Optimization of 8-Plate Multi-Resonant Coupling Structure Using Class-E^2 Based Capacitive-Wireless Power Transfer System

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OPTIMIZATION OF 8-PLATE MULTI-RESONANT COUPLING STRUCTURE
USING CLASS-E\(^2\) BASED CAPACITIVE-WIRELESS POWER TRANSFER SYSTEM

by

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B. Tech. May 2014, Malla Reddy Institute of Technology & Science
M.S. May 2017, Old Dominion University

A Dissertation Submitted to the Faculty of Old Dominion University in Partial Fulfillment of the Requirements for the Degree of

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ELECTRICAL & COMPUTER ENGINEERING

OLD DOMINION UNIVERSITY
May 2022

Approved by:

Shirshak Dhali (Director)
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Shu Xiao (Member)
Vishnukumar Lakdawala (Member)
ABSTRACT

OPTIMIZATION OF 8-PLATE MULTI-RESONANT COUPLING STRUCTURE USING CLASS-E\textsuperscript{2} BASED CAPACITIVE-WIRELESS POWER TRANSFER SYSTEM

Yashwanth C Bezawada
Old Dominion University, 2022
Director: Dr. Shirshak Dhali

Capacitive-wireless power transfer (CPT) effectively charges battery-powered devices without a physical contact. It is an alternative to inductive-wireless power transfer (IPT) which is available in the present market. Compared with IPT, CPT offers flexibility in designing the coupling section. Because of its flexibility, CPT utilizes various coupling methods to enhance the coupling capacitance. Misalignment is a common issue in any WPT system. Among IPT and CPT, IPT has better performance for misalignments, but it requires bulk and expensive ferrite core to attain a high coupling coefficient. This work focuses on designing a CPT system to minimize the impact of misalignments. In this research, a novel 8-plate multi-resonant Class-E\textsuperscript{2} CPT system is developed to improve the performance of the CPT system for misalignments. The proposed CPT model expands the resonant frequency band, which results in better performance for misalignments compared with the regular 4-plate CPT system. The 8-plate coupling structure is designed to charge a 100 Ah drone battery. For this application, the coupling is formed when the drone lands on the capacitive- wireless charging pad. This work also presents the analysis of several dielectric materials with different dielectric constants. A well-designed capacitive coupler can effectively limit harmonics during the interaction between transmitter and receiver. Also, the effect of coupling plate shape is identified on the CPT system. The hardware tests indicate the round-shaped plates have better stability in coupling capacitance with the variation in frequency. The effect of misalignments is studied through the impedance tracking of the Class-E\textsuperscript{2} power converter.
Impedance plots for 50 μH, and 100 μH resonant inductors are used to determine input current peak for each case. Additionally, hardware tests are performed to study the variation of input current and output voltage for a range of frequencies. The test results indicate the efficiency at optimal impedance point for a resonant inductor with 50 μH is 8% higher compared to the CPT with a 100 μH resonant inductor which highlights the effects of the resonant inductor on efficiency. The zero-voltage-switching (ZVS) limits are also identified for varying frequencies and duty cycles. Later in this research, the optimal design of the Class-E rectifier is identified to enhance the power transfer. Several cases were considered to investigate the impact of the secondary inductor on the output voltage and the ZVS property. Hardware tests validate that under optimal conditions the efficiency of the Class-E\(^2\) based CPT system improves by 18% compared with \(A_r \gg 1\). Further work presents the advantages of 8-plate multi-resonant coupling for misalignments. The proposed model has a simple design procedure which enhances the power flow from the inverter to the rectifier section. The hardware results of the proposed 8-plate multi-resonant coupling show an increase in efficiency to 88.5% for the 20.8 W test, which is 18% higher than regular 4-plate coupling. Because of the wider resonant frequency band \([455\text{-}485 \text{ kHz}]\), compared with regular 4-plate coupling, the proposed design minimized the output voltage drop by 15% for 10% misalignment. Even for large misalignments, 8-plate improves the CPT performance by 40% compared with 4-plate coupling.
This thesis is dedicated to my parents, brother, uncle, and friends.
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### NOMENCLATURE

<table>
<thead>
<tr>
<th>Acronym</th>
<th>Description</th>
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<tbody>
<tr>
<td>WPT</td>
<td>Wireless Power Transfer</td>
</tr>
<tr>
<td>IPT</td>
<td>Inductive-Wireless Power Transfer</td>
</tr>
<tr>
<td>CPT</td>
<td>Capacitive-Wireless Power Transfer</td>
</tr>
<tr>
<td>EV</td>
<td>Electric Vehicle</td>
</tr>
<tr>
<td>AC</td>
<td>Alternating Current</td>
</tr>
<tr>
<td>DC</td>
<td>Direct Current</td>
</tr>
<tr>
<td>EMI</td>
<td>Electromagnetic Interference</td>
</tr>
<tr>
<td>PWM</td>
<td>Pulse Width Modulation</td>
</tr>
<tr>
<td>SLLD</td>
<td>Switch-Inductor-Inductor-Diode</td>
</tr>
<tr>
<td>CCM</td>
<td>Continuous Conduction Mode</td>
</tr>
<tr>
<td>ZVS</td>
<td>Zero-Voltage-Switching</td>
</tr>
<tr>
<td>MOSFET</td>
<td>Metal-Oxide-Semiconductor Field-effect Transistor</td>
</tr>
<tr>
<td>UAV</td>
<td>Unmanned Ariel Vehicles</td>
</tr>
<tr>
<td>PZT</td>
<td>Lead Zirconate Titanate</td>
</tr>
<tr>
<td>MPPT</td>
<td>Maximum-PowerPoint-Tracking</td>
</tr>
<tr>
<td>VRF</td>
<td>Variable Resonant Frequency</td>
</tr>
<tr>
<td>FRF</td>
<td>Fixed Resonant Frequency</td>
</tr>
<tr>
<td>ZCS</td>
<td>Zero Current Switching</td>
</tr>
<tr>
<td>ZDS</td>
<td>Zero Derivative Switching</td>
</tr>
<tr>
<td>PCB</td>
<td>Printed Circuit Board</td>
</tr>
<tr>
<td>SS</td>
<td>Series-Series</td>
</tr>
<tr>
<td>SP</td>
<td>Series-Parallel</td>
</tr>
<tr>
<td>PS</td>
<td>Parallel-Series</td>
</tr>
<tr>
<td>SiC</td>
<td>Silicon Carbide</td>
</tr>
</tbody>
</table>
## TABLE OF CONTENTS

LIST OF TABLES .......................................................................................................................... xi
LIST OF FIGURES ....................................................................................................................... xii

1. **INTRODUCTION** .................................................................................................................. 1
   1.1. Capacitive coupling plate structures .............................................................................. 3
   1.1.1. Two-plate coupling structure .................................................................................. 3
   1.1.2. Four-plate coupling horizontal structure ................................................................. 4
   1.1.3. Four-plate vertical structure .................................................................................... 5
   1.1.4. Six-plate coupling structure ................................................................................... 6
   1.2. Compensation networks used in CPT system ............................................................... 7
       1.2.1. Series L ................................................................................................................ 7
       1.2.2. LC compensation ................................................................................................ 8
       1.2.3. LCL compensation .............................................................................................. 8
       1.2.4. LCLC compensation .......................................................................................... 9
   1.3. Topologies used for CPT system .................................................................................. 10
       1.3.1. PWM converter topology ................................................................................... 11
       1.3.2. Power amplifier-based topology ......................................................................... 12
       1.3.3. Full-bridge inverter topology ............................................................................ 13
   1.4. Advantages of CPT over IPT ..................................................................................... 14
   1.5. Limitations of the CPT ............................................................................................. 15
   1.6. Thesis contribution .................................................................................................. 16
   1.7. Thesis organization .................................................................................................. 17

2. **DISCUSSION ON CAPACITIVE COUPLING & ANALYSIS ON RESONANT POINTS** .... 19
   2.1. Applications of capacitive coupler with dielectric as medium: .................................. 19
       2.1.1. Analysis on capacitive coupler with different dielectric materials .................. 21
   2.2. Comparison on capacitive plates of different structure .............................................. 24
       2.2.1. Design of coupling plates ................................................................................ 24
       2.2.2. Results of different coupling plates using single active switch converter ........ 26
   2.3. Analysis on resonant points of SLLD for MPPT ......................................................... 30
       2.3.1. Experiment verification ..................................................................................... 34
   2.4. Conclusion ............................................................................................................... 37

3. **OPTIMIZATION OF CLASS-E² BASED CPT SYSTEM** ................................................... 38
   3.1. Discussion on Parameter Variation and Maximum Power Transfer ............................. 38
       3.1.1. Theoretical Analysis of Class-E² Capacitive-Wireless Power Transfer System ... 40
       3.1.2. Class-E inverter and it’s ZVS Limits ................................................................... 41
       3.1.3. Coupling Section with Resonant Network ......................................................... 42
       3.1.4. Class-E rectifier and its Equivalent Circuit ....................................................... 43
       3.1.5. Analysis on Optimal Impedance Tracking of Class-E² Converter .................... 45
       3.1.6. Discussion on the Impact of Duty and Frequency on ZVS Property ................. 55
   3.2. Optimization of class-E rectifier using finite secondary inductance ........................... 59
       3.2.1. Analysis on Class-E² converter with finite primary and secondary inductance ... 60
       3.2.2. Consideration of Class-E inverter and matching network ............................... 60
3.2.3. Analysis on Class-E rectifier with finite inductance ...................................................... 61
3.2.4. Results and discussions ...................................................................................................... 63
3.3. Conclusion ............................................................................................................................. 66

4. A NOVEL 8-PLATE MULTI-RESONANT COUPLING STRUCTURE FOR DRONE CHARGING .................................................................................................................. 68
   4.1. Methods to minimize the effects of Misalignments .............................................................. 68
   4.2. Implementation of multi-resonance to minimize the impact of misalignment .................. 69
      4.2.1. Frequency splitting in WPT systems .......................................................................... 69
      4.2.2. Class-E² with 8-plate multi-resonant coupling structure .............................................. 71
      4.2.3. Impedance analysis of 8-plate multi-resonant coupling structure ............................... 73
   4.3. Simulation and hardware test analysis of 4-plate and 8-plate multi-resonant coupling structure ............................................................................................................................ 77
      4.3.1. Simulation tests ............................................................................................................. 77
      4.3.2. Experimental validation ............................................................................................... 82
   4.4. Conclusion ............................................................................................................................. 90

5. Summary and Conclusion ......................................................................................................... 91
   5.1. Summary and conclusion .................................................................................................... 91
   5.2. Recommended future work .................................................................................................. 93

REFERENCES .................................................................................................................................. 94

VITA ............................................................................................................................................... 109
## LIST OF TABLES

<table>
<thead>
<tr>
<th>Table</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1. Coupling capacitance measured under different frequencies</td>
<td>25</td>
</tr>
<tr>
<td>2. Specified parameters of SLLD circuit</td>
<td>27</td>
</tr>
<tr>
<td>3. List of three primary secondary inductor-pair cases</td>
<td>33</td>
</tr>
<tr>
<td>4. SLLD resonant frequency for respective coupling capacitance</td>
<td>33</td>
</tr>
<tr>
<td>5. Specification of SLLD circuit parameters</td>
<td>34</td>
</tr>
<tr>
<td>6. Comparison of output voltages under VRF and FRF controllers</td>
<td>36</td>
</tr>
<tr>
<td>7. Parameters of Class-E$^2$ power converter for CPT</td>
<td>46</td>
</tr>
<tr>
<td>8. Hardware results for various duty cycle and frequencies of Class-E$^2$</td>
<td>58</td>
</tr>
<tr>
<td>9. Class-E$^2$ Circuit parameters</td>
<td>63</td>
</tr>
<tr>
<td>10. Cases based on additional resonant inductor values</td>
<td>75</td>
</tr>
<tr>
<td>11. Parameters of class-E$^2$ power converter with 4-plate, and 8-plate multi-resonant coupling</td>
<td>79</td>
</tr>
</tbody>
</table>
LIST OF FIGURES

1. Block diagram of CPT system ................................................................. 2
2. Two-plate coupling structure ................................................................. 3
3. Four-plate horizontal coupling structure ............................................... 4
4. Four-plate vertical plate structure ............................................................ 6
5. Six-plate coupling structure ................................................................. 7
6. Compensation networks for CPT systems ............................................... 10
7. PWM converter topology ........................................................................ 11
8. Power amplifier-based topology with vertical plate structure .................. 12
9. Full-bridge converter .............................................................................. 14
10. Dielectric material properties under different application conditions .... 20
11. Conformal bumper and the wall as capacitor plates ................................ 21
12. Variation of coupling capacitance w.r.t the distance ................................ 23
13. Structure and specification of coupling plates ......................................... 25
14. Topology of SLLD circuit ..................................................................... 27
15. Testbed of SLLD circuit ......................................................................... 27
16. Waveforms of SLLD circuit components during test: driving signal of MOSFET (top, orange), voltage across receiving-side inductor (middle, green), and voltage across coupling capacitor (bottom, purple). ................................................................. 29
17. Output voltage comparison with square-shape PZT plates and 17.5 nF film capacitors .... 29
18. Output voltage comparison of three cases at coupling section .............. 30
19. The resonant tank of SLLD circuit to be analysed .................................. 31
20. The Bode plots of three inductor-pair cases .......................................... 33
21. Specified SLLD circuit and its testbed ................................................................. 35
22. Output voltage of SLLD circuit with different coupling capacitances under VRF controller. 35
23. Class-E\textsuperscript{2} converter with compensation network ........................................ 40
24. Circular capacitive-coupling plates manufactured for hardware tests in lab. .................. 43
25. Class-E\textsuperscript{2} converter with equivalent rectifier circuit ..................................... 45
26. Impedance variation w.r.t frequency for different resonant inductance ......................... 48
27. Simulink plots of input impedance w.r.t frequency (a) \( L_r = 100 \mu\text{H} \), (b) \( L_r = 50 \mu\text{H} \), and (c) \( L_r = 18 \mu\text{H} \) .......................................................... 50
28. Variation of current through input side, MOSFET, and resonant network with respect to the frequency ........................................................................................................ 51
29. Variation of current through input side, MOSFET, and resonant network for different resonant inductor values .............................................................. 52
30. Hardware setup of the CPT system ........................................................................... 53
31. Hardware results of Class E\textsuperscript{2} converter (a) Input current vs. Frequency, (b) Output Voltage vs. Frequency ................................................................. 54
32. Waveform across the shunt capacitor (a) ZVS (b) early ZVS ....................................... 56
33. Waveform across the shunt capacitor for frequency (a) 90 kHz, (b) 130 kHz, (c) 180 kHz . 56
34. Waveform across the shunt capacitor for duty cycle (a) \( D = 0.3 \), (b) \( D = 0.5 \), (c) \( D = 0.7 \). 57
35. Class-E\textsuperscript{2} converter with impedance marking .................................................. 60
36. Front view and side view of circular capacitive plates .................................................. 61
37. Equivalent circuit of Class-E rectifier (a) diode in reverse bias, (b) diode in forward bias ... 62
38. Hardware setup of the Class-E\textsuperscript{2} based CPT system ........................................ 64
39. Hardware results of the Class-E\textsuperscript{2} converter (a) Input current w.r.t switching frequency, (b) Output voltage w.r.t switching frequency

40. Voltage waveforms across the MOSFET for (a) $L_S=18 \, \mu\text{H}$, (b) $L_S=100 \, \mu\text{H}$

41. Drone charging using 8-plate coupling plates

42. Idea of 8-plate multi-resonant coupling structure

43. Class-E\textsuperscript{2} with 8-plate multi-resonant coupling structure

44. Equivalent circuit of class-E\textsuperscript{2} based CPT

45. Resonant impedance plots of 4-plate and 8-plate multi-resonant coupling with and without misalignment for three cases

46. 4-plate Class-E\textsuperscript{2} CPT Simulink model

47. 8-plate multi-resonant Class-E\textsuperscript{2} CPT Simulink model

48. Simulation results of (a) 4-plate and (b) 8-plate multi-resonant

49. Class-E inverter design for multi-resonance

50. Class-E rectifier design for multi-resonance

51. Capacitive-wireless charging setup for drone

52. Experimental set of 8-plate capacitive coupling

53. Output voltage of class-E\textsuperscript{2} CPT with and without misalignments

54. Waveform of the class-E\textsuperscript{2} with 8-plate multi-resonant coupling (Blue: Gate signal, Purple: Voltage across MOSFET, Orange: Voltage across diode)

55. Output voltage of class-E\textsuperscript{2} CPT with and without misalignments
CHAPTER 1
INTRODUCTION

Due to the enhancement in power ratings of semiconductor devices, wireless power transfer (WPT) technology expanded its number of applications that can transfer kilowatts of power up to several centimeters in range. WPT is an effective and convenient method to charge battery-powered devices [1] - [5]. This approach eliminates the bulky and wear-prone cables. Wireless power transfer follows the principle of electromagnetic induction. A typical WPT setup consists of an inverter, followed by the coupling section, and a rectifier. The coupling section and the medium associated with it play a significant role in determining the power transfer [6] - [8]. WPT is employed utilizing several methods such as inductive coupling, resonant coupling, and capacitive coupling. Each method has its pros and cons. Selection of the coupling section is dependent on the power ratings and transmission range [9].

