An Investigation on the Characteristics of Conical Coils for Wireless Power Transfer Systems

by

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Abstract

Historically, there have been many efforts to transfer power over the air medium and eliminate the necessity of conducting wires. Recent success stories on this direction has empowered unforeseen applications in many fields ranging from biomedical, wearable, food industry, and consumer products to electric vehicles.

In this work, conical coil structure is presented as a viable solution to improve the efficiency of magnetically coupled wireless power transfer (WPT) systems. It is shown, both qualitatively and quantitatively, that the use of conical inductors in place of traditional planar coils increases the self-resonance frequency of transmitter resonator while maintaining the flux linkage to the receiver side. Electromagnetic (EM) and circuit simulations predict an efficiency increase using conical coils in wireless power transfer link. The measurement results of a prototype 3-coil structure built based on the conical structure confirm the validity of simulation results.

The analysis, design, and characterization comparison of WPT systems that use both planar and conical coils is also presented. A power transfer efficiency of up to 53.9% is achieved for a 4-coil WPT system employing conical coil resonators that are 50 cm apart (which translates into the separation distance of \( \sim1.5\times \) the diameter of the resonators). In contrast, the same system using planar resonators achieves a power transfer efficiency of 32.8%. Thus, employing conical coils improves the power transfer efficiency by a factor of up to 1.6×.

Finally, an adaptive control mechanism to improve the efficiency of magnetically-coupled resonators (MCRs) is presented. To minimize the degradation in power transfer efficiency, the proposed system dynamically adjusts the capacitance of MCRs as the distance between the transmitter (TX) and receiver (RX) coils changes. The control unit operates in a self-sufficient manner through rectifying a portion of the AC signal present on TX and RX coils. A proof-of-concept circuit operating at 13.56 MHz is designed in a 0.13 \( \mu \)m CMOS technology and the simulation results confirm the validity of the proposed scheme.
Preface

I, Parinaz Hadadtehrani, am the principle contributor of all chapters. This manuscript would not have been possible with strong support of my research supervisor Professor Shahriar Mirabbasi. I would like express my deepest gratitude to Dr. Mirabbasi. I would like also to thank Dr. Reza Molavi for his ongoing support and guidance during this project. Dr. Pouya Kamalinejad and Soroosh Dehghaniy also helped with simulations and measurements of this research work.

Some of the chapters in this thesis are written based on the following conference material.

Conference papers:


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<td>A4WP</td>
<td>Alliance For Wireless Power</td>
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<tr>
<td>AC</td>
<td>Alternating Current</td>
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<tr>
<td>ADC</td>
<td>Analog To Digital Converter</td>
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<td>DC</td>
<td>Direct Current</td>
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<td>EV</td>
<td>Electric Vehicles</td>
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<td>EMF</td>
<td>Electro Motive Force</td>
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<td>EM</td>
<td>Electromagnetic</td>
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<td>FCC</td>
<td>Federal Communication Commission</td>
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<td>HFSS</td>
<td>High Frequency Simulation System</td>
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<td>IPT</td>
<td>Inductive Power Transfer</td>
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<td>MRI</td>
<td>Magnetic Resonance Imaging</td>
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<td>MCR</td>
<td>Magnetically Coupled Resonators</td>
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<td>MCU</td>
<td>Microcontroller Unit</td>
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<td>NFC</td>
<td>Near-field Communications</td>
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<td>PDL</td>
<td>power delivered to the load</td>
</tr>
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<td>PDI</td>
<td>Power Distance Index</td>
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<td>PMA</td>
<td>Power Matters Alliance</td>
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<td>PTE</td>
<td>Power Transfer Efficiency</td>
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<td>RFID</td>
<td>Radio Frequency Identifications</td>
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<tr>
<td>RX</td>
<td>Receiver</td>
</tr>
<tr>
<td>SPS</td>
<td>Solar Power Satellite</td>
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<tr>
<td>TX</td>
<td>Transmitter</td>
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<td>WPC</td>
<td>Wireless Power Consortium</td>
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<td>WPT</td>
<td>Wireless Power Transfer</td>
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Chapter 1  Introduction

Energy transfer from the generator to the load is facilitated by the propagation of electric ($E$) and magnetic ($H$) field in space surrounding the wire [1][2]. This statement is implied from the Poynting theory, which represents the electromagnetic power per unit area as $S = E \times H$, an expression first developed by John Henry Poynting and Oliver Heaviside [3][4]. Wireless energy transfer is the flow of energy beyond the boundaries of conductors and into the air medium. The transfer of data and information over air medium using antenna has been in use for several decades [5, 6]. However, due to excessive loss, the transfer of energy through the air medium was buried for almost a century after its introduction by N. Tesla in 1891.

The idea of wireless power transfer (WPT) employing resonance effects was brought to life by Soljacic et al. team in 2007, when their work demonstrated that a light bulb could be wirelessly lit over a six feet distance [7]. WPT systems have recently attracted many researchers and scientist due to its many applications in a variety of fields such as wearables and implantable devices, electric vehicles (EVs) and charging applications. In near future, most devices joining Internet-of-Things (IoT) will be powered through wireless applications [8].

1.1  Historical Overview of Wireless Energy Transfer

The interconnectivity of electric and magnetic fields was first discovered by Hans Oersted in the beginning of 19th century when he demonstrated the deflection of compass needle due a neighboring electric current [5]. In 1823 André Ampère showed that electric current in a coil produces a magnetic field in its vicinity of the coil due to electromagnetic induction. Shortly after in 1831, Michael Faraday showed how this energy is transferred back from magnetic field to current, or an electric field [5]. It was in 1873 that William Maxwell published his theoretical work on interchangeable characteristics of electric and magnetic fields [5]. Electric and magnetic fields have the capability of conducting through air, water and earth,
therefore by using any of these mediums wire can be eliminated and system would be called “wireless” [6].

One of the early attempt in this domain was made by Morse. Figure 1(a) demonstrates Morse’s setup composed of two long wires placed along the two sides of river. Wire on one side was excited by batteries and four plates (see Figure 1a) were placed inside the water to transfer electricity between plates and through water. However, this system suffered from significant loss due to the diffusion in the water and very small current was detected using huge batteries [9].

![Figure 1a) Morse's system of telegraphing across canal [6] b) Preece system of two wire loops TX and RX [6]](image)

Another attempt using current coils and harvesting the associated magnetic field was made by William Preece in 1892 across Bristol Channel at about the same time as Morse did his experiment. As shown in Figure 1b) two very large and lengthy loops of wire were set up as transmitter and receiver, and he managed to extract tiny currents induced on the receiver loop placed 5 km away [10]. In between 1880 -1890 Edison conducted several experiments transmitting power from a distant tower and detecting electrical signals on a sailing ship as well as a moving train [11].
Experiments carried between 1857-1894 by Heinrich Hertz had a profound effect in paving the way to our today’s capability to transfer information through the air [10]. One of his experimental setups as shown in Figure 2, used a transmitter (right) with a spark gap and two metal plates AA’ behaving similar to a capacitor. Transmitter sparks would produce high frequency waves in air which could, in turn, generate sparks in the ring-shaped receivers across the room (left) without any physical connection to it [12]. Heinrich Hertz’s pioneering work on the generation and propagation of electromagnetic waves then solidified electromagnetic theory proposed by Maxwell.

A breakthrough in wireless communication came about by Nicola Tesla’s experiments on the verge of 20th century. The use of alternating current enabled Tesla to explore high frequency and high-voltage transfer techniques for wireless transmission. He proposed that transferring energy through air was possible using oscillating electromagnetic waves radiating outward from the transmitter. One of his important inventions called Tesla Coils (1891), used transmitter and receiver coils operating at their resonance frequency, a configuration more recently referred to as magnetically-coupled resonators (MCR) [13]. In MCR systems, energy is transferred between the two coils with little loss, and each coil has exact equal amount of inductance (magnetic field) and capacitance (electric field), to ensure maximum power transfer subject to ohmic loss.
Several years later, another major step in using WPT was delivered by W.C Brown who developed an antenna that was capable of receiving and rectifying microwave signals, a device he called “rectenna” [14]. In 1975 Brown demonstrated the highest amount of microwave power transfer (MPT) till that date. The transfer of power from a parabolic antenna of 26 m diameter to a rectenna array of 3.5 x 7.2, placed one mile away, achieved an 82.5% rectified efficiency [15]. Brown's work was the basis for the first solar power satellite (SPS) system [16].

Near-field and far-fields refer to regions of space surrounding EM transmitters, e.g. coils or antenna, as characterized by the properties of electric and magnetic fields. This regimes have been defined in alternative ways depending on the application and the required accuracy [17]. In far-field regime, electric and magnetic waves are correlated and are typically presented as planar electromagnetic wave [5]. However, in near-field electric and magnetic field act independently and have their own characteristics. Figure 3 shows different regions and their dependencies on distance $r$.

<table>
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<tr>
<th>THREE REGION MODEL</th>
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<tr>
<td>$\frac{1}{r}$</td>
<td>$\frac{1}{r^2}$</td>
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<tr>
<td>FAR FIELD</td>
<td>FRAUNHOFER ZONE</td>
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<td>FRAUNHOFER ZONE</td>
<td>NEAR FIELD</td>
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<td>NEAR FIELD</td>
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<td>FAR FIELD</td>
<td>INDIRECT FIELD</td>
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<td>QUASI-STATIONARY</td>
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*Figure 3 Two- and three-region models describing the surroundings of an electromagnetic source [17]*

Abovementioned MPT and SPS systems are example applications of WPT energy transfer in far-field regime. In 1996 Zierhofer et al. [18] presented their work showing the impact of device geometry on the amount of coupling between magnetically coupled coils. In this work authors showed how the spreading of the coil winding along their radii (for planar or spiral coil) improves the coupling coefficient as compared to concentrating them on outside region of the coil.
On the other hand, Radio Frequency Identifications (RFID) technique which came to existence in 1973 is an example of near-field applications [19]. HF RFID operates in the unlicensed Industrial-Scientific-Medical (ISM) frequency band at 13.56 MHz and its transmission range is typically in the order of 10 cm to 1 m and are mainly used in the inventory control of stores [20].

In 2004, Near-field Communications (NFC) forum was founded by Nokia, Philips and Sony. The forum chose the ISM band of 13.56 MHz as the operating frequency, and built a short range (about 10 cm) data link used for contactless payments using smart phones [21].

