

Communication

Design of Four-Plate Parallel Dynamic Capacitive Wireless Power Transfer Coupler for Mobile Robot Wireless-Charging Applications

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Abstract: A detailed theoretical design of an electric resonance-based coupler for dynamic wireless power transfer (DWPT) at the mobile robot level is presented. The scattering matrix of the coupler was derived by transforming and multiplying transmission matrices for each circuit network in a practical equivalent circuit that accounted for loss resistance. This theoretical approach was validated through equivalent circuit models, yielding results consistent with 3D full-wave simulations and showing an error rate of less than 1%. Additionally, a null-power point characteristic, where efficiency sharply decreases when the receiver moves outside the transmitter's range, was observed. The detailed theoretical design of the practical equivalent circuit for electric resonance-based dynamic WPT couplers is expected to contribute to the design of couplers for various specifications in future applications.

Keywords: WPT (wireless power transfer); dynamic WPT; CPT (capacitive wireless power transfer); PEC (practical equivalent circuit); mutual capacitance



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1. Introduction

Wireless power transfer (WPT) technology is considered essential for ensuring a continuous power supply to all IoT devices. From smartphones, wearable devices, and home appliances to mobility units, such as automobiles, IoT devices are increasingly integrating more functions, placing them at the forefront of modern technological innovation. WPT, which eliminates the need for physical connections, offers significant experiential advantages to users. Furthermore, it operates across various environments, including smart homes, smart cities, industrial automation, and autonomous vehicles. In these settings, WPT overcomes limitations related to charging time and location, making it a critical enabler for the intelligent operation of devices through an uninterrupted power supply [1–5].

Particularly, DWPT technology for dynamic wireless charging enables continuous charging even during movement, addressing challenges related to battery capacity and charging infrastructure. As such, it has been actively studied for electric vehicle (EV) applications [6–8]. However, DWPT technology is not limited to EVs. It is a necessary technology that should be developed to provide a continuous power supply across all power-consuming devices, including industrial robots, equipment, smart appliances, and IoT devices.

Since the power requirements vary across applications, appropriately sized WPT couplers must be designed. For instance, EV wireless charging, which typically demands outputs ranging from several kW to tens of kW, requires WPT coupler coils with widths exceeding 0.2 m [9–11]. In contrast, couplers for mobile robots, which operate at an output level of approximately 24–50 W, should be practically smaller than 0.2 m [12,13]. Furthermore, a coreless track-based wireless charging system capable of transmitting and receiving over 100 W has been studied to support the free movement of robots [14].

These technologies are based on inductive wireless power transfer (IPT) using magnetic induction and magnetic resonance. Initially proposed by an MIT research team in 2007, IPT has since been extensively studied [15]. This approach, like the aforementioned studies, achieves high efficiency at close transmitter–receiver distances. However, the strong electromagnetic fields generated by IPT systems cannot be overlooked due to their potential impacts on other electronic devices and human health [16,17]. Additionally, the induced voltage caused by magnetic fields can lead to malfunctions in other devices [18–21]. Given these potential risks, international guidelines regulating electromagnetic exposure limits have been updated [22]. Based on these standards, studies have also been conducted to analyze the electromagnetic fields induced within the human body [23].

In contrast, capacitive wireless power transfer (CPT) using electric resonance is free from these issues [24]. Notably, designing a DWPT system based on IPT requires complex structures [25,26], whereas CPT allows for relatively simpler and more efficient system designs [27,28].

The structure of a four-plate CPT coupler is categorized into parallel and stacked configurations [29]. Generally, the parallel structure features two transmitting metal plates and two receiving metal plates arranged to face each other [30]. In the stacked structure, the two transmitting plates are stacked vertically, and the receiving plates are arranged in the same manner, with the two plates on each side layered on top of each other [31]. These two configurations can be designed and applied according to specific objectives [32,33].

Dynamic wireless capacitive power transfer (DCPT) systems also apply these structures [34–36]. As illustrated in Figure 1, mutual coupling is formed between the transmitter plate and receiving plate, both composed of metal, facilitating power transfer through electric field coupling.