The present market utilizes the inductive-wireless power transfer (IPT) to follow the Qi standards for mobile and laptop wireless charging [10]. Higher power applications like electrical vehicles (EV) follow the society of automotive engineer (SAE) J2954 standard with three power levels WPT1 (3.7 kW), WPT2 (7 kW), and WPT3 (11 kW) [11] - [12]. The operation of the IPT is similar to transformers, except IPT is operated at high frequency. IPT uses an inductive coil for low power transfer, whereas for the high power-high frequency it uses ferrite cores to achieve a high coupling co-efficient [13]-[14]. Although IPT has the capability to achieve efficiency of over 90% with a transmission range up to tens of centimeters [15], the requirement of ferrite core makes it bulky and expensive. An alternative to IPT is the capacitive-wireless power transfer (CPT). The operation of CPT is similar to that of IPT; it consists of a high-frequency inverter that converts
DC-AC voltage, coupling section with/without compensation network, and the high-frequency rectifier shown in Figure 1.

![Block diagram of CPT system](image)

Figure 1: Block diagram of CPT system

Unlike IPT, CPT utilizes capacitive plates instead of bulky and expensive ferrite cores. The capacitive plates provide flexibility in designing the coupling structure [15]-[16]. Coupling plates can be of any shape, and the necessity of the flat nature of the coupling plates can also be eliminated through the use of capacitive films. For example, in [17] the coupling structure is designed to adjust its shape based on the conformal bumper shape. The flexibility of change in shape could eliminate the uneven air gaps that lead to an increase in coupling capacitance. Also, the electric field is confined between the coupling plates, which limits the electric field radiation outwards. The flexibility in designing coupling structures will improve the coupling co-efficient, resulting in an increase in power transfer. The shape of the coupling structure can also reduce the edge effects related to it [18]. The edge effect is dependent on the perimeter of the coupling plates; it can be reduced by considering the appropriate shape. Typically, circular plates have a lower edge effect compared to square coupling plates [16].
1.1. Capacitive coupling plate structures

The capacitive couplers consist of films or plates that generate the electric field when AC voltage is applied across them. The coupling plates are manufactured using copper or aluminum foil with dielectric material or medium between the plates. Several coupling structures like a two-plate coupler, four-plate/4-plate parallel, four-plate/4-plate vertical, and a six-plate structure were used for previous works that will be discussed below.

1.1.1. Two-plate coupling structure

The CPT system requires at least two plates, one on the transmitter side and the other on the receiver side to form a coupling structure. The two-plate structure is a unipolar, electric field; there will be radiation from the primary end to the secondary end. The mutual capacitance of the coupling structure is solely dependent on the area and distance between the two plates. The two-plate coupling structure is shown in Figure 2. A two-plate coupling structure can be applicable for both short and long transmission ranges. It can also be designed as an asymmetric coupling structure which means coupling plates can be unidentical. Here, the conductive path can be different for each application. For example, in the propulsion of a high-speed train, the metals use a conductive path [19]. In the case of EV charging, the conductive path is a ground [20].

![Figure 2: Two-plate coupling structure](image)
A two-plate coupling structure is popular for its simplicity. Recent works [21]-[22] also indicate that it has fair tolerance to large misalignments. Considering a case where the parasitic capacitance with the ground is low, the two-plate coupling structure has an effective power transfer.

1.1.2. *Four-plate coupling horizontal structure*

The four-plate horizontal plate structure is the most popular and is a commonly used capacitive coupler [23]. This coupling structure forms a bipolar setup. One of the pairs is used to transfer power from primary to secondary, and the other pair can function as a return path [24]-[25]. The four-plate coupling structure is shown in Figure 3; the transmitter side consists of two plates P₁ and P₂, and they form pairs with the two plates P₃ and P₄ on the receiving side. The power transfer for this coupling structure is majorly dependent on the capacitance between the P₁ and P₃, P₂ and P₄. A CPT system with a transmission distance over 100 mm has a self and cross-coupling capacitance. It has a significant impact on the overall power transfer [26].

![Figure 3: Four-plate horizontal coupling structure](image-url)
The equivalent behavior source model of a capacitive coupler is shown [27]. It is a two-port network with the input port on the primary side and the output port on the secondary side. The self-capacitance ($C_1$, $C_2$) and cross-coupling capacitance ($C_M$) can be identified by using the (1), (2) and (3). When the primary and secondary sides are close to each other, the coupling capacitance will be high. In that case, both cross-coupling and self-capacitance can be neglected [28]. A four-plate horizontal plate structure is applied for various applications like EV charging [26], drone charging [29], and medical purposes [30].

$$C_1 = \frac{(C_{AB} + C_{AD})(C_{CB} + C_{CD})}{C_{AB} + C_{AD} + C_{CB} + C_{CD}}$$  \hspace{1cm} (1)$$

$$C_2 = \frac{(C_{AB} + C_{BC})(C_{AD} + C_{CD})}{C_{AB} + C_{AD} + C_{BC} + C_{CD}}$$  \hspace{1cm} (2)$$

$$C_M = \frac{-C_{AD}C_{BC} + C_{AB}C_{CD}}{C_{AB} + C_{AD} + C_{BC} + C_{CD}}$$  \hspace{1cm} (3)$$

1.1.3. Four-plate vertical structure

Low coupling capacitance has the requirement of high frequency, or installation of a high-value inductor is needed for compensation network [31]-[32]. To avoid this case, the four-plate vertical offers high self-capacitance of the coupling plates [32]. It can eliminate the requirement of the external capacitance of the compensation or resonant network. The four-plate vertical structure is shown in figure 4, it consists of the plates $P_1$ and $P_2$ on the transmitting end that is stacked one over the other. Here, the size of the $P_2$ is less than $P_1$. A similar setup is formed on the receiving end using $P_3$ and $P_4$. The self-capacitance of $P_1$-$P_2$ and $P_3$-$P_4$ is high as they are placed close to each other. The coupling capacitance and cross-coupling capacitance is noted as low. The power transfer in this coupling structure is majorly dependent on the mutual and cross-coupling
capacitance. The four-plate vertical structure is invulnerable to angular misalignment. The addition of circular plates will eliminate the angular misalignments completely [32]. The applications of four-plate vertical structures include high-power EV charging [32].

![Diagram of four-plate vertical plate structure](image)

Figure 4: Four-plate vertical plate structure

1.1.4. Six-plate coupling structure

The six-plate coupling structure is a combination of four-plate horizontal coupling and four-plate vertical coupling [33]-[34]. As shown in figure 5, the four-plate horizontal coupling structure is placed between the plates P₅ and P₆. Here, P₁, P₂, P₃, and P₄ work as active plates to transfer power between primary and secondary sides. P₅ and P₆ are used to increase the self-capacitance and also serve as electric field shielding. This coupling structure consists of 15 capacitances. In the main model, only the self and mutual capacitances are considered whereas cross-coupling capacitance is neglected. The relation between the two-port model and full capacitance is discussed in [33].
Compensation networks used in CPT system

Compensation is the key to the maximum power transfer; eliminating reactive power will increase the power delivered to the receiving side. For CPT, several resonant/compensation networks such as series L, LC, LCL, LCLC, CLLC are utilized for previous works [24]-[30]. The selection of compensation network is dependent on the power ratings and coupling capacitance of the CPT system. In this section, series L, LC compensation, LCL compensation, and LCLC compensation networks are discussed.

1.2.1. Series L

Series L is shown in Figure 6 (a). It is the simple compensation topology for the CPT system. As the capacitive coupler has capacitance at the coupling section, a simple series inductor can be added for the compensation. Two inductors, one for the primary side and the other for the secondary side, are used to compensate the coupling capacitance of the four-plate coupling structure [35]-[37]. The resonant inductor can be utilized for EMI suppression function to the input side [38]. For transmission range over 100 mm applications, the CPT requires a bulk inductor with large inductance to compensate for the coupling capacitance. The value of the inductor is
dependent on the switching frequency and the coupling capacitance. Series L is applicable for both high and low power applications but is preferred to use low distance applications i.e., capacitive coupler with high coupling capacitance [28]. The resonant point of the series L network is sensible when it comes to misalignment. The change in coupling capacitance will alter the resonant point which can disturb the load requirements.

1.2.2. *LC compensation*

For high transmission range applications, the coupling capacitance will be low. It requires high AC voltage on transmitting side to meet system power requirements. To overcome this issue, an LC compensation network shown in Figure 6 (b) can be used to increase the equivalent capacitance at the coupling section. The external capacitance from the LC network is connected in parallel with the self-capacitance, which enhances the power transfer. The capacitance of the external capacitors is much larger than the coupling capacitance. This eliminates the requirement of bulk inductors to resonate with the equivalent capacitance of the coupling section. The benefits of the double-sided LC compensation circuit is its feasibility in long distance and high power applications [39]. Due to the external capacitor connected in parallel, the resonance in the system is less sensitive to misalignment.

The disadvantage of a double-sided LC network is the system power is inversely proportional to the coupling coefficient [40]. With the increase in distance, the system power increases, but the system efficiency decreases significantly with the distance. Additionally, high power requires high AC voltage and current stress on the circuit components which leads to safety concerns.

1.2.3. *LCL compensation*
Combining LC and L compensation results in a double-sided LCL compensation topology [41]. Here, the inductance $L_1$ shown in Figure 6 (c) partially compensates for the mutual capacitance $C_{M1}$, $C_{M2}$, and the remaining capacitance is compensated by the LC network. LCL compensation network offers the flexibility to tune the system power by designing the inductance ratio of $L$ and $L_1$ shown in figure 6 (c), but the system power is still inversely proportional to the coupling coefficient. This network will require a high series inductor value for low coupling capacitance.

1.2.4. LCLC compensation

LCLC compensation shown in Figure 6 (d) overcomes the limitations of LC and LCL networks. Here, the additional LC networks $L_1$-$C_1$ and $L_2$-$C_2$ are installed on primary and secondary sides to convert the voltage sources into current sources for the resonant circuits. LCLC topology consists of multiple resonances; for example, $C_1$ resonates with $L$ and $L_1$, which is also used to increase the voltage in the capacitive coupler for sufficient power transfer. The advantages of a double-sided LCLC network are that system power is directly proportional to the coupling coefficient. The system power can be controlled using circuit parameter design without having an impact on the coupling coefficient [42]. Therefore, it can maintain a high coupling coefficient to satisfy high power requirements. Additionally, the LCLC network eliminates the requirement of a high-value inductor due to external capacitors. The disadvantages include complexity, increase in power losses due to more components, and increased cost.

This LCLC compensation can also be extended by cascading multiple LC networks. It serves as voltage gain and current gain networks to meet the system requirement for high transmission range systems [43]-[47]. Several other compensation networks include CLLC. LLC enhance the power transfer [48]- [50]. The general purpose of the compensation network is to
match the equivalent inductance of the L, LC, LCL, LCLC network to match with mutual capacitances.

![Diagram of Compensation Networks for CPT Systems](image)

**Figure 6:** Compensation networks for CPT systems

1.3. *Topologies used for CPT system*

A typical WPT consists of a high-frequency inverter and high-frequency rectifier. Several topologies like PWM converter, high-frequency amplifiers, half-bridge, and full-bridge converters are utilized for CPT systems. The selection of the topology is dependent on the coupling capacitance and the power requirements; these determine the system power capability, efficiency and frequency properties. A proper topology can amplify the AC voltage applied to the coupling plates for high power transfer [51]. Additionally, it requires a high coupling coefficient and an optimal load ratio to maximize system efficiency [52].

The CPT topologies are classified into resonant and non-resonant converters. Non-resonant converters include pulse width modulation (PWM) converters, where the intermediate capacitor is
designed to operate as coupling capacitors. Resonant converters include the high-frequency power amplifiers and half-bridge/full-bridge inverter with compensation network.

1.3.1. **PWM converter topology**

The CPT-based PWM converters include buck-boost, Cuk, sepic, and zeta converters [53]. Figure 7 shows the buck-boost converter, also known as switch-inductor-inductor-diode (SLLD) topology.

![Figure 7: PWM converter topology](image)

It is a modified version of the typical buck-boost converter. Two coupling capacitors make the topology look different from the conventional buck-boost converter. These capacitors store the energy components and provide isolation between the primary and secondary sides. The energy is stored when the switch is turned off and releases energy when it is on. The coupling capacitance and the switching frequency should be large enough to transfer sufficient power and to operate the PWM converter in continuous-conduction mode (CCM). The PWM converter is non-sensitive to parameter variation. In real-time applications, the coupling capacitance varies due to misalignment; in that case, PWM converter is more applicable as along as coupling capacitance
and frequency is high enough to operate in CCM. Due to the requirement of the high coupling capacitance, the system efficiency is relatively high.

The disadvantages of the PWM converter include the requirement of high coupling capacitance and a transmission range limited to 1 mm. To improve the coupling capacitance further, it requires high dielectric material which limits the number of applications. Soft-switching is one of the constraints, zero-voltage switching (ZVS) cannot be achieved for all load conditions [54]. As PWM topologies use a single active switch, the voltage stress experienced by the MOSFET is high for high power circuits. The power level of the system is limited because of the single-active switch. Therefore, to increase the power level, multiple PWM converters can be connected in parallel. Interleaved control strategy is applied to improve the performance of those systems [55].

1.3.2. Power amplifier-based topology

Class D, Class E, Class EF, and class ϕ converters fall into the category of CPT-based power amplifier topologies. The most commonly used power amplifier for capacitive-wireless power transfer is the class-E converter shown in Figure 8.