In 2007, a research team led by Prof. Soljacic at MIT demonstrated wireless powering of 60 W light bulb over a 2 meter distance employing MCR systems [7]. In 2009 [19], Sony’s scientist developed a WPT system to power a 60 W television over a 50 cm distance with an efficiency of 60% [22].

In 2009, Philips research team published their work on general characteristics and limitations of inductive WPT systems also providing design guidelines for IPT consumer products [23]. They also investigated the effects of impedance matching for two-coil systems over different distances. It was concluded, that IPT systems are not well-suited for large transmission distances and/or when using low quality factors TX and RX coils. They also presented an inductive power pad system employing local detection which enables user to arbitrary position mobile phones on the charging.

In 2011 RamRakhyani, et al. [24] presented the design and optimization of four-coil resonance-based systems and achieved twice as much efficiency as compared to the traditional two-coil WPT systems. With coil diameters of 64 mm and 22 mm, for high-quality and low-quality coils, respectively, and over a transmission distance of 20 mm, researchers at UBC achieved 82% power transfer efficiency (PTE) operating at 700 kHz resonance frequency. Utilizing high-quality-factor four-coil this work managed to decouple the effects of source and load resistance from the transmitter and receiver coils leading to a significant efficiency improvement.
As an alternative solution the design and optimization of 3-coil WPT system is presented by Kiani et al. [25] in 2011 and high power delivered to the load (PDL) and power transfer efficiency (PTE) was achieved over a larger distance. A three-coil system, a trade-off between the traditional two-coil and the recently introduced four-coil WPT, still provides an additional degree of freedom, i.e. the extra coupling coefficient to design the required impedance transformation. The advantage of 3-coil system over four-coil is that in the latter system the load impedance is transformed to a very high reflected resistance across driver coil, in effect limiting available power and reducing PDL. At 13.56 MHz frequency and 12 cm distance this work achieved PTE of up to 35% and PDL.

In [26] the use of relay resonators to extend the transmission range of WPT systems is investigated. To also overcome the sensitivity of MCR systems to TX-RX misalignments, [25] proposes the curled placement of planar resonators to allow the guidance of magnetic flux in space for more efficient transfer of power (Figure 4).

![Figure 4 Repeater resonators placed in circular pattern [26]](image)

Today, WPT technology finds new applications in many areas at an astonishing rate. This includes areas such as automotive, household application, Internet-of-Things devices, wearable devices, remote wireless sensors, and many applications in biomedical and health care. Markets and Markets research estimated the growth of WPT market to $13.11 billion by 2020 from a value of $0.72 billion in 2014 [27].
On the commercial front, Toyota has collaborated with WiTricity (a spin company of the original MIT team) to produce WPT systems in its electric vehicles, and Qualcomm Inc. has used its patented WiPower technology in smart watches [28]. Smartphone manufactures, such as Samsung and LG have also integrated WPT technology in their cell phones starting 2013 [28].

Recently there has been many researches on effect of different geometries of coils on increasing quality factor and effectively power transfer efficiency of WPT system [29, 30, 31].

A recent review of applications, challenges and future trends in contactless non-resonance and resonance WPT system is presented in [32]. Authors introduced a power-distance index (PDI), as the product of transferred power amount and the transmission distance to enable a fair comparison among different WPT requirements and system capabilities. Figure 5 compares industry requirements in terms of PDI and efficiency.

Figure 5 PDI versus efficiency for contactless WPT application [32]

In order to standardize the implementation of WPT systems, there are a number of consortiums such as Wireless Power Consortium (WPC), Qi, Power Matters Alliance (PMA), and alliance for Wireless Power (A4WP) actively working to establish standards and protocols.
1.2 Motivation and Organization of the Thesis

In near future, cordless devices using WPT systems will be prevalent in many industries [33]. Maximizing the energy transfer between wireless links and improving their efficiencies is an attractive topic of many recent research works.

TX (RX) resonators constitute the main components of WPT systems and directly control the energy transfer mechanism. By operating both TX and RX at the same resonance frequency, system losses are minimized, hence higher transfer efficiency is achieved. An important factor that affects the performance of the WPT systems is the impedance matching and is addressed in this thesis. Impedance matching minimizes the reflection of energy waves back to the TX and ensures that the transmitted power is efficiently absorbed on the RX side.

Smart home concept is gaining more popularity each year. WPT provides a solid infrastructure to power up nearby devices and creates many more possibilities of smart devices around the house. Currently, there exists a great demand for every device that is controlled through either IoT or smart home to be powered by a WPT system which may be embedded inside the structure of house. For example, inductive coils can be installed behind the dry walls, floors or ceilings throughout the entire house and be used to power different devices in their close proximity. In order to increase the range and accessibility of WPT, typically, large size coils have to be installed. This research has focused on 50 to 75 cm distance, which would be a reasonable distance for smart home applications and is suitable for charging devices that are in the close proximity of the transmit coils.

The Focus of this research has been to increase the efficiency of WPT links both when TX – RX distance is fixed (using a new conical coil geometry discussed in Chapters 3 and 4) as well as when their separation dynamically varies over time (using an adaptive frequency tuning technique in Chapter 5). The rest of thesis is organized as follows.
In Chapter 2, the fundamental circuit and system parameters required to characterize WPT systems are discussed. Also, a brief review of state-of-the-art WPT systems configuration and their measured performances are presented.

In Chapter 3, conical coil structure is presented as an alternative geometrical solution to implement resonator coils and it is shown that a higher performance is achievable in terms of coils quality factor and self-resonant frequency.

In Chapter 4, the analysis, design and characterization of four-coil planar and conical WPT system are presented. It is experimentally shown that conical structure achieves higher power transfer efficiency when compared to its planar counterpart.

In Chapter 5, an adaptive control mechanism is presented to improve the efficiency of WPT system. The proposed system dynamically adjusts the capacitance of resonators as distance between TX and RX coils changes.

The combination of coil structures and circuit techniques proposed in chapters 3 – 5 can result in implementation of more efficient WPT systems for mid-range applications, where TX-RX separation is larger than coil geometries by a factor of 2 to 10.
Chapter 2  Review of Wireless Power Transfer

As discussed in previous chapter, in WPT systems conducting wires are replaced by medium of air and energy is transferred via propagation of EM waves. WPT systems are typically classified based on their propagation mechanism [34] as shown in Figure 6. For the first category, called electromagnetic radiation, EM waves are traveling in far-field region such as radio frequency links (RF) and optical links. Examples of far-field techniques range from the use of laser beams to operate drone aircrafts to the application of radio waves to power RFID tags [35], [36]. In spite of high attainable power efficiencies over larger distances, the major limitation of these techniques is the required mechanism to properly align the TX and RX and maintain the directionality. On the other hand, near-field techniques such as the inductive power transfer (IPT) or resonance-based wireless power systems which rely on the magneto-static effects of electromagnetic field in short-to-mid range distances constitute the second propagation technique [7]. Third classification of WPT is capacitive coupling which operates based on the principles of electric field propagation. In this chapter the principles of capacitive and inductive coupling techniques are explained followed by detailed discussion of MCR systems. Then characteristics of WPT system and their performance metrics are explained. In the final section of this chapter some of the state-of-the-art work in the field are being reviewed.

Figure 6 WPT systems Classifications
2.1 Capacitive Coupling

WPT operation using capacitive coupling [37] or electrostatic induction relies on energy transfer through the dielectric material residing between the two plates of capacitor as shown in Figure 7. Capacitive coupling operation is generally confined to the near-field region of TX and RX plates. Energy is transferred by the buildup of electric field generated by the displacement current jumping from one plate to the other. The recent Wireless Capacitive Power Transfer [38] (WCPT) standard works on the same transmission principles. Advantages of this method include low cost, spatial flexibility and operation for a wide range of frequencies [34]. For an input voltage of $V_{TX}$ connected to the transmitter, the induced voltage on receiver side, $V_{RX}$ is given by:

$$V_{RX} = V_{TX} \frac{R_{load}}{R_{load} + \frac{1}{j\omega (\frac{1}{C_1} + \frac{1}{C_2})}}$$

where $R_{load}$ is the impedance of the load. In order to achieve significant output power, large capacitance and high voltage, high frequency TX source are desired according to (1). These requirements limit capacitive coupling technique to applications where the transmission distance is typically small and in order to increase the distance between the plates, the generated electric field has to be from a very high voltage and high frequency AC power.

![Figure 7 Simplified circuit diagram of capacitive coupling WPT](image)

In [34] authors suggest the use of one-pulse switch systems to maximize the frequency content and increase the transfer efficiency. Other benefits of capacitive coupling include its simplicity and low electromagnetic
radiation. Furthermore, such systems also have capability of combining data and power transfer using same interface [39].

2.2 Inductive Coupling

Inductive coupling techniques operate on the basis of magnetic induction, generally confined to the near-field region of the operating coils. As shown in Figure 8 consider two loops of wire in close proximity of each other.

![Figure 8 Mutual inductance in LC resonator.](image)

According to Faraday’s and Lenz’s Laws, time variations of the current in the primary loop (TX in Figure 8) gives rise to the magnetic flux $\phi_{TX}$ linking the two coils, and induces an electromagnetic force ($\varepsilon$), with an opposite polarity, in the secondary coil (RX):

$$\varepsilon = -\frac{d\phi_{TX}}{dt} = -M \frac{dI_{TX}}{dt}$$  \hspace{1cm} (2)
\[ d\phi_{TX}/dt, \] is the time derivate of magnetic flux through the coils and \( M \) is the mutual inductance between the them. The induced emf in RX loop produces a current \( I_{RX} \) in the opposite direction to the primary current, \( I_{TX} \), and induces magnetic flux \( \phi_{RX} \) in RX loop. \( I_{RX} \), current be calculated using Biot-Savart’s law as follow:

\[
B = \frac{\mu_0}{4\pi} \oint \frac{I_{TX} d_i \times r}{|r|^3}
\]

where \( B, \mu_0, I_{TX}, \text{and } r \) are the magnetic flux density, permeability of free space, current in the TX coil, and displacement vector, respectively. Mutual inductance is derived from eq. (2) and (3). Now with fundamental blocks of inductive coupling explained, common classifications are discussed.

