



Design, Analysis and Realization of Compact Energy Efficient RF Wireless Power Transfer Systems

by

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January 2022

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New Delhi 110020



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by

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Submitted

**in partial fulfillment of the requirements for the degree of
Doctor of Philosophy**

to the

Indraprastha Institute of Information Technology Delhi

January 2022

Certificate

This is to certify that the thesis titled “**Design, Analysis and Realization of Compact Energy Efficient RF Wireless Power Transfer Systems**” being submitted by Mr. Shalin Verma to the Indraprastha Institute of Information Technology Delhi, for the award of the degree of Doctor of Philosophy, is an original research work carried out by him under my supervision. In my opinion, the thesis has reached the standards fulfilling the requirements of the regulations relating to the degree.

The results contained in this thesis have not been submitted in part or full to any other university or institute for the award of any degree/diploma.



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Date: January 2022

Place: New Delhi

Dedicated to my Family and Friends

Acknowledgements

First and foremost, I would like to express my deepest gratitude to my supervisor Dr. Mohammad Hashmi for his tremendous support, guidance, and encouragement throughout these five years. I have been indeed lucky to have the opportunity to work with Dr. Hashmi. He has always pushed me for independent research, problem-solving, and collaboration with other esteemed professors. It has helped me a lot to pursue independent research and excel in my research career. I am looking forward to continuing research with him in the future.

My family's support has been enormous throughout this journey. It was my parents who insisted on pursuing research. Their love and support were phenomenal to encourage me for smooth completion of PhD. They have always stood strong for me which gave me more confidence to overcome tough moments. Thank you, mom and dad for everything.

A special mention for my doctoral committee members Dr. Anand Srivastava and Dr. Vivek Bohra, for their feedback and comments. A special thanks to all administrative and academic staff for all kinds of infrastructure and financial support.

Any journey is incomplete without friends. I would like to thank all my friends Dr. Dinesh Rano, Deepayan Banerjee, and Rahul Gupta, for being a support and help throughout these years.



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List of Abbreviations

WPT	Wireless Power Transfer
RF	Radio Frequency
IoT	Internet of Things
RFID	Radio Frequency Identification
EV	Electric Vehicles
WBAN	Wireless Body Area Network
WSN	Wireless Sensor Network
TX	Transmitter
RX	Receiver
IMN	Impedance Matching Network
DB	Dual Band
MSL	Microstrip Line
PTE	Power Transfer Efficiency
EM	Electromagnetic
PBG	Photonic Band-gap
EBG	Electromagnetic Band-gap
DGS	Defected Ground Strucutre
BSF	Band Stop Filter
NF	Near Field
IPT	Inductive Power Transfer
CMT	Coupled Mode Theory
GMA	Geometric Mean Area
SMD	Surface Mounted Device
SMA	Sub Miniature Version A
PCB	Printed Circuit Board
ROF	Roll-Off Factor

SRR	Split Ring Resonator
OSIR	Open Ended Stepped Impedance Resonator
DMS	Defected Microstrip Line
DUT	Device Under Test
CLIN	Coupled Line
WIPT	Wireless Information and Power Transfer
LED	Light Emitting Diode
USRP	Universal Software Radio Peripheral
QAM	Quadrature Amplitude Modulation
BER	Bit Error Rate
UI	User Interface
FoM	Figure of Merit
SIMO	Single Input Multiple Output
MIMO	Multi Input Multiple Output

Abstract

Nowadays, wireless power transfer (WPT) has become a top area of research due to ease and portable outcome in the present technological scenario. This creates an increased demand globally in various technological sectors such as consumer electronics, bio-medical implants, automotive industry and so on. In consumer electronics, wireless power transfer is used in wireless charging of electronic devices. For example, mobile devices are required to be plugged in manually for charging which limits mobility and disrupts use when charge is depleted. In addition, due to compactness in the overall size of the portable device, the connectors become a larger fraction of device size. As a result, the use of wireless power transfer technique eliminates the connector from portable electronic devices which improves both size and reliability. Furthermore, in biomedical implants, the use of non-rechargeable batteries needs replacement at the end of their life span by a costly surgery. Also, the bulky nature of non-rechargeable batteries creates an obstacle in the design of compact implantable devices. These concerns raise the requirement of wireless power transfer technique in modern applications.

Furthermore, wireless power transfer is classified into two different types i.e. near field and far field WPT. This thesis mainly focuses on resonant inductive type near field WPT systems operating in MHz frequency range. This type of WPT system transfers power at higher efficiency and over longer distances than non-resonating WPT systems. In addition, these systems can also transfer power at mid-range separation distance where the distance of power transfer is greater than the size of the resonator. Most importantly, resonant inductive near field WPT is realized using different coil mechanisms i.e. litz/solid wire and printed circuit board (PCB) based spiral coils and planar microstrip line structures such as defected ground structures (DGS). Firstly, the coil mechanism based on litz/solid wire uses two similar coils separated by a distance. This type of coil mechanism has a specific number of turns to obtain a higher value of mutual inductance. The higher the mutual inductance, the greater will be the coupling coefficient which results in power transfer with higher efficiency. Moreover, the litz/solid wire-based coil mechanism mostly operates at single frequency and transfers power at longer distances with optimum efficiency. Secondly, the coil mechanism utilizing printed spiral on the top plane of the PCB can be designed to operate both in single as well as dual frequency band. In addition, it is useful in enabling on-body communication. Finally, the planar microstrip line structure using defect in the ground plane (DGS) is used. The DGS possesses good slow wave characteristics and high electrical length due to which the operating frequency decreases for the fixed physical length. This makes the overall circuit size more compact. Further miniaturization in the circuit size is achieved by adding an external capacitor in the excitation gap of defect. Moreover, DGS

helps in achieving the most compact overall circuit size in comparison to both the litz/solid wire and printed spiral based coil mechanism. Furthermore, WPT is achieved by utilizing two similar DGSs placed in a back-to-back manner and separated by a particular distance. The DGS can be tuned to operate at single and multi-band frequency range. This enables power transfer at different frequency bands simultaneously.

Here, it is important to note that among all the coil mechanisms as explained above, the DGS based WPT system looks more attractive due to compact size at mega-hertz frequency range. The miniaturized size of DGS based WPT system enables the use in biomedical implants such as defibrillators which checks the abnormal heart rhythm, cochlear implants which helps patients to hear for better understanding of speech, retinal prosthesis which restores perception for blind patients and so on. The compact and light weight characteristic of the DGS-WPT system alongwith the rechargeable battery used in biomedical implants puts a barrier on non-rechargeable, big and bulky batteries.

Finally, the objective of this thesis is to address the concerns of size, cost, efficiency and separation distance of the wireless power transfer system. Other objectives include designing of the WPT system operating on multiple frequency bands to enable simultaneous power and data transfer and enabling power transfer using single input-multi output (SIMO) and multi input-multi output (MIMO) wireless topology. Both SIMO and MIMO wireless topologies allow the charging of batteries of multiple portable electronic devices simultaneously.

Publications

Currently Under-Review Article:

- UR[1] S. Verma, D. Rano, and M. S. Hashmi “Dual-Band SIMO and MIMO Based Defected Ground Structure for Wireless Power Transfer Applications,” *IEEE Transactions on Instrumentation and Measurement*.

Peer-Reviewed Published Journal Articles:

- J[1] S. Verma, D. Rano and M. S. Hashmi, “Enhancing the performance of defected ground structure type near-field radiofrequency WPT system by coupled-line impedance matching,” *IET Microwaves, Antennas and Propagation*, vol. 14, no. 12, pp. 1431-1439, Oct. 2020, doi: 10.1049/iet-map.2020.0217.
- J[2] S. Verma, D. Rano, S. Malhotra and M. S. Hashmi, “Measurements and Characterization of a Newly Developed Novel Miniature WIPT System,” *IEEE Transactions on Instrumentation and Measurement*, vol. 70, no. 2004211, pp. 1-11, 2021, doi: 10.1109/TIM.2021.3075537.

Articles Published in Conference Proceedings:

- C[1] S. Verma, D. Rano and M. S. Hashmi, “A novel miniaturized band stop filter using fractal type defected ground structure (DGS),” *IEEE Asia Pacific Microwave Conference (APMC)*, pp. 799-802, 2017, doi: 10.1109/APMC.2017.8251569.
- C[2] S. Verma, D. Rano, M. Hashmi and V. Bohara, “A High Q Dual E-Shaped Defected Ground Structure for Wireless Power Transfer Applications,” *Asia-Pacific Microwave Conference (APMC)*, pp. 1435-1437, 2018, doi: 10.23919/APMC.2018.8617254.
- C[3] S. Verma, E. Sareen and M. S. Hashmi, “A Miniaturized Dual-Band Right Triangle Defected Ground Structure Band Stop Filter for Energy Harvesting Applications,” *IEEE MTT-S International Microwave and RF Conference (IMaRC)*, pp. 1-3, 2018, doi: 10.1109/IMaRC.2018.8877313.
- C[4] S. Verma and M. S. Hashmi, “A Compact Dual Band Bandstop Filter Using Triangular Shaped Defected Ground Structure and Split Ring Type Defected Microstrip Line,”

- IEEE International Conference on Advanced Networks and Telecommunications Systems (ANTS)*, pp. 1-3, 2018, doi: 10.1109/ANTS.2018.8710139.
- C[5] S. Verma, D. Rano and M. Hashmi, “A Novel Dual Band Defected Ground Structure for Short Range Wireless Power Transfer Applications,” *IEEE Wireless Power Transfer Conference*, pp. 188-191, 2019, doi: 10.1109/WPTC45513.2019.9055660.
- C[6] D. Kupreyev, K. Dautov, M. Hashmi and S. Verma, “Design of a Compact DGS based Dual-band RF WPT System for Low-Power Applications,” *IEEE Asia Pacific Conference on Antennas and Propagation (APCAP)*, pp. 48-49, 2019, doi: 10.1109/APCAP47827.2019.9471919.
- C[7] S. Malhotra, S. Verma, V. Bohara and M. Hashmi, “Dual-band WPT system using Semi-H DGS for Biomedical Applications,” *IEEE Asia-Pacific Microwave Conference (APMC)*, pp. 720-722, 2019, doi: 10.1109/APMC46564.2019.9038176.
- C[8] K. Dautov, M. Hashmi, S. Verma and R. Gupta, “Highly-efficient Compact Size DGS-based Wireless Power Transfer for Low-Power Sensor Nodes,” *IEEE MTT-S International Microwave and RF Conference (IMARC)*, pp. 1-4, 2019, doi: 10.1109/IMaRC45935.2019.9118659.

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Chapter 1

Introduction

There have been increasing requirements of near field WPT techniques to meet the demands of emerging applications involving IoT, RFIDs, biomedical devices, electric vehicles (EV), modern electronic gadgets, wireless sensor network (WSN), wireless body area network (WBAN), etc. [1]-[16]. Fig. 1.1 depicts the block diagram of a wireless power transfer (WPT) system depicting near field coupling. It is apparent that the AC input signal is converted to DC in an AC/DC conversion block. Then a high-efficiency switching amplifier converts the DC voltage into an RF signal that is used to drive the transmitter (TX) resonator. The impedance matching network (IMN) ensures that the amplified signal is efficiently coupled to the TX resonator. The magnetic or electric field provided by the TX resonator is coupled to the receiving (RX) resonator. The RX resonator receives the coupled field and transfers the energy to the RF/DC rectifier block through an IMN. Finally, the rectified signal is used to power the load or charge a battery.

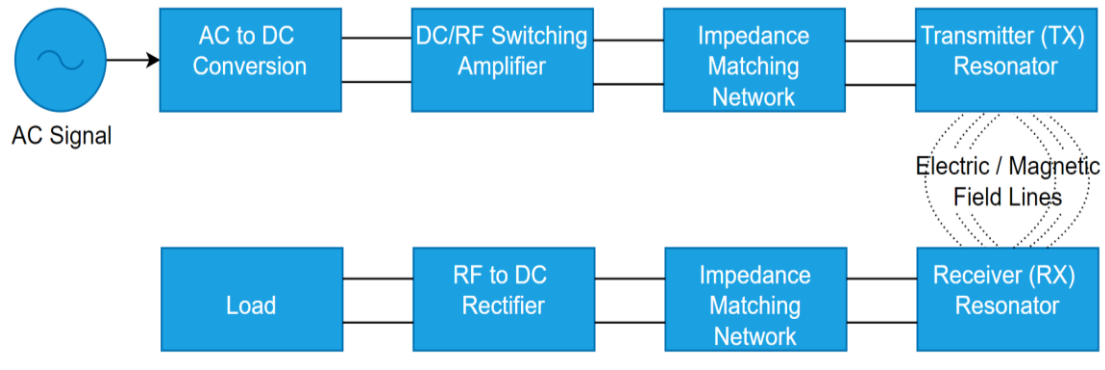


Fig. 1.1. The block diagram of wireless power transfer system depicting near field coupling [19].

Furthermore, the WPT systems based on near field coupling can be categorized as inductively or capacitively coupled for short range applications, and resonant inductively coupled for mid-range applications. The short-range applications which involve inductive coupling, transfer power in lower frequency region with high efficiency. Whereas, capacitive coupling is limited to power transfer at a smaller distance in comparison to inductive coupling with high efficiency [17]. For the emerging applications, usually operating at high frequencies, the resonance-based techniques are preferred as it concentrates power at the resonance frequency

and thereby enabling enhanced performance of WPT systems. Also, they provide the advantage of immunity to neighboring environment and line-of-sight transfer requirement [18].

Near field wireless power transfer (WPT) can be achieved by a variety of different techniques. Firstly, short range power transfer by the means of printed spiral resonator based on strong resonant coupling is shown in [20]. Printed spiral resonators utilizing high Q-factor resonators still have the limitation of power transfer distance that does not exceed the maximum dimension of the resonators. For example, in [21], an H-shaped DGS for WPT is shown by designing H-shaped resonator on an area of $25\text{ mm} \times 25\text{ mm}$ to transfer power with efficiency of 85% at a transmission distance of 5 mm at 1 GHz. These performance metrics have been further improved using strong resonant coupling to achieve WPT efficiency of 70% at effective transmission distance of 9 mm at 1.5 GHz [22]. WPT by the means of strongly coupling is achieved by utilizing high Q intermediate resonators to obtain high power transfer efficiency. However, the use of intermediate resonators increases the system complexity with the reduced power transfer distance.

Secondly, Mid-range power transfer is achieved by semi-H shaped DGS resonators at single frequency band [23]. Mid-range wireless power transfer focusses power at a distance greater than the size of the resonator. Initially, to achieve mid-range power transfer, H-shaped DGS resonator is designed which transfers power at short range. The H-shaped DGS resonator takes an area of $20\text{ mm} \times 20\text{ mm}$ and achieves power transfer with 68% efficiency at effective transmission distance of 13 mm at 300 MHz. From H-shape DGS resonator, Semi-H shape DGS resonator is modelled by taking half of the H-shape DGS resonator into consideration. The Semi-H shape DGS resonator focusses power in mid-range. In this case, the resonators take a board size of $21\text{ mm} \times 21\text{ mm}$, and transfers power with an efficiency of 73% at a transmission distance of 25 mm at 300 MHz.

Lastly, Mid-range power transfer is achieved for two frequency bands i.e. dual-band wireless power transfer. Dual band wireless power transfer (DB-WPT) is achieved by cascading two symmetrical DGS resonators loaded with two different capacitors [24]. Each DGS resonator perturbs the field propagating through the microstrip line (MSL) and resonates at a different frequency. In this work, both the DGS resonators transfers power at two different frequencies i.e. 0.3 GHz and 0.7 GHz simultaneously. Since coupling coefficient at two frequencies of interest is different, for example at 0.3 GHz, coupling coefficient (k_1) is lower while at 0.7 GHz, coupling coefficient (k_2) is higher. As a result, power transfer efficiency (PTE) at two frequencies comes out to be different i.e. at 0.3 GHz, PTE is 71% whereas at 0.7 GHz, PTE is 72%. Also, the mutual coupling between the transmitter and receiver varies at two different frequencies due to change in the values of self-inductances at two different frequencies results in choosing an

optimal distance where power is transferred with optimum efficiency. This work enables power transfer at a distance of 16 mm for two different frequencies simultaneously.

In this dissertation, we have focused on resonant inductive near field WPT system from the biomedical perspective. In addition, we have also designed the WPT system for portable electronic devices and board to board applications. The comprehensive study of various aspects of these scenarios provides sufficient clues that there is a need to address several aspects to incorporate the technological advancements in such a system.

1.1 Literature Survey

1.1.1 Defected Ground Structure

Defected Ground Structure (DGS) is widely used nowadays in active and passive devices. Recent advancements [23]-[30] show that DGS technique enables compactness of the resonator which leads to the development of compact near-field wireless power transfer (WPT) systems. In addition, DGS is used for enhancing the operating bandwidth and gain, reducing the mutual coupling between two networks, suppressing the higher order harmonics and unwanted cross-polarization, and also producing notched band to stop interference with any band. Moreover, this type of structure is integrated on the ground plane with planar transmission line such as microstrip line, coplanar waveguide, and conductor backed coplanar wave guide.

The following section gives a brief summary on DGS by focusing on evolution of the structure, the definition and working principle, types, and equivalent circuit modelling.

1.1.1.1. Evolution

The concept of Defected Ground Structures (DGS) evolved in recent years primarily from the studies of Photonic Band Gap (PBG) structures in electromagnetics. The PBG's employed in electromagnetic (EM) applications, are now referred to as Electromagnetic Band Gap (EBG) structures [31]. They are actually artificial periodic structures exhibiting an unusual property of preventing EM waves from propagating through them over a range of frequencies, termed as "stopband/Band gap" and allowing EM waves to propagate through them over a range of frequencies, termed as "passband." Fig. 1.2 indicates the "band-gap" in transmission caused by an EBG structure.

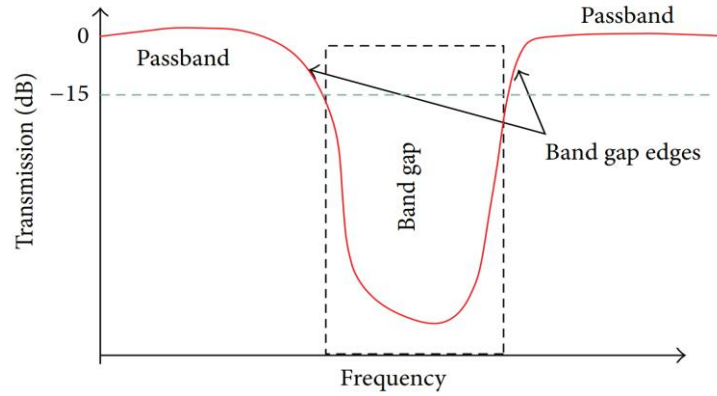


Fig. 1.2. Band-Gap Transmission caused by an EBG structure [32].

For a printed microwave circuit and transmission line, its ground plane appears to be the most suitable choice to implement an EBG as a printed pattern. In 1999, a group of researchers further simplified the geometry and discarded the periodic nature of the pattern. They simply used a unit cell of dumbbell shape to achieve considerable stopband in C and X-bands for a microstrip line which they called it a “PBG unit structure.” In their subsequent article, the same structure was termed as “Defected Ground Structure (DGS)”. Therefore, a DGS may be regarded as a simplified variant of a printed EBG on a ground plane.

1.1.1.2. Definition and Working Principle

The compact geometrical slots embedded on the ground plane of microwave circuits are referred to as Defected Ground Structure (DGS). Furthermore, it can be defined in the following ways:

- A DGS can be described as a unit cell EBG or an EBG with limited number of cells and a period.
- The DGS slots are resonant in nature.
- They have different shapes and sizes with different frequency responses and equivalent circuit parameters.

The working principle of Defected Ground Structure is described as below: -

- The defects on the ground plane perturbs the current distribution of the ground plane.
- This perturbation modifies the characteristics of a transmission line (or any structure) by including some parameters (slot resistance, slot capacitance, and slot inductance) to the line parameters (line resistance, line capacitance, and line inductance).
- In other words, any defect etched in the ground plane under the microstrip line changes the effective capacitance and inductance of microstrip line by adding slot resistance, capacitance, and inductance.

Thus, it influences the guided wave characteristics and is found to exhibit (1) bandgap properties as revealed due to EBG structures; and (2) a slow wave effect, which helps in compacting the printed circuits.

1.1.1.3. Types

The study of DGS can be classified into two categories depending on their configurations:

- Single unit cell DGS and
- Uniform or Non-Uniform periodic arrangement of unit cells.

Here, Fig. 1.3 shows some basic DGS classifications and geometries.

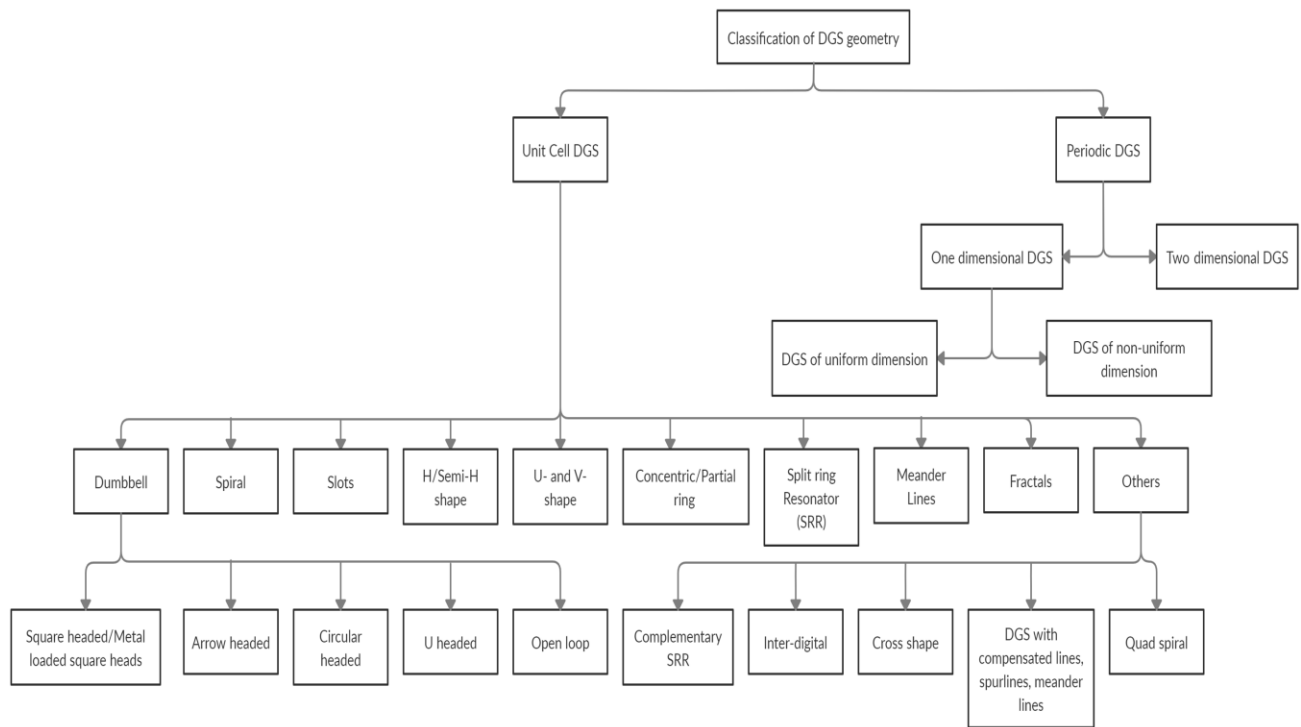


Fig. 1.3. Basic Classification of DGS.

• Single Unit Cell DGS

The DGS geometries reported in literature include simple shapes such as rectangular dumbbell, circular dumbbell, spiral, “U”, “V”, “H”, cross, concentric rings, etc. [31]. Also, different complex structures like split ring resonators, or fractals have been examined. Fig. 1.4 shows some of the most commonly used unit cell DGS’s. These DGS cells are utilized to implement filters, suppress unwanted surface waves, and control harmonics in microstrip antennas, compact microwave circuits and other microwave applications. In addition, these geometries have been explored with the aim of achieving improved performance in terms of stopband and passbands, compactness and ease of design.

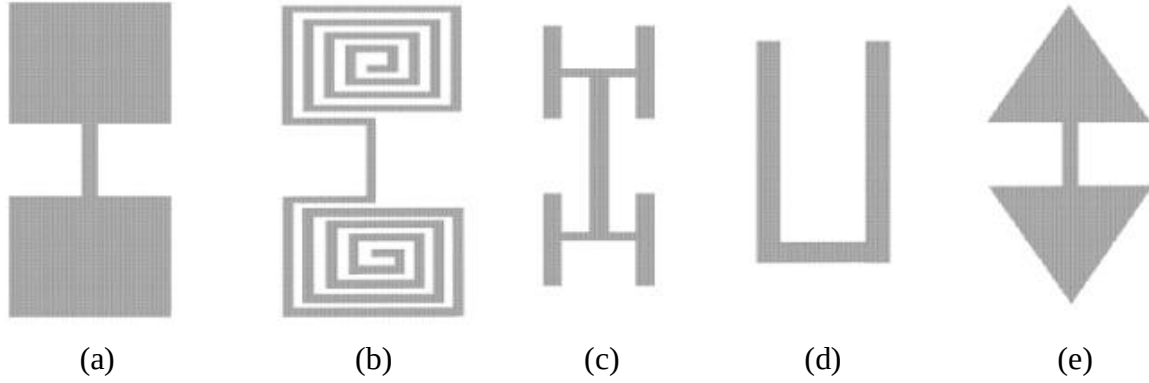


Fig. 1.4. Some Basic Geometries of DGS's. (a) Dumbbell shape. (b) Spiral shape. (c) H-shape. (d) U-shape. (e) arrow head dumbbell.

Amongst all the unit cell DGS's, the first unit cell DGS that was reported by Park et al. and subsequently by Kim et al., was a simple dumbbell-shaped DGS. The dumbbell-shaped DGS consists of two rectangular slots connected by a narrow slot, thus resembling a dumbbell-shaped lattice, etched out on the metallic ground plane of a $50\ \Omega$ microstrip line, as shown in Fig. 1.5(a). The simulated S-Parameter response of the dumbbell DGS operating at a fixed frequency is shown in Fig. 1.5(b) [31]. It can be seen that a simple unit cell DGS shows a distinct stop-band property for the wave propagating through the microstrip line.

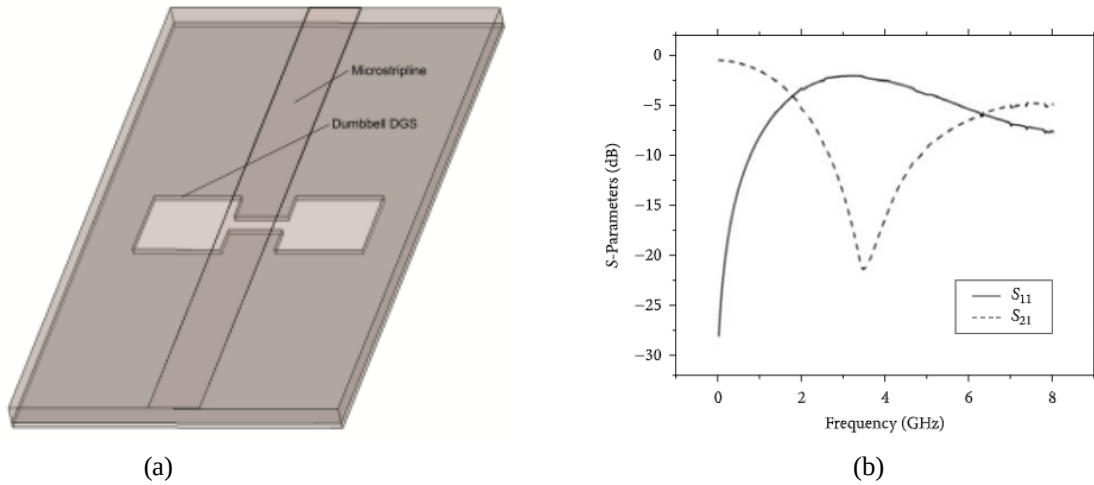


Fig. 1.5. Dumbbell Shaped DGS. (a) Geometry. (b) Simulated S-Parameter response [31].

- **Uniform periodic and non-uniform periodic DGS**

The unit cells are repeated in periodic manner and this helps in achieving deeper, steeper and wider stopband characteristics. A DGS formed by repetition of unit cells is referred to as a Periodic DGS. For example, in Fig. 1.6(a), a five-cell uniform periodic DGS with a $50\ \Omega$ microstrip line printed on its other side is shown. The microstrip line with a periodic structure has a finite pass band and rejection band as low-pass filters. In addition, the increased slow-wave

effect and the additional equivalent components are important properties of periodic structure that can be realized and the circuit size can be reduced using these properties.

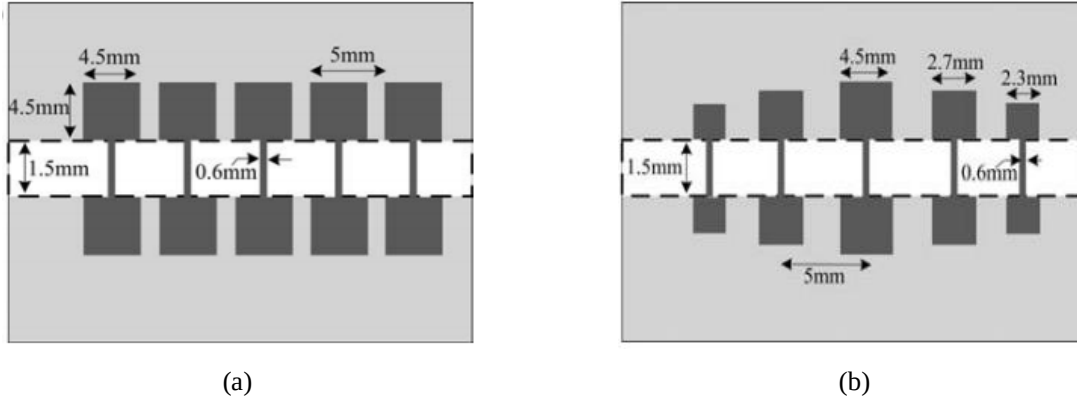


Fig. 1.6. Types of DGS. (a) Uniform periodic DGS. (b) Non-Uniform periodic DGS [31].

Another category is non-uniform periodic DGS as discussed here. By utilizing non-uniform periodic DGS, a better performance in terms of enhancing the stopband-bandwidth and suppression of ripples is obtained. In Fig. 1.6(b), the cell dimensions are varied as $e^{(1/n)}$ relative to the central unit cell, where n is a positive integer indicating the number of the elements on either side including the central one. For example, in Fig 1.6(b), $n = 3$ and the amplitude distribution of the units are:

$$e^{1/3} e^{1/2} e^1 e^{1/2} e^{1/3} \approx 1:396; 1:649; 2:718; 1:649; 1:396$$

Here, the central dumbbell, chosen to have its amplitude “ e^1 ” has 4.5 mm side length. Then the other dumbbell cells have side lengths of 2.7 mm and 2.5 mm, respectively following the exponential distribution mentioned above. Moreover, in comparison to uniform periodic DGS as shown in Figure 1.6(a), the non-uniform variant provides a much wider stopband feature along with suppressed ripples in the passband.

