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A Hybrid Compensation Topology for Battery Charging System Based on IPT Technology

Junfeng Yang ¹, Xiaodong Zhang ^{1,*}, Xu Yang ¹, Qiujiang Liu ¹ and Yi Sun ²

¹ School of Electrical Engineering, Beijing Jiaotong University, Beijing 100044, China; yangjunfeng@bjtu.edu.cn (J.Y.); 17117425@bjtu.edu.cn (X.Y.); 14117383@bjtu.edu.cn (Q.L.)

² China Electrotechnical Society, Beijing 100055, China; sunyi@ces.org.cn

* Correspondence: xdzhang@bjtu.edu.cn; Tel.: +86-51687108

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Abstract: Based on the double-sided LCC (DLCC) compensation topology circuit, a battery charging method is proposed to meet various charging requirements. Firstly, mathematical model was obtained by modeling primary and secondary sides of DLCC. The current gain and voltage gain of the inductive power transfer (IPT) system are derived. Then, taking into account the smooth conversion of charging mode, the parameter configuration conditions for constant current (CC) output and constant voltage (CV) output are designed systematically. Finally, after choosing parameters, the CC and CV modes can be achieved by adding one switch and an auxiliary capacitor. With few additional components and non-sophisticated control, both cost and complexity can be significantly reduced. An experimental prototype with 64 V charging voltage and 1A charging current is built. The experimental results show that the charging voltage and current fluctuation of the system are small and the method can meet the above requirements.

Keywords: inductive power transfer (IPT); LCC compensation network; constant current (CC); constant voltage (CV)

1. Introduction

With the advantages of convenience, safety, and being environment-friendly, inductive power transfer (IPT) has attracted considerable attention in low-power biomedical implants [1], medium-power consumer electronics [2], high-power electric vehicles (EVs) [3–5], and so on [6–8]. The promising technology utilizes a time-varying electromagnetic field to transfer energy without cable. Recently, high performance lithium-ion batteries have been widely used in various energy storage applications due to their high energy density, self-discharge rate, and negligible memory effect [9,10]. The charging method of batteries is one of the key technologies to realize popularization. A typical charging profile of a battery comprises constant output current (CC) charging mode and constant output voltage (CV) charging mode [5,11]. To be specific, initially, the battery is charged in CC mode which means the output current is constant, and the battery voltage increases rapidly. When the battery voltage reaches a specified voltage, the charger changes to the CV mode with the constant output voltage until the battery current reaches nearly zero. The equivalent resistance of the battery varies significantly during the entire charging profile, which makes the design of the IPT charger complicated. To prolong the battery lifetime and recycle time, an IPT charger should provide expected output current and voltage efficiently for different charging stages of the battery.

In order to satisfy the battery charging curve (from CC mode to CV mode), lots of approaches for IPT system have been proposed. These approaches can be roughly divided into two categories: control schemes and hybrid compensation topologies. According to the different control objects, control methods can be classified into primary side control and secondary side control. For secondary side

control, the regulation of output current and voltage depends on an extra DC–DC converter [12]. This approach will lead to the change of the secondary side reflection impedance, which makes the energy picked up by the magnetic coupler vary greatly and its output stabilization ability is limited. A front-end converter added before the primary-side inverters is also applied in [13], however, this approach requires additional components and increases the associated energy losses. Besides, for continued regulation, a wireless communication between the primary side and the secondary side is needed. The high-frequency inverter (HFI) is another critical and indispensable converter to regulate the output voltage of the source. Load-independent output current or output voltage can be realized by phase-shift control (PSC) [14] or variable frequency control (VFC) [9,15–17]. However, it is difficult to achieve zero-voltage switching under light load condition with PSC and the frequency bifurcation phenomenon which would lose controllability and stability, which may occur under variable load with VFC. Obviously, the additional power converter makes the whole system not only inefficient, bulky, and uneconomical, but also more complicated.

Besides the sophisticated control strategies which have been aforementioned, some researchers have focused their research on the new topologies to realize load-independent output current and voltage without control strategies. There are four basic compensation topologies: series–series (SS), series–parallel (SP), parallel–series (PS), and parallel–parallel (PP). Based on the four topologies, some dual hybrid topologies to realize CC and CV modes are put forward [13], [18–20]. In [18], it is realized by changing topology between SS and SP with different parameters, but the circuit is complicated, containing a center-tapped loosed transformer, four switches, and two additional capacitors. In [19], an additional capacitor and two switching at the primary side form a hybrid topology of SS–PS. In [20], after full discussion of series connected and parallel connected compensation circuits, two hybrid topologies are proposed by switching between either SP and PP topology or SS and PS topology. Nevertheless, not only are three switches and an additional inductance introduced, but the wireless link is also required.