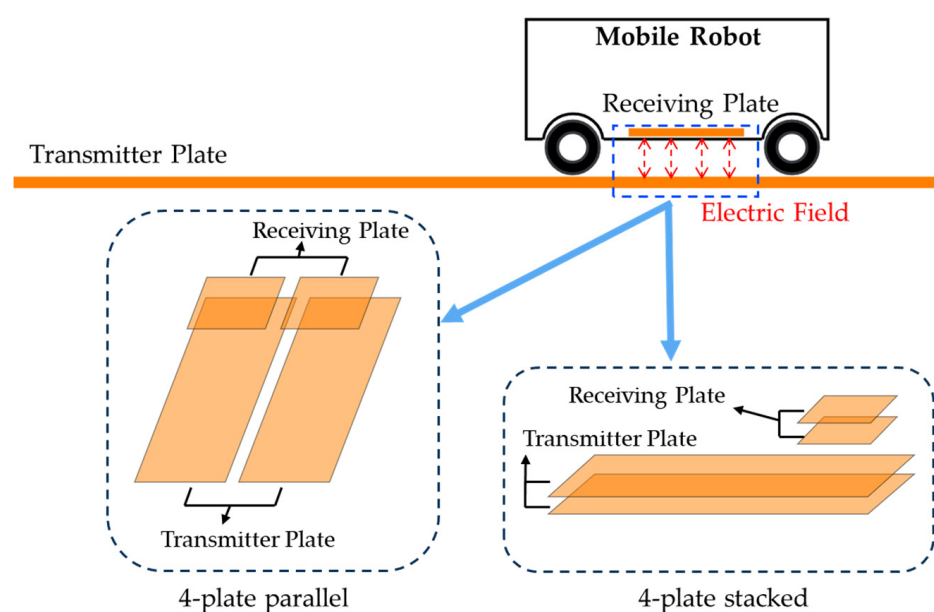


Figure 1. Dynamic wireless-charging systems for mobile robots.

Despite extensive research on WPT applications, the equivalent circuit analysis required to understand the electromagnetic characteristics of couplers remains insufficient. From a safety perspective, WPT systems can experience unnecessary power losses due to structural properties or interference from external objects. To address this, technologies using additional sensors for external object detection have been studied [37–39].

However, these techniques are primarily applied to IPT systems. For CPT systems, a method for detecting external objects or receiver position changes by evaluating the symmetry of the capacitance matrix has been proposed [40]. Ultimately, an equivalent circuit analysis is essential for examining coupling variations caused by changes in the electromagnetic characteristics of couplers. In addition, studies have demonstrated stable power transfer in experiments involving biological entities and uniform performance in multi-transmitter and receiver systems [41,42].

This study provides a theoretical design of a practical equivalent circuit model for a basic DCPT coupler, accounting for loss resistance. The results offer validated findings that can support the design of couplers for various applications. The contributions of this study can be summarized as follows:

1. This study presents high-accuracy validation results through a theoretical design based on the practical equivalent circuit (PEC) model of a four-plate parallel DCPT coupler, compared with electromagnetic simulation results.
2. It provides an analysis of the correlation between S_{21} and mutual capacitance in the four-plate parallel DCPT coupler.

This study aimed to theoretically design a four-plate parallel DCPT coupler using an equivalent circuit approach. For this purpose, the basic equivalent circuit of a CPT coupler was transformed into a practical equivalent circuit that accounted for coupler losses. Each parameter of the equivalent circuit was defined based on the calculation of self-capacitance and mutual capacitance from a model designed using the 3D full-wave simulation software. The transmission matrix was used to calculate the power transfer efficiency of the equivalent circuit, including the impedance-matching network. The theoretical calculations were validated through equivalent circuit calculations and compared with 3D electromagnetic simulation results. Furthermore, the correlation between the theoretically calculated mutual capacitance and power transfer efficiency was analyzed to provide a detailed understanding of the four-plate parallel DCPT structure.

2. Model of Four-Plate Parallel DCPT Coupler

2.1. Theoretical Design Method of Equivalent Circuit

A four-plate CPT system using electric resonance is typically configured as shown in Figure 2. It consists of four metal plates for wireless power transfer, which are electrically connected to the transmitter and receiver circuits on either side. A DC voltage source is converted into an AC voltage at the transmitter, passing through an impedance-matching circuit. The AC voltage source generates electric field coupling between the metal plates, enabling power transfer. To achieve maximum coupling at the desired operating frequency, the impedance-matching circuits on both the transmitter and receiver sides must be accurately designed. This ensures high wireless power transfer efficiency by minimizing the reflected power at the receiver through the alignment of source and load impedances. Therefore, a vector network analysis of the coupler, including impedance matching, is essential. This study aimed to analyze the four-plate parallel DCPT coupler, incorporating an impedance-matching circuit.

The equivalent circuit of the coupler structure, where mutual coupling occurs, is shown in Figure 3, as derived from the structure in Figure 2. In Figure 3a, self-capacitance C_1 is formed between the two metal plates on the transmitter side, while self-capacitance

C_2 is formed between the two metal plates on the receiver side. Since the transmitter and receiver are not electrically connected in the circuit, mutual capacitance C_M is established between them. When this equivalent circuit is transformed into a T-model, it is represented as shown in Figure 3b. In this configuration, the metal plates on the transmitter and receiver sides can be expressed as being connected in the circuit, where C_M appears in parallel with C_1 and C_2 . Consequently, the combined capacitance can be calculated as a parallel sum and represented in a T-type equivalent circuit, as shown in Figure 3c.