1.1.1.4. Equivalent Circuit Modelling

A quantitative analysis is needed to measure the performance of a DGS or to extend the design by cascading multiple units of the same DGS. To apply the proposed DGS section to a practical circuit design, it is necessary to extract the equivalent circuit parameters. In order to derive the equivalent circuit parameters of DGS unit at the reference plane, the S-parameters vs Frequency should be calculated by full-wave electromagnetic (EM)-simulation to explain the cutoff and attenuation pole characteristics of the DGS section. The circuit parameters for the derived equivalent circuit can be extracted from the simulation result which can be fit for the one-pole Butterworth-type low-pass response. Now, the DGS can be modelled using the three types of equivalent circuit given below:

- (a) Transmission line modeling
- (b) LC and RLC circuit modeling and
- (c) Quasi-Static modeling [31].

Amongst the three types of equivalent circuits given above, the second one appears to be more general and it is relatively straightforward to model a DGS in terms of equivalent parallel LC or RLC circuit.

- **LC and RLC Equivalent Circuits**

Fig. 1.7(a) and Fig. 1.7(b) show the equivalent circuit of the DGS and one-pole Butterworth prototype of the LPF. The equivalent circuit of DGS is represented in terms of parallel LC or RLC circuit which has been discussed below.

For realizing a parallel LC equivalent circuit, an example is given using a dumbbell shaped DGS geometry. In dumbbell shaped DGS, two rectangular defected areas and one connecting slot correspond to the equivalently added inductance (L) and capacitance (C), respectively. As the etched area of the two rectangular head increases, the effective series inductance increases due to the increase in surface current. This increase in series inductance results in decrease of the cutoff frequency. On the other side, a rectangular shaped slot is etched below the microstrip feed line in the ground plane which acts as a parallel plate capacitor giving rise to capacitive effect. The capacitance decreases with the increase in the distance of the etched slot and this causes the location of the attenuation pole to shift to higher resonance frequency. The attenuation pole results due to presence of the etched rectangular shaped gap in the ground plane below the microstrip line. Here, the slot capacitance (C) is in parallel with the inductance (L) which results in resonance at a certain frequency.

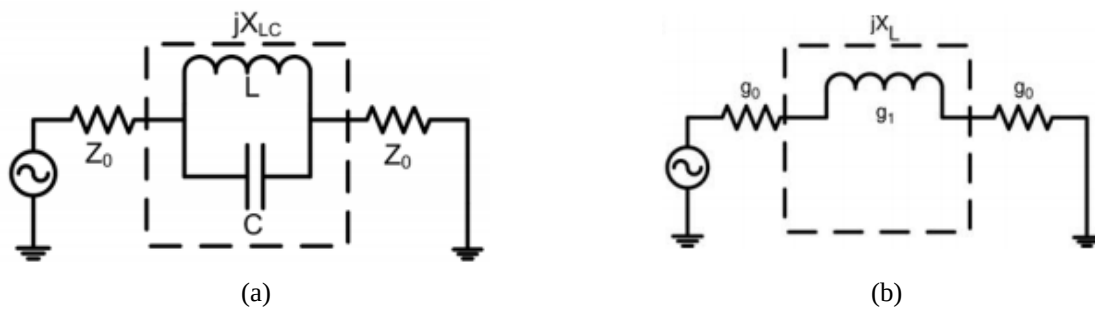


Fig. 1.7. Circuit Diagram. (a) Equivalent LC Circuit. (b) One-Pole Butterworth prototype low pass filter.

The extraction of equivalent L and C values is described below [31]. The reactance of the equivalent circuit of Fig. 1.7(a) is represented by:

$$X_{LC} = \frac{1}{\omega_0 C} \left(\frac{\omega_0}{\omega} - \frac{\omega}{\omega_0} \right) \quad (1.1)$$

where ω_0 is the angular resonant frequency and ω is angular frequency. The reactance of LPF in Fig. 1.7(b) is given by

$$X_L = \omega_1 Z_0 g_1 \quad (1.2)$$

where ω_1 is the normalized angular frequency, Z_0 is input and output port impedances, and g_1 is the ‘prototype element’ as described in [31].

From (1.1) and (1.2) [31] at the cut-off frequency, we have,

$$X_{LC} \big|_{\omega=\omega_c} = X_L \big|_{\omega_1=1} \quad (1.3)$$

$$C = \frac{\omega_c}{Z_0 g_1 (\omega_0^2 - \omega_c^2)} \quad (1.4)$$

$$L = \frac{1}{4\pi^2 f_0^2 C} \quad (1.5)$$

Where f_0 and f_c are resonance (attenuation pole) and cutoff frequency which can be obtained from EM simulation results.

The parallel LC equivalent circuit modeling does not account for any losses caused by radiation or conductor/dielectric losses. A more realistic model takes an equivalent loss resistance R into account as shown in Fig 1.8. This loss resistance R can be extracted from simulated S_{11} employing the following relation.

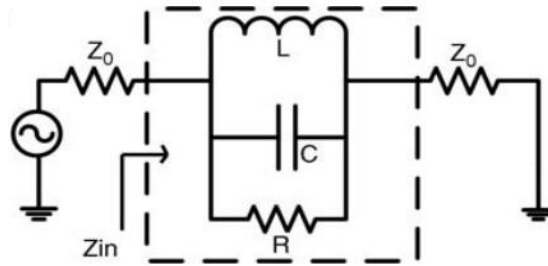


Fig. 1.8. Parallel RLC Equivalent Circuit [31]

$$R = \frac{2Z_0}{\sqrt{\frac{1}{|S_{11}(\omega)|^2 - \left[2Z_0 \left(\omega C - \frac{1}{\omega L}\right)\right]^2} - 1}} \quad (1.6)$$

where,

$$S_{11}(\omega) = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} \quad (1.7)$$

and the equivalent C and L are expressed in Equations 1.4 and 1.5, respectively.

1.1.2. Near Field Wireless Power Transfer (NF-WPT)

Wireless power transfer (WPT) system is an emerging technology that is useful to recharge the battery wirelessly for various portable and biomedical implant devices, battery free sensors, passive RF identification, near-field communications, and many others in near field region. In the near field WPT region, first, the fundamental principle and general block diagram of inductive NF-WPT is explained. Then, three different types of WPT techniques such as inductive power transfer, resonant Inductive wireless power transfer, and capacitive power transfer are briefly summarized.

1.1.2.1 Basic Principle

The fundamental theory of Inductive near field WPT systems is governed by Faraday and Ampere's laws as shown in Fig. 1.9 [33]. The Ampere's law states that, current I generate a magnetic field H . From this law, the magnetic field links the secondary coil and according to Faraday's law, a voltage V is induced.

Ampere's law can be mathematically expressed as [33]:

$$I = \oint H \cdot dl \quad (1.8)$$

This equation states that the line integral of the magnetic field intensity around a closed loop is equal to the current flowing through it.

Faraday's law, on the other hand, is expressed by:

$$V = -N_2 \frac{d\phi}{dt} \quad (1.9)$$

Where N_2 is the number of turns of the secondary coil.

The negative sign in (1.9) is described by Lenz's law, which states that the current flow in the secondary coil (when a load is connected) will be such that it creates a magnetic field that opposes the primary magnetic field.

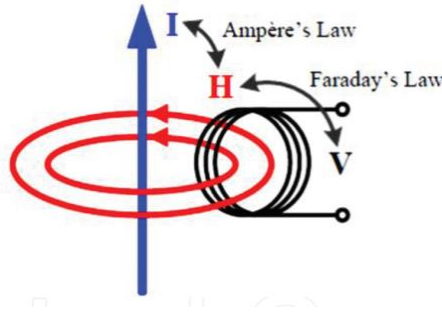


Fig. 1.9 Fundamental Theory of Near Field Inductive WPT

1.1.2.2. Block Diagram

A typical Near Field WPT system is normally constructed by winding the primary and secondary coils into separate halves of ferrite cores parted by an air gap as shown in Fig. 1.10 [33]. Since direct physical contacts are eliminated in such a system, as a result magnetic link is used to provide a reliable and efficient power transfer across an air gap. Here, the magnetic link is the magnetic flux (ϕ_m) which couples from one half of the core to the other half and provides a mutual inductance M that couples energy from the primary to the secondary side. The mutual inductance M is a function of geometry and can be found by simulation, measurement, or modelling the physical structure.

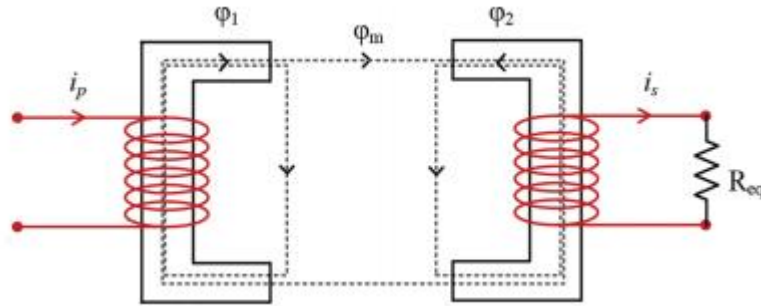


Fig. 1.10 Typical Flux Linkage in Inductively Coupled WPT System.

Also, it can be seen that a loosely coupled wireless power transfer system results in large leakage fluxes which cannot be ignored. Apart from the leakage fluxes, the proportion of the primary coils flux that links with the secondary coil is known as the primary coupling coefficient expressed by [33]:

$$k_1 = \frac{\phi_m}{\phi_1} \quad (1.10)$$

Likewise, the secondary coupling coefficient is given by:

$$k_2 = \frac{\phi_m}{\phi_2} \quad (1.11)$$

These two values are typically combined to give the overall system coupling coefficient as:

$$k = \sqrt{k_1 k_2} \quad (1.12)$$

The mutual inductance between the primary and the secondary coils then is given by:

$$M = k\sqrt{L_p L_s} \quad (1.13)$$

Where L_p and L_s are the self-inductances of the primary and the secondary coils.

1.1.2.3. Inductive Coupling

The Inductive power transfer (IPT) happens when a primary coil of an energy transmitter generates predominantly varying magnetic field, and then the linkage magnetic flux induces the voltage at the secondary coil of the energy receiver. Here, the transmitting coil is excited at the oscillating frequency and the power is transmitted to the receiving coil. The induced voltage in the receiving coil enables wireless power transfer as shown in Fig. 1.11. This induced voltage can be used for charging a wireless device or storage system. Also, it is important to know that operating frequency of inductive coupling is typically in the kHz range.

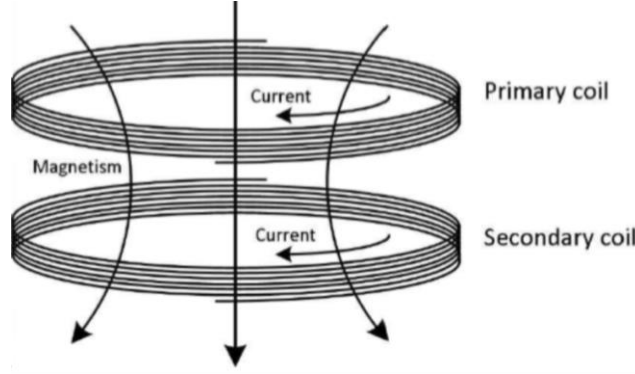


Fig. 1.11. Inductive Coupling [19]

Some of the important applications in the field of Inductive coupling WPT such as,

- In [34], an improvement in the transfer distance to 5 m between primary and secondary coils using magnetic dipole-type coils with ferrite cores based on the inductive coupling was shown. In addition, the experimental results show that the transfer efficiencies are 29%, 16%, and 8% for transfer distances 3, 4, and 5 m at 20 kHz, respectively.
- Ye et al. [35] designed a new approach of inductive coupling that can transfer energy with a relatively greater distance than the traditional method. They proposed a new U-coil system method that consists of three coils, namely, a primary coil, a secondary coil, and a relay coil placed between the primary and secondary coils and that passes the energy

between them. The researchers describe this method (i.e., U-coil WPT) to further improve energy efficiency than the two-coil method. This method achieves 100 cm of energy transfer between transmitter and receiver coils with working frequency of 85 KHz. Compared with the two-coil method, the power transfer efficiency of the U-coil method improved to 66%.

1.1.2.4. Resonance Inductive Coupling

Magnetic resonance coupling as shown in Fig. 1.12 is based on evanescent wave coupling which generates and transfers electrical energy between two resonant coils through varying or oscillating magnetic fields. As two resonant coils, operating at the same resonant frequency are strongly coupled, high energy transfer efficiency can be achieved. Magnetic resonance coupling also has the advantage of immunity to neighboring environment and line-of-sight transfer requirement.

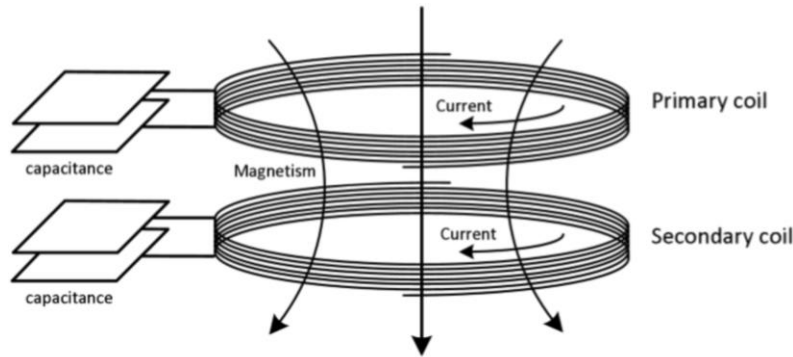


Fig. 1.12. Magnetic Resonance Coupling [19]

As the magnetic resonance coupling typically operates in the megahertz (MHz) frequency range, the quality factors are normally high. In addition, with the increase of power transfer distance, the high-quality factor helps to mitigate the sharp decrease in coupling efficiency, and thus power transfer efficiency. Consequently, extending the effective power transfer distance to meter range is possible.

Some of the important applications in the field of Resonant Inductive coupling WPT such as,

- In reference [36], a variable coupling method to accomplish high efficiency in a resonance coupled WPT system was proposed. The WPT system comprises four identical coils. Two coils are found on each side of the transmitter (source) and receiver (load). In addition, both resonators were sensibly adjusted to reach the identical resonant frequencies. Therefore, a 6.7 MHz self-resonant frequency is generated in all coils. By applying the matching condition, the efficiency of WPT is improved. The matching condition is achieved by changing the coupling factors among the internal resonator and load. By applying the matching techniques experimentally, the energy transfer efficiencies are developed to 46.2% and 29.3% at a distance of 60 and 100 cm,

respectively. However, the maximum efficiency was computed at 15 cm and was found to be 92.5%. Experimental outcomes revealed that the energy transfer efficiency increases dramatically by modifying its coupling factors across distance variations.

- Cannon et al. [37] presented a lumped capacitor circuit for the WPT technique where the source or transmitter operates at a resonant frequency of 8.3 MHz and power transfer from transmitter coil to receiver coil is 28.2 mW. The power transfer at the given value corresponds to 1.68 Volts at the transmitter coil. Finally, the results disclosed that when the receiver coil is placed 17 cm apart from the transmitter coil, the voltage drops from 3 V to 0.57 V, thereby corresponding to 3.3 mW energy transferred at the receiver coil. Here, the wireless power transfer is achieved by a large transmitter and either one or two small receiver coils via magnetic resonant coupling. In addition, a circuit design is established to identify the scheme with one receiver coil and expanded to identify the method with two receiver coils.

Note: Our work involves near field resonant inductive WPT only. So, the above explanation does not include Capacitive near Field WPT.

1.1.3. Performance Metrics of Wireless Power Transfer

In WPT, certain metrics such as frequency, transmitted power, efficiency, and transfer distance can be considered key parameters in determining the overall system performance. These key parameters are explained in detail in the following.

1.1.3.1. Frequency

The design of near-field WPT operating frequency has been implemented with regulatory constraints and a number of technical limitations. The operating frequency varies differently in both the coupling methods i.e. inductive coupling and magnetic resonant coupling.

- In inductive coupling, the operating frequency varies in low frequency (LF) bands. For example, in [38], the authors adopted a frequency range of 20–40 kHz when the planned distance from coil to coil is roughly 10 cm.
- In magnetic resonant coupling, the operating frequency varies in high-frequency (HF) bands. For example, in [39], a frequency range of 0.1–50 MHz was considered. In the magnetic resonant coupling method, high efficiency with a larger distance between the transmitter and receiver can be achieved when their resonance frequencies are the same.

The experimental results in [40] show that frequency control can increase the transfer distances. For example, when the frequency band is 2 MHz, the transfer distance can be approximately 50 cm, whereas the transfer distance can improve to 99.5 cm when the frequency band is adjusted to

30 MHz. Consequently, when the operating frequency increases to several GHz, the transfer efficiency can be improved and the receiving coil size can be minimized.

1.1.3.2. Efficiency of Power Transfer

Power Transfer efficiency is defined as the power received at the receiver coil divided by the total input at the transmitter coils. The efficiency of the WPT system is categorized on the basis of two types:

- **Near field and Far field Effect**

The near field consists of two distinct coupling i.e. inductive coupling and magnetic resonant inductive coupling [41]. The inductive coupling technique achieves an energy transfer efficiency of 70–90 %. The efficiency decreases with the distance between primary and secondary coils. To obtain high efficiency, accurate alignment between primary and secondary coils is required. The magnetic resonant coupling technique has a medium efficiency of 40–60% and also decays with distance. However, the transfer efficiency drops as it differs from the resonance frequency.

In the far Field, the power is transferred wirelessly using microwaves and LASER. The efficiency of the microwave and laser is less than the previous two techniques by approximately 30%. Furthermore, it decreases extremely quickly as the separation between the receiving and transmitting coils increases. The Fig. 1.13 shows the transfer efficiency of the different WPT techniques for near and far fields [41].

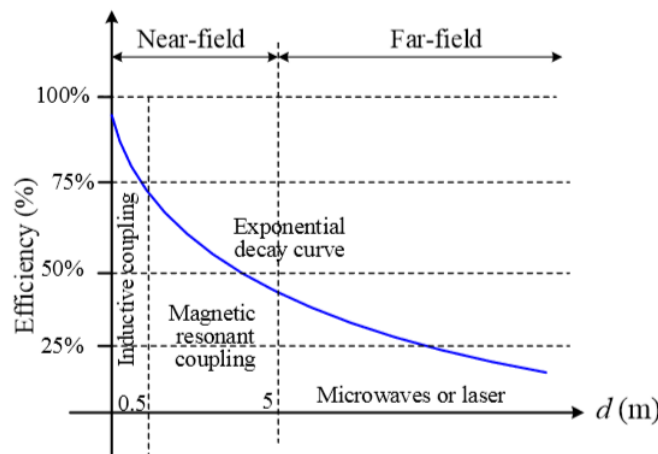


Fig. 1.13. Comparison of Efficiency for different WPT techniques

- **U-Factor Method**

The efficiency of the near field inductive and magnetic resonant coupling methods can be increased with a high-quality factor of the coils and the coupling coefficient between coils. The

quality factor (Q) and coupling coefficient (k) are two important parameters which are responsible for the efficiency of WPT system. Both of these parameters are explained by U-factor method which is derived from coupled mode theory (CMT) to represent the efficiency of WPT system. Here, efficiency is a function of U-factor. The U-factor is defined by,

- The product of coupling coefficient and unloaded quality factor represented mathematically by [19]:

$$U = kQ \quad (1.14)$$

- The ratio of coupling to voltage reflection coefficient represented mathematically by [19]:

$$U = \frac{k}{\sqrt{\Gamma_1 \Gamma_2}} \quad (1.15)$$

Where Γ_1 and Γ_2 depicts the loss of transmitter coil and receiver coil respectively.

To obtain higher efficiency, either of the two parameters i.e. coupling coefficient or unloaded quality factor should be greater. For example, in inductive power transfer, the lower separation distance between transmitter and receiver results in higher coupling coefficient, therefore higher power transfer efficiency whereas, in resonant inductive power transfer, the separation distance is greater than inductive power transfer. This causes the coupling coefficient to decrease with distance. In order to compensate the decrease in coupling coefficient, higher unloaded quality factor is required to obtain higher power transfer efficiency.

1.1.3.3. Transfer Distance

The transfer distance varies from one type to another near-field WPT, where it depends on the uses of the power transfer technique. The inductive coupling technique has limited transfer distance with high transfer efficiency, whereas for magnetic resonant coupling, the energy can be transferred at midrange from source to load relative to the inductive coupling method. However, in this method, the transfer efficiency will be reduced. In addition, some of the works have been reported recently targeting mid-range power transfer using magnetic resonant coupling such as:

- Hekal et al. [23] have proposed the coupled defected ground structure (DGS) WPT systems. In [23], a single band resonant coupled defected ground structure based WPT system is proposed. This system shows power transfer at 300 MHz operating frequency using H-shape and Semi-H shape DGS at transfer distance of 13 mm and 25 mm respectively. The H-shape DGS is fabricated with board size of 20 x 20 mm² and transfers power at a distance of 13mm. This occurs as a result of equivalent inductance in two parallel rectangular loops is half of one of them. In addition, current is flowing in two rectangular loops in opposite directions which causes the mutual coupling between transmitter and receiver to reduce. These effects cause the power transfer distance to be

reduced to half. Whereas Semi-H shape DGS takes a board size of $21 \times 21 \text{ mm}^2$ and transfers power at a distance of 25mm. This occurs as a result of current flow in single rectangular loop which causes the inductance to become twice. Furthermore, it leads to higher mutual coupling between transmitter and receiver in comparison to H-shape DGS. The higher mutual coupling and double inductance value causes increased power transfer distance which is greater than the size of the DGS resonator.

- Tahar et al. [24] have proposed the coupled defected ground structure (DGS) resonators taking an area of $30 \text{ mm} \times 15 \text{ mm}$, while achieving a WPT efficiency of more than 71% at a separation distance of 16mm for both bands (300 MHz and 700 MHz).

Now, the next section focusses on different types of equivalent circuit of the WPT system.

1.1.4. Equivalent Circuits of WPT System

The equivalent circuits shown below are discussed for symmetric WPT system where the transmitter size and receiver size are taken to be similar.

1.1.4.1. Coupled Resonance

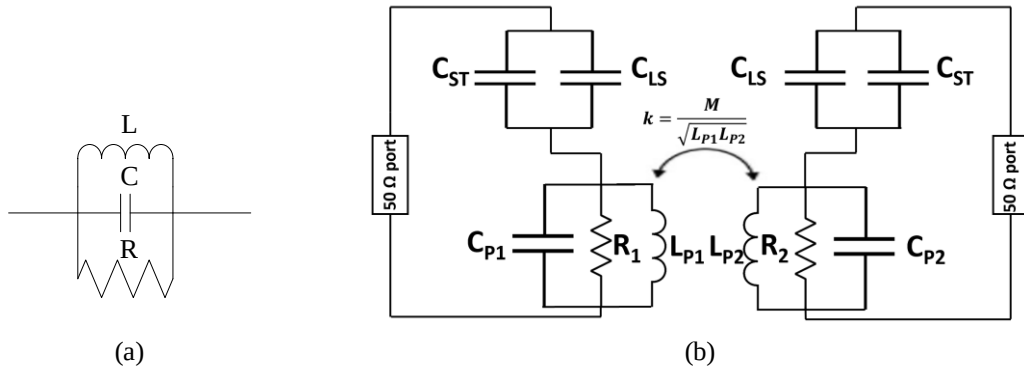


Fig. 1.14. Equivalent Circuit. (a) Transmitting/Receiving Coil. (b) Resonant Coupled WPT System.

Here, Fig. 1.14(a) represents the parallel RLC equivalent circuit of transmitting and receiving coil and Fig. 1.14(b) shows a resonant coupled WPT system which consists of transmitter (represented by parallel combination of R_1 , L_{P1} , and C_{P1}) coil and receiver (represented by parallel combination of R_2 , L_{P2} , and C_{P2}) coil [19]. Here, the external capacitor C_{P1} and C_{P2} used, forms the compensation network which are responsible for resonance at the desired frequency. In addition, the transmitter and receiver are mutually coupled to each other defined by magnetic coupling coefficient k . Furthermore, the transmitter and receiver are separated at a distance and individually connected in series with symmetric immittance inversion circuit which is depicted by open stub C_{ST} and external capacitor C_{LS} . The immittance inversion circuit provides impedance matching for maximum power transfer from transmitter to receiver.

Lastly, the overall circuit consisting of transmitter, receiver, and Immittance inversion circuit are terminated by 50 Ω port.

1.1.4.2. π - Type

The coupled resonant circuit shown in Fig. 1.14(b) can be further represented in a simplified form by π - type equivalent circuit as shown in Fig. 1.15 [19]. The expressions shown in (1.16) and (1.17) calculate the equivalent circuit parameters of the π model (L_m , L_a and L_b). Here, L_a and L_b represents the inductance introduced by the transmitter coil and receiver coil. The transmitter coil and the receiver coil are symmetric so as to resonate at the design frequency. In addition, both the coils are mutually coupled with each other which is given by mutual inductance L_m . Furthermore, the source (transmitter) end and load (receiver) end are connected with immittance inversion (C_{S1} and C_{S2}) and compensation (C_{P1} and C_{P2}) network to ensure appropriate impedance matching and operating frequency. The overall circuit including the transmitter, receiver, and their immittance inversion and compensation network is terminated by R_S and R_L . The source resistance R_S and load resistance R_L are 50 Ω SMA feeding ports. The source and load impedances are taken as 50 Ω so that the wireless power transfer system can be conveniently measured with 50 Ω based measurement equipment. A design equation is given below to mathematically represent π - type equivalent circuit in reference to the coupled WPT system.

$C_{S1} = C_{S2} = C_{ST} + C_{LS}$, where C_{ST} represents capacitance of stub and C_{LS} represents external capacitor. C_{LS} is used when stub does not provide enough capacitance for impedance matching.

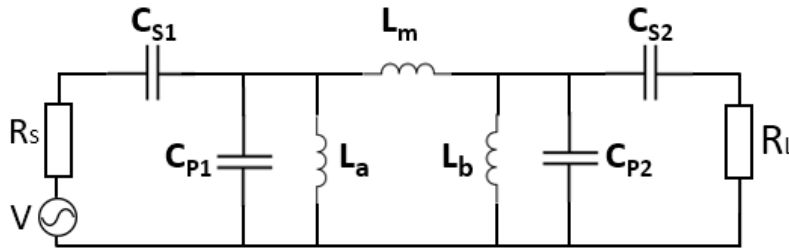


Fig. 1.15. π -Type Equivalent Circuit

$$L_1 = L_a \parallel L_m \rightarrow L_1 = L_{P1} - \frac{M^2}{L_{P2}}$$

$$L_2 = L_b \parallel L_m \rightarrow L_2 = L_{P2} - \frac{M^2}{L_{P1}} \quad (1.16)$$

$$L_a = \frac{L_{P1}L_{P2} - M^2}{L_{P2} - M} \quad L_b = \frac{L_{P1}L_{P2} - M^2}{L_{P1} - M} \quad L_m = \frac{L_{P1}L_{P2} - M^2}{M} \quad (1.17)$$

1.1.4.3. *J*-Inverter

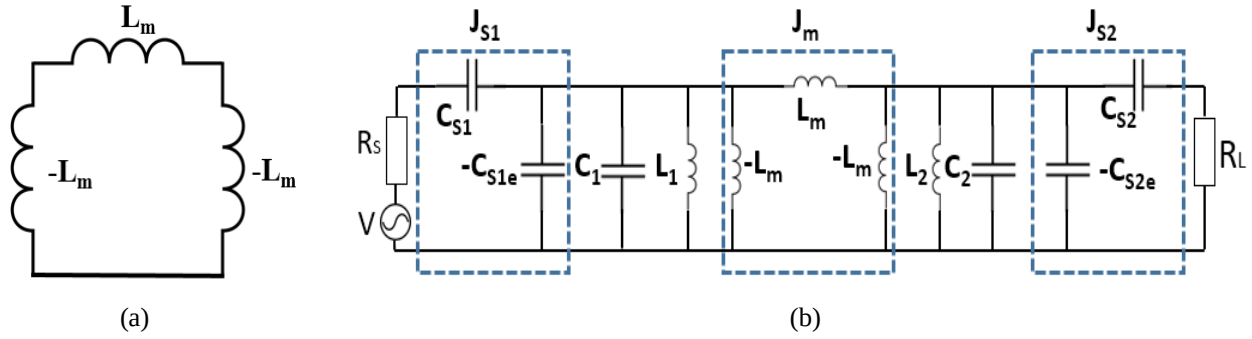


Fig. 1.16. Equivalent Circuit: (a) *J*-Inverter Circuit. (b) Coupled WPT system using *J*-Inverter.

The parallel RLC equivalent circuit of transmitter and receiver as shown in Fig. 1.14(a) can be modelled as admittance (*J*) inverter circuit as shown in Fig. 1.16(a). In Fig. 1.16(b), two resonant coils which are resonating at the same frequency is shown. The resonant coils (L_1C_1 and L_2C_2) are modelled as parallel resonators, since the input impedance of each resonant coil is the maximum at the resonant frequency. The transmitting and the receiving coil are connected to source and load via admittance (*J*) inverter with inverter values J_{s1} and J_{s2} respectively. In addition, two coils (L_1 and L_2) are also connected to each other via an admittance inverter with an inverter value of J_m . For better understanding of the equivalent circuit in Fig. 1.16(b), Fig. 1.17 shows equivalent circuits of the admittance (*J*) inverters [19].