Another hybrid topology combining high-order compensation topologies with seven additional components is proposed in [21–23]. In [21], an auxiliary capacitor and two switches are employed at the receiver side, which the basic topology adopts S–LCC with a series (S) compensation network on the primary side and an inductor-capacitor-capacitor (LCC) compensation network on the secondary. In [22], CC and CV outputs are realized by switching between double-sided LCC (DLCC) and LCC–S modes. The output of HFI can achieve zero phase angle (ZPA) when using two switches and an auxiliary capacitor. Based on double-sided inductor-capacitor-inductor (DLCL) compensation alteration, CC mode and CV mode are achieved in [23]. The compensation tank consists of one additional switch, one auxiliary capacitor, and two auxiliary inductors. What is more, the process of switching is not analyzed without considering continuous charging under two modes.

Because of the lack of freedom, the four basic compensations may not be able to meet the required output, which is difficult to optimize design. Thus, higher order compensation topologies are better to apply to implement CC and CV output. Although some have been used in aforementioned work, each has its own shortcomings. Since the LCC topology has three parameters design freedoms [24–26], it will be better to implement this by changing the components of the topology. This paper develops a new IPT charging method to meet various charging requirements, including five sections. In Section 2, simplified circuits of DLCC topology and the output characteristics are theoretically analyzed. Detailed systematic analysis of parameters to achieve constant voltage and current output is also presented. Considering the smooth transition between two modes, the parametric design method of LCC network is carried out in Section 3. In Section 4, an experimental prototype is implemented and experimental results are presented in detail to evaluate the performance of the designed IPT system. The experimental results are in good agreement with the performance of the method, indicating that the method proposed in the paper is effective and feasible.

2. Analysis of Circuits

2.1. System Topology

As Figure 1 shows, the typical IPT charging system consists of three parts; power supply, resonant tank, and load. The power supply consists of a constant voltage source and a high-frequency full-bridge square-wave inverter. The resonant tank includes primary and secondary compensation networks, and magnetic coupled coils. The load includes a full bridge rectifier, a filter, and a battery load.

Figure 1. Wireless battery charging system.

In this paper, the resonant tank is what we have focussed on. The input voltage of the resonator is the square wave obtained by the high frequency inverter (HFI). Owing to the filtering function of compensation network, the current of primary coil is nearly sinusoidal, containing only the fundamental component. As a result, the fundamental harmonic approximation (FHA) is employed. When the duty cycle of the inverter is 50%, the relationship between the inverters' AC output RMS voltage U_{in} and the input DC voltage U_d can be expressed in the equation as follows:

$$U_{in} = \frac{2\sqrt{2}}{\pi} U_d \quad (1)$$

The relationship between input voltage U_O , current I_O of full bridge rectifier, and charging voltage U_B , I_B at the power receiver can be expressed as follows:

$$\begin{cases} U_O = \frac{2\sqrt{2}}{\pi} U_B \\ I_O = \frac{\sqrt{2}\pi}{4} I_B \end{cases} \quad (2)$$

The equivalent resistance of the battery load to the resonant tank is:

$$R_O = \frac{8}{\pi^2} R_B \quad (3)$$

To facilitate the analysis of the inductive power transfer characteristics, the fundamental wave equivalent only containing the compensation topology and the coil system was considered. U_{in} , I_{in} , U_O , I_O , R_O derived from Figure 1 were adopted to analyze at the primary side and the secondary side.

2.2. Realization of CC Mode and CV Mode

Varied from T-type network, the inductor-capacitor-inductor (LCL) compensation network is promoted widely for its good output characteristics. By adding another capacitance to the coil branch, which decreases the device pressure of compensate inductance and improves the power transmission capability, the LCC compensation network was built. According to the T-type compensation, the equivalent circuit of double-side LCC is put forward in Figure 2. M is the mutual inductance of transmitting and receiving coils, and the parasitic resistance of the capacitance and inductance coil is ignored. U_{in} is the AC input voltage source after HFI. L_P and L_S represent the self-inductance of

the transmitting and receiving coils, respectively. L_1 , C_1 , and C_2 form the primary side LCC resonant compensation network, and L_4 , C_3 , and C_4 form the secondary side LCC resonant compensation network. In order to facilitate the analysis, L_2 was used instead of the branch composed of C_2 and L_p , and L_3 was used instead of the branch composed of C_3 and L_s .

Figure 2. Circuit of DLCC compensation topology.

Firstly, the primary side was analyzed. Assuming that the secondary impedance is Z_s and the primary side impedance reflected from the secondary side is Z_r , the overall input impedance of HFI is

$$Z_{in} = j\omega L_1 + \left(\frac{1}{j\omega C_1}\right) \left(\frac{1}{j\omega C_2} + j\omega L_p + Z_r\right) = \frac{Z_r + j\omega(L_2 A_1 - C_2 Z_r^2 + L_1 A_1^2 + L_1 A_2^2)}{A_1^2 + A_2^2} \quad (4)$$

where

$$Z_r = \frac{(\omega M)^2}{Z_s} A_1 = 1 - \omega^2 C_1 L_2 A_2 = \omega C_1 Z_r L_2 = L_p - \frac{1}{\omega^2 C_2} \quad (5)$$