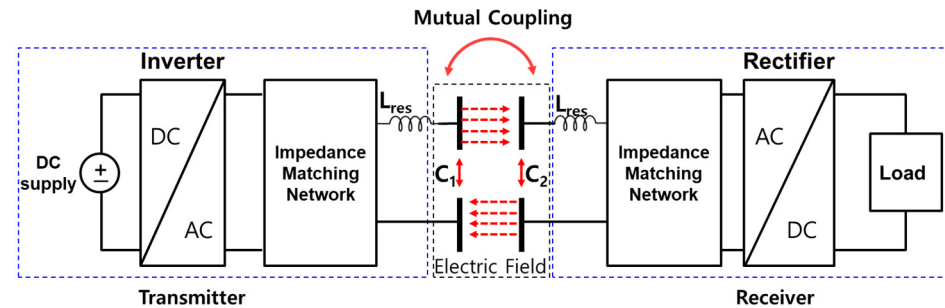


Figure 2. Topology of CPT system.

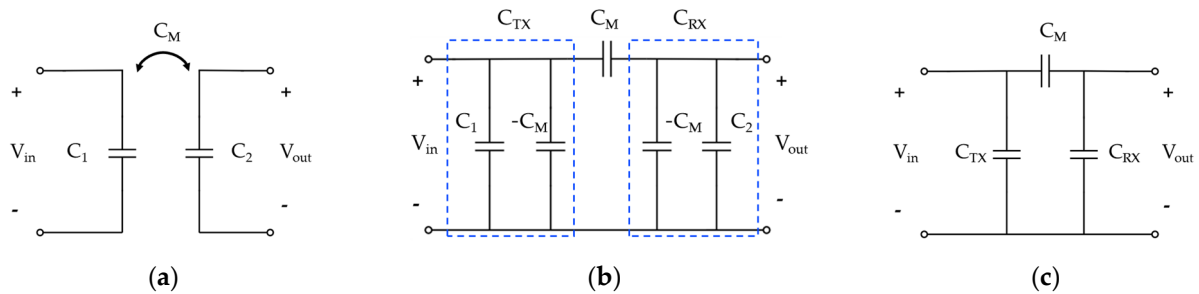


Figure 3. Equivalent circuits of CPT coupler. (a) Basic equivalent circuit; (b) practical equivalent circuit; (c) simplified practical equivalent circuit.

This study provides a theoretical design of the practical equivalent circuit (PEC) by incorporating a lumped inductor for resonance and an impedance-matching circuit. The S-parameter matrix is derived by multiplying the transmission matrices of each circuit network. Among the parameters of the equivalent circuit, the self-capacitances C_1 and C_2 and the mutual capacitance C_M are obtained using the 3D electromagnetic simulation tool Ansys Q3D. The parameters of the impedance-matching circuit are determined based on the impedance of the coupler designed in the 3D full-wave simulation tool ANSYS HFSS.

The transmission coefficient of the coupler is compared between the theoretical calculations of the equivalent circuit model and the results of 3D full-wave analysis using ANSYS HFSS. According to previous studies, equivalent circuit modeling and simulation validation alone can yield results consistent with experimental measurements [43]. Therefore, this study focuses on providing equivalent circuit analysis through simulation-based validation, excluding experimental measurements.

2.2. Simulation Model of Four-Plate Parallel DCPT

Figure 4 illustrates the four-plate parallel DCPT model analyzed in this study, with the specifications detailed in Table 1. The transmitter comprises plates P1 and P2, while the receiver consists of plates P3 and P4. Considering its design on a PCB substrate, the model uses a 0.035 mm thick copper sheet and a 1.52 mm thick dielectric substrate. In Figure 4a, the width L of both the transmitter and receiver substrates is 90 mm, the length TW of the

transmitter substrate is 165 mm, and the length RW of the receiver substrate is 70 mm. The transmitter and receiver are separated by a gap of 10 mm.