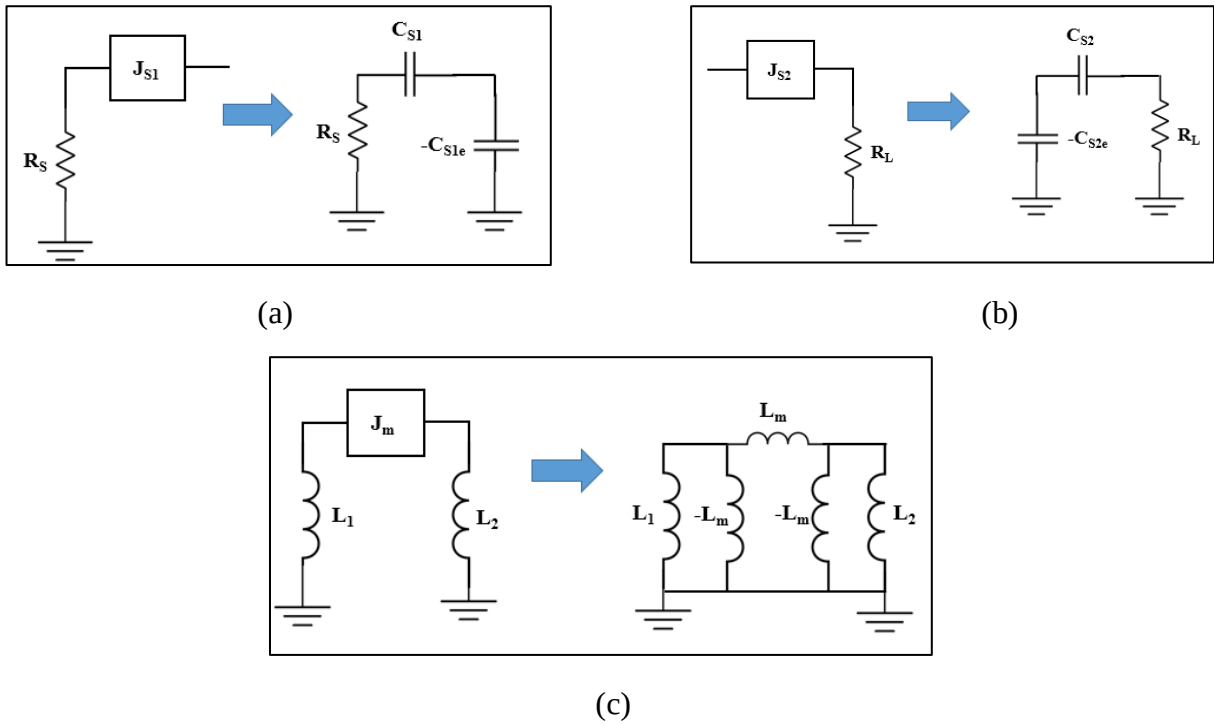


Fig. 1.17. Equivalent Circuit of *J*-Inverters.

The equivalent circuit shown in Figs. 1.17(a) and 1.17(b) are valid provided the following conditions are satisfied.

$$\begin{aligned} C_{S1} &= \frac{J_{s1}}{\omega_0 \sqrt{1 - (J_{S1} R_S)^2}} & C_{S2} &= \frac{J_{s2}}{\omega_0 \sqrt{1 - (J_{S2} R_L)^2}} \\ C_{S1e} &= \frac{C_{S1}}{1 + (\omega_0 C_{S1} R_S)^2} & C_{S2e} &= \frac{C_{S2}}{1 + (\omega_0 C_{S2} R_L)^2} \end{aligned} \quad (1.18)$$

where ω_0 is the angular resonant frequency of the resonators.

The derived circuit parameters (C_1 , L_1 and C_2 , L_2) are related to J -inverter and these parameters are listed in (1.16) and (1.19) respectively.

$$C_{P1} = C_1 - C_{S1e} \quad C_{P2} = C_2 - C_{S2e} \quad (1.19)$$

Moreover, J_m can be determined using the equation in (1.20). This proves inter resonator independent coupling.

$$J_m = \frac{1}{\omega_0 L_m} \quad (1.20)$$

Most importantly, perfect impedance matching at input port leads to maximum power transfer from source to load. In Fig. 1.16(b), the condition for perfect impedance matching at input port is computed by solving input admittance at inverter input as shown by equation (1.21).

$$Y_{S1} = Y_{in} = \frac{J_{S1}^2 J_{S2}^2 R_L}{J_m^2} = \frac{1}{R_S} \quad (1.21)$$

Furthermore, the WPT system will have maximum power transfer from source to load at resonance frequency if the condition shown in (1.22) is satisfied.

$$J_{S1}^2 J_{S2}^2 = \frac{J_m^2}{R_S R_L} \quad (1.22)$$

Therefore, it can be seen that J -Inverter circuit provides an easier platform to model the WPT system. Since the design equations for WPT system are more complex in nature. As a result, it becomes time consuming and error prone to calculate the design equations. At this point, J -inverter equivalent circuit helps in obtaining the conditions for perfect impedance matching or the maximum power transfer by providing concise and analytic design equations.

1.2. Problem Statement

After an extensive research survey of near field resonant inductive type near field WPT system design, it was identified that the existing designs have severe limitations in most of the performance metrics namely compactness [20], separation distance [21], impedance matching [23], efficiency [42] and misalignment [43]. A compact size WPT system is required for biomedical implantable devices which are in the order of milli-metre (mm) to few centi-metres (cms) [44]. To obtain compactness in the overall size of the WPT system, defected ground structure (DGS) is used. DGS comes in various shapes such as rectangular/circular dumbbell, H, semi-H, spiral, split ring, fractal, triangular, circular and so on [31]. Each DGS shape has different inductance and capacitance value to obtain resonance characteristics. Moreover, further miniaturization in the circuit size can be achieved by utilizing an external capacitor in the ground plane of the DGS structure.

To maintain a separation distance between the transmitter and the receiver greater than the overall size of the DGS resonator is the biggest challenge. In [21][23], it can be seen that H-shape WPT system transfers power at a distance less than the size of the DGS resonator. The separation distance is directly related to the size of the resonator. In symmetric resonator, where the transmitter and receiver resonator have similar size, the separation distance is equivalent to the diameter of both the transmitter and receiver structures [19]. On the other hand, in asymmetric resonator, where the transmitter and receiver resonator have different size, the separation distance is equivalent to the square root of the geometric mean area (GMA) [19]. Furthermore, to overcome this challenge, the shape of the DGS resonator needs to be modified such that mutual inductance between the TX and RX resonator is high and the current flowing in the DGS resonator is in single direction.

Impedance matching in a near field WPT system becomes a challenge when it comes to factors such as the complexity and robustness. Both of these factors can be observed in [23] where an external capacitor is soldered between top and bottom plane to obtain perfect matching. In literature, the complexity and robustness can be overcome by using double stub impedance matching approach [24]. However, this approach is only shown for WPT system operating at two frequencies and not for single band WPT system.

Efficiency is another challenge while designing the near field WPT system. It is apparent from [42] that the efficiency is low at critical coupling distance. To overcome this, a proper balance of two key parameters i.e. coupling coefficient (k) and quality factor (Q_U) requires to be maintained [23]. The coupling coefficient (k) is defined as the ratio of the mutual inductance between the TX and RX DGS resonators to the self inductance of individual resonator. On the other hand, the quality factor (Q_U) of an individual DGS resonator is directly dependent on the self inductance and inversely dependent on the resistance. Therefore, the maximum efficiency is extracted by optimizing the dependent and inversely dependent parameters of both coupling

coefficient and quality factor. Furthermore, in a WPT system, the efficiency is degraded due to the misalignment in the position of the RX resonator with respect to the TX resonator [43]. There are different methods proposed in the literature to handle the effects of misalignment in order to maximize WPT system efficiency [26].

1.3. Major Contributions of this Thesis

This thesis has been pursued to contribute into this direction and to address most of the concerns such as the design of compact DGS resonator for WPT applications, development of miniaturized and highly efficient resonant inductive near field WPT system, building a new type of impedance matching technique for a WPT system, designing the system to operate on triple band in order to achieve wireless information and power transfer (WIPT) and lastly, forming a single input multiple output (SIMO) and multi input multiple output (MIMO) WPT system for portable electronic devices.

A. Miniaturized DGS resonator for WPT Applications

Initially, different types of compact DGS resonators based on single band and dual band have been developed. The single band DGS resonators are based on fractal type geometry. Whereas, the dual band DGS resonators consists of cascaded right triangle, a combination of triangular DGS and split ring type defected microstrip line (DMS).

The major contributions are listed below:

- To design and develop miniaturized DGS resonators by utilizing different types of geometries.
- Utilizing closed-form design equations to extract values for the equivalent circuit.
- Verification of the results obtained from both circuit simulation and EM simulation to ensure the effectiveness of the design.

The corresponding publications are [C1, C3, C4].

B. Compact Highly Efficient Resonant Inductive WPT Systems

Further, two compact resonant inductive type near field WPT system are designed using DGS resonators. The first one consists dual E shape based DGS resonator operating on a single band. Whereas, the second one, utilizes a unit cell U shape based DGS resonator operating on two different bands. Both dual E shape and U shape based DGS resonators are utilized to design a WPT system operating on single and dual frequency bands.

The major contributions are listed below:

- To design and develop a compact WPT system utilizing high quality factor E shape DGS resonator.
- For E shape DGS WPT system, the parameter values of equivalent circuit are computed using the closed form design equations and the theory of J-inverter.
- For U shape DGS WPT system, the parameter values of equivalent circuit are computed using the closed form design equations and the theory of impedance matrix.

The corresponding publications are [C2, C5].

C. Impedance Matching technique: Coupled Line

In this chapter, a coupled line based impedance matching technique to enhance the efficiency and robustness of the near field resonant inductive DGS based WPT system is discussed. In the WPT system, the triangular shaped DGS is utilized which provides higher inductance value in comparison to other types of DGS resonator such as rectangular, square and circular. A complete design procedure to develop the WPT system using coupled line matching have been shown.

The major contributions are listed below:

- To design and develop a DGS based WPT system in conjunction with coupled line.
- To develop a systematic analytical approach for impedance matching and a simplified design procedure for the WPT system.
- To develop an equivalent circuit model consisting of parallel LC network and coupled lines for the evaluation of the proposed WPT system design technique.

The corresponding publication is [J1].

D. Wireless Information and Power Transfer (WIPT) System: Triple Band Operation

In this chapter, a triple band near field resonant inductive DGS WPT system has been shown. The lower most frequency is used to transfer power and upper two frequencies provide uplink and downlink of data transfer. The power transfer is verified using an LED experiment where an LED bulb is lit at critical coupling distance. On the other hand, the data transfer is demonstrated using universal software radio peripheral (USRP) where a string of data is transmitted at critical coupling distance. The string received at the receiver is verified using bit error rate and constellation diagram.

The major contributions are listed below:

- To design and develop a tri-band DGS wireless information and power transfer (WIPT) system.
- To develop measurement and characterization setup for power and data transfer using the WIPT system.
- To practically realize power and information transfer using measurement setup.
- To focus on challenges experienced during the measurement of WIPT system.

The corresponding publication is [J2].

E. Single Input Multiple Output (SIMO) and Multi Input Multiple Output (MIMO) WPT System

In this chapter, a SIMO and MIMO dual band DGS-WPT systems are discussed. Both the systems are designed to charge multiple portable electronic devices. The efficiency of both the WPT systems are compared while maintaining similar size of transmitter resonator and receiver resonator.

The major contributions are listed below:

- To design and develop two different types of WPT systems.
- To develop a equivalent circuit using the theory of impedance/admittance matrix and quasi-static approximation for the evaluation of WPT system design technique.
- To identify cross-coupling in both the SIMO and MIMO WPT systems.
- Comparing the efficiency of both the WPT systems with similar overall dimensions.

The corresponding publication is [UR1].

1.4 Dissertation Outline

The rest of the dissertation is divided into 7 chapters. In this section, we discuss the contents of each chapter briefly.

Chapter 2: In this chapter, different types of compact DGS resonators based on single band and dual band have been designed and developed. The single band DGS resonators are based on fractal type geometry. Whereas, the dual band DGS resonators consists of cascaded right triangle, a combination of triangular DGS and split ring type defected microstrip line (DMS). Moreover, closed form design equations have been utilized to extract the values of the equivalent circuit. The results obtained from the EM simulation performed in CST Microwave studio are verified with the circuit simulation results obtained from ADS design tool to ensure the effectiveness of the design. Furthermore, the prototype is fabricated and measurements are performed. The design exhibits excellent agreement between the simulated and measurement results.

Chapter 3: In this chapter, we discuss two compact resonant inductive type near field WPT systems using DGS resonator. The first one consists of high quality (Q) factor dual E shape based DGS resonator operating on a single band. Whereas, the second one, utilizes a unit cell U shape based DGS resonator operating on two different frequency bands. Moreover, in dual E shape DGS WPT system, the parameter values of equivalent circuit are computed using the closed form design equations and the theory of J-inverter. Whereas, in U shape DGS WPT system, the parameter values of equivalent circuit are computed using the closed form design equations and the theory of impedance matrix. Furthermore, the EM simulated results are validated using circuit simulation and measurement results to ensure the correctness of the proposed technique.

Chapter 4: Here, a coupled line based impedance matching technique to enhance the efficiency and robustness of the near field resonant inductive DGS based WPT system is discussed. In the WPT system, the triangular shaped DGS is utilized which provides higher inductance value in comparison to other types of DGS resonator such as rectangular, square and circular. A complete design procedure to develop the WPT system using coupled line matching have been shown. A novel equivalent circuit model consisting of parallel LC network and coupled lines is also developed for the evaluation of the proposed WPT system design technique. Moreover, the fabricated prototype operating at 300 MHz developed on Rogers RO4350B substrate achieves a peak efficiency of 80% at a transmission distance of 17 mm. In addition, the figure of merit is evaluated and compared with other state of the art designs.

Chapter 5: In this chapter, a triple band near field resonant inductive DGS WPT system operating at 300, 433 and 700 MHz has been shown. The 300 MHz frequency is used to transfer power and 433 MHz and 700 MHz frequencies provide downlink and uplink of data transfer. The power transfer is verified using an LED experiment where an LED bulb is lit at critical coupling distance. On the other hand, the data transfer is demonstrated using universal software radio peripheral (USRP) where a string of data is transmitted at critical coupling distance. The string received at the receiver is verified using bit error rate and circular QAM constellation diagram. Furthermore, the measurement carried out on a prototyped system shows an excellent agreement between the measured and simulation results and provides power transfer efficiencies of 70.1%, 66%, and 65% at 300 MHz, 433 MHz, and 700 MHz respectively at a separation distance of 23 mm.

Chapter 6: In this chapter, a SIMO and MIMO dual band resonant inductive DGS-WPT systems operating at 300 MHz and 700 MHz are discussed. Both the WPT systems are characterized

using a novel equivalent circuit realized using the theory of impedance/admittance matrix and quasi-static approximation for the evaluation of the proposed WPT system design technique. The measurement carried out on a prototyped system shows an excellent agreement between the measured and simulation results. The power transfer efficiency of SIMO WPT system at 300 MHz and 700 MHz is 68.7% and 70%. While, in MIMO WPT system, the efficiency is equivalent to 75.1% and 73.2% at 300 MHz and 700 MHz respectively. The separation distance is fixed at 18 mm in both the WPT systems. Furthermore, the efficiency of both the WPT systems are compared while maintaining similar size of transmitter resonator and receiver resonator.

Chapter 7: In the last chapter of this dissertation, we conclude our present research work and have listed the future research in this direction.

Chapter 2

Compact DGS Resonator for WPT Applications

This chapter is based on the following research papers:

- S. Verma, D. Rano and M. S. Hashmi, “A novel miniaturized band stop filter using fractal type defected ground structure (DGS),” *IEEE Asia Pacific Microwave Conference (APMC)*, pp. 799-802, 2017, doi: 10.1109/APMC.2017.8251569.
- S. Verma, E. Sareen and M. S. Hashmi, “A Miniaturized Dual-Band Right Triangle Defected Ground Structure Band Stop Filter for Energy Harvesting Applications,” *IEEE MTT-S International Microwave and RF Conference (IMaRC)*, pp. 1-3, 2018, doi:10.1109/IMaRC.2018.8877313.
- S. Verma and M. S. Hashmi, “A Compact Dual Band Bandstop Filter Using Triangular Shaped Defected Ground Structure and Split Ring Type Defected Microstrip Line,” *IEEE International Conference on Advanced Networks and Telecommunications Systems (ANTS)*, pp. 1-3, 2018, doi: 10.1109/ANTS.2018.8710139.

In this chapter, three types of compact DGS resonators are shown individually. The first one resembles the Sierpinski Carpet fractal geometry. On the other hand, the second one consists of two cascaded right triangular defects. Subsequently, a triangular defect in the ground plane in conjunction with split ring defect type in the microstrip line is shown. The design and development of each DGS resonator is shown below:

2.1. Sierpinski Carpet fractal DGS resonator

Here, the concept of Sierpinski Carpet in conjunction with DGS has been utilized and then modified to achieve significant size reduction in the proposed band stop filter. The design process essentially involves two stage iteration i.e. iteration 1 and iteration 2 as shown in Fig. 2.1. In the first iteration, four square shaped lattices are etched in the ground plane to achieve band stop property, for example around 300 MHz in this case. In the second iteration, more

ground etching is done to enhance the electrical length and subsequently the adjustment of the desired frequency. In the second iteration, more ground etching is done to enhance the electrical length and subsequently the adjustment of the desired frequency.

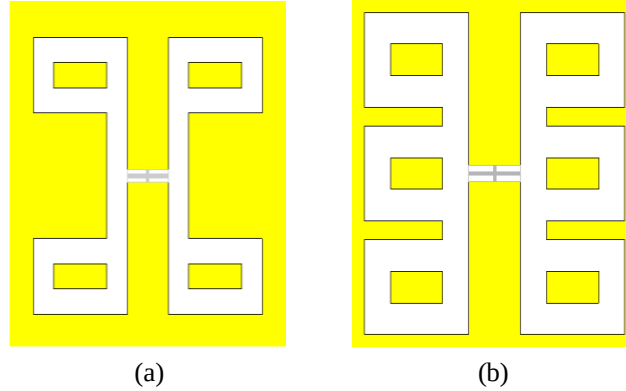


Fig. 2.1.1. Layout of DGS band stop filter. (a) First Iteration. (b) Second Iteration.

Now, it is important to mention that the fractals have two main properties i.e., the self-similarity and the space filling [45]. The space filling property makes the overall circuit size more compact due to the fact that the electrical length of the filter circuit increases and that leads to the increase in capacitance. This in turn helps in the reduction of the resonant frequency. As the number of iterations increases, the space filling property of the structure causes the pole location to shift to lower frequencies [46]. The addition of fractal in the structure while moving from first to second iteration leads to addition of inductance and the electrical length. In addition, fractals increase the net equivalent capacitance in the circuit. The overall effect is thus the decrease in resonant frequency due to the increase in inductance, capacitance, and electrical length. As a result of which, the filter type using fractal DGS geometry results in greater degree of compactness [47] [48] when compared to a simple DGS structure.

2.1.1. Equivalent Circuit

The fractal based DGS is modeled by an equivalent parallel resonant LC circuit depicted in Fig. 2.2. The value of capacitance (C_p) is in Pico farads (pF) whereas the value of the inductance (L_p) is in nano henrys (nH). The capacitance and the inductance are mathematically expressed by (2.1) and (2.2).

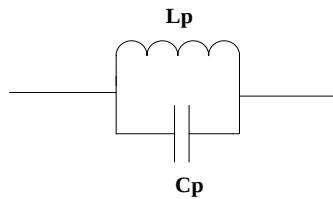


Fig. 2.2 Equivalent LC Model

$$C_p = \frac{5f_c}{\pi(f_0^2 - f_c^2)} \quad (2.1)$$

$$L_p = \frac{250}{C_p(\pi f_0)^2} \quad (2.2)$$

Where f_c is the 3dB cutoff frequency and f_0 is the pole frequency [49]. It is important to reiterate that the increase in the width of the defect in the substrate leads to minor increase in the capacitance [50] and this essentially helps in slightly reducing the resonant frequency. Similarly, the reduction in the width of the defect in the substrate will lead to increase in the resonant frequency. In summary, the resonant frequency can be adjusted slightly by varying the width of the defect in the substrate.

2.1.2. Design

The fractal DGS cell comprises of two stage iteration as shown in Fig. 2.3. The S-parameter analysis of the layout and the equivalent circuit of each iteration is performed. Furthermore, the flow chart described in the Fig. 2.3 presents the design procedure for Fractal type DGS. This design procedure follows consecutive iteration to obtain the design frequency at 300 MHz.

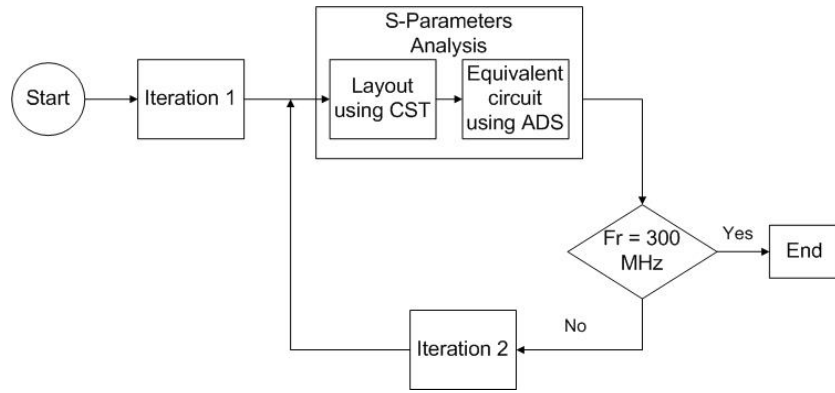


Fig. 2.3 Fractal DGS band stop filter design flowchart.

The fractal DGS cell, depicted in Fig. 2.4, is designed on FR-4 substrate with $\epsilon_r=4.7$, substrate thickness = 1.524 mm, and metal thickness = 35 μ m. The top layer of the substrate is a 50 Ω microstrip feed line of length L_F and width W_F as shown in Fig. 2.4(a). Here, the width of the feed line is altered to achieve impedance matching at 50 Ω . The increase in the width of the feed line leads to decrease in the impedance whereas the reduction in the width of the feed line results in the increase of the impedance. The DGS resonator represents a two port network i.e. Ports P1 and P2, which are SMA connector extending from bottom layer, i.e. ground plane, to the top layer. The width of the ports are of the size of feed line. The ground plane of fractal DGS

resonator as shown in Fig. 2.4(b) is described by four square shaped defect of width W_D , length L_D , length of fractal L , width of fractal W , Length of the slot L_S , distance of the slot g , width of inner defect W_1 , and the length of inner defect L_1 whose values are defined in Table 2.1. The defects consisting of each square shaped lattice gives rise to inductance which essentially leads to the increase in the series inductance.

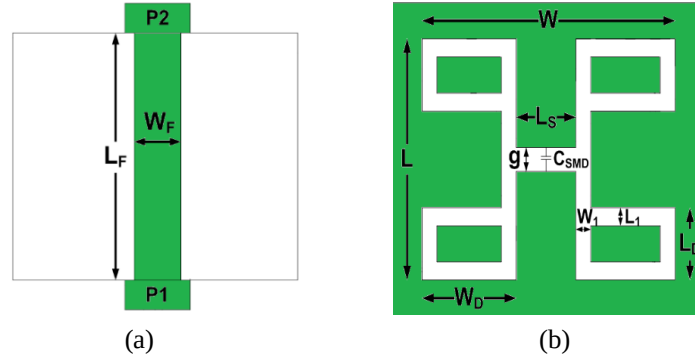


Fig. 2.4 Geometry of First Iteration DGS resonator. (a) Top Plane. (b) Bottom Plane.

The equivalent series inductance of individual patch is added to the microstrip line which is responsible for fixed cut-off frequency. As a result, by adding more number of self-similar square loops in the structure leads to decrease in the cut-off and the resonant frequencies. Furthermore, a rectangular shaped slot is etched below the microstrip feed line in the ground plane which acts as a parallel plate capacitor giving rise to capacitive effect. The capacitance decreases with the increase in the distance ' g ' of the etched slot and this causes the location of the attenuation pole to shift to higher resonance frequency. The attenuation pole results due to presence of the etched rectangular shaped gap in the ground plane below the microstrip line. Here, the attenuation pole location equals the resonance frequency of equivalent parallel LC circuit shown in Fig. 2.2 [51].

Table 2.1: Design Parameters of the proposed Fractal type defected ground structure band stop filter at 300 MHz operating frequency.

Design Parameters	L	W	W_D	L_D	W_1	L_1	g	L_S	L_F	W_F	C_{SMD} (pF)
Dimensions (mm)	11	11.2	4.6	3	1	1	0.5	2	14	3	33

Now, in order to obtain more miniaturization, the rectangular slot in the ground plane is enclosed by a surface mounted (SMD) ceramic capacitor C_{SMD} as shown in Fig. 2.4(b). By using this, the physical length of the circuit can be minimized leading to a more compact structure [23]. Fig. 2.5 depicts second iteration which consists of six square loops in the ground plane of fractal DGS resonator. The dimensions of the fractals are same in this iteration as well to maintain the self-similarity. Here, it is relevant to note that the desired frequency was obtained after 2 iterations and hence further iteration was not required.

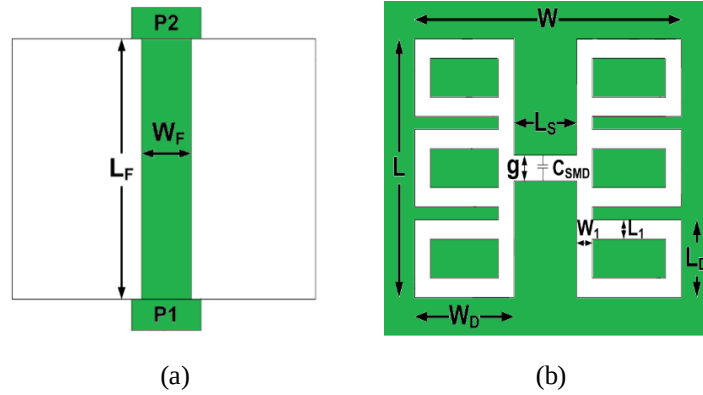


Fig. 2.5 Geometry of Second Iteration DGS resonator. (a) Top Plane. (b) Bottom Plane.

The simulation results after first and second iteration is shown in Fig. 2.6. The first iteration causes frequency variation from the design frequency, i.e. 300 MHz, and this difference occurs as a result of the decrease in the equivalent series inductance and capacitance. Therefore, to mitigate the effects of frequency variation, the second iteration is used and this apparently shifts the frequency backwards to the desired value of 300 MHz.

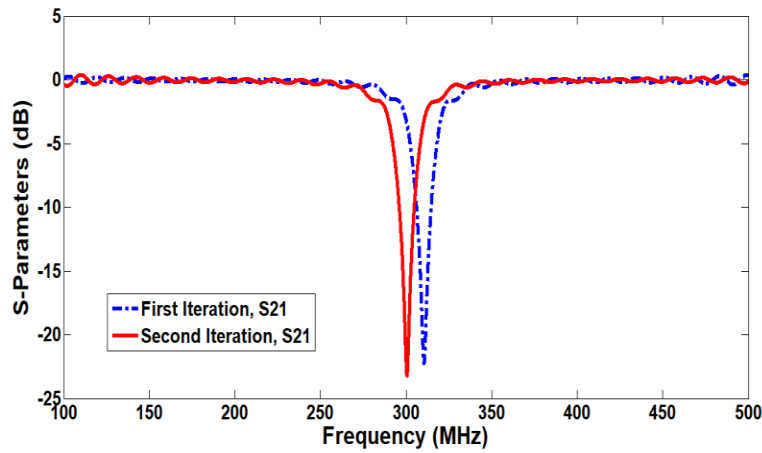


Fig. 2.6 Simulated Performance after the two iterations

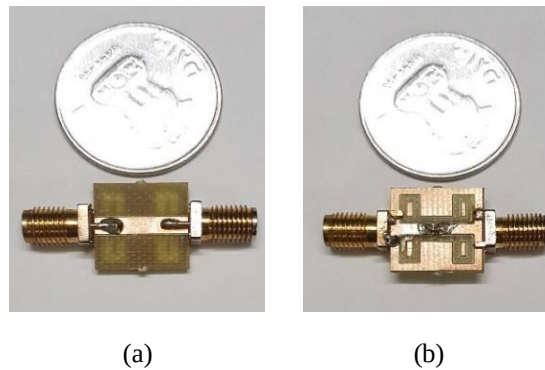


Fig. 2.7 Prototype of Iteration-1 Fractal DGS band stop filter. (a) Top. (b) Bottom.

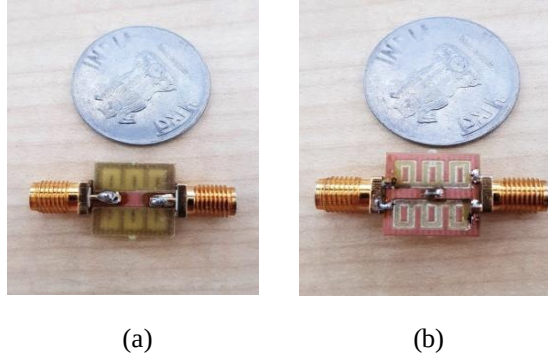


Fig. 2.8 Prototype of Iteration-2 DGS band stop filter. (a) Top. (b) Bottom.

Two prototypes for 300MHz, one each for iteration-1 and iteration-2, in Figs. 2.7 and 2.8 respectively have been developed on FR4 substrate and measured using Keysight VNA over the frequency range of 100-500MHz in order to show the performance enhancement achieved in the measurement results.

The EM simulated and measured results after first iteration are compared in Fig. 2.9. It is apparent that there are two clear differences in both the S11 and S21 values.

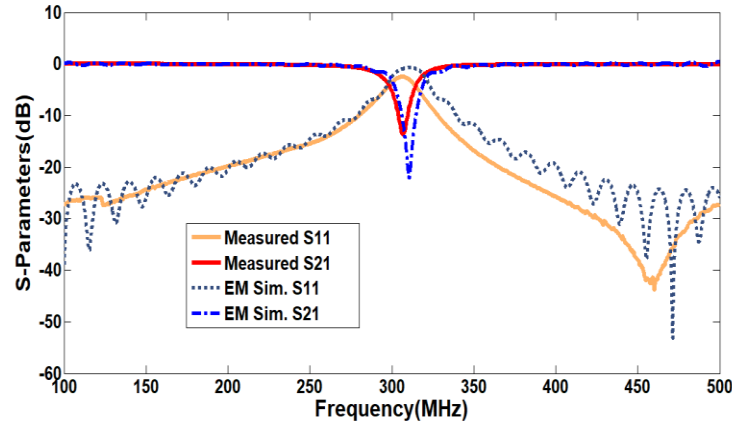


Fig. 2.9. Simulated and Measured S-Parameter of the DGS band stop filter.

Firstly, for example, the S21 in both the simulated and measured cases are shifted from the desired frequency of 300MHz. Secondly, it's measured of value -14dB at 307MHz is much lower than the EM simulated value of -20dB at 310MHZ. In order to address both these anomalies, another iteration is carried out as mentioned in the design section. In the second iteration, the required capacitance is of SMD type from Kemet with package size 0402 (1mm x 0.5mm) having value of 33pF. Owing to the fact that capacitor present in the ground plane leads to impedance mismatch at different frequencies due to the value of SMD capacitor chosen, resonates at design frequency. As a result, ripples will appear for those frequencies where impedance mismatch takes place. Furthermore, an attempt was made to design the fractal based DGS with dimension of 10 x

10 mm², however due to very narrow spacing between the pins of SMA connector in the ground plane resulted in impedance mismatch which deteriorated the overall response of the filter.

