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Analysis and Design of Capacitive Power Transfer System with Robustness to Load Variations and Capacitor Plate Misalignment

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Abstract: This paper proposes a capacitive power transfer system (CPT), which is robust to misalignment of coupling capacitor plates and load variations. The CPT system needs to transmit power at high frequencies. Therefore, it is essentially sensitive to variations in the resonant network, namely resonant component tolerances and load variations. It is required to maintain ZVS and a constant output voltage against these variations. The basic concept of the proposed CPT system is to design the load-independent (LI) resonant dc-dc converter and insert a coupling capacitance into the resonant network. If the coupling capacitance is sufficiently high compared to the resonant capacitance, the variation of the reactance component can be ignored. Therefore, the proposed CPT maintains the LI characteristic, namely, achieving ZVS and constant output voltage against load variations and plate misalignment without requiring control or feedback communication. This paper gives analytical expressions of the proposed system and the design procedure. Experimental results showed the effectiveness of the proposed CPT system along with its design procedure, as well as the validity of the analysis.

Keywords: capacitive coupling; load-independent; misalignment; wireless power transfer

1. Introduction

The appearance of wide bandgap semiconductor devices enables the operating frequency of power converters to be higher and is a trigger to accelerate the development of high-frequency resonant converters. Wireless power transfer (WPT) systems are regarded as resonant converter family. Therefore, a high-frequency WPT system, whose transmission frequency is in the megahertz range, is also attracting many researchers and developers [1–20]. Examples of applications of high-frequency WPT systems include EVs [21,22], implantable devices [4,5], drones [6,7], and industrial robot arms [8].

A capacitive power transfer (CPT) system transmits power through the electric field between the plates. The electric field for power transmission is mainly limited to the area between the plates [5–7,9–12,18–21]. Therefore, CPT systems can be expected to be highly safe in metallic environments [4,5]. CPT systems use the coupling capacitor as the resonant capacitor. Since large capacitance is difficult to realize, the operating frequency shifts, requiring high-frequency inverters such as class-E, EF, or Φ .

CPT systems can be equivalently modeled as high-frequency non-isolated resonant converters, which are highly sensitive to component tolerances in the resonant network [20]. Therefore, the effects of load variations and coupling-capacitor plate misalignment must be considered in the design of CPT systems. CPT systems are often



required to simultaneously maintain constant output voltage and zero-voltage switching (ZVS) under load variations and coupling-capacitor plate misalignment. Generally, feedback control is adopted to achieve the requirements [10–12]. However, because wireless communication is mandatory for feedback, communication delays and packet losses occur, which complicate and make the control system design particularly difficult at high frequencies. Therefore, it is required to design CPT systems, achieving constant output voltage and ZVS without wireless communication feedback.

In recent years, load-independent (LI) technology has been attracting the attention of many researchers and developers [13–19]. Several magnetic-resonant LI WPT systems have been proposed, which maintain ZVS and a constant output voltage against load variations without any control. By applying perfect resonance to the transmission frequency and a post-regulator at the receiver side, magnetic-resonant LI WPT systems can also be made robust to coil misalignment [17]. There is, however, no CPT system that is robust to load variations and plate misalignment.

This paper proposes a design strategy of the CPT system with robustness to load variations and capacitor-plate misalignment. First, we design the LI non-isolated high-frequency resonant converter. To the authors' knowledge, there is no design guideline for LI non-isolated resonant converters to date. In addition, the coupling capacitance is inserted into the resonant network. When the inserted coupling capacitance is sufficiently higher than the resonant capacitance, the reactance component variations of the resonant network can be ignored despite capacitor-plate misalignment. As a result, the CPT system works with the LI operation, keeping the constant output voltage and ZVS at any load resistances and capacitor-plate positions, without any control and wireless communication. As a concrete example, a CPT system composed of an LI Class-EF inverter, an LCL compensation circuit, and a Class-D rectifier is designed along with its analytical expressions. Additionally, the condition of coupling capacitance to maintain the output voltage and ZVS is predicted numerically. By carrying out experiments, we show the effectiveness of the CPT system along with the validity of the analysis and its design procedure. In the experiments, the implemented CPT system achieved 88.5% power delivery efficiency at 135 W output in the rated state.

2. Design Concept

2.1. CPT and Its Problems

CPT systems can be modeled as resonant non-isolated dc-dc converters [1–3,21,23]. Figure 1 shows the CPT system, where it is modeled as the class-E-D dc-dc converter [23]. In CPT systems, the coupling capacitance usually acts as the resonant capacitance. Because it is difficult to achieve a sufficiently high value for the coupling capacitance, the transmission frequency is high, in the megahertz range. Therefore, high-frequency inverters, such as the class-E, class-EF, and class-Φ, are often used for the dc-ac inverter. Figure 1b shows example waveforms of the class-E-D CPT system. When the load resistance and coupling capacitance are at their rated values, the system achieves ZVS and the rated output voltage. However, when there are variations in load resistance or resonant capacitance due to misalignment of the coupling capacitor plates, the CPT system cannot maintain the rated output voltage and ZVS. This is because resonant converters are sensitive to load variations and resonant frequency.

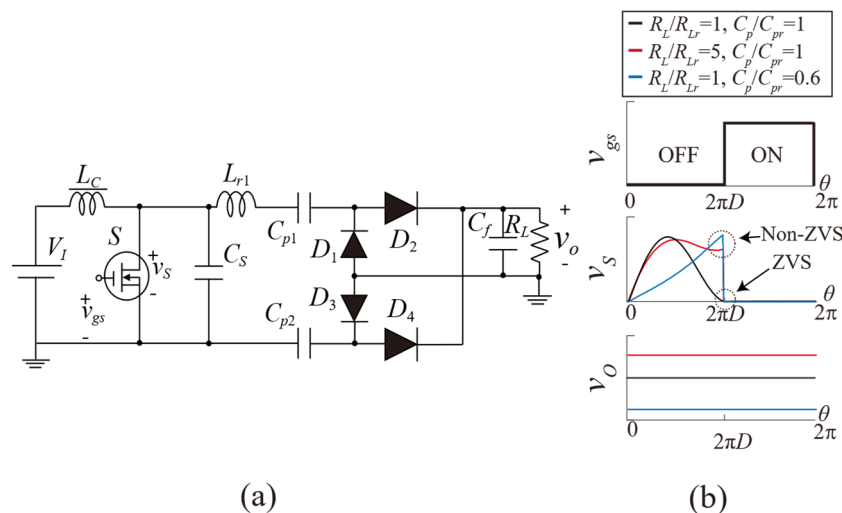


Figure 1. Conventional CPT system using class-E inverter and class-D rectifier. (a) Circuit topology. (b) Waveform example.

2.2. Proposed CPT Design Strategy

Recently, the LI technology of high-frequency inverters has attracted the attention of many researchers and developers [8,13–19]. The LI operation is one of the special modes of operation for a class-E inverter, which can be achieved by setting the proper component values [13,14]. In the LI operation, it is theoretically ensured that the amplitude of the output voltage is constant and the inverter always maintains ZVS against load variations.

Our design strategy is to apply the LI operation to the CPT system. However, in the CPT system, the coupling capacitance varies due to misalignment of the coupling capacitor plates. Because the LI operation is still sensitive to variations in the reactance component of the resonant network, the output voltage variation and non-ZVS occur due to misalignment.

Our second CPT design idea is to make the resonant capacitance much lower than the coupling capacitance. The series connection of the resonant and coupling capacitances makes the system robust to coupling-capacitor plate misalignment.

By combining LI technology and low resonant capacitance, the CPT system, which is robust to both load variations and plate misalignment, can be obtained.

3. Analysis and Design of LI Non-Isolated Resonant Converter

Figure 2 shows a circuit topology of the proposed non-isolated resonant converter for CPT applications, consisting of a class-EF inverter, LCL filter, and class-D rectifier. In the proposed system, L_1 is virtually divided into three inductances, the inverter resonant inductance L_a , the additional inductance L_b for phase shift of the resonant current, and the LCL filter inductance L_2 .