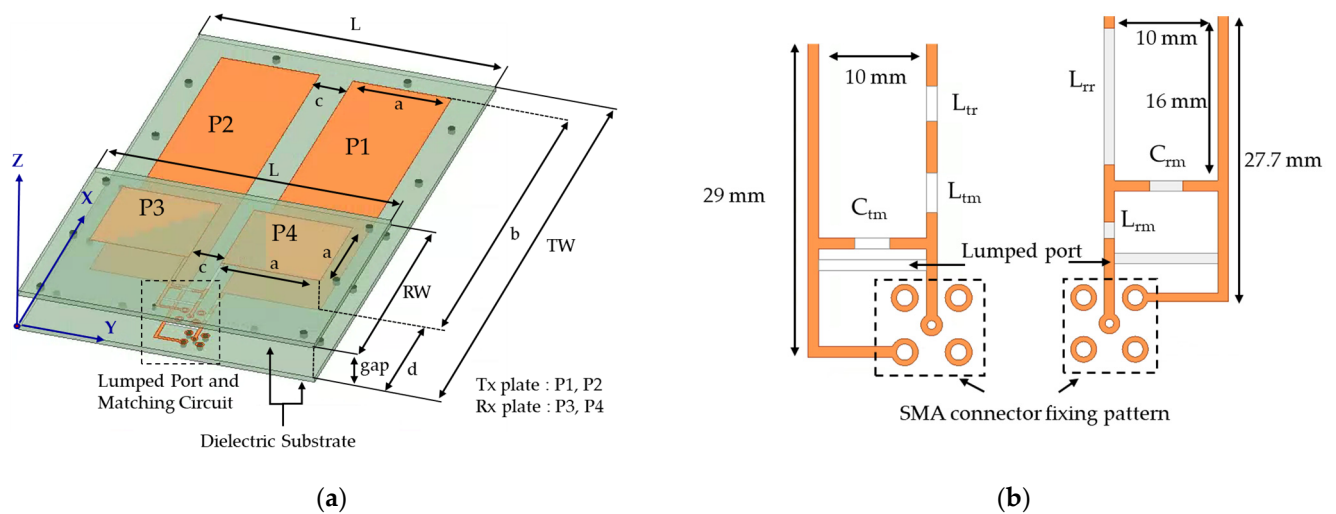


Figure 4. Configuration of 4-plate parallel DCPT coupler. (a) Overall structure; (b) lumped port and matching circuit.

Table 1. Specifications of 4-plate parallel DCPT coupler.

Components	Notation	Value
Width of Tx and Rx substrate	L	90 mm
Length of Tx substrate	TW	165 mm
Length of Rx substrate	RW	75 mm
Width of P1~P4	a	30 mm
Length of P1, P2	b	120 mm
Distance between metal plates of P1 and P2	c	10 mm
X-axis length from the metal plate to the origin	d	35 mm
Distance between Tx and Rx	gap	10 mm

Figure 4b depicts the circuit components, including the ports and matching circuits for the transmitter and receiver shown in (a). The transmission line width is 1 mm, and the positions of passive components reflect their actual sizes. On the left is a top view of the transmitter circuit, while on the right is a bottom-up view of the receiver circuit. The hole at the edge of the substrate is a support-fastening part used to secure the coupler during manufacturing. To accommodate practical design considerations, additional soldering pads for securing SMA connectors are included.

The simulation results of this model are used to determine the self-capacitance and mutual capacitance, which are parameters of the equivalent circuit model. The mutual capacitance and transmission characteristics are observed as the receiver moves along the x-axis. Observations are made from −30 mm to 120 mm, covering the range where the receiver fully moves out of the area corresponding to the transmitter's metal plates. Additionally, to observe the efficiency under misalignment, the transmission characteristics are observed in the range of −50 mm to 50 mm along the y-axis. The component values of the matching circuit, along with the self-capacitance and mutual capacitance calculation results, are described in conjunction with the equivalent circuit model.

2.3. Equivalent Circuit Model of Four-Plate Parallel DCPT

Figure 5 illustrates the equivalent circuit, including the CPT impedance-matching circuit. Figure 5a represents the equivalent circuit that incorporates the parasitic components

of the elements and the inherent loss resistance of the coupler. The parasitic capacitances C_{tmp} , C_{trp} , C_{rmp} , and C_{rrp} are negligible. Similarly, the parasitic resistances R_{tm} , R_{tr} , and R_{rr} are extremely small and are, therefore, combined with the loss resistance of the coupler for calculation. Consequently, the equivalent circuit shown in Figure 5b is used for further calculations.