It is apparent from the measured results in Fig. 2.10 that the second iteration overcomes the effect of frequency variation and also leads to a good impedance bandwidth in comparison to the first iteration. For the sake of completeness, this plot also includes the equivalent circuit simulation results in addition to the EM simulated and measured results.

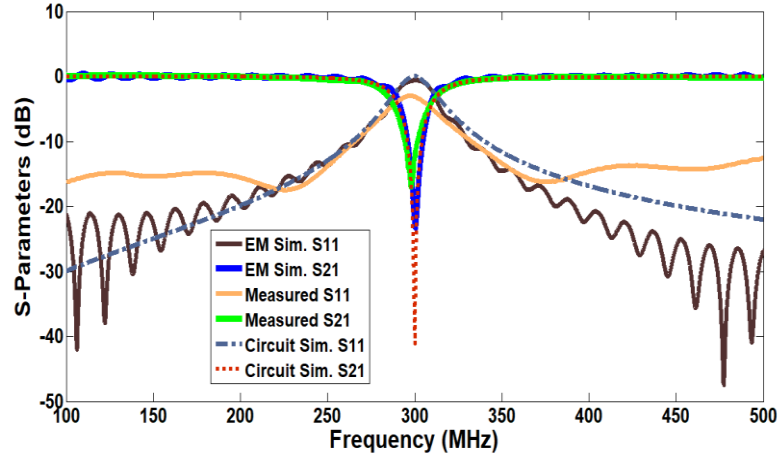


Fig. 2.10 Simulated and Measured S-Parameter of the DGS band stop filter.

Clearly, the measured value of S21 around -20dB fulfills the needs of most of the practical situations encountered in applications such as energy harvesting applications. After second iteration, the band-stop frequency is around 298MHz and improvements can be obtained by using better substrate. Other techniques to improve this shift include tuning of the capacitor or the width of the defects. Alternatively, for more precise applications, another iteration can be incorporated. Finally, the design proposed in this paper is compared with other reported designs in Table 2.2. The proposed design in this paper exhibits the most compact size. For example, the proposed filter operating at 300MHz and designed on a low cost FR4 substrate takes an area of 14 x 14 mm² and this improves the size reported in a recent paper [23] by 33%.

Table 2.2: Comparison with Other DGS Filter Geometries.

DGS Geometries	Frequency (GHz)	Size of Filter (mm ²)
[47], Minkowski fractal	6.2	30x30
[23], H and Semi-H shaped DGS	0.3	21x21
[50], U and V slot DGS	3.5-4.3	L=22, W= not defined
[52], Dumbbell-shaped DGS	5.3-10	11.4x4.5 (slot length only, filter length not defined)
This work, Sierpinski Carpet Fractal DGS	0.3	14x14

2.2. Dual-Band Right Triangle DGS Resonator

The right triangle based DGS cell, depicted in Fig. 2.11, is realized on Rogers (RO4350B) with $\epsilon_r = 3.66$, substrate thickness = 1.524 mm, and metal thickness = 35 μm . The top layer of the substrate is a 50 Ω microstrip feed line of length L_F and width W_F as shown in Fig 2.11(a). The ground plane of the DGS resonator depicts a right triangle defect of width W_T , height H_T , width of slot W_S , length of slot L_S , overall width W and length L , as shown in Fig 2.11(b). Table 2.3 presents the design parameters of the right triangle DGS resonator operating at 0.93 GHz. The dimension of each parameter is selected to fulfill the requirements of physical size restriction and targeted operating frequency. To satisfy both these requirements, the electrical length is maximized by increasing the length of the defect in the ground plane. By increasing the length, the resonance frequency shifts downwards towards lower frequency range. Further shift in the frequency towards the design frequency can be achieved by adding a loading capacitor C_{SMD1} in the slot in the ground plane as shown in Fig. 2.11(b). In addition, the design parameters were optimized using CST Microwave Studio.

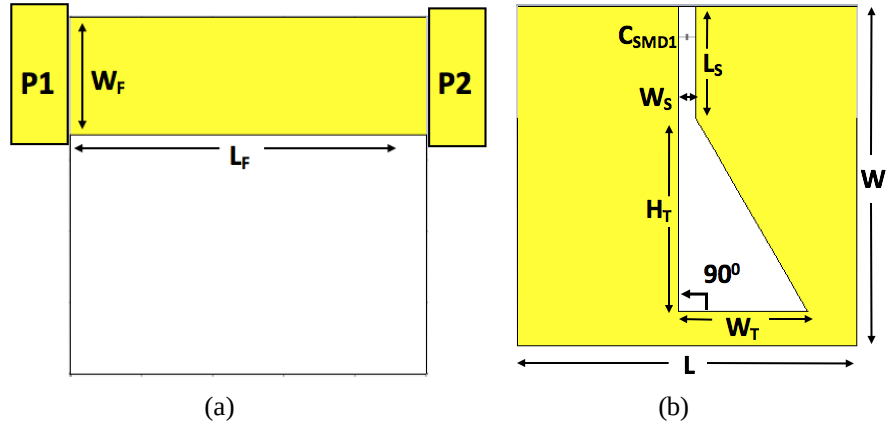


Fig. 2.11. Geometry of right triangle DGS resonator: (a) Top Plane, (b) Bottom Plane.

Table 2.3: Design Parameters of the proposed right triangle DGS Resonator.

Design Parameters	L	W	H_T	W_T	W_S	L_S	L_F	W_F	C_{SMD1} (pF)
Dimensions (mm)	10	10	5.71	3.79	0.5	3.29	10	3.28	4.4

The right triangle based DGS can be modeled by an equivalent parallel resonant LC filter depicted in Fig. 2.2. The value of inductance (L_P) is in nano henrys (nH) whereas the value of the capacitance (C_P) is in Pico farads (pF). The inductance and the capacitance can be mathematically expressed by (2.3) and (2.4) [24]. Here, the impedance matrix $imag(Z_{11})$ is determined at the

resonance frequency of interest in CST Microwave Studio. Subsequently, the roll-off factor of the filter is expressed by (2.5).

$$L_p = \frac{\text{imag}(Z_{11})}{2\pi f_{01}} \quad (2.3)$$

$$C_p = \frac{1}{4\pi^2 f_{01}^2 L_p} \quad (2.4)$$

$$\text{Roll-off factor} = \frac{f_{c1}}{f_{01}} \quad (2.5)$$

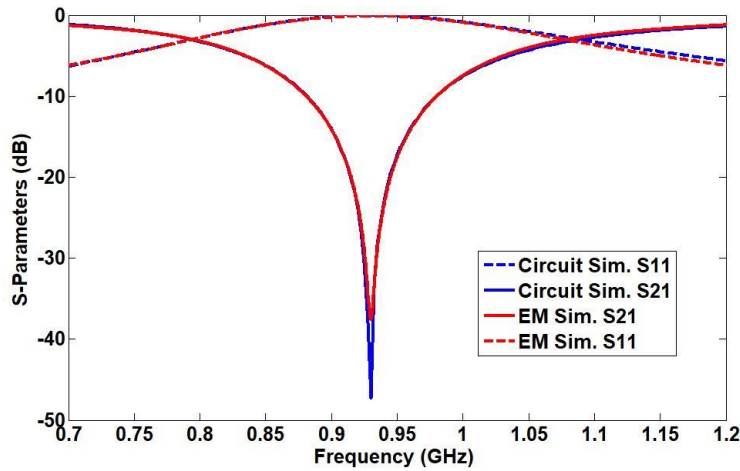


Fig. 2.12. S-Parameter plot of single band DGS resonator.

The Roll-off factor (ROF) determines the sharpness of the filter response. As the roll-off factor reaches close to the maximum value of 1, the sharpness of the filter increases [47]. It can be seen in Fig. 2.12 that the EM simulation of the right triangle DGS resonator and the S-parameter simulation results from the equivalent circuit are in good agreement and shows a very good DGS band-stop characteristics. From this plot, the -3dB cutoff frequency f_{c1} and the pole frequency f_{01} can be deduced. In present scenario, the value of $f_{c1} = 0.794$ GHz and $f_{01} = 0.93$ GHz. Furthermore, the design can also be modified, as expressed in the subsequent section, to show resonance at two frequencies.

2.2.1. Design of Dual-Band Cascaded Resonator

In this section, the right triangle dual band DGS resonator is represented by equivalent circuit shown in Fig. 2.13. Since the structure is designed to resonate at two frequencies i.e. 0.93 GHz and 2.44 GHz, as a result, at $f_{01} = 0.93$ GHz, C_{p1} shorts L_{p1} . While at $f_{02} = 2.44$ GHz, C_{p2}

shorts L_{P2} . Here, R_0 and R_1 represents 50Ω SMA feeding ports. The feeding ports are the load terminations in the equivalent circuit.

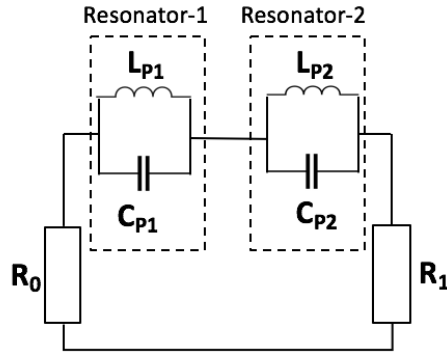


Fig. 2.13. Equivalent Circuit of dual band resonator.

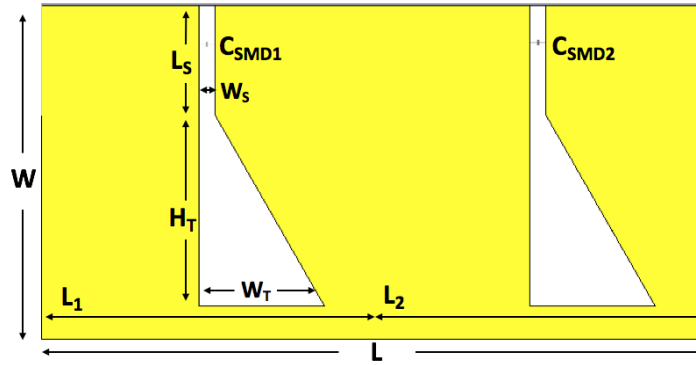


Fig. 2.14 Dual-Band defected ground structure (DGS) resonator.

Furthermore, the dual band DGS resonator is formed by cascading two single band right triangle DGS band-stop filter (BSF) as shown in the Fig. 2.14. The dimensions are the same as the single-band DGS-BSF except the overall length L ($L=L_1+L_2$) of the dual band DGS resonator increases from 10 mm to 20 mm. Here, two different loading capacitors $C_{SMD1} = 4.4$ pF and $C_{SMD2} = 0.4$ pF are added to the excitation gap of two symmetric right triangle DGS resonator. The values of the loading capacitors are chosen such that the two resonators resonate at two different frequencies i.e. 0.93 GHz and 2.44 GHz.

A prototype of the proposed dual-band right triangle DGS band-stop filter, shown in Fig. 2.15(a), has been developed on Rogers 4350B and measured using Keysight VNA over the frequency range of 0.5-3.5 GHz in order to show the performance enhancement achieved in the measurement results. The measurement setup, in Fig. 2.15(b), is connected to the two ports of the right triangle DGS resonator structure to obtain the S-Parameter response.

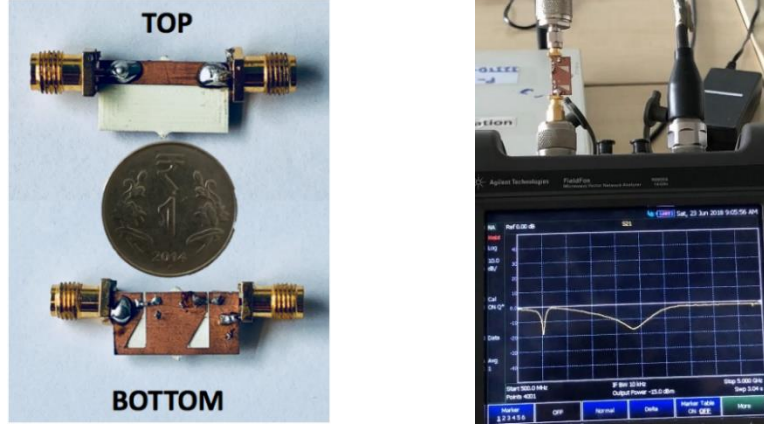


Fig. 2.15. Prototype of the dual band right triangle DGS-BSF. (a) Top and bottom side. (b) Measurement setup.

The EM simulation and measured results are depicted in Fig. 2.16. The EM simulated results show resonance at two frequencies i.e. $f_{01} = 0.93$ GHz with $|S_{21}| = -34$ dB and $|S_{11}| = -0.24$ dB and $f_{02} = 2.44$ GHz with $|S_{21}| = -41$ dB and $|S_{11}| = -0.1$ dB. Whereas, in the measurement result, the dual band DGS resonator resonates at $f_{01} = 0.931$ GHz with $|S_{21}| = -20$ dB while $|S_{11}| = -1.56$ dB. Whereas at $f_{02} = 2.5$ GHz with $|S_{21}| = -19.2$ dB while $|S_{11}| = -2.39$ dB. Here, the resonant frequency f_{02} shows a wideband response in comparison to the narrow band response shown by resonance frequency f_{01} . This occurs as a result of lower value of capacitance of resonator 2 which increases the bandwidth at f_{02} whereas, the value of capacitance at resonator 1 is higher which leads to narrower bandwidth at f_{01} .

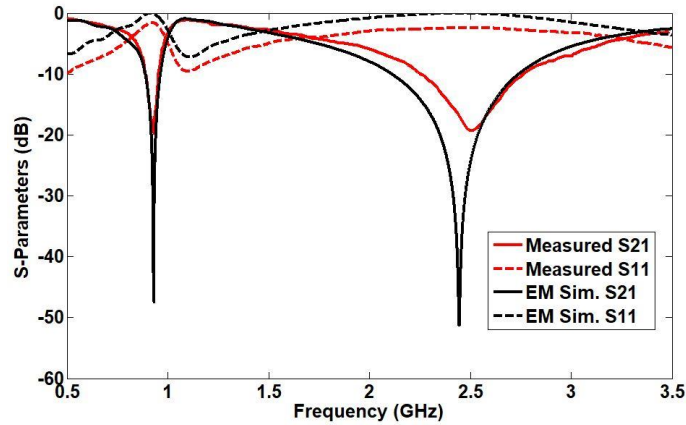


Fig. 2.16. Simulated and Measured S-Parameter of the dual band DGS resonator.

It is important to note that higher mutual coupling between two resonators results in return loss of the pass band frequencies, greater than -10 dB. The higher mutual coupling occurs as a result of closer value of capacitance C_{SMD1} and C_{SMD2} at two frequencies respectively, which further leads to increase in bandwidth. In order to ease out the complexity of high mutual coupling, the spacing between the two resonators can be increased which leads to lower mutual coupling and bandwidth. In addition, it can be seen that there is a slight variation in the two frequencies of the

measured result in comparison to the EM simulation result. This occurs as a result of tolerances in the loading capacitor or the fabrication complexity due to compact size and offset from the given dimensions during the etching process of the defect in the ground plane. To overcome the shift, capacitor values of low tolerance and high precision fabrication instruments can be utilized. The Table 2.4 compares different types of dual band stop filter designs which shows that right triangle dual band stop filter has the most compact size of 200 mm².

Table 2.4: Comparison with other Dual Band Stop Filter Designs

Dual Band Stop Filter Design	f_{01} (GHz)	f_{02} (GHz)	Size of the filter (mm ²)
[24], Circular shape DGS	0.3	0.7	30x15
[53], Microstrip-DGS resonator	1.7	2.3	34x29.4
[54], Split Ring Resonator (SRR) and Complementary SRR	2.55	5.05	14x20
[55], Open ended Stepped Impedance Resonator (OSIR)	0.9	2.1	28.9x29.3
This work, Right Triangle DGS	0.93	2.44	20x10

2.3. Triangular shaped DGS and Split ring type DMS Dual Band Resonator Design

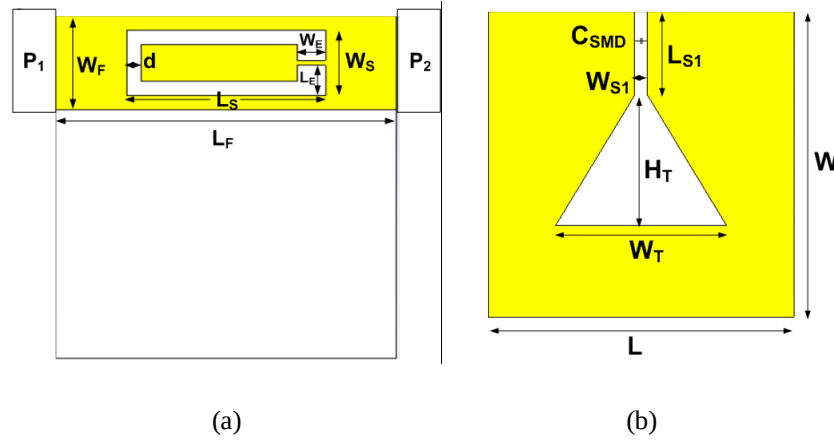


Fig. 2.17. Geometry of DB-BSF resonator: (a) Top Plane. (b) Bottom Plane.

The split ring type DMS cell, depicted in Fig. 2.17, is realized on Rogers (RO4350B) with $\epsilon_r = 3.66$, substrate thickness = 1.524 mm, and metal thickness = 35 μm . The top layer of the substrate is a 50 Ω microstrip feed line of length L_F and width W_F as shown in Fig 2.17(a). The microstrip feed line is etched with a patterned defect of width W_S , length L_S , slot width d , length of edge L_E , and width of edge W_E . The patterned defect depicts a split ring with two protruded rectangular edges to obtain resonance at desired frequency i.e. 5.86 GHz. Here, the protruded

rectangular edges give rise to additional inductance in comparison to the inductance obtained by split ring defect. As a result, there is an increased shift in the resonance frequency towards lower frequency range. Furthermore, the triangular shaped defected ground structure is etched on the bottom plane of the defected microstrip feed line, as shown in Fig. 2.17(b). The ground plane depicts triangular defect of height H_T , width W_T , slot width W_{S1} , and slot length L_{S1} . Here, the triangular shaped defect gives rise to effective inductance whereas excitation slot gives rise to effective capacitance. Since, the inductance and capacitance obtained by triangular defect and excitation slot is not enough to show resonance at the desired frequency i.e. 2.4 GHz. As a result, further shift in the frequency towards the design frequency can be achieved by adding a loading capacitor C_{SMD} in the slot in the ground plane. The Table 2.5 presents the design parameters of the triangular shaped defected ground structure and the split ring type defected microstrip line operating at 2.4 GHz and 5.86 GHz. The dimension of each parameter is selected to fulfill the requirements of physical size restriction and targeted operating frequency. To satisfy both these requirements, the electrical length is maximized by increasing the length of the defect in the top plane and ground plane. By increasing the length, the resonance frequency shifts downwards towards lower frequency range. In addition, the design parameters were optimized using CST Microwave Studio.

Table 2.5: Design Parameters of the proposed Dual-band DGS-DMS Resonator

Design Parameters	Dimensions (mm)	Design Parameters	Dimensions (mm)
L	12	L_E	1.05
W	12	d	0.5
W_F	3.28	H_T	5.12
L_F	12	W_T	6.71
W_S	2.28	W_{S1}	0.5
L_S	7	L_{S1}	3.28
W_E	1	C_{SMD} (pF)	0.4

Now, the split ring type defected microstrip line and triangular shaped defected ground structure is represented by parallel LC equivalent circuit, as shown in Fig. 2.13. The value of inductance L_{P1} and L_{P2} is in nano henrys (nH) whereas the value of the capacitance C_{P1} and C_{P2} is in pico farads (pF). The capacitance and the inductance can be mathematically expressed by (2.6) and (2.7) [56].

$$C_{P1} = \frac{\omega_{c1}}{2Z_0(\omega_{01}^2 - \omega_{c1}^2)}$$

$$C_{P2} = \frac{\omega_{c2}}{2Z_0(\omega_{02}^2 - \omega_{c2}^2)} \quad (2.6)$$

$$L_{P1} = \frac{1}{4\pi^2 f_{01}^2 C_{P1}}$$

$$L_{P2} = \frac{1}{4\pi^2 f_{02}^2 C_{P2}} \quad (2.7)$$

Here, ω_{c1} represents -3dB cut-off frequency of resonator-1, ω_{01} represents resonance frequency of resonator-1, ω_{c2} represents -3dB cut-off frequency of resonator-2 and ω_{02} represents resonance frequency of resonator-2. Since the structure is designed to resonate at two frequencies i.e. 2.4 GHz and 5.86 GHz, as a result, at $f_{01} = 2.4$ GHz, the bottom plane DGS becomes active and C_{P1} shorts L_{P1} . While at $f_{02} = 5.86$ GHz, the top plane DMS switches to active state and C_{P2} shorts L_{P2} . Here, R_0 and R_1 represents 50Ω SMA feeding ports. The feeding ports are the load terminations in the equivalent circuit.

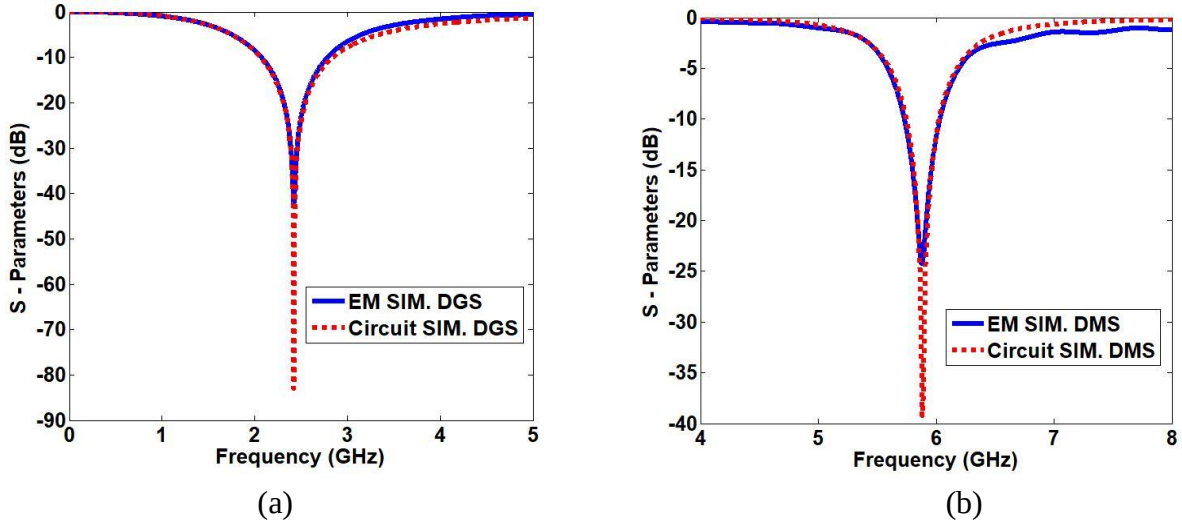


Fig. 2.18. EM simulation and circuit simulated transmission response. (a) Defected Ground Structure. (b) Defected Microstrip Line.

Subsequently, individual performance of each unit cell defect i.e. defected microstrip line and defected ground structure is analyzed. First, a split ring based defected microstrip line without etching in the ground plane is realized. Subsequently, triangular shaped defect in the ground plane with 50Ω feed line on the top plane is analyzed. The Fig 2.18(a) and Fig 2.18(b) depicts the EM simulation and circuit simulation results for triangular DGS and split ring type DMS. For triangular DGS, the EM simulation and circuit simulation results show resonance at lower frequency i.e. 2.4 GHz whereas for split ring type DMS, both the simulation results show resonance at higher frequency i.e. 5.86 GHz.

Finally, a prototype of the proposed dual-band band stop filter using triangular shaped DGS and split ring type DMS, shown in Fig. 2.19(a), has been developed on Rogers 4350B and

measured using Keysight VNA over the frequency range of 50KHz - 8GHz in order to show the performance enhancement achieved in the measurement results. The device under test (DUT), in Fig. 2.19(b), is connected by two ports P1 and P2 to obtain the S-Parameter response.

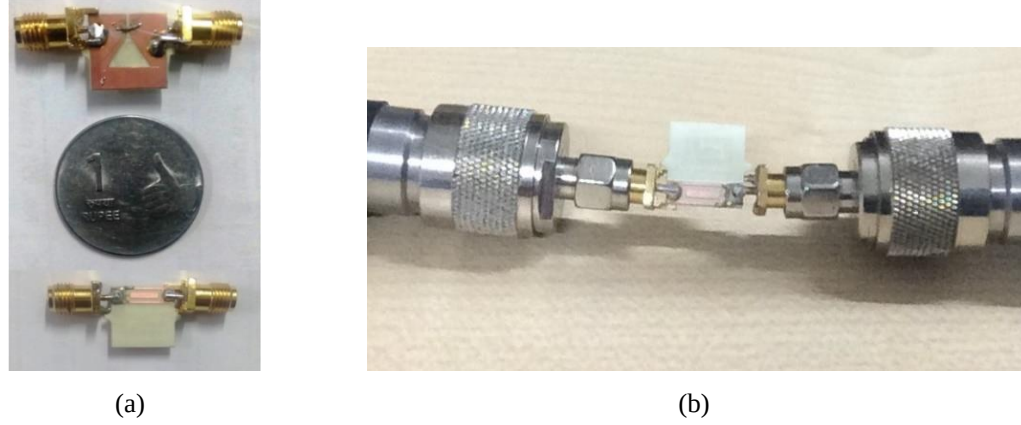


Fig. 2.19. Prototype of the dual band DGS-DMS resonator. (a) Bottom and top side. (b) Device under test (DUT).

The EM simulation and measured results are depicted in Fig. 2.20. The EM simulated results show resonance at two frequencies i.e. $f_{01} = 2.4$ GHz with $|S_{21}| = -46$ dB and $|S_{11}| = -0.07$ dB and $f_{02} = 5.86$ GHz with $|S_{21}| = -30$ dB and $|S_{11}| = -0.62$ dB. Whereas, in the measurement result, the dual band DGS resonator resonates at $f_{01} = 2.5$ GHz with $|S_{21}| = -16$ dB while $|S_{11}| = -1.89$ dB. Whereas at $f_{02} = 5.85$ GHz with $|S_{21}| = -49.5$ dB while $|S_{11}| = -1$ dB.

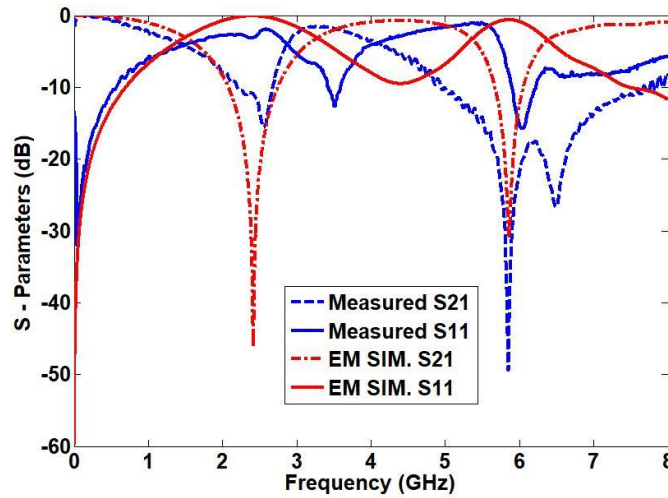


Fig. 2.20. Simulated and Measured S-Parameter of the dual band DGS resonator.

In addition, it can be seen that there is a slight variation in the two frequencies of the measured result in comparison to the EM simulation result. This occurs as a result of tolerances in the loading capacitor or the fabrication complexity due to compact size and offset from the given dimensions during the etching process of the defect in the ground plane. To overcome the shift, capacitor values of low tolerance and high precision fabrication instruments can be utilized. The

Table 2.6 compares different types of dual band stop filter designs which shows that DMS-DGS based dual band stop filter has the most compact size of 144 mm².

Table 2.6: Comparison with other Dual Band Stop Filter Designs

Dual Band Stop Filter Design	f_{o1} (GHz)	f_{o2} (GHz)	Size of the filter (mm²)
[24], Circular shape DGS	0.3	0.7	30x15
[53], Microstrip-DGS resonator	1.7	2.3	34x29.4
[54], Split Ring Resonator (SRR) and Complementary SRR	2.55	5.05	14x20
[55], Open ended Stepped Impedance Resonator (OSIR)	0.9	2.1	28.9x29.3
This work, DGS-DMS	2.4	5.86	12x12

2.4. Conclusion

In this chapter, we proposed three different types of compact DGS resonators namely the Sierpinski carpet fractal resonator, two cascaded right triangular resonator design, and a resonator based on defected ground structure (DGS) and a defected microstrip line (DMS). The Sierpinski fractal based resonator design operating at 0.3 GHz takes an area of 14 x 14 mm². On the other hand, two cascaded right triangular defects operating at dual band frequencies 0.93 GHz and 2.44 GHz take an area of 20 x 10 mm². Finally, the DGS-DMS based resonator design operating at 2.4 GHz and 5.86 GHz frequencies takes an area of 12 x 12 mm². The DGS-DMS based dual band resonator technique achieves a more compact size in comparison to two cascaded right triangle dual band DGS resonator design. Furthermore, these three different types of resonator techniques achieve the most compact circuit size in comparison to the state-of-the-art resonator designs. In addition, an excellent agreement between the measurement and EM simulated results is a testament to the robustness of the proposed design technique.

Chapter 3

Compact Highly Efficient Resonant Inductive WPT Systems

This chapter is based on the following research papers:

- S. Verma, D. Rano, M. Hashmi and V. Bohara, “A High Q Dual E-Shaped Defected Ground Structure for Wireless Power Transfer Applications,” *Asia-Pacific Microwave Conference (APMC)*, pp. 1435-1437, 2018, doi: 10.23919/APMC.2018.8617254.
- S. Verma, D. Rano and M. Hashmi, “A Novel Dual Band Defected Ground Structure for Short Range Wireless Power Transfer Applications,” *IEEE Wireless Power Transfer Conference*, pp. 188-191, 2019, doi: 10.1109/WPTC45513.2019.9055660.

