A New Integration Method for an Electric Vehicle Wireless Charging System Using LCC Compensation Topology: Analysis and Design

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Abstract—There is a need for charging electric vehicles (EVs) wirelessly since it provides a more convenient, reliable, and safer charging option for EV customers. A wireless charging system using a double-sided LCC compensation topology is proven to be highly efficient; however, the large volume induced by the compensation coils is a drawback. In order to make the system more compact, this paper proposes a new method to integrate the compensated coil into the main coil structure. With the proposed method, not only is the system more compact, but also the extra coupling effects resulting from the integration are either eliminated or minimized to a negligible level. Three-dimensional finite element analysis (FEA) tool ANSYS MAXWELL is employed to optimize the integrated coils; and detailed design procedures on improving system efficiency are also given in this paper. The wireless charging system with the proposed integration method is able to transfer 3.0 kW with 95.5% efficiency (overall DC to DC) at an air gap of 150 mm.

Index Terms—Wireless power transfer, coil design, electric vehicle charging.

I. INTRODUCTION

For centuries, engineers are seeking to transfer power wirelessly [1-5]. Nowadays wireless power transfer (WPT) technology is widely used in powering biomedical implants, tooth brush, cell phones, laptops, and even plug-in hybrid electric vehicles (PHEVs) and electric vehicles (EVs). The power level ranges from several milliwatts to tens of kilowatts [6-16, 18-21]. The inductive based wireless charging for PHEVs and EVs is highly acknowledged since it provides a more convenient, reliable, and safer charging option compared to conducting charging. The research in this field is of a great variety, such as coil design, power electronics converters and control method, compensation topologies, foreign object detection, and safety issues. Among them, coil design and compensation topologies are two main research areas.

Coils are the essence of an inductive based wireless charging system. They determine the power transfer capability and transfer efficiency. One important property of the coils is the geometry as it closely relates to the coupling coefficient of the coil structure and the quality factor of each coil. Reference [8] gives the design and optimization procedures for circular coils and demonstrates a 2 kW 700-mm-diameter pad. Reference [9] presents a flux-pipe coil structure and claims that a charging system can transfer 3.0 kW power with 90% efficiency at an air gap of 200mm. However, the flux-pipe coil conducts double-sided flux paths and one of the flux paths is wasted. In order to solve this problem, a bipolar coil structure is developed in [10]. The bipolar coil structure offers high efficiency and good misalignment tolerance. An 8 kW wireless charging system is built and tested in [11]. With the optimized bipolar coil structure, the charger can transfer power with 95.66% efficiency when fully aligned and 95.39% efficiency with a 300 mm horizontal misalignment. More recent research on coil design can be found in [22]. The researchers embed a coplanar coil into the primary coil system to improve the overall coupling of the system, which increases the efficiency with the SS compensation topology.

Compensation topology is another important aspect in a wireless charging system as it increases the power transfer ability, minimizes the VA rating of the power source, and helps achieve soft switching of the power electronics devices [23]. There are four basic compensation topologies: SS, SP, PS, and PP where the letters “P” and “S” stand for the way how the resonant capacitor is connected to the coil, i.e., “P” represents parallel connection and “S” represents series connection. More advantageous compensation topologies are put forward in [12-13]. The double-sided LCC compensation topology is outstanding since not only is its resonant frequency independent of coupling coefficient and load condition, but also it is highly efficient [13]. However, one drawback of the double-sided LCC compensation topology is its large volume due to the compensated coils.

To overcome the volume limitation, reference [14] first puts forward the idea of integrating the compensated coil into the main coil system. The authors integrate a bipolar compensated coil into a bipolar main coil system. As it is shown in Fig. 1, five extra coupling effects appear after the integration and the
coupling effect of the two coils at the same side of the wireless charging system are studied. A 6 kW wireless charging system with 95.3 % efficiency was designed and tested in [14]. Further detailed analysis on both the coupling effect of the same-side coils and the coupling effect of the cross-side coils can be found in [15-16]. References [14-16] successfully make the system more compact and highly efficient; however, the method of integration complicates the design of a wireless charging system using double-sided LCC compensation topologies.

In order to simplify the design and analysis while keep the advantages of compactness and high efficiency, this paper proposes a new integration method for a wireless charging system using LCC compensation topology. In this method, the five extra coupling effects are either eliminated or minimized to a neglected level, which greatly simplifies the design and analysis. Additionally, a new method of improving system efficiency is given. Consequently, the system keeps outstanding performance and is able to deliver 3.0 kW power with 95.5% DC-DC efficiency at an air gap of 150 mm.

