An LC/S Compensation Topology and Coil Design Technique for Wireless Power Transfer

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Abstract—Wireless power transfer (WPT) has attracted a lot of attention these years due to its convenience, safety, reliability, and weather proof features. First and foremost, the consistency of mutual model and T model of loosely coupled transformer (LCT) was deduced. The application scenarios of these two models were then concluded so as to choose suitable model in circuit analysis. Then, a new WPT compensation topology, which is referred to as LC/S compensation topology and consists of one inductor and two capacitors, is proposed. The constant-current-output (CCOut) characteristic of the newly proposed topology is analyzed in detail on the basis of the discussion about LC and CL resonant tank. The equivalent resistance of the rectifier, filter, and resistor circuit is also analyzed to simplify circuit analysis. Then, the current and voltage stress on each component and the system performance under imperfect resonant condition are studied with the help of MATLAB. The LCT is deliberately designed by the finite element analysis software ANSYS Maxwell as well because the coupling coefficient, primary, and secondary self-inductance have a significant impact on system efficiency, power level, and density. The LCT design approach employed in this paper can be extended to magnetic design of almost all WPT systems. Theoretical analyses are verified by both Pspice simulation and practical experiments. Practical output currents with transient loads show an excellent CCOut characteristic of LC/S compensation topology.

Index Terms—Constant-current-output (CCOut), LC/CL resonant tank, loosely coupled transformer (LCT), magnetic design, wireless power transfer (WPT).

I. INTRODUCTION

C ompared to conventional plug-in or brush and bar-contact-based power transfer method, wireless power transfer (WPT) has some inherent advantages, such as convenience, safety, reliability, and weather proof [1]. All these abovementioned advantages make various applications of WPT, including contactless power supplies for professional tools, contactless battery charging across large airgaps for electric vehicles, compact electronic devices, mobile phones, and medical implants [2].

WPT has been widely studied in recent 30 years. There are many research fields in WPT, such as compensation topology and circuit analyses, coil design techniques for large gap and misalignment tolerance, optimization for high efficiency, control methods, foreign object detection and safety issues. Among these fields, compensation network and circuit analyses play fundamental role since its determination of resonant frequency, power factor, output characteristics. According to the literature published, four basic compensation topologies, SS, SP, PS and PP, where the first S or P stands for series or parallel compensation of the primary coil and the second S or P stands for series or parallel compensation of the secondary coil, are mostly researched for various applications [3]–[10]. Taking SS compensation topology as an example, there are two ways to design the resonant capacitor. The first way is to design the capacitor to resonate with the leakage inductance [3], [6], [7], which could achieve a higher ratio between the active power and reactive power. The specific define of the ratio can refer to [11]. The other way is to resonate with the coil self-inductance [8]–[10], which could maximum the transferred power at a certain coil current. Generally, the original input of the WPT system is a dc voltage source. If the operation frequency is required to be some specific value, and the loosely coupled transformer (LCT) has been manufactured, which means that the LCT parameters cannot be changed, the maximum output power is determined, no matter which one of the four basic compensation topologies is chosen and no matter which way to design the resonant capacitor is adopted. Therefore, the maximum output power cannot be altered unless another LCT with different parameters is employed once the original input voltage and operation frequency of the system are determined. However, it is quite difficult to manufacture an LCT, which highly satisfy our requirements, making the system design hard.

Besides these basic compensation topologies, several novel compensation topologies with some quite good characteristics are proposed, such as LCC [1], LCL [2], SP/S [12], [11], S/SP [13]–[16], [17]–[21], and [22]–[26]. A new compensation topology called SP/S, which can transfer rated power even with high misalignment (up to 25% of the secondary coil width) is proposed in [12]. This paper focuses on misalignment tolerance because the application scenario is dynamic electric vehicle charge, where high misalignment is inevitable. Another novel compensation topology called S/PS, which overcomes some drawbacks of SS and SP is proposed in [13]. The experiment results have validated its advantage of zero phase angle (ZPA)
and insensitivity. However, similar to four basic compensation topologies, the maximum output power of S/SP compensation topology cannot be changed unless another LCT with different parameters is taken.

The design of a new unity-power-factor inductive-power-transfer pickup using an LCL tuned network for application in high-power system is proposed in [17]. An improved LCL compensation topology utilizing a coupling structure between the resonant inductors in the T-type network to mitigate the problems of LCL compensation topology is proposed in [18]. In fact, the LCL compensation is an ideal symmetrical T-type compensation network, which has been analyzed in detail in [25]. It is the essence of LCL compensation network. Both the double-sided LCL compensation network employed in [19] and the secondary LCL inductively coupled power transfer pickup adopted in [20] are compliant with the essence. The LCL network discussed in [2] and [21] is totally different from that in [17]–[20]. In [2] and [21], the LCL/P (primary LCL and secondary parallel) compensation topology can be divided into a primary series inductance, which is going to be optimized, and a PP compensation topology. The key point of these two literatures is the optimization of the primary series inductance under the premise of determined PP compensation parameters. They are not typical LCL compensation topologies.

The LCC compensation topology is widely researched and applied in last several years because it can easily achieve ZPA and ZVS, which means low volt-ampere (VA) rating and high system efficiency. Besides, with the tuning method presented in [22], the resonant frequency is irrelevant with the coupling coefficient between the primary and secondary coils and is also independent of the load condition, indicating the system can work at a constant switching frequency. Moreover, the system is of the characteristic of load-independent output current, which will simplify control circuit design. To reduce system size and improve its power density, Li et al. [23] integrated the additional inductors with the main coils. A new compensation circuit design procedure with the consideration of high-order current harmonics is elaborated in [24]. The inverter zero-current switching is achieved. An LCC compensation topology from the perspective of ideal symmetrical T-type compensation network is analyzed in [25]. It is very helpful to understand the essence of LCC compensation topology. Finally, LCC compensation topology has a lot of desirable characteristics. The only drawback is that it needs six compensation elements, which will increase the system size, cost, and power loss.

Except compensation network and circuit analyses, coil design is also very important because it is directly related to the coupling coefficient, which has a great impact on the power transferred by the WPT systems. In [27], the design and optimization methodology of a circular magnetic core is presented. In [28], two kinds of coil that are called unipolar coil and DD coil respectively are designed, optimized, and compared. However, it is still hard to design a DD-type LCT under specified size constraint since several coupling-related variables are still not discussed.

This paper proposes a new compensation topology, the diagram of which is shown in Fig. 1. The primary part of the compensation topology is a series inductor $L_1$ and a parallel capacitor $C_1$ while the secondary part is just a series capacitor $C_2$. We named the newly proposed compensation topology LC/S in terms of the nomenclature of LCC and SS compensation topology. LC/S compensation topology can easily achieve ZPA and ZVS simultaneously. Additionally, the resonant frequency is irrelevant with the coupling coefficient and the load condition. Further, the output current is load-independent, which is referred to as constant-current-output (CCOut) characteristic in this paper. In other words, all the main advantages of doubled-sided LCC compensation topology can be obtained by LC/S compensation topology whereas the latter only needs three compensation elements, resulting in smaller system size, higher power density, and lower cost. Compared to conventional SS compensation topology, it can free the design from the constraints imposed by the LCT parameters. The maximum output power of LC/S compensated system can be easily changed by altering the value of $L_1$ and $C_1$. In theory, the output power can reach infinity though the original input voltage and operation frequency is set to some specific value. All these characteristics will be further analyzed in Section III.

In Section II, the consistency of mutual inductance model and T model of LCT are deduced. The corresponding application scenarios of these two models are concluded as well. It is very helpful to choose a suitable LCT model in circuit analysis. In Section IV, the relationship between three coupling-related variables against coupling coefficient is discussed with the help of finite element analysis (FEA) software ANSYS Maxwell. Then, an LCT is designed with the size constraint of $340 \times 210 \times 100$ mm$^3$ under the guideline of former discussion. To verify the correctness of the theoretical analyses before, Psipce simulations are carried out in Section V, whereas corresponding practical experiments are conducted in Section VI. Both the simulation and experiment results have shown high consistency with theoretical analyses. Finally, the main points of the study is concluded in Section VII.