In this chapter, two types of compact defected ground structure (DGS) wireless power transfer (WPT) systems are shown individually. The first one is based on dual E shaped DGS resonator operating on single frequency band. On the other hand, the second one consists of U-shaped DGS resonator operating on two different frequencies. In both the scenarios, the resonators are placed back-to-back separated by a critical coupling distance to transfer power wirelessly. The design and development of each DGS-WPT system is shown below:

3.1. Dual E-shaped DGS-WPT System Design

The dual E-shaped DGS cell, depicted in Fig. 3.1, is realized on Rogers (RO4350B) with $\epsilon_r=3.48$, substrate thickness = 1.524 mm, and metal thickness = 35 μ m. The top layer of the substrate is a 50 Ω microstrip feed line of length L_F and width W_F as shown in Fig 3.1(a). The ground plane of the DGS resonator depicts six rectangular shaped defects with the corresponding dimensions in Fig. 3.1(b). Table 3.1 presents the design parameters of the dual E-shape DGS resonator operating at 270 MHz. The dimension of each parameter is selected to fulfill the requirements of physical size restriction and targeted operating frequency. To satisfy both these requirements, the electrical length is maximized by increasing the length of the defect in the

ground plane. By increasing the length, the resonance frequency shifts downwards towards lower frequency range. Further shift in the frequency towards the design frequency can be achieved by adding an external capacitor (C_{SMD}) in the slot in the ground plane as shown in Fig. 3.1(b). In addition, the design parameters were optimized using CST Microwave Studio.

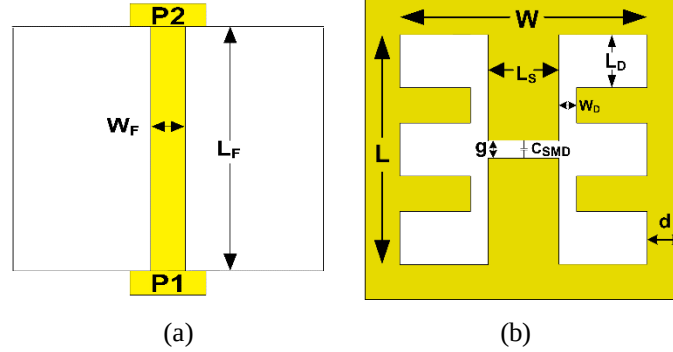


Fig. 3.1. Geometry of dual E- DGS resonator. (a) Top Plane. (b) Bottom Plane.

Table 3.1: Design Parameters of the proposed dual E-shape DGS Resonator.

Design Parameters	L	W	W_D	L_D	d	g	L_S	L_F	W_F	C_{SMD} (pF)
Dimensions (mm)	18	18	2.4	5.2	1	0.5	4	20	3.40	18

The dual E-shape DGS can be modeled by an equivalent parallel resonant RLC circuit shown in Fig. 3.2. Here, $C_p = C_g + C_{SMD}$ is the parallel combination of gap capacitance C_g and external capacitor C_{SMD} . The inductance L_p is contributed by the loops of length H , width W and thickness d . The equivalent resistance R_{DGS} represents the lossy behavior of the gaps and the substrate. Then the design parameters in Table 3.1 is used to compute values of the resistance, capacitance, and inductance of the equivalent circuit [24]. Subsequently, S-parameter simulation is carried out to evaluate the developed equivalent circuit.

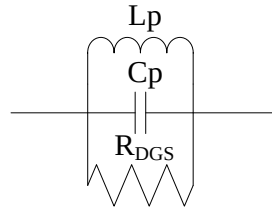


Fig. 3.2. Equivalent RLC Circuit

It can be seen in Fig. 3.3 that the EM simulation of the dual E-shaped resonator and the S-parameter simulation results from the equivalent circuit are in good agreement and shows a very good DGS-band stop characteristics. From this plot, the -3 dB cutoff frequency ω_c and the pole frequency ω_0 can be deduced. For example, the values f_c and f_0 in these plots do not meet the chosen frequency of 270 MHz. Instead f_0 is 287 MHz and f_c is 256 MHz and these are offset by a

significant value from the design frequency. In order to use the equivalent circuit in the analysis and design of WPT system, the DGS resonator needs further optimization. Then the optimized dimensions need to be used for extraction of parameters of equivalent circuit. The process requires repetition until the bandstop features meet the intended frequency. In this work, this process was repeated during the design of WPT system.

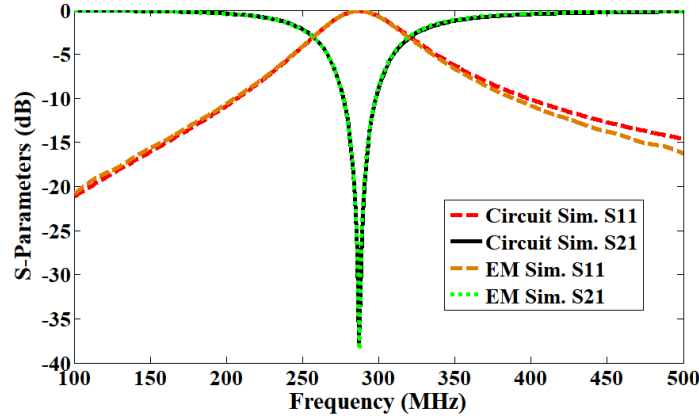
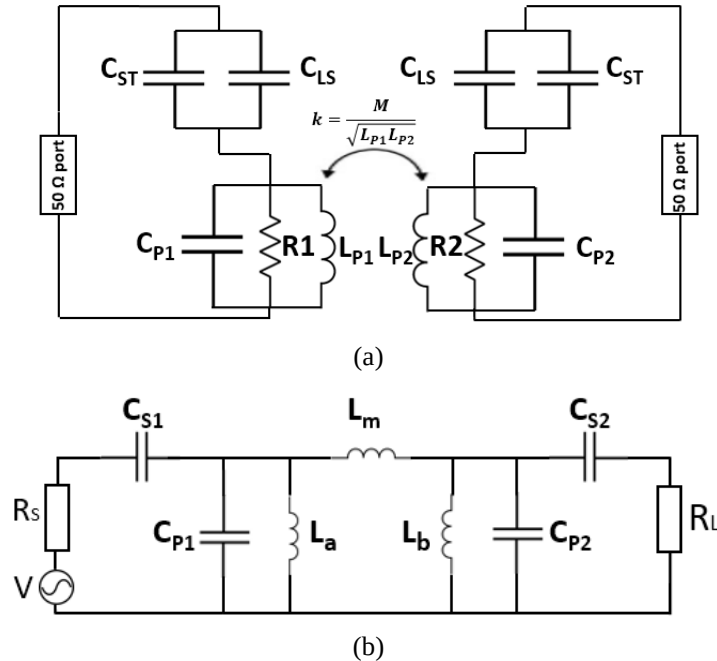


Fig. 3.3. S-parameter plot of Dual E-shaped DGS bandstop filter

Now, the proposed WPT system consists of two symmetrical DGS resonator coupled magnetically with each other. They are distance apart by ‘ h ’ and the power transfer takes place as a result of the intersection of magnetic fields. The proposed design includes two dual E-shaped DGS resonators with open stub in addition to 50Ω feed line at the top plane. The stub used here acts as a capacitor to provide impedance matching. The PCB layout of dual E-shaped coupled resonator is modelled by an equivalent circuit as shown in Fig. 3.4.



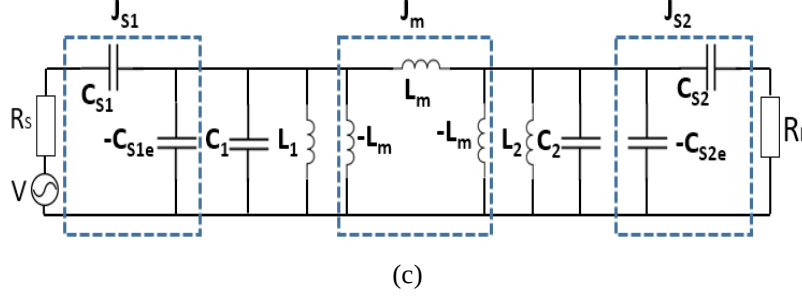


Fig. 3.4. (a) Equivalent circuit of coupled resonator WPT system. (b) The π equivalent circuit. (c) J -inverters based simplified equivalent circuit [24].

The equivalent circuit can be analyzed with J -(admittance) inverter as shown in Figs. 3.4(b)-(c) to determine the requirements for impedance matching [24]. Here, the source and load impedances, R_s and R_L , are taken as 50Ω so that the proposed wireless power transfer system can be measured with 50Ω standard measurement instrument. The transmitting and receiving DGS resonator are connected to source and load impedances via J - inverter. The mutual inductance, L_m , and the self-inductances, L_a and L_b , are expressed in (17), whereas the capacitor terms are expressed in (18). The expressions in (17) and (18) in conjunction with Fig. 3.4(c) can be used to find the input admittance of the inverter in (21). The condition in (21) is necessary condition for achieving maximum WPT efficiency [24]. Alternatively, it can be said that the impedance matching at the resonant frequency leads to the optimal efficiency of the WPT system. Now, the simulation of the back-to-back coupled dual-E shape based WPT system shows split, as can be seen in Fig. 3.6, in the resonance frequency in the over coupled region due to the proximity of the receiver and the transmitter. Now, in the case of magnetic coupling in the WPT system, the higher resonant frequency corresponds to the electric wall, f_e , and the lower one corresponds to the magnetic wall, f_m . This leads to the expression of coupling coefficient (3.1) [21]. It is one of the most important factor in the WPT system.

$$k = \frac{L_m}{L_p} = \frac{f_e^2 - f_m^2}{f_e^2 + f_m^2} \quad (3.1)$$

The other critical parameter in the design of a high performing WPT system is the quality factor, Q_U , of the uncoupled resonator given in (3.2).

$$Q_U = \frac{R_{DGS}}{X_L} = \frac{R_{DGS}}{w_0 L} \quad (3.2)$$

The parameters Q_U and the coupling coefficient k regulate the efficiency η_{opt} of the WPT system. These are related through the U-factor obtained from the coupled mode theory (CMT).

$$U = kQ_U \quad (3.3)$$

$$\eta_{opt} = \frac{U^2}{\left(1 + \sqrt{1 + U^2}\right)^2} \quad (3.4)$$

Apparently both high k and high Q_U is required for high efficiency of the WPT system. However, it is not straightforward to achieve both at the same time. For example, at larger distance the coupling factor k is small and that requires extremely high quality factor Q_U for maintaining high WPT efficiency. However, the requirement of high Q_U increases the design complexity of WPT system due to large number of turns i.e. the slots in the ground plane but also provides higher resistance (R_{DGS}) due to the narrow spacing between the slots. In order to ease the requirements of high Q_U , there is a need of tightly coupled resonators i.e. the resonators should be in close proximity. The drawback of tight coupling results into the phenomenon of frequency splitting and this leads to high efficiency at the two frequencies but with a significantly reduced power transfer. The level of power transfer can be increased by incorporating impedance matching. One aspect to be kept in consideration is to avoid the critical coupling condition as it eliminates the frequency splitting. In such a situation the input impedance of transmitter equals the internal resistance of WPT system (R_S) and it leads to a maximum power transfer but at the efficiency of only 50%.

A prototype, shown in Fig. 3.5(a), of the proposed WPT system has been developed on Rogers 4350B to evaluate its concept and the performance. The measurement is carried out using Keysight VNA over the frequency range of 100-500MHz. The measurement setup, in Fig. 3.5(b), has the two symmetric dual E-shape DGS resonators placed back-to-back on a customized platform. Here, one of the resonators acts as a transmitter while the other as receiver.

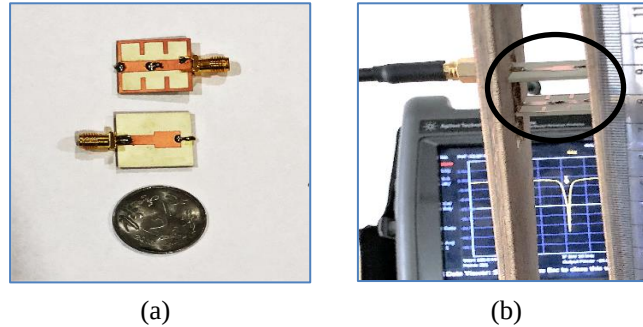


Fig. 3.5. Prototype of dual E-shape DGS resonator WPT system. (a) Bottom and top side. (b) Measurement setup.

The EM simulated and measured results are depicted in Fig. 3.6. The EM simulated result shows resonance at $f_0 = 270$ MHz with $|S_{21}| = -0.76$, while $|S_{11}| < -25$ dB. Whereas, in the measurement result, the WPT system resonates at $f_0 = 269$ MHz with $|S_{21}| = -1.5$ dB, while $|S_{11}| = -27$ dB.

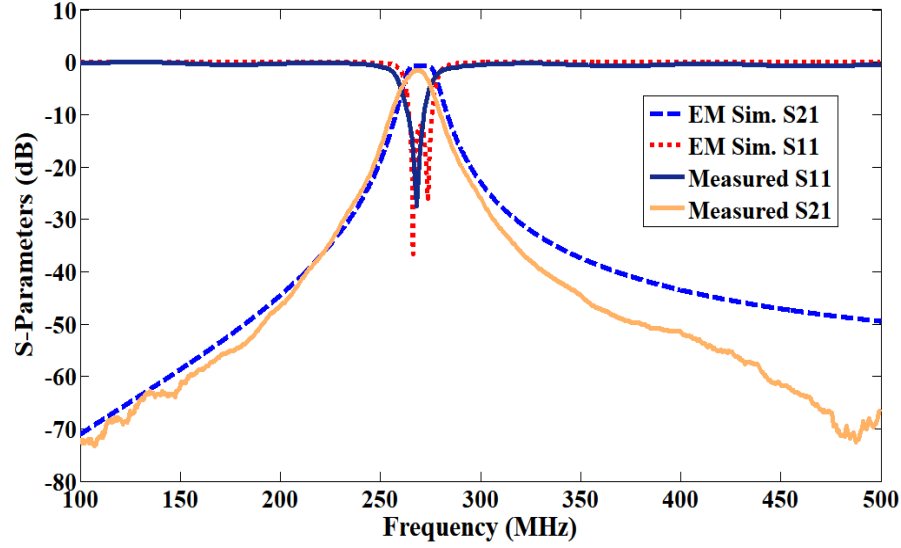


Fig. 3.6. Simulated and Measured S-Parameter of the WPT system.

The simulation and measured result have a slight shift in the resonance frequency from the desired frequency probably due to some tolerances in the loading capacitor or the in-house fabrication. To overcome the shift, capacitor values of low tolerance and frequency tuning can be incorporated. Furthermore, the measured efficiency of the proposed WPT system, depicted in Fig. 3.7, is calculated from the ratio of the power received at the output to the power transmitted at the input port given by $\eta = \frac{|S_{21}|^2}{1-|S_{11}|^2}$. Moreover, the proposed design is the most compact when compared to other designs in literature as can be inferred from Table 3.2. In addition, it achieves the simulated and measured efficiency of 85% and 71% respectively at a distance of 11mm.

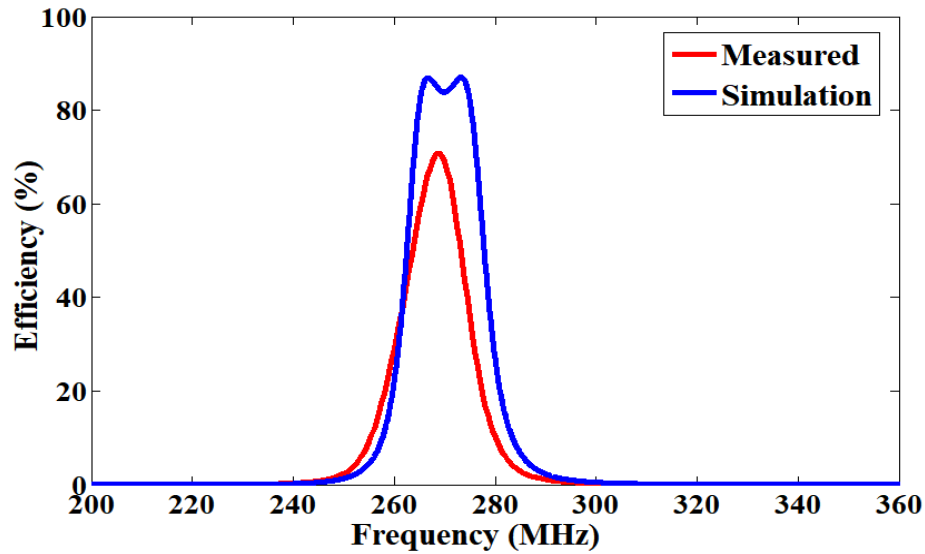


Fig. 3.7. Comparison between measured and simulation results.

Table 3.2: Comparison with other compact WPT Systems.

WPT System	Frequency (GHz)	Size (mm ²)	h (mm)	η (%)	Q-factor
[23]	0.3	20x20	13	68	-
[57]	0.28	20x10	10	65	-
[24]*	0.3 and 0.7	30x15	16	71 and 72	300 and 350
[58]	0.05	40x40	50	68.5	339
This work, Dual-E shape	0.27	20x20	11	71	470

*dual band

3.2. U-Shaped DGS-WPT System Design

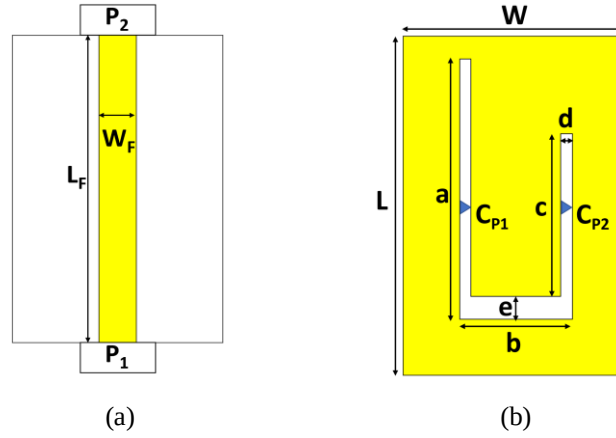


Fig. 3.8. Geometry of dual band DGS resonator: (a) Top Plane. (b) Bottom Plane.

The DGS cell, depicted in Fig. 3.8, is realized on Rogers (RO4003C) with $\epsilon_r = 3.55$, substrate thickness = 0.813 mm, and metal thickness = 35 μm . The top layer of the substrate is a 50 Ω microstrip feed line of length L_F and width W_F as shown in Fig 3.8(a). The dimensions of the ground plane of the DGS resonator depicts three rectangular loops of lengths ‘ a ’, ‘ b ’, and ‘ c ’ and width ‘ d ’, ‘ e ’ respectively as shown in Fig. 3.8(b). Since the two rectangular loops of lengths ‘ a ’ and ‘ c ’ behaves as a open loop resonator, as a result, both the loops resonate at two different frequencies. Resonances at two different frequencies depend on the individual loop lengths ‘ a ’ and ‘ c ’ shown later in this section. Now, the two resonances are tuned by loading the two loops with two different capacitors C_{P1} and C_{P2} to show resonance at the desired frequencies of interest. Here, Table 3.3 presents the design parameters of the dual band DGS resonator operating at 2.4 GHz and 5.8 GHz. The dimension of each parameter is selected to fulfill the requirements of physical size restriction and targeted operating frequency [22]. To satisfy both these requirements, the electrical length is maximized by increasing the length of the defect in

the ground plane. By increasing the length, the resonance frequency shifts downwards towards lower frequency range. In addition, the design parameters were optimized using CST Microwave Studio.

Table 3.3: Design Parameters of the proposed dual band DGS Resonator

Design Parameters	L	W	a	b	c	d	e	L_F	W_F	C_{P1} (pF)	C_{P2} (pF)
Dimensions (mm)	15	10	11.5	5	7.2	0.5	1	15	1.8	2.2	0.4

The dual band DGS resonator is represented by equivalent circuit shown in Fig. 2.13. Since the structure is designed to resonate at two frequencies i.e. 2.4 GHz and 5.8 GHz, as a result, at $f_{01} = 2.4$ GHz, C_{P1} shorts L_{P1} . While at $f_{02} = 5.8$ GHz, C_{P2} shorts L_{P2} . Here, R_0 and R_1 represents 50Ω SMA feeding ports. The feeding ports are the load terminations in the equivalent circuit. Moreover, the value of inductance L_{P1} and L_{P2} are in nano henrys (nH) whereas the value of the capacitance C_{P1} and C_{P2} are in pico farads (pF). The capacitance and the inductance can be mathematically expressed by (2.6) and (2.7). The values of capacitances and inductances extracted from (2.6) and (2.7) are (2.2 pF, 1.84 nH) and (0.4 pF, 1.88 nH) at 2.4 and 5.8 GHz respectively. Furthermore, these values are substituted in the equivalent circuit to show dual band band-stop response. The results obtained for S-parameter from the circuit simulation and the EM simulation as shown in Fig. 3.9. Apparently, both simulation setup agrees with the frequencies of interest.

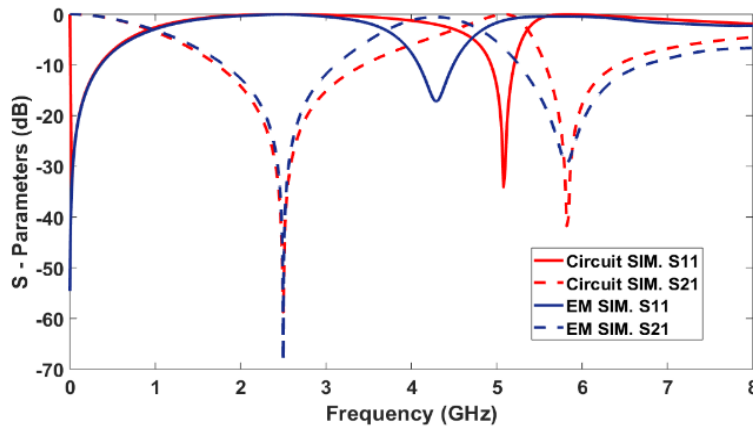


Fig. 3.9. S-Parameter plot of dual band DGS resonator.

The surface current distribution pattern shown in Fig. 3.10 depicts the respective rectangular slot sections responsible for resonance at two frequencies. Here the loops lengths ' a ' and ' c ' responsible for resonances at 2.4 GHz and 5.8 GHz show reasonably lower effect of mutual coupling among each other. At 2.4 GHz, the coupling effects between slot ' a ' and ' c ' are slightly greater in comparison to mutual coupling seen at 5.8 GHz where slot lengths ' c ' and ' a ' show independent operating characteristics.

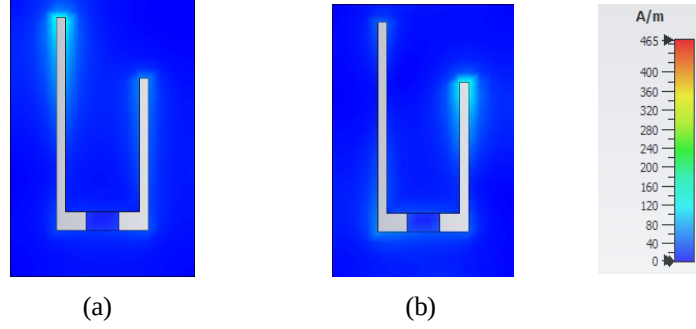


Fig. 3.10. Surface Current Distribution on the DGS resonator surface at two frequencies (a) at 2.4 GHz (b) 5.8 GHz.

This effect of mutual coupling at 2.4 GHz can be further studied by Fig. 3.11 where longitudinal length of rectangular slot ' a ' is varied between 7.5 mm and 11.5 mm to show the variation at lower and upper resonance frequencies. It can be seen that decreasing the slot length ' a ' result in small variation at 5.8 GHz. This variation at 5.8 GHz is seen due to mutual coupling. Here, the effects of mutual coupling are minimized by loading the slot lengths ' a ' and ' c ' with two external capacitors of different values.

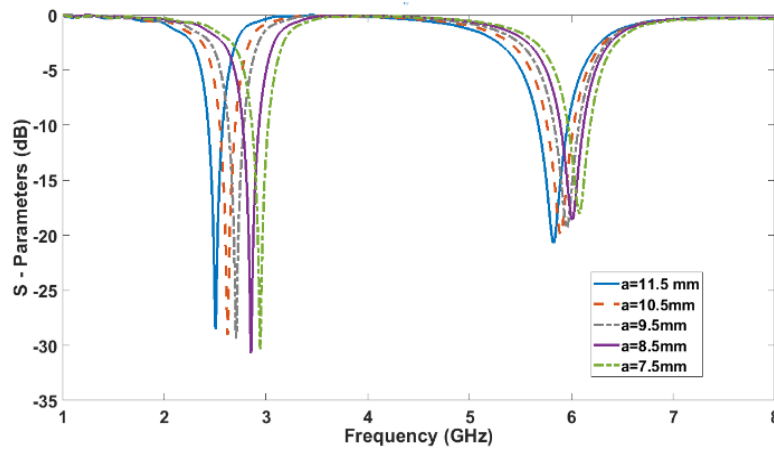


Fig. 3.11. Parametric Sweep for different values of longitudinal slot length ' a '.

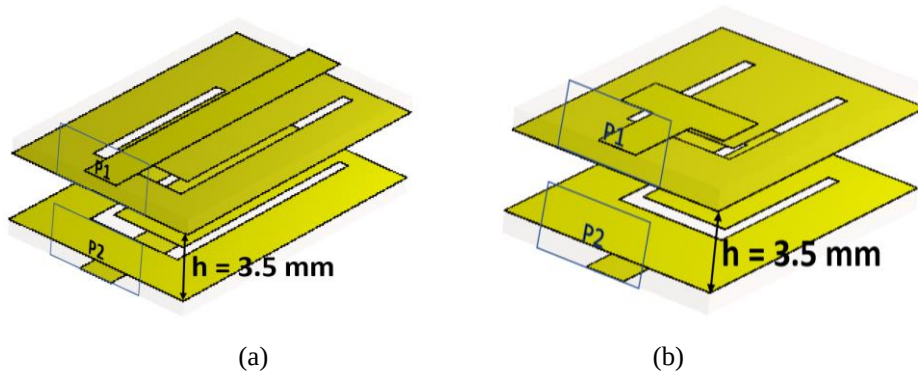


Fig. 3.12. Coupled dual band WPT system. (a) without stub. (b) with stub.

The proposed dual band wireless power transfer system as shown in Fig. 3.12(a) consists of two symmetrical defected ground structure (DGS) resonators placed back-to-back with one end of the microstrip feed line terminal connected to the 50 Ω port and the other microstrip feed line terminal in both the DGS resonators is left open. Both the DGS resonators are distance apart by ' h ' and the power transfer takes place as a result of the intersection of magnetic fields. Most importantly, the separation distance ' h ' results in impedance mismatch from desired value of input impedance ($Z_{in} = 50 + j0$). This causes the design frequency at $f_{01} = 2.4$ GHz to shift towards the frequency less than f_{01} and the design frequency at $f_{02} = 5.8$ GHz shift towards the frequency greater than f_{02} . Now, in order to obtain the design frequency at f_{01} , the rectangular slot lengths ' a ' is reduced from 11.5 mm to 8.5mm as shown in Fig. 3.13(b) which decreases the inductance value from 1.84 nH to 1.3 nH. In addition, at the design frequency f_{02} , the inductance value reduces from 1.88 nH to 1.12 nH which occurs as a result of decrease in the overall length of the U - slot defect due to decrease in slot length ' a ' with fixed length of slot ' c '. This increases the resonance frequency at f_{02} to a greater degree. Now, in order to obtain the desired resonance frequency at f_{02} , the microstrip feed line is replaced with open stub on the top plane for frequency tuning as shown in Fig. 3.12(b). Moreover, optimizations are performed in CST microwave studio to achieve resonances at 2.4 GHz and 5.8 GHz. Table 3.4 represents the design parameters of the proposed WPT system.

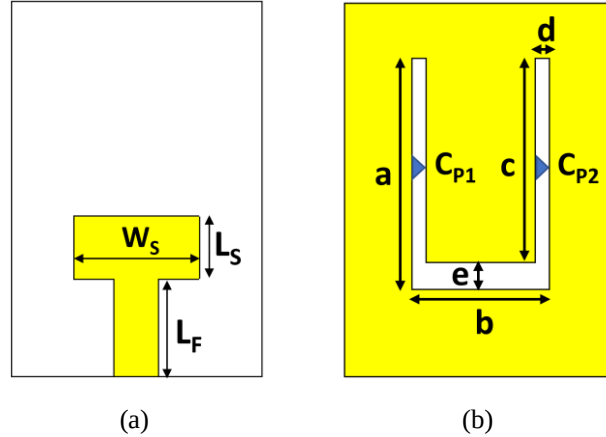


Fig. 3.13. Dual band WPT system layout. (a) Top Plane. (b) Bottom Plane.

Table 3.4: Design Parameters of the proposed dual band DGS WPT System.

Design Parameters	W_S	L_S	L_F	a	e	L_{P1}	L_{P2}	C_{P1}	C_{P2} (pF)	C_S (pF)
Dimensions (mm)	5	2.5	3.9	8.5	1	1.3	1.12	2.2	0.4	0.5

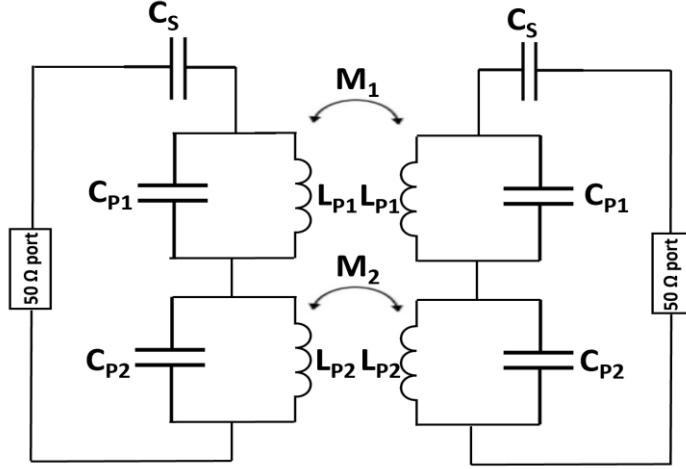


Fig. 3.14. Transformer Modelling of Dual Band Coupled WPT System.

Now, the WPT system is analyzed using transformer modelling as shown in Fig. 3.14. The model is realized by coupling the dual band defected ground structure (DGS) band stop filter (BSF) back-to-back with a separation distance (h). Here, M_1 and M_2 represents mutual inductance between the resonators operating at 2.4 GHz and 5.8 GHz. The mutual inductance is defined by coupling coefficient (k) and is mathematically shown in (3.5) and (3.6) [27].

$$k_1 = \frac{M_1}{L_{P1}}, \quad k_2 = \frac{M_2}{L_{P2}} \quad (3.5)$$

$$M_1 = \frac{\text{imag}(Z_{12})}{2\pi f_1}, \quad M_2 = \frac{\text{imag}(Z_{12})}{2\pi f_2} \quad (3.6)$$

where k_1 and k_2 represents coupling coefficient at two frequencies and $\text{imag}(Z_{12})$ represents imaginary part of impedance element Z_{12} . Furthermore, to obtain maximum power transfer, impedance matching at the two frequencies of interest is achieved by incorporating open stub of length L_S and width W_S which acts as a series capacitor. The open stub is positioned at the end of microstrip feed line of length L_F as shown in Fig. 3.13(a). The length of open stub is carefully selected in order to provide matching at two frequencies. The capacitance (C_S) corresponding to the length of the open stub is shown using (3.7).

$$C_S = \frac{\tan(\beta l_{st})}{2\pi f_0 Z_0} \quad (3.7)$$

Where, $\beta = \frac{2\pi}{\lambda_g}$ rad/m and $\lambda_g = \frac{300}{f_0(\text{GHz})\sqrt{\epsilon_r}}$ mm

The efficiency of the WPT system is computed from the U -factor method which is given by the product of coupling coefficient (k_1 and k_2) and unloaded quality factor (Q_{u1} and Q_{u2}).

$$U_1 = k_1 Q_{u1} \text{ and } U_2 = k_2 Q_{u2} \quad (3.8)$$

$$Q_{u1} = \frac{\omega_{01} L_{P1}}{R_1} \text{ and } Q_{u2} = \frac{\omega_{02} L_{P2}}{R_2} \quad (3.9)$$

Here, U_1 and U_2 depicts the U-factors computed at two different operating frequencies and the maximum efficiency using both the U -factors are calculated using the equation in [27]. In addition, the resistances R_1 and R_2 represents the radiation loss, dielectric loss, and conductor loss in the defect at f_{01} and f_{02} respectively [24].