Two common classification of inductive couplings based on coupling distance \( d \) (in Figure 8) between the coils (TX and RX) are tightly and loosely coupling. Tightly-coupled scenario is referred to the case where distance \( d \) is small compared to the size of coupled coils. As an example, products that are following Qi standards are designed based on tightly-coupled operation. Tightly-coupled operation is attractive to Qi designers due to a higher achievable coupling coefficient \( k \). The significance of having higher \( k \) for wireless transmission is explained in section 2.3.2. Another benefit of using tightly-coupled coils is their lower electromagnetic emission compared to loosely-coupled coils. IPT systems that operate based on tightly-coupled coils have been used in electric vehicles (EV) for a couple of decades [40] [41]. As a drawback, in IPT systems efficiency drops very rapidly with increasing distance, in order of \( \frac{1}{r^3} \) according to Biot-Savart’s equation (3).

Loosely-coupled regime of operation for IPT systems occurs when distance \( d \) between TX and RX is large compared to the dimensions of coils or when two closely-spaced coils are misaligned [42]. Loosely-coupled coils when operating at the resonance frequency of each coil are referred to MCRs, as defined in previous chapter. Both IPT and MCR systems operate based on principal of magnetic induction, however, there are few differences between them based on their respective system architecture and transfer characteristics [42]. According to [42], MCR systems typically operate at their Self-Resonance Frequency (SRF), i.e.
resonance achieved through the use of their intrinsic parasitic capacitance to maximize the operating frequency. SRF of MCRs is generally in the range of few to tens of MHz, depending on the geometry of coils used. As will be seen shortly, higher SRF leads to an increase in the quality factor of coils significantly improving the efficiency.

Therefore, MCR systems can improve impedance matching and reduces losses, which leads to overall efficiency increase in larger separation distance. On the other hand IPT systems run in kHz frequency range and has quality factor of below 10 [43]. IPT systems operate with two coil system and in order to reach operating frequency has to implement capacitors in series or parallel combinations with TX and RX coils. Secondary coil in IPT is tuned to operating frequency, therefore slight changes in distance between these coils will drastically decreases the magnetic coupling. Next section goes to more detail about how resonance based system or MCR functions and analyzes its system performance [42]. Another advantage of MCR systems is having freedom of employing more than two coils in the WPT system. By having extra coil in both primary and secondary side of system, source and receiver loads will be decoupled from the system, resulting in increase of the overall efficiency.

2.2.1 Magnetically Coupled Resonators

MCR systems mainly consists of three to four self-resonating, high quality-factor, coupled resonators that transmit energy in the near-field regime. Robustness of the functionality to variable parameters such as the distance between transmitting and receiving coils and changes in the load conditions are important criteria to determine the versatility of the implemented systems [44, 35]. MCR system operate by storing the energy in the electromagnetic field linking the transmitter and receiver coils, thus minimizing the radiation loss. In coupled resonators any energy that is not consumed by the receiver side returns to the transmitter coil. The major source of energy loss in these systems is due to the AC resistance in the coil wires.
Supply power from the commercial home outlets is a 50 – 60 Hz AC source, which is much lower than the frequency required by WPT system. Figure 9 illustrates general block diagram of components required for a typical WPT system. Power inverter together with the power amplifier is responsible to generate the higher frequency AC power required by MCRs. On the receiver side, power is rectified and converted back to the DC.

In order to better understand MCR system consider the RLC tank resonator, with inductor and capacitors connecting in series. In this system energy is stored and transformed, back and forth, between the magnetic-field of the inductor and the electric field of the capacitor. Resonant system can be associated with pendulum operation, where kinetic energy is transferred to the potential energy through pendulum (maximum potential energy occurs when pendulum comes to halt at certain height and when it starts moving, the energy is transferred back from potential to kinetic energy.

In RLC circuit, resonance is defined as, collapsing magnetic field of inductor induces current that charges capacitor and then discharging of capacitor will construct the magnetic field in the inductor again. Existence of any resistance in the both mechanical and resonator system manifests itself as damping effect.

Figure 9 4-coil MCR wireless power system.
Transferring sufficient power is one main requirement of the most WPT systems. Figure 10 presents the 4-coil WPT system. In this design two coils are placed on the primary side (TX) and two on the secondary side (RX). L₁-L₄, Cₛ₁-Cₛ₄, Rₛ₁-Rₛ₄ represents the four values of TX and RX inductance, tuning capacitors, and resistances due to the coil losses in the inductors, respectively. As also depicted in Figure 10, d is the distance between TX and RX, and k₁₂, k₂₃, k₃₄ are the coupling coefficient of the respective coils in the system. By applying KVL to the four loops, all currents and voltages may be calculated and the transfer function of the system can be found (V₁/Vₛ) as a function of operating frequency ω. It is worth noting that in the context of RF circuits, Scattering (S) parameters are often used for analysis and measurement of circuits.

2.1 Maximum Power Transfer Efficiency

Maximum Power transfer efficiency of MCR WPT system is based on resonators. Most losses in this system depends on resistance of the coil and AC losses. Therefore, the power efficiency of higher than 50% is achievable. In a magnetically-coupled WPT system, power transfer efficiency heavily depends on the Q-factor of the two coils and the coupling coefficient between them.
Derivation of equation (4) is presented in appendix A. Four coil system has been an alternative solution for many WPT applications due to its ability to decouple the low Q-factor of source and load coils (due to loading effects) from the low coupling coefficient between TX and RX. Efficiency of four-coil system is presented by [24]:

$$\eta = \frac{k^2 Q_1 Q_2}{1 + k^2 Q_1 Q_2}$$  \hspace{1cm} (4)$$

$$\eta = \frac{(k^2 Q_1 Q_2)(k^2 Q_2 Q_3)(k^2 Q_3 Q_4)}{[(1 + k^2 Q_1 Q_2)(1 + k^2 Q_3 Q_4) + k^2 Q_2 Q_3)(1 + k^2 Q_2 Q_3 + k^2 Q_3 Q_4)]}$$ \hspace{1cm} (5)$$

In comparison with efficiency of 2-coil system eq. (4) 4-coil system eq. (5) can achieve higher efficiency, because here $Q_1 - Q_4$ are much smaller than $Q_2 - Q_3$ due to the loading effect of source and receiver side. Coupling coefficient between 2-3 coils is much smaller than coupling coefficient between 1-2 and 3-4 coils $K_{23} \ll K_{34}, K_{12}$ because in MCR system distance between two middle coils is increase. Approximate model is shown as follow [24]:

$$\eta \approx \frac{k^2 Q_2 Q_3}{1 + k^2 Q_2 Q_3}$$ \hspace{1cm} (6)$$

Advantage of using 4-coil system is to increase the transmission distance for a given PTE, however, this system reduces the PDL significantly and is not practical for the applications that requires high power at the RX side unless very large input voltage is applied [25]. To improve the PDL, 3-coil WPT systems are proposed in [25]. The three parameters to characterize any WPT resonators system are the quality factor, coupling coefficient and the resonance frequency, which are being presented in the next sections.
2.1.1 Quality Factor and Resonance Frequency

The two important factors directly effecting the efficiency of resonators are the maximum amplitude of current and its relative width when graphed versus $\frac{\omega}{\omega_o}$ as shown in Figure 11 by spice simulation.

![Alternating current graph across resonator](image)

**Figure 11 Alternating current graph across resonator**

Maximum amplitude of the current through this RLC tank is reached at the resonance frequency given by [25]:

$$\omega L - \frac{1}{\omega C} = 0 \text{ or } \omega_o = \frac{1}{\sqrt{LC}} \quad (7)$$

$L, C, \omega, \omega_o$ represent the inductance, capacitance, driving frequency of system and resonant frequency of the system, respectively. At resonance frequency voltage across capacitor and inductor cancel each other due to the two reactance be equal in magnitude and opposite in sign. Therefore, system is reduced to a simple resistance and voltage source [45].

The width of the current graph is presented by a quantity called $Q$-factor or quality-factor of the system. It is the ratio of energy stored in $L$ and $C$ of the system over the losses as shown in eq. (8).
Q-factor of the RLC system can be compared with stiffness of spring in pendulum system. In other words, Q represent how fast the energy decays when source of oscillation is removed, and according to Figure 11, sharper peaks correspond to higher Q-factors [46].

\[ Q = 2\pi f \frac{\text{Energy stored}}{\text{Power loss}} \]  \hspace{1cm} (8)

### 2.1.2 Coupling Coefficient

The self-inductance \( L \) of the coil is a property that demonstrate the measure device opposition to any current changes through the conductor [47]. This opposition to the current change produces a back emf \( (v = \frac{dp}{dt}) \) in the circuit [47]. Back emf is related to the rate of change of magnetic field. Current inside the coil generate back emf within itself due to the changing magnetic field created by turning of conductor in the coil. When this induce back emf created by the turns of inductor induces a magnetic field in an adjacent conductor then these two inductors are sharing the magnetic flux or magnetic coupling between them. This is presented by Mutual Inductance \( M \). Maximum value of mutual inductance exist between two coils are presented by eq. (9):

\[ M^2 = L_1 L_2 \]  \hspace{1cm} (9)

Maximum mutual inductance that can exist between two coils is \( M = \sqrt{L_1 L_2} \) meaning all the magnetic flux from first coil is linked to the second coil [47]. The ratio of actual mutual inductance to the maximum achievable mutual inductance is called the coupling coefficient, \( k \) as shown in eq. (10). Coupling coefficient can also be considered as the fraction of primary coil voltage that can be induced on the secondary (EMF).

\[ k = \frac{M}{\sqrt{L_1 L_2}} \]  \hspace{1cm} (10)

In [48] authors presented a closed-from expression for coupling coefficient as a funtion of their geometry, i.e. their radii \( r_1 \) and \( r_2 \), and the distance \( d \) between them:
\[ k = \frac{1}{[1 + 2^\frac{3}{2}(d/\sqrt{r_1 r_2})^2]^{\frac{3}{2}}} \]  

(11)

2.2 Frequency Splitting

Contrary to common perception, when coils 2 and 3 in a 4-coil WPT system (with a fixed input frequency) are brought to close vicinity of each other, the overall efficiency sharply decreases. This is mainly due the change in the effective value of inductance in each coil due to the coupling from the neighboring coil, an effect known as frequency splitting. In other words, Frequency splitting occurs due to change in the system impedance [35]. It has been analytically, and experimentally shown that the existence of a neighboring resonator results in the generation of two distinct peaks equally distanced from the original resonance frequency as shown in Figure 12 [49]. Figure 12 is provided through spice simulation. This is a critical issue in WPT systems where one (or both) coils move and they may be brought to close proximity of each other, significantly affecting the resonance frequency. In Chapter 5, an adaptive matching technique is proposed to alleviate the problem using a switchable capacitor scheme.