The ω is the angular frequency. The current flowing through the transmitting coil can be expressed as

$$I_p = \frac{U_{in}}{(1 - \omega^2 C_1 L_1) Z_r - j\omega(\omega^2 C_1 L_2 L_1 - L_1 - L_2)} \quad (6)$$

The symmetrical T-type is employed as the primary side compensation network. The resonance condition of the parameters for the equivalent circuit are given by

$$j\omega L_1 + \frac{1}{j\omega C_1} = 0 \quad (7)$$

$$j\omega L_p + \frac{1}{j\omega C_1} + \frac{1}{j\omega C_2} = 0 \quad (8)$$

From (4) and (6), the input impedance Z_{in} and the current I_p at the resonance condition can be derived as

$$Z_{in} = \frac{(\omega L_1)^2}{Z_r} \quad (9)$$

$$I_p = \frac{U_{in}}{j\omega L_1} \quad (10)$$

It can be seen that the input impedance is pure resistance when Z_r is resistive. Zero phase angle (ZPA) between output voltage and output current of the HFI can be achieved. The current I_p of the primary transmitting coil is only related to the input voltage U_{in} and L_1 with the load-independent characteristic. In order to reduce the communication link between the primary and the secondary side, the primary side adopts symmetrical LCC compensation network to provide a constant voltage source for the receiving coil and the induced voltage of receiving coil is

$$U_s = j\omega M I_p \quad (11)$$

Constant current output and constant voltage output is achieved by changing the topology of the secondary side. From Figure 2, there are three components in the T-type network besides the inductance of receiving coil. The output characteristics are different under asymmetric conditions. In order to get constant current output and constant voltage output, the impedances of X_{C3} , X_{C4} , and X_{L4} for the left side, the down side, and the right side in a general LCC compensation network are introduced here. The impedance of receiving coil X_{LS} is set as reference quantity. Then the T-type provides three freedom degrees to adjust the transmission characteristic: λ_1 is the ratio of the capacitance C_3 in the left side to X_{LS} , λ_2 is the ratio of the capacitance C_4 in the down side to X_{LS} , and λ_3 is the ratio of the inductance L_4 in the right side to X_{LS} . λ_1 , λ_2 , and λ_3 are defined as follows:

$$\lambda_1 = \frac{X_{C3}}{X_{LS}} \quad \lambda_2 = \frac{X_{C4}}{X_{LS}} \quad \lambda_3 = \frac{X_{L4}}{X_{LS}} \quad (12)$$

Similar to the primary side, the receiver side equivalent impedance is

$$Z_s = j\omega L_s + \frac{1}{j\omega C_3} + \left(\frac{1}{j\omega C_4}\right)(j\omega L_4 + R_o) = \frac{B_1 X_{LS} + jB_2 X_{LS} R_o}{jB_3 X_{LS} + R_o} \quad (13)$$

where

$$B_1 = \lambda_2 - \lambda_3 - \lambda_1 \lambda_2 + \lambda_1 \lambda_3 + \lambda_2 \lambda_3 \quad B_2 = 1 - \lambda_1 - \lambda_2 \quad B_3 = \lambda_3 - \lambda_2 \quad (14)$$

The voltage and current of load are derived from the following formula:

$$I_o = \frac{U_s}{(1 - \omega^2 L_3 C_4) R_o - j\omega(\omega^2 L_3 L_4 C_4 - L_3 - L_4)} \quad (15)$$

$$U_o = \frac{U_s R_o}{(1 - \omega^2 L_3 C_4) R_o - j\omega(\omega^2 L_3 L_4 C_4 - L_3 - L_4)} \quad (16)$$

By substituting (12) into (15) and (16), we have

$$I_o = \frac{\lambda_2 U_s}{(\lambda_1 + \lambda_2 - 1) R_o - j(\lambda_3 - \lambda_2 + \lambda_1 \lambda_2 - \lambda_1 \lambda_3 - \lambda_2 \lambda_3) X_{LS}} \quad (17)$$

$$U_o = \frac{\lambda_2 U_s R_o}{(\lambda_1 + \lambda_2 - 1) R_o - j(\lambda_3 - \lambda_2 + \lambda_1 \lambda_2 - \lambda_1 \lambda_3 - \lambda_2 \lambda_3) X_{LS}} \quad (18)$$

To achieve constant current output (CC mode) independent of load, the coefficient of R_o in (17) should be equal to zero, and we get

$$\lambda_1 = 1 - \lambda_2 \quad (19)$$

The relationship between C_3 , C_4 , and L_s in CC mode can be expressed as

$$j\omega L_s + \frac{1}{j\omega C_3} + \frac{1}{j\omega C_4} = 0 \quad (20)$$

Then, the current I_o can be calculated as

$$I_o = \frac{U_s}{j(1 - \lambda_1)\omega L_s} = \frac{U_s}{j\lambda_2\omega L_s} \quad (21)$$