Fig. 3.15 shows a comparison between the circuit simulations using ADS and the EM simulations using CST Microwave studio of the dual band DGS-WPT system. The magnitude of reflection ($|S_{11}|$) coefficient and transmission ($|S_{21}|$) coefficient show good agreement.

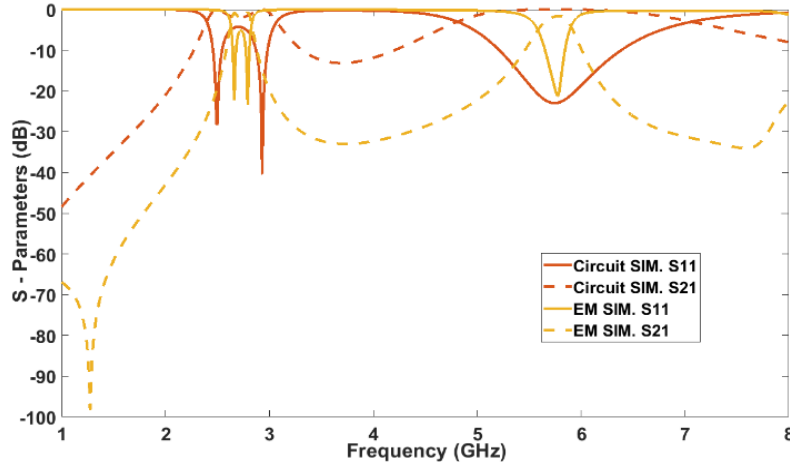


Fig. 3.15. Comparison of EM simulation and Circuit simulation S-parameter plot of dual band DGS WPT system.

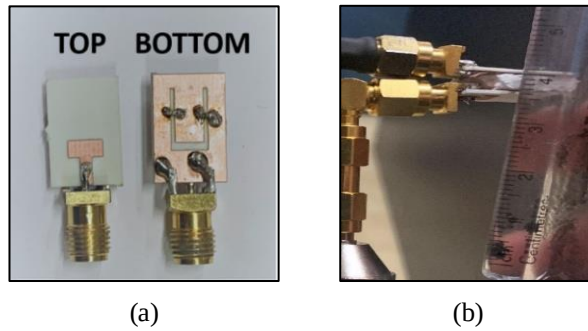


Fig. 3.16. Prototype of U shape DGS resonator WPT system. (a) Top and Bottom side, (b) Measurement setup.

A prototype, shown in Fig. 3.16(a), of the proposed WPT system has been developed on Rogers 4003C to evaluate its concept and the performance. The measurement is carried out using Keysight VNA over the frequency range of 1-8GHz. The measurement setup, in Fig. 3.16(b), has two symmetric U shape DGS resonators placed back-to-back and separated by plastic foam in

order to fix the distance to 3.5 mm. Here, one of the resonators acts as a transmitter while the other as receiver.

In Fig. 3.17, the EM simulated results show resonance at two frequencies i.e. $f_{01} = 2.4$ GHz with $|S_{21}| = -1.97$ dB and $|S_{11}| = -12$ dB and $f_{02} = 5.8$ GHz with $|S_{21}| = -1.55$ dB and $|S_{11}| = -20$ dB. On the other hand, in the measurement result, the dual band WPT system resonates at $f_{01} = 2.2$ GHz with $|S_{21}| = -2.4$ dB while $|S_{11}| = -12$ dB. Whereas at $f_{02} = 5.3$ GHz with $|S_{21}| = -1.8$ dB while $|S_{11}| = -23.3$ dB. The simulation and measured result have a slight shift in the resonance frequency from the desired frequency probably due to parasitic capacitance effects at the operating frequencies. Also, it can occur due to some tolerances in the loading capacitor or the soldering loss. To overcome the shift, the distance between the loop lengths ‘a’ and ‘c’ can be increased, capacitor values of low tolerance and frequency tuning can be incorporated. Furthermore, it can be seen that frequency splitting phenomenon in addition to ripples occur in the measured S-parameter plot. These results due to angular offset near the SMA connector and narrow separation distance between the two DGS resonators.

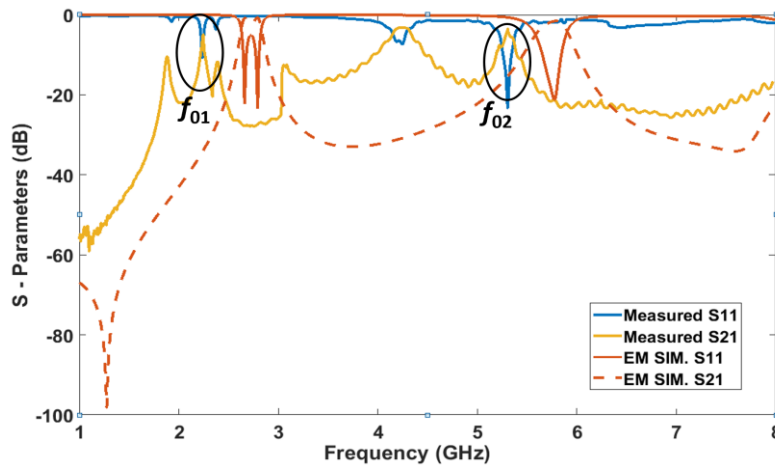


Fig. 3.17. Simulated and Measured S-Parameter of the WPT system.

Furthermore, the Table 3.5 shows the comparison of different types of dual band and single band WPT system. The performance of the proposed WPT system is evaluated based on frequency, measured efficiency (η) and the $h/\sqrt{\text{Size}}$ ratio. The proposed design achieves the measured efficiency of 60% and 67% at f_{01} and f_{02} respectively.

Table 3.5: Comparison with other compact WPT Systems.

WPT System	Frequency (GHz)	Size (mm ²)	h (mm)	η (%)	$h/\sqrt{\text{Size}}$
[42] *	0.11	15x10	3	30	0.073
[22] *	1.43	20x20	3.5	85	0.148
[24]	0.3 and 0.7	30x15	16	71 and 72	-
[59]	0.0903 and 0.1388	50x50	40	70 and 69	0.8
This work	2.4 and 5.8	15x10	3.5	60 and 67	0.285

*Single Band WPT System

3.3. Conclusion

In this chapter, we proposed two different types of compact highly efficient resonant inductive WPT systems. The first one is based on dual E-shaped based DGS resonators operating at 0.27 GHz frequency and utilizes an area of $20 \times 20 \text{ mm}^2$. These resonators possess a high unloaded Q factor of 470 to obtain high WPT system efficiency of 71 %. This WPT system is modeled using a parallel RLC resonant circuit and J-inverter circuit in conjunction with coupled mode theory (CMT) and S-parameters. On the other hand, the second WPT system proposed is based on U shaped DGS resonators operating at 2.4 GHz and 5.8 GHz frequencies and takes an area of $15 \times 10 \text{ mm}^2$. The equivalent circuit of the U-shape DGS WPT system is realized using transformer modeling and achieves a measured efficiency of 60% and 67% at 2.4 GHz and 5.8 GHz. Finally, both the WPT systems are compact and achieve improved performance metrics in comparison to the other reported designs in the literature.

Chapter 4

Impedance Matching Technique: Coupled Line

This chapter is based on the following research paper:

- S. Verma, D. Rano and M. S. Hashmi, “Enhancing the performance of defected ground structure type near-field radiofrequency WPT system by coupled-line impedance matching,” *IET Microwaves, Antennas and Propagation*, vol. 14, no. 12, pp. 1431-1439, Oct. 2020, doi: 10.1049/iet-map.2020.0217.

In this chapter, a new coupled-line-based impedance matching technique to enhance the efficiency and robustness for resonant coupled DGS-WPT system is proposed. In this impedance matching technique, a symmetric coupled lines on the top plane of the DGS resonator are utilized. The utilization of coupled line in impedance matching provides wideband matching performance which can be convenient in the case of any mismatch during the development of equivalent circuit of the WPT system and then in the practical realization [60]. This is one of the key benefits of coupled-line matching over other matching techniques such as the open stub technique, exhibiting narrow band performance, used in the design of WPT systems [23][27]. Furthermore, the coupled-line matching technique ensures perfect matching for wide range of input impedance (Z_{in}) to source impedance (Z_s) owing to the presence of even- and odd-mode characteristic impedances and the possibility to readily engineer them according to the need. This results in an improved and more versatile impedance matching and the eventual enhancement in the efficiency and other performance metrics. Furthermore, the choice of the triangular-shaped defect in the ground plane of the DGS resonator is regulated by the possibility of higher inductance which aids in reaching the resonance readily while keeping the goal of miniaturization as a perspective.

The next section revisits the modelling and design of triangular shaped DGS and realization of its filter characteristics. Section 4.2 presents the realization of the WPT system incorporating coupled lines and the adopted design strategy using the equivalent circuit model. Subsequently, experimental evaluation of the proposed technique and comparisons with the existing state-of-the-art is included in Section 4.3.

4.1. Design and Analysis of Triangular Shaped DGS Resonator

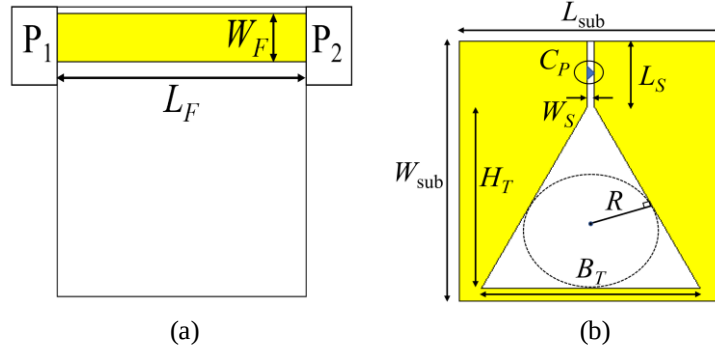
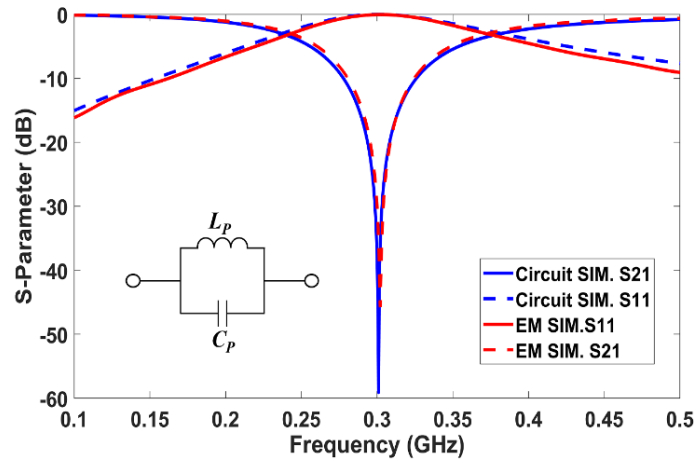
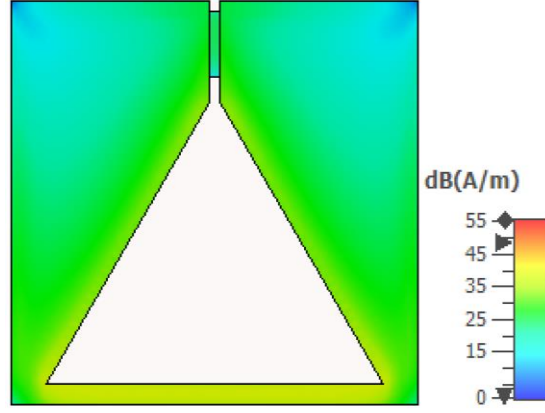


Fig. 4.1. Single Band DGS Resonator. (a) Top Plane. (b) Bottom Plane.

The triangular shaped DGS resonator, depicted in Fig. 4.1, operating at a frequency of 300 MHz is realized on Rogers RO4350B having $\epsilon_r = 3.66$, substrate thickness $h_{\text{sub}} = 1.524$ mm, and metal thickness $t = 35$ μm . The top layer of the substrate is a $50\ \Omega$ microstrip feed line, of length L_F and width W_F , and is connected to lumped ports P_1 and P_2 as shown in Fig 4.1(a). The ground plane of overall length L_{sub} and width W_{sub} , shown in Fig. 4.1(b), consists of a triangular defect of height H_T and base B_T with internal angle of 60° , and excitation slot of length L_S and width W_S . The excitation slot of the resonator is mounted with an appropriate external capacitor C_P to achieve the desired operating frequency. Furthermore, the EM simulation is carried out in CST microwave studio based on two different scenarios. Firstly, without C_P , the DGS resonator achieves resonance characteristics at 1.6 GHz. In the second case, the C_P is incorporated so as to fulfill the requirements of physical size (20×20 mm²) restriction and target frequency. Subsequently, optimizations are carried out by increasing the height of the triangular defect in the ground plane so that the electrical length increases and this in turn reduces the resonant frequency. The optimized design parameters of the chosen triangular DGS resonator are listed in Table 4.1.



(a)



(b)

Fig. 4.2. Band-stop Performance of triangular DGS resonator: (a) S-parameter plot of EM and circuit simulations. (b) EM simulated surface current distribution on the DGS surface.

Table 4.1: Design Parameters of Single Band DGS Resonator

Design Parameters	L_F	W_F	L_{sub}	W_{sub}	H_T	B_T	L_S	W_S	L_P (nH)	C_P (pF)
Dimensions (mm)	20	3.28	20	20	14	16.6	5	0.5	24.64	11.3

The developed resonator is then modelled as an equivalent parallel resonant LC circuit, shown in Fig. 4.2(a), to aid in the expedited design and optimization. The inductance L_P , contributed by the current flowing along the periphery of triangular shaped DGS, and the capacitance C_P , representing the capacitor mounted on the excitation slot, are computed using (4.1) and (4.2) [57], [61]. These are key elements in creating the band-stop (ω_0) characteristics for the resonator.

$$L_P = \frac{\mu_0 P}{2\pi} \left[\ln \left(\frac{2P}{R} \right) + 0.25 - \ln \left(\frac{P^2}{A} \right) \right] \quad (4.1)$$

Here,

μ_0 = Magnetic Permeability of free space, $4\pi \times 10^{-7}$ H/m.

P = Perimeter of triangle.

R = Radius of metal.

$$C_P = \frac{1}{\omega_0^2 L_P} \quad (4.2)$$

$$R_{DGS} = \text{real}(Z_{11}) \quad (4.3)$$

$$Q_U = \frac{\omega_0 L_P}{R_{DGS}} \quad (4.4)$$

In the above expression, μ_0 is the magnetic permeability of free space ($4\pi \times 10^{-7}$ H/m), P is the perimeter of the triangular DGS, R is the inscribed circle radius, and A is the area of the DGS triangle. In general, L_P and C_P are in the range of nano henry (nH) and pico farad (pF). The resistance R_{DGS} , expressed in (4.3), represents the losses in the DGS resonator [62]. This is obtained from one port Z-parameter (Z_{11}) EM simulation by removing the microstrip feed line from the top plane and then substituting a lumped port in the excitation slot. Finally, the unloaded quality factor (Q_U) of the resonator, expressed in (4.4), relies on the R_{DGS} and enables the determination of the losses in the resonator and the identification of efficiency of WPT system [23]. The values of inductance and capacitance extracted from (4.1) and (4.2) are substituted in the LC equivalent circuit and simulation is carried out in ADS (Advanced Design System) for comparing the results obtained in ADS and CST environments at the operating frequency of 300 MHz.

The S-parameter plot obtained from both circuit simulation and the EM simulation, Fig. 4.2(a), compare favorably and thus the reliability of the developed circuit model is validated. In addition, the EM simulation is used to obtain the surface current distribution, of electromagnetic wave on the surface of the triangular shaped DGS as shown in Fig. 4.2(b). It can be seen that the surface current flowing around the periphery of the triangular defect perturbs the field propagating through the microstrip feed line at the operating frequency.

4.1.1. Investigation of DGS geometries

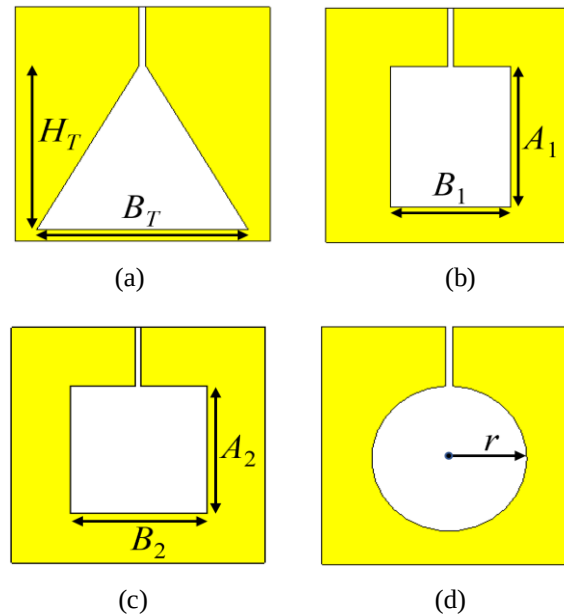


Fig. 4.3. Layout of four different DGS geometries. (a) Triangular geometry. (b) Rectangular geometry. (c) Square geometry. (d) Circular geometry.

The layout of four different DGS shapes namely triangular geometry, rectangular geometry, square geometry, and circular geometry) are shown in Figs. 4.3(a)-(d). Here, the comparison among four different DGS geometries are performed on the basis of similar design parameters. The design parameters of the proposed triangular DGS resonator such as substrate material, overall length L_{sub} and width W_{sub} , microstrip feed line of length L_F and width W_F , excitation slot of length L_S and width W_S , and external capacitor C_P are utilized to design the other three DGS resonators. In addition, the area of rectangular defect ($A_1 \times B_1 = 12 \times 9.7 \text{ mm}^2$), square defect ($A_2 \times B_2 = 10.8 \times 10.8 \text{ mm}^2$), and circular defect ($\pi r^2 = \pi (6.1)^2 \text{ mm}^2$) is almost similar to the area occupied by the triangular defect ($0.5 \times H_T \times B_T = 116.2 \text{ mm}^2$).

The EM simulated $|S_{21}|$ plot, as shown in Fig. 4.4, demonstrates that the triangular defect has the highest inductance value in comparison to rectangular, square, and circular shaped defects in similar area as can be deduced from the smallest resonant frequency for the triangular DGS. This high inductance value of the triangular defect is due to the tapered side width and length which increases the physical length of the defect to a greater degree. The increased physical length of the defect results in the increased electrical length and therefore aids in achieving resonance more readily.

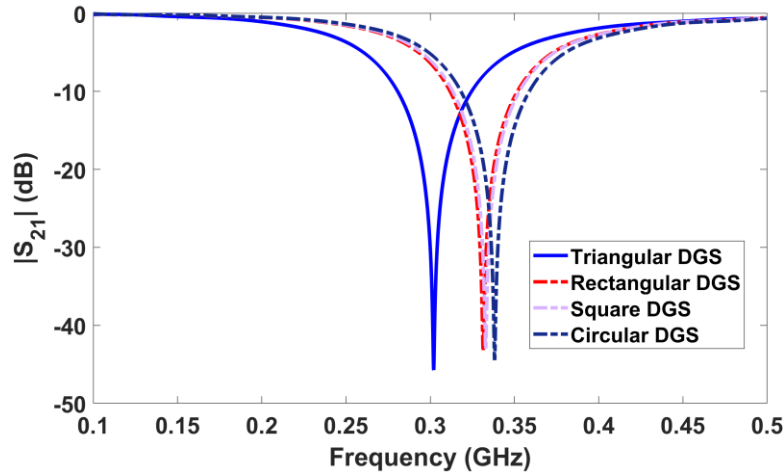


Fig. 4.4. Comparison of EM simulated $|S_{21}|$ plot of triangular, rectangular, square, and circular shaped DGS resonator.

4.2. WPT System Design

The proposed DGS based WPT system makes use of the modified resonator shown in Fig. 4.5(a). It consists of a microstrip feed line followed by a symmetric defected coupled-line of length L , width W , and gap S on the top plane. The length of microstrip feed line on the top plane in this design is reduced to half the total length of the substrate. This makes sure that the excitation slot of DGS resonator lies exactly below the center of substrate which acts as a location for analyzing the input impedance (Z_{in}). The impedance matching network is also

substituted at that point in the eventual WPT system. Moreover, an external capacitor C_S mounted between the energized coupled-line segment and the passive coupled-line segment provides an additional capacitance for impedance matching. The C_S is required when the capacitance realized by the coupled-line is not sufficient for perfect match at the desired resonant frequency. Both the C_S and the coupled-line of length L provide the frequency tuning capability and thus help in finely adjusting the resonance frequency close to the operating frequency. Furthermore, the ground plane, shown in Fig. 4.5(b), is exactly similar to the ground plane of the original resonator as shown in Fig. 4.1(b) and hence the design parameters for bottom plane remains same except the capacitor C_P as explained in the design procedure.

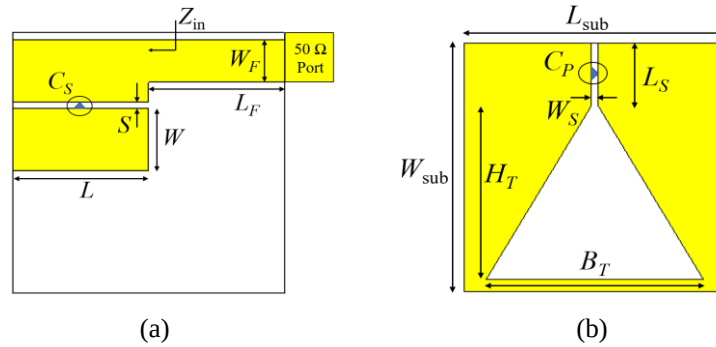


Fig. 4.5. Use of coupled-lines in the modified DGS resonator structure in the proposed WPT system.
(a) Top plane. (b) Bottom plane.

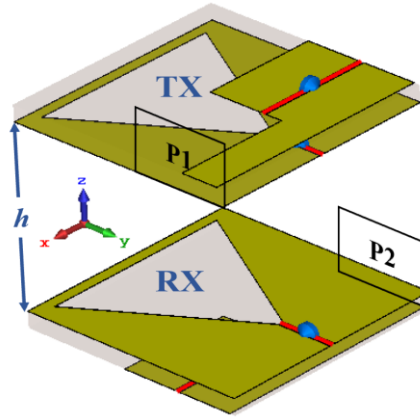


Fig. 4.6. Back-to-back view of the proposed DGS based WPT system including the direction of flow.

The proposed WPT system is realized by placing two modified resonators back-to-back separated by a distance h as shown in Fig. 4.6. Apparently, the resonators are symmetric and have coupled-line on the top plane. One end of the microstrip feed line is connected to ports P1 and P2 while the other end of both the resonators are left open. The separation distance h between the two resonators transforms the band-stop characteristics of the DGS resonators into band-pass characteristics. It facilitates power transfer from TX to RX at the operating frequency (300 MHz in this instance). The power transfer between the two DGS resonators follows the

principle of non-radiative near field resonant inductive coupling. This can also be inferred from Fig. 4.7 which shows the EM simulated magnetic field current distribution plot of the proposed DGS based WPT system at different phases Φ (0° , 90° , 180° , and 270°). In the plot, for each 90° variation in the phase, the corresponding length $l = \lambda/4$ (using the formula $\Phi = \beta l$ and substituting $\theta = 90^\circ$ and $\beta = 2\pi/\lambda$). Therefore, for each length, $l = \lambda/4$, $I_{\max}$ (maximum current) becomes $I_{\min}$ (minimum current) and vice versa. For example, at $\Phi = 90^\circ$, the current in the TX goes to $I_{\min}$ from $I_{\max}$ at $\Phi = 0^\circ$. On the other hand, in RX end, at $\Phi = 90^\circ$, the current goes to $I_{\max}$ from $I_{\min}$ at $\Phi = 0^\circ$.

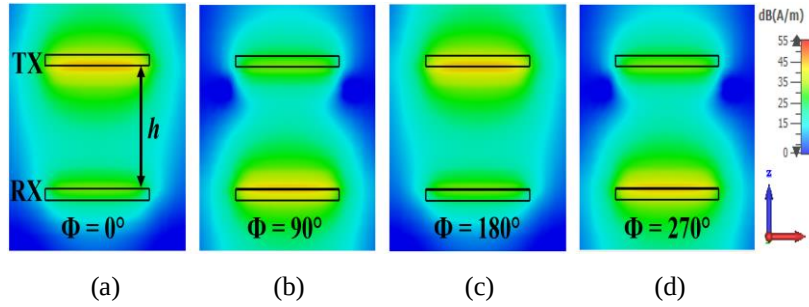


Fig. 4.7. EM simulated magnetic field distribution of the proposed DGS based WPT system at 300 MHz at plane Y = 0

(a) $\Phi = 0^\circ$ (b) $\Phi = 90^\circ$ (c) $\Phi = 180^\circ$ (d) $\Phi = 270^\circ$

4.2.1. Design Procedure

The proposed WPT system includes the effect of DGS in the coupled-lines and therefore the associated design process of this modified resonator and hence the full WPT system is different as depicted in Fig. 4.8. The design flow considers both the scenarios whether C_s is required or not for achieving perfect impedance matching. The process is explained in simplified steps below by considering TX. Similar steps are required for resonator employed in the design of RX.

Step 1: Specify the design requirements of the WPT system i.e., the operating frequency f_0 , size ($L_{\text{sub}} \times W_{\text{sub}}$) and choose the substrate parameters such as relative permittivity (ϵ_r), loss tangent ($\tan\delta$), height (h_{sub}), thickness (t) and conductivity (σ).

Step 2: Using CAD tool such as ADS, determine the physical parameters of coupled-line namely the width (W), the gap (S) to fulfill the coupling factor (C), and the length (L). For practical WPT systems, C is in the range of $-10\text{dB} > C > -20\text{dB}$ and depends on fabrication complexity and impedance matching issues. For the determination of L , assume $L = L_{\text{sub}}/2$.

Step 3: Consider the initial value of transmission distance $h = L_{\text{sub}}/2$. Then the transmission distance is varied until maximum power transfer efficiency at critical coupling.

Step 4: Utilize the physical parameters of coupled line computed in Step 2 along with the defect in the ground plane of the TX and RX.

Step 5: Tune L and W of coupled-line until the desired value of Z_{in} ($= 50 + j0$) is obtained. For length, maintain $L_{sub}/2 \leq L \leq L_{sub}$ while tuning to achieve the imaginary part of Z_{in} equal to zero. For $R_{in} > 50 \Omega$, increase the W of coupled-line whereas for $R_{in} < 50 \Omega$, decrease the W of coupled-line to obtain $Re(Z_{in}) = 50 \Omega$. Furthermore, if the desired value of Z_{in} is not obtained by tuning L and W then external capacitor C_S is required to achieve perfect matching.

Step 6: Once Z_{in} is matched to the port impedance of 50Ω , use the optimized physical parameters obtained from the EM simulation to extract electrical parameters. Finally, utilize these electrical parameters to generate the equivalent circuit for system evaluation and performance validation.

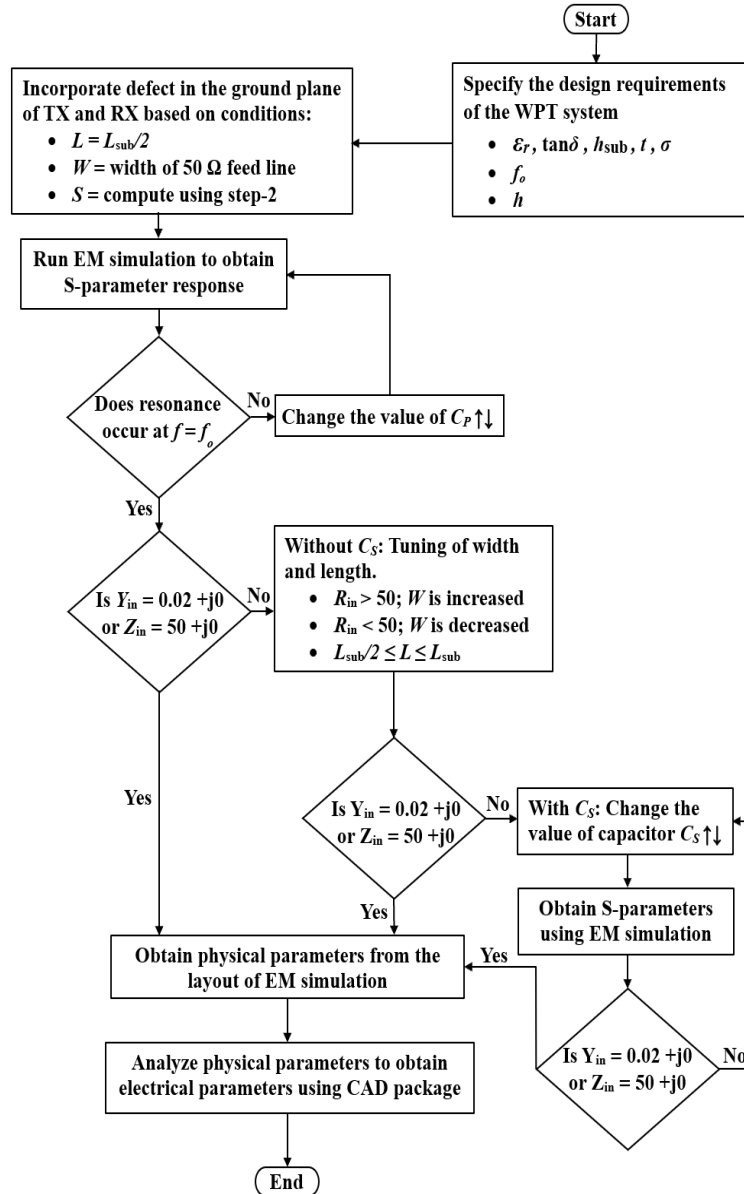


Fig. 4.8. WPT system design-flow using coupled-line with DGS.

