resonant frequencies, whose analytical expressions have been derived, can be found to realize constant current output and constant voltage output respectively under zero-phase angle condition. Besides, an effective tuning method has been presented for zero voltage switching realization to reduce the switching loss. Moreover, a parameter design procedure has been summarized for the wireless charger with a simplified structure and high efficiency. Finally, the method is validated through experiments on a 3.5kW prototype realizing constant current and constant voltage outputs with a peak efficiency of 97.3%.

INDEX TERMS
Inductive power transfer, LCC-S compensation, constant current and constant voltage charging mode.

I. INTRODUCTION
IPT has received significant focus in recent years due to its advantages of safety and convenience. It has been used in fields such as wireless charging of mobile phone and electric toothbrushes in consumer electronics [1], and wireless supply to implantable devices in medical apparatus [2]. Furthermore, wireless battery charger has been designed for electric vehicle (EV) in higher power applications [3], [4].

In the above-mentioned applications, the lithium-ion (Li-ion) battery has been widely adopted due to their characteristic in terms of high energy density, compact size, and reliability [5], [6]. To charge the Li-ion batteries, there are at least two stages named constant current (CC) charge and constant voltage (CV) charge [7]. Generally, the battery charging process starts with a CC mode and the battery voltage starts to rise. When the voltage rises to a specific level, the charging process is shifted to the CV mode. Subsequently, the charging process ends when the charging current decreases to a certain value. To prolong the battery life and ensure the reliability, the output voltage and current curve of the charger should be regulated accurately according to the battery charging profile, which is usually taken for granted but given less attention in conducted charging scenario. However, it is not easy to design an IPT based wireless charger that can implement a CC-CV charging profile due to the wide range of load variations.

In an IPT based wireless charging system, three components are commonly seen including high frequency (HF) inverter, the loosely magnetic coupler, and the compensation networks [8]. The compensation networks are especially
crucial to the overall performance because it forms the resonant tanks with the loosely coupled inductors to reduce the VA rating of the system while improving the power transfer capability. Additionally, the equivalent load of the battery varies throughout the charging process affecting the value of the input impedance, which will influence the performance of the entire system. By means of the compensation topology design, the input impedance can be resistive regardless of load thus achieving zero phase angle (ZPA). In this case, the reactive power within the resonant circuit is minimized during the entire charging process, which reduces the switching loss of the HF inverter.

In general, there are four basic capacitive compensation topologies, depending on the way of the compensation capacitors being connected, namely series-series (SS), series-parallel (SP), parallel-series (PS), and parallel-parallel (PP) [9]. High-order symmetrical compensation topologies formed by LCL and LCC networks are promoted due to their superior performance [10], [11]. Additionally, the topologies adopted high order networks on the primary side and basic capacitive compensations on the secondary side have also been studied, including LCC-S, LCC-P, etc. However, the above compensation topologies are usually designed to realize either constant-current or constant-voltage output characteristic [12], which cannot match the requirement of the ideal battery charging profile. To have the wireless charger achieve both the CC and CV mode output, quantities of efforts have been carried out in recent years. Generally, these methods can be roughly divided into three categories.

1) Hybrid Compensation topologies. Since a single topology seems difficult to implement both CC and CV mode, the combination of compensation topologies with different output characteristics draw much attention, especially considering the load-independent characteristic is preferable in both modes. For example, SS and SP can achieve CC and CV outputs respectively with specific parameters, while the PS and PP can achieve load-independent CC and CV modes respectively [13]. Based on the four basic topologies, Qu et al. [14] and Auvigne et al. [15] employ a combination of two of the four basic topologies such as SS combined with PS, SP combined with PP to perform battery charging. Furthermore, high order compensation topologies are also applied in hybrid topologies. SS combining with S-LCC in [16] and LCC-LCC combining with LCC-S in [17] are proposed to realize CC and CV modes. However, changes in the working topology will inevitably require the use of several switches, which causes the extra power loss as well as complicates the control strategy. Moreover, the introduction of a hybrid circuit on the receiving side will increase the complexity of the receiving system, increase the system cost as well as reduce the portability of the system, which is contrary to the lightweight principle of the receiving side.

2) Control schemes of back-end converter or active bridges. Usually, a DC-DC converter is adopted as the back-end stage of the charger to regulate the output voltage and output current to attain the required charging profile [21]. To reduce the number of the stages of the charging system, the phase shift control (PSC) scheme is designed and carried out by active bridges of the inverter and semi-active bridges of the rectifier to regulate the output characteristic [18]–[20]. Furthermore, variable frequency control (VFC) is proposed by the author of [22], [23] to fulfill the required output accommodating to the load. However, a back-end stage undoubtedly complicates the power circuit topology, while the complicated characteristics of the resonant tank call for the elaborated design of the PSC or the VFC scheme. Therefore, the compensation topology with both properties of load-independent CC and CV output is preferable [24], which simplifies the overall topology and the control scheme.