From (21), the condition of constant current is also independent of the compensation inductance λ_3 . Accordingly, by substituting (10) and (11) into (21), the ratio of the output current I_o to the input voltage U_{in} , which is marked as the transconductance gain G_{CC} , is derived as

$$G_{CC} = \frac{M}{j(1 - \lambda_1)\omega L_1 L_s} \quad (22)$$

Transconductance gain G_{CC} is load-independent constant at a certain input voltage U_{in} and a certain frequency f . It can be easily deduced that the CC mode can be implemented at this coefficient regardless of the variable load. The value of Z_S can be simplified as

$$Z_s = \frac{(1 - 2\lambda_1 + \lambda_1^2)X_{LS}}{j(\lambda_1 + \lambda_3 - 1)X_{LS} + R_O} \quad (23)$$

Taking $\lambda_2 = \lambda_3$, which means the down hand capacitor C_4 and the right-hand inductance L_4 is under resonant condition, the pure resistance of Z_S can be realized and the solution can be obtained as follows:

$$Z_s = \frac{(1 - 2\lambda_1 + \lambda_1^2)X_{LS}}{R_O} \quad (24)$$

To realize the constant voltage output (CV mode) against variable resistive load R_O , the coefficient of X_{LS} in (18) should be set to zero, and we get

$$\lambda_3 = \frac{(\lambda_1 - 1)\lambda_2}{\lambda_1 + \lambda_2 - 1} \quad (25)$$

From (25), it can be seen that if the sum of λ_1 and λ_2 is equal to 1, the λ_3 cannot be regulated and the CV mode cannot be realized. Therefore, the (20) cannot be maintained. The relationship between C_3 , C_4 , L_4 , and L_S in CV mode can be expressed as

$$(j\omega L_S + \frac{1}{j\omega C_3})(j\omega L_4 + \frac{1}{j\omega C_4}) - \frac{L_4}{C_4} = 0 \quad (26)$$

Then, the output voltage U_O can be calculated as

$$U_O = \frac{\lambda_2 U_S}{\lambda_1 + \lambda_2 - 1} \quad (27)$$

The voltage gain G_{CV} is given by substituting (10), (11) into (27) as

$$G_{CV} = \frac{\lambda_2 M}{(\lambda_1 + \lambda_2 - 1)L_1} \quad (28)$$

The coefficients of B_1 , B_2 , B_3 in (14) and the Z_S in (13) can be simplified as

$$B_1 = 0 \quad B_2 = 1 - \lambda_1 - \lambda_2 \quad B_3 = \frac{\lambda_2}{1 - \lambda_1} \quad (29)$$

$$Z_s = \frac{(1 - \lambda_1 - \lambda_2)(1 - \lambda_1)X_{LS}R_O[\lambda_2 X_{LS} + j(1 - \lambda_1)R_O]}{\lambda_2^2 X_{LS}^2 + (1 - \lambda_1)^2 R_O^2} \quad (30)$$

From (19), it can be seen that by reasonably configuring the parameters of λ_1 and λ_2 , the load-independent constant current output can be achieved. The CC mode is independent of L_4 but L_4 has an effect on the receiver equivalent impedance. From (25), by reasonably configuring the parameters of λ_1 , λ_2 , and λ_3 , the load-independent constant voltage output can be achieved, and the receiver equivalent impedance is related to these three variables.

2.3. Parameters Analysis

In order to analyze the influence of key parameters on the system characteristics and provide a theoretical basis for the reasonable configuration of parameters, the effects of f , λ_1 , λ_2 , and λ_3 on the output of IPT system were analyzed as follows. For the size of transmitting and receiving coils fixed in practical application, L_P and L_S were set as a reference value to design other parameters. In order

to provide a constant voltage source for the secondary side, the symmetrical T-type compensation network is adopted in the primary side. The original parameters of the analysis are shown in Table 1.

Table 1. System specification and parameter values.

Symbol	Value	Symbol	Value
L_1	19.95 μH	L_S	147.6 μH
L_P	603.8 μH	L_4	106.24 μH
C_1	183.02 nF	C_3	80.94 nF
C_2	5.94 nF	C_4	33 nF
U_i	40 V	R_O	40–100 Ω
M	34.21 μH	f	85 kHz

From (20) and (26), the resonant frequency f_{CC} in CC mode and f_{CV} CV mode can be derived as

$$f_{CC} = \frac{1}{2\pi} \sqrt{\frac{C_3 + C_4}{L_S C_3 C_4}} \quad (31)$$

$$f_{CV} = \frac{1}{2\pi} \sqrt{\frac{D_1 \pm \sqrt{D_1^2 - 4D_2}}{2L_S L_4}} \quad (32)$$

where

$$D_1 = \frac{L_S}{C_4} + \frac{L_4}{C_3} + \frac{L_4}{C_4} \quad D_2 = \frac{L_S L_4}{C_3 C_4} \quad (33)$$