Figure 3 shows example waveforms of the converter for the fixed output resistance. We adopt the LI operation for the class-EF inverter. Therefore, the inverter always maintains ZVS at any load resistance. The LI Class-E and Class- Φ inverters inherently operate in the CV mode [8], whereas the LI Class-EF inverter operates in the CC mode. In isolated LI WPT systems, the output-voltage gain can be adjusted by the transformer turns ratio. However, such gain adjustment is unavailable in the proposed CPT. Therefore, the Class-EF inverter was selected so that its load-independent output current could be converted by the LCL network into a CV output with the required voltage gain.

The class-D rectifier rectifies the AC voltage into the DC output voltage. As a result, the CV DC output can be obtained against load variations without any control.

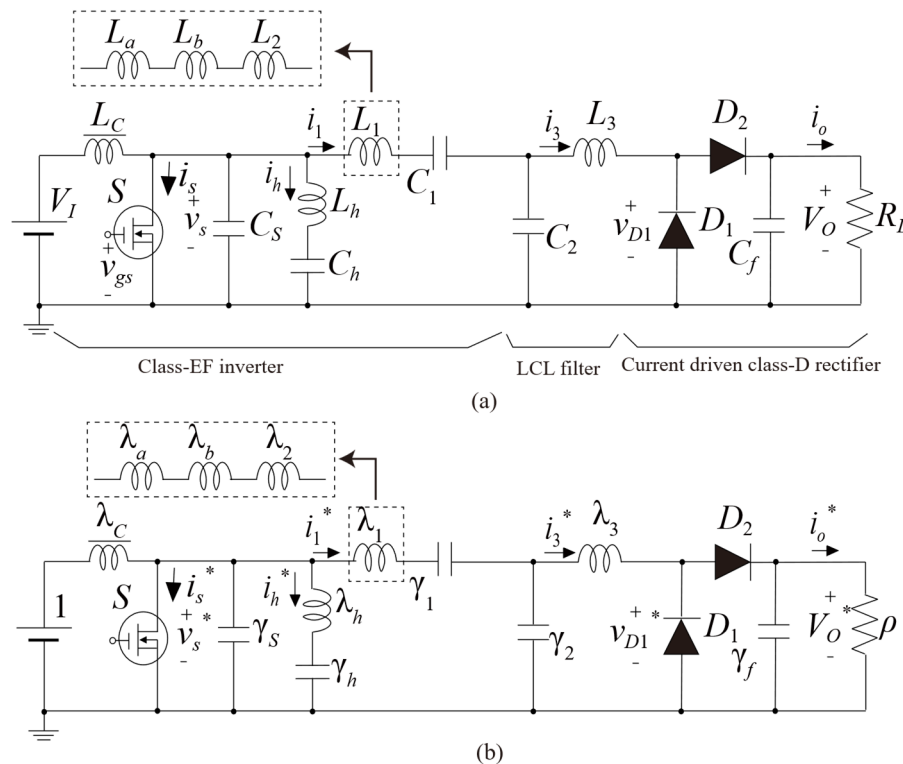


Figure 2. Proposed dc-dc converter. (a) Circuit topology. (b) Normalized circuit model.

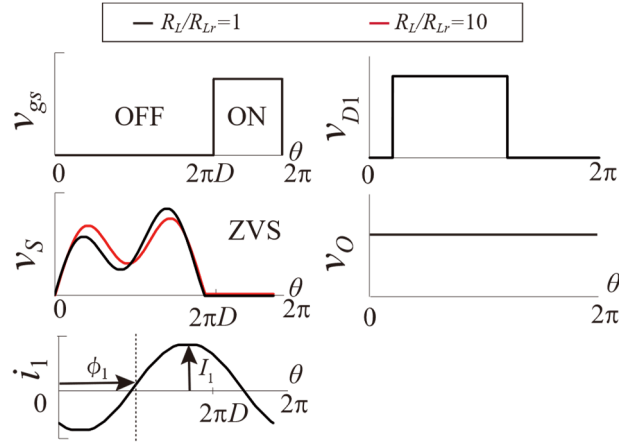


Figure 3. Example waveforms of the proposed dc-dc converter.

This paper presents analytical expressions for the proposed dc-dc converter. For simplicity of the analysis, the following assumptions are given.

- (1) The quality factor of the resonant structure $L_a + L_b - C_1$ is sufficiently large. Thus, the pure sinusoid of the current flowing through $L_a + L_b$ is assumed.
- (2) The diode is ideal, and the parasitic capacitance is negligible.
- (3) The switching operation of the inverter and diode is ideal. The inverter switches at $\theta = 0$ and $2\pi D$. The parasitic output capacitance of the switch S is absorbed into the shunt capacitance C_s .
- (4) All passive components, including the parasitic capacitance of the inverter switch, exhibit linear behavior.

In this analysis, circuit parameters are normalized as

$$\lambda_k = \frac{\omega L_k}{R_{Lr}}, \gamma_k = \omega C_k R_{Lr}, \quad (1)$$

$$\rho_k = \frac{R_k}{R_{Lr}}, x_k = \frac{X_k}{R_{Lr}}, \quad (2)$$

where k is the label of the passive components and the subscript r indicates the rated values. Additionally, all voltages and currents are normalized as

$$v_k^* = \frac{v_k}{V_I}, i_k^* = \frac{R_{Lr} i_k}{V_I}, \omega_k^* = \frac{\omega_k}{\omega}, \quad (3)$$

where $*$ denotes normalization. By applying the above normalizations, all the parameters and variations become dimensionless, which gives the generality of the analytical expressions.

Figure 4 shows the circuit-topology transformations following the converter analysis. Figure 4a shows the normalized circuit model of the proposed converter. The class-D rectifier can be equivalently expressed as a purely resistive component R_D [24], as shown in Figure 4b, namely

$$\rho_D = \frac{2\rho_L}{\pi^2}. \quad (4)$$

The input resistance and reactance of the LCL filter with the rectifier resistance are expressed as

$$\rho_{eq} = \frac{\rho_D}{(H_1 + \gamma_2 H_2)^2 + (\gamma_2 \rho_D)^2}, \quad (5)$$

$$x_{eq} = \frac{j[(H_1 + \gamma_2 H_2)(H_1 \lambda_3 - \lambda_2) + \gamma_2 \rho_D H_1]}{(H_1 + \gamma_2 H_2)^2 + (\gamma_2 \rho_D)^2}, \quad (6)$$

respectively, where

$$H_1 = 1 - \lambda_2 \gamma_2, \quad (7)$$

$$H_2 = \lambda_3 - \lambda_2. \quad (8)$$

For the LI operation of the inverter, it is mandatory that the equivalent reactance is constant against load variations. When the following conditions

$$H_1 = 1 - \lambda_2 \gamma_2 = 0, \quad (9)$$

$$H_2 = \lambda_3 - \lambda_2 = 0, \quad (10)$$

are satisfied, the equivalent resistance and the equivalent reactance are

$$\rho_{eq} = \frac{1}{\gamma_2^2 \rho_D} = \frac{\lambda_2^2}{\rho_D} = \frac{\pi^2 \lambda_2^2}{2\rho_L}, \quad (11)$$

$$x_{eq} = 0. \quad (12)$$

It is seen from (12) that the equivalent reactance is independent of the converter load resistance. Therefore, the proposed converter can be equivalently converted to Figure 4c. We can adopt the LI design for the class-EF inverter.

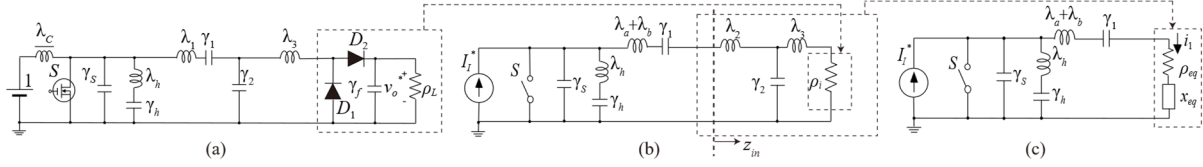


Figure 4. Circuit transformation process of the proposed converter. (a) Normalized circuit model. (b) Equivalent inverter with LCL filter. (c) Equivalent inverter.