II. APPLICATION SCENARIOS OF MUTUAL INDUCTANCE MODEL AND T-TYPE MODEL OF LCT

There are two well-known models of the LCT, which are called mutual inductance model and T model of LCT are shown in Fig. 2. In Fig. 2(a), $U_P$ and $U_S$ are port voltage phasors of primary and secondary coil, respectively. $I_P$ and $I_S$ are corresponding coil current phasors. $L_P$ and $L_S$ are self-inductances of primary and secondary coil, respectively, whereas $M$ is mutual inductance between the two coils. $\omega$ is operation frequency. In Fig. 2(b), $U_P$ and $I_P$ are identical to that
in Fig. 2(a). \( U_{S-P} \) and \( I_{S-P} \) are primary-converted port voltage and coil current of secondary winding, respectively. \( L_{P1} \) and \( L_{S1-P} \) are primary leakage inductance and primary-converted secondary leakage inductance, respectively, whereas \( L_{M-P} \) is primary-converted mutual inductance.

According to basic circuit theory, the consistency of mutual inductance model and T model of LCT can be simply verified. An important note is that the application scenarios of these two models of LCT are different. In mutual inductance model, the inductance is shown in self-inductance format whereas in T model, the inductance is displayed in leakage and mutual inductance format. As a result, mutual inductance model is more suitable for the condition that the self-inductance needs to be compensated and T model is more applicable for the situation that the leakage inductance needs to be compensated. In LC/S compensation topology, to ensure resonant frequency is irrelevant with coupling coefficient, the self-inductance needs to be compensated. As a result, mutual inductance model is more suitable.

III. TOPOLOGY DESIGN

The proposed LC/S compensation topology and corresponding power electronics circuit components are shown in Fig. 3. \( Q_1 \sim Q_4 \) are four power MOSFETs in the primary side. \( D_1 \sim D_4 \) are the secondary-side rectifier diodes. \( L_P \) and \( L_S \) are the self-inductances of the transmitting and receiving coils, respectively. \( L_1 \) and \( C_1 \) are series compensation inductor and parallel compensation capacitor in primary, respectively. \( C_2 \) is secondary series compensation capacitor. \( k \) is the coupling coefficient between the primary and secondary coils. \( u_{AB} \) is the output voltage of the inverter, and \( u_{ab} \) is the input voltage of the rectifier. \( i_p \) and \( i_S \) are coil current of primary and secondary, respectively.

Fig. 3. LC/S compensation topology for WPT.

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A. CCOut Characteristics of LC/S Compensation Topology

Fig. 4 shows an LC resonant tank, where the resonant angular frequency of \( L_{LC} \) and \( C_{LC} \) equals the angular frequency of the sinusoidal voltage source \( \omega_{S-LC} \), i.e.,

\[
\omega^2_{S-LC} = \frac{1}{L_{LC}C_{LC}}. \tag{2}
\]

On the basis of Kirchhoff’s voltage and current law, the current through \( Z_{LC} \) (\( I_{Z-LC} \)) under resonant condition can be yielded

\[
I_{Z-LC} = -jU_{LC} \sqrt{\frac{C_{LC}}{L_{LC}}} = -j \frac{U_{LC}}{\omega_{S-LC}L_{LC}}. \tag{3}
\]

In accordance with formula (3), if \( U_{LC} \) is magnitude-constant, the output current of LC resonant tank \( I_{Z-LC} \) will be magnitude-constant as well, having nothing to do with the load. The LC resonant tank shows a CCOut characteristic.

Fig. 5 shows a CL resonant tank, where the resonant angular frequency of \( C_{CL} \) and \( L_{CL} \) equals the angular frequency of the sinusoidal voltage source \( \omega_{S-CL} \), i.e.,

\[
\omega^2_{S-CL} = \frac{1}{L_{CL}C_{CL}}. \tag{4}
\]
The angular frequency of the resonant tank under resonant condition is represented by a current-controlled voltage source. The output voltage of the resonant tank, which is the first-order harmonic of the square voltage inverted at operation angular frequency, making 
\[ U_{AB} = j\omega L P - C_1'' M P \]

Fig. 6. Analytical circuit of LC/S compensation topology.

Similar to the analysis about LC resonant tank, the output voltage of CL resonant tank under resonant condition \( U_{Z-CL} \) can be obtained
\[ U_{Z-CL} = -jI_{CL} \sqrt{\frac{L_{CL} C_{CL}}{C_{CL}}} = -jI_{CL} \omega S_{-CL} L_{CL} \ldots (5) \]

On the basis of (5), if \( I_{CL} \) is magnitude-constant, the output voltage of CL resonant tank \( U_{Z-CL} \) will be also magnitude-constant, having nothing to do with the load. The CL resonant tank shows a constant-voltage-output (CVO) characteristic.

The CCOut characteristic of LC/S compensation topology is going to be analyzed with the first-order harmonic of the square voltage waveform at switching frequency. The parasitic resistances of all components are neglected for simplicity of the analysis. The accuracy of the approximations will be verified by circuit simulation and experiments in later sections. The analytical circuit of Fig. 3 is shown in Fig. 6.

\( U_{AB} \) is the first-order harmonic of the square voltage inverted from the dc voltage \( U_{dc} \). The angular frequency of \( U_{AB} \) \( \omega_0 \) is referred to as operation angular frequency hereafter. \( C_1' \) and \( C_1'' \) are split by \( C_1 \). The relationship between them can be expressed by the following equation:
\[ C_1' + C_1'' = C_1 \ldots (6) \]

\( C_1' \) resonates with \( L_1 \) at operation angular frequency, making the output current of the first LC resonant tank sinusoidal and amplitude-constant, whose value can be calculated by (3). It is rewritten as
\[ I_{LC} = -jU_{AB} \sqrt{\frac{C_1'}{L_1}} = -j \frac{U_{AB}}{\omega_0 L_1} \ldots (7) \]

\( C_1'' \) resonates with \( L_P \), likewise at operation angular frequency. Since the output current of the LC resonant tank, which is also the input current of the CL resonant tank, is sinusoidal and amplitude-constant, according to CVO characteristic of CL resonant tank, the output voltage of the CL resonant tank is still sinusoidal and amplitude-constant. The output voltage \( j\omega_0 M I_S \), which is represented by a current-controlled voltage source in Fig. 6, can be calculated according to (5) and (7).
\[ -j\omega_0 M I_S = -U_{AB} \frac{C_1'L_P}{L_1 C_1''} = -U_{AB} \frac{L_P}{L_1} \ldots (8) \]

Then, the secondary coil current \( I_S \) can be obtained
\[ I_S = -j \frac{U_{AB}}{\omega_0 M} \frac{C_1'L_P}{L_1 C_1''} = -j \frac{U_{AB} L_P}{\omega_0 M L_1} \ldots (9) \]

According to (1) and (9), the average power consumed by the equivalent resistance can be deduced
\[ P_{RE} = \frac{8}{\pi^2} \frac{U_{AB} - R_{LMS}^2}{\omega_0^2 M^2 L_1} = \frac{8}{\pi^2} \frac{U_{AB} - R_{LMS}^2 L_P^2}{\omega_0^2 M^2 L_1} R_L \ldots (10) \]

where \( U_{AB} - R_{LMS} \) is the RMS value of \( U_{AB} \). In fact, the power described by (10) is consumed by \( R_L \). Then, the direct current through \( R_L \) (I_{RL}) can be deduced
\[ I_{RL} = \frac{2\sqrt{2} U_{AB} - R_{LMS}}{\pi} \frac{C_1'L_P}{\omega_0 M L_1} = \frac{2\sqrt{2} U_{AB} - R_{LMS} L_P}{\omega_0 M L_1} \ldots (11) \]