![Figure 12 Frequency splitting of two resonators for k values ranging from 0 (blue in center) to 0.2 in 0.05 increments. V_1=1v](image)

2.3 AC Resistance

Electrical circuits operating at high frequency AC current encounter power losses different from those operating at DC (DC losses are mainly in the form of heat). These losses are due to eddy currents created inside the conductors. Eddy currents are loops of electric currents induced by magnetic induction, which oppose the existing current in the conductor as shown in Figure 13. Eddy current create secondary magnetic flux inside the conductor, which oppose the original magnetic flux created by the applied AC current.
Secondary induced magnetic flux generated by eddy current reduces overall flux or field in the center of the conductor [50]. There are two major sources of eddy current effecting the performance of WPT system, namely, skin effect and proximity effect. Both of these effects cause uneven distribution of current through the cross section of a circular conductor. As a result the current density will be higher towards the outside edge of the conductor.

Figure 13 Eddy current and skin effect

2.3.1 Proximity Effect

Proximity effect is when two current carrying conductors are place next to each other they induce eddy currents in the adjacent conductor in the process changing the magnetic flux and current density distribution in both conductors. In the high frequencies when skin effect creates small skin depth the effect of proximity effect increases. Proximity effect causes significant power loss in high frequency AC inductors [51].

2.3.2 Skin Effect

Skin effect caused by AC current causes an exponential decrease in the intensity of current towards the inside of the conductor. As a result only a thin layer of outside of conductor carries most of the current and
its depth farther reduces by increasing the frequency of the AC current. The thin layer is presented by skin depth $\delta$ [52] Figure 14.

$$\delta \approx \frac{1}{\sqrt{\pi \mu \sigma f}}$$  \hspace{1cm} (12)

Eddy current in the conductor causes an increase of magnetic flux concentration in the center and leading to increasing of inductance and consequently reactance. Result would be less current density in the center of the conduct.

Skin effect is mainly due to the magnetic induction and is a direct consequence of Maxwell equations.

![Figure 14 Skin depth $\delta$ created due to skin effect](image)

2.4 Use of Litz Wire

This class of conductors provide a solution to increase the effective wire cross section at high frequencies while reducing both skin effect and proximity effect [53]. Figure 15 shows how Litz wire is constructed by bundling multiple strands of wire that are twisted and connected in parallel. The diameter of individual strands of the Litz wire is smaller than the skin depth of regular wire conductor (dependent on frequency of operation). Therefore, multiple insulated strands significantly reduce the skin effect and AC loses. Twisting of the wire has two effect firstly it redistributes the magnetic field of the conductor causing to reduce eddy current. If considering a
single strand, twisting also reduces the presents of that strand in the outside path reducing proximity effect with the adjacent wires in the process.

In [54] using simulation and calculation methods, skin and proximity effect of Litz wire is calculated for a WPT system. Two method of calculating skin and proximity effects are the exact 2-D and approximate 1-D methods, both using Bessel functions.

Due to the infinitesimal radius of strands in Litz wire, it may contain thousands of strands inside and would not be practical to use them in design environments such as finite element simulation software [54]. Therefore, according to [54], Litz wire has been homogenized and overall effect is used in the simulations. Figure 16 illustrates wire resistance versus frequency graph. According to Figure 16 proximity effect is the main contributor to the AC losses of Litz wire and on the other hand skin effect is mainly constant by increase of frequency.
2.5 Review of Conferences and Transaction Papers in WPT

This section briefly reviews several WPT research works used for a variety of applications ranging from EV to adaptive biomedical circuits and Microrobots.

2.5.1 Investigation in Efficiency of MCR WPT systems

In [55] using magnetic resonance coupling theory, authors have derived an expression relating the maximum efficiency of two-coil system to the air gap distance between the them:

$$\eta_{21}(\omega_o) = \frac{(Z_o - R)^2}{L_m^2 \omega_o^2} = \frac{L_m^2 \omega_o^2}{(Z_o + R)^2}$$  \hspace{1cm} (13)

$$L_m = \frac{\mu_o}{4\pi} \int_0^{2\pi} \int_0^{2\pi} \frac{r^2 \cos(\theta_1 - \theta_2)}{\sqrt{2r^2 + g^2 - 2r^2 \cos(\theta_1 - \theta_2)}} d\theta_1 d\theta_2$$  \hspace{1cm} (14)

where $R, L_m, \omega, Z_o$ are the internal resistance of antenna, mutual inductance, characteristics impedance and the resonance frequency, respectively. As implied from (14), $L_m$ is a function of air gap distance, $g$.

The TX and RX are implemented as two short-type identical helical antenna that are tuned to the same resonance frequency (using similar external series capacitor). Using this structure, the efficiency of the
resonator system is characterized by adjusting the air gap between TX and RX. This efficiency is measured using $S_{21}$ measurement performed using a vector network analyzer (VNA).

![Diagram showing power flow and antennas](image)

**Figure 17 Parameters of helical antennas and experimental setup [52]**

The measured mutual inductance of coils are in close agreement with calculations using Neumann formula Eq. (14), and the coupling coefficient $k$ is calculated using the method of moments.

It is noteworthy to mention that the geometrical constraints of the coils such as their radii and number of turns in the coils, also directly impact the maximum achievable efficiency.

### 2.5.2 IPT systems in Electric Vehicle

MCR systems operate at MHz frequency range and flux linkage is typically through the air medium. However, in order to adopt this technology for EV and meet the few hunder kilowatt power required for EV operation, large size coils are required which essentially lower the operational frequency to kHz range. Furthermore, to avoid flux leakage, ferrite cores are used wherever possible. In [56] authors focus on the advancement of IPT technology in power distribution and automated systems. They introduced improved version of DD (D refers to shape of the pad as depicted in Fig. 18) to polarize flux flow in space and lower leakage loss. Compared to the traditional DD structions, these new polarized pads enable designers to capture almost all the transmitted magnetic flux (if TX - RX are placed in a proper orientation) by increasing the flux height. Coupling coefficient of proposed pads are reported to be 20%-25% higher than those of traditional pads.
2.5.3 Investigation in Coil Misalignment Model

As discussed earlier, one of the practical problems of WPT systems, which significantly reduces power transfer efficiency is TX – RX coil misalignment. In [57] authors presented a theoretical framework that allow users to predict the performance of WPT systems, eliminating the need for time-consuming 3D simulations to provide first cut numbers due to orientation imperfections. In this work both types of lateral and angular misalignments are investigated using both cylindrical and circular planar coil geometries. In order to derive relationship between the efficiency and spatial misalignments, Biot-Savart Law is used to calculate the magnetic field generated by TX coil. Using Faraday’s Law the voltage generated across RX coil is then obtained. These steps are performed for i) perfectly aligned coils ii) laterally misaligned and eventually iii) angularly misaligned, i.e. tilted coils. For example the following expression demonstrates the dependence of transfer efficiency on angular misalignment represented by angle $\phi$ angle in Eq. (17).

Experimental data are shown to be in close agreement with that of analytical results.
\[
\eta = \frac{P_{RX}}{P_{TX}} = \frac{\mu_0^2 N_{TX}^2 N_{RX}^2 \omega^2 b^2 a^2 \pi^2 \cos^2}{16 R_{TX} R_{RX} (a^2 + d^2)^3}
\]

2.5.4 Novel Adaptive Impedance Matching System

One of the main disadvantages of MCR WPT systems is the performance sensitivity to the matching of input frequency and coil resonance frequency. For high-Q resonators, specifically, when system is not running at the resonance frequency, the power efficiency drops sharply. This frequency mismatch can simply be due to components ageing, capacitances slightly change over time [58], or the existence of other neighboring structures changing the effective inductance of the coils. In order to prevent this efficiency degradation, several impedance-matching techniques have been proposed in the literature. In [59], authors proposed the addition of a novel matrix of capacitors, which can connect in series or parallel to the tank, in conjunction with a dynamic impedance tracking algorithm to dynamically track the optimum capacitance combination. According to the author, this scheme is capable of covering wider range of frequency mismatches by finding the optimum capacitance with lower resistance compared to traditional matching networks. Another advantage of the proposed system is its ability to maintain the original resonance frequency that system was tuned in which eliminates the need for any frequency change to the source. Figure 19 shows the block diagram of the proposed system. There is a real time impedance-tracking system that is monitoring current on the power transmitter side, bottom circuit. An analog-to-digital converter (ADC) is used to send the sampled current to the microcontroller unit (MCU), which computes the required tank impedance in order to maximize the output signal. This impedance is then translated to a set of control signals connected to the array of switchable series/parallel capacitances that tune the circuit. In this work, seven capacitor matrices have been implemented to produce a power transfer efficiency of 88%.
2.5.5 Implementation of MCR WPT in Microrobots

This paper presents the use of MCR WPT systems (in place of the conventional batteries) in order to power a 4cm, 2.1 g Harvard Ambulatory Micro robots [60]. These biologically-inspired robots operate using piezoelectric actuators, which require an input power of 10mW to 1W to operate. 3-coil resonance inductive coupling technique is used due to its high efficiency capabilities.

Figure 19 Real time impedance matching system [59]

Figure 20 Robot moving 5cm above blue TX coil. [60]
A single TX coil, top picture of Figure 20, is able to provide the required power to operate the robot. As shown on the bottom picture of Figure 20, the implementation of multiple such TX coils along the path of robot allows to maintain a constant power to it extending its physical range of operation.

![Figure 21 Equivalent circuit diagram of 3-coil WPT system for micro robots [60]](image)

Circuit representation of system is shown in Figure 21, Class A power amplifier is connected in series with a bi-directional coupler and the TX coil inductor, $L_1$. The combination of coupler, RF detector and microcontroller unit in a feedback loop is used to track the frequency and maximize the delivered power. The setup shown in Figure 20 (bottom) is shown to be able to autonomously operate miniature robots.
Chapter 3  On the Use of Conical Helix Inductors in Wireless Power Transfer Systems

3.1 Introduction

As seen in previous chapter, inductors typically implemented in the shape of planar or helical coils, are responsible to generate the magnetic flux linkage between the TX and RX. Physical characteristics of the coils such as their radius (or the geometric mean of their radii, if their radius is different) and center-to-center separation, play an important role in determining the amount of flux linkage between TX and RX. For large transmission distance it is essential to use relatively large coils to maintain the flux linkage at acceptable levels.