3) Dual-frequency switching control. As mentioned in [25], the aforementioned four basic topologies cannot implement both CC and CV modes while realizing ZPA at the same time due to limited parameter design freedom. Therefore, high order topologies have been investigated to meet this requirement. Qu and Chu [26] analyze the output characteristics of LCC-LCC by numeral calculation and realize the required load-independent CC and CV outputs at two ZPA frequencies. Nevertheless, the additional compensation capacitance and inductance introduced on the receiving side will inevitably increase the complexity and reduce the portability of the receiving side, which has the same shortcomings as the first method. To adapt to the principle of lightweight in the receiving side as well as fit the transfer characteristic, the authors of [27]–[29] realize the transformation of CC or CV mode by analyzing the characteristics of LCC-S by circuit equivalence. Analogously, the work by Lu et al [30], [31] analyzes the features of high order topologies and summarizes some of the topologies that can implement CC and CV mode. However, in Lu’s design method of achieving CC and CV outputs, the analytical expressions of the specifications and the design parameters do not directly link to the actual component due to the segmentation processing in the model. By way of illustration, under certain output characteristic conditions, since the relationship between self-inductance and coupling is unpredictable, the iterative coupling parameters may be impractical to implement. Meanwhile, although the ZPA constraint is handy for solving the high order resonant networks, the zero voltage switching (ZVS) is preferable for the MOSFET-based inverter. Due to the adoptions of the equivalent components to achieve parameter calculations [29]–[31], the specific ZVS tuning method are difficult to obtain, therefore, related research is still scarce. Since ZVS heavily affected the efficiency, it is necessary to explore it to improve the performance of the system. Furthermore, Table 1 is introduced in order to briefly show the difference between this work and the existing papers about LCC/S topology. It can be clearly show that this work can not only simplify the realization of CC and CV mode, but also proposes a tuning method in clearer way of ZVS which guarantees the high efficiency of system.

Focused on the LCC-S topology, aimed at solving the problem of the indecipherable relationship between
TABLE 1. Comparison between this work and existing papers about LCC-S system.

| LCC/S proposed in | [27] | [28] | [29] | This work |
|-------------------|------|------|------|----------|
| ZPA in CV mode    | YES  | NO   | YES  | YES      |
| ZPA in CC mode    | NO   | YES  | YES  | YES      |
| Analysis of ZVS   | YES  | YES  | NO   | YES      |
| Simplified system design | YES | YES | NO | YES |

CC-CV parameters and the unclear implementation conditions of ZVS, this paper promotes a parameter design procedure to achieve load-independent current output and load-independent voltage output while ZVS operation is guaranteed in a wide range. Firstly, the T-type and 7-type networks as two basic compensated units are introduced with their characteristics of load-independent output.

Secondly, the analytical expressions of achieving CC and CV mode for LCC-S topology have been derived respectively. Subsequently, the input impedance is analyzed for ZPA realization. Thirdly, the influence of various circuit elements on the input impedance angle is analyzed. A parameter design procedure has been proposed, which ensures the wireless charger operating at ZVS mode above 1/2 rated power in CC mode, and above 1/4 rated power in CV mode to achieve high efficiency in the entire charging process.

The rest of the paper is organized as follows. The modeling and theoretical analysis of load-independent current and voltage outputs of LCC-S compensation are mentioned in Section II. The proposed parameter design method is introduced in Section III. In section IV, both the simulation and experimental results of a 3.5kW prototype are present to validate the proposed method. Finally, Section V concludes this paper.

II. ANALYSIS OF CC AND CV MODE

A. T-TYPE AND 7-TYPE COMPENSATION NETWORKS

The T-type compensation networks as shown in Fig. 1 are widely used in circuit analysis due to the superior characteristics such as load-independent output voltage [32], [33]. According to KVL and ohm’s law, we can get the following formulas.

\[ G_v = \frac{U_{OUT}}{U_{IN}} = \frac{Z_3}{(Z_1 + Z_3) + \Delta/Z_R} \]  
\[ G_i = \frac{I_{OUT}}{I_{IN}} = \frac{Z_3}{\Delta + (Z_1 + Z_3) \cdot Z_R} \]

where \( \Delta = Z_1Z_2 + Z_1Z_3 + Z_2Z_3 \), \( G_v \) is the voltage gain and \( G_i \) is the transconductance gain. From (1) and (2) separately, the condition to get the load-independent \( G_v \) and \( G_i \) are as follows:

\[ \Delta = Z_1Z_2 + Z_1Z_3 + Z_2Z_3 = 0 \]  
\[ Z_1 + Z_3 = 0 \]

In the case of load-independent \( G_i \), the output current characteristic is independent of \( Z_2 \), therefore the network only consists of \( Z_1 \) and \( Z_3 \) which can be regarded as 7-type. When the input source is a sinusoidal voltage source and the above specific conditions are met, T-type and 7-type can achieve load-independent voltage output and load-independent current output respectively.

B. ANALYSIS OF LCC-S TOPOLOGY

Fig. 2 shows the LCC-S compensation topology. The circuit consist of a power supply, an inverter, a primary LCC topology and secondary S topology, a rectifier and a load network. The primary compensation network consists of resonant inductor \( L_1 \), and two resonant capacitor \( C_1 \) and \( C_1 \). The secondary compensation network is composed of resonant capacitor \( C_2 \). The magnetic coupler consist of \( L_1, L_2 \) and mutual inductance \( M \).