To achieve ZPA and reduce system VA rating, \( C_2 \) should be chosen deliberately. According to [5], the load impedance of the secondary can be calculated as a lumped impedance \( (Z_S) \), whose value is given by
\[ Z_S = jX + R_E \ldots (12) \]

where \( X = \omega_0 L_S - 1/(\omega_0 C_2) \). The loading effect of the secondary can be calculated as a parallel impedance \( (Z_r) \). This impedance is dependent on the transformer coupling and operation frequency, which is given by
\[ Z_r = \frac{\omega_0^2 M^2}{Z_S} \ldots (13) \]

Substitute (12) into (13), the reflected impedance is
\[ Z_r = \frac{\omega_0^2 M^2}{jX + R_E} \ldots (14) \]

Then, the input impedance \( Z_{in} \) can be yielded
\[ Z_{in} = \left[ (Z_r + j\omega_0 L_P) \right] / \left( \frac{1}{j\omega_0 C_1''} \right) / \left( \frac{1}{j\omega_0 C_1'} + j\omega_0 L_1 \right) \ldots (15) \]

where the operator “/” represents the parallel calculation of impedance. Considering that \( L_1 \) and \( L_P \) resonate with \( C_1' \) and \( C_1'' \), respectively, both at operation angular frequency, \( Z_{in} \) can be simplified as
\[ Z_{in} = \frac{\omega_0^2 L_1}{\omega_0^2 L_P C_1' - j\omega_0 Z_r C_1''} \ldots (16) \]
IjωL will be studied. \( Z \) represents the equivalent impedance of \( R \). Smaller \( Z \) leads to ZVS. It is also found that \( Z \) can be deduced

\[
\alpha = -\left(\arctan\frac{\omega M^2 + \frac{1}{\omega^2} (X L P C_1 + \omega L P R E C_1)}{L P C_1^2 R E}\right).
\]

To achieve ZPA, the following equation should be satisfied:

\[
\omega_0 M^2 C_1 = X L P C_1'.
\]

Then, the capacitance \( C_2 \) that leads to ZPA can be deduced

\[
C_{2-ZPA} = \frac{L P C_1'}{\omega_0^2 (L S L P C_1 - M^2 C_1)}.
\]

In fact, what we really want to achieve is ZVS. Therefore, the input impedance angle \( \alpha \) should be positive. Assuming the desirable input impedance angle is \( \beta \), according to (18), the corresponding \( C_2 \) leads to ZVS with an input impedance of \( \beta \) can be obtained

\[
C_{2-ZVS} = \frac{L P C_1'}{\omega_0^2 L P L S C_1' - \omega_0^2 M^2 C_1 + \omega L P R E C_1' \tan \beta}.
\]

In terms of (21), it is found that the larger the \( \beta \), the smaller the \( C_{2-ZVS} \), and the easier achievement of ZVS. It is also found that \( C_{2-ZVS} \) is related to both \( M \) and \( R E \). Smaller \( M \) and larger \( R E \) result in smaller \( C_{2-ZVS} \). To ensure ZVS can be achieved under all conditions, \( C_{2-ZVS} \) should be chosen when \( M \) are smallest and \( R E \) are largest.

B. Normalized Stress of the Inductor and Capacitor

The normalized stresses of an inductor (\( \sigma_L \)) and a capacitor (\( \sigma_C \)) are respectively defined as follows:

\[
\sigma_L = \frac{S_L}{P_{RL}} = \frac{2I_{L}^2\omega L}{P_{RL}} \quad \text{(22)}
\]

\[
\sigma_C = \frac{S_C}{P_{RL}} = \frac{2U_{C}^2\omega C}{P_{RL}} \quad \text{(23)}
\]

where \( S_L \) and \( S_C \) are apparent powers of the inductor and capacitor. \( P_{RL} \) is the power consumed by the system resistive load \( R_L, I_L \), and \( U_C \) are the current through the inductor and the voltage over the capacitor, both are RMS values. \( \omega \) is system operation angular frequency. \( L \) and \( C \) correspond to the inductance and capacitance.

The stress analytical circuit of LC/S compensation topology is shown in Fig. 8, where \( Z \) represents the load impedance of the secondary, and \( Z \) stands for the reflected impedance of \( Z \). \( Z \) represents the equivalent impedance of \( C_1, L_P \) and \( Z \) is the system input impedance. Then, \( Z \), \( Z \), \( Z \), and \( Z \) can be deduced

\[
\begin{align*}
Z_S &= \frac{\omega^2 M^2}{Z_S} \\
Z_r &= \frac{\omega^2 M^2}{Z_S} \\
Z_{C1} &= \frac{(Z_r + j\omega L_P) / \frac{1}{j\omega C_1}}{Z_S} \\
Z_{in} &= Z_{C1} + j\omega L_1.
\end{align*}
\]

Therefore, the currents through \( L_1, L_P, L_S \) and the voltages over \( C_1, C_2 \) can be yielded

\[
\begin{align*}
I_{L1} &= \frac{U_{AB}}{Z_{in}} \\
I_P &= \frac{U_{C1}}{Z_r + j\omega L_P} \\
I_S &= \frac{j\omega M I_P}{Z_S} \quad \text{(25)} \\
U_{C2} &= \frac{I_S}{j\omega C_2} \\
U_{C1} &= \frac{U_{AB}}{Z_{C1} + j\omega L_1}.
\end{align*}
\]

The power consumed by \( R_E \) equals that consumed by \( R_L \), which means

\[
P_{RL} = P_{RE} = I_S^2 R_E \quad \text{(26)}
\]

where \( I_S \) is the RMS value of \( I_S \). In accordance to previous equations, the normalized stresses of the employed inductors and capacitors can be obtained

\[
\begin{align*}
\delta_{L1} &= \frac{2I_{L1}^2\omega L_1}{I_S^2 R_E} \\
\sigma_{LP} &= \frac{2I_P^2\omega L_P}{I_S^2 R_E} \\
\sigma_{LS} &= \frac{2I_{LS}^2\omega L_S}{I_S^2 R_E} \quad \text{(27)} \\
\sigma_{C1} &= \frac{2U_{C1}^2\omega C_1}{I_S^2 R_E} \\
\sigma_{C2} &= \frac{2U_{C2}^2\omega C_2}{I_S^2 R_E}.
\end{align*}
\]

\( C_1 \) is chosen to make \( C_1' \) and \( C_1'' \) resonate with \( L_1 \) and \( L_P \), respectively, both at operation angular frequency, and \( C_2 \) is chosen in terms of (21) to achieve an input impedance of 25°. The normalized stresses on \( L_1, L_P, L_S, C_1 \) and \( C_2 \) against \( L_P, L_S, L_1, k, \omega \), and \( R_L \) are going to be studied. \( U_{AB} \) is not considered because generally, it is given and amplitude-constant. In the following simulations of this section, \( U_{AB} \) is set to be (100 * 4/\pi) V.
The normalized stresses versus $L_P$ and $L_S$ are studied first of all when $L_1, k, \omega_0,$ and $R_L$ are $310 \mu H$, $0.45$, $(2\pi \times 85 \text{ k}) \text{ rad/s}$, and $60 \Omega$, respectively. The results are shown in Fig. 9. Fig. 9(a) indicates that the normalized stress on $L_1$ increases with $L_P$ but decreases with $L_S$. The normalized stress keeps relatively small as long as a relatively large $L_P$ and a relatively small $L_S$ are not employed concurrently. Fig. 9(b) is similar to Fig. 9(d). In light of these two figures, both $L_P$ and $L_S$ should be relatively large to avoid big stress on $L_P$ or $C_1$. Fig. 9(c) implies a proportional relation between the normalized stress and $L_S$ whereas the stress on $L_S$ keeps invariable when $L_P$ varies. Compared with the stress on $L_1, C_1$ and $L_P$, the stress on $L_S$ is quite small. Fig. 9(e) suggests that the stress on $C_2$ is also quite small. Considering all above aspects, together with power transfer capability, coupling between coils and system cost, as well as VA rating balance between $L_P$ and $L_S$, they are chosen to be around $230 \mu H$.