II. ANALYSIS

The double-sided LCC compensation topology is proven to be highly efficient for WPT applications in electric vehicles and plug-in electric vehicles [13-16]. Fig. 1 shows a wireless charging system using the double-sided LCC compensation topology. The DC input is transformed to a high frequency AC charging system using the double-sided LCC compensation topology involves four magnetic components: CL-filter, the power is qualified to charge the battery packs. Afterwards, it is converted back to DC by the rectifier consist of four diodes. After being further filtered by the CL-filter, the power is qualified to charge the battery packs.

As can be seen from Fig. 1, a double-sided LCC compensation topology involves four magnetic components: $L_1$, $L_2$, $L_3$, and $L_4$ where $L_1$ and $L_2$ are the main coils while $L_3$ and $L_4$ are the compensated coils. $L_1$ and $L_2$ are air-core inductors; however, the large dimension of air-core inductors is not practical in reality. In order to make the system more compact, we propose to integrate the compensated coils into the main coil system. Once they are integrated, extra couplings appear. As shown in Fig. 1, six couplings exist and the coupling coefficients are chosen to represent their coupling effects, where “$k$” stands for the coupling coefficient between the main coils $L_1$ and $L_2$ while “$k_{102}$” represents the coupling coefficient between compensated coils $L_1$ and $L_2$: “$k_{f1}$” is the coupling coefficient between $L_1$ and $L_3$ and similarly, “$k_{23}$” is the coupling coefficient between $L_2$ and $L_4$.

The same-side coupling coefficients; “$k_{102}$” and “$k_{23}$” are the coupling coefficient between $L_1$ and $L_2$, the coupling coefficient between $L_2$ and $L_3$, respectively, “$k_{102}$” and “$k_{23}$” are also known as cross-side coupling coefficients. Moreover, in Fig. 1 the coupling coefficient $k$ is represented by a solid line while the other five is in dashed or dash-dot lines. WPT relies on the main coupling to transmit power wirelessly between the transmitting side and the receiving side; therefore, the main coupling between $L_1$ and $L_2$ is of great significance and the other five couplings are redundant. The goal is to maximize the main coupling and eliminate the five extra couplings or minimize their coupling effects to a negligible level.

Fig. 2 presents the proposed integrated coil structure. Bipolar coils are selected as the main coils $L_1$ and $L_2$ because of the outstanding performance and good misalignment tolerance [11]. Due to the onboard space limitation, the size of the secondary side is smaller than that of the primary side. Unipolar coils are chosen for compensated coils $L_3$ and $L_4$. Since both the primary side and the secondary side have similar coil structures, analysis will be first focused on one side. On the primary side, the bipolar coil $L_1$ performs as a magnetic dipole and the magnetic flux excited by $L_1$ passes through the compensated coil $L_3$. The amount of the magnetic flux can be expressed as $\psi_{11}$,

$$\psi_{11} = \int \vec{B}_1 \cdot d\vec{S}_1$$

where “$\vec{B}_1$” stands for the magnetic flux density and “$\vec{S}_1$” represents the area of the primary compensated coil $L_3$. The compensated coil is placed in the center of the coil structure in a different layer. As a main characteristic of a magnetic dipole, the magnetic flux it excites is symmetrical in space; therefore, the amount of the magnetic flux flows into the compensated coil equals to that flows out of it. As a result, the net magnetic flux passing through $L_3$ is zero. Since $\psi_{11}$ is zero, the coupling coefficient is zero so the coupling effect is eliminated,

$$k_{f1} = \frac{\psi_{f1} \psi_{11}}{\psi_{f1} \psi_{11}} = 0$$

where $\psi_{f1}$ is the amount of magnetic flux excited by $L_1$ passes through $L_3$, $\psi_{11}$ and $\psi_{f1}$ represent the amounts of their self-magnetic flux. Similarly, the net amount of magnetic flux which is excited by $L_2$ and passes through $L_4$ is zero so the coupling coefficient $k_{f2}$ is zero. Moreover, the net amount of magnetic flux which is excited by $L_1$ and passes through $L_2$ is zero and the net amount of magnetic flux which is excited by $L_2$ and passes through $L_1$ is also zero. The coupling coefficients of
The proposed coil structures: (a) Overview, (b) Side view, and (c) Primary coils.

extra-side coils \( k_{1/2} \) and \( k_{2/2} \) are both zero. When fully aligned, the four coupling effects are eliminated. Furthermore, when misalignment between the primary side coil structure and the secondary side coil structure occurs, the four inductance values almost do not change. Since the relative positions of the same-side coils are fixed, the same-side coupling coefficients \( k_{1/1} \) and \( k_{2/2} \) are still zero. Though the relative positions of the extra-side coils change, the extra-side coupling effects can be neglected. It is because the compensated coil inductance values are multiple times smaller than the main coil inductance values and the air gap between the main coil and compensated coil is considered large when compared with coil dimensions. Additionally, the coupling coefficient \( k_{1/2} \) can be minimized to a negligible level by three-dimensional FEA tool when fully aligned, which will be discussed in next section. When misalignment occurs, \( k_{1/2} \) will be further decreased because the relative positions of the compensated coils change.