For simplicity, the equivalent circuit of LCC-S is shown in Fig. 3, which is based on First Harmonic Approximation...
Moreover, in order to facilitate the derivation of the CC mode, the CV mode, and the input impedance, the mutual inductance model is adopted. The mutual inductance can be expressed as

\[ M = k \sqrt{L_1 L_2} \]  

(5)

where \( k \) is the coupling coefficient. The first harmonic components, based on the equivalent circuit in Fig. 3 can be written as

\[ U_{IN} = \frac{4}{\pi} V_{IN} \sin(\omega t), U_{OUT} = \frac{4}{\pi} V_{OUT} \sin(\omega t + \theta) \]

\[ I_2 = \frac{2}{\pi} I_0 \sin(\omega t + \theta), R_{OUT} = \frac{8}{\pi^2} R_f \]  

(6)

\( V_{AB}, V_0, I_2 \) represents the value for input dc-link voltage of full bridge inverter, required voltage of battery, and required current of battery, \( \theta \) is the phase between \( U_{IN} \) and \( U_{OUT} \). \( R_f \) is the equivalent resistance of battery and \( R_{OUT} \) is the equivalent ac resistance of \( R_f \).

### C. IMPLEMENTATION OF CV MODE

The LCC network is usually designed as primary compensation for IPT system due to the simplicity of achieving a constant current through the primary coil. When the secondary coil is coupled, a stable induced voltage can be generated on the secondary side. Therefore, LCC-S topology is usually used for CV mode.

As shown in Fig. 4, the components marked in yellow are designed in specific resonance mode to meet the output characteristics. In addition, the parameters marked in red mean that they remain constant. It can be seen from Fig. 4(a) that in order to form a stable current through the primary coil, it means that \( L_{f1} \) and \( C_{f1} \) need to form a 7-type to achieve load-independent current output, the relationship between \( L_{f1} \) and \( C_{f1} \) can be obtained as

\[ j\omega_{CV} L_{f1} + \frac{1}{j\omega_{CV} C_{f1}} = 0 \]  

(7)

Then the relationship between \( I_1 \) and \( U_{IN} \) can be calculated as

\[ I_1 = \frac{U_{IN}}{j\omega_{CV} L_{f1}} \]  

(8)

It can be seen that \( I_1 \) is only controlled by \( U_{IN} \) when \( L_{f1} \) remain unchanged. Therefore, the equivalent controlled voltage source of the secondary side ‘\( j\omega_{CV} M_{I1} \)’ is a constant value depending on \( U_{AB} \). By applying Kirchhoff’s law to the Fig. 4(a), the voltage equation for the secondary circuit model can be written as

\[ j\omega_{CV} M_{I1} = j\omega_{CV} L_2 I_2 + \frac{I_2}{j\omega_{CV} C_2} + U_{OUT} \]  

(9)

It can be easily deduced from (9) that in order to make \( U_{OUT} \) independent of secondary parameters, \( L_2 \) and \( C_2 \) should satisfy

\[ j\omega_{CV} L_2 + \frac{1}{j\omega_{CV} C_2} = 0 \]  

(10)

When (10) is met, \( U_{OUT} \) is equal to the controlled voltage source and controlled by \( U_{IN} \). The total voltage gain can be calculated as

\[ G_c = \frac{U_{OUT}}{U_{IN}} = \frac{j\omega_{CV} M_{I1}}{U_{IN}} = \frac{j\omega_{CV} M \cdot u_{IN}}{U_{IN} \cdot j\omega_{CV} L_{f1}} = \frac{M}{L_{f1}} \]  

(11)

### D. IMPLEMENTATION OF CC MODE

The resonance conditions for achieving CC mode are illustrated in Fig. 4(b). Assuming that a constant load-independent current output on the secondary side has been realized, the value of the primary controlled voltage source ‘\( -j\omega_{CC} M_{I1} \)’ stays constant. In another word, the output voltage of the primary T-type compensation network is constant. In this case, according to (3) and (4), the components \( L_{f1}, C_{f1} \) and \( L_{e1} \) constituting the T-type should satisfy

\[ -\omega_{CC}^2 L_{f1} L_{e1} + L_{f1}/C_{f1} + L_{e1}/C_{f1} = 0 \]  

(12)

\[ j\omega_{CC} L_{e1} = j\omega_{CC} L_{1} + \frac{1}{j\omega_{CC} C_{1}} \]  

(13)

where \( L_{e1} \) is the equivalent to \( L_1 \) and \( C_1 \), defined as (13).