Under the condition that $L_P$ and $L_S$ both equals $230 \mu H$, the correlations between $L_1, k$, and $\sigma_{L_1}, \sigma_{L_P}, \sigma_{C_1}$, and $\sigma_{C_2}$ are displayed in Fig. 10. $\omega_0$ and $R_L$ are identical to those corresponding to Fig. 9. The normalized stress on $L_S$ is not covered because in terms of Fig. 8 and (27), it keeps constant when $L_1$ and $k$ vary. The stress on $L_S$ can be calculated as $4.832$ based on the given parameters.

The profile of Fig. 10(a) is highly similar to Fig. 10(c), suggesting that the normalized stresses on $L_1$ and $C_1$ decrease with both $L_1$ and $k$. If a relatively small $L_1$ and $k$ are adopted simultaneously, the stress will surge to a very high value, which
should be avoided. On the basis of Fig. 10(b), the stress on $L_P$ is negatively correlated with $k$ and it has no relation to $L_1$. The stress on $C_2$ is negatively associated with $k$ and it is unrelated to $L_1$ as well. However, the decline rate of the stress on $L_P$ against $k$ decreases with $k$, whereas the decline rate of the stress on $C_2$ against $k$ increases with $k$. Moreover, the stress on $C_2$ is very small regardless of the values of $L_1$ and $k$. In light of the discussion above, the bigger the $L_1$, the higher the $k$, the smaller the stress. Nevertheless, in accordance to (11), $L_1$ is mainly determined by the output current of the WPT system. In other words, $L_1$ cannot be altered in terms of the stress on it. To reduce the stress on $L_1$, increasing the coupling coefficient will be a feasible approach as the stress decreases with the coupling coefficient. Therefore, the LCT optimization design will be done in Section IV to maximize the coupling coefficient between two coils.

With the same method, the normalized stress on each element against operation angular frequency $\omega_0$ and resistive load $R_L$ is obtained and shown in Fig. 11. All the calculations are conducted under the condition that $L_P$, $L_S$, $L_1$, and $k$ are equal to 230, 230, 310 $\mu$H, and 0.45, respectively.

According to Fig. 11(a), (b), and (d), a bigger $\omega_0$ will be better because it leads to smaller normalized stresses on $L_1$, $L_P$, and $C_1$. However, in terms of Fig. 11(c) and (e), the conclusion is completely opposite because the normalized stress on $L_S$ and $C_2$ both increases with $\omega_0$. Therefore, a tradeoff must be done to make none of the normalized stresses too big. The operation frequency of our practical WPT system is set to be 85 kHz, the recommended system operation frequency in J2954TM, a standard frequency proposed by the Society of Automotive Engineers for wireless charging. The normalized stresses on $L_1$, $L_P$, and $C_1$ increases with $R_L$, whereas the normalized stresses on $L_S$ and $C_2$ decrease with $R_L$. This is just a conclusion, which cannot be utilized because the resistive load is uncontrolled.

According to the requirements of the project and the discussion before, the parameters of LC/S compensated system are chosen and listed in Table I. With these parameters, the normalized stresses on employed inductors and capacitors can be obtained and given by Table II. The unit of normalized stress is V·A/W.

The optimization effect is limited because system efficiency depends on many factors, such as input voltage, coupling coefficient, dimension of employed Litz wire, output current, rated load, system size limitation, and so on. To obtain highest system efficiency, a comprehensive optimization, taking all efficiency-related factors into consideration, should be done under the premise of diverse limitations. This optimization is quite complicated, and it will be done in the future.

### C. Stress Comparison Between LC/S and Double-Sided LCC Compensated System

To compare the component stress between WPT systems with different compensation topologies, the following two rules must be obeyed. First, the input voltage, operation frequency, resistive load, and output power should be identical. Next, both systems should be optimized.

Fig. 12 is the analytical circuit of double-sided LCC compensation topology. The parameter tuning method can refer to [22]. On the basis of the first rule given before, $U_{AB}$, $\omega_0$, $R_E$, and $I_{RE}$ (the peak value of the current through $R_E$) in Fig. 12 should be 127.32 V, $(2\pi \times 85 \text{ k}) \text{ rad/s}$, 48.63 $\Omega$, and 1.71 A. What follows is the optimization of the double-sided LCC compensation topology.
The optimization of a double-sided LCC compensated system can be divided into two decoupled steps. First, design the LCT for higher coupling coefficient and figure of merit (FOM) under various constraints. Second, design the compensation circuit parameters to make VA rating of $L_P$ and $L_S$ close. The LCT
can be derived by Maxwell simulation, which and Δ compensation system are summarized in Table III. L and C represent the primary compensation inductance and capacitance, respectively.

When ΔL changes by 5%, the current decreases 1.84% and ΔC deviates from its resonant value, the current changes 0.4%. When L1 equals the resonant value, the current becomes 3.28 A, which is 5% larger than the resonant value, the current becomes 3.28 A.

### Table III

<table>
<thead>
<tr>
<th>Parameter</th>
<th>C1 (nF)</th>
<th>C2 (nF)</th>
<th>L1/2 (μH)</th>
<th>L2 (μH)</th>
<th>k</th>
<th>Lf1 (μH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Value</td>
<td>26.0</td>
<td>36.8</td>
<td>97.1</td>
<td>36.1</td>
<td>26.4</td>
<td>60.0</td>
</tr>
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### Table IV

<table>
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<tr>
<th>Component</th>
<th>Lf1</th>
<th>Lp</th>
<th>Ls</th>
<th>L1/2</th>
<th>C1</th>
<th>C2</th>
<th>C1/2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Normalized Stress</td>
<td>2.47</td>
<td>4.45</td>
<td>4.44</td>
<td>2.13</td>
<td>4.14</td>
<td>1.84</td>
<td>2.57</td>
</tr>
</tbody>
</table>

optimization will be interpreted in detail in Section IV while the optimization results are employed here in advance. Based on the optimization results and system dimension limitations, Lp, Ls, and k can be derived by Maxwell simulation, which are 230 μH, 230 μH, and 0.45, respectively.

The following equation should be fulfilled to make the VA rating of Lp identical to that of Ls

\[ L_{f2} = \sqrt{\frac{M^2 R_p E_{Ls}^2 L_s}{\omega_0^2 L_p^2}}. \] (28)

Substitute the specific values of the corresponding variables into (28), Lf2 can be derived as 97.1 μH. In terms of circuit fundamentals, the following equations can be yielded:

\[ L_{f1-o} = \frac{U_{AB-RMS} k \sqrt{L_p L_s}}{\omega_0 L_{f2} I_{RE} \sqrt{1 + (\tan \beta)^2}} \] (29)

\[ L_{f1} = L_{f1-o} + \frac{U_{0} L_{f1-o} - 2 L_{f2}^2}{R_E k^2 L_p L_s} \tan \beta \] (30)

where Lf1-o stands for the primary compensation inductance when the input impedance angle is zero. Lf1 represents the primary compensation inductance when the input impedance angle is 0. For fair comparison and realization of ZVS, the input impedance angle of double-sided LCC compensated system is set to 25°. Therefore, Lf1-o and Lf1 can be calculated as 134.8 and 216.7 μH, respectively. The employed parameters of double-sided LCC compensated system are summarized in Table III. Where U_{AB} is peak value. With the parameters given in Table III, the normalized stresses on employed inductors and capacitors can be obtained, tabulated in Table IV. The unit of normalized stress is voltamperes per Watt.

According to Tables II and IV, the stresses on L1, Lp, and Ls of LC/S are larger than the stresses on Lf1, Lf2, and Ls of double-sided LCC. However, double-sided LCC has another inductor, causing additional power loss. Therefore, in terms of the stresses on employed inductors, LC/S and double-sided LCC have their own advantages. The stress on C1 of LC/S is the biggest one among all component stresses in Tables II and IV, which should be especially paid attention to. Nevertheless, the stresses on C1 and C2 of double-sided LCC are also considerable. On the whole, the total stress on the capacitors of double-sided LCC are larger than that of LC/S.