Fundamental analysis of the LI class-EF inverter is shown in [15,17]. Because the pure sinusoid of the current flowing through $L_a + L_b$ is assumed, we have

$$i_1^*(\theta) = I_1 \sin(\theta + \phi_1) \quad (13)$$

where I_1 is the amplitude of i_1 , θ is the angular time, and ϕ_1 is the phase difference between the output current and the drive signal, as shown in Figure 3. From [19,21], the voltage across the shunt capacitance C_S can be expressed as

$$v_s^*(\theta) = \begin{cases} -\frac{1}{\gamma_s} \left\{ \frac{A_1 \sin(\sqrt{\alpha+1}\omega_h^* \theta) + B_1 [1 - \cos(\sqrt{\alpha+1}\omega_h^* \theta)]}{\sqrt{\alpha+1}\omega_h^*} - \frac{\theta I_1^*}{\alpha+1} - \frac{(1-\omega_h^{*2})I_1^* [\cos(\theta + \phi_1) - \cos \phi_1]}{1 - (\alpha+1)\omega_h^{*2}} \right\}, & (0 \leq \theta \leq 2\pi D) \\ 0, & (2\pi D \leq \theta \leq 2\pi) \end{cases} \quad (14)$$

where $\omega_h^* = \frac{1}{\sqrt{\lambda_h \gamma_h}}$ is the normalized harmonic resonant frequency, and $\alpha = \frac{\gamma_h}{\gamma_s}$ the ratio of capacitances. In addition, A_1 and B_1 are unknown constants.

When the class-EF inverter satisfies

$$v_s^*(\theta = 2\pi D) = 0, \quad (15)$$

$$\frac{dI_1^*}{d\rho_L} = 0, \quad (16)$$

the inverter works with the LI operation.

In order to satisfy the conditions of Equations (15) and (16), the following conditions

$$\phi_1 = -\pi D, \quad (17)$$

$$\omega_h^* = \frac{1}{2(1-D)}, \quad (18)$$

$$2\alpha \sin\left(\frac{\pi D \sqrt{\alpha+1}}{1-D}\right) + \frac{\pi D \sqrt{\alpha+1}}{1-D} [1 + \cos\left(\frac{\pi D \sqrt{\alpha+1}}{1-D}\right)] = 0, \quad (19)$$

should be satisfied.

With the above LI-operation conditions, the normalized current amplitude of the inverter can be obtained analytically as

$$I_1^* = \left\{ -\frac{2(1-D)^2\alpha}{(\sqrt{\alpha+1})^3} \left[\frac{\sin(\frac{\sqrt{\alpha+1}\pi D}{1-D} + \pi D)}{\sqrt{\alpha+1} + 2(1-D)} - \frac{\sin(\frac{\sqrt{\alpha+1}\pi D}{1-D} - \pi D) - \sin(\pi D)}{\sqrt{\alpha+1} - 2(1-D)} \right] \right. \\ \left. + \frac{\pi(1-D)D}{\alpha+1} \left[-\frac{\cos(\frac{\sqrt{\alpha+1}\pi D}{1-D} + \pi D) - \cos(\pi D)}{\sqrt{\alpha+1} - 2(1-D)} + \frac{\cos(\frac{\sqrt{\alpha+1}\pi D}{1-D} - \pi D) - \cos(\pi D)}{\sqrt{\alpha+1} + 2(1-D)} \right] \right. \\ \left. - \frac{2[\pi D \cos(\pi D) - \sin(\pi D)]}{\alpha+1} \right\}^{-1} \quad \pi\gamma_s = H_{inv}\gamma_s \quad (20)$$

where H_{inv} is a variation about ω_h and α , which are determined uniquely for a given D . Figure 5 shows H_{inv} as a function of the duty ratio.

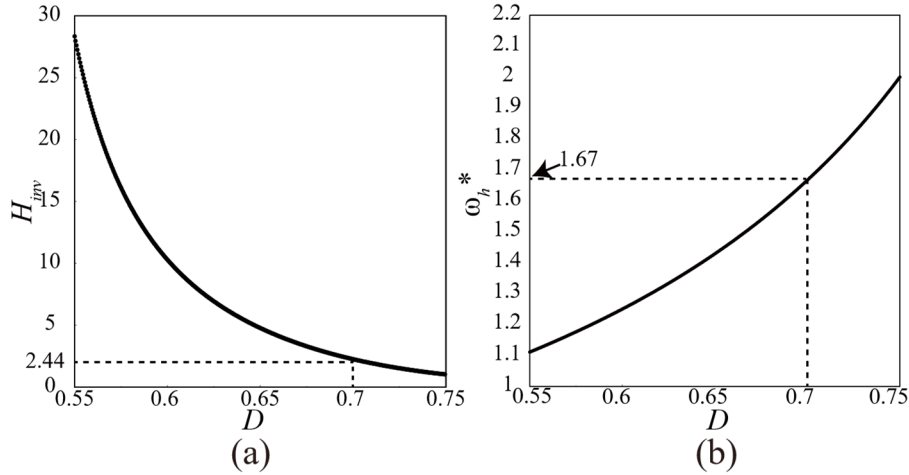


Figure 5. Parameter curves for designing the LI class-EF inverter. (a) H_{inv} as functions of D . (b) ω_h^* as functions of D .

In addition, there are no power losses in the ideal circuit model in Figure 2. Namely,

$$V_I^* I_I^* = \frac{\rho_{eq} I_1^{*2}}{2} = \frac{V_O^2}{\rho_L} \quad (21)$$

where, V_O^* is the normalized output voltage of the proposed converter. From (11), (20), and (21), the normalized output voltage of the proposed converter can be obtained as

$$V_O^* = \frac{1}{2} \pi H_{inv} \lambda_2 \gamma_s, \quad (22)$$

It is seen from (22) that the output voltage is independent of the load resistance. We also see that the output voltage can be adjusted by λ_2 . This is an explanation that the LCL network has a function to adjust the output voltage.

4. CPT System Analysis

In our design concept, the CPT system is constructed by adding a coupling capacitance to the resonant capacitance of the resonant converter in series, as shown in Figure 6. The equivalent capacitance of the resonant and coupling capacitors is expressed as

$$\gamma_{eq} = \frac{1}{\frac{1}{\gamma_1} + \frac{1}{\gamma_{p1}} + \frac{1}{\gamma_{p2}}}. \quad (23)$$

Therefore, the normalized resonant angular frequency including the coupling capacitance is

$$\omega_1^{*'} = \frac{1}{\sqrt{\gamma_{eq} \lambda_a}}. \quad (24)$$

From (23), if the coupling capacitance is sufficiently high compared to the resonant capacitance, the equivalent capacitance is dominated by the resonant capacitance. Equivalent capacitance can be regarded as constant when the variation due to misalignment is absorbed. Consequently, ZVS and CV output are achieved regardless of load variations and plate misalignments by combining the LI operation and high coupling capacitance. The proposed method is applicable only when both coupling capacitances C_{p1} and C_{p2} are sufficiently larger than the resonant capacitance C_1 . If either coupling capacitance is comparable to or smaller than C_1 , the equivalent capacitance becomes strongly dependent on the coupling capacitance, and the LI operation, including ZVS and constant-voltage output, can no longer be guaranteed.

High coupling capacitance can be realized in several environments, such as underwater conditions [9,24,25]. In this paper, we implement a CPT system with high relative permittivity, short plate distances, and large plate areas [20].

Figure 7 shows the upper limit of $Q_{eq} = 1/(\gamma_{eq}\rho_{eq})$ as a function of the normalized resonance frequency ω_1^* for fixed allowable voltage variation obtained from (11), (23), and (24), where the variations of ω_1^* and Q_{eq} express those of the resonant capacitance including the coupling capacitances and inverter output resistance. When the Q_{eq} value is lower than the line, the output voltage variations are suppressed within the target variation range. For example, when the resonant frequency variation due to the coupling capacitance variation is 5%, namely $\omega_1^* = 1.05$, the variation ranges of output voltage are less than 1%, 3%, 5%, and 7% when the Q_{eq} value is less than 20, 24, 31, and 35, respectively.

The small variations in V_o mean that the CPT can nearly achieve the LI operation. It is obviously seen from Figure 7 that low Q_{eq} is preferred for obtaining the robustness against coupling-capacitor plate misalignment. However, lowering the Q value requires increasing C_1 . Conversely, the smaller the value of C_1 , the smaller the variation range of the resonant capacitance. In other words, there is a trade-off relationship about C_1 between low Q_{eq} and the variation range of ω_1^* .