However, there exists a subtle trade-off between the size of the coil and its useful bandwidth. To address the inherent limited bandwidth of large planar coils as shown in Figure 22-(a), an alternative inductor geometry which offers comparable performance with a higher bandwidth is favorable. A practical candidate
to enhance the operational frequency of the coils is the conical geometry as shown in Figure 22-(b). Conical inductors are used in a wide range of applications including military, measurement, magnetic resonance imaging (MRI), optical transceivers, and electromagnetic levitation melting [61]. Incorporation of conical inductors to replace their planar counterparts is specifically advantageous in mid to long range wireless power transfer systems for which large coils are typically required to achieve reasonable efficiency values.

This chapter proposes the use of conical helix inductors to improve the efficiency of multi-coil WPT systems. Section 3.2 discusses and compares the benefits of employing conical geometry in place of planar structure in multi-coil WPT systems as well as electromagnetic simulation results of multi-coil structures. Circuit-level simulations and measurement results for a 3-coil WPT system are presented in Section 3.3. Concluding remarks follows in Section 3.4.

3.2 Conical Inductors for Wireless Power Transfer

Resonance-based wireless power transfer systems traditionally use two to four planar coils to enable efficient energy delivery. The efficiency of resonance-based magnetically-coupled WPT systems is given by [23], [24]:

\[
\eta = \frac{k_{TX, RX}^2 \cdot Q_{TX} \cdot Q_{RX}}{C + k_{TX, RX}^2 \cdot Q_{TX} \cdot Q_{RX}}
\]  

(16)

where \( k_{TX, RX} \) is the coupling coefficient between TX and RX coils and \( Q_{TX} \) and \( Q_{RX} \) are the quality factor of TX and RX coils, as defined by \( \frac{L_{TX} \omega}{R} \) and \( \frac{L_{RX} \omega}{R} \) respectively. Constant \( C \) varies from 4 for the two-coil to 1 for four-coil systems [23] [25]. In a two-coil system, the primary (also called source) and the secondary (also called load) coils perform as TX and RX, respectively. In three-coil system, an intermediate high quality factor resonator which is tightly coupled to source (load) coil serves as the TX (RX). Finally, in a four-coil system, the two high quality factor resonators coupled to the source and load coils perform as TX and RX respectively [24]. As shown by Eq. (16), the overall efficiency of the power transfer system could
improve by either increasing the coupling coefficient, namely, $k_{TX,RX}$ or increasing the quality factor of the coils. Coupling coefficient between two spiral coils can be estimated by [62]:

$$ k = \frac{1}{\left[1 + \frac{2^2 d}{\sqrt{r_1 r_2}}\right]^2} $$  \hspace{1cm} (17)

where $r_1$ and $r_2$ are the radii of the transmitter and receiver coils, respectively, and $d$ is the power transfer distance (i.e., the center-to-center distance between the aligned TX and RX coils). It can be seen from Eq. (17) that doubling the transmission distance reduces $k_{TX,RX}$ by a factor of $\approx 8$. Therefore, to maintain the efficiency over a longer distance it is imperative to increase the size of TX and RX coils, hence keeping the denominator of the expression in Eq. (17) constant. This increase in the size of the coils lends itself to a lower frequency of operation due to the increased parasitic capacitance of the coils. This limitation is further highlighted for magnetically coupled resonators (MCRs) in resonance-based three- and four-coil power transfer systems, where to maximize $Q_{TX}$ and $Q_{RX}$ the system is typically operated at frequencies close to the self-resonance frequency (SRF) of the MCRs [35]. Due to the higher achievable SRF for comparable size and performance, conical geometry is an attractive alternative candidate as compared to its planar counterpart. More specifically, the cone-shaped inductor owes its enhanced SRF to a lower stray capacitance value [61]. For a tightly wound coil, this value is inversely proportional to the separation between the adjacent windings [24]. The separation is typically larger for conical coils and therefore, results in a lower stray capacitance. One drawback of operating at higher frequencies is that the resistance of the coils may also rise due to the skin and proximity effects. However, the wire resistance typically grows less than linearly with frequency and may further be reduced by the use of special wires such as Litz wire. Therefore, operating the MCRs at higher frequencies potentially translates to an improved $Q_{TX}$ and $Q_{RX}$ values. As suggested by Eq. (16), higher values of $Q_{TX,RX}$ could improve the overall power transfer efficiency of the WPT system. This is provided that using conical coils maintains (or improves) the coupling
factor $k_{TX, RX}$. To investigate the effects of conical geometry on coupling coefficient, several combinations of conical and planar spiral coils are designed and compared using the electromagnetic

(EM) simulation performed in HFSS® environment as shown in Figure 23. To meet the physical constraints of a given problem, number of turns and lateral dimensions of coils are kept similar. Note that as shown in Figure 23, in conical coils TX (RX) distance is measured between the closest cross section of each coil. As can be seen from the HFSS simulation results structure-3, namely face-to-face configuration, shows a higher coupling coefficient among conical structures. This slight increase in the simulated coupling coefficient can be attributed to the concentration of flux linkage on the larger side of the coil as depicted in H-field simulations of Figure 22-(b). Face-to-face configuration is, therefore, selected as a structure of choice for the subsequent experiments in this work.
3.3 Simulation and Measurement Results

Given that the three-coil systems offer a reasonable trade-off in terms of complexity, power efficiency, transmission distance and impedance matching when compared to their two-coil and four-coil counterparts [63], in this work we will focus on three-coil structures. Two proof-of-concept prototypes of a three-coil WPT systems are designed and their performances are verified through simulation and measurement results. The first prototype employs three planar coils with a TX resonator placed close to the source coil and is used as the reference design as shown in Figure 24 (planar system). As suggested by the EM simulations of the previous section, the second prototype replaces planar TX resonator coil and the load coil with two face-to-face conical coils as depicted in Figure 24 (conical system). For the coils in each prototype, the inductance L, quality factor Q, SRF, and coupling coefficient for different transmission distances are extracted from HFSS EM simulation. These parameters are then fed to a circuit simulator (Spectre) where
the stray capacitance \( C_{S_{ls}} \) in Figure 24 and series resistance values \( R_{S_{ls}} \) Figure 24) are modeled based on the HFSS-extracted quality factor of the coils as shown in Figure 25. Figure 26 compares the simulated performance of both three-coil WPT systems (planar and conical) for several values of \( k_{23} \) ranging from 0.01 to 0.4 representing various distances between the TX resonator and the load coil. Figure 26-(a) provides the maximum magnitude of \( S_{21} \) for 50\( \Omega \) source and load resistances. The two systems are simulated at 4.3 MHz and 4.7 MHz, respectively, which is close to the SRF of the two resonators as determined by EM simulations. Also, as shown in the figure for the conical resonator, two scenarios are considered for the series resistance of the coil, namely, WoR and WR. In WoR,

![Graph showing quality factor vs frequency for planar and conical coils](image)

*Figure 25 Simulation results for the quality factor of planar and conical coils*

The series resistance is considered constant when the frequency is increased from 4.3 MHz to 4.7 MHz (this is to mimic the ideal case where there is no deterioration in the resistance, for example due to skin effect). In contrast, in WR, the AC component of the resistance is a function of frequency and is modeled according to the specification of 40-44 AWG Litz wire which is used for measurement setup. As shown, for both scenarios, \( S_{21} \) magnitude of the conical system shows an improvement over its planar counterpart. Figure 26-(b) shows the power transfer efficiency enhancement (in percentage) of the WoR and WR conical systems over the planar one for the same range of \( k_{21} \) variations.
In view of the EM and circuit simulation results, the two three-coil WPT systems comprising planar and conical coils are built as shown in Figure 27. Litz wire (40-44 AWG) is used for all the coils to minimize their AC resistances at MHz range. The source coils in both structures are 4-turn and measures 11 cm in diameter. An Agilent E5061B VNA is used to measure the S-parameters of TX resonators and determine its inductance and SRF in each setup. The TX resonator coil is 15 turns and measures 32 cm in diameter. The resonator measures 127 \( \mu \)H for both the planar and conical cases by manually trimming the end of spirals. The measured SRF of the coils are 4.34 MHz and 4.72 MHz, respectively, which is in close agreement with EM simulation results. The input frequency in each case is set to the corresponding SRF of the coils. Figure 28 shows the measured power transfer efficiency of the planar and conical systems. Due to the sensitivity of the SRF to the small variations in the self-capacitance of the coils, it is difficult to tune the input RF amplifier to the resonance frequency of TX coils.
Figure 27 Measurement setup for the three-coil planar (top) and conical (bottom)

Figure 28 Measured efficiency of conical and planar systems
This imperfection manifest itself as smaller efficiency improvements compared to the theory. However, as depicted in the figure, the efficiency of the conical system improves as predicted by theory and simulation results. The transmission distance of 25 cm to 75 cm corresponds to the under-coupled region (see Figure 26). The transmission distance d for the planar system is measured as the center-to-center distance of the two planes that house the coils. For the conical system, this distance is measured from the center of the outer layer of one conical coil to the corresponding point on the other conical coil which represents the worst case scenario. To also demonstrate the usefulness of conical coils at lower operating frequencies, two identical 220 pF capacitors are placed in parallel with each TX resonator. Figure 29 shows the $S_{21}$ measurement results of both spiral and conical resonators. Conical coil resonates at slightly higher frequency, 938 kHz, compared to planar resonator, 908 kHz due to its higher SRF. The measured Q-factor of the conical resonator, 195 as measured by $\frac{\omega_o}{\Delta\omega}$ where $\omega_o$ is the resonance frequency and $\Delta\omega$ the 3-dB bandwidth from $S_{21}$ plot, is also higher than its spiral counterpart at 183.