Therefore, it can be inferred from (1) and (12) that the controlled voltage source of the primary side ‘\( -j\omega_{CC} M_{I2} \)’ is a constant value depending on \( U_{AB} \). Obviously, secondary current \( I_2 \) is directly controlled by \( U_{AB} \). The total transconductance gain can be expressed as

\[ G_1 = \frac{I_{OUT}}{U_{IN}} = \frac{1}{(\omega_{CC}^2 L_{f1} C_{f1} - 1)j\omega_{CC} M} \]  

(14)
E. IMPLEMENTATION OF ZPA IN CC AND CV MODES

In order to achieve ZPA, the input impedance of the circuit needs to be calculated. The input impedance of the secondary network $Z_S$ can be derived as

$$ Z_S = j\omega L_2 + 1/j\omega C_2 + R_{OUT} \quad (15) $$

Equivalent impedance of the secondary side reflecting to primary side can be calculated as

$$ Z_R = \frac{(\omega M)^2}{Z_S} \quad (16) $$

Thus the overall input impedance of the circuit can be calculated as

$$ Z_IN = j\omega L_1 + \frac{1}{j\omega C_1} + j\omega L_1 + 1/j\omega C_1 + Z_R \quad (17) $$

As shown in [27], in the condition of $Z_1 = Z_2 = -Z_3$, the input impedance is purely resistive when the load is resistor. Obviously in CV mode, the secondary side reflection impedance is resistive according to (10), (15) and (16). Moreover, (7) also satisfies $Z_1 = -Z_3$, and the ZPA can be achieved by satisfying $Z_1 = Z_2$. Therefore, conditions for implementing ZPA in CV mode can be derived as

$$ \begin{align*}
\omega_{CV} & = 1/\sqrt{L_1 C_1} = 1/\sqrt{L_2 C_2} \\
\omega_{CV} & = j\omega_{CV} L_1 + j\omega_{CV} C_1
\end{align*} \quad (18) $$

As analyzed above, ZPA and CV output can be achieved by operating at a specific frequency $\omega_{CV}$ solved by (18).

However, no explicit expression of a specific frequency can be drawn directly from (12) and (13) for the CC mode to realize ZPA. In this case, the input impedance of the resonant circuit should be specially designed as resistive. To correlate the CC operating frequency with $\omega_{CV}$, $C_1$, $C_2$ can be expressed by $1/L_1 \omega_{CV}^2$, $1/(L_1 L_2) \omega_{CV}^2$, $1/L_2 \omega_{CV}^2$ according to (18). Meanwhile, the operating frequency that achieves ZPA in CC mode named $\omega_{CC}$ can be calculated from (12)(13)

$$ \omega_{CC} = \sqrt{\left(\frac{1}{L_1 C_1} + \frac{1}{L_2 C_2}\right)} $$

$$ \omega_{CC} = \sqrt{\omega_{CV}^2 \frac{L_1}{L_1 + L_1} - (L_1 - L_1)} \quad (19) $$

For a given primary side compensation parameter, the relationship between $\omega_{CC}$ and $\omega_{CV}$ can be expressed as

$$ \omega_{CC} = \omega_{CV} \sqrt{(\delta_{1,2} + (\delta_{1,2} + (1 - L_f/L_1)) / \delta_{1,2}} \quad (20) $$

where $\delta_{1,2} = L_f \pm \sqrt{L_f L_1}$. It is appeared that the load-independent CC output can be achieved at two frequencies, where

$$ \alpha_{1,2} = \sqrt{(\delta_{1,2} + (\delta_{1,2} + (1 - L_f/L_1)) / \delta_{1,2}} \quad (21) $$

The input impedance of CC mode can be calculated from (17) by submitting the $\omega_{CC}$ with $\omega_{CV}$ in (22), as shown at the bottom of the next page.

In order to achieve ZPA in the CC mode, the imaginary part of $Z_{inCC}$ needs to be zero. For case 1, $\omega_{CC1} = \alpha_1 \omega_{CV}$, substituting (20) into (22) and solving, when $\text{Im}(Z_{inCC}) = 0$, $L_f$ and $L_1$ should satisfy

$$ L_{f,1,1} = L_1 \frac{k^2}{(k - 1)^2} \quad (23) $$

For case 2, $\omega_{CC1} = \alpha_2 \omega_{CV}$ the condition that $\text{Im}(Z_{inCC}) = 0$ can also be obtained according to the above method. Consequently, the conditions that $L_{f,1}$ and $L_1$ should satisfy

$$ L_{f,1,2} = L_1 \frac{k^2}{(k + 1)^2} \quad (24) $$

Substituting (23) and (24) into (21) and simplifying, two possible cases are given as (25)

$$ \begin{align*}
\text{Case 1} & \quad \left\{ \begin{array}{l}
L_{f,1,1} = L_1 \frac{k^2}{(k - 1)^2} \\
\omega_{CC1} = \frac{1}{1 - k} \omega_{CV} \\
U_{OUT1} = \frac{(1 - k)^2}{k(1 - k)^2} L_2 U_{IN} \\
I_{OUT1} = \frac{(1 - k)^3/2}{k^2 j \omega_{CV} \sqrt{L_1 L_2}} U_{IN}
\end{array} \right. \\
\text{Case 2} & \quad \left\{ \begin{array}{l}
L_{f,1,2} = L_1 \frac{k^2}{(k + 1)^2} \\
\omega_{CC2} = \frac{1}{1 + k} \omega_{CV} \\
U_{OUT2} = \frac{(1 + k)^2}{k} L_2 U_{IN} \\
I_{OUT2} = \frac{(1 + k)^3/2}{k^2 j \omega_{CV} \sqrt{L_1 L_2}} U_{IN}
\end{array} \right. 
\end{align*} \quad (25) $$