When the coupling capacitance is pre-designed before the CPT design, the variation range of the coupling capacitance can be determined. Additionally, the minimum value of ρ_{eq} can usually be obtained from the design specification. The proper value of C_1 to obtain the LI operation in the specified variation ranges of load resistance and coupling-capacitance plate positions can be determined by referring to this figure.

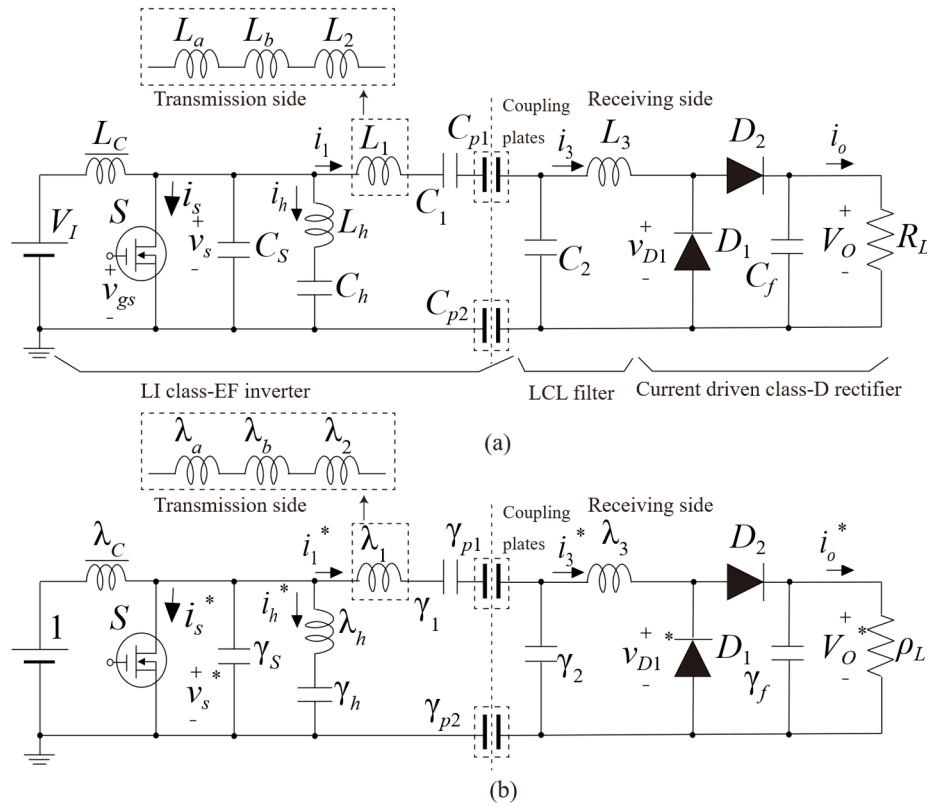


Figure 6. Proposed CPT system. (a) Circuit topology. (b) Normalized circuit model.

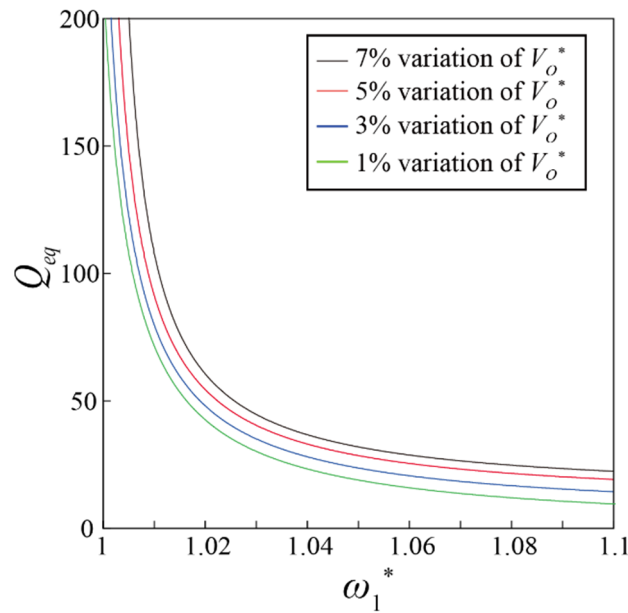


Figure 7. Boundary lines of Q_{eq} at each variation rate of V_o^* with respect to the normalized resonant frequency ω_1^* .

5. Circuit Design and Implementation

5.1. Design Specification

In this design, we give design specifications as input voltage $V_I = 200$ V, operating frequency $f = 6.78$ MHz, rated output voltage $V_{Or} = 45$ V, rated load resistance $R_{Lr} = 15 \Omega$, and rated coupling-capacitor plate distance $d_r = 0$ mm. The switch-off duty ratio is set to $D = 0.7$ to obtain maximum power output capability of the class-EF inverter [20]. The range of variation of load resistance is $R_{Lr} \leq R_L \leq 10R_{Lr}$. In addition, the positional range of the coupling-capacitor plate is $0 \text{ mm} \leq d \leq 150 \text{ mm}$ in the horizontal direction, as shown in the experimental circuit photograph in Figure 8.

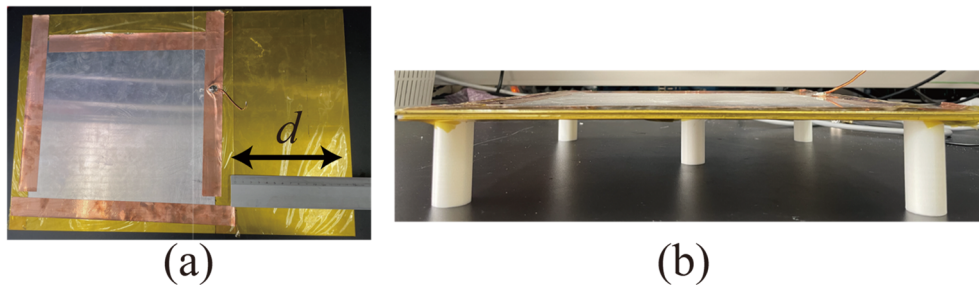


Figure 8. Photo of the coupling capacitor. (a) Photo taken from above (b) Photo taken from the side.

5.2. Coupling Capacitance

As described in Section I, it is necessary to make coupling capacitance as high as possible. A typical coupling capacitance is made by inserting a dielectric between two metal plates. A high capacitance can be obtained by increasing the plate area and permittivity.

In this paper, each coupling capacitance plate size is $300 \text{ mm} \times 300 \text{ mm} \times 2 \text{ mm}$ ($L \times W \times H$). An aluminum plate (A6061) was used as the metal plate. To increase the capacitance value, a Kapton film (3-1966-01 50H) with a relative permittivity of $\epsilon_r = 3.4$ was attached to each plate with copper tape, and the conductors were soldered. Figure 8 shows a photo of the coupling capacitors.

Figure 9 shows the measured capacitance and equivalent series resistance (ESR) of the coupling capacitor as a function of plate distance. The coupling capacitance is 9.5 nF at $d = 0 \text{ mm}$ and 4.5 nF at $d = 150 \text{ mm}$. The capacitance decreases by approximately 43% in the specified range. On the other hand, the ESR increases as the increase in the plate distance. The purpose of this design is to achieve constant output voltage and ZVS in the specified range of coupling capacitance plate misalignment for $0 \text{ mm} \leq d \leq 150 \text{ mm}$.

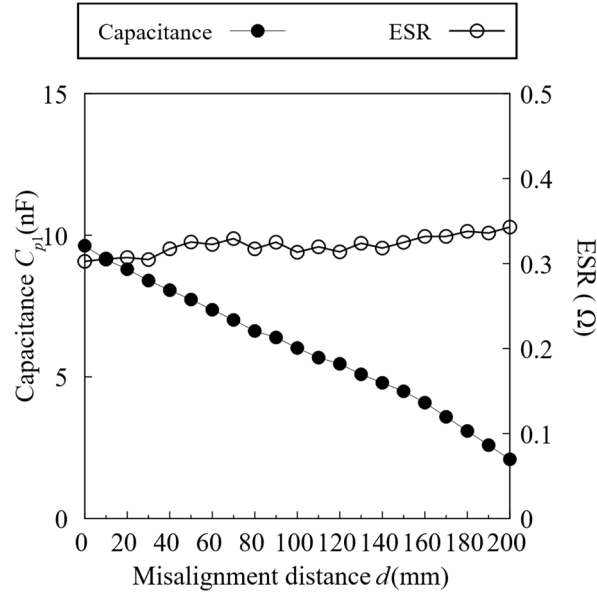


Figure 9. Measured capacitance and ESR of the coupling capacitor as functions of the distance d .