![Figure 8: Power amplifier-based topology with vertical plate structure](image-url)
From Figure 8, the resonant capacitor is modified to serve as capacitive couplers. They form an isolation between the primary and secondary sides and act as the resonant network in the CPT system. The working principle of the CPT-based class-E amplifier is the same as the conventional one. As the inductor L and \( C_1 \) form the series resonance, it behaves as a double-sided L-compensation network. The advantages of the CPT-based power amplifier include flexibility on transmission distance \([56]-[58]\). Unlike PWM converters, it eliminates the requirement of high coupling capacitance. As the power amplifier is operated at a high frequency, the size of the passive components (such as resonant inductors) can be reduced. Additionally, the power amplifier has the ability to maintain ZVS for various load conditions \([59]\).

The disadvantage of the CPT-based power amplifier is its sensitivity to parameter variation. Most WPT systems have misalignment that could vary the coupling capacitance significantly. Due to the series LC circuit, the resonant point can vary. All these variations will reduce the system power; in that case, ZVS property can also be disturbed, which will cause hard switching.

1.3.3. Full-bridge inverter topology

Full-bridge converters are suitable for high-power applications. A full-bridge inverter is an effective topology to generate high-voltage AC signals. It consists of four MOSFET switches which are controlled by PWM signals \([60]\). To regulate the power, it is convenient to adjust the switching frequency and duty ratio \([60]\). The secondary side consists of a full-bridge rectifier with four diodes to convert AC-DC i.e., passed to the load.
1.4. **Advantages of CPT over IPT**

The CPT has three advantages over IPT: low eddy current loss, low cost and weight, and good misalignment performance. When the IPT is operated at high frequency, it generates eddy-current losses in the nearby metal, leading to a significant increase in temperature for both low and high-power applications. Specifically for EV charging, that has a load requirement of 10s of kW generates magnetic field can induce a large amount of heat in the nearby metal object whereas in the CPT system electric fields are used to transfer power that eliminates the concerns of eddy-current losses [37][44][47]. IPT requires the two coils that work as a loosely coupled transformer. Due to the low coupling co-efficient, high current must be passed through the coils to transfer enough power [61]. With a requirement of high current, considering the skin depth of the conductor, a large amount of Litz-wire is required to build the coils at high frequency which increases the cost and weight of the coupling section [62]. To achieve a high coupling co-efficient, large ferrite cores are required. Air core is used in recent works to eliminate magnetic losses [63], but this requires a large space, making it difficult to use.

Previous works show that compared with IPT, CPT has less impact on the performance of the system due to misalignment. For example, in a CPT system with a metal size of 610 mm*610
mm can achieve an efficiency of 89.4% of the properly aligned capacitive plates with 300 mm misalignment [62]. However, in an IPT system with a coil dimension of 600 mm * 800 mm, the system power drops to 56% of the properly aligned value with 310 mm displacement [64]. Therefore, CPT is suitable for EV charging applications over the IPT when misalignment is a factor.

1.5. 

Limitations of the CPT

Inductive-wireless power transfer is popular in the present market. Over the past two decades, significant work has already been done on IPT, whereas CPT started gaining attention in the past decade. There are a few limits CPT needs to overcome: low power density, low efficiency, and strong magnetic field emissions. In an IPT system, an inductive coupler with 450 mm * 450 mm can achieve 7.7 kW power transfer for EV application, resulting in a power density of 38.0 kW/m2 [65], whereas in a CPT system with a capacitive coupler of dimensions 914 mm *914 mm, it can achieve 1.87 kW power transfer, with a power density of 2.2 kW/m2. From the mentioned works with long-distance applications, the power density of a CPT system is lower than the IPT. It is because of the low coupling capacitance (in pico-Farads) for transmission range over 100 mm. The effective method to increase the power density is by increasing the switching frequency and the plate voltage. The addition of a compensation network can enhance the voltage applied to the coupling section [45]-[55].

From [64], the IPT system achieved an efficiency of 97% from source to load, whereas CPT can attain 91.6% for EV application [66]. As the CPT system is operated at high frequency (in MHz), the power losses at the inductor are a major contributor in downgrading efficiency. The IPT systems generate magnetic fields to transfer the power. The magnetic fields can be conformably shielded using ferrite and aluminum plates. It will minimize the leakage field to the
safety level around the coupling section [42], but the electric fields tend to pass through the metal [21]. Capacitive plates require high AC voltage to generate electric fields; the 4-plate vertical structure and six-plate coupling structure provide better shielding [32]-[33].

1.6. Thesis contribution

Several works on capacitive-wireless power transfer advanced the topic over the past decade with few limitations. Electric field shielding is among them. The electric fields generated by capacitive couplers are difficult to shield compared with magnetic fields. Another limitation includes the low power density of the capacitive couplers. To enhance the power density, additional circuitry is required to boost the AC voltage to the coupling plates. With these limitations, CPT is often used in applications with a low transmission range or low power. The concentration of the dissertation is on wireless drone charging. The UAVs are used for many practical purposes like surveillance [67]-[72], delivery [73]-[74], agriculture [75]-[77] and healthcare [78]-[81], etc. However, the weight of the UAV is a trade-off for battery and flying time. To increase reliability, WPT is applied in UAVs to recharge the battery to improve the working time. Previously, different approaches were applied for drone charging; one of them focused on combining high operating frequency with the data transmission. These CPT systems can both charge and transmit data simultaneously over the same channel. It transfers data at a rate of 170-kbps with 1-MHz of operating frequency and achieves a power efficiency of 70%. The alternative design [76] minimizes the data rate to 119-kbps and enhances power efficiency to 90.5%. [82] used a single active switch class-D inverter operated at high-frequency 6.78-MHz, but it requires LC filter and impedance transmission circuits such as step-up and step-down transformer for drone charging.
Misalignment is a common issue in the WPT system. As the Class-E converter is sensitive to parameter variations, the circuit behavior is noticed for misalignments. The impedance analysis on the class-$E^2$ based CPT system with a 4-plate capacitive coupler presents the impact of change in the resonant point of the 4-plate coupling structure. It provides a scope to design an optimal series-series resonant network for the 8-plate multi-resonant coupling. The conditions of secondary inductor of the class-E rectifier to improve power flow from the receiving plates to the load are identified. The design of the secondary inductor can enhance the output power and maintains zero-voltage-switching (ZVS) at optimal conditions.

Misalignments cause power transfer to plummet. Most previous works on this issue focused on designing an appropriate controller for misalignments or through the cascade of a compensation network. In our research, a novel 8-plate multi-resonant coupling is adopted to improve the output voltage for misalignments. The proposed method uses multiple series-series compensation networks to improve power flow for misalignments. This coupling design widens the resonant frequency band which helps in improving power flow for misalignments. The 8-plate multi-resonant coupling minimizes the reactance of the coupling section and offers better performance for larger misalignments.

1.7. Thesis organization

The remainder of the thesis is organized as follows: Chapter 2 discusses the selection of optimal dielectric material based on the variation of coupling capacitance with transmission distance. The effects of coupling plate shape on the CPT are realized through simulations and hardware tests. This chapter also presents analysis of the resonant point of the modified buck-boost converter or SLLD for CPT applications. Chapter 3 analyzes the impedance variations of the CPT-based class-$E^2$ converter for several resonant inductors which also explains the change in
resonance points due to misalignments. This chapter also displays the conditions to improve the power flow from receiving plates to load through the selection of secondary inductor. Chapter 4 introduces a novel 8-plate multi-resonant coupling which widens the resonant frequency band to improve the performance of class-E\(^2\) for misalignments. Chapter 5 summarizes with the conclusion and possible future works.
Chapter 2

DISCUSSION ON CAPACITIVE COUPLING & ANALYSIS ON RESONANT POINTS

This chapter presents the work on coupling capacitors and analyzing the resonant points of a single-switch converter. CPT offers low coupling capacitance (in pico-farads) for high transmission range applications. It requires a circuit to generate a high frequency-high voltage, and a dielectric with high breakdown voltages must be considered. For this case, the electric field radiates outwards, making it difficult to shield for high-power long-distance applications [18]. In the case of a transmission range over 100 mm, IPT is suitable to achieve high efficiency. However, for a low transmission range (in 10s mm), CPT is more suitable. In a high-power EV application with a transmission range less than 2mm [17]-[84], CPT offers an efficiency of over 90%. Additionally, it contains the electric field between the coupling section that cancels out the external fields, increasing the safety range.

1.8. Applications of capacitive coupler with dielectric as medium:

Capacitive plates offer flexibility in designing coupling sections. The coupling section can be designed to achieve maximum power transfer. [85] presents work on the capacitive ball-joint structure; for this application, CPT offers constant coupling over the rotatable angular range with an efficiency over 80%. In the case of IPT, the mutual inductance is varied for a similar setup [86] that provides the benefit of CPT over IPT. CPT offers low cost, minimizes size, and increases efficiency for low transmission range applications. The dielectric constant is directly proportional to the coupling capacitance; the coupler can be designed to maintain the medium as a complete dielectric for low transmission range. For example, [87] uses the car window as the dielectric medium for capacitive coupling. This setup offers high coupling capacitance and eliminates both the requirement to operate high-frequency and the high AC voltage applied to the coupling plates.
The properties of various dielectric materials for the capacitive coupling section are introduced in [88]. With the increase in frequency, the demonstrated capacitance decreases [18], this will impact the circuit behavior and thus the performance of the CPT system under high-frequency operation. Figure 10 presents the dielectric material used for various frequencies and capacitance values.

Figure 10: Dielectric material properties under different application conditions

The coupling interface is an essential part of the CPT system. For many applications, transmitting plates are in a fixed position, whereas the receiving plates are movable. As mentioned
earlier, EV wireless charging can be implemented using a conformal bumper [84]. Figure 11 shows
the dielectric material in the form of foam attached to the wall on the transmitting side. The
receiving side plates are attached to the rigid bumper. Due to the uneven nature of the bumper,
there might be more air gap than the calculated air gap between the flat plates. Additionally, the
transmitted electric field is not consistent throughout the area of the capacitor plates. To confront
this issue the dielectric foam is attached to the metal foil to match the shape of the rigid surface
when the bumper comes in contact with the wall [84]. This foam can also reduce the probability
of damage to the vehicle’s bumper. Additionally, this setup offers high coupling capacitance.

Figure 11: Conformal bumper and the wall as capacitor plates

1.8.1. *Analysis on capacitive coupler with different dielectric materials*

For the analysis of different dielectric materials, the study is conducted with an electric
vehicle battery as load. For this setup, the coupling plates are attached to the EV bumper, and the
area of the capacitive plates is assumed to be 1000 mm * 250 mm. The dielectric constant is
proportional to the distance between coupling capacitor plates. The base distance to achieve
specific coupling capacitance will be different for each dielectric material. Based on theoretical
analysis and simulation study, the nominal coupling capacitance is 50 nF to achieve rated output with a high coupling coefficient. The base distance to achieve capacitance of 50 nF for each dielectric material is 100mm for lead zirconium titanate (also called PZT), 50mm for PbLaZrTiO3, and 13mm for BaSrTiO3. To notice circuit behavior output voltage, power efficiency, and the resonant frequency shift should be analyzed for different transmission range.

To identify the variation of coupling capacitance with respect to the transmission distance three dielectric materials are considered here, i.e., PZT (with $\varepsilon_r = 2400$), PbLaZrTiO3 (with $\varepsilon_r = 1000$), and BaSrTiO3 (with $\varepsilon_r = 300$). Figure 12 presents the relationship between the coupling capacitance and the distance between two coupling plates for three dielectric materials. For short distances, the demonstrated capacitances drop sharply. For longer distances, the change in coupling capacitance is minor.

![Graph](image-url)
The quality factor and capacitive film of the coupling plates significantly impact the overall efficiency of the system. The power losses at capacitive interference, Inductors, MOSFET, and Schottky diode define the efficiency of the CPT system. The power dissipated at coupling capacitor (P_{CC}) is calculated using (4):
\[ P_{CC} = V_{RMS} \omega C \tan(\delta) \]  
\[ \tan(\delta) = \frac{1}{Q} \]  
\[ Q = \frac{1}{\omega C_{R} R_{C}} \]

1.9. **Comparison on capacitive plates of different structure**

1.9.1. **Design of coupling plates**

To identify the impact of coupling plate shape, round and square lead zirconate titanate (PZT) plates are considered as shown in Figure 13. Figure 13 (a) shows the structure of the vertical layer. For the design of coupling capacitors shown in Figure 13 (b), PZT plates with a dielectric constant of 2400 are used as a medium, and copper foil is used as capacitive foil because it is 0.6 times less resistant than aluminum foil [90]. For mobile charging applications, the distance between capacitive plates is considered as 2mm, which is about the thickness of a cell phone’s back cover. Table 1 lists the measured capacitance of coupling plates made of PZT material and film capacitors under different frequencies.

![Diagram](image)
Table 1: Coupling capacitance measured under different frequencies

<table>
<thead>
<tr>
<th>Testing Frequency</th>
<th>Film capacitor (Marked as 25nF)</th>
<th>Round PZT Capacitance</th>
<th>Film capacitor (Marked as 17.5nF)</th>
<th>Square PZT Capacitance</th>
</tr>
</thead>
<tbody>
<tr>
<td>100 Hz</td>
<td>25.07 nF</td>
<td>25.13 nF</td>
<td>17.39 nF</td>
<td>17.27 nF</td>
</tr>
<tr>
<td>1 kHz</td>
<td>25.06 nF</td>
<td>23.88 nF</td>
<td>17.38 nF</td>
<td>16.84 nF</td>
</tr>
<tr>
<td>10 kHz</td>
<td>25.04 nF</td>
<td>23.67 nF</td>
<td>17.37 nF</td>
<td>16.55 nF</td>
</tr>
<tr>
<td>100 kHz</td>
<td>25.10 nF</td>
<td>20.95 nF</td>
<td>17.37 nF</td>
<td>12.24 nF</td>
</tr>
</tbody>
</table>

Table 1 presents coupling capacitance concerning the frequency. It is noted that the capacitance of PZT plates is reduced drastically compared to the capacitance of film capacitors, along with an increase in measurement frequency. Also, it is found that when the frequency increases from 100 Hz to 100 kHz, the round-shape PZT plates reduce coupling capacitance by 16.7%, which is much smaller than the square shape PZT plates of 29.2%.
1.9.2. Results of different coupling plates using single active switch converter

The selection of circuit topology for C-WPT has a significant effect on system performance. Here, a single-switch topology in [17] is considered for its potential low power losses and simple controller and circuit design. Unlike other topologies such as power amplifier or bridge-style topologies, a single switch minimizes the switching losses in the inverter circuit. The PWM converter-based CPT systems are not sensitive to parameter variation. The efficiency of the PWM converter-based CPT is high because of its large coupling capacitance [89]. As this topology has only one active switch, the power rating of the system is limited. The initial part of the work is on PWM converter-based CPT; the topology of a Single Switch-Inductor-Inductor-Diode (SLLD) circuit is adopted for the analysis of capacitive-wireless power transfer.

The SLLD circuit is a modified version of the buck-boost DC-DC converter. This circuit has a relatively simple design, which makes it easy to analyze the impact of each component in the topology. Additionally, it eliminates DC bias at the coupling section, which allows AC voltages to pass through. The SLLD is applicable for both low power and high-power applications [84]. The topology is divided into two sections – transmitter and receiver – as shown in Figure 14. The transmitter section consists of a MOSFET and a primary inductor. The high-frequency inverter generates AC voltage signals to energize the resonant tank. The receiver section consists of the secondary inductor and a diode rectifier to generate DC output. The two inductors (\(L_t\) and \(L_r\)) and the capacitive coupling plates form a resonant tank. The bipolar coupling section is used, and it is a four-capacitor coupling plate as shown in Figure 14.
Figure 14: Topology of SLLD circuit.