An important aspect to consider during the design of WPT system is that a shift in the resonance frequency from the design frequency is very common. It is due to the effect of mutual coupling between transmitter and receiver caused by the separation distance. This issue can be overcome by tuning the C_P until the design frequency (f_0) is obtained.

4.2.2. Equivalent Circuit modelling

The proposed WPT system can be modelled by its equivalent circuits shown in Figs. 4.9 and 4.10. Apparently, there are slight differences in these two and the subsequent descriptions provide clarity in this regard. The circuit in Fig. 4.9 consists of two symmetric DGS resonators modelled by equivalent parallel LC circuit and coupled-lines. Both the resonators are coupled through mutual inductance L_m and hence the coupling coefficient k [23].

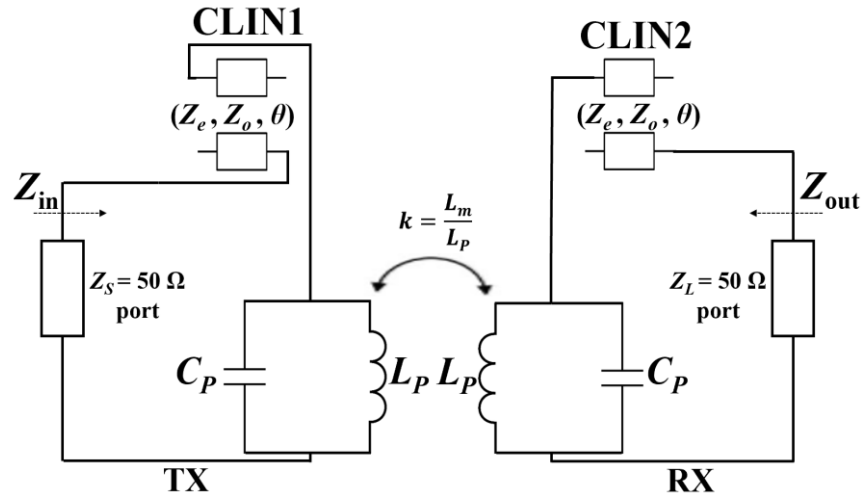


Fig. 4.9. Equivalent circuit of DGS WPT system with coupled-lines based impedance matching.

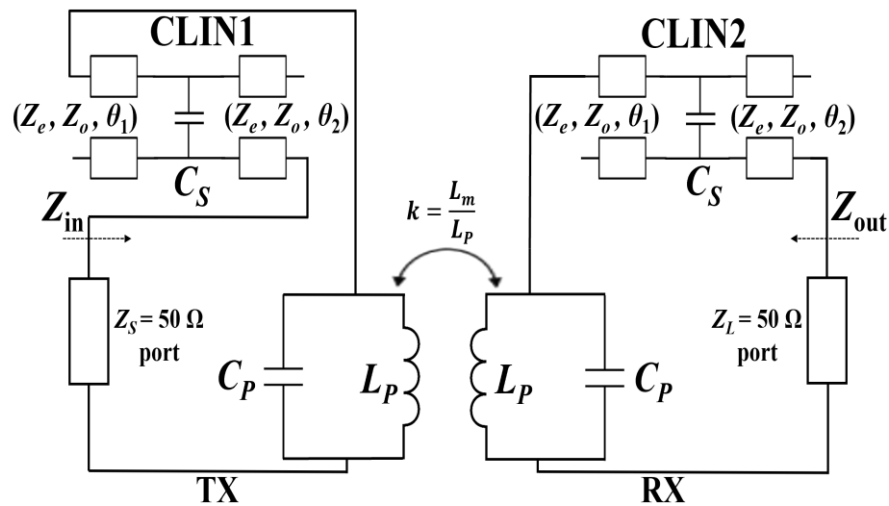


Fig. 4.10. Equivalent circuit of DGS WPT system using coupled-lines impedance matching and external capacitor (C_S).

In these equivalent circuits, the coupled-lines (CLIN1 and CLIN2) between the port terminations and the TX-LC circuit and RX-LC circuit takes care of the impedance mismatches between the Z_{in} and the Z_S . The equivalent circuit model in Fig. 4.10 depicts the scenario when external capacitor C_S is required for matching. Once again, the arrangement of two similar coupled-lines are used between the port termination and both TX-LC and RX-LC circuits. The two coupled-lines are connected in parallel by capacitor C_S so that the electrical length (θ_1 and θ_2) of both the lines combine to give the total electrical length (θ). For design ease, the electrical length of both the coupled-lines are distributed in equal proportion.

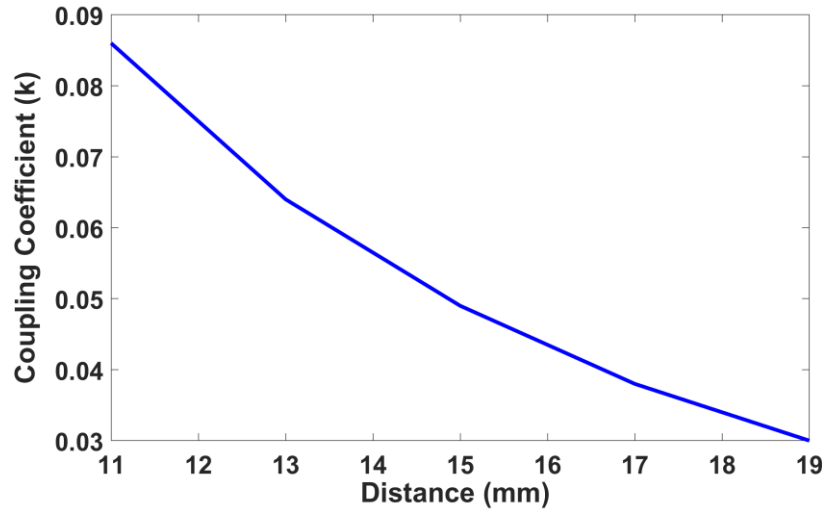
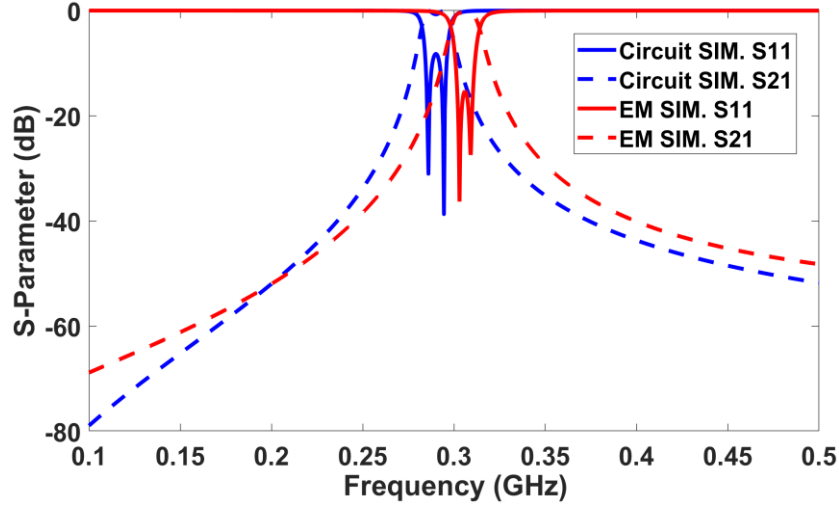


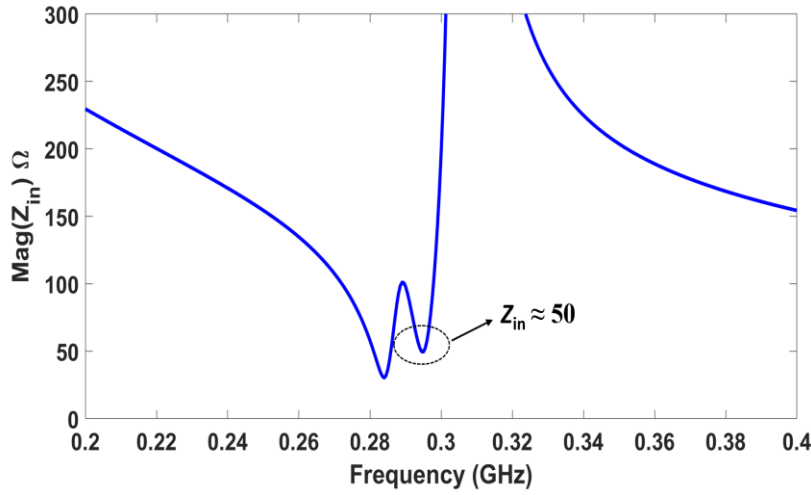
Fig. 4.11. Variation of coupling coefficient w.r.t. distance.

The coupling coefficient (k) between the TX and RX resonator is plotted with respect to the transmission distance in Fig. 4.11. The value of coupling coefficient is recorded for increasing value of transmission distance from 11 mm to 19 mm at the operating frequency. It can be seen from the plot that the coupling is maximum at 11 mm and minimum at 19 mm and it is in line with the established theory. Moreover, keeping the overall size of the WPT system in perspective, the range of separation distance is kept between 11 mm and 19 mm due to the fact that the WPT system is in over-coupled region within this range. The coupling becomes very large below 11 mm, whereas it reduces and eventually reaches close to zero at a separation distance greater than 19 mm. Both these scenarios lead to degradation of the WPT system efficiency. In this analysis, the points in the range nearer to the critically coupled distance have been considered.

Subsequently, the circuit model is simulated in ADS. It can be seen in Fig. 4.12(a) that the circuit simulation results, although exhibits similar shape, experience some variation from the EM simulation results. It occurs due to slight overlap of triangular defect and excitation slot in the ground plane with the coupled line on the top plane. The effect of triangular defect and excitation slot on the coupled line has not been considered in the circuit simulation.



(a)



(b)

Fig. 4.12. EM simulation and circuit simulation plot of the proposed WPT system. (a) Comparison of EM simulation and circuit simulation S-parameters. (b) Magnitude of input impedance vs frequency.

In essence, this overlap leads to reduction in the physical length and the electrical length of the coupled line. Furthermore, the overlap increases the even mode impedance (Z_e) substantially and the odd mode impedance (Z_o) marginally. Apparently, the reduction in the physical length result in the decrease of gap capacitance. As a consequence, the reduced gap capacitance along with the increased even- and odd-mode impedances lead to an increase in the resonance frequency and therefore we observe the shift between the circuit simulation and EM simulation results. Moreover, the shift in the circuit simulation result can also be attributed to anomaly in substrate parameters, such as losses, not taken into consideration while developing the equivalent circuit model. In addition, small frequency split can also be observed in both the EM simulation and circuit simulation results and these convey the transformation from the over coupled to critically coupled region.

Furthermore, a swept frequency response for input impedance (Z_{in}) using the circuit model gives the outcome presented in Fig. 4.12(b). It can be inferred that the desired impedance of 50Ω is obtained at 295 MHz and once again it demonstrates the effectiveness of the model in cross validation of performance.

In addition, the proposed equivalent circuit is very effective in achieving desired resonant frequency quickly. The impedance matching obtained using coupled line and the external capacitor C_S is depicted in Fig. 4.13. The value of external capacitor C_S is varied between 1.2 pF and 2.4 pF to study the wideband matching performance of the proposed WPT system. It is apparent that the increase in C_S leads to shift in the resonance frequency downwards while still maintaining decent power transfer. Just for clarity, an optimum wideband matching performance ($|S_{11}| < -30$ dB), not given here, is maintained between 283 MHz and 300 MHz for three different C_S i.e. 2pF, 1.6 pF, and 1.2 pF but degrades at larger C_S of 2.4 pF.

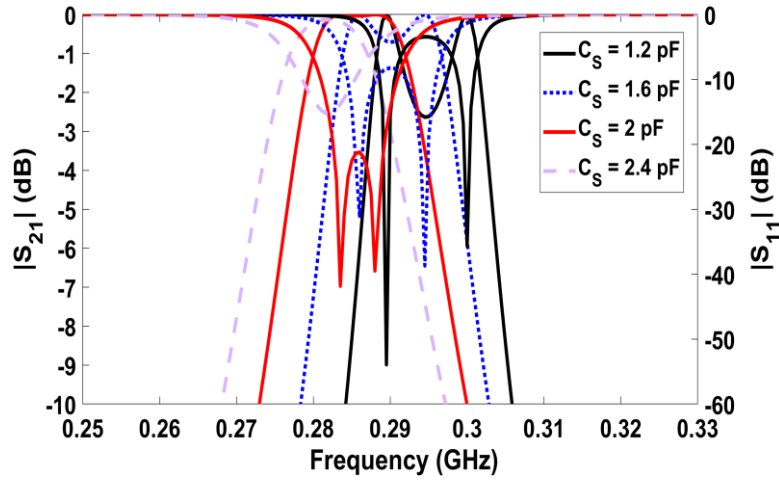


Fig. 4.13. Circuit simulated S-parameter for various C_S values

To understand the effectiveness of the proposed design and modeling approach, the WPT system is also designed at frequency 700 MHz. The performance and parameters were easily optimized using the proposed approach and the Table 4.2 lists them for ready reference.

Table 4.2: Design Parameters of Single Band WPT System using Triangular shaped DGS

Design Parameters	$f_0 = 300$ MHz	$f_0 = 700$ MHz	Design Parameters	$f_0 = 300$ MHz	$f_0 = 700$ MHz
	Optimized Values			Optimized Values	
L_{sub} (mm)	20	20	L_S (mm)	5	5
W_{sub} (mm)	20	20	W_S (mm)	0.5	0.5
L_F (mm)	10	10	H_T (mm)	14	11
W_F (mm)	3.28	3.28	B_T (mm)	16.6	14.2
L (mm)	10	8	L_P (nH)	24.64	19.8
W (mm)	4.8	2	C_P (pF)	9.1	2
S (mm)	0.5	0.8	C_S (pF)	1.6	-

$Z_e (\Omega)$	45	79.3	$R_{DGS} (\Omega)$	0.088	0.159
$Z_o (\Omega)$	30	49.88	Q_U	530	549.5
$\rho = Z_e/Z_o$	1.5	1.59	L_m (nH)	0.971	0.92
θ (deg)	6.09	11	k	0.0379	0.046
$R_{in} (\Omega)$	49.7	49.5	h (mm)	17	17
$X_{in} (\Omega)$	0.4	0.9	η (%)	83	94

4.3. Measurement Results

The prototype in Fig. 4.14(a) depicts the bottom and top plane of the proposed resonator. The realization of WPT system requires back-to-back connection of two such resonators. One of the resonators acts as a TX and the other as RX. The resonators are separated by plastic foam, as shown in the measurement setup of Fig. 4.14(b), to maintain a distance h of 17mm between them for optimal efficiency. This optimal distance was achieved in simulation. The measured S-parameter results obtained for the WPT system are plotted in Fig. 4.15. The result clearly shows resonance characteristics at $f_0 = 300.5$ MHz with $|S_{21}| = -1$ dB and $|S_{11}| = -21$ dB. The measured result, although excellent, show slight variation in the resonance frequency from the operating frequency of 300 MHz. This occurs due to the tolerances of the capacitors C_P and C_S (part numbers CBR04C919B5GAC and CBR04C169B5GAC) and the parasitic effects inherent in the designed system. In addition, other factors such as the inset of plastic foam which adds on to the permittivity, and the in-house fabrication which introduce losses such as soldering loss, connector loss, etc. also introduce some anomaly.

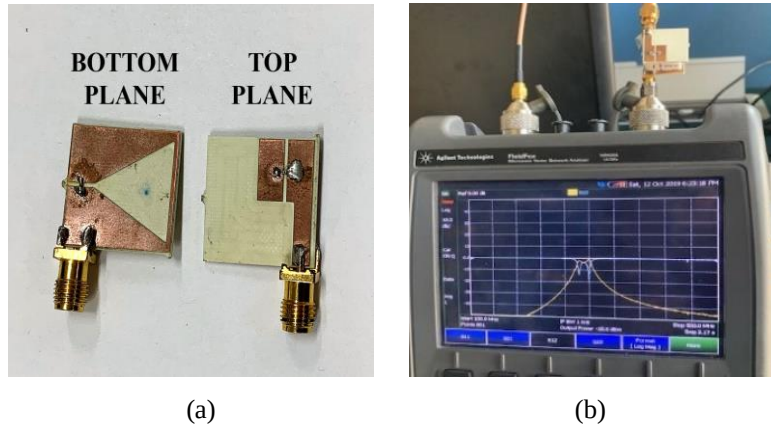


Fig. 4.14. Prototype and measurement. (a) Bottom and top side. (b) Measurement setup.

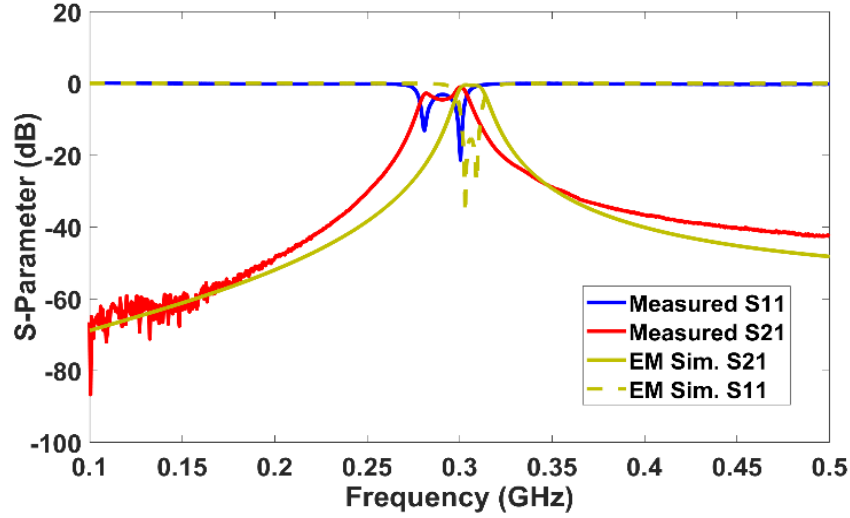


Fig. 4.15. Measured and EM simulated S-parameters

To overcome the shift in the frequency, capacitor values with low tolerances and frequency tuning should be used. Moreover, it can be seen that the obtained bandwidth from measurement is larger when compared to the EM simulation. This once again can be attributed to the losses in the connector, substrate, and metal which decrease the quality factor of the WPT system.

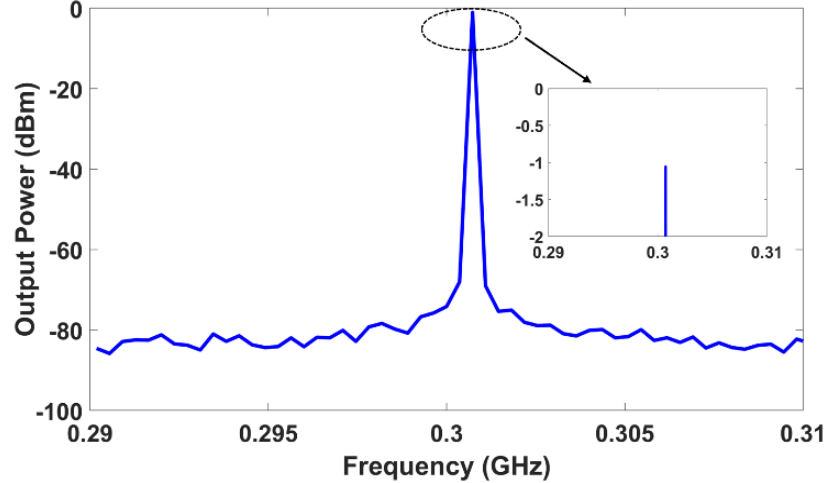


Fig. 4.16. Received output power measured using power spectrum analyzer at the design frequency.

The measured efficiency (η_{WPT}) of the developed WPT system, computed using (4.5) [24], is 80% at 300.5 MHz. Subsequently, the measured efficiency of the WPT system at the operating frequency is validated using Anritsu MS2720T spectrum master. Here, an input power of 0 dBm at 300.5 MHz is fed to the transmitter DGS resonator and the spectrum of the output power is recorded at the receiver DGS resonator as shown in Fig. 4.16. The received output power at 300.5 MHz is -1 dBm over 17 mm separation distance respectively. The ratio of output

power and input power gives similar efficiency as computed by (4.5) and thus provides confidence in the measured efficiency value obtained using VNA setup. This obtained efficiency is excellent and far superior to other recently reported DGS based WPT systems listed in Table 4.3. This inference is made by comparing and trading-off various parameters such as resonant frequency, size, power transmission distance and efficiency.

$$\eta_{WPT} = \frac{|S_{21}|^2}{1 - |S_{11}|^2} = \frac{\text{Transferred Power (TP)}}{1 - |S_{11}|^2} \quad (4.5)$$

$$FoM = \eta \times \frac{h}{\sqrt{\text{Size}}} \quad (4.6)$$

The proposed WPT system is also compared on the basis of figure of merit (FoM) calculated using (4.6) [24]. The FoM provides a platform for estimating the performance of various WPT systems by considering the effects of distance, efficiency, and the size of the WPT system. It can be seen in Table 4.3 that the proposed design exhibits an FoM of 0.68 which is significantly improved when compared to other reported systems operating in the same frequency range.

Table 4.3: Performance comparison of proposed WPT system with the state-of-the-art designs.

WPT system	Matching technique	Frequency (GHz)	Size (mm)	Distance (mm)	Efficiency (%)	Figure of Merit (FoM)
[27]	Open stub	0.27	20 X 20	11	71	0.390
[23]	Open stub	0.3	20 X 20	13	68	0.442
[63]	Open stub	0.88	23 X 12	8	70	0.34
[22]	Open stub	1.43	20 X 20	3.5	85	0.148
This work	Coupled-line	0.3	20 X 20	17	80	0.68

4.3.1. Misalignment Study

The WPT system can be greatly impacted by lateral and angular misalignments between the TX and RX. In order to understand this impact, the TX and RX are laterally shifted along the horizontal (X-axis) direction and then along the vertical (Y-axis) direction as depicted in Figs. 4.17(a) and 4.17(b) respectively. The effect of lateral shift on efficiency is studied and the obtained results are shown in Figs. 4.18(a) and 4.18(b).

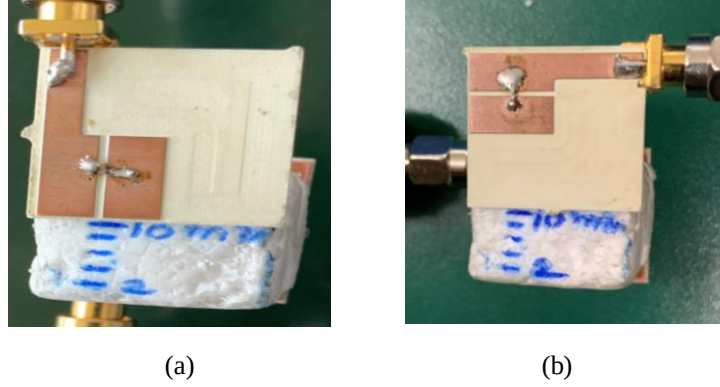
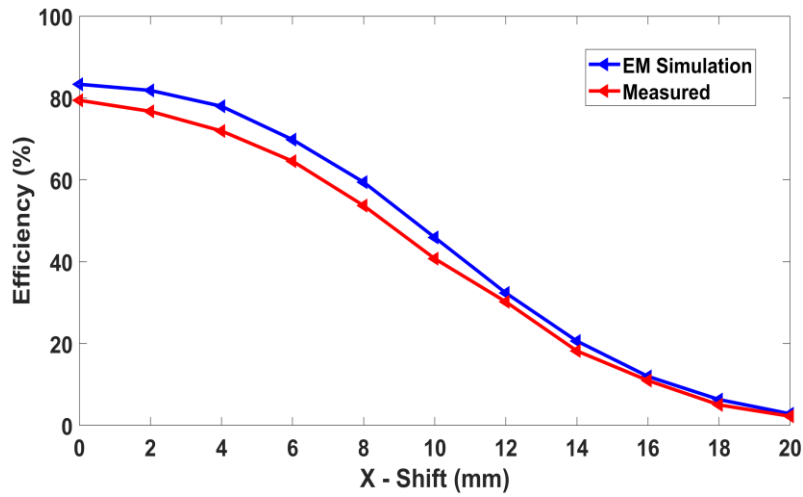
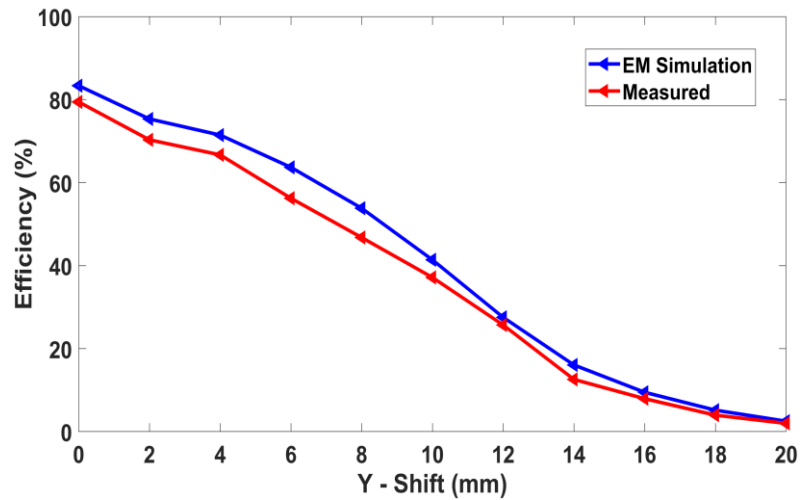


Fig. 4.17. Misalignment of prototype. (a) Lateral shift in X direction. (b) Lateral shift in Y direction.



(a)



(b)

Fig. 4.18. Effect of lateral misalignment. (a) Shift in X-direction. (b) Shift in Y-direction.

For Fig. 4.18(a), the result is recorded when the RX is shifted along X-axis between 0 mm and 20 mm with a step-size of 2 mm. At 0 mm, the RX is in the line of sight and is perfectly

aligned with the TX whereas 20 mm is the maximum misalignment. It is apparent in Fig. 4.18(a) that the efficiency is maximum at no misalignment (i.e., misalignment of 0 mm) and degrades with the increase in misalignment until it approaches zero at 20 mm in this case. The results of misalignment along Y-axis, shown in Fig. 4.18(b), conveys that the maximum and minimum efficiency follow similar pattern.

It can be deduced from Figs. 4.18(a) and 4.18(b) that the degradation of efficiency with shift in X- and Y-directions are almost similar and consistent. These results identify that the proposed design achieves efficiency of above 50% even with lateral misalignment of 6-8 mm and this is far superior than an earlier reported H-shaped DGS based WPT system with identical lateral shift in X-direction [23]. Furthermore, the proposed system exhibits similar efficiency when compared to H-shaped DGS based WPT system when misalignment occurs in Y-direction. The consistency in the achieved results for the proposed WPT system can be attributed to the excellent matching achieved by incorporating the coupled-line based impedance transformation technique.

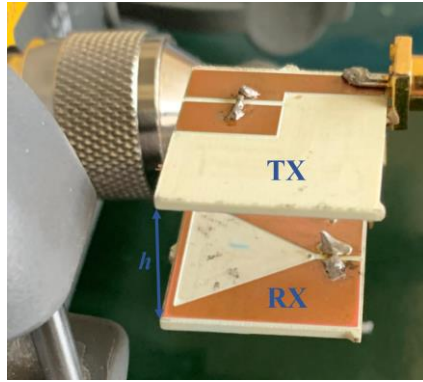


Fig. 4.19. 3-D representation of angular rotation of RX position at 90°.

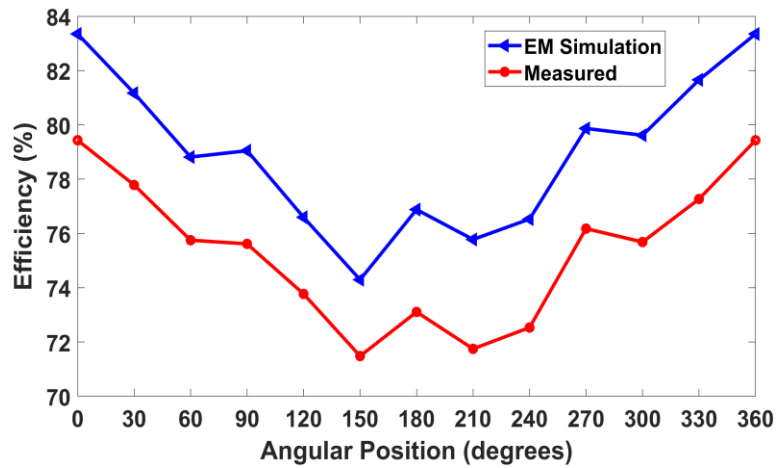


Fig. 4.20. Effect of angular misalignment of RX on efficiency

Subsequently, the angular misalignment for various angular positions between 0° and 360° of TX and RX are studied. One such case of angular orientation of RX position at 90° is shown in Fig. 4.19. It can be seen in Fig. 4.20 that the EM simulated efficiency and the measured efficiency follows similar pattern with some anomaly due to soldering loss, connector loss, etc. during the prototyping process. It also conveys that the efficiency is not greatly impacted by the angular misalignment of TX and RX. It once again significantly improves the outcome reported for the H-shaped DGS WPT system [23].

In brief, it is safe to convey that the proposed system is immune to the variation in angular position of the RX considering that there is no major drop in system efficiency. This is due to the fact that the maximum area of coupled-lines of the RX are in the line of sight with the TX unlike the two cases of lateral shifts. Notably, it can be seen that the measured efficiency for the 90° orientation is close to 76% and this is excellent for any practical situation requiring TX and RX to be misaligned by 90° .

4.4. Conclusion

In this chapter, a new technique using symmetric coupled lines for impedance matching and miniaturization of WPT systems has been proposed. First, the theoretical foundations are developed which are subsequently validated through EM simulation, circuit simulation and measurement results. The proposed technique is frequency scalable and exhibits reduced complexity in fabrication through the introduction of surface-mounted capacitor between the coupled lines on a top plane to obtain perfect matching. The incorporation of the defect in the ground plane solves two different purposes namely miniaturization of the overall WPT system and the achievement of the desired resonant frequency. The reported DGS based WPT system using coupled lines takes an area of $20 \times 20 \text{ mm}^2$ and achieves power transfer at 17 mm with 80% system efficiency. The designed system possesses superior FoM as compared to recent works and thus significantly improves the existing state of the art.

Chapter 5

Wireless Information and Power Transfer (WIPT) System: Triple Band Operation

This chapter is based on the following research paper:

- S. Verma, D. Rano, S. Malhotra and M. S. Hashmi, “Measurements and Characterization of a Newly Developed Novel Miniature WIPT System,” *IEEE Transactions on Instrumentation and Measurement*, vol. 70, no. 2004211, pp. 1-11, 2021, doi: 10.1109/TIM.2021.3075537.