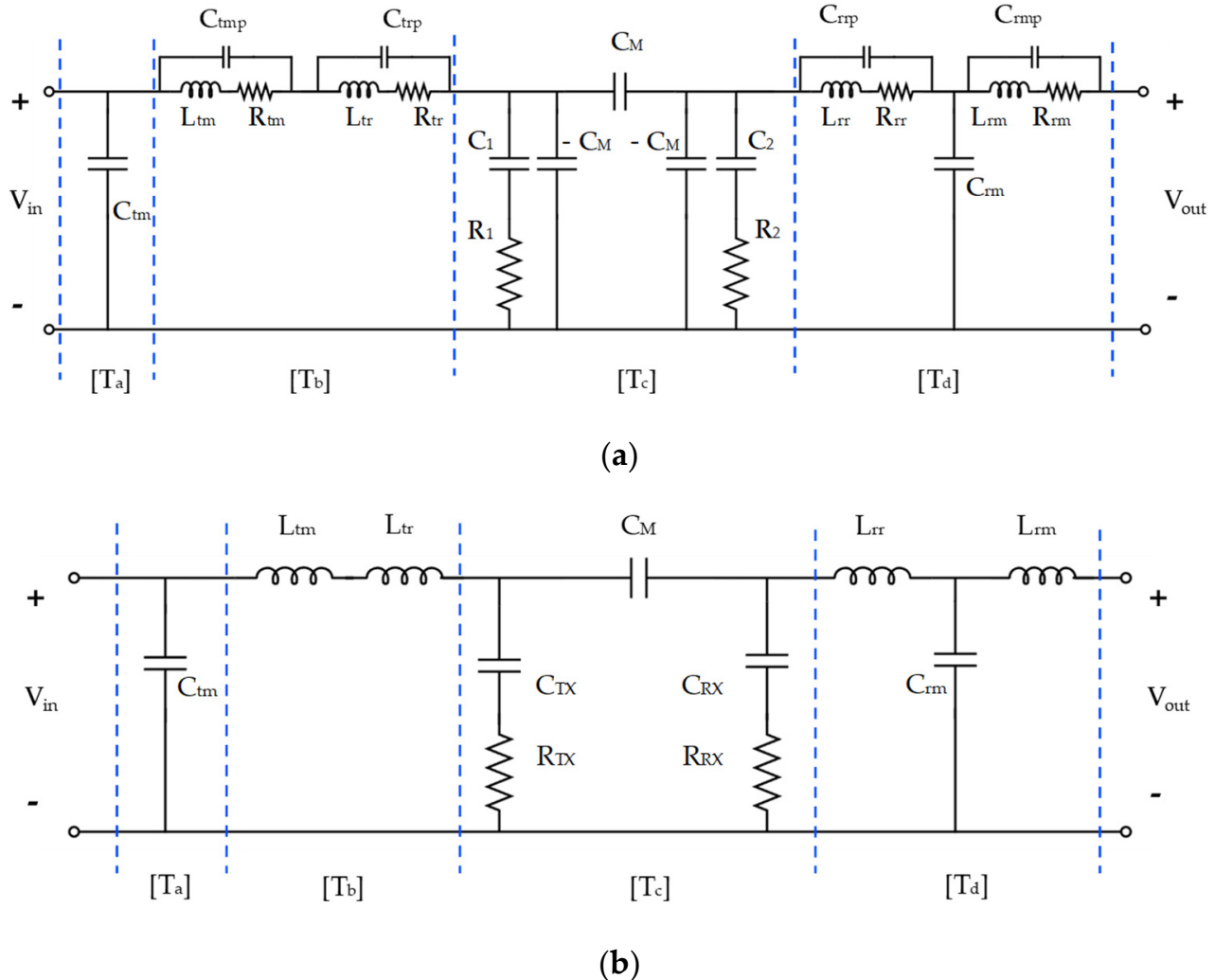


Figure 5. Practical equivalent circuit model of four-plate parallel DCPT coupler. (a) Equivalent circuit indicating parasitic components; (b) simplified practical equivalent circuit (PEC).

The self-capacitances C_{TX} and C_{RX} and the mutual capacitance C_M of the four-plate parallel CPT can be calculated using Equation (1) [44]. Descriptions and values for each parameter in Figure 5b are provided in Table 2. The values of the resonant inductors and matching circuit components are fixed to a single value for use in electromagnetic analysis. The values in Table 2 are the results of matching when the receiver is in its original position, as shown in Figure 4.

When the receiver moves along the x-axis or y-axis, changes in the impedance and coupling degree may cause variations in self-capacitance and mutual capacitance. The self-capacitance of the coupler is calculated based on the impedance obtained from electromagnetic analysis. This is determined using formulas related to the resonant frequency and LC circuit characteristics. The capacitance values are derived by substituting the imaginary part of the Z-parameters in the target frequency range, and these values are applied to the equivalent parameters. Since the obtained values include the effects of mutual capacitance,

a separate process to exclude mutual capacitance, as described in the calculation using Equation (1), is necessary.

Table 2. Equivalent parameters of 4-plate parallel DCPT.

Component	Notation	Value
Coupler self-capacitance of Tx	C_1	3.504 pF
Coupler self-capacitance of Rx	C_2	1.292 pF
Coupler intrinsic resistance	R_{TX}, R_{RX}	1 Ω
Tx's lumped L	L_{tr}	56 μ H
Rx's lumped L	L_{rr}	350 μ H
Tx's impedance-matching L	L_{tm}	4.3 μ H
Rx's impedance-matching L	L_{rm}	650 nH
Tx's impedance-matching C	C_{tm}	100 pF
Rx's impedance-matching C	C_{rm}	750 pF
Mutual capacitance	C_M	0.43 pF

The PEC model considers the loss resistance of the coupler structure. R_{TX} and R_{RX} are determined through an optimization process by comparing the power transfer efficiency of the equivalent circuit model with the electromagnetic simulation. These values are then applied to the circuit analysis for cases where the receiver moves.

$$\begin{cases} C_M = \frac{C_{24}C_{13} - C_{14}C_{23}}{C_{13} + C_{14} + C_{23} + C_{24}} \\ C_{TX} = C_1 - C_M = C_{12} + \frac{(C_{13} + C_{14}) \cdot (C_{23} + C_{24})}{C_{13} + C_{14} + C_{23} + C_{24}} - C_M \\ C_{RX} = C_2 - C_M = C_{34} + \frac{(C_{13} + C_{23}) \cdot (C_{14} + C_{24})}{C_{13} + C_{14} + C_{23} + C_{24}} - C_M \end{cases} \quad (1)$$

L_{tr} and L_{rr} are resonant inductors, and significant differences in their values arise due to the disparity in the self-capacitance of the transmitter and receiver couplers. This basic impedance difference also leads to variations in the configuration of the impedance-matching circuit. For the transmitter, a series inductor is connected first, followed by a parallel capacitor, to achieve impedance matching from the perspective of input impedance. This configuration differs from a typical LC-matching network. However, other forms of impedance-matching circuits may also be employed, and this is not a strict limitation.