Figure 3 shows the constant current output and constant voltage output can be achieved under different parameters with specific resonant frequency. The output characteristics vary with f for different loads, but at a specific resonant frequency, the output current or voltage are load independent. In Figure 3a, λ_1 , λ_2 , and λ_3 are set at 0.28, 0.72, and 0.72, respectively. It can be seen that there is a resonant frequency at about 85 kHz for CC mode and 115.6/54.5 kHz for CV mode. The constant current output is 1.1 A and the constant voltage output is 54 V and 56 V, respectively. Moreover, changing the value of inductor L_4 to make $X_{L4} = X_{C3}$, there are other resonant frequencies. CC mode at 85 kHz with 2.8 A constant current output, and CV mode at 90.9/43.5 kHz with 100/24 V constant voltage output. It is shown that different L_4 does not affect the resonant point of the CC mode. The resonant frequency to realize CV mode is different, with frequency bifurcation phenomenon. What is more, if the parameters are not carefully chosen, the outputs of the system are very sensitive to frequency. It is unfavourable to IPT system. This method needs to vary resonant frequency, which increases the cost of control [9,17].

The inductance L_S is related to the design of the magnetic coupler, i.e., the receiving coil. Considering the transmission capacity and efficiency, the receiving coil is fixed and can not be designed too small. In order to reduce the size and cost of compensation network parameters, parameter λ_1 was introduced. The current output curve versus λ_1 with different loads in DLCC compensation network is shown in Figure 4. For a given λ_2 , there is a λ_1 so that the output current is independent of the load, as shown in Figure 4. When λ_2 equals 0.72 and λ_1 equals 0.28, the constant current output is 1.48 A. The larger deviation of λ_1 from 0.28, the more obvious the change of current with load, and the smaller the current is. When λ_2 equals 0.28 and λ_1 equals 0.72, the constant output current is 9 A. With the deviation increase of λ_1 , the current output has the same trend, but the stability is worse. It is shown that with the specific resonant frequency and source voltage, the constant current output is proportional to λ_1 and is inversely proportional to λ_2 .

Figure 5 shows the current output curve versus the different normalized parameters of λ_1 , λ_2 , and λ_3 with different load conditions in CC mode. The four subgraphs in Figure 5 correspond to 20 Ω , 40 Ω , 60 Ω , and 80 Ω , respectively. The sensitivity of parameters to the output current was studied. It can be seen from the brown dotted line in Figure 5 that the current output is independent of λ_3 . Under different loads, the variation of λ_1 has a relatively small influence on current output and it is not

sensitive. With the increase of load, the sensitivity of output current to λ_2 decreases and λ_1 increases. However, the sensitivity of λ_2 is greater than that of λ_1 overall, obviously.

(a)

(b)

Figure 3. Relationship between I_O , U_O and f . (a) $\lambda_3 = \lambda_2 = 1 - \lambda_1 = 0.72$; (b) $\lambda_3 = \lambda_2 = 1 - \lambda_1 = 0.28$.

Figure 4. Relationship between I_O and λ_1 , λ_2 .

Figure 5. I_O versus normalized parameters.

The voltage output curve versus λ_2 with different loads is shown in Figure 6. It is shown that different parameters can achieve different load-independent constant voltage outputs. The same load-independent constant voltage output can also be achieved by different parameters. When the output voltage is set to 64 V, there are two sets of implementation parameters, i.e., $\lambda_1 = 0.28$, $\lambda_2 = 0.36$, $\lambda_3 = 0.72$, and $\lambda_1 = 0.44$, $\lambda_2 = 0.28$, $\lambda_3 = 0.56$. When the output voltage is set to 100 V, it can be achieved by parameters $\lambda_1 = 0.54$, $\lambda_2 = 0.28$, $\lambda_3 = 0.72$. Figure 7 shows constant voltage output capability under different parameters. With the increase of λ_1 , the output voltage capacity becomes larger, but λ_1 is limited to less than 0.72. The smaller the λ_1 is, the smaller the sensitivity of output voltage U_O to parameter λ_1 is and the more stable it is. Under the same condition of λ_1 , the output capacity increases with the increase of λ_2 . With the increase of λ_2 , the starting voltage satisfying CV mode increases, and the range of λ_1 satisfying constant voltage decreases. It cannot meet the requirements of low constant voltage output.

Figure 6. Relationship between of U_O and λ_1 , λ_2 .

Figure 7. U_O versus λ_1, λ_2 in CV mode.