It can be clearly seen from (25) that when $L_2$ and $L_1$ are close and $k$ is less than 1, the voltage gain of case

\[FIGURE 5. The distribution of the coupling coefficient and coupler parameters for $G_v$ and $G_t$ (a) case 1 (b) case 2.\]
TABLE 2. Specification of the IPT battery charger.

| Parameters | Discription | Rate |
|------------|-------------|------|
| P | Power rating | 3.5kW |
| V<sub>in</sub> | Input voltage | 330DC |
| V<sub>o</sub> | CV mode output voltage | 400V |
| I<sub>o</sub> | CC mode output current | 8.75A |

2 by \( G_{v,\text{case}2} \approx 1/k + 2 + k > 3 \) while \( G_{v,\text{case}1} \) can be less than 1. Similarly, the transconductance gain \( G_i \) of case2 is larger than that of case 1 under the same conditions. In addition, (25) indicates that if we define the equivalent load resistance at the CC/CV mode switching moment as \( R_{\text{equ}} = U_{\text{OUT}}/I_{\text{OUT}} \), the value of \( R_{\text{equ,case}2} \) is larger than that of case 1. Therefore, the above factors together with the actual requirements should be considered comprehensively in selecting these two cases.

F. COMPARISON OF THE TWO POSSIBLE CC DESIGN CASES

Take the application of the EV charger as an example, the \( G_v \) for the IPT system is usually below 1.5 while the \( G_i \) is below 0.1 [11], [12]. Therefore, \( G_v \) and \( G_i \) are set as 1.5 and 0.1 respectively as the design requirements for the cases study here. According to (25), the coupling coefficient of magnetic couplers, the square root of the self-inductance ratio, and the square root of the product of the self-inductance are selected as the benchmark for gain comparison. The \( G_v \) and \( G_i \) can be depicted as Fig. 5. Specifically, Fig. 5(a) illustrates that if case 1 is adopted, the targeted design value of \( G_v \) and \( G_i \) as highlighted lines can be obtained when the coupling coefficient is around 0.3. Fig. 5(b) illustrates that if case 2 is used, the expected value of \( G_i \) is found when the coupling coefficient is around 0.2. However, the expected value of \( G_v \) is not achievable since the minimum voltage gain is 4. Therefore, the case 1 is more suitable for the EV wireless charger while case 2 can be used in applications where high voltage gain is required.

G. VERIFICATION OF LOAD-INDEPENDENT CC AND CV MODES

To verify the required output characteristics, a typical IPT charging system with specifications listed as Table 2, has been taken as an example. All the parameters of the resonant tank can be calculated by the aforementioned design rules, which have been included in the captions of Fig. 6. In the case of varying frequency and different load, the curves of \( G_v \), \( G_i \), and phase angle of the input impedance \( Z_{IN} \) are drawn as Fig. 6(a), (b), and (c) respectively. It can be observed from Fig. 6(a) that \( G_v \) is a constant at the CV frequency of 85kHz regardless of the load variation, while ZPA achieved as Fig. 6(c). Similarly, the \( G_i \) is a constant at the CC frequency of 106.1kHz regardless of the load variation and achieves ZPA.

III. PARAMETERS DESIGN PROCEDURE

A. INFLUENCE OF VARIOUS PARAMETERS ON THE INPUT IMPEDANCE ANGLE

It has been proved that the converter operates in ZPA condition when the resonant circuit parameters are designed in full compliance with the above rules. However, to decrease the switching loss, zero voltage switching (ZVS) operation is preferable than ZPA operation for a MOSFET based inverter. That is to say, the input impedance of the resonant tank driven by the inverter should be inductive. Therefore, the influence of each parameter on the input impedance should be analyzed to have the whole circuit maintain inductive in the entire charging process.

The input impedance angle (\( \alpha \)) expression has been derived as (17). It is well-know that \( \alpha \) is equal to zero when the circuit is resistive. Besides, positive \( \alpha \) represents inductive input impedance, otherwise capacitive impedance.

For the sake of intuitively showing the impact of parameter variation, \( C^* \), \( L^* \) and \( k^* \) is defined to represent the original value of \( C \), \( L \) and \( k \) at resonance. Thus, the relationship between \( \alpha \) and \( C_2 \) in CV mode can be simplified.

\[
\alpha = \arctan\left(\frac{C_2/C_2^*-1}{\omega_{CV}C_2R_{OUT}}\right)
\]

Similarly, according to the relationship between \( \omega_{CV} \) and \( \omega_{CC} \) in case 1, the phase \( \alpha \) can be rewritten as

\[
\alpha = \arctan\left(\frac{1 - C_2/C_2^*}{\omega_{CV}C_2R_{OUT}\sqrt{1/(1-k)}}\right)
\]