Table 2 lists the specified parameters of a testing circuit, and Figure 15 shows the setup of the hardware testbed.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Load Resistance ($R_L$)</td>
<td>100 Ω</td>
</tr>
<tr>
<td>Input Resistance ($R_S$)</td>
<td>0.5 Ω</td>
</tr>
<tr>
<td>Input Capacitor ($C_{in}$)</td>
<td>100 nF</td>
</tr>
<tr>
<td>Transmitting side Inductor ($L_t$)</td>
<td>220 µH</td>
</tr>
<tr>
<td>Receiving side Inductor ($L_r$)</td>
<td>10 µH</td>
</tr>
<tr>
<td>Output Capacitor ($C_{out}$)</td>
<td>1 µF</td>
</tr>
</tbody>
</table>

Figure 15: Testbed of SLLD circuit.
Hardware tests are performed to notice the effects of coupling plates on the CPT system. Initial tests were performed to observe the output voltage from the SLLD circuit: 1) using film capacitors as a coupling section and 2) using square PZT plates as a coupling section. From Figure 16, it is noted the coupling section with PZT generates harmonics in current and voltage waveforms. Due to the harmonics at the coupling section, the output voltage is a distorted curve as shown in Figure 17. Also, from Table 2, it is noticed that coupling capacitance reduces with the increase in frequency. The reduction levels are different in PZT and film capacitors. The noise and excessive reduction in capacitance lead to a voltage drop at load as shown in Figure 17.

(a) using film capacitors as coupling section

(b) using square PZT plates as coupling section
Figure 16: Waveforms of SLLD circuit components during test: driving signal of MOSFET (top, orange), voltage across receiving-side inductor (middle, green), and voltage across coupling capacitor (bottom, purple).

Figure 17: Output voltage comparison with square-shape PZT plates and 17.5 nF film capacitors

For further analysis, three cases with different combinations of PZT plates and film capacitors are considered for hardware tests. The bi-polar coupling section consists of two coupling capacitances, and each capacitance is formed by a pair of PZT plates or a film capacitor:

Case #1 – “Round PZT plates” + “Square PZT plates”,

Case #2 – “Square PZT plates” + “Film capacitor”,

Case #3 – “Round PZT plates” + “Film capacitor”.

Figure 18 shows the output voltage measurement of these three cases. Results demonstrate that the film capacitors can minimize noise in the circuit. The coupling section with a combination of round plates and film capacitor has less distortion and voltage drop than the other two cases. From the results, it is noticed that the round plates have better performance than the square ones. A 7% improvement in peak output voltage is observed by comparing case 2 and case 3. It is caused due to the edge effects. The edge effects are dependent on the perimeter of the coupling plate.
Round plates have less perimeter compared with square plates. It can be proved using finite element analysis of the PZT plates.

Figure 18: Output voltage comparison of three cases at coupling section.

1.10. **Analysis on resonant points of SLLD for MPPT**

To design a high-efficient CPT system the following aspects should be considered: circuit topology, power semiconductor devices, switching frequency, and other auxiliary parts. Apart from the semiconductor losses, the two important factors that influence the efficiency of the CPT system are: 1) the control method to achieve the resonant point, and 2) the dielectric material (media) that influences the quality of the capacitive coupling plates. Therefore, in this work, we studied the parameter specification and resonant tank of a single-switch power converter for designing the optimal control.

To study the behavior of the hybrid-connected resonant tank, two resonant points are formalized in (7) and (8) for the series- and parallel- resonant points, respectively. (9) shows the quality factor of the resonant tank. Figure 19 presents the resonant tank of the SLLD circuit to be
analyzed, which consists of two coupling capacitances ($C_C$), one transmitting inductor ($L_t$), and one receiving inductor ($L_r$).

![Diagram of resonant tank](image)

**Figure 19:** The resonant tank of SLLD circuit to be analysed.

\[
\begin{align*}
    f_{\text{parallel}} &= \frac{\sqrt{2}}{2\pi\sqrt{(L_t+C_C+L_r+C_C)}} \\
    f_{\text{series}} &= \frac{\sqrt{2}}{2\pi\sqrt{(L_r+C_C)}} \\
    Q_r &= \frac{1}{\omega_r R_c C_C} = \frac{1}{2\pi f_r R_c C_C}
\end{align*}
\]  

Whereas,

- $C_C$: the coupling capacitance of both coupling plates; $f_r$: the resonant frequency (series or parallel) of resonant tank.

- $Q_r$: the quality factor of resonant tank; $RC$: the damping resistance in resonant tank.

These two resonant points are generated from the series branch and the parallel branch in the resonant tank, respectively. When $L_r \geq L_t$, both the series and parallel resonance points are close to each other; the equivalent impedance will be high for this case. Thus, the resonance
phenomenon does not exist. When \( L_f \ll L_t \), the two resonant points are far from each other thus creating a peak magnitude at the output.

The maximum power transfer can be achieved by adjusting the resonant points, such as a VRF controller in [24]. By analyzing the characteristic of a resonant tank, the power delivered is minimal at the parallel resonant point and is maximum at the series resonant point. When both series and parallel resonant points are close to each other, the resonance effect of the circuit is lost. To overcome this issue, the separation of two resonant points must be reflected in the design. Figure 20 shows the Bode plots of three inductor-pair cases: 1) \( L_t = L_f \), 2) \( L_t > L_f \), and 3) \( L_t \gg L_f \). It is noted that a small inductance of \( L_f \) could separate the two resonant points for MPPT control. It is important to choose an appropriate value for the receiving-side inductor. If the value is too low, it could lead to high AC voltage signals at the receiving-side inductor, which would increase voltage stress and power loss at the components on the receiving side. Here, the secondary inductance of 10 \( \mu \)H is considered for the receiving-side inductor, which is about 20 times smaller than the transmitting-side inductor of 220 \( \mu \)H.

Figure 20 presents the bode plot of series and parallel resonant points of the coupling section. Three cases are considered to identify the separation of resonant points. For each case, the secondary inductance is varied. From the theoretical and simulation results, case 3 presents the optimal condition. The secondary inductance is considered low to separate the series and parallel resonant points. This increases the performance of the system. In case 1, both the series and parallel resonant points overlap with each other. Therefore, the output voltage does not bound with the increase in switching frequency, which makes the CPT system unstable. For case 3, it has a series resonant point at which the output voltage can be maximized is proven with the hardware tests.
Table 3: List of three primary secondary inductor-pair cases

<table>
<thead>
<tr>
<th>Cases</th>
<th>Inductance of $L_t$</th>
<th>Inductance of $L_r$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$L_t = L_r$</td>
<td>220 µH</td>
</tr>
<tr>
<td></td>
<td></td>
<td>220 µH</td>
</tr>
<tr>
<td>2</td>
<td>$L_t &gt; L_r$</td>
<td>220 µH</td>
</tr>
<tr>
<td></td>
<td></td>
<td>100 µH</td>
</tr>
<tr>
<td>3</td>
<td>$L_t &lt; L_r$</td>
<td>220 µH</td>
</tr>
<tr>
<td></td>
<td></td>
<td>10 µH</td>
</tr>
</tbody>
</table>

With the secondary inductor specification, the resonant points for three coupling capacitances are presented in Table 4. These coupling capacitances are different due to the distance of the coupling plates.

Table 4: SLLD resonant frequency for respective coupling capacitance

<table>
<thead>
<tr>
<th>Coupling Capacitance</th>
<th>Parallel Resonant Frequency (kHz)</th>
<th>Series Resonant Frequency (kHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>20 nF</td>
<td>111.10</td>
<td>516.8</td>
</tr>
<tr>
<td>15 nF</td>
<td>121.35</td>
<td>583.43</td>
</tr>
<tr>
<td>10 nF</td>
<td>148.58</td>
<td>714.28</td>
</tr>
</tbody>
</table>

Figure 20: The Bode plots of three inductor-pair cases.
1.10.1. *Experiment verification*

The effect of resonant point separation was experimentally verified by performing hardware tests. Figure 21 shows the setup of the hardware testbed. The parameters of the testing circuit are specified in Table 5.

<table>
<thead>
<tr>
<th>Items</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage</td>
<td>15 V</td>
</tr>
<tr>
<td>Load Resistance (R&lt;sub&gt;L&lt;/sub&gt;)</td>
<td>100 Ω</td>
</tr>
<tr>
<td>Input Capacitance (C&lt;sub&gt;in&lt;/sub&gt;)</td>
<td>100 nF</td>
</tr>
<tr>
<td>Transmitting side Inductor (L&lt;sub&gt;t&lt;/sub&gt;)</td>
<td>220 µH</td>
</tr>
<tr>
<td>Receiving side Inductor (L&lt;sub&gt;r&lt;/sub&gt;)</td>
<td>10 µH</td>
</tr>
<tr>
<td>Output Capacitor (C&lt;sub&gt;out&lt;/sub&gt;)</td>
<td>1 µF</td>
</tr>
</tbody>
</table>

(a) prototype of specified SLLD circuit
(b) Setup of testbed

Figure 21: Specified SLLD circuit and its testbed.

Figure 22 shows the output voltage of the SLLD circuit for different coupling capacitances. The input voltage for the hardware test is 15 VDC. Operating frequency range is [20 kHz, 700 kHz], which is sufficient to separate both series- and parallel-resonant points. By separating resonant points and with specified SLLD parameters, the maximum output voltage can be obtained when the operating frequency is close to the series resonant point. Here, the coupling plates are arranged far from each other to exclude the mutual capacitance and other effects. The main motive of the work is to identify the resonant points and the impact of VRF on the output voltage.

Figure 22: Output voltage of SLLD circuit with different coupling capacitances under VRF controller.
Considering the coupling capacitance of 17.5 nF from PZT plates as a rated capacitance in SLLD circuit design, the switching frequency keeps a constant of 537.3 kHz for the typical FRF control. By applying the VRF controller, the switching frequency can autonomously move to the actual resonant point, which reflects a certain coupling capacitance for a distance between coupling plates. Table 5 summarizes and compares the effect of VRF and FRF controllers on the output voltage. The result of tests verified that the VRF controller can realize MPPT control in the SLLD circuit dynamically, which has an effective improvement on the voltage holding and power delivery capability over broader variable coupling distance.

Comparing the results from Tables 4 and 6, there is a shift in natural resonant points. It is because of the parasitic elements under high frequency. By considering the shift in resonant point, the operating frequency of FRF is adjusted to 480 kHz. That is noticed from the baseline test with 17.5 nF. The resonant point shift will not impact the VRF controller performance in practice since the controller tracks the actual resonant points onboard. From Table 6, it is noted that the affect of VRF control becomes distinct when the coupling distance is farther from its rated value. It will improve the power delivery capability of the SLLD circuit for a high transmission range in CPT application i.e., for low coupling capacitance values.

Table 6: Comparison of output voltages under VRF and FRF controllers

<table>
<thead>
<tr>
<th>Coupling Capacitance</th>
<th>10nF</th>
<th>15nF</th>
<th>20nF</th>
</tr>
</thead>
<tbody>
<tr>
<td>VRF Controller</td>
<td>17.5V</td>
<td>19.68V</td>
<td>21.23V</td>
</tr>
<tr>
<td>(at 620kHz)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>FRF Controller</td>
<td>16.25V</td>
<td>19.68V</td>
<td>20.20V</td>
</tr>
<tr>
<td>(f_{sw} = 537.3kHz)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Improvement in voltage level</td>
<td>7.69%</td>
<td>0</td>
<td>5.10%</td>
</tr>
<tr>
<td>FRF Con. Adjusted ($f_{sw}$ adj = 480kHz)</td>
<td>15.29V</td>
<td>19.06V</td>
<td>21.05V</td>
</tr>
<tr>
<td>Adjusted Improvement in voltage level</td>
<td>14.45%</td>
<td>3.25%</td>
<td>0.86%</td>
</tr>
</tbody>
</table>

1.11. Conclusion

Chapter 2 presents the work on the coupling section. To improve the performance of the coupling section, an investigation is performed to identify the material with optimal dielectric coefficient, breakdown voltage, and cost. To investigate dielectric, the plot of the coupling capacitance using PZT, PbLaZrTiO3, and BaSrTiO3 as the dielectric is presented with different transmission range. From the results, PZT is identified as the ideal dielectric material for near-field CPT systems. Further analysis is done to determine the impact of coupling plate shape on power transfer. The hardware results using SLLD circuit indicate that the circular plates perform better than the square plates due to minimizing edge effects.

In addition to coupling plates, the identification of resonance in the circuit contributes significantly to improving power transfer. The SLLD has a series and parallel impedance which requires the separation to enhance the power transfer. As the series and parallel resonant points are dependent on the primary and secondary of the SLLD topology, the value of the secondary inductor is varied to enhance the power. This work presents the conditions to design the CPT topology. Because of the requirement of the high coupling capacitance and the noises across the coupling section, SLLD limits the applications. For this case, Class-E$^2$ is considered for the following chapters.
Chapter 3

OPTIMIZATION OF CLASS-E$^2$ BASED CPT SYSTEM

This chapter covers the key portion of this dissertation. It presents the circuit behavior of the Class-E$^2$ converter with capacitive coupling. Further analysis is performed to optimize the Class-E$^2$ for achieving high efficiency. For this dissertation, the CPT system is designed considering the drone battery as an application, with a rating of 100 Ah and a transfer distance within a 1 cm range. For near-field applications, the capacitive-wireless power transfer method has proven efficient and safe [24]. Inverter and rectifier topologies are key to the WPT system. The inverter function is to provide AC voltage to the coupling section, but it has a crucial role in enhancing efficiency. In the previous chapter, the analysis of the coupling section is performed using SLLD topology; it operates efficiently for high coupling capacitance. However, the SLLD limits the number of applications because of the requirement of high coupling capacitance. Misalignment is a common problem in the WPT systems. The coupling coefficient drops significantly because of the misalignment. In the case of low coupling capacitance, Class-E$^2$ is an effective topology over SLLD. Though Class-E$^2$ is sensitive to parameter variation, the resonant tracking should overcome this issue [91].

1.12. Discussion on Parameter Variation and Maximum Power Transfer

For long-distance, loosely coupled CPT systems, the mutual capacitance is limited. In such cases, it requires increasing the plate voltage for sufficient power transfer [92]. To achieve high power transfer, both IPT and CPT systems require a compensation network. The switching frequency of the system is operated close to the resonant frequency to eliminate the reactive power. Thus, addition of an LC network enhances the system's performance. The compensation network also amplifies the AC signal to the coupling section. Misalignment is one of the important issues
in WPT. It affects the overall impedance of the system by adding the reactance to the circuit. The reactive power is minimized by tuning the switching frequency close to the resonant frequency. Several works have noticed the influence on power factor and the impedance characteristics with the variation in load resistance, coupling capacitance, and the addition of external capacitors. Reference [93] has proposed a bi-lateral LC-compensation network. It is noted that the power factor varies from 0.902 to 0.977 when the load resistance changes from 30.2 Ω to 70.8 Ω and coupling capacitance from 310 pF to 1490 pF, respectively. The optimal load of the double-sided LC-compensated CPT is analyzed in [94]. The optimal load condition only improves the AC-AC efficiency of the resonant network; DC-DC efficiency was only 78.4% with 109.3 W output power. To quantify the connection between the optimal load condition and the external capacitance, the impedance characteristics of the full-bridge inverter and rectifier are presented in [39]. To improve the efficiency of the system, a parameter design method is proposed which limits the external capacitor to ensure that the voltage stress is in a reasonable range. Recent works [95] include power transfer tracking for a range of load variations from 5 Ω to 1 k Ω, the maximum efficiency for 10 W is found to be 70%.