Taking into account the primary side, the input phase angle between the output current I_{in} and voltage U_{in} of the HFI is the main concern and is analyzed in the following. For the DLCC compensation network, the input impedances of HFI are different in CC and CV modes for different values of λ_1 and λ_2 . The receiver equivalent impedance Z_S can be calculated by (23) in CC mode and (30) in CV mode. By substituting (23) and (30) into (9), the input impedances Z_{CC} in CC mode and Z_{CV} in CV mode can be obtained as

$$Z_{CC} = \left(\frac{L_1}{M}\right)^2 \frac{(1 - 2\lambda_1 + \lambda_1^2)X_{LS}}{j(\lambda_1 + \lambda_3 - 1)X_{LS} + R_O} \quad (34)$$

$$Z_{CV} = \left(\frac{L_1}{M}\right)^2 \frac{(1 - \lambda_1 - \lambda_2)(1 - \lambda_1)X_{LS}R_O[\lambda_2 X_{LS} + j(1 - \lambda_1)R_O]}{\lambda_2^2 X_{LS}^2 + (1 - \lambda_1)^2 R_O^2} \quad (35)$$

The input phase angle θ_i can be derived as

$$\theta_{CC/V} = \frac{180^\circ}{\pi} \arctan \left[\frac{\text{Im}(Z_{CC/V})}{\text{Re}(Z_{CC/V})} \right] \quad (36)$$

where the “Im” is imaginary components of the variable and the “Re” is resistive components. The results of the input phase angle in CC mode (θ_{CC}) and CV mode (θ_{CV}) are shown in Figures 8 and 9.

Figure 8. Phase angle of input impedance in CC mode.

Figure 9. Phase angle of input impedance in CV mode.

According to (24), the constant current output can make the secondary impedance pure resistance by the change of λ_3 . From Figure 8, when λ_3 equals 0.72 (marked as λ_{3_R}), no matter what the values of λ_1 and λ_2 are, the input impedance is pure resistance, which means no reactive power in the IPT system. The input impedance of the HFI is inductive when λ_3 is less than λ_{3_R} and capacitive when λ_3 is greater than λ_{3_R} . Additionally, it is also shown that the phase angle decreases with the increase of load.

The phase angle of the IPT system in CV mode is shown in Figure 9. The input impedance cannot be pure resistance. When λ_2 equals 0.72 (marked as λ_{2_R}), the phase difference is 90 degrees. The input impedance of the HFI is inductive when λ_2 is less than λ_{2_R} and capacitive when λ_2 is greater than λ_{2_R} . The larger the difference between λ_2 and λ_{2_R} , the smaller phase angle between the output current I_{in} and voltage U_{in} of the HFI. Hard switching will occur and the switching loss will increase if the input impedance of the IPT system is capacitive. Therefore, λ_2 less than λ_{2_R} is adopted in practical application.

According to the transconductance gain G_{CC} (22) and the voltage gain G_{CV} (28), the relationship between battery charging current, charging voltage, and source voltage of IPT system can be expressed as

$$G_{BC} = \frac{8}{\pi^2} G_{CC} = \frac{8}{\pi^2} \frac{M}{j(1 - \lambda_1)\omega L_1 L_5} \quad (37)$$

$$G_{BV} = G_{CV} = \frac{\lambda_2 M}{(\lambda_1 + \lambda_2 - 1)L_1} \quad (38)$$

The current output in CC mode and voltage output in CV mode are related to the output voltage of HFI under the condition of fixed parameters. By adjusting the output voltage of HFI, the output voltage and current can be controlled, realizing the charging target of multi-voltage and current levels.

3. Parameters Design

From the above analysis, the constant voltage output and constant current output can be realized by configuring different parameters using DLCC compensation. The parameter λ_1 related to C_3 and the parameter λ_2 related to C_4 for achieving constant current output and constant voltage output were studied, respectively. For a battery that needs continuous charging, it is necessary to take into account the smooth conversion of charging mode. Assume the charging voltage and current of the battery are U_B and I_B . At the particular moment t , the charging mode is switched from CC mode to CV mode, and the switching load is R_O' . The voltage of the R_O in CC mode at time t_- is marked as $U_{CC}(t_-)$, and the voltage in CV mode at time t_+ is marked as $U_{CV}(t_+)$. There are two ways to change the charging mode

by changing one parameter. With the constant input voltage, the voltages of batteries in CC mode and CV mode are derived, respectively, as

$$U_{CC}(t_-) = I(t_-)R'_O = \frac{U_S R'_O}{(1 - \lambda_1)\omega L_S} = \frac{U_S R'_O}{\lambda_2 \omega L_S} \quad (39)$$

$$U_{CV}(t_+) = \frac{\lambda_2 U_S}{\lambda_1 + \lambda_2 - 1} = \frac{\lambda_3 U_S}{1 - \lambda_1} = \left(1 - \frac{\lambda_3}{\lambda_2}\right) U_S \quad (40)$$

where

$$R'_O = \frac{8}{\pi^2} \frac{U_B}{I_B} \quad (41)$$

During switching operation, $U_{CC}(t_-)$ should be equal to $U_{CV}(t_+)$. If it is achieved by changing λ_2 , the parameters are related by

$$\frac{U_S R'_O}{(1 - \lambda_1)\omega L_S} = \frac{\lambda_3 U_S}{1 - \lambda_1} \quad (42)$$