![Figure 29 S21 measurement results of TX resonators in parallel with 220 pF capacitors.](image)

### 3.4 Conclusion

A conical coil geometry is shown to improve the efficiency and flexibility of resonance-based WPT systems. For similar coil specifications, EM simulation results show an increase in the SRF and quality factor of the conical structure (as compared to the traditional planar coils) hence improving the overall
power transfer efficiency. Furthermore, the conical structure shows less efficiency variations as the orientation between TX and RX varies. Measurement results of a prototype 3-coil structure over a transmission distance of 40 cm confirms the advantages of the proposed structure.
4.1 Introduction

Near-field WPT system being tested in this chapter is the 4-coil inductively coupled working in inductive resonance coupling as presented in Figure 30 is studied. Four-coil structure is chosen to maximize power transmission efficiency of the system.

Three fundamental characteristic of resonant systems are quality factor, coupling coefficient and resonant frequency of the system, which have to be precisely measured. In addition to these three factors, geometry and shape of inductor has a significant effects in directing electromagnetic flux from primary to secondary coils [18]. In order to design a high efficient WPT system the conical coil presented in previous chapter and its characteristics are measured and compared with planar spiral coil. As discussed in Chapter 2, the quality factor of TX and RX resonators plays important role in transmitting energy to the load. The major advantage of 4-coil structure is decoupling the source $R_{Source}$ and load resistance $R_L$ from the resonators TX and RX, by doing so degrading of quality factor is prevented and more power efficiency is achievable. Measuring a high quality factor resonators has to be performed with great precision to avoid loading effect of measurement device such as VNA.
In this chapter advantage of using conical inductor, qualitatively discussed in Chapter 3, are proved by accurate measurement fundamental characteristics of resonant circuit and comparing the result with the planar spiral coil. In the first part authors demonstrate more accurate method proposed by [64] to determine Q-factor. Inductive resonance circuit is designed and used to measure accurate quality factor. High frequency simulation system (HFSS) is used to confirm the results presented by theory and measurement.

### 4.2 Theory

Majority of WPT systems operated on resonators, in any non-ideal resonator system losses occur. Quality of stored energy to loss energy in the system is characterized by quality factor. Resonator circuits are in WPT are always connected to the external circuit for example rectifier power amplifier or VNA to measure S-parameters, which considered coupling network. Therefore, presents of external circuit will affect the overall quality factor of the resonator. For high quality factor resonators indirect method of measurement is develop to avoid loading effect described above, such as capacitive or inductive coupling. Capacitive coupling method presented in [65] requires capacitors on the order of \( \frac{1}{100} \) of the value of the tank capacitance and is unattainable in case of resonators, due to the very low capacitance value of the resonators (in pf order). Inductive coupling is more practical in this case.

The overall Q-factor of the system consisting of intrinsic Q of coils and Q of external \( (Q_e) \) circuit in this case VNA is known as the “loaded” quality factor \( (Q_L) \). The unloaded quality factor of the system is related
to the loaded Q-factor by the coupling coefficient of the system, which is the ratio of flux transferred between primary and secondary coils. Measuring Q-factor is difficult common procedure to measure quality factor is using the resonance frequency and 3-dB bandwidth $S_{21}$ parameter measured through VNA.

Figure 31 A resonator and an external source: (a) Thevenin’s equivalent circuit and (b) Norton’s equivalent circuit.

4.3 Measurement Technique

Resonators constitute the basic building blocks for the majority of WPT systems. The efficiency of a magnetically coupled resonating WPT system at resonance is given by:

$$\eta = \frac{k^2 \cdot Q_1 \cdot Q_2}{n + k^2 \cdot Q_1 \cdot Q_2} \text{ for } 1 \leq n \leq 4$$

where $k$, $Q_1$ and $Q_2$ are the coupling factor and unloaded quality factor of coils, respectively. To demonstrate the advantage of conically shaped coils compared to their planar coils, a reliable method of measuring $k$ and $Q$ is desirable. To understand the challenges of measuring high $Q$ tank circuits, consider the distributed model of an RLC resonator connected to the measurement device, e.g., a network analyzer, as shown in Figure 31-(a). If the source is represented by its Norton’s equivalent (Figure 31-(b)), the loading effect of the external conductance, $G_S$, on the tank overall (loaded) Q-factor, $Q_L$ is

$$\frac{1}{Q_L} = \frac{1}{Q_o} + \frac{1}{Q_S}$$

where $Q_o$, unloaded Q factor of the tank, and $Q_S$ are defined as:
$$Q_S = \frac{R_S}{\omega_0 L_0} \quad \text{(20)}$$
$$Q_o = \frac{R_o}{\omega_0 L_o} \quad \text{(21)}$$

The ratio of power dissipated in the source conductance to the power dissipated in the resonator is called coupling coefficient, $\kappa$. The voltage across both $G_s$ and $G_o$ is equal to $V_s'$; hence, the coupling coefficient can be written as:

$$\kappa = \frac{V^2 G_s}{V^2 G_o} = \frac{G_s}{G_o} = \frac{Q_s}{Q_o} \quad \text{(22)}$$

When an equal amount of power is dissipated in the external circuit as in the resonator itself, the coupling is said to be critical and the coupling coefficient in this case is $\kappa = 1$.

An under-coupling means that more power is dissipated in the resonator than in the external circuit, while an over-coupling means that more power is dissipated in the external circuit than in the resonator [64]. By combining eq. (19) and eq. (22), the expression for unloaded Q-factor, $Q_o$, can be written as:

$$Q_o = Q_o (1 + \kappa) \quad \text{(23)}$$

During the measurement, the resonator is loaded by the external circuit (network analyzer), and the measurement produces $Q_L$. Since the value of $R_0$ at the resonant frequency is not known, a measurement procedure must be devised to get a value for the coupling coefficient, $\kappa$. To facilitate the measurement of $\kappa$, an indirect method of connecting the resonator to the source must be sought. The inductive coupling method, as shown in Figure 32, is used in this work. The input impedance of the setup shown in Figure 32 can be expressed as:
\[ Z_{in} = j\omega L_p + \frac{(\omega M)^2}{R_o} \left( 1 + jQ_o \left( \frac{\omega}{\omega_o} - \frac{\omega}{\omega_o} \right) \right) \]  

(24)

As discussed in [64], \( L_p \) must be much smaller than \( L_0 \) to make reliable measurements. Therefore, the series resistance associated with \( L_p \) can be neglected. The locus of the input reflection coefficient, as shown in Figure 33, is a circle at frequencies close to the resonant frequency. To analytically show this, first note that the effect of \( L_p \) can be found from the Smith chart in Figure 33 as being equal to the normalized reactance at the point corresponding to \( f_s \). The non-zero value of \( L_p \) moves the input reflection coefficient from the short circuit point on the Smith chart to that of \( f_s \). Similar to a step-down transformer, the input resistance, at the resonant frequency is

\[ R_{in} = \frac{(\omega M)^2}{R_o} \]  

(25)

In the relatively narrow frequency range of operation in the vicinity of the self-resonant frequency, the input resistance, \( R_{in} \), is approximately constant with respect to frequency. In addition, in this limited range, \(|\Delta\omega| = |\omega - \omega_0| \leq 1\); therefore, in the range of the self-resonant frequency, the second term eq. (24) can be simplified as: \( \frac{\omega}{\omega_o} - \frac{\omega}{\omega_o} \approx \frac{2\Delta\omega}{\omega_0} \). Through \( S_{11} \) measurement, for the frequencies far from the resonant frequency, \( S_{11} \) and the input impedance approaches zero, i.e., \( Z_{in} |\Delta\omega \rightarrow \infty = 0 \). Consequently, the input reflection coefficient at these frequencies becomes \( \Gamma_{in} |\Delta\omega \rightarrow \infty = \Gamma_0 = -1 \). Therefore, \( \Gamma_{in} - \Gamma_0 \) takes a simple form.
\[ \Gamma_{in} - \Gamma_0 = \frac{2}{(1 + \frac{1}{\kappa})(1 + \frac{j2Q_L \Delta \omega}{\omega_0})} \]  \hspace{1cm} (26)

At the resonant frequency, \( \Delta \omega = 0 \), \( |\Gamma_{in} - \Gamma_0| \), reaches a maximum corresponding to the diameter of the circle:

\[ d = |\Gamma_{in} - \Gamma_0|_{\text{max}} = \frac{2\kappa}{(1 + \kappa)} \]  \hspace{1cm} (27)

Therefore, through the measurement of the circle diameter on Smith chart, the coupling coefficient can be calculated as:

\[ \kappa = \frac{d}{2 - d} \]  \hspace{1cm} (28)

To measure \( Q_L \), note that the phase of (26), \( \beta \), is given by:

\[ \tan \beta = -2Q_L \frac{\Delta \omega}{\omega_0} \]  \hspace{1cm} (29)

For a given value of \( \beta \), two frequencies \( f_L \) and \( f_R \) can be measured and the loaded Q factor is calculated by:
\[ Q_L = \frac{f_0}{f_R - f_L} \tan \beta \]  

(30)

Knowledge of \( \kappa \) and loaded Q-factor, \( Q_L \), facilitates the calculation of the unloaded Q-factor, \( Q_0 \), according to (23). To characterize the self-inductance of each coil, the measurement frequency must be selected such that \( \omega \ll \omega_0 \) to minimize the effect of the inter-winding parasitic capacitance. From the one-port \( S_{11} \) measurements, \( L \) is calculated as:

\[ L = \frac{\text{Im}(Z_{11})}{2\pi f} \]  

(31)

To measure the coupling factor, \( k \), between the coils the mutual inductance between the two coils, \( L_a \) and \( L_b \), must be measured. As noted in [66], there is a one-to-one relationship between the coupling factor, \( k \), and the coupling coefficient \( \kappa \). From two port measurements, the value of \( M \) and \( \kappa \) can be extracted as:

\[ M = \frac{\text{Im}(Z_{12})}{2\pi f} \]  

(32)

And

\[ k = \frac{M}{\sqrt{L_1L_2}} \]  

(33)

4.4 Experimental Results

Two 4-coil WPT systems using planar and conical resonators, shown in Figure 34(a)-(d), were implemented. To make a fair comparison between the two structures, similar physical dimensions for planar and conical coils are chosen as summarized in Table 1. To improve the efficiency, large resonator coils of 15-turn Litz wire are implemented for both setups. S-parameter measurements is made with an Agilent E5061B VNA. Additionally, an E4417A Agilent power meter was used to measure both input and output power for efficiency measurement. A full two-port VNA calibration is performed to compensate for the
measurement cables. $L_p$, shown in Figure 32, is implemented as a small 3-turn inductor coil with a radius of 5.1 cm and is connected to port 1 of the VNA.