The S-parameters of the equivalent circuit shown in Figure 5b can be calculated through the transformation of the transmission matrix. The overall S-parameters of the equivalent circuit can be obtained by multiplying the transmission matrices of each circuit network. The transmission matrix for each circuit network in the equivalent circuit is expressed as shown in Equation (2). When expressed in impedance form, it is represented as Equation (3).

$$\begin{cases} [T]_a = \begin{bmatrix} 1 & 0 \\ Y & 1 \end{bmatrix} \\ [T]_b = \begin{bmatrix} 1 & Z \\ 0 & 1 \end{bmatrix} \\ [T]_c = \begin{bmatrix} 1 + \frac{Y_2}{Y_3} & \frac{1}{Y_3} \\ Y_1 + Y_3 + \frac{Y_1 Y_2}{Y_3} & 1 + \frac{Y_1}{Y_3} \end{bmatrix} \\ [T]_d = \begin{bmatrix} 1 + \frac{Z_1}{Z_3} & Z_1 + Z_2 + \frac{Z_1 Z_2}{Z_3} \\ \frac{1}{Z_3} & 1 + \frac{Z_2}{Z_3} \end{bmatrix} \end{cases} \quad (2)$$

$$\left\{ \begin{array}{l} [T]_a = \begin{bmatrix} 1 & 0 \\ j\omega C_{tm} & 1 \end{bmatrix} \\ [T]_b = \begin{bmatrix} 1 & j\omega(L_{tm} + L_{tr}) \\ 0 & 1 \end{bmatrix} \\ [T]_c = \begin{bmatrix} -\frac{C_2}{C_M} & -\frac{1}{j\omega C_M} \\ j\omega \left(3C_M + 2(C_1 + C_2) + \frac{C_1 C_2}{C_M} \right) + R_{TX} + R_{RX} & 2 + \frac{C_1}{C_M} \end{bmatrix} \\ [T]_d = \begin{bmatrix} 1 - \omega^2 L_{rr} C_{RX} & j\omega(L_{rr} + L_{rm} - \omega^2(C_{RX} L_{rr} L_{rm})) \\ j\omega C_{RX} & 1 - \omega^2 C_{RX} L_{rm} \end{bmatrix} \end{array} \right. \quad (3)$$

According to the formula for converting a transmission matrix into a scattering matrix, the equations can be simplified using the following relationships. By representing the elements of the 2×2 transmission matrix as t and applying the substitution in Equation (4) to simplify each element of the converted scattering matrix, it can be expressed as shown in Equation (5). To represent the scattering matrix for each circuit network, as described in Equation (3), the parameters required to express it in the form of Equation (5) are calculated and presented in Table 3. Consequently, as shown in Equation (6), the scattering matrix of the entire equivalent circuit is calculated by multiplying the scattering matrices of each circuit network.

$$t_n = t_{11} - t_{22}, \quad t_m = t_{11} + t_{22}, \quad t_i = \frac{t_{12}}{Z_0} - t_{21}Z_0, \quad t_j = \frac{t_{12}}{Z_0} + t_{21}Z_0 \quad (4)$$

$$[S] = \begin{bmatrix} \frac{t_n + t_i}{t_m + t_j} & \frac{2(t_{11}t_{22} - t_{12}t_{21})}{t_m + t_j} \\ \frac{2}{t_m + t_j} & \frac{t_i - t_n}{t_m + t_j} \end{bmatrix} \quad (5)$$

Table 3. Parameters of the S-matrix of each detailed circuit network.

Matrix	Parameter
$[S]_a$	$\begin{cases} t_n = 0 \\ t_m = 2 \\ t_i = -j\omega C_{tm}Z_0 \\ t_j = j\omega C_{tm}Z_0 \end{cases}$
$[S]_b$	$\begin{cases} t_n = 0 \\ t_m = 2 \\ t_i = \frac{j\omega(L_{tm} + L_{rm})}{Z_0} \\ t_j = \frac{j\omega(L_{tm} + L_{rm})}{Z_0} \end{cases}$
$[S]_c$	$\begin{cases} t_n = -\frac{C_1 + C_2}{C_M} - 2 \\ t_m = \frac{C_1 - C_2}{C_M} + 2 \\ t_i = -\frac{1}{j\omega C_M Z_0} - Z_0 \left(j\omega C_M \left(\left(\frac{C_1}{C_M} + 2 \right) \left(\frac{C_2}{C_M} + 2 \right) - 1 \right) + R_{TX} + R_{RX} \right) \\ t_j = -\frac{1}{j\omega C_M Z_0} + Z_0 \left(j\omega C_M \left(\left(\frac{C_1}{C_M} + 2 \right) \left(\frac{C_2}{C_M} + 2 \right) - 1 \right) + R_{TX} + R_{RX} \right) \end{cases}$
$[S]_d$	$\begin{cases} t_n = \omega^2 C_{RX} (L_{rm} - L_{rr}) \\ t_m = 2 - \omega^2 C_{RX} (L_{rm} + L_{rr}) \\ t_i = \frac{j\omega(L_{rm} + L_{rr} - \omega^2 L_{rr} L_{rm} C_{RX})}{Z_0} - j\omega C_{RX} Z_0 \\ t_j = \frac{j\omega(L_{rm} + L_{rr} - \omega^2 L_{rr} L_{rm} C_{RX})}{Z_0} + j\omega C_{RX} Z_0 \end{cases}$