5.3. Component-Value Derivations

From the specifications and [15], the important parameters for the LI class-EF inverter are obtained as

$$\phi_1 = -\pi D \approx -2.20 \text{ rad}, \quad (25)$$

$$\omega_h = \frac{1}{\omega \sqrt{L_h C_h}} = \frac{1}{2(1-D)} \approx 1.67, \quad (26)$$

$$\alpha = \frac{\gamma_h}{\gamma_s} \approx 0.764, \quad (27)$$

$$I_1 \approx 15.3 \times f C_s V_I. \quad (28)$$

From (28) and (22), we have

$$V_O \approx 151 \times f^2 C_s L_2 V_I. \quad (29)$$

The filter capacitance of the class-D rectifier is $C_f = 1 \text{ } \mu\text{F}$, which needs to be high enough so that the output voltage is DC with no ripple [15,17]. From (4), the equivalent resistance R_{Dr} of the class-D rectifier at rated load is obtained as

$$R_{Dr} = \frac{2R_{Lr}}{\pi^2} = \frac{R_{Lr}}{4.93} \approx 3.04 \Omega. \quad (30)$$

From (29) and the specified input and output voltage, we have

$$\gamma_s \lambda_2 = \frac{V_O}{3.82 V_I} \approx 0.059. \quad (31)$$

In this experiment, we used GS66502B as the switch device. According to the datasheet, C_S must be at least 30 pF, and L_2 is limited to a maximum of 1 μH . For this experiment, we selected $L_2 = 225 \text{ nH}$ and $C_S = 144 \text{ pF}$ as the appropriate core size and inductance value.

From (9) and (10) of the LCL filter conditions, L_3 and C_2 are determined as

$$L_2 = L_3 \approx 225 \text{ nH}, \quad (32)$$

$$C_2 = \frac{1}{\omega^2 L_2} = \frac{1}{39.5 \times f^2 L_2} \approx 2.44 \text{ nF}. \quad (33)$$

The equivalent resistance R_{eq} of the LCL filter and class-D rectifier is obtained from (5) as

$$R_{eqr} = \frac{\pi^2 \omega^2 L_2^2}{2R_{Lr}} = \frac{195 \times f^2 L_2^2}{R_{Lr}} \approx 30.3 \Omega. \quad (34)$$

From (27), C_h is obtained as

$$C_h = 0.764 \times C_s \approx 110 \text{ pF}. \quad (35)$$

By substituting (35) into (26), L_h is obtained as

$$L_h = \frac{1}{110 \times f^2 C_h} \approx 1.80 \mu\text{H}. \quad (36)$$

We now have the variation range of C_p from the measurement. Therefore, the variation of C_1 is obtained when the value of C_{eq} is given. Consequently, the variation ratio of ω_{1r}^* is determined as $\omega_{1r}^* = \omega_{1,max}^*/\omega_{1,min}^* = \sqrt{C_{eq,min}/C_{eq,max}}$. We also obtain the variation range of R_{eq} from (11) as $R_{eq,min} = 3.03 \Omega < R_{eq} < R_{eq,max} = 30.3 \Omega$ for the specified variation range of load resistance $15 \Omega < R_L < 150 \Omega$. Therefore, in the given specifications, the maximum value of Q_{eq} is obtained from $Q_{eq,max} = 1/(\omega C_{eq,min} R_{eq,min})$ for given C_1 .

Figure 10 shows $Q_{eq,max}$ as a function of ω_{1r}^* as a black line, where the red line, which is the same as that in Figure 7, is superimposed. As shown in Figure 10, the output voltage variation can be kept within 5% along the black line that lies below the red line. Therefore, we can pick the Q_{eq} value from $Q_{eq} < 65$, and select $Q_{eq} = 50$ in this design. Thus, C_{eq} is calculated as

$$C_{eq} = \frac{1}{\omega Q_{eq} R_{eq}} = \frac{1}{31.4 \times f R_{eqr}} \approx 155 \text{ pF}. \quad (37)$$

Thus, C_1 and L_a are obtained as

$$C_1 = \frac{1}{\frac{1}{C_{eq}} - \frac{1}{C_{p1}} - \frac{1}{C_{p2}}} \approx 160 \text{ pF}, \quad (38)$$

$$L_a = \frac{1}{\omega^2 C_1} \approx 3.56 \mu\text{H}. \quad (39)$$

From the LI conditions in (25)–(27), and (27) of [15], L_b is obtained as

$$L_b = \frac{1}{\pi \omega I_1} \int_0^{2\pi} v_s(\theta) \cos(\theta + \phi) d\theta \approx 1.31 \mu\text{H}, \quad (40)$$

where details of the integral calculation are shown in the Appendix A. Therefore, L_1 is obtained as

$$L_1 = L_a + L_b + L_2 \approx 5.09 \mu\text{H}. \quad (41)$$

Table 1 gives the analytical and measured component values of the proposed CPT. The measured C_s does not include the parasitic capacitance of the switches. Also, L_1 does not include the inductance of the conductors of the coupling capacitors. In the experiment, GS66502B from GaN Systems, STPS5H100 from STMicroelectronics, and UCC27512 from Texas Instruments were used as inverter switch, rectifier diodes, and gate driver, respectively. C0G multilayer ceramic capacitors were used for C_s , C_h , C_1 , C_2 , and C_f . In addition, iron powder cores made of Micrometals Material No. 6 were used for coils L_h , L_1 , and L_3 . Figure 11 shows a photo of the implemented CPT system.

Table 1. Element values of the proposed circuit.

Symbol	Analytical	Measured	Symbol	Analytical	Measured
L_c	-	120 μH (ESR: 113 m Ω)	C_1	160 pF	160 pF
L_h	1.80 μH	1.80 μH (ESR: 156 m Ω)	C_{p1}	-	9.5 nF (ESR: 303 m Ω)
L_1	5.09 μH	4.69 μH (ESR: 450 m Ω)	C_{p2}	-	9.5 nF (ESR: 300 m Ω)
L_3	225 nH	224 nH (ESR: 52 m Ω)	C_2	2.44 nF	2.43 nF
C_s	144 pF	68 pF	C_f	-	1 μF
C_h	110 pF	108 pF			

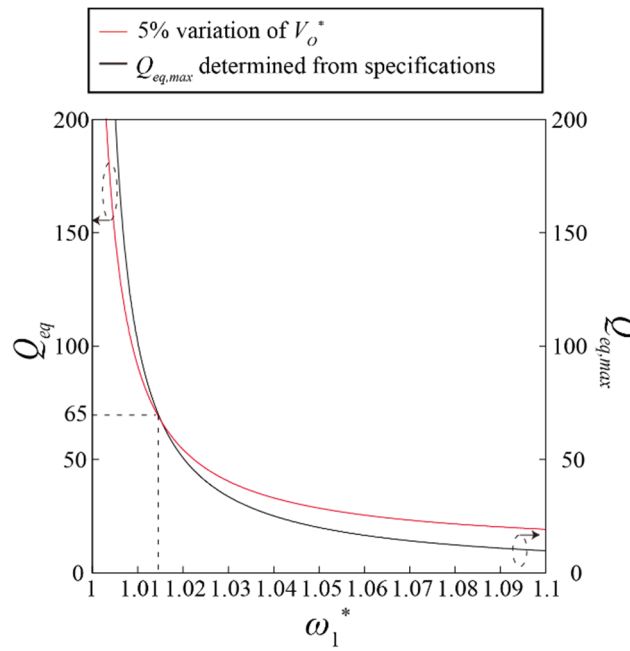


Figure 10. Q_{eq} determined from measurement results and the variation rate of ω_1^* , and the boundary line of Q_{eq} for a 5% variation in V_o^* versus the normalized resonance frequency ω_1^* .