### Table V

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<th>Parameters</th>
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<td>Lp</td>
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<tr>
<td>Ls</td>
<td>200 μH</td>
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<tr>
<td>Coupling coefficient k</td>
<td>0.4</td>
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<tr>
<td>L1</td>
<td>200 μH</td>
</tr>
<tr>
<td>Operation angular frequency ω₀</td>
<td>(2π × 85) rad/s</td>
</tr>
<tr>
<td>C1</td>
<td>35.059 nF</td>
</tr>
<tr>
<td>C2</td>
<td>25.779 nF</td>
</tr>
<tr>
<td>U_{AB}</td>
<td>127.32 V</td>
</tr>
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</table>

### Table VI

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Lp</td>
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<tr>
<td>Ls</td>
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<td>Coupling coefficient k</td>
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<td>25.779 nF</td>
</tr>
<tr>
<td>U_{AB}</td>
<td>127.32 V</td>
</tr>
</tbody>
</table>

D. System Performance Under Imperfect Resonant Condition

According to the analysis in Section III-A, the system shows an excellent CCOut characteristic under perfect resonant condition. However, in practical system, it is hard to achieve perfect resonant condition due to various reasons. It is necessary to analyze the system performance under imperfect resonant condition.

Generally speaking, imperfect resonant condition mainly results from two aspects: Inaccurate component parameters and variation of coils’ relative position. In accordance with the discussion before, minor changes on L1, C1, k, and C2, named ΔL1, ΔC1, Δk, and ΔC2, respectively, are introduced to analyze system performance under imperfect resonant condition. It is important to note that L1, C1, k, and C2, together with U_{AB}, Lp, Ls, and L_E comprise a perfect resonant system and their values are listed in Table V. For simplicity, we supposed that ΔL1, ΔC1, Δk, and ΔC2 do not appear simultaneously.

Fig. 13 shows the load current profiles versus R_E and ΔL1, ΔC1, Δk, and ΔC2. In Fig. 13(a), when L1 equals the resonant value, the current through R_E is 2.98 A, which does not vary with R_E. When L1 is 5% larger or smaller than the resonant value, the current changes with R_E, both from 2.97 A (20Ω) to 2.86 A (100Ω). In fact, if L1 increases or decreases by a same percentage, the corresponding currents through R_E are conjugate symmetric, and the absolute values are identical. Strictly speaking, if L1 deviates from its resonant value, the CCOut characteristics do not exist anymore. However, if the deviation is within 5%, the variation of current versus R_E is so small that can be neglected. It can still be regarded as a CCOut system. Further, when L1 changes by 5%, the current decreases by 4.0% when R_E is 100Ω and 0.3% when R_E equals 20Ω, displaying quite a good CCOut characteristic against variation of L1.

The profile of load current against R_E and ΔC1 is displayed in Fig. 13(b). When C1 equals the resonant value, the current through R_E (2.98 A) is invariant as R_E changes. When C1 is 5% larger than the resonant value, the current become 3.28 A.
(20 Ω) and 2.78 A (100 Ω). When $C_1$ is 5% smaller than the resonant value, the current falls into 2.69 A (20 Ω) and 2.39 A (100 Ω). No matter $C_1$ is 5% larger or smaller than the resonant value, the current will change with $R_E$. However, compared to the variation of $R_E$, the variation of the current is small enough that can be neglected.

On the basis of Fig. 13(c), when the coupling coefficient increases by 20%, the load current decreases by 16.8%. When the coupling coefficient decreases by 20%, the current increases by 25.2%. It is obvious that the system output current changes with the coupling coefficient. This is a common problem of almost all kinds of compensation topologies. As far as we are concerned, this problem can be solved by closed-loop control (altering the dead time of inverted square-wave voltage) or LCT design. According to Fig. 13(c), it can also be found that the system output current does not change with $R_E$ when the coupling coefficient is fixed to a certain value, no matter it is larger or smaller than the perfect resonant value. Though the system output current changes with the coupling coefficient. Once the coupling coefficient is determined, the system output current keeps constant regardless of the load. This characteristic is of great importance since it decouples the system output current and the load, which will significantly simplify the design of system control circuit.

Fig. 13(d) describes the relationship between the load current, $\Delta C_2$ (represented by percentage of $C_2$) and $R_E$. No matter $C_2$ rises or drops and what the value of $R_E$ is, the current through $R_E$ does not change (the little variation shown in Fig. 13(d) is caused by calculation error). It means that $C_2$ has no impact on system output current.

Precisely speaking, any change of $L_1$ or $C_1$ will impair the CCOut characteristic of LC/S compensation topology. However, even though the change makes up a percentage of 5%, the variation of the current through $R_E$ is still so small, especially compared to the variation of $R_E$, that can be neglected. Therefore, in practical application, the imperfect resonant system can still be considered as a CCOut system, as long as the change is within a certain range. Additionally, the change of $k$ or $C_2$ would not change the CCOut characteristics of the system. It is very appealing since it makes it much easier to design a WPT system with CCOut characteristics.

Fig. 14 shows the input impedance angle profiles versus $R_E$ and $\Delta L_1$, $\Delta C_1$, $\Delta k$, and $\Delta C_2$. When the system is under resonant condition, which means all $L_1$, $C_1$, $k$, and $C_2$ are identical to perfect resonant values, the input impedance angle always equals zero regardless of the value of $R_E$. However, if the system is not under perfect resonant condition, no matter which one of $L_1$, $C_1$, $k$, and $C_2$ deviates from its perfect resonant value, the input impedance angle changes with $R_E$.

In terms of Fig. 14, the input impedance angle shows a positive correlation with $\Delta L_1/L_{1-p}$, $\Delta C_1/C_{1-p}$, and $\Delta k/k_P$, whereas it exhibits a negative relationship with $\Delta C_2/C_{2-p}$. With the decrease of $R_E$, the variation of input impedance angle with $\Delta L_1/L_{1-p}$ and $\Delta C_1/C_{1-p}$ becomes slower, but the variation of input impedance angle with $\Delta k/k_P$ and $\Delta C_2/C_{2-p}$ becomes rapider. By comparing the influence of $\Delta L_1/L_{1-p}$, $\Delta C_1/C_{1-p}$, $\Delta k/k_P$, and $\Delta C_2/C_{2-p}$ on the input impedance angle, it can be concluded that $\Delta C_1/C_{1-p}$ has the largest effect on the input impedance angle while $\Delta C_2/C_{2-p}$

![Fig. 13. Load current profiles versus (a) $\Delta L_1/L_1$ and $R_E$. (b) $\Delta C_1/C_1$ and $R_E$. (c) $\Delta k/k$ and $R_E$. (d) $\Delta C_2/C_2$ and $R_E$.](image-url)
has the smallest impact. As a consequence, the capacitor with higher accuracy should be employed to $C_1$ to get better system performance. On the contrast, the capacitor with lower accuracy can be adopted in $C_2$ to reduce the system cost while keep an acceptable system performance. What needs to be emphasized is that all the conclusions related to Fig. 14 are just applicable to this particular condition. Nevertheless, the analysis methodology proposed here is always applicable, despite of parameter variations.

IV. LCT DESIGN

Budhia et al. [27] presented the design and optimization methodology of a circular magnetic core. Zhang et al. [28] studied the structure and interoperability of DD and unipolar coils, respectively, shown in Fig. 15(a) and (b), for electric vehicle (EV) wireless charging system. Simulation results suggested that DD coil is better than unipolar coil from the perspective of coupling coefficient and system efficiency. Budhia et al. [30] conducted a comparative study on DD and unipolar coils. The simulation results indicated that DD coil is better than unipolar coil in terms of misalignment tolerance, flux path height, and charge zone. Based on the analysis above, DD coil is finally selected. The principles to determine the distance between two D coils in the same side, the length, and width of inner air lump, marked by $D$, $L_1$, and $W_1$ in Fig. 15(a), respectively, are not discussed in [28]. Therefore, it is going to be discussed and aimed at designing desirable LCT more easily. This discussion is conducted with the help of FEA software ANSYS Maxwell while the criterion for convergence is set as 1%.