In addition to the previous works, the portion of this chapter focuses on optimal impedance tracking. It presents the circuit’s behavior for a range of frequencies. This method can identify the maximum power transfer tracking points for a range of frequencies. Impedance tracking is performed for various resonant inductors. The gain, bandwidth, and efficiency were observed for each resonant inductor. Additionally, the impact of the duty cycle on the CPT system was noted.
1.12.1. *Theoretical Analysis of Class-\(E^2\) Capacitive-Wireless Power Transfer System*

The mathematical model of the CPT system is designed to analyze the behavior of the circuit. For high-frequency power circuits, power losses of semiconductor components contribute significantly to the system’s efficiency. With the advancement in MOSFET technology, the operational limits of devices have expanded. This expansion has led to the use of single-active switch topologies for high-power WPT systems. Single-active switch topologies are introduced in [84], these topologies can handle power up to 1~3 kW. With minimal circuitry, these topologies are cost-effective, reliable, efficient, and easy to control. In [84], before Class-\(E^2\), a modified buck-boost converter known as a switch-inductor-inductor-diode (SLLD) has provided legitimate results. It has sensitivity issues because of the cable’s length. It adds low inductance to the resonant network, which leads to the requirement for a high switching frequency. Compared to buck-boost or SLLD, Class-\(E^2\) has shown better performance; it eliminates the noises using a resonant network. The Class-\(E^2\) converter is shown in Figure 23; it is the combination of a Class-E inverter and Class-E rectifier. It consists of a simple LC resonant network to enhance the power transfer. Analysis of the Class-\(E^2\) converter can be divided into three sections: Class-E inverter, resonant network with coupling capacitors, and Class-E rectifier.

![Class-\(E^2\) converter with compensation network.](image)

Figure 23: Class-\(E^2\) converter with compensation network.
1.12.2. *Class-E inverter and it’s ZVS Limits*

The input section consists of a single active switch Class-E inverter that functions as a DC-AC voltage converter. Shunt capacitance connected across the switch achieves the ZVS property. Frequency is one of the key factors for the calculation of the shunt capacitors; a variation of frequency will impact the ZVS property based on the limits of the capacitor. When the operating frequency is lower than the resonant frequency, the ZVS property is maintained to an extent, but the capacitor starts discharging below a resonant frequency, resulting in the loss of ZVS property. In this case, decreasing the duty cycle could help to regain the ZVS, but with low duty cycle, voltage spikes are noticed at the MOSFET, thereby increasing conduction losses. Variation of the duty cycle is critical and will be examined in the discussion section. For high frequencies, i.e., over a certain frequency, the ZVS condition is lost. The choke inductor is connected in series at the input section to handle the sudden changes in current. As DC current flows through the choke inductor \(L_c\), it acts as a short circuit, and the DC link will behave as an open circuit. Selection of the choke inductor is crucial: it can resonate with the shunt capacitor, resulting in a second resonant point. The impact of shunt capacitance on the resonant point is recorded. To operate Class-E inverter in ZVS condition, the frequency limits are presented based on the impedance region (i.e., inductive/capacitive) of the resonant network. Here, the resonant network needs to be inductive to achieve ZVS [96]. This requires the circuit to be operated at frequencies higher than \(f_1\). When the equivalent impedance of the resonant network with shunt capacitance goes to the capacitive region, the Class-E inverter loses the ZVS property for frequencies over \(f_2\). \(f_1\) and \(f_2\) are the lower and upper-frequency limits to achieve the ZVS conditions, which are calculated using (10) and (11).

\[
f_1 = \frac{1}{2\pi \sqrt{2 \cdot L_c C_1}} \tag{10}
\]
\[ C_1 = \frac{\left( \frac{C_i}{2} \right) * C_i}{\frac{C_i}{2} + C_i} \]  

(11)

\[ f_2 = \frac{1}{2\pi \sqrt{2 * L_r C_{eq}}} \]

\[ C_{eq} = \frac{C_s C_1}{C_1 + C_s} \]  

(12)

The operating frequency should be within \([f_1, f_2]\) to achieve ZVS. Equations (10), (11), and (12) conclude \(f_1 < f_2\). \(C_1\) and \(R_1\) are the series capacitance and resistance of the rectifier section; further explanation of \(C_1\) and \(R_1\) is provided in Figure 25 later in this section. The impedance of both the resonant network and the rectifier section influences the ZVS condition and the current through MOSFET. Parameters of the Class-E inverter are calculated by using (13) and (14).

The minimum choke inductance is calculated by using

\[ L_f = 2 \left( \frac{\pi^2}{4} + 1 \right) \frac{R_i}{f} \]  

(13)

The reactance of the shunt capacitor is calculated by using

\[ Z_{C_s} = \frac{1}{\omega C_s} = 0.54466R_i \]  

(14)

1.12.3. Coupling Section with Resonant Network

The resonant network is the critical part of the CPT system. It influences the power delivered to the rectifier. Achieving resonant points is one of the objectives of the CPT design. Shunt capacitance connected across the diode \((C_i)\) and switch \((C_s)\) have minimal impact on the resonant point. As mentioned earlier, the shape of the coupling plates is selected based on the
application; it determines the power transfer ability of the CPT system. From the previous chapter, circular plates have shown better performance over square plates by minimizing the edge effect. Additionally, square plates are vulnerable to circular displacement. Figure 24(a) presents the setup of the coupling plate pair. Overall, 4-plates are utilized to form two pairs as a capacitive coupling interface. The diameter of the plates is 5 cm with a thickness of 1 mm like the setup used in chapter 2. A copper foil is used as an electrode, and piezoelectric (PZT is also known as lead zirconium titanate) is used as a dielectric. Because of a high dielectric value with a 2 mm distance, coupling capacitance is measured as 40 nF. From the coupling structure shown in Figure 24, the primary and secondary side coupling capacitance \((C_C - C_M) \approx C_C\). Here, the coupling capacitance is high enough to ignore the mutual capacitance between the plate pairs. High coupling capacitance allows the Class-E\(^2\) converter to operate at a resonant frequency of 111 kHz. Additionally, low-value inductors can be considered for the design of a resonant network.

![Figure 24: Circular capacitive-coupling plates manufactured for hardware tests in lab.](image)

1.12.4. Class-E rectifier and its Equivalent Circuit

The rectifier circuit consists of a single active diode connected to a shunt capacitor for achieving ZVS on the receiving section. The low pass LC filter is connected at the output end to
eliminate AC signals to the load. The values of the LC circuit are dependent on the load resistance presented in (15) and (16) i.e., (13) and (14) from [97].

The inductance of the resonant inductor is calculated by

\[ L_r = \frac{1}{4\pi^2 f^2 C_c} \]  

(15)

The shunt capacitance of the diode for 0.5 duty cycle is

\[ C_d = \frac{1}{\pi \omega R_L} \]  

(16)

One of the key points to consider is that, with the additional compensation circuitry, the resonant point is also dependent on the other parts of the circuit such as the rectifier and the switch. To be precise, the resonant frequency depends on the shunt capacitor connected across the diode.

For analysis, the Class-E rectifier is modified into an equivalent circuit as shown in Figure 25. As the choke inductor \( L_s \) acts as a short circuit and the output capacitor \( C_{out} \) acts as an open circuit, they are ignored for the resonant section. For analysis purposes, the Class-E rectifier is modified to its equivalent circuit. Load resistance and shunt capacitor are modified to series connections [40]. \( C_i \) and \( R_i \) is calculated by using (17), (18), and (19), i.e., (4.7), (4.27), and (4.20) from [98] for the duty cycle \( D = 0.5 \).
Figure 25: Class-\(E^2\) converter with equivalent rectifier circuit.

\[
\phi = \tan^{-1} \left( \frac{1 - \cos(2\pi D)}{2\pi(1 - D) + \sin(2\pi D)} \right)
\]

\[
\frac{C_i}{C_d} = \frac{\pi}{\pi(1 - D) + \sin(2\pi D) - \frac{1}{4} \cos(2\phi) \sin(4\pi D) - \frac{1}{2} \sin(2\phi) \sin(2\pi D)^2 - 2\pi(1 - D) \sin(\phi) \sin(2\pi D - \phi)}
\]

\[
\frac{R_i}{R_L} = 2\sin(\phi)^2
\]

In the following section, the impedance equations are derived to find the optimal impedance point, and the variation in current through the choke inductor, switch, and resonant inductor is observed to identify the Maximum Powerpoint.

1.12.5. Analysis on Optimal Impedance Tracking of Class-\(E^2\) Converter

Class-\(E^2\) converter is a simple circuit with fewer components. The impact of load variation is shown in [35]. It is also important to investigate the impact of different resonant inductors on the overall impedance of the circuit, and the optimal impedance point is identified for each case.
The analysis is performed by observing the impedance curves for a range of frequencies. The impedance characteristics also determine the current flow through the choke inductor, MOSFET, and the primary resonant inductor for a range of frequencies. This method contributes to determining the Maximum PowerPoint. The output of the CPT system is dependent on the resonant network as it controls the power flow to the receiving section. For this work, the coupling plate pairs are placed far from each other, so the mutual capacitance is ignored. From Figure 25 and equation (20), the equivalent capacitance of the resonant network is the series combination of coupling plates and equivalent capacitor $C_i$. Resonant inductance on primary and secondary sections is considered equal resulting in a single resonant point. The work from [24] has examined two different values on primary and secondary inductors to realize multiple resonant points. Besides, the impedance of multi-resonant points will be larger than the single-resonant point, which limits the maximum current flow towards the rectifier section. Circuit parameters are listed in Table 7.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Remark</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_{in}$</td>
<td>DC link/Input capacitor</td>
<td>100 nF</td>
</tr>
<tr>
<td>$L_f$</td>
<td>Choke inductor</td>
<td>18 $\mu$H</td>
</tr>
<tr>
<td>$C_s$</td>
<td>Shunt capacitor at switch</td>
<td>44 nF</td>
</tr>
<tr>
<td>$C_c$</td>
<td>Coupling capacitor</td>
<td>40 nF</td>
</tr>
<tr>
<td>$L_r$</td>
<td>Resonant inductor</td>
<td>18 $\mu$H/50 $\mu$H/100 $\mu$H</td>
</tr>
<tr>
<td>$C_d$</td>
<td>Shunt capacitor at diode</td>
<td>44 nF</td>
</tr>
<tr>
<td>$L_s$</td>
<td>Secondary inductor</td>
<td>18 $\mu$H</td>
</tr>
<tr>
<td>$C_{out}$</td>
<td>Output Capacitor</td>
<td>10 $\mu$F</td>
</tr>
<tr>
<td>$R_L$</td>
<td>Load resistance</td>
<td>2.2 $\Omega$</td>
</tr>
</tbody>
</table>
\[ C_{eq} = \frac{2}{C_c} + \frac{1}{C_l} \]  \hspace{1cm} (20)

\[ Z_r = R_i + j \left[ 2 \cdot \omega L_r - \frac{1}{\omega C_{eq}} \right] \]  \hspace{1cm} (21)

\[ |Z_r| = \sqrt{R_i^2 + \left( 2 \cdot \omega L_r - \frac{1}{\omega C_{eq}} \right)^2} \]  \hspace{1cm} (22)

\[ |Z_{in}| = R_{Lf} + \frac{|Z_r|}{|Z_r| + \frac{1}{\omega C_s}} \]  \hspace{1cm} (23)

The input impedance \( Z_{in} \) is derived in terms of resonant network and shunt capacitor presented in (23). The plots are presented considering the input voltage as 25 V for both theoretical and simulation analysis. As discussed earlier, the choke inductor (\( L_i \)) does not influence the resonant point for high inductance values. However, for low inductances, the choke inductor resonates with the shunt capacitor at the switch, resulting in the second resonant point. This is proved by performing hardware tests. However, from (23), the DC resistance/equivalent series resistance (ESR) of the choke inductor has an influence on overall input impedance. Different resonant inductors (\( L_r \)) are considered to investigate the impact of the resonant network on the input impedance \( Z_{in} \). The remaining parameters of the Class-E\(^2\) converter remain the same for each case. Using (23), the relation between the input impedance \( Z_{in} \) with respect to frequency is realized in Figure 26 (a). The impedance of the resonant network (\( Z_r \)) for different values of the inductor (i.e., \( L_r = 18 \mu H/50 \mu H/100 \mu H \)) can be tracked from Figure 26 (b). From Figure 26, the input impedance adjacent to the resonant point is measured to be the least, the frequency at which maximum power is transferred. It is defined as an optimal resonant point. For each case, i.e., for various resonant inductor (\( L_r \)) values, the optimal resonant point is observed in Figure 26 (a).
The theoretical results were compared with the simulation using an impedance analyzer from the Simulink. The impedance analyzer is connected across the switch to measure the input impedance for all cases, i.e., \( L_r = 18 \mu H/50 \mu H/100 \mu H \) presented in Figure 27. Comparing Figures 26 and 27, the input impedance graph generated from Simulink closely matches the theoretical results. The optimal resonant point from the simulation matches with theoretical equations. It is important to mention that simulations are performed considering high secondary inductance (\( L_s = 1 \text{ mH} \)), which minimizes the multi-resonant points. The input impedance and resonant network impedance drop significantly at the resonant frequency for each case. The variation in phase angle concerning the frequency shows whether the input impedance is inductive or capacitive. The impedance increases with respect to frequency when the phase angle is positive and decreases when the phase angle is negative; this leads to the input impedance being capacitive at a higher
frequency. The impedance spike adjacent to the resonant frequency in Figure 27 is due to the diode and components at the receiving section.
Figure 27: Simulink plots of input impedance w.r.t frequency (a) $L_r = 100 \ \mu H$, (b) $L_r = 50 \ \mu H$, and (c) $L_r = 18 \ \mu H$.

\[ I_{DC} = \frac{V_{DC}}{Z_{in}} \]  \hspace{1cm} (24)

\[ I_r = \frac{I_{DC}Z_r}{Z_r + Z_{CS}} \]  \hspace{1cm} (25)

\[ I_{CS} = I_{DC} - I_r \]  \hspace{1cm} (26)

From Figure 28, input current ($I_{DC}$), switching current ($I_{CS}$), and resonant current ($I_r$) can be tracked with respect to the frequency. The plot is realized for resonant inductor 50 $\mu H$. Clearly, at a resonant frequency, the input current ($I_{DC}$) is at maximum due to the low input impedance. The maximum current passed to the secondary section results in maximum power across the load. From the comparison of $Z_{in}$ with $I_{DC}$, and $Z_r$ with $I_r$, current curves are inverted versions of the
impedance curves. As mentioned above, for frequencies over the resonant point, the input impedance drops significantly, but the resonant network’s impedance increases as the operating frequency moves away from the resonant point resulting in low output power delivery. In this case, most of the current is passed to the switching section which is shown in Figure 28.

Figure 28: Variation of current through input side, MOSFET, and resonant network with respect to the frequency.