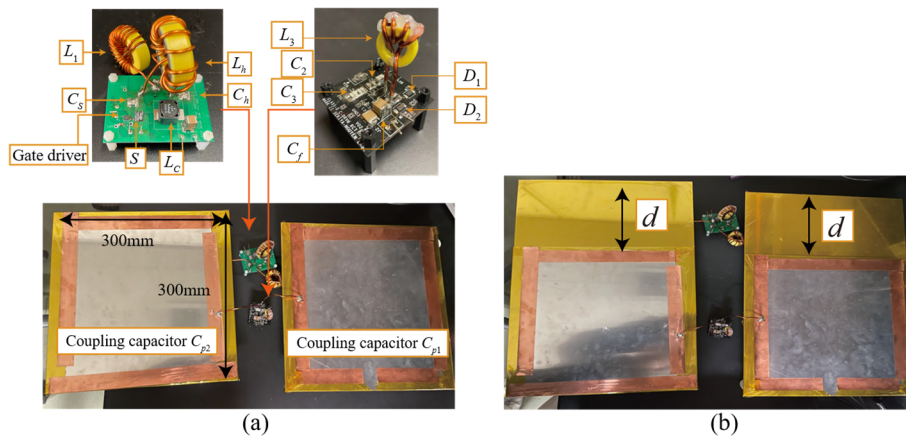


Figure 11. Photo of the implemented circuit. (a) $d = 0$. (b) $d = 150$ mm.

6. Experimental Results

6.1. Waveform Evaluation

Figure 12 shows superimposed analytical and experimental waveforms of the designed CPT system. It is seen from Figure 12a that, at rated load and positional deviation distance of 0 mm, ZVS, the target phase difference for i_1 , and the target output voltage were achieved. It is seen from Figure 12a,b that the inverter of the CPT achieved ZVS regardless of load resistances. Furthermore, the amplitude and phase of the inverter current i_1 were kept constant against load variations. In addition, the output voltage of the CPT was also kept constant. From these results, we can confirm the success in the LI design, which shows the validity of the conditions of LI operation (15), (16) and LCL filter (9) and (10).

Similarly, it is seen from Figure 12a,c that the CPT system achieved ZVS and CV output even with plate misalignment of the coupling capacitor. This is because the coupling capacitance is sufficiently high compared to resonant capacitance. Consequently, the operation is almost the same as the rated condition even though coupling-capacitance plate misalignment occurs.

It is seen from Figure 12a,d that the CPT system also achieved ZVS and CV output against both load variations and coupling capacitor plate misalignment. Because of high coupling capacitance, the CPT worked with the LI operation even though the plate misalignment occurred. As a result, the CPT could achieve ZVS and the output voltage was the same as in the other cases.

It is confirmed from these results that the output voltage of the designed CPT system was always constant. In addition, the inverter switch also satisfied ZVS against load variations and plate misalignments.

Figure 13 shows the experimental waveforms with the sudden load variation. It can be seen from this figure that i_1 and v_o were stable even if the sudden load variation occurred, for both the rated load and the sudden load change of 10 times the rated load. No spike appeared in the experimental waveforms in response to sudden load changes.

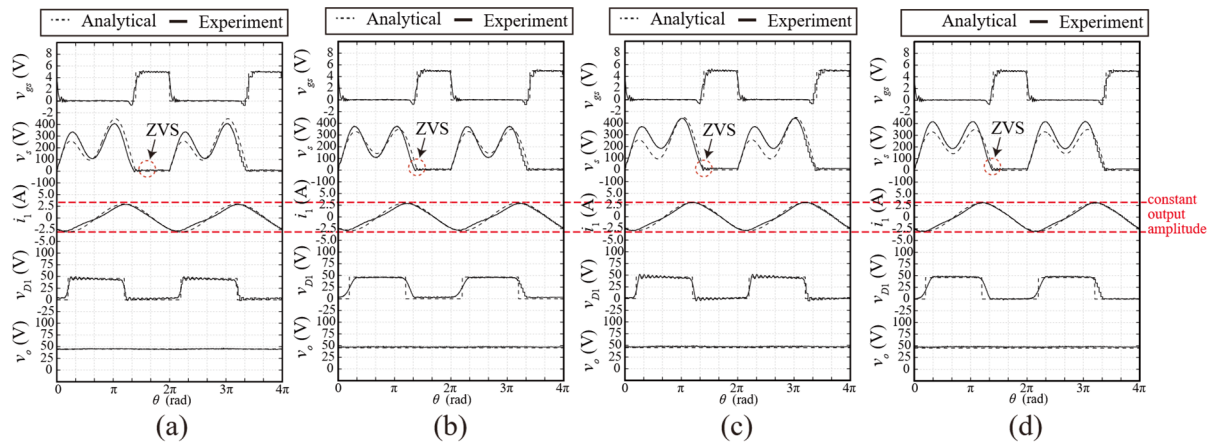


Figure 12. Superimposed analytical and experimental waveforms of the proposed CPT system. (a) $R_L/R_{Lr} = 1, d = 0$. (b) $R_L/R_{Lr} = 10, d = 0$. (c) $R_L/R_{Lr} = 1, d = 150$ mm. (d) $R_L/R_{Lr} = 10, d = 150$ mm.

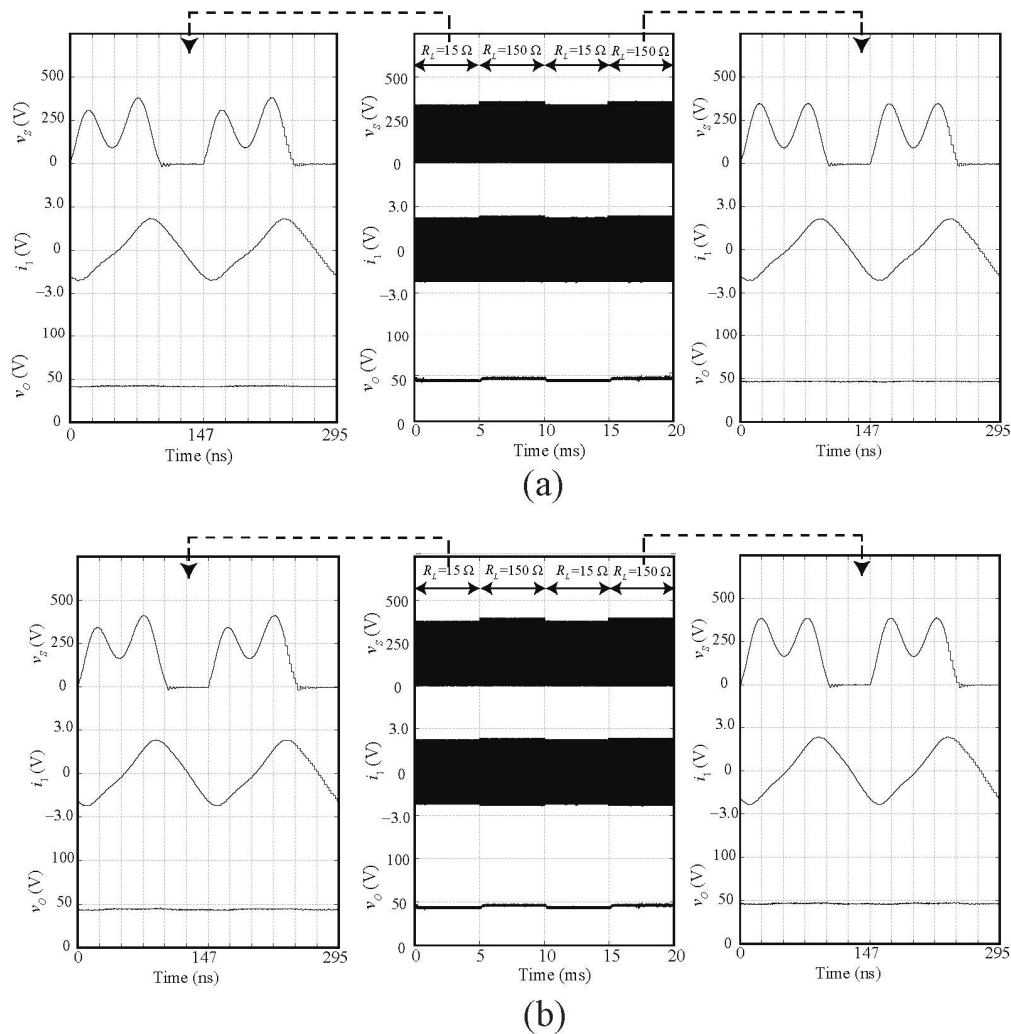


Figure 13. Step response waveforms under sudden load changes. (a) $d = 0$. (b) $d = 150$ mm.