Roughly speaking, higher LCT coupling coefficient leads to smaller component stress and less reactive power, hence lower power loss and increased system efficiency. Except coupling coefficient, system transfer factor (STF) $Q$, the geometric average of primary and secondary coil power factors, also has great impact on system performance. Generally speaking, higher $Q$ results in less power loss and higher system efficiency [31]. Therefore, a composite variable— FOM, the product of coupling coefficient and STF—is proposed to comprehensively reflect the quality of an LCT [32], [33].

The power factor of a coil ($Q_L$) is defined as follows:

$$Q_L = \frac{\omega L}{\text{ESR}_{\text{c1}} + \text{ESR}_{\text{h1}} + \text{ESR}_{\text{ecl}}} \quad (31)$$

where $\omega$ is system operation frequency while $L$ represents coil inductance. $L$ can be obtained by simulation. ESR$_{\text{c1}}$, ESR$_{\text{h1}}$, and ESR$_{\text{ecl}}$, respectively, stand for the equivalent series resistance caused by copper loss, hysteresis loss, and eddy current loss. ESR$_{\text{ecl}}$ is omitted due to extremely large resistivity of employed ferrite. ESR$_{\text{h1}}$ can also be omitted in this study since the coil current is relatively small and the equivalent cross-sectional area of ferrite is relatively large. This has been verified by
Table VI gives some fixed parameters adopted in the simulation. The variation ranges of three coupling-related parameters \((D, L_I, W_I)\) are listed in Table VII. Only primary parameters are listed in Tables VI and VII since the secondary parameters are totally identical to primary ones. It is best to optimize the LCT with consideration of all five parameters (length of ferrite bar, width of ferrite bar, \(D\), \(L_I\), and \(W_I\)). However, the five-dimensional (5-D) optimization problem will take much time and it is unacceptable for practical application. To obtain time-acceptable and relatively optimal solution, the 5-D optimization problem is divided into three 3-D problems (respectively analyzed in Sections IV-A, B, and C as follows). The optimization results have verified the feasibility of the division.

### Table VI

<table>
<thead>
<tr>
<th>Primary parameters</th>
<th>Values</th>
</tr>
</thead>
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<tr>
<td>Airgap</td>
<td>50 mm</td>
</tr>
<tr>
<td>Litz wire diameter</td>
<td>3 mm</td>
</tr>
<tr>
<td>Turn of a single (D) coil</td>
<td>10</td>
</tr>
<tr>
<td>Space between wires</td>
<td>0</td>
</tr>
<tr>
<td>Thickness of ferrite bars</td>
<td>8 mm</td>
</tr>
</tbody>
</table>

* A single \(D\) coil refers to the part circled by blue dashed line in Fig. 15(a).

### Table VII

<table>
<thead>
<tr>
<th>Symbols</th>
<th>Values</th>
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<tbody>
<tr>
<td>(D)</td>
<td>0.1 mm to 60 mm</td>
</tr>
<tr>
<td>(L_I)</td>
<td>0.1 mm to 60 mm</td>
</tr>
<tr>
<td>(W_I)</td>
<td>0.1 mm to 60 mm</td>
</tr>
</tbody>
</table>

* The minimum distance/length/width is set to be 0.1 mm for the consideration of insulation between two \(D\) coils in the same side.

simulation with ANSYS Maxwell. Detailed verification process is not encompassed in the paper due to space limitation. ESR, approximately equals resistivity of employed Litz wire multiplied by the length.

Fig. 16. Simulation results of ferrite bar size optimization when \(D\) equals 0.1 mm.

#### A. Distance Between Two \(D\) Coils in the Same Side

The influence of \(D\) to coupling coefficient is studied under the condition that both \(L_I\) and \(W_I\) are 30 mm, and \(D\) varies from 0.1 to 60 mm with an interval of 10 mm (the first interval is 9.9 mm). With respect to each scenario, the length and width of the ferrite bar are first optimized while the thickness is constant. The simulation results of ferrite bar size optimization when \(D\) equals 0.1 mm is shown in Fig. 16. FOM increases first and decreases later with either length or width of the ferrite bar. It achieves maximum when the dimension of ferrite bar is \(150 \times 64 \times 8 \text{ mm}^3\).

Through similar optimization approach to each scenario, the profiles of coupling coefficient (CCO), STF, and FOM versus \(D\) are depicted in Fig. 17. CCO increases with \(D\) whereas STF decreases with \(D\), meaning a failure of obtaining largest CCO and STF simultaneously. However, FOM increases with \(D\) all the time, indicating CCO dominates FOM. Compared to the increment of system size, the increment of FOM is not so appealing. Besides, the size constraint of our application is harsh. Therefore, \(D\) is set to be 0 mm.

#### B. Length of Inner Air Lump

Through similar optimization method, the profiles of CCO, STF, and FOM versus \(L_I\) can be derived. They are displayed in
Fig. 17. Profiles of CCO, STF, and FOM versus $D$.

Fig. 18. Profiles of CCO, STF, and FOM versus $L_I$.

Fig. 19. Profiles of CCO, STF, and FOM versus $W_I$.

Fig. 20. Profiles of $i_{MOS}$ and $u_{DS}$ when $R_E$ and $R_L$ are respectively employed.

To get better system performance, $L_I$ should be as large as possible under the premise of satisfying system size constraint. $L_I$ is finally set to be 80 mm.

C. Width of Inner Air Lump

Fig. 19 shows the profiles of CCO, STF, and FOM versus $W_I$. The profiles are obtained under the premise that $D$ and $L_I$ are 0.1 and 80 mm, respectively. Simulation results suggest strong positive correlations between CCO, STF, FOM against $W_I$. $W_I$ is chosen to be 40 mm in practical prototype due to system size constraint.

V. Simulation Verification

All simulations in this section are about electric circuit, having nothing to do with magnetic circuit. The adopted simulation software is Pspice from Cadence Design Systems, Inc. The simulation circuit is based on the circuit shown in Fig. 3, but the dc input voltage source and full-bridge inverter is substituted with a 100-V square wave voltage source. The filter capacitance $C_F$ employed in the simulation is 100 $\mu$F while the rectifier diode is ideal. Unless specially claimed, the other parameters are identical to those given by Table I.

A. Achievement of ZPA

Fig. 20 shows the profiles of the current through the MOSFET ($i_{MOS}$) and the voltage between the drain and source ($u_{DS}$). On the basis of (20), $C_{2-ZPA}$ should be 29.08 nF. It can be found that $i_{MOS}$ is nearly in phase with the first harmonic of $u_{DS}$, demonstrating the validity of (20).

B. ZVS of the MOSFET

Fig. 21 exhibits the profiles of $i_{L1}$ and $u_{AB}$. The MOSFET has realized ZVS, reducing switching losses and system size. The power density and level can also be improved. Compared with Fig. 20, the system input impedance angle can be readily altered by changing secondary series compensation capacitance. It significantly reduces the difficulty of debugging circuit to achieve ZVS and optimum performance.
Fig. 21. Profiles of $i_{L1}$ and $u_{AB}$ when $C_{ZVS}$ are employed.

Fig. 22. Profiles of load current when load suddenly decreases.

### C. CCOut Characteristic of LC/S Compensation Topology

According to the analysis in Section III-A, the system output current does not change, with $R_L$ if other variables keep invariant. Fig. 22 shows the profile of load current when load suddenly decreases. When $R_L$ equals 60 $\Omega$, the current through $R_L$ is 1.083 A. When $R_L$ suddenly decreases to 30 $\Omega$, the current surges to 2.148 A, and then gradually drops to 1.083 A and keeps constant. The transient process continues for 19 ms. The transient time is mostly determined by the steady current through $R_L$, the variation range of the load and the filter capacitance. In terms of (11), the current through $R_L$ is 1.088 A, very close to 1.083 A, verifying the correctness of the formula to calculate load current.