Similar to impedance, current curves are plotted for different resonant inductor values. Here, the input current (I_{DC1}), current through the resonant inductor (I_{r1}), and the current through a switch (I_{Cs1}) are for \(L_r = 100 \, \mu\text{H}\). Similarly, I_{DC2}, I_{r2}, and I_{Cs2} are for \(L_r = 50 \, \mu\text{H}\), and I_{DC3}, I_{r3}, and I_{Cs3} are for \(L_r = 18 \, \mu\text{H}\). The amplitude of the current at a resonant point for each case can be seen in Figure 29 (a), (b) and (c). With different resonant inductor values, the resonant point varies, and the amplitude of the current at the resonant point is also different for each case. It can be noted that, for the high resonant inductor value, the curve is much sharper compared to the CPT with a low resonant inductor value. In these cases, the impedance sensitivity is high for frequency...
variations resulting in significant changes in the load. For high frequencies, the switching current ($I_{C_s}$) curve remains almost the same for different resonant inductor values.

![Graphs showing current variations](image)

Figure 29: Variation of current through input side, MOSFET, and resonant network for different resonant inductor values.

The prototype board of a Class-E² converter was designed; it can handle power up to 100 W. In Figure 30 (a), the Class-E inverter with a resonant inductor ($L_r$) can be seen. Figure 30 (b) consists of a Class-E rectifier with the second resonant inductor ($L_r$), and Figure 30 (c) shows the complete circuit with round capacitive coupling plates. Theoretical and simulation results are validated by performing hardware tests for 5 V input. The input current and output voltage readings were noted simultaneously. Tests were performed for resonant inductors $L_r = 50 \mu H$ and $100 \mu H$, and 2.2 Ω load was connected at the output end.
Hardware tests were conducted for two cases, i.e., for $L_r = 50 \mu\text{H}$ and $L_r = 100 \mu\text{H}$, and their respective resonant frequency points were defined as $F_{R1}$ and $F_{R2}$. Figure 31 realizes the input current ($I_{\text{DC}}$) and output voltage ($V_{\text{out}}$) of the hardware tests. Due to the capacitive nature of the circuit, the input impedance drops significantly for high frequencies, and it can be observed from Figure 31 (a) that the input current increases with the operating frequencies, i.e., far from the resonant point. Even though output voltage is high at several frequency points, maximum
efficiency (87%) is achieved only at the optimal impedance point, i.e., close to the resonant point. However, it can be noticed that output voltage increases at a non-resonant frequency \( (F_{R11}) = 200 \text{ kHz} \) for resonant inductor \( L_r = 50 \mu H \). It occurs due to the low choke inductor (18 \( \mu H \)), which resonates with the shunt capacitor. Similarly, the output voltage increases slightly for \( L_r = 100 \mu H \), at non-resonant frequency \( (F_{R22}) = 90 \text{ kHz} \). For high frequencies, the output voltage decreases gradually to zero due to high resonant impedance \( (Z_r) \). At low frequencies, the input impedance is expected to be inductive, which leads to a high input current. When comparing both cases, the efficiency for \( L_r = 50 \mu H \) is measured to be 87% at the optimal resonant point and 79% for \( L_r = 100 \mu H \). Further, low resonant inductances could be considered, but this would require the circuit to operate at a high operating frequency.

Figure 31: Hardware results of Class E\(^2\) converter (a) Input current vs. Frequency, (b) Output Voltage vs. Frequency.
1.12.6. Discussion on the Impact of Duty and Frequency on ZVS Property

The variation in input current and output voltage is observed for different duty cycles. The relationship between the output voltage with respect to the duty cycle is majorly dependent on the shunt capacitor connected across the switch. As the Class-E converter has the ZVS configuration, it is important to identify the impact of change in frequency or duty cycle to maintain ZVS. The shunt capacitor affects the Class-E\(^2\) converter in two ways: 1) the impedance that controls the current flow to the coupling plates, 2) the ZVS configuration. Considering \(C_{\text{Smax}}\) is the maximum shunt capacitance to achieve ZVS, and \(C_{\text{S}}\) is the shunt capacitance. When \(C_{\text{S}} < C_{\text{Smax}}\), the shunt capacitor voltage reaches zero, comparatively earlier to the closed switch as shown in Figure 32 (a). Figure 32 (b) shows the plot when \(C_{\text{S}} = C_{\text{Smax}}\). For both cases, below the resonant frequency (\(f_2\) from (2)) the capacitor gets discharged, which leads to the flow of current to the ground. This results in low output voltage. For frequencies over \(f_1\) (from (1)) the ZVS condition is lost, which also results in low output voltage as the operating frequency will be away from the resonant frequency. When it comes to the variation of a duty cycle for both cases, each one has different conditions. For \(C_{\text{S}} < C_{\text{Smax}}\), the shunt capacitor starts discharging for the lower duty cycle (i.e., below 0.5). For the duty cycle over 0.5, the output voltage remains constant until the ZVS condition is lost. The ZVS condition for the high duty cycle is dependent on the value of the shunt capacitor. For \(C_{\text{S}} = C_{\text{Smax}}\), ZVS is lost for duty cycle over 0.5. For a low duty cycle, the output remains constant until the shunt capacitor starts discharging.
Figure 32: Waveform across the shunt capacitor (a) ZVS (b) early ZVS.

Hardware experiments were performed for duty cycles 0.3, 0.5, and 0.7 and frequency in a range [90–180 kHz]. The following readings are under the condition $C_s = C_{s_{\text{max}}}$. Figure 33 illustrates the voltage reading across shunt capacitor for various frequencies. From Figure 33 (c), for high frequencies with a duty cycle of 0.5, the ZVS condition is lost. ZVS is maintained for low frequencies until the shunt capacitor starts discharging as shown in Figure 33 (a). Figure 34 presents the voltage waveforms for the various duty cycles. From Figure 34 (a), when switching frequency 130 kHz for duty cycle 0.3 the capacitor discharges, which results in low output voltage. For a high duty cycle, the Class-E inverter loses its ZVS condition as shown in Figure 34 (c).
Figure 33: Waveform across the shunt capacitor for frequency (a) 90 kHz, (b) 130 kHz, (c) 180 kHz.

Figure 34: Waveform across the shunt capacitor for duty cycle (a) D = 0.3, (b) D = 0.5, (c) D = 0.7.
The input impedance with respect to the duty cycle is presented by using equations (12, 17) [98]. With the increase in duty cycle from 0.5, the input impedance drops significantly resulting in a high current. Although it loses the ZVS condition, it does contribute to increasing the output voltage. It is proved by means of performing hardware experiments. Table 8 presents the input current, output voltage for various duty cycles, and frequencies. From Table 8, it can be observed that the output voltage is high when the switching frequency is close to the resonant frequency, i.e., 100 kHz, but ZVS is achieved without discharging the shunt capacitor for operating frequencies over the resonant frequency. For switching frequency 130 kHz, the input current and output voltage can be noticed. The output voltage remains the same for duty cycles 0.3 and 0.5 and increases for 0.7. There is a 1% increase in output voltage for a 50% increase in input currents, which concludes with the drop in efficiency for a high duty cycle.

Table 8: Hardware results for various duty cycle and frequencies of Class-E².

<table>
<thead>
<tr>
<th>Frequency (kHz)</th>
<th>Input Current</th>
<th>Output Voltage</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>D = 0.3</td>
<td>D = 0.5</td>
</tr>
<tr>
<td>90</td>
<td>0.3</td>
<td>0.29</td>
</tr>
<tr>
<td>100</td>
<td>0.38</td>
<td>0.36</td>
</tr>
<tr>
<td>110</td>
<td>0.31</td>
<td>0.46</td>
</tr>
<tr>
<td>120</td>
<td>0.25</td>
<td>0.42</td>
</tr>
<tr>
<td>130</td>
<td>0.17</td>
<td>0.17</td>
</tr>
<tr>
<td>140</td>
<td>0.11</td>
<td>0.1</td>
</tr>
<tr>
<td>150</td>
<td>0.07</td>
<td>0.12</td>
</tr>
</tbody>
</table>
1.13. **Optimization of class-E rectifier using finite secondary inductance**

Misalignment influences several factors like resonant frequency, zero-switching-voltage (ZVS), and the distribution of electric fields Class-E based CPT system [91]. ZVS eliminates the switching losses, to achieve ZVS appropriate shunt capacitor needs to be identified. The shunt capacitance is dependent on the phase angle; and the phase angle is related to coupling capacitance, resonant inductor, and load resistance. [99] proposes a new Class-E converter with an additional shunt capacitor to improve the voltage level at the output end. It is noticed that, with the increase in shunt capacitance the voltage level increases, and the ZVS is maintaining a range of shunt capacitor values. To maintain the ZVS for a wide range of plate distances, [100] presents the Class-E\(^2\) converter: a combination of Class-E inverter and ZCS based Class-E rectifier. This topology absorbs the secondary-side compensation resonance inductance into the equivalent inductance of the Class-E ZCS rectifier. This design maintains the power level with maximum efficiency by keeping ZVS on the inverter and ZCS on the rectifier side. In [91] the impact of varying coupling coefficients on zero-voltage-switching (ZVS) and zero-current-switching (ZDS) in a Class-E-based CPT system is noted. This work identifies the optimal coupling coefficient to achieve ZVS and ZDS at 100kHz to transfer 83.5W power with an efficiency of 92.5%. The mentioned works successfully demonstrate the ZVS property under various conditions, but the consideration of secondary finite inductance of Class-E rectifier is ignored for all cases. The Class-E inverter circuit design will determine the input impedance, specifically primary inductance and shunt capacitor across the switch. The impedance across the inverter controls the input current. The Class-E rectifier consists of minimal circuitry, and a shunt capacitor across the diode included in the resonant circuit, but the impact of finite secondary inductance needs to be identified. Several works [101]-[103] presented the work on a Class-E rectifier with finite secondary inductance. This paper
provides an analysis of several cases that determines the impact of secondary inductance on the Class-E\textsuperscript{2} based capacitive-wireless power transfer.

1.13.1. Analysis on Class-E\textsuperscript{2} converter with finite primary and secondary inductance

The Class-E converter is proven to be efficient for CPT over the other topologies [84]. In this work, the Class-E\textsuperscript{2} converter is used as a combination of Class-E inverter and Class-E rectifier shown in Figure 35. For most cases, the primary and secondary inductance of the Class-E\textsuperscript{2} is considered to be infinite, which excludes both the inductors for the analysis. In this work, several cases are presented considering different A\textsubscript{r} values (A\textsubscript{r} = f\textsubscript{r}/f). Here, f is the operating frequency, f\textsubscript{r} is the resonant frequency of L\textsubscript{S}C\textsubscript{a}.

![Figure 35: Class-E\textsuperscript{2} converter with impedance marking](image)

1.13.2. Consideration of Class-E inverter and matching network

In this work, the Class-E inverter is designed considering the finite primary inductance (L\textsubscript{f}). The shunt capacitor (C\textsubscript{S}) is calculated based on (27) to achieve the ZVS condition. In addition, the DC link (C\textsubscript{in}) is included to absorb voltage spikes at the input end. The considered Class-E inverter offers complete protection at the input end. The coupling section consists of a resonant
inductor (L_r), which acts as a compensation network. The coupling capacitors and resonant inductors form a series resonance at the coupling section.

\[
Z_{cs} = \frac{1}{\omega C_s} = 0.54466 R_l \tag{27}
\]

\[
\hat{f}_s = \frac{1}{2\pi \sqrt{L_r C_s}} \tag{28}
\]

![Figure 36: Front view and side view of circular capacitive plates](image)

1.13.3. Analysis on Class-E rectifier with finite inductance

The Class-E rectifier can be operated in both Zero-Current-Switching (ZCS) and Zero-voltage-switching (ZVS) mode. In [100], a ZCS based Class-E rectifier is considered. The inductor is used to achieve ZCS; it also operates as a compensation inductor. The mentioned topology can maintain ZVS for a wide transmission range. Here, the finite secondary inductor forms resonance with the shunt capacitor across the diode. The theoretical equations of the input impedance (Z_{rec}) and resonant frequency (\(\hat{f}_s\)) of the Class-E of the rectifier are presented below i.e., (30). Figure 37 shows the Class-E inverter in two modes: (i) diode in reverse bias, (ii) diode in the forward bias. The Z_{rec} remains the same for both modes; it only represents the direction of the current flow.
Figure 37: Equivalent circuit of Class-E rectifier (a) diode in reverse bias, (b) diode in forward bias

\[ Z_{rec} = R_{rec} + jX_{rec} \]  \hspace{1cm} (29)

\[ Z_{rec} = \frac{R_L}{(\omega C_d R_L)^2 + (\omega L_s C_d - 1)^2} + j \frac{-\omega_r C_d (R_L^2 + \frac{L_s}{C_d} (\omega L_s C_d - 1))}{(\omega C_d R_L)^2 + (\omega L_s C_d - 1)^2} \]  \hspace{1cm} (30)

The resonant condition is \( \text{Im}[Z_{rec}] = X_{rec} = 0 \); the resultant resonant frequency \( f_r \) is

\[ f_r = \frac{1}{2\pi} \sqrt{\frac{1}{L_s C_d} - \left( \frac{R_L}{L_s} \right)^2} \]  \hspace{1cm} (31)

From (31), \( \frac{1}{L_s C_d} \gg \frac{R_L}{L_s} \) which means \( \frac{R_L}{L_s} \) can be ignored for the calculation of resonant frequency for Class-E rectifier. \( f_r \) determines the resonant frequency of the rectifier section. The resonant point defines the reactive power to be zero. In the next section, the experimental setup is
presented. The results are noted for different secondary inductance, and the ZVS property is observed for different cases.

1.13.4. Results and discussions

The theoretical analysis discussed above presents the optimal design of the Class-E rectifier. The resonant point of the coupling section and Class-E rectifier determines the parameters that minimize the reactive power. This work identifies the difference in the output voltage for various \( A_r >1, =1, <1 \). The impact on ZVS property with the change in secondary inductance is also observed. The hardware setup is designed using the parameters from Table 9, and it is shown in Figure 38. The experimental setup consists of two PCB boards: Class-E inverter and Class-E rectifier with two pairs of circular coupling plates.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Capacitance (( C_{in} ))</td>
<td>100 nF</td>
</tr>
<tr>
<td>Primary Inductance (( L_f ))</td>
<td>9 ( \mu )H</td>
</tr>
<tr>
<td>Shunt capacitor at switch (( C_S ))</td>
<td>22 nF</td>
</tr>
<tr>
<td>Coupling capacitor (( C_C ))</td>
<td>12 nF</td>
</tr>
<tr>
<td>Resonant Inductor (( L_r ))</td>
<td>50 ( \mu )H</td>
</tr>
<tr>
<td>Shunt Capacitor at diode (( C_d ))</td>
<td>22 nF</td>
</tr>
<tr>
<td>Secondary Inductor (( L_S ))</td>
<td>9 ( \mu )H/18 ( \mu )H/100 ( \mu )H</td>
</tr>
<tr>
<td>Output Capacitor (( C_{out} ))</td>
<td>10 ( \mu )F</td>
</tr>
<tr>
<td>Load resistance (( R_L ))</td>
<td>2.2 ( \Omega )</td>
</tr>
</tbody>
</table>
The tests are performed for three cases; the secondary inductance considered are 9 μH, 18 μH, and 100 μH. Figure 39 indicates the readings of the input current and output voltage for different switching frequencies. From the results, for all three cases, the maximum power transfer occurs at the resonant point i.e., 210 kHz. By observing the input current and output voltage for all three cases, the results conclude that the input voltage is 10% less and the output voltage is 15% higher for $A_r = 1$ at the resonant frequency. The highest efficiency noted for the optimal condition i.e., $A_r = 1$ is 70.2%.
Figure 39: Hardware results of the Class-E<sup>2</sup> converter (a) Input current w.r.t switching frequency, (b) Output voltage w.r.t switching frequency

Figure 40 presents the ZVS plots for the two conditions ($A_r > 1$). From the plots, it is noted that the ZVS condition is achieved when $A_r = 1$. Based on the theoretical results, $X_{rec}$ is positive i.e., inductive when $L_s = 9 \mu H/18 \mu H$, $X_{rec}$ is negative i.e., capacitive for $L_s = 100 \mu H$. The hardware results support the theoretical equations. The ZVS of the Class-E inverter is maintained for the inductive impedance [104].
Figure 40: Voltage waveforms across the MOSFET for (a) $L_s = 18 \, \mu H$, (b) $L_s = 100 \, \mu H$

1.14. **Conclusion**

Chapter 3 presents a method to track the maximum power-point and further improves the power by optimizing the design of the Class-E rectifier. The impedance characteristics of the single-active switch are studied to track the maximum power-point. Resonant inductor ($L_r$) and coupling plates are used as a compensation network for the Class-E$^2$ power converter. The optimal impedance point for different resonant inductor is investigated. Change in the resonant inductor
does impact the amplitude of the input impedance. Theoretical equations and simulations conclude the optimal impedance point is close to the resonant point. The reactance of the circuit at the optimal impedance point is relatively low, resulting in high output voltage with maximum efficiency. The prototype PCB board is developed to verify the theoretical and simulation results. Hardware tests are performed for 2.5 W; maximum efficiency of 87% is achieved at the optimal impedance point, i.e., close to resonant point $F_{R1} = 100$ kHz for $L_r = 50 \, \mu H$ and 79% at 70 kHz for $L_r = 100 \, \mu H$. This work helps to track the maximum power point by tracking the variation of current with respect to the frequency. From the hardware tests there is an increase of 9% in efficiency for $L_r = 100 \, \mu H$, and an increase of 17% for $L_r = 50 \, \mu H$, compared to prior studies in [35]. Besides, the ZVS property was analyzed for various duty cycles and frequencies. The limits of the ZVS can be observed for both frequency and duty cycle.