The five extra coupling coefficients are either eliminated or minimized to a negligible level, the following analysis will be only focused on the main coupling coefficient \( k \) between \( L_1 \) and \( L_2 \). The input DC voltage is converted to a square wave voltage of a designed frequency and enters into the primary resonant tank, where the voltage and its resulting current resonate at the resonant frequency. The resonant tank also serves as a filter and almost all high-frequency components are filtered out. Therefore, first harmonic analysis (FHA) is applied in analyzing the double-sided LCC compensation topology. Here “\( V_{ac} \)” in (3) refers to the fundamental component of the square wave voltage. Furthermore, the battery pack is considered to be a constant voltage source when designing the circuit. Since the output power is constant, the battery pack is modeled as a load resistor \( R_e \) in (4). The root mean square (rms) output current of the double-sided LCC compensation topology is constant once the input voltage is fixed. Therefore, the load resistor can be converted equivalently to \( R_{eq} \) at the input side of the rectifier [17].

\[
v_{ac} = \frac{2\sqrt{2}}{\pi} V_{ac} \angle 0^\circ \tag{3}
\]

\[
R_{eq} = \frac{V_{ac}^2}{P_{out}} \tag{4}
\]

\[
R_e = \frac{8}{\pi^2} R_{eq} = \frac{8}{\pi^2} \frac{V_{ac}^2}{P_{out}} \tag{5}
\]

In Fig. 3, \( L_1 \) resonates with \( C_1 \) and as a result, \( i_1 \) is constant once \( V_{ac} \) is fixed. “\( j\omega M_i \)” can be treated as a constant voltage source for the secondary resonant tank. \( C_2 \) is chosen to compensate \( L_2 \) such that the series impedance of \( L_2 \) and \( C_2 \) is equal to that of \( L_2^* \). \( L_2^* \) and \( C_2 \) resonate with \( C_2^* \). Therefore, the output current \( i_2 \) is also constant, which is desired for battery charging.

\[
\omega L_{i_1} - \frac{1}{\omega C_{i_1}} = 0 \tag{6}
\]

\[
\omega L_2 - \frac{1}{\omega C_2} - \frac{1}{\omega C_{i_2}} = 0 \tag{7}
\]

\[
i_1 = \frac{V_{ac}}{j\omega L_{i_1}} = \frac{2\sqrt{2}}{\pi} V_{ac} \angle -90^\circ \tag{8}
\]

\[
i_2 = \frac{k\sqrt{L_1 L_{i_1} V_{ac}}}{j\omega L_2} = \frac{2\sqrt{2}}{\pi} \frac{k\sqrt{L_1 L_{i_1} V_{ac}}}{\omega L_{i_2}} \angle -90^\circ \tag{9}
\]

As can be seen from equation (9), the output current \( i_2 \) lags the input voltage \( V_{ac} \) by 90°. Therefore, the voltage \( V_{i_2} \) in Fig. 3, which is defined as the voltage at the input side of the rectifier also lags \( V_{ac} \) by 90°. Due to the symmetry of the double-sided LCC compensation network, \( L_2^* \) resonates with \( C_2^* \) while \( L_1 \) and \( C_1 \) are combined to resonate with \( C_1^* \).

\[
\omega L_{i_2} - \frac{1}{\omega C_{i_2}} = 0 \tag{10}
\]

\[
\omega L_1 - \frac{1}{\omega C_1} - \frac{1}{\omega C_{i_1}} = 0 \tag{11}
\]

By substituting equations (6), (7), (10), and (11) into the impedance equations, the secondary side impedance \( Z_{sec} \), the reflected impedance \( Z_{refl} \), and the input impedance \( Z_{in} \) can be simplified as:

\[
Z_{ac} = j\omega L_{sec} + \frac{1}{j\omega C_{sec}} + \frac{1}{j\omega C_{refl}} \left( j\omega L_{i_2} + R_e \right) = \frac{\omega^2 L_{i_2}^2}{R_{eq}} \tag{12}
\]
Fig. 3. Resonant circuits with double-sided LCC compensation topology

\[
Z_{\text{refl}} = -\frac{j\omega M_i}{i_1} = \frac{\alpha^2 M^2}{Z_{\text{sec}}}, \quad M = k\sqrt{L_1L_2} 
\]