The parameter λ_3 can be obtained as

$$\lambda_3 = \frac{8}{\pi^2} \frac{U_B}{\omega L_S I_B} \quad (43)$$

Therefore, for a given battery known as constant charging voltage U_B and the constant charging current I_B , the right-hand compensation inductor L_4 can be derived firstly by changing λ_2 as

$$L_4 = \frac{8}{\pi^2} \frac{U_B}{\omega I_B} \quad (44)$$

According to (2) and (21), the value of λ_1 can be derived by

$$\lambda_1 = 1 - \frac{2\sqrt{2}}{\pi} \frac{U_S}{\omega L_S I_B} \quad (45)$$

After we get λ_1 and λ_3 , the parameter λ_2 for CC mode and λ_2 can be given directly by solving (19) and (25) as

$$\lambda_{2_CC} = 1 - \lambda_1 \quad (46)$$

$$\lambda_{2_CV} = \frac{(\lambda_1 - 1)\lambda_3}{\lambda_1 - \lambda_3 - 1} \quad (47)$$

If the charging profile is achieved by changing λ_1 , the value of λ_3 should be chosen according to

$$\frac{U_S R'_O}{\lambda_2 \omega L_S} = \left(1 - \frac{\lambda_3}{\lambda_2}\right) U_S \quad (48)$$

By solving (48), the parameter λ_3 is given by

$$\lambda_3 = \lambda_2 + \frac{8}{\pi^2} \frac{U_B}{\omega L_S I_B} \quad (49)$$

From (49), λ_2 should be calculated firstly by solving (26)

$$\lambda_2 = \frac{2\sqrt{2}}{\pi} \frac{U_S}{\omega L_S I_B} \quad (50)$$

By substituting (50) into (49), L_4 is given as following

$$L_4 = \frac{2\sqrt{2}U_S + \frac{8}{\pi^2}U_B}{\omega I_B} \quad (51)$$

The parameter λ_1 for CC mode and λ_2 can be given by solving (19) and (25) as

$$\lambda_{1_CC} = 1 - \lambda_2 \quad (52)$$

$$\lambda_{1_CV} = 1 + \frac{\lambda_2 \lambda_3}{\lambda_2 - \lambda_3} \quad (53)$$

After we get λ_1 and λ_2 , the corresponding compensation capacitance C_3 and C_4 can be obtained by solving (12). A design method of DLCC compensated IPT system is proposed in this section to realize CC output and CV output. For a given input voltage and an output charging profile, two methods with different C_3 , C_4 , and L_4 can be obtained by the following flowchart shown in Figure 10.

Figure 10. Flowchart of parameters design.

According to the comprehensive analysis of the parameter characteristics in Section 2, more than one set of parameters can be achieved for a specific output voltage in the CV mode. As shown in Figure 10, two methods achieve constant voltage output by changing the capacitance parameters: the parallel capacitance of C'_4 or C'_3 . The left branch of the flowchart is the first method by changing C_4 only, and the right branch is the second method by changing C_3 only. The two methods are implemented in different ways and require different components. The first method requires an additional switch and an auxiliary capacitor to realize the change from CC mode to CV mode. The λ_3 of CC mode makes the IPT system realize ZPA. From (21), it can be seen that the value of λ_3 in CC mode does not affect the constant current output. If the ZPA of the IPT system is not considered, λ_3 and λ_2 can be calculated in CV mode firstly, and change λ_1 to achieve the conversion between the two modes by an auxiliary capacitor and an additional switch. However, this method increases the reactive power of the system. To realize ZPA in CC mode by changing λ_1 , λ_3 also needs to be changed at the same time. Comparing with the above two methods, two parameters in the CV mode have changed, i.e., two

additional switches and an auxiliary capacitor and an auxiliary inductor are needed. This method is not adopted for the increasing of cost and complexity.

The input voltage U_d is 40 V and the operating frequency of IPT system set to 85 kHz. For a given receiving coil set to be 106.24 μH , two sets (two for each set) of parameters were designed according to Figure 11. The values of inductance and capacitance are listed in Tables 2 and 3. Both methods need an additional switch operation to change from CC mode to CV mode, but the first method can realize ZPA in CC mode. From the analysis of the parameters in the Section 2, it can be seen that the stability of the system is affected by the sensitivity of different parameters to operational frequency and parameter offset. It can be seen from Figures 3 and 4 that the frequency offset and parameter offset has less influence on the current output when $\lambda_1 = 0.28$. From Figure 7, it can be seen that smaller λ_1 is beneficial to the stability of constant voltage output. Consequently, the next part of the experimental platform adopts the first method.

Figure 11. Two methods of charging mode conversion.

Table 2. Theoretically calculated parameter values of first method.

λ_1	λ_2 for CC Mode	λ_2 for CV Mode	λ_3
0.28	0.72	0.36	0.72
C_3 80.94 nF	C_4 33 nF	C_4 66 nF	L_4 106.24 μH

Table 3. Theoretically calculated parameter values of second method.