Further, to enhance the power delivered to the load, the Class-E rectifier is modified to minimize the reactive power of the circuit. The output voltage and the ZVS property are influenced by the selection of the secondary inductance. Three conditions i.e., $A_r < 1$, $= 1$, $>1$ are considered. From the results, at the optimal condition i.e., $A_r = 1$, the ZVS property is maintained, and maximum efficiency of 70.2% is achieved at the resonant frequency of 210 kHz.
Chapter 4

A NOVEL 8-PLATE MULTI-RESONANT COUPLING STRUCTURE FOR DRONE CHARGING

1.15. Methods to minimize the effects of Misalignments

The misalignments are common in the WPT system. In practice, the capacitive plates attached to the drone may not align precisely with a charging pad. For this problem, several approaches were applied to minimize the effect of misalignments in the CPT system. One of the methods is by identifying the optimal coupling structure. The vertical 4-plate coupling structure is designed to minimize the effects of circulating misalignment compared with the 4-plate horizontal coupling structure [32]. The output power using round plates is over 20% for rotational misalignments over square plates, but the 4-plate vertical structure demands a high electric field to generate identical power, which leads to safety concerns around the coupling plates. Another coupling setup [104] uses a capacitively coupled matrix to absorb misalignments under variable coupling conditions, but the implementation is complicated as it requires complex control logic to adjust the plates and their respective parasitic and main capacitances. Another method uses a two-port compensation network to handle misalignment on the secondary side. The compensation network also includes a hybrid inductive and capacitive WPT system [105]. The IPT and CPT coupler in this system are employed to compensate for the misalignments, but system performance is limited to 10% misalignments. The third method utilizes a closed-loop control strategy for dynamic capacitive-wireless power transfer [106]. This approach implements the dynamic reactive compensation under misalignments. This method requires an active variable reactance rectifier, but it is recommended for high-power applications.
The effects of misalignment in the CPT system are minimized through a novel 8-plate multi-resonant Class-E power converter. This method follows a unique approach that uses a simple coupling structure with multiple series LC compensation networks connected in parallel. In [19], the impedance analysis of the Class-E converter is presented. It concludes that the optimal impedance points for each resonant inductor overlap with a resonant point for the MPPT. The proposed design utilizes different resonant inductors to generate multiple resonant points which offer benefits over the regular 4-plate coupling structure for misalignments. This chapter introduces the proposed 8-plate multi-resonant coupling and its benefits over the regular 4-plate coupling.

1.16. Implementation of multi-resonance to minimize the impact of misalignment

1.16.1. Frequency splitting in WPT systems

Identifying resonance in the circuit is key to enhancing efficiency, it minimizes the reactance and improves the performance. The basic compensation networks of the WPT include series-series (SS), series-parallel (SP), parallel-series (PS), and parallel-parallel which are applied to eliminate the reactive component at the coupling section [107]. The selection of these compensation networks relies on the coupling capacitance and the topologies. For example, the resonant power converter such as Class-E utilizes series-series/series-parallel compensation networks in most cases. The disadvantages of the compensation network include the sensitivity of the resonance due to misalignments that require controllers to maintain the output voltage [108]. An alternative approach is to connect the resonant networks in a cascade [109], which requires more components (like an inductor) and increases the complexity of the circuit.

The charging setup for the drone is shown in Figure 41. The charging pad consists of four transmitting plates placed to match the sequence of drone legs as shown. The receiving plates are
attached to the drone legs as shown in Figure 41. Here, the drone is projected to land on the charging pads using the reference points of the charging station. It forms the coupling.

As mentioned earlier, misalignment shifts the resonant points. Because of misalignment, transferred power changes from a single peak curve to a double peak curve while the driving frequency moves away from the resonant frequency. This phenomenon is known as frequency splitting. The coupled model theory is adapted in [110], frequency splitting is noticed in multiple coupling systems. Many researchers focused on optimizing the WPT system performance by minimizing the frequency splitting, as it is affecting the power transfer and efficiency [111] - [112]. As the frequency splitting is impractical to suppress in a WPT system, [111] uses a control method which makes it complex.

In our proposed method, frequency splitting is utilized to form multi-resonance through the series-series compensation network shown in Figure 42. The 8-plate multi-resonant coupling structure is the modified version of the 4-plate coupling. The 8-plate multi-resonant coupling
structure is formed by adding a branch with resonant inductors in series with a capacitive coupling pair connected in parallel to the regular 4-plate coupling structure shown in Figure 42. The new compensation inductors with capacitive coupling pair are designed to form two resonant points in addition to the resonant point formed by regular 4-plate coupling with a series-series compensation network.

![4-plate coupling equivalent circuit](image1)

![Addition of two resonant branches to 4-plate coupling](image2)

![8-plate multi-resonant coupling equivalent circuit](image3)

Figure 42: Idea of 8-plate multi-resonant coupling structure

1.16.2. *Class-\(E^2\) with 8-plate multi-resonant coupling structure*
A Class-E\textsuperscript{2} based CPT system is utilized for this work. As mentioned earlier, the advantages of Class-E include a single active switch with zero-voltage-switching (ZVS). It is applicable for high frequencies, and a series inductor forms a resonance with the coupling plates resulting in a series-series compensation network. A Class-E inverter circuit is sensitive to parameter variation. For the applied input voltage, the power flow from the primary to secondary circuit depends on the resonant inductor and coupling capacitance. To maintain ZVS for parameter variation such as coupling capacitance or load, [113] uses a control strategy and a proper shunt capacitor. Few other works [114] - [115] focused on maintaining ZVS by designing the appropriate impedance matching network. Even in this dissertation, a similar method is applied to achieve ZVS for the parameter variations.

![Diagram of Class-E\textsuperscript{2} with 8-plate multi-resonant coupling structure](image)

Figure 43: Class-E\textsuperscript{2} with 8-plate multi-resonant coupling structure

The 8-plate multi-resonant coupling structure is applied to the Class-E\textsuperscript{2} power converter shown in Figure 43. \( L_r \) is the regular resonant inductor, \( L_{r1} \) and \( L_{r2} \) are the additional resonant inductors utilized to achieve multi-resonance. The resonant inductors \( L_{r1} \) and \( L_{r2} \) are installed to operate as compensation inductors for low coupling capacitance due to misalignments. The order
of resonant inductor values in the 8-plate multi-resonant coupling structure are as follows: \( L_{r2} > L_{r1} > L_r \). Using the designed parameters, the 8-plate multi-resonant CPT system without misalignment achieves resonance at two branches. For the misalignment of 0% -20%, the additional inductors can compensate by achieving resonance in one of the new branches based on the coupling capacitance. The reactive power at the coupling section is minimized through the parallel connection of series LC branches. The proposed design can increase the power flow for all operating frequencies compared with the regular 4-plate coupling structure.

1.16.3. Impedance analysis of 8-plate multi-resonant coupling structure

[28] uses the equivalent rectifier model for the impedance analysis of the Class-E\(^2\) based CPT system. A similar method is applied to plot the impedance curve for Class-E\(^2\) converter both regular 4-plate coupling and 8-plate multi-resonant coupling structure. Figure 44 shows the equivalent circuit of the Class-E\(^2\) based 8-plate multi-resonant coupling. To compare 4-plate and 8-plate coupling, the resonant impedance equations for 8-plate multi-resonant coupling are derived for the impedance tracking with respect to the frequency.

![Figure 44: Equivalent circuit of Class-E\(^2\) based CPT](image-url)
\[ Z_{r1} = j(X_{lr1} - X_{cc}) \]  
(32)

\[ Z_r = j(X_{lr} - X_{cc}) \]  
(33)

\[ Z_{r2} = j(X_{lr2} - X_{cc}) \]  
(34)

\[ Z_{rec} = R_l - jX_{cl} \]  
(35)

where \( X_{lr1} = \omega \ast L_{r1}, X_{lr} = \omega \ast L_r, X_{lr2} = \omega \ast L_{r2}, X_{cc} = \omega \ast C_c \), and \( \omega = 2 \ast \pi \ast f \).

The resultant equation of the resonant impedance of 8-plate multi-resonant Class-E\(^2\) based CPT system with respect to the transmitting side is:

\[ Z_{res} = R_l + j \left( \frac{Z_{r1} \ast Z_r + Z_{r2} \ast Z_r}{Z_{r1} + Z_r + Z_{r2} + Z_r - X_{cl}} \right) \]  
(36)

\[ |Z_{res}| = \sqrt{R_l^2 + \left( \frac{Z_{r1} \ast Z_r + Z_{r2} \ast Z_r}{Z_{r1} + Z_r + Z_{r2} + Z_r - X_{cl}} \right)^2} \]  
(37)

It is impractical to maintain constant output for misalignments in the WPT system [24], but the impact can be minimized. This work adopts a method that does not require controllers to reduce the drop in output voltage for misalignments. Figure 45 presents the resonant impedance plots of 4-plate coupling with SS compensation network and an 8-plate multi-resonant coupling structure with and without misalignments for three cases. Each case uses different resonant inductor values shown in table 1. For the 4-plate coupling structure, due to series resonance, the impedance is minimal only at the resonant point (also the operating frequency) i.e., 470 kHz, whereas the 8-plate multi-resonant coupling structure widens the resonant frequency band based on additional resonant inductor values and provides better performance for misalignments.
Table 10: Cases based on additional resonant inductor values

<table>
<thead>
<tr>
<th>Case</th>
<th>$L_r$</th>
<th>$L_{r1}$</th>
<th>$L_{r2}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>a</td>
<td>33 μH</td>
<td>34 μH</td>
<td>36 μH</td>
</tr>
<tr>
<td>b</td>
<td>33 μH</td>
<td>35 μH</td>
<td>40 μH</td>
</tr>
<tr>
<td>c</td>
<td>33 μH</td>
<td>40 μH</td>
<td>45 μH</td>
</tr>
</tbody>
</table>

While designing the Class-E$^2$ based CPT system, the operating frequency will be the resonant frequency. Therefore, the operating frequency of the Class-E$^2$ based CPT system is 470 kHz for 4-plate and 490 for 8-plate coupling. The idea is to keep the operating frequency constant for misalignment such that additional controller setup is avoided. The performance of 4-plate and 8-plate multi-resonant coupling is evaluated for misalignments. For 4-plate coupling, the resonant point shifts to 495 kHz from the original resonant point for 10% misalignment. The shift in the resonant point increases the resonant impedance by 10 times at the operating frequency for 4-plate coupling. It affects the power flow and results in the decline of output voltage. For the misalignment, due to the wide resonance frequency band, the proposed 8-plate multi-resonant coupling resonant impedance is smaller at their respective operating frequencies compared with the 4-plate coupling. The 8-plate multi-resonant coupling provides better power flow with and without misalignments. Additionally, through the proposed coupling design, the impedance away from the resonance point (i.e., at 350 kHz or 600 kHz) is almost half of the impedance of a regular 4-plate with an SS compensation network. The following section presents the results of simulation and hardware tests.
Figure 45: Resonant impedance plots of 4-plate and 8-plate multi-resonant coupling with and without misalignment for three cases.

Impedance plots shown in Figure 45 exhibit the design procedure of the 8-plate multi-resonant coupling. The misalignment drops the coupling capacitance; the additional resonant inductors’ ($L_{r1}$, $L_{r2}$) values must be greater than the base resonant inductor ($L_r$) to compensate for the drop. Each case carries a set of different $L_{r1}$, $L_{r2}$ values to present the behavior of the 8-plate multi-resonant coupling structure. Figure 45 (a) presents the setup for three resonant inductors with
a difference of 1 μH and 3 μH in reference to \( L_r \). The impedance plots exhibit smoother curves for both 8-plate coupling with and without misalignment, but the width of the resonant frequency band is small compared to the other two cases. Comparing cases from a to c, the impedance peak within the resonance frequency band increases. As the additional resonant points are far from the original resonant frequency, case c generates a high impedance peak between the additional and original resonant points. Case b exhibits a wider resonant frequency band and a minimal impedance peak within the band. For this design, the selection of the additional resonant inductors has a key role in generating a smooth resonant impedance curve for misalignments.

1.17. **Simulation and hardware test analysis of 4-plate and 8-plate multi-resonant coupling structure**

1.17.1. **Simulation tests**

The theory is supported by performing simulation tests using Simulink. The coupling setup of the drone is shown in Figure 41. The coupling is formed when the drone lands on the charging station. As the receiving plate is positioned on top of the transmitting plate, the transmission distance will be minimal, which results in a high coupling capacitance. The four coupling pairs are positioned far from each other, which eliminates the cross and self-coupling capacitance. The equivalent model of the coupling plates will be the ideal capacitor for the simulation. This setup forms a series LC network at each resonant branch. The Simulink models of the 4-plate Class-E\(^2\) and 8-plate multi-resonant Class-E\(^2\) power converters are shown in Figures 46 and 47.
The PZT plates used for the proposed 8-plate setup have a different width compared to the previous test. The measured coupling capacitance of the new coupling plates is 3.45 nF. This
results in new parameters of the Class-E\textsuperscript{2} based CPT system presented in table 10. As the Class-E\textsuperscript{2} power converter can operate at high frequencies, the resonant inductance $L_r$ is considered as 33 μH, which results in a resonant frequency of 470 kHz. The Class-E rectifier is designed to ensure the equivalent impedance of the receiving section holds in the inductive region for the operating frequency. The additional resonant inductors, $L_{r1}$ and $L_{r2}$, are considered based on 10% and 20% misalignment. The parallel connection of the resonant branches offers low impedance at the coupling section and results in an increase in power flow to the receiver side.