Accordingly, the influence of resonant network parameters on \( \alpha \) is calculated and summarized in Table 3. It can be observed that when \( C_2 \) is larger than the resonance value, the circuit is always inductive in CC mode while remaining capacitive in CV mode. Therefore, adjusting \( C_2 \) cannot guarantee that the circuit remains inductive in the entire charging process. Besides, the impedance characteristic cannot be concluded directly from Table 3 when the capacitance of \( C_1 \) or \( C_1' \) is adjusted. Fortunately, increasing the value of \( L_1 \) can make the resonant tank inductive since the term of \( (L_1' - L_1^*) \) are identical in both CC and CV modes’ impedance expression. Therefore, ZVS can be realized for the entire charging process by tuning \( L_1' \).
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**FIGURE 6.** (a) $G_v$ of the IPT converter. (b) $G_i$ of the IPT converter. (c) Phase of $Z_{in}$. (When $L_1 = 252\mu F$, $L_2 = 244.6\mu F$, $L_{f1} = 75.7\mu F$, $C_{f1} = 44.54nF$, $C_1 = 20.23nF$ and $C_2 = 41.33nF$, $k = 0.3585$).

| TABLE 3. The relationship between compensation network parameters and $\alpha$. |
|---------------------------------------------------------------|
| Parameters | $\alpha$ in CV mode | $\alpha$ in CC mode |
| $C_1$ | $\alpha = \text{arctan}\left(\frac{(C_1^* - C_1)}{\alpha_{CV} R_{out}}\right)$ | $\alpha = \text{arctan}\left(\frac{(C_1 - C_1^*)/(2k-1) \times A}{(k-1)^2 R_{out} \alpha_{CV} L_1 C_1^* / \sqrt{1/(1-k)}}\right)$ |
| $C_2$ | $\alpha = \text{arctan}\left(\frac{C_2}{C_2^* - 1}\right)/\alpha_{CV} R_{out}$ | $\alpha = \text{arctan}\left(\frac{1 - C_2}{C_2^*}\right)/\alpha_{CV} R_{out}$ |
| $C_{f1}$ | $\alpha = \text{arctan}\left(\frac{(C_{f1}^* - C_{f1}^*)}{\alpha_{CV} M^2 R_{out}}\right)$ | $\alpha = \text{arctan}\left(\frac{(C_{f1}^* - C_{f1}) \times D}{\alpha_{CV} C_{f1} R_{out} k^2 (k-1)^3 / \sqrt{1/(1-k)}}\right)$ |
| $L_{f1}$ | $\alpha = \text{arctan}\left(\frac{(L_{f1} - L_{f1}^*)}{R_{out} L_{f1}^2 \sqrt{1/(1-k)}}\right)$ | $\alpha = \text{arctan}\left(\frac{L_{f1} - L_{f1}^*}{R_{out} \sqrt{1/(1-k)}}\right)$ |
| $k$ | $\alpha = 0$ | $\alpha = \text{arctan}\left(\frac{(k^2 - k^2) C_1}{k (k-1) / \sqrt{1/(1-k)} R_{out}}\right)$ |

$A = ((R_1^2 k^4 \alpha_{CV}^2 + (R_{out} - 3 L_2 \alpha_{CV}^2) k^3 + L_2^2 k^2 \alpha_{CV}^2 - 2 R_{out}^2 k + R_{out}^2) C_1 - (2k-1)(R_{out}^2 k - R_{out}^2 - L_2^2 k^2 \alpha_{CV}^2) C_1^*)$ $B = (-\alpha_{CV}^2 C_{f1} k^4 L_2^2 + R_{out}^2 C_{f1} / \alpha_{CV}^2 C_{f1}^2 - 2 R_{out}^2 / \alpha_{CV}^2 C_{f1}^*)$ $D = (L_2^2 k^2 \alpha_{CV}^2 C_{f1} + L_2^2 \alpha_{CV}^2 k^2 (C_{f1} - 2 C_{f1}^*) + R_{out}^2 k^2 C_{f1} + R_{out}^2 k (C_{f1} + C_{f1}^*) + R_{out} C_{f1}$)

**B. ANALYSIS ON THE IMPACT OF TUNING PARAMETERS ON SYSTEM CHARACTERISTICS**

In order to confirm the feasibility of the tuning method, the influence of parameter changes on the entire system is further analyzed. Firstly, based on the above-mentioned system parameters, the trend of the impedance angle of the system when each parameter is adjusted as shown in Fig. 7. It can be observed that in both CC and CV modes, a slight decrease of $C_1$ or the increase of $L_{f1}$ will make the whole circuit inductive no matter in $P_0$ or $0.5P_0$. This is basically consistent with the aforementioned analysis. Therefore, by adjusting $L_{f1}$, it is possible to satisfy ZVS in CC and CV modes.

Secondly, since the output voltage and current play a vital role in the system, the influence caused by varying parameters also needs to be considered. The normalized output current and voltage with varying normalized parameters are shown in Fig. 8. It is quite clear that the change of $C_1$ has an insignificant influence on the gain of voltage while causing great influence on the gain of the current. In contrast, the varying $L_{f1}$ has a small effect on the output voltage and current. Therefore, ZVS can be realized by adjusting $L_{f1}$ to ensure that the overall performance is not greatly affected. It should be noted that a very wide range of ZVS can be achieved by offsetting the $L_{f1}$ in a large value, and the transmission characteristics of the system will be affected non-negligibly. When considering the ZVS range, it needs to be related to the fluctuation of the transmission characteristics. It is essential to consider them together and make a trade-off before selecting the parameters.