(13)

where \( M \) in equation (13) represents the mutual inductance between \( L_1 \) and \( L_2 \). In this lossless analysis, the input power equals to the output power. Therefore, the expression of the system power is given by

\[
P_n = P_{\text{out}} = \frac{V_{\text{AC,in}}}{Z_n} = \frac{8k\sqrt{L_1L_2}V_{\text{AC,in}}}{\pi \omega L_1L_2} 
\]

(15)

The voltage at the input side of the rectifier \( v_i \), the input current \( i_1 \), and the current in the secondary main coil \( i_2 \) can be written as

\[
v_i = i_1 \cdot R_n = \frac{2\pi}{\omega} V_{\text{AC,in}} \angle -90^\circ 
\]

(16)

\[
i_1 = \frac{k\sqrt{L_1L_2}V_i}{j\omega L_1L_2} = \frac{2\pi}{\omega} \left( \frac{k\sqrt{L_1L_2}V_{\text{AC,in}}}{\omega L_1L_2} \right) \angle 0^\circ 
\]

(17)

\[
i_2 = \frac{V_{\text{AC,in}}}{j\omega L_2} = \frac{2\pi}{\omega} \frac{V_{\text{AC,in}}}{\omega L_2} \angle 0^\circ 
\]

(18)

From equations (3), (9), (16), and (17), it is obvious that \( v_i \) is in phase with \( i_1 \) and \( v_i \) is in phase with \( i_2 \). Zero phase angle (ZPA) is achieved and no reactive power is needed from the power source. Moreover, \( v_i \) leads \( v_o \) by 90\(^\circ\) and the energy resonates between the primary side and secondary side back and forth.

III. DESIGN PROCEDURES

1. Main Coil Simulation

Based on equation (15), the coupling coefficient \( k \) between the main coils \( L_1 \) and \( L_2 \) is closely related to the power transfer ability of a wireless power transfer system with the double-sided LCC compensation topology. The value of \( k \) depends on the coil geometry. Therefore, once the dimensions of the coils are fixed, it is preferred to maximize \( k \) by changing the coil widths in both the primary and the secondary sides. Since the application is for passenger electric vehicles, the space is quite limited on the vehicle. So, the receiving coil installed on the car is better to be smaller than the transmitting coil placed on the ground. Fig. 4(a) shows the overview of the proposed main coil structure. We decide to choose the dimension of the transmitting coil at the primary side as “600 mm\(^2\)*450 mm\(^2\)*4 mm” and that of the receiving coil at the secondary side as “400 mm\(^2\)*300 mm\(^2\)*4 mm”. Ferrite plates and aluminum shields are used for magnetic and electric shielding. They are also made successively larger to provide the needed shielding effects. The primary side ferrite plate has a dimension of “640mm\(^2\)*496mm\(^2\)*8mm” while the secondary is “480 mm\(^2\)*352 mm\(^2\)*8 mm”. Aluminum shields are modelled with a primary dimension of “711.2 mm\(^2\)*558.8 mm\(^2\)*2 mm” and a secondary dimension of “508 mm\(^2\)*406 mm\(^2\)*2 mm”. As shown in Fig. 4(a), there are two design variables: the primary coil width and secondary coil width. By varying the two variables, the coupling coefficient \( k \) can be maximized. In our case, \( k_{\text{max}} \) is 0.1764 with the primary coil width of 112 mm and the secondary coil width of 48 mm. Fig. 4(b) shows the simulation results.

2. Main Coil Winding and Measurement

Two design requirements must be met: (1) the designed resonant frequency is 85 kHz and (2) the maximum efficiency is achieved at the desired output power. Additionally, there are eight components in the double-sided compensation topology. If each component has a little variance from the designed value, it will be hard to meet the two requirements at the same time. Therefore, it is highly recommended to build and measure each component right after simulation.

3. Optimization of \( L_{12} \) and \( L_{22} \) at maximum efficiency

According to equation (15), once the resonant frequency, the system power, the main coupling coefficient, and the inductance values of the main coils are determined, the product of the compensated coil inductance values is fixed. A number of combinations could contribute to the same product and it is desired to optimize the values based on the efficiency curve of the system. The system is required to achieve the highest efficiency at the desired output power and the compensated

Fig. 4. (a) Overview of the main coil structure and (b) Coupling coefficient \( k \) with different coil widths.
inductance values can be optimized at that point. Once the output power is fixed, the system efficiency is determined by the power loss, which is closely related to the equivalent series resistance (ESR) of each component. The ESRs of the inductors can be calculated by their quality factors and inductance values and the ESRs of the capacitors rest with their dissipation factors. We can write expressions for the ESRs as

\begin{align}
R_{c1} &= DF \cdot \frac{1}{j\omega C_1}, \quad R_{c2} = DF \cdot \frac{1}{j\omega C_2}, \quad R_{s} = DF \cdot \frac{1}{j\omega C_3}, \quad R_{s2} = DF \cdot \frac{1}{j\omega C_4}, \\
R_{c12} &= DF \cdot \frac{1}{j\omega C_5}
\end{align}