$$[S]_{TOTAL} = [S]_a \cdot [S]_b \cdot [S]_c \cdot [S]_d \quad (6)$$

3. Results and Discussion

Figure 6 compares the S21 graphs of PEC and HFSS when the Rx movement distance is 0 mm. The comparison is divided into before and after impedance matching. The S21 before impedance matching was close to 0, and the PEC results were the same. After impedance matching, the resonant frequency and the maximum value of S21 for both PEC and electromagnetic simulation were 6.35 MHz and 0.86, respectively, showing identical results. As previously described, the C_{TX} and C_{RX} of the equivalent circuit were accurate values calculated through simulation, and C_M was adjusted through the optimization process of the equivalent circuit model. Additionally, the coupler loss resistances R_{TX} and R_{RX} were also tuned to obtain accurate graphs.

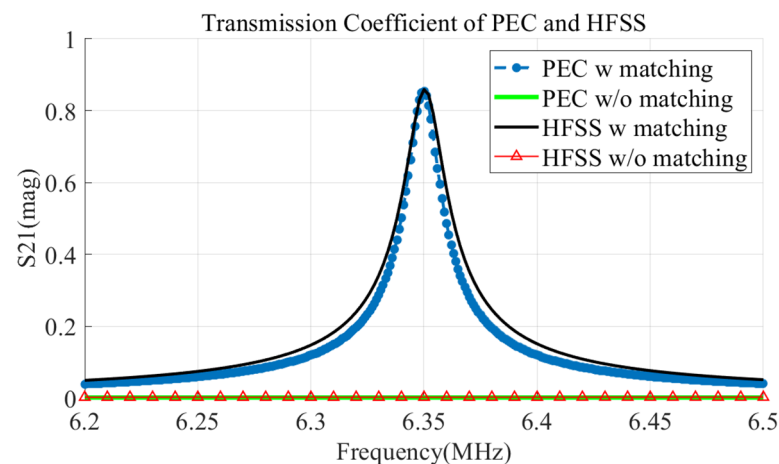


Figure 6. Comparison of transmission coefficient by impedance matching.

Figure 7 shows the maximum S21 and the magnetic and mutual capacitances as the receiving unit moves along the x-axis from -30 mm to 120 mm. The magnetic and mutual capacitances tended to remain constant when the receiving unit was within the metal plate area of the transmitting unit, between 0 mm and 90 mm. Outside this area, as inferred from Equation (1), a decrease in mutual capacitance corresponded to an increase in magnetic capacitance. Additionally, when compared with the maximum S21, the mutual capacitance showed a similar trend within the metal plate area of the transmitting unit. However, the maximum S21 sharply decreased in the regions beyond the metal plate area of the transmitting unit, between -30 mm to 0 mm and 90 mm to 120 mm.

Figure 8 shows the results of observing the maximum value of S21 and mutual capacitance as the Rx moves between -50 mm and 50 mm along the y-axis. As the mutual capacitance approached zero, S21 decreased with a similar slope. The point where S21 approached zero can be distinguished as the null-power point. The locations where null-power points occurred were at -35 mm and 32.5 mm, and it was observed that the value of S21 recovered thereafter. Table 4 summarizes the maximum values of S21 and capacitance at key positions of the Rx along the y-axis. As noted in previous studies, null-power points tend to occur when the mutual capacitance is close to zero. However, the results presented in this study indicate that the value of S21, which had decreased in regions where the mutual capacitance was less than 0.1 , also showed signs of recovery.

As discussed in a previous study [29], this phenomenon occurs when the mutual capacitance C_M formed between the transmitter and receiver converges to zero. The C_M is determined by the interaction between the main coupling capacitances C_{23} and C_{14} and the cross-coupling capacitances. This relationship can be expressed as shown in Equation (7). The cross-coupling capacitances correspond to C_{13} and C_{24} in Equation (1).