**C. PARAMETER TUNING OF ZVS AND SYSTEM DESIGN PROCESS**

Specifically, to guarantee ZVS operation in both modes, the turn-off current of MOSFET must be large enough to discharge the junction capacitors within the dead-time, which can be expressed as follows: [34]

$$I_{OFF} \geq \frac{4C_{OSS} U_{AB, max}}{t_d}$$ (28)
TABLE 4. Compensation circuit parameter for the proposed system.

| Parameters                                      | Simulation      | Measurement     |
|------------------------------------------------|-----------------|-----------------|
| Wire diameter, N. of strands                    | 0.1 mm 1500     | 0.1 mm 1500     |
| Outer diameters and turns of the primary winding| 550 mm*500 mm   | 550 mm*500 mm   |
| Outer diameters and turns of the secondary winding| 550 mm*500 mm   | 550 mm*500 mm   |
| Air Gap                                         | 120 mm          | 120 mm          |
| Self-inductance of primary coil $L_1$           | 252 μH          | 246.38 μH       |
| Self-inductance of secondary coil $L_2$         | 244.6 μH        | 249.3 μH        |
| Coupling coefficient $k$                        | 0.3585          | 0.349           |
| Resonant frequency $f_{CV}$ and $f_{CC}$        | 85 kHz, 106.126 kHz | 85 kHz, 105.5 kHz |
| Primary compensation inductor $L_f$             | 78.7 μH         | 75.8 μH         |
| Primary parallel compensation capacitor $C_f$   | 44.54 nF        | 49.51 nF        |
| Primary compensation capacitor $C_1$            | 20.23 nF        | 19.97 nF        |
| Secondary compensation capacitor $C_2$          | 14.33 nF        | 14.06 nF        |

where $C_{oss}$ is the junction capacitance, $U_{AB,max}$ is the max input voltage, and $t_d$ is the dead time.

In order to analyze the current at the switching time more accurately, it is necessary to further calculate the influence of high-order harmonics. As stated in [11], for high-order harmonics calculation, the interaction between the primary and secondary sides can be ignored because the couplers can be regarded as a high-order filter. Moreover, the sum of the value of high-order harmonics at the switching time can be expressed as:

$$I_{OFF} - I_{OFF, high\_order} = I_{IN} \sin \alpha$$  \hspace{1cm} (30)

To sum up, a practical parameter design procedure is summarized as Fig. 9 for the LCC-S compensated IPT charging system to realize the CC charge with subsequent CV charge profile. First of all, the resonant frequency of the CV mode $f_{CV}$ is selected by considering the volume and the loss of the resonant components [3]. Next, the size of the magnetic coupler is affected by the parameter limitation in the targeted application. The value of $L_1$, $L_2$ and $k$ can be obtained by using electromagnetic field analysis software. Then $L_f$ and $f_{CC}$ can be determined by using (25). The value of $C_1$, $C_2$, and $C_f$ can be calculated by using (18) respectively. Then, an LCC-S compensated IPT charger can realize input ZPA.
with both output CC or CV mode by changing the operation frequency. Last but not least, by weighing the ZVS range as well as the output characteristic fluctuations, the final adjustment value of $L_{f1}$ is derived by equation (30).

### IV. EXPERIMENTAL VALIDATION

According to the proposed design procedure, the tuned value of $L_{f1}$ for ZVS operation can be calculated based on the previously designed IPT charger as shown in Fig. 9. Besides, a prototype has been built to verify the theoretical analysis of the expected load-independent CC and CV characteristics with ZVS realization. All the designed parameters and actual measurement parameters of the prototype can be founded in Table 4. When constructing the prototype, the current and voltage stress of the components can be obtained by (7, 10, 12, 25), and the selection of components is carried out accordingly to ensure the reliability of the system.

Varying frequency should be adjusted according to the charging mode in practical situation. Therefore, it is necessary to adopt the switching module to achieve the CC and CV mode transition. As shown in Fig. 10, the charging mode depends on the feedback signal of battery voltage $V_0$. When the battery voltage is lower than the preset voltage $V_{ref}$, 400V in this paper, the switching of the current mode is activated. Once the voltage of battery exceeds 400V, the current mode is switched OFF, and the voltage module is switched ON.

The experimental setup is shown in Fig. 11. The full-bridge inverter is composed of four MOSFETs (C2M0040120D), and the secondary rectifier uses fast-recovery diodes (C5D50065D). The output filter capacitor is 10 $\mu$F. Moreover, multiple capacitors in parallel are adopted to acquire the voltage and current rating of the LCC-S compensation tank. Texas Instruments Incorporated TMS320F28335 DSP is used to implement closed-loop control. The battery voltage and current are separately measured by LV-25P voltage sensor and HO 25-P/SP33 n current sensor. The sensed signal of the secondary side can be feedback wirelessly with no difficulty by using a mature module (NRF24L01). A resistive load is used to emulate the battery. Besides, the waveforms of the circuit are captured by Tektronix TBS 2000.