(19)

where \(Q_1, Q_2, Q_3, \) and \(Q_4\) stand for the quality factors of \(L_1, L_2, L_3,\) and \(L_4\) respectively. \(Q_1\) and \(Q_2\) can be measured by LCR meters since they have already been built in the last step. \(Q_3\) and \(Q_4\) can be approximated at 200 based on their inductance values, which are multiple times smaller than the main coil inductance values. \(DF\) is the dissipation factor of the capacitors. Once the compensated inductance values are given, the ESRs of the capacitors rest with their dissipation factors.

Fig. 5(a) and (b) gives the circuit diagram with ESRs added into the system. The system is divided into primary side circuit and the secondary side circuit. Therefore, the efficiency analysis can be done separately. The overall efficiency is the product of primary side circuit efficiency \(\eta_{pri}\) and secondary side efficiency \(\eta_{sec}\). On the secondary side, \(\eta_{sec}\) is the product of \(\eta_{s1}, \eta_{s2},\) and \(\eta_{s3}\). As shown in Fig. 5(a), \(\eta_{s1}\) refers to efficiency of block 1, which is the ratio of the real power obtained by the load over the real power entering into block 1; \(\eta_{s2}\) is defined as the efficiency of block 2, which is the ratio of the real power entering into block 1 over the real power absorbed by block 2; \(\eta_{s3}\) is efficiency of block 3, which is the ratio of the real power absorbed by block 2 over the real power transferred from the primary side of the circuit. The expressions of \(\eta_{s1}, \eta_{s2},\) and \(\eta_{s3}\) are given by

\begin{align}
\eta_{s1} &= \frac{R_{c1}}{R_{c1} + R_{s}} \\
\eta_{s2} &= \frac{R_{c12} + \frac{1}{j\omega C_5}}{(R_{c1} + R_{s}) + \frac{1}{j\omega C_2}} = \frac{R_{c12} + \frac{1}{j\omega C_5}}{(R_{c1} + R_{s}) + \frac{1}{j\omega C_2}} \\
\eta_{s3} &= \frac{R_{c12} + \frac{1}{j\omega C_5}}{Re(Z_{filt}) + R_{c12}}
\end{align}

(20)

where \(Z_{filt}\) is the reflected impedance from the secondary side circuit and \(Z_{block5}\) is the impedance of block 5. We can plot the efficiency over equivalent resistance curve by applying equations (21) and (22) in Mathcad. By adjusting the inductance values of \(L_4\) and \(L_5\), the efficiency can be maximized at the desired equivalent resistance. For our case, the desired output power is 3.0 kW and the battery pack is 300 volts, therefore, the efficiency is maximized at 24.32 ohms. Fig. 6 gives the efficiency curve obtained from Mathcad and the desired inductance values are determined as \(L_4 = 46\mu\text{H}\) and \(L_5 = 38\mu\text{H}\).
4. Maxwell Simulation for $L_{f1}$ and $L_{f2}$

As analyzed in the previous section, a unipolar compensated coil is preferred in a coil structure with bipolar main coils. In order to further decrease the coupling effects between the main coil and the compensated coil, it is more desired to place the compensated coil in the region with weaker magnetic field densities. By substituting $L_{f1}$ and $L_{f2}$ into (8) and (18), the currents flowing through $L_1$ and $L_2$ can be determined and set as the excitation currents in the Maxwell model. Vacuum boxes are built up at the expected positions of the compensated coils, which are shown in Fig. 7(a). Fig. 7(b) gives the simulation results of magnetic flux densities in the vacuum boxes.

Clearly, the magnetic field is weaker in the central areas of the two main coils, which are circled by the dashed lines in Fig. 7(b). The compensated coils should be placed in those areas. With the desired inductance values, the compensated coils are built and simulated in ANSYS MAXWELL.

The compensated coils with the desired values are placed in the system as shown in Fig. 8(a) and the coupling coefficient $k_{f1f2}$ is 0.05, which is good enough for the experiment. However, it can be further optimized. Since Fig. 7(b) indicates a larger area with weaker magnetic flux density on the primary side, so the primary compensated coil has enough area to be placed. By rotating the primary compensated coil from 0° to 180° as shown in Fig. 8(b), the lowest coupling case is achieved. Fig. 9(a) shows the coupling coefficient $k_{f1f2}$ of the compensated coils and the same-side coupling coefficients $k_{f1}$ and $k_{f2}$. It is obvious that $k_{f1f2}$ is minimized as 0.04 at 90° rotating angle while the same-side coupling coefficients are almost equal to zero with the angle rotating. Fig. 9(b) demonstrates that the self-inductance values of the two compensated coils keep almost constant with the angle rotating.