![Image](72x532 to 540x669)

![Image](144x210 to 468x407)

*Figure 34 a) 4-coil conical measurement setup b) 4-coil planar measurement setup*

For Q measurements, $L_p$ is placed at a distance of 3.5 cm from the device-under-test (DUT). The measured input reflection coefficient, $S_{11}$, is shown for both conical and planar coils in Figure 35.

![Image](529x51)

*Figure 35 Input reflection coefficient (a) Smith chart and (b) magnitude for Q*

The loaded self-resonant frequency, $f_0$, for each setup corresponds to the frequency at which $S_{11}$ reaches its minimum. In order to measure the loaded Q-factor using (34), $\beta$ is chosen to be 45° and the values of $f_R$ and $f_L$ are extracted from the VNA measurements. Measurement results shown in Table I verify the increase
in the self-resonant frequency of conical coils as compared to planar coils. This increase is primarily due to the lower fringe capacitance in conical structures. An increase in the measured unloaded Q-factor of conical coils further confirms the usefulness of the conical structure. The simulation results and measurement values of the coupling factor, k, for different distances between TX and RX coils are plotted in Figure 36. As can be seen in the figure, there is close agreement between measurement and Ansoft HFSS simulation results. All power efficiency measurements reported are defined as:

\[ \eta = \frac{P_{out}}{P_{in}} = \frac{P_{out}}{(1 - |S_{11}|^2)P_{ave}} \quad (35) \]

where \( P_{out} \) is the RF power delivered by the WPT structures, \( P_{in} \) is the RF power delivered to the input of the structures, \( P_{ave} \) is the available RF source power, and \((1 - |S_{11}|^2)\) is the input mismatch loss. Compensation for input mismatch loss is included because the design will use matching circuit. The efficiency measurements for the 4-coil structure is displayed in

Table 2. The efficiency of the conical structure in the worst case scenario (0.5 m distance of outer winding between the conical structures) is higher than that of its planar counterpart.

<table>
<thead>
<tr>
<th>DUT Coil</th>
<th>Inner radius (cm)</th>
<th>Pitch (cm)</th>
<th>Inductance (µH)</th>
<th>Self-resonant frequency (MHz)</th>
<th>Unloaded Q factor</th>
</tr>
</thead>
<tbody>
<tr>
<td>Planar ((L_a))</td>
<td>15</td>
<td>—</td>
<td>137</td>
<td>4.13</td>
<td>145</td>
</tr>
<tr>
<td>Conical ((C_a))</td>
<td>15</td>
<td>0.3</td>
<td>129</td>
<td>4.77</td>
<td>259</td>
</tr>
</tbody>
</table>
### Table 2 Performance Comparison of 4-Coils WPT

<table>
<thead>
<tr>
<th>Structure</th>
<th>Frequency of operation (MHZ)</th>
<th>Distance of middle coils (m)</th>
<th>S11 (dB)</th>
<th>Power Efficiency (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{ex1}, L_a - L_b, L_{ex2}$</td>
<td>4.37</td>
<td>0.5</td>
<td>4.97</td>
<td>32.8</td>
</tr>
<tr>
<td>$L_{ex1}, C_a - C_b, L_{ex2}$</td>
<td>4.84</td>
<td>0.5 (max)</td>
<td>2.71</td>
<td>53.9</td>
</tr>
<tr>
<td>$L_{ex1}, C_a - C_b, L_{ex2}$</td>
<td>4.84</td>
<td>0.5 (mid)</td>
<td>3.21</td>
<td>49.5</td>
</tr>
<tr>
<td>$L_{ex1}, C_a - C_b, L_{ex2}$</td>
<td>4.84</td>
<td>0.5 (min)</td>
<td>3.89</td>
<td>43.0</td>
</tr>
</tbody>
</table>

**Figure 36** The coupling factor of planar and conical coils as a function of coils

### 4.5 Conclusion

The effect of coil geometry on the performance of wireless power transfer (WPT) systems is studied. Using an inductive coupling technique, the unloaded quality factor of the inductors is measured, confirming an improvement in the value of self-resonant frequency and quality factor.
of conical coils. A power transfer efficiency of up to 53.9\% is achieved for a 4-coil WPT system employing conical coil resonators for a TX and RX separation of 50 cm.
Chapter 5  An Adaptive Magnetically-Coupled Wireless Power Transmission System

5.1 Introduction

Resonators pulling effect causes efficiency degrading, when placed in close proximity of each other. To address this issue adaptive impedance matching and frequency tuning techniques are used. Specifically, the variations in the transmission distance can significantly impact the dynamics of the system. Splitting of the resonance frequency into even and odd modes is a known effect for short range power transmission which could significantly deteriorate the performance of systems with a fixed input frequency. In [35], [59] adaptive frequency tuning techniques are suggested where a directional coupler connected between the input RF amplifier and the transmission coil measures the amount of incident and reflected powers and accordingly tunes the frequency of RF source to minimize the reflection component. Such techniques benefit from the fact that magnetic coupling is a near-field effect and the power not absorbed by the load would reflect back to the source.

Figure 37 HFSS model of critically coupled and over-coupled 4-coil power transmission system.
An alternative approach to alleviate the impact of distance variations is suggested in [44] where an impedance matching network is added to both source and load coils. The addition of matching network at both ports facilitates the matching of impedance hence maximizing the power transfer to the load. However, the implemented prototype relies on a control system that operates based on extensive measurement data taken using VNA and a Matlab algorithm developed to determine the state of programmable switches that set the amount of capacitance on both TX and RX sides.

5.2 Four Coil System Analysis

Resonant-based power delivery systems typically use four coils as shown in Figure 37 to accomplish the task. The primary coil (also called source loop) is connected to the RF source, hence loaded by its internal resistance. To compensate the resulting low-quality factor, a high-quality factor secondary coil, namely TX coil in Figure 37, is tightly coupled to the primary coil. A proper value of capacitor is added to each coil to make them resonate at the same frequency.

![Figure 38 Simulated S21 for the critically coupled and over-coupled resonators.](image)

This combination structure is suitable for mid-range power transmission where the coupling factor is poor, 0.01 to 0.001. In [8], it is shown that using the same combination structure at the receiver side, the low coupling coefficient of RX and TX, $k_{23}$, is decoupled from the low-quality factor of source and load loops as expressed by the efficiency expression:
\[ \eta = \frac{k_{23}^2 \cdot Q_2 \cdot Q_3}{C + k_{23}^2 \cdot Q_2 \cdot Q_3} \]  

(36)

Where \( k_{23} \) is the coupling coefficient between the secondary (TX) and ternary (RX) coils and \( Q_2 \) and \( Q_3 \) are the quality factor or TX and RX coils, respectively. It is noteworthy to observe that efficiency is independent of low quality factor of source and load loops, namely \( Q_1 \) and \( Q_4 \). It has been shown that as the distance between the transmitter and receiver structure decreases, i.e., the coupling factor \( k_{23} \) increases, the resonance frequency of each coil is disturbed from the value determined by \( \omega_o = \sqrt{\frac{1}{L_i C_i}} \), where \( L_i \) and \( C_i \) are the inductance and capacitance of that coil. This effect, also known as frequency splitting, is illustrated in Figure 38 where the frequency of resonance splits into two distinct modes as the coupling factor exceeds a value known as the critical coupling. A static resonating structure is well-suited for energy transfer below the critical coupling point, the under-coupled regime, although its efficiency would quickly drop with increasing distance between the TX and RX coils. However, beyond the critical coupling point, the system requires a dynamic means of monitoring the distance, i.e., the effective coupling coefficient, and tuning the circuit accordingly. Otherwise, any small changes in the transmission distance de-tunes the LC tanks and causes significant drop of overall efficiency. There are two major adaptive methods to cope with the resonance frequency splitting in over-coupled regime, namely adaptive frequency tuning and adaptive impedance tuning [44]. Once a drop in the efficiency of the system is detected, adaptive frequency tuning scheme manipulates the frequency of the power amplifier in an effort to adjust the frequency of transmission to resonance frequency of the over-coupled coils (odd or even mode). This scheme despite eliminating the need for any impedance matching network, comes with an inherent disadvantage. In most applications, the wireless power transmission system needs to operate within a certain frequency band (ISM band). In order to accommodate a reasonable distance variation (corresponding to wide range of coupling coefficient), while maintaining a high efficiency, adaptive frequency tuning scheme normally requires a tuning range which far exceeds the narrow bandwidth of the relevant ISM band. This shortcoming severely limits
incorporation of frequency tuning scheme in most applications where Federal Communication Commission (FCC) regulations have to be observed. Adaptive impedance tuning scheme on the other hand, operates at a designated frequency (within the ISM band) and manipulates the value of switchable capacitors to maintain a high efficiency. More specifically, upon detection of a drop in efficiency in the over coupled region, capacitors are tuned or added to the LC tank to push the resonance frequency of the LC resonator towards the frequency of operation (see Figure 38).

5.3 Proposed Adaptive System

The overall 4-coil power transmission system including the monitoring and control circuitry employed in this work is modeled as shown in Figure 39. As shown in the figure, the proposed system consists of four magnetically-coupled LC resonators represented by \( L_i C_i \) (\( i = 1; 2; 3; 4 \)) and their respective coupling coefficients. For a practical scenario, cross-coupling coefficients \( k_{13}; k_{24} \) and \( k_{14} \) are neglected [24]. Resistors \( R_{is} \) model the loss of each tank. For TX and RX loops, a bank of switchable capacitors \( C_{SWs} \), controlled by series switches \( SW_i \), is incorporated to enable the proposed adaptive tuning scheme.

![Figure 39 Schematic diagram of the proposed adaptive 4-coil wireless power transmission system.](image)
A small portion, 10% in this design, of the total induced voltage on the TX and RX coils, represented by $V_{sense,X,Y}$ are employed to track the distance variations between the TX and RX. This voltage can be extracted from the tank by way of connecting across one turn (or a portion) of the coil inductor as shown in the schematic of Figure 39. As the coupling between the TX and RX coils increases above the critical coupling value, the AC voltage induced on the secondary coil reduces. This voltage drop triggers the monitoring circuitry and it accordingly, sets the state of the switches in the capacitor array such that the impedance matching at fixed resonance frequency of 13.56 MHz improves and AC voltage across the tank increases to acceptable levels. The block diagram of the proposed monitoring circuitry is presented in Figure 40.