6.2. Performance Evaluations

Figure 14 shows the output voltage as functions of load resistances and coupling capacitance plate-shift distance. We see from Figure 14a that the output voltage was constant, independent of load resistance. In addition, it is seen from Figure 14b that the output voltage remained within the red dotted line even within the specified range $0 \text{ mm} \leq d \leq 150 \text{ mm}$, indicating it is within the 5% variation range as shown in Figure 10.

On the other hand, the output voltage varied as d increased outside the specified range ($d > 150 \text{ mm}$) because the composite capacitance C_{eq} changed, which is similar to Figure 7. When the load resistance R_L is large, the Q-factor of the resonant filter of the inverter increases, making it more sensitive to changes in the resonant frequency. From the results, we can state that the value of C_1 should be determined at the largest resistance case.

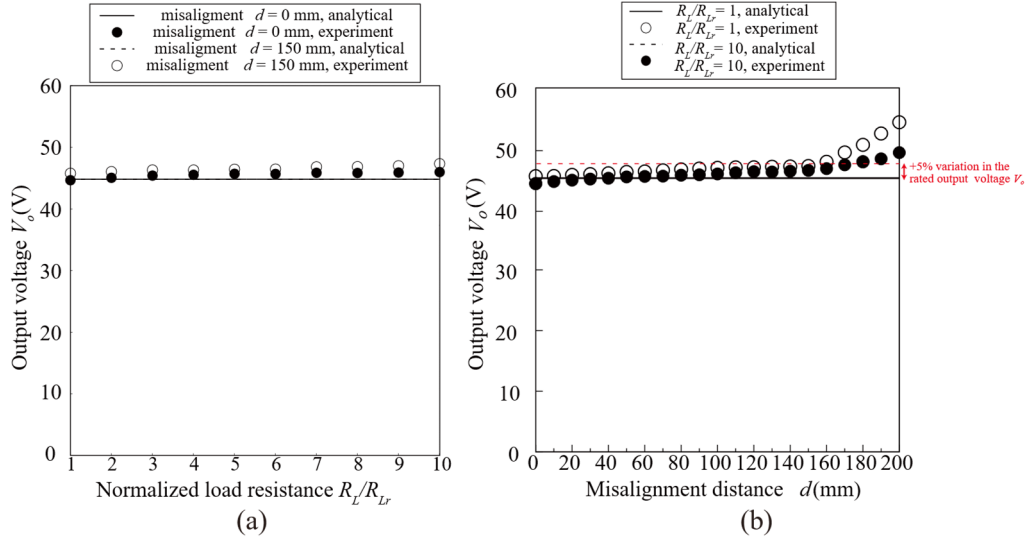


Figure 14. Output voltage of the proposed CPT system. (a) As a function of normalized load resistance R_L/R_{Lr} . (b) As a function of the distance d .

Figure 15 shows the power-delivery efficiency as functions of load resistances and plate-shift distance. In Figure 15a, the power-delivery efficiency decreased as the load resistance increased. This is because the equivalent load resistance R_{eq} of the class-EF inverter becomes smaller as the load resistance increases and the power loss in the ESR of the circuit element becomes dominant. From Figure 15b, it can be seen that the power-delivery efficiency was hardly affected by the misalignment of the coupling capacitor. This is because the coupling-capacitance plate misalignment does not significantly affect the power loss factors. It can be confirmed from Figure 15 that the theoretical efficiency agreed with the measured efficiency quantitatively.

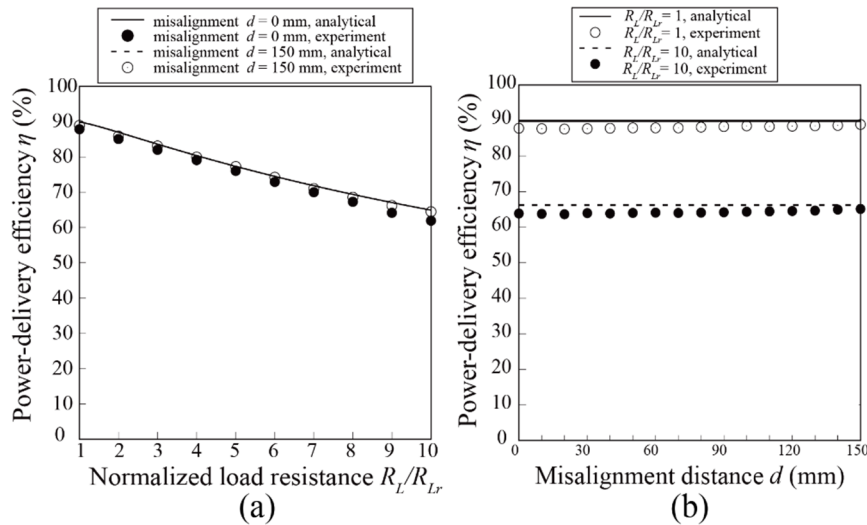


Figure 15. Power-delivery efficiency of the proposed CPT system. (a) As a function of normalized load resistance R_L/R_{Lr} . (b) As a function of the distance d .

Figure 16 shows the analytical power-loss breakdown. The total power loss measured in the experiment is also plotted in Figure 16. The subscript “P” in the boxes in Figure 16 indicates the power loss coefficient. Details of the analytical calculations for each loss are provided in Appendix B. Figure 16 shows that the transmitter-side power losses, particularly P_S , P_{Lh} , P_{L1} , and P_{Cp} remained nearly constant across the entire load range. This is because the LI class-EF inverter exhibits CC operation, causing the conduction losses of each component to remain nearly constant with respect to load variations. This is a reason why the power-delivery efficiency decreased at light loads. Conversely, the receiver side operated in the CV mode because of the LCL filter. Therefore, the power losses decreased as the load resistance increased.

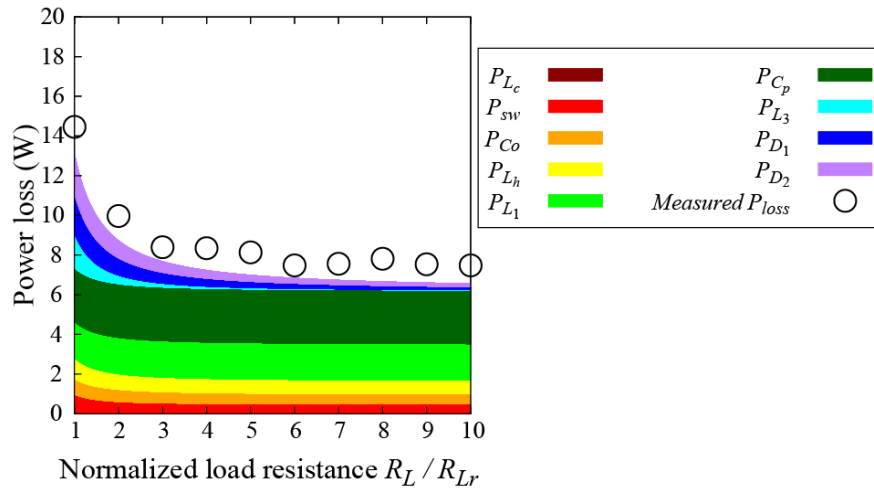


Figure 16. Analytical power-loss breakdown against load variation and measured P_{loss} of the proposed system.

7. Comparisons with Previous Works

There have been CPT systems that are robust against load variations and coupling capacitor misalignment. Table 2 gives a comparison with previous works. Refs. [10,11] achieve robust operation against load variations and misalignment of the coupling capacitor by utilizing wireless communication. On the other hand, ref. [19] achieves CV and ZVS against load variations without wireless communication through a LI design, but it cannot accommodate misalignment. Ref. [20] achieves ZVS against misalignment without wireless communication, but cannot accommodate CV operation or load variations. The proposed CPT system has the advantage of being able to achieve both ZVS and CV against load variations and misalignment of the coupling capacitors. Furthermore, this study does not require wireless communication for control between the transmitter side and the receiver side.

Conventional post-regulation using a linear or switching regulator can provide tighter output-voltage regulation. However, it requires an additional power-conversion stage, resulting in additional power loss and circuit complexity. In contrast, the proposed method achieves CV operation directly in the CPT power stage without an additional post-regulator, which is advantageous for low-to-medium-power applications in terms of efficiency and implementation simplicity.

Table 2. Comparison with previous works.