λ_1 for CC Mode	λ_1 for CV Mode	λ_2	λ_3
0.72	0.44	0.28	0.56
C_3 33 nF	C_3 51.51 nF	C_4 80.94 nF	L_4 82.63 μH

4. Experimental Verification

To verify the analysis above, an experimental prototype shown in Figure 12 was built. The value of L_4 can be directly calculated by battery parameters and is independent of other parameters through the first method, as shown in Figure 10. So, the experimental platform is set up according to Figure 11. The parameters of the compensation networks are provided in Table 2. The dc input voltage U_d is 40 V. The primary and secondary coils are both round structures, the diameters are 75 mm and 35 mm, respectively. The air gap is set to 8 mm. Switches $S_{1,2,3,4}$ of full bridge inverter are IPW90R120C3 and secondary rectifier diodes are DSEI2 $\times$ 101. The output filter uses C_f of 560 μF . The internal resistance of battery can be equivalent to the non-linear resistance [15,16]. In this experiment, an electronic load with wide range was used to simulate the equivalent behavior of batteries during charging. For simplicity, the whole charging profile is set to two sections: charging current I_B 1A and charging voltage U_B 64 V.

Figure 12. Prototype of the WPT charging system.

Firstly, the results of CC mode were analyzed. With different load resistance, the output waveforms u_{in} , i_{in} of the inverter, and the current i_p are shown in Figure 13. The transient waveforms u_B and i_B with load resistance switching from $40\ \Omega$ to $60\ \Omega$ are shown in Figure 14. It can be seen that under the symmetrical T-type compensation parameters, the waveform of i_{in} is always nearly in phase with u_{in} . Nearly zero reactive power is generated in this case of well compensation. The current flowing through the transmitting coil is independent of the load. In Figure 14, the charging current I_B of IPT is 1.0 A at $R_B = 40\ \Omega$ and 994 mA at $60\ \Omega$. The fluctuation margin of charging current is less than 1% and the charging process is basically constant. After the circuit is stabilized, the load voltage increases with the increase of the load.

Figure 13. Experimental waveforms of u_{in} , i_{in} , and i_p in CC mode when (a) $R_B = 40\ \Omega$ and (b) $R_B = 60\ \Omega$.

Figure 14. Transient waveforms switch from $40\ \Omega$ to $60\ \Omega$ at CC mode.

The results of CV mode are shown in Figures 15 and 16. As shown in Figure 11, the IPT system operates in CV mode when switch S is ON, in which the capacitor changes from C_4 to C_{4_CV} by paralleled connection of C_4' to C_4 . The value of equivalent capacitor is given as

$$C_{4_CV} = C_4 + C_4' \quad (54)$$

Figure 15. Experimental waveforms of u_{in} , i_{in} in CV mode when (a) $R_B = 70 \Omega$ and (b) $R_B = 100 \Omega$.

Figure 16. Transient waveforms from 70Ω switch to 100Ω at CV mode.

The capacitance C_4 changes from $33 \mu\text{H}$ to $66 \mu\text{H}$. Figure 15 shows the output waveforms u_{in} , i_{in} of the inverter, and the output waveforms u_B , i_B of load under different resistance. The waveform of i_{in} lags behind u_{in} , which illustrates that IPT is charging with reactive power at CV mode. In Figure 16, the charging voltage u_B of IPT system is 63.9 V at $R_B = 70 \Omega$ and 64 V at 100Ω . It remains constant after the circuit is stabilized, and the load current decreases with the increase of the load. The fluctuation margin of charging voltage is less than 1% and the charging process is basically constant.

Figure 17 shows the charging mode changing from CC mode to CV mode, while the battery voltage u_B and current i_B change smoothly during the mode transition. In CC mode, the charging current i_B varies from 1.11 A at $R_B = 40 \Omega$ to 1.09 A at $R_B = 64 \Omega$, and the fluctuation margin of i_B is 1.8%. The charging voltage u_B varies from 63.5 V at $R_B = 67 \Omega$ to 64.9 V at $R_B = 100 \Omega$, and the fluctuation margin of u_B is less than 2.5%. The whole charging profile, including charging current I_B and charging voltage U_B versus R_B , is shown in Figure 18. At the beginning of charging, the battery current is maintained at 1 A . With the increase of the equivalent battery load, the battery voltage increases to the required 64 V and then remains stable.

Figure 17. Transient waveforms of i_{in} , u_B and i_B from CC mode to CV mode.

Figure 18. Measured charging profile versus R_B .

5. Conclusions

In this study, a DLCC compensated IPT system was designed to achieve load independent constant current output and constant voltage output for battery charging applications. The characteristics of LCC compensate network with three design freedoms has been analyzed in detail. According to the theoretical analysis, two parametric design methods of realizing CC mode and CV mode have been proposed. With one switch and an additional capacitor, a smooth mode transition can be achieved. The fluctuation margins of charging current and charging voltage are both less than 2.5%, which meets the charging requirements of batteries. The experimental results agree well with the performance of the proposed method which demonstrates that it is feasible to a given charging characteristics of battery.

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