![Figure 40 Adaptive control circuitry.](image)

It consists of a rectifier, a voltage reference, a sample-and-hold (S&H), two comparators, and digital produces a proportional DC voltage at its output. This DC voltage serves as the supply voltage for the rest of monitoring circuitry. This value of rectified voltage also provides information on the deviation of resonance frequency from its nominal value, $\omega_0$. In other words, as the receiver moves closer to the
transmitter, the induced voltage on the secondary coil drops due to the creation of the $S_{21}$ dip between the two new resonant modes (see Figure 38). This drop manifests itself as a proportional drop in the amplitude. Therefore, the comparison of $V_{\text{rect}}$ against a well-defined voltage, $V_{\text{ref}}$ provided by the reference voltage generator block [67] allows detection of one such scenario. Note that in order to ensure the proper functionality of the system, the comparator uses $V_{\text{rect}}$ as its supply voltage, while it receives an attenuated version of this supply voltage, namely $V_{\text{comp}}$ as its input. Upon detection of the voltage drop comparator triggers the controller, which in turn, activates a successive process of adding/removing capacitors ($C_{\text{SWI}}$), to the coil loops.

More specifically, the controller stores the value of the comparison voltage $V_{\text{comp}}$ using the S&H unit and switches in one capacitor. If this addition improves the resonance such that $V_{\text{comp,new}} > V_{\text{comp,old}}$ (which is monitored by a second comparator comp2 the controller keeps switching in more capacitors until it meets the condition of $V_{\text{comp}} > V_{\text{ref}}$. Otherwise, the controller swaps its direction and starts switching off one capacitor at a time. The process continues until the desired threshold (i.e., $V_{\text{ref}}$) is met. Details of the control process is shown in the flowchart of Figure 41. It should be noted that for an n-bit capacitor array, the adaptive tuning system is guaranteed to converge to the desired result (i.e., $V_{\text{comp}} > V_{\text{ref}}$). The current drawn by the adaptive control circuitry can be reduced to approximately 0.6% of the TX and RX loop currents. Therefore, the proposed scheme practically does not degrade the quality factor of the TX and RX coils.
5.4 Simulation Results

The complete setup was designed and simulated using electromagnetic (EM) and circuit analysis tools. The coils were designed in HFSS® environment and their specification as well as their performance metrics, at the frequency of operation, are summarized in Table 3.

<table>
<thead>
<tr>
<th>No. of Turns</th>
<th>Turn Spacing</th>
<th>Track Width</th>
<th>Outer Diameter</th>
<th>Q</th>
</tr>
</thead>
<tbody>
<tr>
<td>L1, L4</td>
<td>1</td>
<td>NA</td>
<td>0.2 cm</td>
<td>6.76 cm</td>
</tr>
<tr>
<td>L2, L3</td>
<td>6</td>
<td>1.7 cm</td>
<td>0.2 cm</td>
<td>25.88 cm</td>
</tr>
</tbody>
</table>
The primary loop and coil are placed at a distance of 5 cm to have a coupling factor of \( k_{12} = 0.1 \). This value of \( k_{12} \) is large enough to allow an efficient power transfer at critical coupling point while it does not impact the resonance frequency of each coil, i.e. effect of mutual coupling on resonance frequency is negligible. Capacitors \( C_{1,3} \) and \( C_{2,4} \) (see Figure 39) are chosen to be 115 pF and 11.5 pF, respectively, to set the resonance frequency of each loop to the 13.56 MHz ISM band. The adaptive control circuitry was designed in a 0.13 \( \mu \)m CMOS technology with an emphasis on low power consumption to avoid degrading the quality factor of the TX and RX coils. All the building blocks are designed for on-chip integration and the voltage levels are kept within the safe levels of the CMOS process. To demonstrate the effectiveness of the proposed tuning technique, the response of system to variations in the value of \( k_{23} \), i.e., change in the transmission distance, is simulated. The TX and RX are initially placed at a distance of 33 cm corresponding to their critical coupling factor value of \( k_{23} = 0.03 \). The transient waveforms of circuit are shown in Figure 42. \( V_{rect} \) is initially at approximately 4 V (top plot) and \( V_{Load} \) exhibits large swings across \( R_{Load} = 50 \) (bottom plot). At \( t = 0.8 \) s, \( k_{23} \) switches to 0.1 which represents a new distance of 20 cm between RX and TX. The performance of circuit with and without the proposed adaptive tuning is evaluated. As can be seen from the top plot, when no adaptive tuning is present, \( V_{rect} \) drops significantly and \( V_{Load} \) swing is deteriorated by 40%. However, in the presence of the proposed adaptive tuning, the drop in \( V_{rect} \) forces \( V_{comp} \) to fall below the \( V_{ref} \) input of the comparator which, in turn, switches the output of comparator (middle plot). The controller accordingly sets the SW1 signal and switches in a fixed capacitance of 1 pF to the LC tank. This addition shifts the S-parameter curve to the left and places the odd mode of resonance at 13.56 MHz (see Figure 38 curve with tuning). Therefore, the resulting resonance increases both \( V_{Load} \) and \( V_{rect} \) as illustrated in Figure 42 and significantly improves the overall system efficiency.
Figure 42 Transient settling behavior of the proposed adaptive tuning scheme.

For values of $C_{SW}$ in increments of 1 pF, Figure 43 shows the magnitude of $S_{21}$ at the desired frequency of 13.56 MHz versus different values of $k_{23}$. As shown in the figure, a rather flat efficiency is obtained for a wide range of $k_{23}$ variations (from 0.01 to 0.5, corresponding to a transmission distance of 0.54 meter).
5.5 Conclusion

An adaptive resonance-based magnetically coupled wireless power transmission system was presented. The system is based on the 4-coil structure and dynamically tunes the resonance frequency of TX and RX coils in response to variations of the coupling coefficient and hence, enables efficient power delivery to a larger coverage area. Unlike the adaptive frequency tuning technique, the proposed scheme is suitable for narrow-band operation. The circuitry operates in a self-sufficient manner through internally generating its supply voltage and is designed with an emphasis on low power consumption, fast real-time response and simplicity.
Chapter 6  Conclusions and Future Work

6.1 Conclusions

Wireless power transfer systems are currently at forefront of many research works. In this dissertation, a study on the characteristics of conical coil WPT systems, an alternative geometrical solution to the traditional planar coil structures is presented. The maximum efficiency of resonance-based multi-coil WPT systems is a function of coils quality factor and coupling between TX and RX [24] [35]. In this work, it is shown that for a given distance and similar lateral dimensions, conical coils provide a higher quality factor compared to their planar counterpart while providing nearly the same coupling factor values. This result, verified by EM simulation and measurement results, improves the kQ product, and, consequently, the efficiency of WPT systems according to Eq. (4).

VNA measurements results show that the quality factor and self-resonance frequency of conical coil is higher than planar counterpart for given physical constraints. Alternatively, the quality factor of conical and planar coils are measured using Ginzton’s impedance method as described in chapter 4. Both results confirm the viability of achieving higher efficiencies for conical coil structures. Three and four coil prototypes of WPT systems incorporating planar and conical coils are constructed and measurement results confirm the expected improvement in efficiency.

Also an adaptive impedance matching technique to minimize effect of distance variation between the TX and RX coils is presented in chapter 5. It is suggested that by dynamic adjustment of the resonator capacitance the amount of energy reflection to TX can be reduced. This adoption ensures the operation of both TX and RX coils at the source frequency improving the efficiency of the link.
6.2 Future Work

A robust WPT solution must provide the maximum efficiency both when TX-RX distance is fixed as well as when one move with respect to the other. Combining the benefits of proposed technique, and working out the implementation details of the proposed adaptive impedance matching technique using FPGAs, Microcontrollers, PCB designs, etc., would be an essential step to further validate the usefulness of proposed techniques.

In future work, the mechanical structure of conical coils and their winding implementation has to be improved. A thorough electromagnetic study of mutual coupling and self-resonance frequency, i.e. maximum quality factor, of WPT conical coil provides invaluable insight into the operation of WPT coils and the underlying energy transfer through the magnetic link.

Finally, a step-by-step methodology to optimize, design and implement static and dynamic WPT systems is of great importance and can be the subject of research studies combining the benefits of multiple design techniques.
Reference


Appendix A

Calculation of Power Transfer Efficiency (\(\eta\)) for Two Coil Resonance Circuit

KVL equations for the loops in a two-coil WPT system are:

\[
\begin{align*}
V_{in} &= R_1 I_1 + j\omega_0 M I_2 \\
0 &= j\omega_0 M I_1 + R_L I_2 \\
\begin{bmatrix}
R_S + R_1 & j\omega_0 M \\
-j\omega_0 M & R_2 + R_L
\end{bmatrix}
\begin{bmatrix}
I_1 \\
I_2
\end{bmatrix} &=
\begin{bmatrix}
V_{in} \\
0
\end{bmatrix}
\end{align*}
\]

The expressions for the input and output power can be derived using above equations:

\[
\begin{align*}
P_{OUT} &= \frac{1}{2} R_2 |I_2|^2 = \frac{\omega^2 M^2 R_2 |V_{in}|^2}{2(R_1 R_2 + \omega_0^2 M^2)^2} \\
P_{IN} &= \frac{1}{2} R_1 |I_1|^2 = \frac{R_2 |V_{in}|^2}{2(R_1 R_2 + \omega_0^2 M^2)}
\end{align*}
\]

Using coupling coefficient and quality factor expressions in (A-6) \(K\) and \(Q_1, Q_2\) will lead to the efficiency equation.
\[ K = \frac{M}{\sqrt{L_1 L_2}}, \quad Q_1 = \frac{\omega L_1}{R_1}, \quad Q_2 = \frac{\omega L_2}{R_2} \]  

(A-6)

Power efficiency of two coil WPT is presented by (A-7).

\[ \eta = \frac{P_{OUT}}{P_{IN}} = \frac{\omega^2 M^2}{R_1 R_2 + \omega_0^2 M^2} = \frac{k^2 Q_1 Q_2}{1 + k^2 Q_1 Q_2} \]  

(A-7)