Table 11: Parameters of Class-E\textsuperscript{2} power converter with 4-plate, and 8-plate multi-resonant coupling

<table>
<thead>
<tr>
<th>Remarks</th>
<th>Parameters</th>
<th>Values for 4-plate</th>
<th>Values for 8-plate</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC link/Input capacitor</td>
<td>$C_{in}$</td>
<td>100 nF</td>
<td>100 nF</td>
</tr>
<tr>
<td>Choke/primary inductor</td>
<td>$L_p$</td>
<td>33 μH</td>
<td>33 μH</td>
</tr>
<tr>
<td>Shunt capacitor at MOSFET</td>
<td>$C_s$</td>
<td>1 nF</td>
<td>1 nF</td>
</tr>
<tr>
<td>Coupling capacitor</td>
<td>$C_c$</td>
<td>3.75 nF</td>
<td>3.75 nF</td>
</tr>
<tr>
<td>Resonant inductors</td>
<td>$L_r$</td>
<td>33 μH</td>
<td>33 μH</td>
</tr>
<tr>
<td></td>
<td>$L_{r1}$</td>
<td>37 μH</td>
<td></td>
</tr>
<tr>
<td></td>
<td>$L_{r2}$</td>
<td>40 μH</td>
<td></td>
</tr>
<tr>
<td>Shunt capacitor at diode</td>
<td>$C_d$</td>
<td>1 nF</td>
<td>1 nF</td>
</tr>
<tr>
<td>Secondary Inductor</td>
<td>$L_s$</td>
<td>33 μH</td>
<td>33 μH</td>
</tr>
<tr>
<td>Output capacitor</td>
<td>$C_{out}$</td>
<td>1 μF</td>
<td>1 μF</td>
</tr>
<tr>
<td>Load resistor</td>
<td>$R_L$</td>
<td>2.4 Ω</td>
<td>2.4 Ω</td>
</tr>
</tbody>
</table>
The simulations were done for a range of frequencies [350 kHz, -600 kHz] for both 4-plate and 8-plate multi-resonant coupling with ideal coupling (\(C_c = 3.7\)) and the misaligned coupling (\(C_c = 3.45\) nF). Figure 48 (a) and (b) presents the output voltage readings of the regular 4-plate and 8-plate multi-resonant Class-E\(^2\) power converter. The Class-E\(^2\) power converter with 4-plate coupling has a single peak; it varies with respect to the resonance at the coupling section. The MPPT occurs when the operating frequency is equal to the resonant frequency of the coupling branch. The impedance away from the resonant frequency increases significantly which results in low power transfer. One of the approaches to handle the misalignment is by increasing the resonant frequency band. Decreasing the quality factor can widen the resonant frequency band, but the circuit with a low-quality factor will decrease efficiency.

The 8-plate multi-resonant coupling structure overcomes this problem. As mentioned earlier, due to the parallel LC resonant branches, the impedance at the coupling section reduces significantly, improving the power flow for operating frequencies far from resonant frequency compared with the regular 4-plate coupling. It can be seen in Figure 48 (a), the 4-plate coupling with coupling capacitance 3.7 nF has a narrow resonant frequency band i.e., [465 kHz -470 kHz]. The resonant band shifts to [490 kHz -495 kHz] for a coupling capacitance of 3.45 nF. Through the proposed method of 8-plate multi-resonant coupling, the simulation results show that the resonant frequency band expands by 5 times presented in Figure 46 (b). This design approach improves the performance of the CPT system for misalignments. The selection of additional resonant inductors has a significant impact on the dip in the output voltage at a frequency between the multiple resonant points seen in Figure 46 (b).

For the 4-plate coupling without misalignment i.e., for \(C_c = 3.7\) nF, the peak output voltage of 8.6 V is noted at resonant frequency 470 kHz. With the misalignment i.e., for \(C_c = 3.45\) nF, the
output voltage at 470 kHz is reduced to 6.3 V. Due to the misalignment of a 4-plate coupling structure, the output voltage drops by 25% at the operating frequency. For 8-plate multi-resonant coupling, the peak output voltage of 8.24 V is noted at a 480 kHz frequency. For a similar misalignment, the output voltage only drops to 7.5 V at 480 kHz, resulting in less than 10% output voltage drop. This presents the benefit of the proposed coupling method. Additionally, the output voltage of the 8-plate multi-resonant is twice the output voltage of the 4-plate for the frequencies far from the resonant point shown in Figure 46. Even for large misalignments, the 8-plate multi-resonant coupling results in a performance over the 4-plate coupling.

![Figure 48: Simulation results of (a) 4-plate and (b) 8-plate multi-resonant](image-url)
1.17.2. Experimental validation

The hardware experiments are performed with the prototype boards of the Class-E inverter and Class-E rectifier with multiple resonant inductors. The new PCB boards: class-E inverter and Class-E rectifier are designed to function as both 4-plate, and 8-plate multi-resonant depending on the installation of resonant inductors at the compensation network. The PCB schematic and board of the Class-E inverter with driver circuit is presented in Figure 49. For multi-resonance, four parallel branches with different inductors are included as shown in Figure 49 (a). The orange blocks in Figure 49 (a) and 50 (a) indicate the inductors at their respective position.

Figures 49 (b) and (c) display the PCB boards with a 4-plate and 8-plate multi-resonant coupling. Silicon Carbide (SiC) MOSFET is used as the circuit is operated at high frequencies; the SCT3160KL SiC MOSFET has low turn-on and turn-off time. The IR2125 low-side driver circuit is used to strengthen the gate signal.

(d) PCB schematic of Class-E inverter
The PCB schematic and hardware boards of the Class-E rectifier are presented in Figure 50. The Class-E rectifier is a simple circuit with fewer components. A SiC Schottky diode with low voltage drop is considered for the Class-E rectifier model. A 33 μH inductor is utilized to handle the large current ripples. The secondary inductor plays a crucial role in enhancing the power flow from the receiving plates to the load discussed in chapter 3.
Figure 51 presents the capacitive-wireless charging setup for the drone. The charging station consists of four coupling plates connected to the Class-E multi resonant inverter. The receiving plates are attached to the Class-E rectifier installed on the drone for charging the 100 W battery. The coupling setup shown in Figure 51 is similar to the mobile wireless charging using Qi inductive coils. Through the reference position of the charging station, the drone can land such that the four legs of the drone are positioned on the coupling plates of the charging station. In real-time, even using the reference point, the positioning of the drone plates on the charging plates may not be precise. It results in misalignment of the capacitive couplers leading to a drop in power flow. The proposed method minimizes the impact and increases the power flow to achieve sufficient input voltage to the DC-DC converter attached in cascade with the Class-E rectifier. The white cables in Figure 51 are connected to the rectifier side, and the red cables are attached to the inverter side.
The experimental setup is shown in Figure 52. The hardware tests are performed for different operating frequencies with an input of 20 V. The input impedance varies with respect to the operating frequency resulting in the variation of input power. The maximum output power of 20.8 W is generated at 470 kHz frequency using Class-E$^2$ 8-plate multi-resonant coupling structure.
Figures 53 (a) and (b) display the test results of 4-plate and 8-plate multi-resonant coupling without and with 10% misalignment. The coupling capacitance of new PZT plates is noted as 3.7 nF. For misalignment, coupling capacitance is noted as 3.45 nF. Hardware results of 4-plate coupling displayed in Figure 53 (a) conclude that the 4-plate has a narrow resonant frequency band [460 - 465 kHz]. The maximum output voltage of Class-E$^2$ through 4-plate coupling is noted as 6.15 V. For the misalignment, the resonant frequency band shifts to [490-495 kHz]. Through the proposed 8-plate multi-resonant coupling, Figure 53 (b) displays a wider resonant frequency band [455-485 kHz], where the output voltage drop within the resonant frequency band is less than 10% of the peak voltage 7.05 V. Through the 8-plate multi-resonant coupling, for 10% misalignment, the resonant frequency band shifts to [475-505 kHz]. Due to the wider resonant frequency band, the 8-plate multi-resonant coupling has better performance over the 4-plate.

The output voltage at the operating frequency of 460 kHz for 4-plate coupling without misalignment is noted as 6.15 V. At the same operating frequency, the output voltage with 10% misalignment is noted as 3.92 V. Through the 4-plate coupling, the output voltage drops by 37% for 10% misalignment. On the other side, using an 8-plate coupling, the output voltage at the operating frequency is noted as 7.05 V, whereas for 10% misalignment it is noted as 5.5 V. A 22% voltage drop is noticed because of 10% misalignment using 8-plate multi-resonant coupling. The proposed method minimizes the impact of the misalignment by 15% compared with the 4-plate coupling. Additionally, the 8-plate output voltage is double the 4-plate at an operating frequency far from the resonant frequency, meaning it provides better performance for larger misalignments compared with 4-plate coupling.
Figure 54 presents the waveform plots without and with 10% misalignment from the hardware tests. From the plots it can be noticed that the ZVS is maintained for the ideal case i.e., when there is not misalignment. As the Class-E is sensitive to parameter variation, the ZVS condition is lost for misalignments at the same operating frequency, but the ZVS condition is
maintained when the operating frequency is matched to resonant frequency for 10% misalignment as shown in Figure 53 (c)

![Waveform images](image.png)

(a) Without misalignment  
(b) with 10% misalignment  
(c) Shifting operating frequency to match resonant frequency for 10% misalignment

Figure 54: Waveform of the Class-E$^2$ with 8-plate multi-resonant coupling (Blue: Gate signal, Purple: Voltage across MOSFET, Orange: Voltage across diode)

Figure 55 displays the efficiency tracking of 4-plate and 8-plate multi-resonant CPT with and without misalignment. At their respective resonant frequency, the efficiency is expected to be high for both 4-plate and 8-plate multi-resonant coupling. The efficiency of the 4-plate Class-E$^2$ CPT system is measured as 70%, whereas 8-plate multi-resonant coupling is noted as 88.5% at
460 kHz. The designed multi-resonant coupling enhances efficiency by 19.8% for the corresponding components with 4-plate coupling. Additionally, the efficiency of 8-plate remains over 60% for a range of frequencies. The parallel connection of the series LC resonant network minimizes the impedance of the coupling section, which improves the power flow.

(a) 4-plate coupling

(b) 8-plate multi-resonant coupling

Figure 55: Output voltage of Class-E² CPT with and without misalignments
1.18. Conclusion

This chapter introduces a novel 8-plate multi-resonant coupling for a Class-E² based CPT system. The advantages of the proposed model are presented by comparing the performance results with a regular 4-plate coupling. The proposed model increases the overall efficiency by 18.5% compared with the 4-plate coupling. The 8-plate multi-resonant coupling expands the resonant frequency band, and it enhances the performances for misalignments. The hardware results present a 14% improvement of output voltage for misalignments compared with 4-plate coupling. The 8-plate multi-resonant coupling is a parallel connection of four LC networks that suppresses the equivalent reactance. Because of low reactance at the compensation network, even for large misalignment, the proposed model displays a 40% rise in output voltage.
Chapter 5

SUMMARY AND CONCLUSION

1.19. Summary and conclusion

Capacitive-wireless power transfer gained attention in the research field because of its safe, reliable, cost-efficient, and compact coupling nature. The coupling section has a significant impact on the power transfer, and misalignment is one of the common issues in the wireless power transfer system. Misalignment minimizes the coupling capacitance and adds reactive power to the circuit. A novel 8-plate multi-resonant coupling is developed to overcome the effects of misalignment. The proposed design is suitable for a 100 Ah drone battery. The coupling is achieved when the drone lands on the charging pad. With a short transmission range, the CPT system is operated in kHz frequency for the coupling capacitance in nano-farads. The focus of this research is to utilize single-active switch topologies to achieve high efficiency.

Through simulations, PZT (Lead Zirconate Titanate (piezoelectric ceramic material)) is identified as an optimal dielectric material with satisfactory breakdown voltage and cost. The impact of coupling plate shape was studied. The hardware results indicate that the circular plates perform better over the square plates because of low edge effects. Additionally, a drop in coupling capacitance for circular plates is 13% lower compared with the square plates for varying frequencies. The resonance is the key for the maximum power transfer, and further work was focused on the identification of resonant points in the CPT system. In a single-active switch modified buck-boost converter (SLLD), a series and parallel resonance is identified. These resonant points are dependent on the primary and secondary inductor of the SLLD. To separate the series and parallel resonance, the secondary inductor values are lowered to a shift resonant point. After the separation of resonant points, the performance of the SLLD for frequency resonant
frequency (FRF) and variable resonant frequency (VRF) indicates that the VRFs increase the output voltage by 14% compared to FRF for varying coupling capacitance.

Further work focuses on the Class-E\textsuperscript{2} resonance power converter, as the SLLD performance is limited to high coupling capacitance and operating frequencies. The impedance plots present the maximum PowerPoint in the CPT for misalignments. The optimal impedance point for each resonant inductor is investigated; the effect on the magnitude of the input impedance for different resonant inductors is presented. Theoretical equations and simulations conclude that the optimal impedance point is adjacent to the resonant point. The hardware results present an 8% increase in efficiency for 50 μH resonant inductors compared with 100 μH for low-power tests. The reactance in the circuit at the optimal impedance point is relatively low which results in high output voltage with maximum efficiency. The dependency of the secondary inductor for power delivered to the load was studied. The Class-E rectifier is modified to minimize the reactive power of the circuit. The three conditions $Ar < 1$, $= 1$, $>1$ are considered to illustrate the influence of the secondary inductor on the output voltage and the ZVS property of the 4-plate Class-E\textsuperscript{2} based CPT system. From the results, at the optimal condition i.e., $Ar = 1$, the ZVS property is maintained, and maximum efficiency of 70.2% is achieved at the resonant frequency of 210 kHz which is 17% higher compared to other conditions.

The simulation and hardware tests were done to analyze the performance of the 8-plate multi-resonant coupling for the misalignments. The proposed coupling design enhances efficiency to 88.5% using Class-E\textsuperscript{2} based CPT for a 20.8 W hardware test. Through the multi-resonance, the resonance frequency band of the Class-E\textsuperscript{2} based CPT system is increased from 5 kHz to 30 kHz. The proposed 8-plate multi-resonant coupling improves the output voltage of the Class-E\textsuperscript{2} based CPT system by 15% compared with the regular 4-plate coupling for 10% misalignments. Even at
ideal coupling i.e., without misalignment, 8-plate multi-resonance coupling efficiency is 18.3% higher than regular 4-plate coupling. Because of the parallel connection of series-series compensation, the reactance of the coupling section is suppressed. This results in better performance of the 8-plate multi-resonant coupling for large misalignments of 40% compared with 4-plate coupling.

1.20. **Recommended future work**

The 8-plate multi-resonant presents promising results using Class-\(E^2\) based CPT. This work provides a scope to work on multi-resonant coupling through several power transfer methods. In the proposed topology, three resonant points were generated through the addition of resonant inductors. For the WPT system with many misalignments, future work can focus on increasing the number of resonant points to minimize the output voltage drop.

The Class-E inverter ZVS is achieved for ideal coupling but loses ZVS at operating frequency for misalignments. Future work can also focus on extending ZVS limits for misalignments through multi-resonant coupling. A possible approach to extend the ZVS limits is to reassign the appropriate secondary inductor that sets the equivalent circuit close to the inductive region.

In this work the coupling plate pairs are positioned far from each other, so the self and mutual capacitance are neglected. The coupling with adjacent placement of capacitive plate pairs adds the self and mutual capacitance which will affect the multi-resonant coupling design. This can be another area of future investigation.
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PUBLICATIONS

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