### D. Easily-Changed Output Power

In four conventional compensation topologies, to change the output power, not only the compensation components, but also the LCT need to be replaced. However, in LC/S compensation topology, the output power can be altered by solely changing $L_1$ and $C_1$. Fig. 23 shows the output power profiles with different combinations of $L_1$ and $C_1$ while all other parameters are invariant. The simulation results illustrate the characteristic of easily-changed output power of LC/S compensation topology. Moreover, on the basis of (10), the theoretical output powers are 170.6 W ($L_1 = 200 \mu H, C_1 = 32.8$ nF), 109.2 W ($L_1 = 250 \mu H, C_1 = 29.3$ nF), and 71.0 W ($L_1 = 310 \mu H, C_1 = 26.6$ nF). The calculation results are very close to simulation results, verifying the correctness of previous theoretical analysis. The calculation results are a little larger than the simulation results due to component parameter error and nonideal filter.

### E. Efficiency Comparison Between LC/S and Double-Sided LCC Compensation Topologies

Simulations with LC/S and double-sided LCC compensation topologies are conducted for efficiency comparison between them. The employed parameters are identical to those given by Table I and Table III except input voltage and RFRC. The input voltage in the simulation is a 100-V square wave, whereas RFRC is replaced by a 48.6-$\Omega$ equivalent resistor.

The power loss of an inductor can be divided into two parts, iron loss, and copper loss. Iron loss includes hysteresis loss and eddy current loss. Nevertheless, the resistivity of PC47, employed as the iron core material of the compensation inductor in proposed WPT system, is very large, leading to a very small eddy current loss, which can be omitted. Therefore, iron loss only refers to hysteresis loss here. What is going to be discussed is the method to calculate the equivalent series resistance of hysteresis loss (ESR$_{hl}$) of an inductor.

When the ferrite core of an inductor is unsaturated, the inductance ($L$) is defined as the ratio of flux linkage ($\Psi$) to the current ($i$), described by the following equation:

$$L = \frac{\Psi}{i} = \frac{NBS}{i}$$  \hspace{1cm} (32)

where $N$, $B$, and $S$ are coil turns, flux density and section area of magnetic circuit, respectively. Then, the flux density of an inductor can be obtained through dividing $L$ by $N$. In fact, the obtained flux density is the $B$ field in the air, not the $B$ field in the ferrite. However, the airgap is quite small and ETD cores are employed. The section area of magnetic circuit in the air is approximately equal to that in the ferrite. As a result, the flux density in the ferrite equals that in the air providing the $B$ field uniformly distributes in the ferrite. Based on the profiles of relative core losses versus frequency and flux density given by the manufacturer, the hysteresis loss and original equivalent series resistance of hysteresis loss (ESR$_{hl-o}$) of an inductor can be derived. Nevertheless, the hypothesis that $B$ field uniformly distributes in the ferrite is incorrect. The nonuniform magnetic distribution leads to an obvious increment of hysteresis loss [34]. Therefore, an extra coefficient $k_{NMD}$ is added to roughly reflect the impact, i.e.,

$$ESR_{hl} = k_{NMD}ESR_{hl-o}.$$  \hspace{1cm} (33)

$k_{NMD}$ is obtained in light of practical measurement for each specific $B$ field.
With the parameters listed in Table I, the peak current through $L_1$ (1.23 A) can be obtained. In terms of the data book of EPCOS corporation [35], the calculated hysteresis loss power of $L_1$ is 0.51 W, and corresponding ESR of $L_{11}$ is 0.674 Ω. According to practical measurement and simple calculation, $k_{NM2}$ can be yielded as 1.91. Based on (33), ESR$_{hl}$ can be finally obtained as 1.287 Ω. According to practical measurement and simple calculation, $k_{NM2}$ can be yielded as 1.91. Based on (33), ESR$_{hl}$ can be finally obtained as 1.287 Ω. According to practical measurement and simple calculation, $k_{NM2}$ can be yielded as 1.91. Based on (33), ESR$_{hl}$ can be finally obtained as 1.287 Ω. According to practical measurement and simple calculation, $k_{NM2}$ can be yielded as 1.91. Based on (33), ESR$_{hl}$ can be finally obtained as 1.287 Ω. According to practical measurement and simple calculation, $k_{NM2}$ can be yielded as 1.91. Based on (33), ESR$_{hl}$ can be finally obtained as 1.287 Ω.

The equivalent series resistance of copper loss (ESR$_{cl}$) of an inductor can be approximately replaced by its dc resistance because the operation frequency is only 85 kHz and appropriate Litz wire is used, substantially mitigating skin effect and proximity effect [19], [36], [37]. Roughly speaking, the dc resistance of an inductor ($R_{DC-L}$) is proportional to coil turns ($N_L$), which can be described as follows:

$$R_{DC-L} = k_1 N_L$$  \hspace{1cm} (34)

where $k_1$ is a coefficient. The inductance of an inductor ($L_L$) is proportional to the square of coil turns. It can be expressed by the following equation:

$$L_L = k_2 N_L^2.$$  \hspace{1cm} (35)

On the basis of (34), (35) and the parameters of a practical inductor, $k_1$ and $k_2$ can be obtained, which are $2.22 \times 10^{-3}$ Ω/turn and $1.53 \times 10^{-7}$ H/turn$^2$, respectively. Then, the ESR$_{hl}$ of $L_1$, $L_{11}$, and $L_{12}$ can be derived. They are 0.1, 0.084, and 0.056 Ω, respectively. The total ESR of an inductor can be got by adding the ESR$_{hl}$ and ESR$_{cl}$ together. The ESRs of $L_1$, $L_{11}$, and $L_{12}$ are 1.387, 1.097, and 0.549 Ω, respectively.

The equivalent series resistance of a capacitor (ESR$_c$) is negatively associated with its capacitance. According to the datasheet of the capacitor employed in proposed WPT system, the product of ESR$_c$ and corresponding capacitance is $2 \times 10^{-9}$ Ω·F. Therefore, the ESRs of $C_1$ and $C_2$ of LC/S compensated system are 0.075 and 0.093 Ω, while the ESRs of $C_{1f}$, $C_1$, $C_2$, and $C_{2f}$ of double-sided LCC compensated system are 0.077, 0.054, 0.055, and 0.076 Ω, respectively. The simulation results are shown in Fig. 24. Through simple calculation, the efficiencies of the WPT system with LC/S and double-sided LCC compensation topologies are 92.0% and 93.7%, respectively. Both LC/S and double-sided LCC have achieved high efficiencies though double-sided LCC shows a little higher.

F. Equivalent Resistance of RFRC

The current through secondary coil ($i_S$), as well as system output power ($P_o$), are shown in Fig. 25 when $R_L$ and $R_E$ (the equivalent resistor of RFRC) are respectively employed. It can be found that $i_S$ and the average system output power vary little when RFRC is replaced by its equivalent resistance. Therefore, it is valid to replace RFRC with its equivalent resistance when analyzing the circuit.