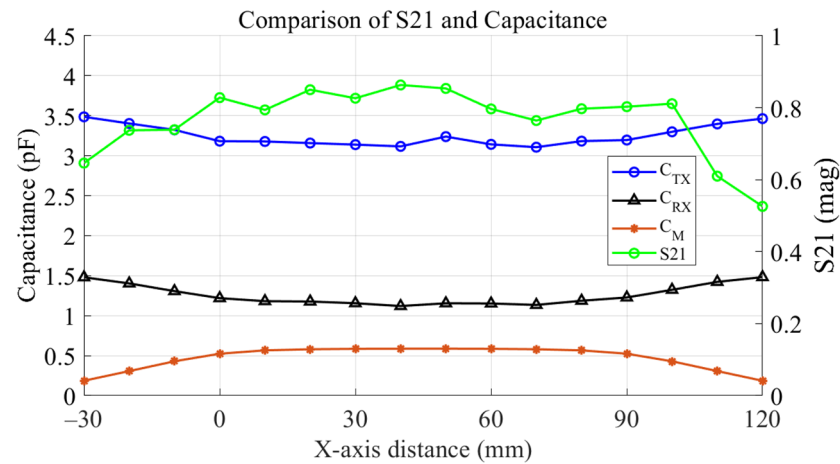


Figure 7. S21 and capacitance by x-axis distance sweep.

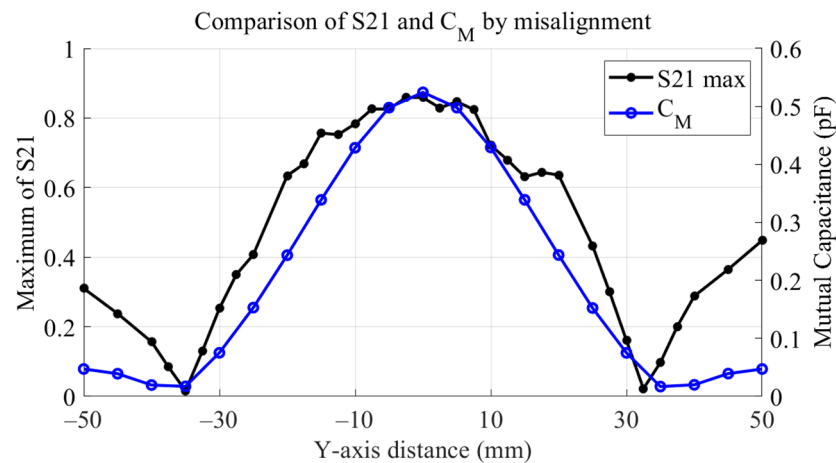


Figure 8. S21 and mutual capacitance by Y-axis distance sweep.

Table 4. Summary of S21 peak and capacitance by y-axis distance sweep.

	Point1 (−50 mm)	Point2 (−35 mm)	Point3 (0 mm)	Point4 (32.5 mm)	Point5 (50 mm)
S21_max	0.31	0.01	0.86	0.02	0.45
C_M (pF)	0.04	0.08	0.52	0.07	0.04
C_{TX} (pF)	3.79	3.89	3.5	3.9	3.86
C_{RX} (pF)	1.7	1.74	1.38	1.75	1.72

In other words, when the product of the main coupling capacitances C_{mc} and the product of the cross-coupling capacitances C_{cc} become equal, the mutual capacitance C_M converges to zero. Based on this principle, it can also be observed that the null-power points in Figure 8 occurred when C_M approached zero.

$$C_{24} \cdot C_{13} = C_{14} \cdot C_{23} \leftrightarrow C_{mc} = C_{cc} \quad (7)$$

4. Conclusions

A theoretical design of a four-plate parallel DCPT coupler applicable to mobile robot-level wireless charging is provided. This design includes the PEC design, incorporating the coupler's loss resistance, and the derivation process of the scattering matrix for each network. The terms of the scattering matrix for each network were expressed in impedance

form, allowing the overall scattering matrix of the proposed PEC to be defined through equation simplification. Valid mutual capacitance values were derived from the scattering matrix calculations of the PEC. This was verified by comparing with the results of 3D full-wave simulations. When the receiver's displacement was 0 mm, the transmission coefficient results from the PEC matched those from the 3D full-wave simulation.

Additionally, changes in electromagnetic characteristics due to the misalignment of the transmitter and receiver of the four-plate parallel DCPT coupler were observed. The S21 and mutual capacitance were compared as the receiver moved between −50 mm and 50 mm along the y-axis. Null-power regions, where the power transmission efficiency sharply decreased, occurred at two points: −35 mm and 32.5 mm. The change in S21 showed a similar trend to the mutual capacitance. The parallel structure of the CPT coupler presented in this study has high mutual capacitance and strong coupling when the transmitter and receiver are aligned, but it is vulnerable to null-power regions.

A theoretical design method for the relatively under-researched DCPT coupler is provided. In the future, theoretical analysis of equivalent circuit models could be an approach to analyzing external factors that induce changes in the electromagnetic characteristics of wireless power transfer couplers. Further research is needed to closely analyze and improve structural alternatives for null-power regions occurring in the four-plate parallel DCPT structure.

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