5. Compensated Coil Winding and Measurement

With the parameters provided by MAXWELL simulation, compensated coils are wound and measured. Again, in order to make the resonate frequency at 85 kHz and the efficiency maximized at the desired output power, capacitance values need to be recalculated with the measured results of $L_1$, $L_2$, $L_{f1}$, and $L_{f2}$ through equations (6), (7), (10), and (11). The capacitance values calculated in this step only slightly vary from those calculated in step 3. Therefore, the efficiency will not be affected.
6. Capacitor Tuning for Zero Voltage Switching

Zero voltage switching (ZVS) not only reduces the switching noise of the MOSFETs, but also helps minimize the switching losses. In this paper, zero voltage turning on of the MOSFETs is achieved. We define the instantaneous current value when MOSFETs $S_1$ and $S_4$ are turned off as $I_{OFF}$. $I_{OFF}$ must

$$I_{OFF} > \frac{2C_{oss}V_{in,max}}{t_{dead}}$$

so that the current value is large enough to charge the parasitic output capacitors of $S_1$ and $S_4$ and discharge the parasitic output capacitors of $S_2$ and $S_3$ [14]. In equation (23), $C_{oss}$ is the parasitic output capacitance of the MOSFET, $V_{in,max}$ is the maximum value of the input voltage, and $t_{dead}$ is the dead time. In this paper, 1200 V C2M0025120D Silicon Carbide (SiC) Power MOSFETs from CREE are used. Its output parasitic capacitance is approximately 400 pF, $V_{in,max}$ is 330 volts, and $t_{dead}$ is 200 ns. Therefore, $I_{OFF}$ should be larger than 1.32 A. During the dead time, the body diodes of the MOSFETs clamp the voltages over the parasitic output capacitors. Therefore, the parasitic output capacitors of $S_1$ and $S_4$ are charged to input voltage value and the parasitic output capacitors of $S_2$ and $S_3$ are discharged to almost 0 V. So, when $S_2$ and $S_3$ are turned on right after the dead time, the voltages across them are almost zero and zero voltage turning on of $S_2$ and $S_3$ is achieved. Similarly, zero voltage turning on of $S_1$ and $S_4$ is achieved after half a cycle. Based on [14], the capacitor $C_2$ is the best choice to tune in order to achieve ZVS. It is demonstrated that $C_2$ is tuned to be 8%~11% larger than the calculated value of $C_2$ in the last step. Once the above six steps are done, it is ready to build the wireless charging system and conduct experiments. The design procedures can be summarized in the flowchart as shown in Fig. 10. In particular, the two design criterions of $k_{max}$ and $k_{off}$ are the authors’ experienced values, which are the worst cases to make the system work as well.

IV. EXPERIMENT

The system specifications are given in Table I. Since the resonant frequency is fixed at 85 kHz, AWG-38 copper wires are selected to minimize the skin effect. The input voltage is 330 V, the output battery pack voltage is 300 V, and the output power is 3.0 kW. Therefore, 800-strand AWG-38 litz wires are used. The ferrite is Ferroxcube 3C95, which has outstanding performance at frequencies below 500 kHz. Aluminum plates serve as the electric shielding. The air gap is 150 mm due to the ground clearance of passenger cars. Based on simulations results, the main coils are built and compensated coils are integrated, which are shown in Fig. 11(a) and (b). Fig. 11(c) presents the whole wireless charging system. It shows that it is highly compact with the compensated coil integrated into the system. The circuit parameters are given in Table II. It is convinced that there are only slight differences between the design values and the measured values.
Table III compares the simulated results and measured results of the six coupling coefficients in the wireless charging system. Clearly, the measured results prove that not only are the same-side coupling coefficients $k_{1f}$ and $k_{2f}$ eliminated, but also the cross-side coupling coefficients $k_{1g}$ and $k_{2g}$, the coupling coefficient of compensated coil $k_{fg}$ are minimized to negligible levels. The maximum main coupling coefficient $k$ is achieved at 0.1877. The measured results of $k$ and $k_{fg}$ match their simulated results since they are relatively large when compared to $k_{1f}$, $k_{2f}$, $k_{1g}$, and $k_{2g}$. The percentages of the errors between the simulated results and the measured results for $k_{1f}$, $k_{2f}$, $k_{1g}$, and $k_{2g}$ are large. It is because their values are very small and only a slight misalignment will result in a big difference in results. Fig. 12 further demonstrates that the center of the compensated coil should be aligned with the center of the main coil so that the same-side coupling coefficient is eliminated. Fig. 13 shows how the measured coupling coefficients $k_{1g}$, $k_{1f}$, and $k_{2g}$ vary with the compensated coils rotating. As shown in Fig. 13(a), $k_{1g}$ is minimized when the primary rotating angle equals 90° and secondary rotating angle is 0°, which verifies the simulation result in Fig. 9(a). The measured results in Fig. 13(b) and Fig. 13(c) indicate that $k_{1f}$ and $k_{2g}$ are small enough to be neglected, further confirming the proposed idea. Fig. 14 presents the measured results of the two compensated coils. The inductance values keep constant at the designed values with the angles rotating, which also confirms the simulation results in Fig. 9(b).