VI. EXPERIMENT VERIFICATION

A WPT prototype with LC/S compensation network, which is shown in Fig. 26, is built to verify the analysis above. It is made up of six parts: DC voltage source, primary PCB board,
TABLE VIII
SOME CRITICAL PARAMETERS OF THE COMPONENTS EMPLOYED IN THE PROTOTYPE

<table>
<thead>
<tr>
<th>Components</th>
<th>Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>MOSFET</td>
<td>FQU10N20CTU</td>
</tr>
<tr>
<td>$L_1$</td>
<td>308.9 μH</td>
</tr>
<tr>
<td>$C_1$</td>
<td>26.7 nF</td>
</tr>
<tr>
<td>$C_2$</td>
<td>19.4 nF</td>
</tr>
<tr>
<td>$L_F$</td>
<td>225.11 μH</td>
</tr>
<tr>
<td>$L_S$</td>
<td>230.8 μH</td>
</tr>
<tr>
<td>$k$</td>
<td>0.4503</td>
</tr>
<tr>
<td>$C_F$</td>
<td>1000 μF</td>
</tr>
<tr>
<td>Rectifier diode</td>
<td>Sirectifier MBR30200PT</td>
</tr>
<tr>
<td>$R_L$</td>
<td>60.3 Ω</td>
</tr>
</tbody>
</table>

TABLE IX
DETAILED PARAMETERS OF THE LCT

<table>
<thead>
<tr>
<th>Components</th>
<th>Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Litz wire</td>
<td>$Φ0.1$ mm² × 500</td>
</tr>
<tr>
<td>Ferrite strip</td>
<td>Manufacturer: EPCOS(TDK)</td>
</tr>
<tr>
<td></td>
<td>Material: PC47</td>
</tr>
<tr>
<td></td>
<td>Size: 128 × 25 × 17.5 mm³</td>
</tr>
<tr>
<td></td>
<td>Number: 2 × 4</td>
</tr>
<tr>
<td>The distance between two D coils</td>
<td>0 mm</td>
</tr>
<tr>
<td>Single D coil of primary</td>
<td>Inner air lump size: 40° × 80 mm²</td>
</tr>
<tr>
<td></td>
<td>Turns: 21</td>
</tr>
<tr>
<td>Single D coil of secondary</td>
<td>Inner air lump size: 40° × 80 mm²</td>
</tr>
<tr>
<td></td>
<td>Turns: 21</td>
</tr>
</tbody>
</table>

Fig. 27. Profiles of output current and voltage of the inverter.

LCT, secondary PCB board, resistive load, and oscilloscope, which are numbered from 1 to 6 in order. Some critical parameters of the components employed in the prototype are listed in Table VIII while the detailed parameters of the LCT are given by Table IX.

A. Achievement of ZPA

Fig. 27 shows the profiles of output current and voltage of the inverter. The experiment was carried out with the parameters listed in Table VIII except $C_2$. According to formula (20), to realize ZPA, the secondary series compensation capacitance should be 29.26 nF. However, in practical experiment, $C_2$ is chose to be 29.09 nF due to availability. In terms of Fig. 27, it can be found that the fundamental harmonic of the voltage and that of the current are almost in phase, validating the correctness of the analysis on ZPA of LC/S compensation network, which is elaborated in Section III-A in detail.

B. ZVS of mosfets

With the parameters given in Table VIII, the ZVS of MOSFETs can be realized, as shown in Fig. 28. Compared to the parameters employed in last experiment, the only difference is $C_2$, which is changed from 29.09 to 19.4 nF. It is very convenient and easy to change the input impedance angle of the WPT system with LC/S compensation topology. This merit can extremely reduce the difficulty of debugging the circuit to achieve ZVS. Additionally, the magnitude of the current in Fig. 28 is larger than that in Fig. 27 since more reactive power is involved. A larger current indicates more power loss in primary compensation components, but a larger current is resulted from a greater leading angle between the voltage and current, which will be more beneficial to realize ZVS. Hence, the minimum angle that make all four MOSFETs realize ZVS will be optimum. In our system, the leading angle is set to be around 30°.

C. CCOut Characteristic of LC/S Compensation Topology

To verify the CCOut characteristic of LC/S compensation topology, which is analyzed theoretically in Section III-A, an experiment with a transient load step is conducted. The experiment results are shown in Fig. 29. First, the resistive load is 40.8 Ω and then changed to 24.36 Ω. On the basis of Fig. 29, the
steady current through 40.8-Ω load is about 1.04 A, indicating a 42.4-V voltage over the filter capacitor $C_F$. When the load is changed to 24.36 Ω, the current surges to 1.7 A, and the voltage over $C_F$ is 41.4 V. The minor difference between these two voltages is caused by reading error. The load current reverted to 1.04 A after 100 ms, demonstrating the CCOut characteristic of LC/S compensation topology. In fact, the steady current through 24.36 Ω load is little more than 1.04 A since the parameters in practical experiments is not precise.

### D. Easily-Changed Output Power

The output power files against input dc voltage are exhibited in Fig. 30. The dashed red profile corresponds to the condition that $L_1$ and $C_1$ equal 308.9 μH and 26.7 nF, respectively, which is referred to as case 1. The solid green profile is gained when $L_1$ and $C_1$ are 201.33 μH and 33.35 nF, respectively, and it is referred to as case 2. The output power is readily altered only by changing the values of primary compensation inductor and capacitor. The specification of the LCT does not need to be changed, making LC/S compensation topology free from the constraints imposed by the LCT parameters. Moreover, according to (11), the output power of case 2 should be 164.8 W. The practical output power is 130.9 W. The difference between experimental and theoretical results is mainly caused by the effect of component power loss, which is not taken into consideration in theoretical analysis, and imprecise parameters. If the two aspects were considered in theory, the experimental results should be highly consistent with the theoretical values.

### E. Efficiency Comparison Between LC/S and Double-Sided LC Compensation Topologies

The loss distribution of LC/S compensated system is shown in Fig. 31 when the parameters listed in Table VIII are employed. The input, output and loss powers are 69.4, 63.8, and 5.6 W, indicating a system efficiency of 91.9%.

The loss distribution of double-sided LC compensated system is shown in Fig. 32. The system operation frequency and compensation parameters of double-sided LC are shown in Table X. The other parameters are identical to those of LC/S compensated system. The input, output, and loss powers are 79.6, 73.7, and 5.9 W, indicating a system efficiency of 92.6%. The system efficiency of double-sided LC is a little higher than that of LC/S, which is in coincidence with simulation results.

With respect to LC/S compensated system, most power is dissipated in control circuit (33%), rectifier diodes (26%), and compensation inductor (18%), as well as inverter MOSFETs (9%). The control circuit loss accounts for the largest proportion since the system power is low. The ratio of control circuit loss is going to drastically decrease when system power increases. The rectifier diode loss also takes up a remarkable portion because the output voltage is relatively low, whereas the conduction voltage drop of employed diode is relatively high. The compensation inductor loss ranks three out of seven loss sources as the ESR of compensation inductor is large. The loss distribution of double-sided LCC compensated system is similar to that of LC/S. The major difference is that secondary compensation inductor loss takes a quite big proportion among the total loss for LC/S.

### F. Equivalent Resistance of RFRC

The input voltage and current of rectifier ($u_{ab}$ and $i_x$) is shown in Fig. 33(a). When RFRC was replaced by its equivalent resistance, the waveforms of $u_{ab}$ and $i_x$ are exhibited in Fig. 33(b). The current waveform is significantly similar to that in Fig. 33(a), including the magnitude. Moreover, the system input powers corresponding to these two scenarios are 71.7 and 72.6 W, respectively, which are very close. Similar current waveforms and close system input powers illustrate the validity of replacing RFRC with its equivalent resistance. This replacement
will significantly reduce the difficulty of circuit analysis while keeping the analysis accuracy.

VII. CONCLUSION

A new compensation topology for WPT system, which is named as LCIS, is proposed in this paper to provide excellent CCOut characteristic. The LCIS compensation topology is free from the constraints imposed by the LCT parameters, which means the system output power can be easily changed without replacing the LCT. This merit makes LCIS compensation topology much better than four conventional compensation topologies because it is time and cost consuming to manufacture a new LCT. Additionally, compared to double-sided LCC compensation topology, which consists of two inductors and four capacitors, the system efficiency of LCIS compensation topology is about 2.5% higher. The increase of system efficiency is mainly resulted from the reduction of compensation inductor, on which the power loss is mostly consumed. Further, the input impedance angle can be readily changed only by altering the value of secondary series compensation capacitor. This will obviously reduce the difficulty of debugging the circuit to obtain ZVS and optimum performance. The method to design the LCT is also improved on the basis of reference [28]. This will highly reduce the difficulty of designing a high-performance LCT.

REFERENCE


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