	[14]	[15]	[23]	[26]	This Work
Transmission frequency	1 MHz	1.2 MHz	1.0 MHz	6.78 MHz	6.78 MHz
Output power	250 W	10 W	5.0 W	277 W	135 W
Coupling capacitance	8.7 pF	22 pF	1.2 nF	8.8 nF	9.5 nF
Wireless communication	Yes	Yes	No	No	No
Output regulation for load variation	Yes	Yes	Yes	No	Yes
Soft switching for load variation	Yes	Yes	Yes	No	Yes
Output regulation for misalignment	Yes	Yes	No	No	Yes
Soft switching for misalignment	Yes	Yes	No	Yes	Yes
Efficiency	85%	76%	81.1%	80%	88.5%

8. Conclusions

This study proposed an analysis and design of a CPT system that is robust to load variations and misalignment of coupling capacitors. The proposed circuit was designed as an LI resonant dc-dc converter to which was added

a coupling capacitor of sufficiently high capacitance. The coupling capacitor is sufficiently high compared to the resonant capacitor, so that the variation of the reactance component of the resonant circuit can be ignored despite the variation of the coupling capacitance. The combination of robustness against coupling capacitance variations and LI design allows the proposed WPT system to always achieve ZVS and maintain constant output voltage without any control and wireless communication. Experimental results showed that the proposed circuit always achieved ZVS and the designed output voltage against load variations and misalignment of the coupling capacitors, demonstrating the effectiveness of the design method.

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Author Contributions

T.K.: conceptualization, methodology, formal analysis, investigation, validation, writing—original draft preparation. A.K.: methodology, formal analysis, writing—review and editing. T.S.: investigation, writing—review and editing. Y.N.: conceptualization, writing—review and editing. H.O.: methodology, writing—review and editing. Y.I.: methodology, writing—review and editing. H.S.: conceptualization, supervision, project administration, writing—review and editing. All authors have read and agreed to the published version of the manuscript.

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The data are not publicly available due to restrictions related to the confidentiality agreement with the collaborating company.

Conflicts of Interest

The authors declare no conflict of interest.

Use of AI and AI-Assisted Technologies

No AI tools were utilized for this paper.

Appendix A. Calculation Details

From [15], the voltage amplitude V_{Lb} across the additional inductance L_b is calculated as

$$\begin{aligned}
 V_{Lb} &= \omega L_b I_1 = \frac{1}{\pi} \int_0^{2\pi} v_s(\theta) \cos(\theta + \phi) d\theta \\
 &= \frac{1}{\pi \omega C_S} \left\langle \frac{A_1}{2\sqrt{\alpha+1}\omega_h} \left\{ \frac{\cos[(\sqrt{\alpha+1}\omega_h+1)2\pi D + \phi] - \cos \phi}{\sqrt{\alpha+1}\omega_h+1} \right. \right. \\
 &\quad \left. \left. + \frac{\cos[(\sqrt{\alpha+1}\omega_h-1)2\pi D - \phi] - \cos \phi}{\sqrt{\alpha+1}\omega_h-1} \right\} \right. \\
 &\quad \left. + \frac{B_1}{2\sqrt{\alpha+1}\omega_h} \left\{ \frac{\sin[(\sqrt{\alpha+1}\omega_h+1)2\pi D + \phi] - \sin \phi}{\sqrt{\alpha+1}\omega_h+1} \right. \right. \\
 &\quad \left. \left. + \frac{\sin[(\sqrt{\alpha+1}\omega_h-1)2\pi D - \phi] - \sin \phi}{\sqrt{\alpha+1}\omega_h-1} - 2[\sin(2\pi D + \phi) - \sin \phi] \right\} \right. \\
 &\quad \left. + \frac{[2\pi D \sin(2\pi D + \phi) + \cos(2\pi D + \phi) - \cos \phi] I_1}{\alpha+1} \right. \\
 &\quad \left. - \frac{(1-\omega_h)I_1}{1-(\alpha+1)\omega_h^2} \left\{ \cos \phi [\sin(2\pi D + \phi) - \sin \phi] - \frac{\sin(4\pi D + 2\phi) - \sin(2\phi)}{4} - \pi D \right\} \right\} \quad (A1)
 \end{aligned}$$

Appendix B. Power Loss Analysis

Figure A1 shows the power-loss model of the proposed CPT system. The power-delivery efficiency is estimated using the analytical waveform. In the power-loss analysis given in this paper, the Equivalent Series Resistance (ESR) of the inductor and coupling capacitor, the on-resistance of the switching device and diodes, and the forward voltage drop across the diodes are considered as power loss factors. These are calculated according to the power loss analysis in [27] as

$$P_{Lc} = r_{Lc} I_1^2, \quad (A2)$$

$$P_{Lh} = \frac{r_{Lh}}{2\pi} \int_0^{2\pi} i_h^2(\theta) d\theta, \quad (A3)$$

$$P_{L1} = \frac{r_{L1}}{2\pi} \int_0^{2\pi} i_1^2(\theta) d\theta = \frac{r_{L1} I_1^2}{2\pi}, \quad (A4)$$

$$P_{Cp} = \frac{r_{Cp}}{2\pi} \int_0^{2\pi} i_1^2(\theta) d\theta = \frac{r_{Cp} I_1^2}{2\pi}, \quad (A5)$$

$$P_{L3} = \frac{r_{L3}}{2\pi} \int_0^{2\pi} i_3^2(\theta) d\theta = \frac{r_{L3} I_3^2}{2\pi}, \quad (A6)$$

where r_{Lc} , r_{Lh} , r_{L1} , r_{Cp} , and r_{L3} are the ESRs of each inductor and coupling capacitor. The on-resistance loss in the switching device S is expressed as

$$P_{SW} = \frac{r_{Son}}{2\pi} \int_0^{2\pi} i_s^2(\theta) d\theta, \quad (A7)$$

where r_{Son} is the on-resistance of the switch. In addition, GaN HEMTs are expected to be used for the switch S to drive the device in the ISM band. Therefore, the hysteresis loss in the output capacitance [28] is included as a loss in the switch and is expressed as

$$P_{Co} = k_o f^{\alpha_o} \left(\frac{V_{Smax}}{V_{DSmax}} \right)^{\beta_o}, \quad (A8)$$

where k_o , α_o , β_o are the coefficients of hysteresis loss, V_{Smax} is the maximum switch voltage, and V_{DSmax} is the rated voltage of the GaN HEMT. The power loss of rectifier diodes is expressed as

$$P_{Dn} = \frac{r_{Dn} I_3^2}{2} + V_{fn} I_3, \quad n = 1 \text{ and } 2, \quad (A9)$$

where r_{Dn} is the on-resistance and V_{fn} is the forward voltage of the diodes. From (A2)–(A9), the sum of all power-loss factors of the proposed system can be expressed as

$$P_{loss} = P_{Lc} + P_{Lh} + P_{L1} + P_{Cp} + P_{L3} + P_{SW} + P_{Co} + P_{D1} + P_{D2}. \quad (A10)$$

Finally, the power-delivery efficiency η is expressed as

$$\eta = \frac{P_o}{P_o + P_{loss}}. \quad (A11)$$

where, the output power P_o of the WPT system is expressed as

$$P_o = \frac{V_o^2}{R_L}. \quad (A12)$$

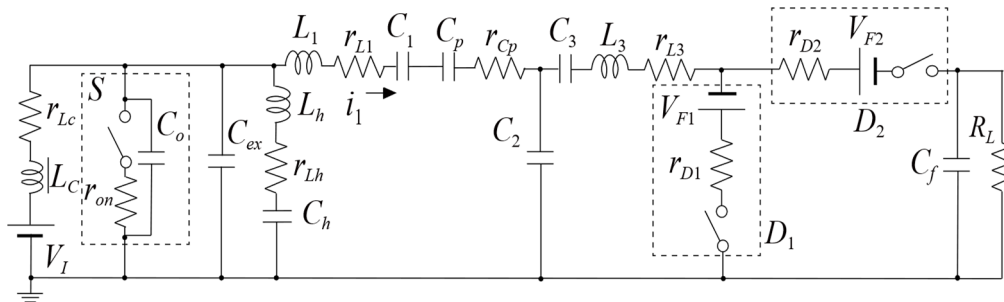


Figure A1. Circuit model of the proposed CPT system for power-loss analysis.

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