### Table I. System Specifications

<table>
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<tr>
<th>Spec/Parameters</th>
<th>Design Value</th>
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<tr>
<td>Input Voltage</td>
<td>330 V</td>
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<tr>
<td>Battery Pack Voltage</td>
<td>300 V</td>
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<tr>
<td>Primary Main Coil Dimension</td>
<td>600 mm × 450 mm × 4 mm</td>
</tr>
<tr>
<td>Secondary Main Coil Dimension</td>
<td>400 mm × 300 mm × 4 mm</td>
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<tr>
<td>Primary Ferrite Plate Dimension</td>
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<td>Secondary Ferrite Plate Dimension</td>
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<td>Primary Shield Dimension</td>
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<tr>
<td>Primary Shield Dimension</td>
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<td>Air Gap</td>
<td>150 mm</td>
</tr>
<tr>
<td>Main Coupling Coefficient</td>
<td>0.1877</td>
</tr>
<tr>
<td>Resonant Frequency</td>
<td>85 kHz</td>
</tr>
<tr>
<td>Maximum Output Power</td>
<td>3.0 kW</td>
</tr>
</tbody>
</table>

### Table II. Circuit Parameters

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Design Value</th>
<th>Measured Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Primary main coil inductance $L_1$</td>
<td>225.35 μH</td>
<td>222.98 μH</td>
</tr>
<tr>
<td>Secondary main coil inductance $L_2$</td>
<td>159.45 μH</td>
<td>157.55 μH</td>
</tr>
<tr>
<td>Primary compensated coil inductance $L_{1f}$</td>
<td>46 μH</td>
<td>44.059 μH</td>
</tr>
<tr>
<td>Secondary compensated coil inductance $L_{2f}$</td>
<td>38 μH</td>
<td>38.04 μH</td>
</tr>
<tr>
<td>Primary resonant capacitor $C_1$</td>
<td>19.59 nF</td>
<td>19.853 nF</td>
</tr>
<tr>
<td>Secondary resonant capacitor $C_2$</td>
<td>31.39 nF</td>
<td>32.71 nF</td>
</tr>
<tr>
<td>Primary compensated capacitor $C_{1g}$</td>
<td>79.68 nF</td>
<td>79.69 nF</td>
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<tr>
<td>Secondary compensated capacitor $C_{2g}$</td>
<td>92.16 nF</td>
<td>92.36 nF</td>
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</tbody>
</table>