As mentioned in Table 2, an 8.75A charging current is required for CC mode until the output voltage reaches 400V for the fabricated prototype. Then 400V should be maintained in the whole CV mode. It means that the equivalent load resistance increases as the charging process proceeds. Fig. 12 shows the measured waveforms of input voltage $U_{AB}$, input current $I_{AB}$ and the output current $I_0$ during CC mode charge in 105.5kHz for two values of load resistance $R_r$ being 25$\Omega$ and 46$\Omega$. It can be seen that the output current is maintained at about 8.7A although the load is varying. When $R_r$ arrives at 46$\Omega$, $V_0$ is grown to 400V, and the charging mode needs to be switched to CV mode. Fig. 13 shows the frequency switching from $f_{CC}$ to $f_{CV}$, which means that the driving signal of H-bridge inverter switches from 105.5kHz to 85kHz. It can be seen that the output voltage $U_0$ and output current $I_0$ change slightly during the switching moment. The waveforms of $U_{AB}$, $I_{AB}$, and $U_0$ in CV mode are shown in Fig. 14(a) and (b) while the $R_r$ being 46$\Omega$ and 90$\Omega$. However, the charging voltage of the battery changes slightly in different $R_r$. Furthermore, the $V_{gs2}$ and $V_{ds2}$ are provided to clearly show the ZVS operation. It can be seen that the ZVS
is achieved in both CC and CV mode, given low switching stresses and high transfer efficiency.

The entire charging process is shown in Fig. 15. The experimental data are consistent with the simulation data, which proves the validity of the parameter design. The efficiency of the whole charging process measured by YOKOGAWA WT1800 is plotted in Fig. 16. It shows that the efficiency is increasing with the output power in CC mode while the efficiency has been above 97% in CV mode. Furthermore, the efficiency improved at the switching point due to the smaller current flowing through the resonant tank in CV mode as a result of impedance changing in varying frequencies.

Based on the above experimental results, the proposed method can achieve high efficient load-independent current or voltage characteristic output. In addition, a series of comparisons among the proposed method and recently related methods in the literature are made, and the results are listed in Table 5.

Firstly, compared with [17], [21], it is seen that the proposed method reduces the usage of active switches, avoids the introduction of a DC-DC transformer at the back end, therefore effectively reduces the complexity of system component is also different in CC and CV mode. As shown in Fig. 17, the higher primary current makes the ratio of copper loss and core loss larger in CC mode than CV mode.

TABLE 5. Comparison of the proposed method and other related work.

| Articles aiming at implementing CC and CV | [21] | [17] | [26] | [31] | [35] | [30] | This work |
|------------------------------------------|------|------|------|------|------|------|----------|
| Number of transmitter-side compensation components | 1    | 3    | 3    | 3    | 4    | 3    | 3        |
| Number of receiver-side compensation components | 1    | 4    | 3    | 3    | 1    | 1    | 1        |
| Number of DC-DC converters | 1    | 0    | 0    | 0    | 0    | 0    |          |
| Number of active switches | 5    | 6    | 4    | 4    | 4    | 4    |          |
| ZPA in CC and CV mode | YES  | YES  | YES  | NO   | YES  | YES  | YES      |
| Analysis of ZVS in CC and CV mode | NO   | YES  | YES  | NO   | NO   | NO   | YES      |
| Tolerance and rationality of system design | NO   | YES  | YES  | NO   | YES  | NO   | YES      |
| Max Power | 3.25kW | 3.3kW | 24W  | 3.5kW | 144W | 3.3kW | 3.5kW    |
| Peak efficiency | 88.05% | 91.8% | ≈93% | 92.9% | 94%  | 90.8% | 97.3%    |
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control. Secondly, the compensated components are significantly reduced, in this regard, which is beneficial to the lightweight of the receiver compared to [26], [31]. Thirdly, the proposed method clearly points out the parameter relationship between the gains and the couplers, which makes the design method simpler and more effective compared to [30]. Fourthly, different from [30], [35], the parameter tuning object and its numerical calculation method to realize ZVS are proposed, making the system work more efficiently. Last but not least, different from the articles listed above which operation efficiency in the entire CC-CV charging process is lower than 94%, the efficiency of the designed prototype maintains above 95% in a wide load range. In short, the proposed approach reduces the number of components required to achieve the CC-CV characteristics of the wireless charging system and simplifies the calculation process, thus reducing the cost and complexity of the system, at the same time, the conditions for achieving ZVS are proposed which improves the efficiency of the system charging process.

V. CONCLUSION

Focused on LCC-S compensation topology for wireless battery charging application, it is feasible to realize load-independent CC output and CV output characteristics respectively at different resonant frequencies. Usually, a specific resonant frequency defined by clear parameter constraints can be designed first for CV output mode with ZPA operation. Subsequently, two load-independent resonant frequencies can be found by analytical expressions for achieving CC output mode with ZPA operation. The two CC mode operation frequencies hold different characteristics in terms of voltage gain and transconductance gain, which needs to be selected properly according to the actual scenarios. Although the CC frequency is inconsistent with the usual resonance frequency, it ensures the realization of ZPA, which can also achieve high efficiency. Besides, it is found that the preferable ZVS operation for the MOSFET-based inverter can be achieved in a wide operating range by tuning the primary compensated inductance. A comprehensive parameter design procedure has been summarized for LCC-S compensation topology to have both load-independent CC and CV output while ZVS operation is guaranteed. The great performance of
the proposed parameter design method in its high efficiency and feasibility has been validated by experiments.

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