Fig. 11. Experiment setup of (a) coils on primary side, (b) coils on secondary side, and (c) wireless charging system.
Fig. 15(a) represents the waveforms of the primary resonant tank’s input voltage $v_{ac}$ and input current $i_{f1}$, the secondary resonant tank’s output voltage $v_o$, and MOSFET gate-to-source voltage $V_{drive}$. In this paper, the purpose of observing $V_{drive}$ is to ensure that the gate-to-source voltage is in its maximum rating all the time since noises may appear during the experiment. Here, $v_{ac}$ leads $i_{f1}$ by 13.4° and the turn-off current $I_{OFF}$ is approximately 4.5 A, which is much larger than 1.32 A and sufficient for charging and discharging the parasitic output capacitors. Fig. 15(b) gives power characteristics of the wireless charging system at the maximum output power. When fully aligned, the system is able to deliver 3.0 kW output power at DC-DC efficiency as high as 95.5%. Further experiments on misalignments are also conducted in the Y-direction, X-direction, and Z-direction separately. Fig. 16(a) to (c) give the respective waveforms. Since the input voltage and the battery pack voltage are fixed at 330 V and 300 V, the waveforms of $v_{ac}$ and $v_o$ remain the same. However, the waveform of $i_{f1}$ gradually deviates from its original waveform shown in Fig. 15(a). When misalignments in different directions occur, the main coupling coefficient $k$ decreases and the output power goes down, as indicated by equation (15). Therefore, the rms value of $i_{f1}$ becomes smaller. Moreover, the mutual inductance $M$ between the main coils decreases and as a result, both the reflected impedance $Z_{refl}$ and input impedance $Z_{in}$ change. The input phase angle between $v_{ac}$ and $i_{f1}$ increases, which causes the turn-off current $I_{OFF}$ to increase and the high-order harmonics components in $i_{f1}$ to rise. The decreased rms value of $i_{f1}$, the increased turn-off current value and high-order harmonics components in $i_{f1}$ result in the waveform distortions when misalignments occur. Fig. 16(a) shows the waveforms when Y-direction misalignment is 150 mm. Since $k$ and $M$ are still comparable to their original values when fully aligned, the waveform distortion of $i_{f1}$ is low. However, when $k$ and $M$ further decreases, the waveform distortion turns high, as shown in Fig. 16(b) and (c). Here, Z-direction misalignment is 50 mm means air gap is increased from 150 mm to 200 mm.
All the experimental results are given by Fig. 17. The solid line in Fig. 17(a) shows the experimental result when the system is fully aligned. It shows great consistency with the simulated efficiency curve in Fig. 6 and the system efficiency is peaked at the maximum output power. The dash-dot line presents the result of the experiment on Z-direction misalignment. The system can transfer 1.76 kW at efficiency of 94.4% when the air gap increases to 200 mm. The solid line with circles gives the system performance on X-direction misalignment. The system delivers 1.57 kW power at an efficiency of 93.8% when the X-direction misalignment increases to 100 mm. The performance is a little weaker and as a result, it is highly recommended to make X-direction compliant with the front-rear direction of the car since it is more convenient for the driver to adjust when the car is parked. The dashed and dotted lines show the system performance in Y-direction. The system performs much better in this direction. When the misalignment even increases to 150 mm, the system can still deliver approximately 2.0 kW at an efficiency of 94.8%. This direction is better to be door-to-door direction since it is hard for drivers to adjust when the car is parked. Fig. 17(a) indicates that when misalignment in different directions occurs, the output power drops 33.33% from the maximum output power in Y-direction, 41.33% in Z-direction, and 47.69% in X-direction. It is because the main coupling coefficient $k$ decreases and other component values remain almost the same, as it is indicated by equation (15). Fig 17(b) shows how the measured main coupling coefficient $k$ decreases as misalignment occurs. When the misalignment increases to 150 mm in Y-direction, $k$ decreases to 0.1244, or 33.72% drop from its original value ($k = 0.1877$); when the misalignment increases to 50 mm in Z-direction, $k$ goes down to 0.1045, or 44.33% drop; when the misalignment increases to 100 mm in X-direction, $k$ declines to 0.1, or 46.72% drop. The differences in the respective dropped percentages between the output power and the main coupling coefficient are caused by the differences in inductance values when misalignment occurs. However, the differences are within 3%, indicating the inductance values only slightly change. Fig. 17(a) and (b) not only verify equation (15), but also validate that extra coupling coefficients are eliminated or minimized to a negligible level. Fig. 17(c) describes how the efficiency varies with misalignments in different directions. It turns out that the efficiency decreases as the misalignment increases. Furthermore, the efficiency curve in Fig. 17(c) shows fairly
In order to achieve that, we will not only analyze the experimental results verify the proposed idea. The compact and good consistency with the main coupling coefficient curve in Fig. 17(b). Experimental results demonstrate the wireless charging system with the new integration method not only has the dual advantage of compactness and high efficiency, but also eliminates the extra coupling effects or minimize them to a negligible level, which greatly simplifies the system analysis and design.

V. CONCLUSIONS

This paper gives a new integrated method of a wireless charging system using double-sided LCC compensation topology. With the compensated coils integrated into the main coil structure, the system becomes much more compact. The proposed compensated coil design further eliminate or minimize the extra coupling effects to a negligible level, making it more straightforward to design a wireless charging system using the double-sided LCC compensation topology. The detailed design procedures to improve system efficiency are also introduced. Both the 3D FEA simulation results and the experimental results verify the proposed idea. The compact and highly efficient wireless charging system is able to deliver 3.0 kW at a DC-DC efficiency of 95.5% with an air gap of 150 mm when fully aligned.

Our future work is to install the designed wireless charger on a vehicle. In order to achieve that, we will not only analyze the additional power loss resulted from ambient objects, such as the EV chassis and the steels buried in the ground, but also optimize the ferrite plates so that minimum ferrite bars are employed to deliver the same amount of power with competitive efficiency.

REFERENCES


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