This chapter focuses on the design of a new defected ground structure (DGS) based miniature tri-band bandstop filter, which significantly advances the state-of-the-art, and its subsequent use in the design of tri-band wireless information and power transfer (WIPT) system. Then an extensive measurement and characterization of the developed WIPT system is carried out to identify the performance metrics. Finally, measurements and evaluation scheme is developed to find out the best and worst case scenarios for the WIPT system. After thorough analysis, it is identified that the proposed tri-band WIPT system is capable of facilitating concurrent power transfer at one frequency and uplink and downlink for the information transfer at the other two frequencies. At this point, it should be noted that the small sized WIPT systems can fulfill the needs of low power applications, such as MBAN and ISM, considering that most of such applications require highly efficient but lightweight WIPT systems.

The next section elaborates the modeling and design of rectangular shaped tri-band DGS resonator and realization of its filter characteristics. The section 5.2 presents the realization of the proposed tri-band WIPT system and investigates the adopted design strategy using equivalent circuit model. It is pertinent to mention that, as a proof of concept, the design frequencies of 300 MHz, 433 MHz, and 700 MHz are selected. The choice of these frequencies are regulated with the fact that many recent WPT designs have been reported at these frequencies and the selection of these frequencies provides an opportunity for ready comparison [9][24][64]. Section 5.3 provides a depiction of the power and data transfer measurement and characterization schemes developed for the proposed WIPT system. Subsequently, power transfer capability of the

proposed design at 300 MHz is experimentally demonstrated using a Light Emitting Diode (LED) based setup and the data transfer Up-link (433 MHz) and down-link (700 MHz) communication are demonstrated using a Universal Software Radio Peripheral (USRP) based setup in Section 5.4. The Section 5.5 provides an elaborated discussion on the challenges faced during the measurement and characterization of the proposed WIPT system and the subsequent solutions to overcome them.

5.1. Design and Analysis of Resonator

This section focuses on the development of the novel design of a tri-band resonator, mimicking band-stop feature, required for the eventual design of the proposed WIPT system. For this purpose, rectangular DGS resonator for achieving tri-band band-stop characteristics at frequencies of 300 MHz, 433 MHz, and 700 MHz is conceptualized and then realized on Rogers 4350B substrate having $\epsilon_r = 3.6$, substrate thickness $h_{sub} = 1.524$ mm, and metal thickness $t = 35$ μm , as shown in Fig. 5.1. The top plane of the rectangular DGS resonator consists of 50 Ω microstrip feed line (MSL), of length L_F and width W_F , and is connected to lumped ports P_1 and P_2 , as depicted in Fig. 5.1(a). The ground plane of the DGS resonator, in Fig. 5.1(b), consists of three rectangular defects of length L_1 , L_2 , and L_3 and width W_1 , W_2 , and W_3 . These defects are connected to node slot of length L_n and width W_n through their respective excitation slot of length L_s and width W_s . Finally, the node slot is connected to common excitation slot of length L_b and width W_b . Moreover, the rectangular shaped defects are arranged in the form of tree structure to obtain overall miniaturization in the circuit size. Further miniaturization in the circuit size and frequency tuning at the three operating frequencies is achieved using external capacitors mounted on the excitation slot of each rectangular defect. Finally, the proposed DGS resonator is optimized, for achieving the desired resonance as identified by EM simulation results in Fig. 5.2(a), using CST Microwave Studio and the relevant design parameters are provided in Table 5.1. In addition, the effectiveness of the resonance of the proposed resonator can also be assessed by looking through the surface current distribution on the DGS surface at the three resonance frequencies as shown in Fig. 5.2(b). It can be inferred that the current flowing around each rectangular defect results in perturbation of the field propagating through the microstrip feed line at the three chosen frequencies. Moreover, the current flowing around the rectangular defect at 300 MHz is greater in comparison to 433 MHz and 700 MHz. This occurs due to larger area of the rectangular defect operating at 300 MHz.

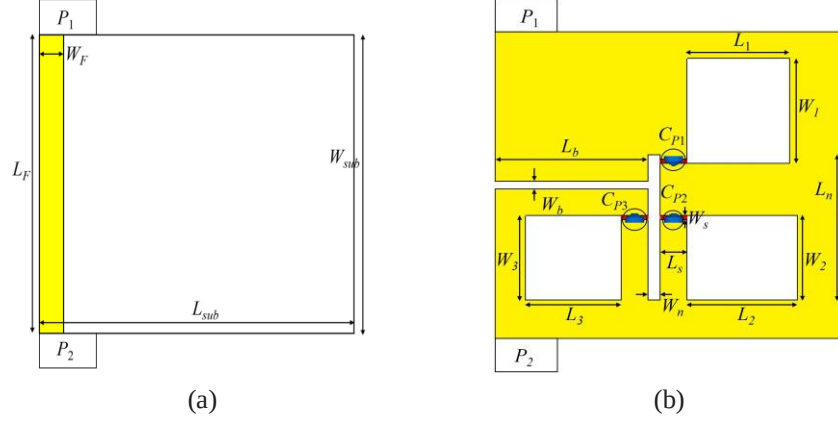


Fig. 5.1. Geometry of the proposed rectangular DGS resonator. (a) Top plane. (b) Bottom plane.

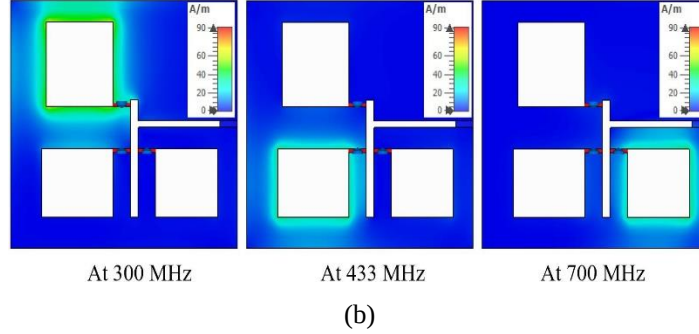
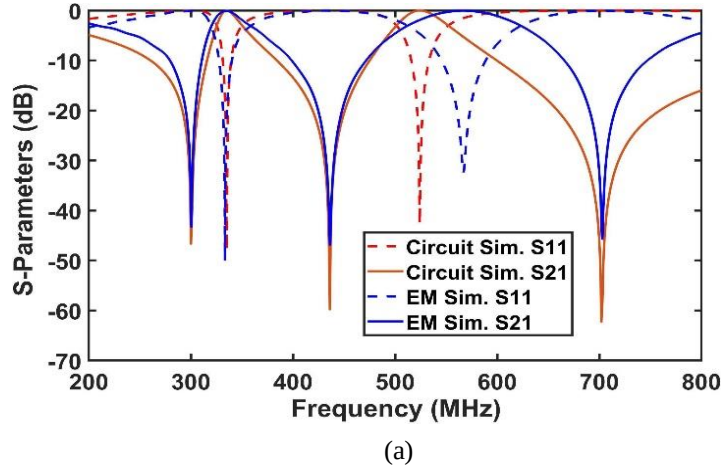


Fig. 5.2. Band-stop performance of rectangular DGS resonator. (a) S-parameter plot of EM and circuit simulations. (b) EM simulated surface current distribution on the DGS surface at the three resonance frequencies.

Table 5.1: Design parameters of tri-band DGS resonator

L_F (mm)	W_F (mm)	L_{sub} (mm)	L_1 (mm)	W_1 (mm)
38	3.3	43	12.6	13
L_2 (mm)	L_3 (mm)	$W_2 = W_3$ (mm)	L_s (mm)	W_s (mm)
13.5	11.7	10.5	3.3	0.5
L_n (mm)	W_n (mm)	L_b (mm)	W_b (mm)	C_{P1} (pF)
18	1.5	18.75	1	10.8

C_{P2} (pF)	C_{P3} (pF)	L_{P1} (nH)	L_{P2} (nH)	L_{P3} (nH)
5	1.6	26	26.67	32.09

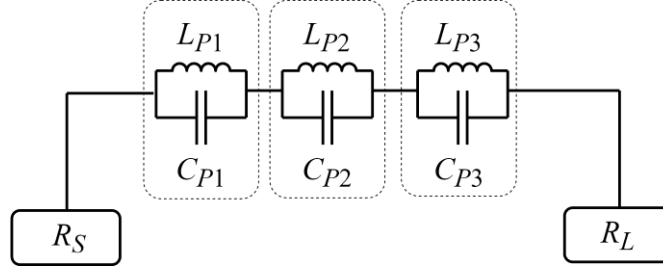


Fig. 5.3. Equivalent LC circuit model.

The EM simulation results can be assessed by developing an LC equivalent circuit, depicted in Fig. 5.3, of the proposed DGS resonator. It will be shown that the development of this step eventually helps in the expedited design and optimization of the proposed WIPT system. The equivalent circuit is designed to resonate at three frequencies, as a result, at $f_{01} = 300$ MHz, C_{P1} shorts L_{P1} , while at $f_{02} = 433$ MHz, C_{P2} shorts L_{P2} , and at $f_{03} = 700$ MHz, C_{P3} shorts L_{P3} . In the equivalent LC circuit, capacitances are the external capacitors (C_{P1} , C_{P2} , and C_{P3}) mounted in the excitation gaps. The inductance (L_{P1} , L_{P2} , and L_{P3}) accounts for the periphery of each rectangular defect in the ground plane calculated using the expression (5.1) - (5.3) [28]. It is imperative to note that the inductance and capacitance are often in the range of nano henry (nH) and pico farad (pF). The inductance of each rectangular defect along with their respective capacitance loaded in gaps combine to give band-stop frequency response. The R_S and R_L represent the source and load resistances. These are taken as 50Ω so that the DGS resonator can be conveniently measured in 50Ω impedance environment.

$$L_{P1} = \frac{1}{4\pi^2 f_{01}^2 C_{P1}} \quad (5.1)$$

$$L_{P2} = \frac{1}{4\pi^2 f_{02}^2 C_{P2}} \quad (5.2)$$

$$L_{P3} = \frac{1}{4\pi^2 f_{03}^2 C_{P3}} \quad (5.3)$$

The values of inductance extracted from (5.1) - (5.3) and the values of external loading capacitors in the gaps are substituted in the equivalent circuit model and the simulation is carried out in ADS (Advanced Design System) for evaluating the equivalent circuit. The results obtained from both circuit simulation and the EM simulation compare favorably, as can be seen in Fig. 5.2(a), and thus the reliability of the developed equivalent circuit model is validated.

5.2. Tri-Band WIPT System Design

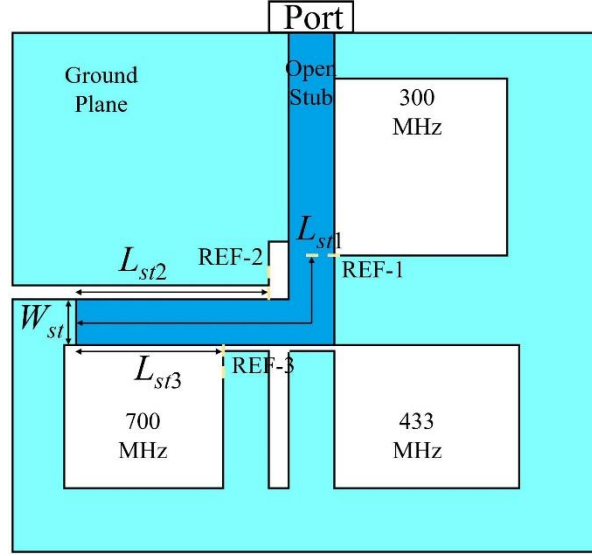


Fig. 5.4. Modified tri-band DGS resonator for the proposed WIPT system, top and bottom plane.

In this section, the necessary modifications, as depicted in Fig. 5.4, required in the proposed resonators for the realization of the WIPT system are presented first. Subsequently, a successful WIPT system is realized and evaluated using a newly developed equivalent circuit model. The top plane of the modified DGS resonator consists of a stub which results from the open-circuited microstrip feed line. The total length of stub $L_{st1} = 22.1$ mm and width W_{st} equals the width of 50Ω microstrip feed line. The total length of stub represents the stub length for frequency f_{01} from the first reference point (REF-1). For f_{02} and f_{03} , an overlapping stub length from two different reference points i.e. REF-2, and REF-3 are taken. The stub length for f_{02} and f_{03} are denoted by $L_{st2} = 14.1$ mm and $L_{st3} = 10.8$ mm. Here, the physical stub length shown is slightly different from the effective stub length at the three resonant frequencies. This is due to the fact that the node slot width " W_n " below the stub line and low capacitive coupling caused by the uniform ground plane area with respect to the stub line is surrounded by defects. The effective stub length for the three frequencies are given by $L_{effst1} = 19.8$ mm, $L_{effst2} = 12.2$ mm, and $L_{effst3} = 2.7$ mm. The values of capacitance (C_{st1} , C_{st2} , and C_{st3}) across the effective stub length (L_{effst1} , L_{effst2} , and L_{effst3}) are deduced in the next section where equivalent circuit model is shown. Furthermore, the top plane of the modified tri-band DGS resonator, in Fig. 5.4, depicts the change in the position of the open-circuited microstrip line in comparison to the microstrip feed line on the top plane of the original resonator in Fig. 5.1(a). The ground plane of the modified DGS resonator, in Fig. 5.4, is exactly similar to the ground plane of the original resonator, in Fig. 5.1(b), except for the external capacitor values. The modified resonator is EM simulated using CST microwave studio and optimization is carried out to achieve the best

possible results for the chosen resonance frequencies. The optimized design parameters of the modified tri-band DGS resonator are given in Table 5.2.

Table 5.2: Optimized design parameters of the modified tri-band DGS Resonator

L_{st1} (mm)	L_{st2} (mm)	L_{st3} (mm)	W_{st} (mm)	L_{effst1} (mm)
22.1	14.1	10.8	3.3	19.8
L_{effst2} (mm)	L_{effst3} (mm)	C_{st1} (pF)	C_{st2} (pF)	C_{st3} (pF)
12.2	2.7	2.58	1.58	0.34
C_{P1} (pF)	C_{P2} (pF)	C_{P3} (pF)	K_1	K_2
8.9	3.8	1.3	0.048	0.058
K_3	h (mm)	η_1 (%)	η_2 (%)	η_3 (%)
0.013	23	76.9	75.5	75.3

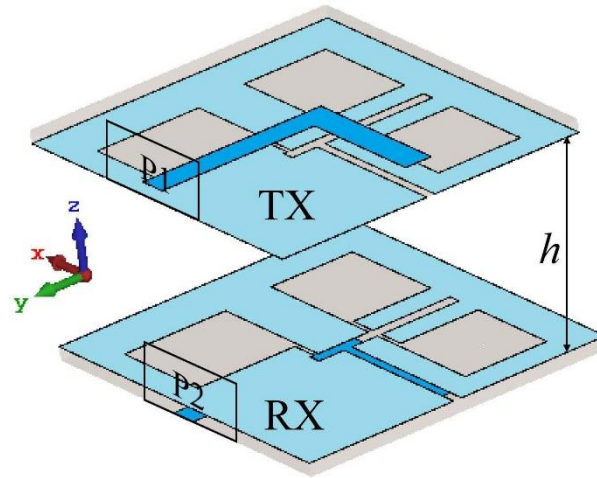


Fig. 5.5. Back-to-back configuration of the resonators for the development of WIPT system.

The basic architecture of the proposed WIPT system is realized by placing two modified tri-band DGS resonators in a back-to-back configuration as shown in Fig. 5.5. Out of the two resonators, one acts as a transmitter (TX) and the other as a receiver (RX). Apparently, the TX and RX resonators are symmetric and have lumped ports P1 and P2 connected at one end of the microstrip feed line. The other end of the microstrip feed line of both the resonators are left open. Moreover, the separation distance (h) between the two resonators transforms the band stop characteristics of the DGS resonators into band pass characteristics. This facilitates power and information transfer from TX to RX at the three operating frequencies. It was identified during EM simulation of the WIPT system that $h = 23\text{mm}$ between the TX and RX depicts critical coupling distance. The critical coupling distance provides the best power and information transfer efficiency in this case.

5.2.1. Equivalent Circuit Model

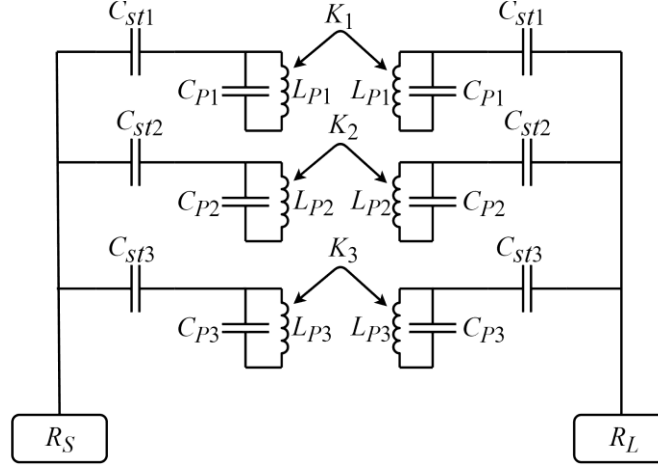


Fig. 5.6. Equivalent circuit of the proposed WIPT system

The developed equivalent LC circuit augmented with effective stub capacitances C_{st} is shown in Fig. 5.6. Apparently, the circuit has different inductance and capacitance values to achieve the desired three frequencies of operation. The TX and RX resonators are similar and as a result two similar equivalent LC circuits make the complete circuit for the proposed WIPT system. The mutual coupling L_m between the TX and RX is expressed in terms of impedance $imag(Z_{21})$ given in (5.4) [65]. The $imag(Z_{21})$ is extracted by performing EM simulation. Prior to EM simulation, the microstrip feed line from the top plane is eliminated and lumped port in the excitation slot of each defect is added successively. This is done for both TX and RX DGS resonator placed in a back-to-back configuration. Moreover, the coupling coefficient K between the TX and RX is related to L_m by (5.5) [27].

The expressions (5.4) and (5.5) can be used interchangeably to calculate and optimize the other design parameters of the WIPT system depending on the final results achieved in EM simulation. For example, in EM simulation, negative value of $imag(Z_{21})$ and higher value of $imag(Z_{21})$ will make the final value of L_m impractical and less significant. In such a scenario, (5.5) can be useful to extract both the values of coupling coefficient and mutual coupling using two peaks f_e and f_m , depicted in Fig. 5.7, resulting from frequency splitting and resonator self inductance L_P . Moreover, the three stub capacitances C_{st1} , C_{st2} , and C_{st3} , obtained from (5.6) [28], provide impedance matching at the three operating frequencies. These are connected between the LC circuit and the source resistance (R_S) and load resistance (R_L).

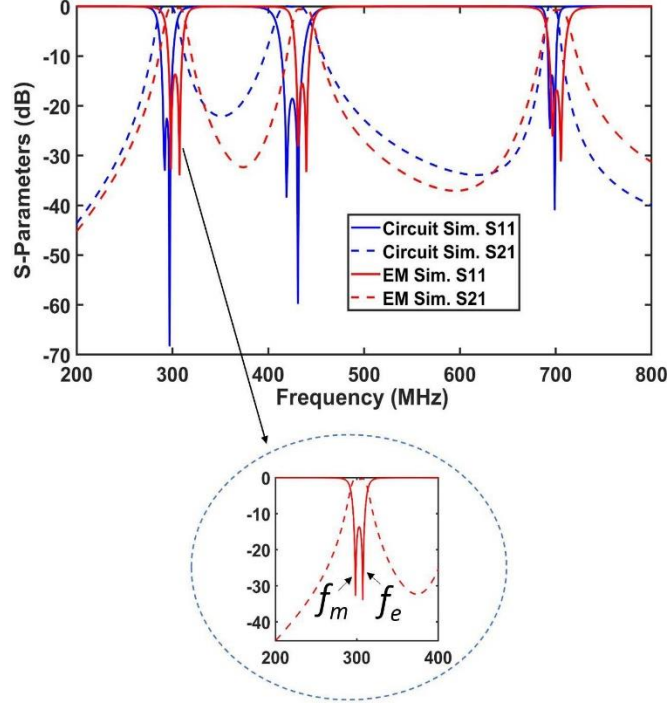


Fig. 5.7. Comparison of EM simulation and circuit simulation S-parameter plot of the proposed WIPT system.

The expressions (5.4), (5.5), and (5.6) depict generic closed-form design equations. These equations are utilized to calculate the design parameters of the proposed WIPT system at the three different frequency bands of operation. Subsequently, a comparison between EM simulation and circuit simulation S-parameter plot is highlighted in Fig. 5.7. It can be seen that the circuit simulation and EM simulation S-parameter plots are in very good agreement with each other. At this point, it is important to mention that this design scheme is scalable and can be readily adopted for other frequencies such as ISM frequency of 868 MHz. In such a scenario, appropriate care will need to be taken with regards to the capacitance and the inductance of each rectangular defect.

$$L_m = \frac{\text{imag}(Z_{21})}{2\pi f_0} \quad (5.4)$$

$$K = \frac{L_m}{L_p} = \frac{f_e^2 - f_m^2}{f_e^2 + f_m^2} \quad (5.5)$$

$$C_{st} = \frac{\tan(\beta L_{effst})}{2\pi f_0 Z_0} \quad (5.6)$$

5.3. Scheme for Power and Information Transfer

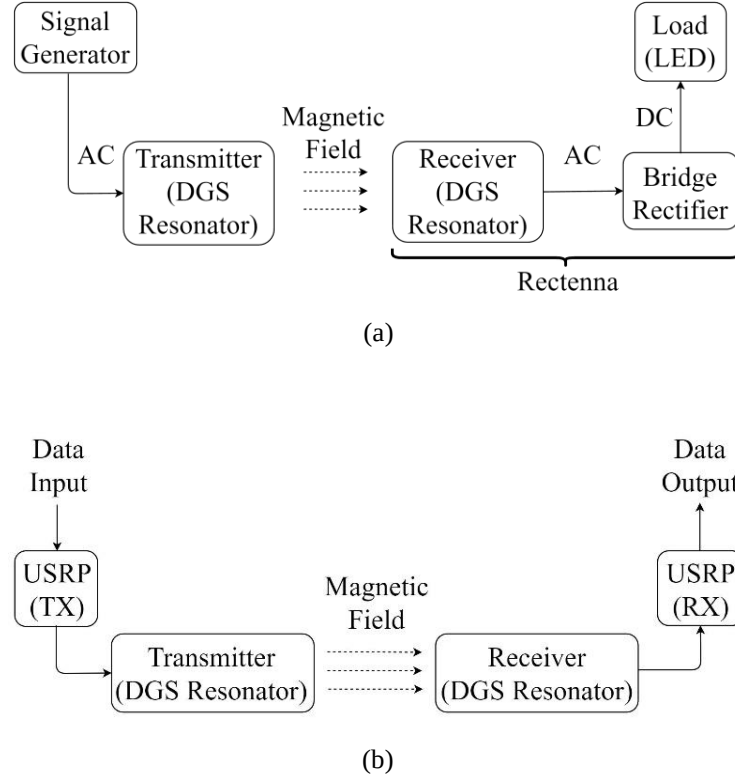


Fig. 5.8. Block diagram showing power and data transfer. (a) Power transfer to an LED. (b) Data transfer using USRP.

In this section, the measurement and characterization setup utilized to obtain power and data transfer using the proposed WIPT system is developed. The Figs. 5.8(a)-(b) provide a block diagram depiction of the power and information transfer schemes of the proposed WIPT system. The practical realization of these schemes are elaborated in the measurement section. Figure 5.8(a) highlights the process of power transfer to operate an LED. It begins with an RF signal generator, which provides the RF signal at the first design frequency i.e. 300 MHz in this case. The RF signal is transmitted from the TX resonator to the RX resonator through magnetic resonance coupling. The received RF signal then passes through full wave bridge rectifier, developed using Schottky diode, and get converted to DC. The rectified signal, i.e. the DC output, powers the load consisting of a Light Emitting Diode (LED).

The data or information transfer scheme is depicted in Fig. 5.8(b). It is developed using two USRPs, one working as TX-USRP while the other working as RX-USRP, and two resonators working as TX and RX respectively. Once again, the principle of magnetic resonance coupling between the TX resonator and RX resonator enables the information transfer between the transmitter and receiver. In this case, the other two design frequencies of 433 MHz and 700

MHz has been chosen for information transfer purposes. The data input to the TX resonator is obtained from the TX-USRP and the output from the RX resonator is sensed using RX-USRP. Here, for demonstration purposes, a string utilizing quadrature amplitude modulation (QAM) scheme generated by TX-USRP has been considered as the input data. Subsequently, the data received at the output of RX-USRP at two different design frequencies are verified using front-end National Instruments Lab-View tool.

5.4. Measurement Results

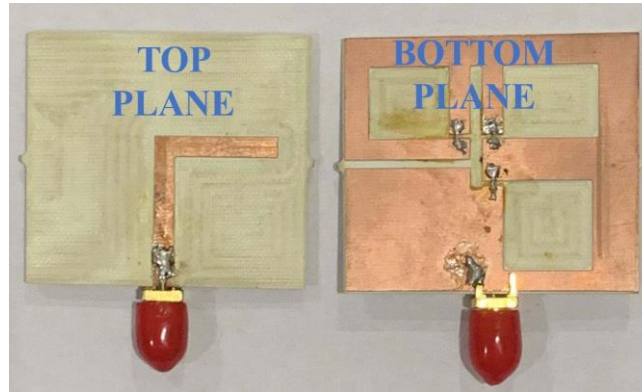


Fig. 5.9. Fabricated prototype of the proposed DGS resonator.

This section includes the thorough measurement and characterization of the developed WIPT system. The fabricated prototype shown in Fig. 5.9 depicts the top and the bottom planes of the proposed DGS resonator to be employed in the WIPT system. The eventual WIPT system is realized by placing two such resonators in a back-to-back configuration where one works as the TX while the other as RX. Based on simulation outcome for optimum performance described in the earlier section, the TX and RX are kept apart at a distance of $h = 23$ mm by utilizing plastic foam during the measurement as shown in Fig. 5.10.



Fig. 5.10. Measurement setup.

The plastic foam is inserted partially so as to align the TX and RX for optimal efficiency. Care is taken that no portion of the defected ground is concealed by the foam.

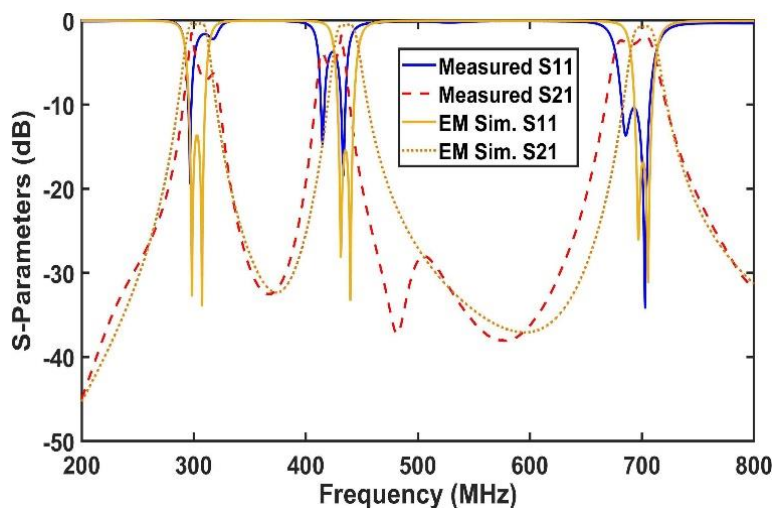


Fig. 5.11. Measured S-parameter plot.

As a first measurement step, the prototyped WIPT system is characterized for S-parameters using a measurement setup developed around Keysight Vector Network Analyzer (VNA) operating over the frequency range of 200-800 MHz. The measured S-parameter plot shown in Fig. 5.11 depicts resonance characteristics at 298 MHz with $|S_{21}| = -1.54$ dB, while at 432 MHz with $|S_{21}| = -1.8$ dB. Both the frequencies at 298 MHz and 432 MHz show $|S_{11}|$ with approximate value of -20 dB. In addition, the third frequency at 702 MHz represent $|S_{21}| = -1.87$ dB and $|S_{11}| = -34$ dB. Moreover, it can be seen that there is a slight shift in the resonance frequency of the measured results ($f_{01} = 298$ MHz, $f_{02} = 432$ MHz, and $f_{03} = 702$ MHz) from the originally chosen design frequencies of 300 MHz, 433 MHz, and 700 MHz. The shift in the resonant frequencies can be attributed to the tolerances in the values of the external capacitor C_P (part numbers: CBR04C899B5GAC, CBR04C389B5GAC, and CBR04C139B5GAC) and the parasitic effects inherent in the designed system. In addition, other factors which also introduce this anomaly include the in-house fabrication which account for losses such as soldering loss, connector loss, etc. Here, the effect of tolerances and external factors on the design frequencies can be reduced to a greater degree by introducing a shift in the design frequency in the EM simulation. For instance, if the design frequency is 300 MHz, the value of the capacitor is selected so that the system resonates at a slightly greater frequency i.e. 302 MHz. As a result, the effect of tolerance and external factors at 302 MHz will bring a shift the resonance frequency closer to design frequency i.e. 300 MHz. A similar technique is followed for other two design frequencies i.e. 433 MHz and 700 MHz. Furthermore, the measured S-parameters can be utilized to assess the power transfer efficiency η_{WPT} , expressed by (4.5), of the WIPT system. Here, $|S_{21}|$

and $|S_{11}|$ are the respective magnitudes of the transmission and reflection coefficients at the three design frequencies respectively.



Fig. 5.12. Spectrum analyzer-based measurement setup for efficiency evaluation.

As a second measurement step, the efficiency of the fabricated WIPT system at the three resonance frequencies is evaluated using Anritsu MS2720T spectrum master as shown in Fig. 5.12. To measure the efficiency of the WIPT system, an input power of -10 dBm is fed to the TX resonator and the spectrum of the output power is recorded individually at the RX resonator for the three frequencies at separation distance of 23 mm. The output spectrum at the three design frequencies are shown in Figs. 5.13-5.15. Apparently, in Fig. 5.13, the received output power at 298 MHz is -11.55 dBm. Similarly, in Fig 5.14 and Fig 5.15, the received output powers are -11.8 dBm at 432 MHz and -11.89 dBm at 702 MHz respectively. Then the expression in (5.7) is used to compute the efficiency of the WPT system (η_{WPT2}) using the spectrum analyzer.

$$\eta_{WPT2} = \frac{P_{out}}{P_{in}} \times 100\% \quad (5.7)$$

The measured efficiency obtained from VNA setup (η_{WPT1}) compare favorably with the measured efficiency obtained using the spectrum analyzer setup (η_{WPT2}) at the three resonance frequencies. In both cases, the measured efficiency at 298 MHz is 70.1 %, 66 % at 432 MHz, and 65 % at 702 MHz.

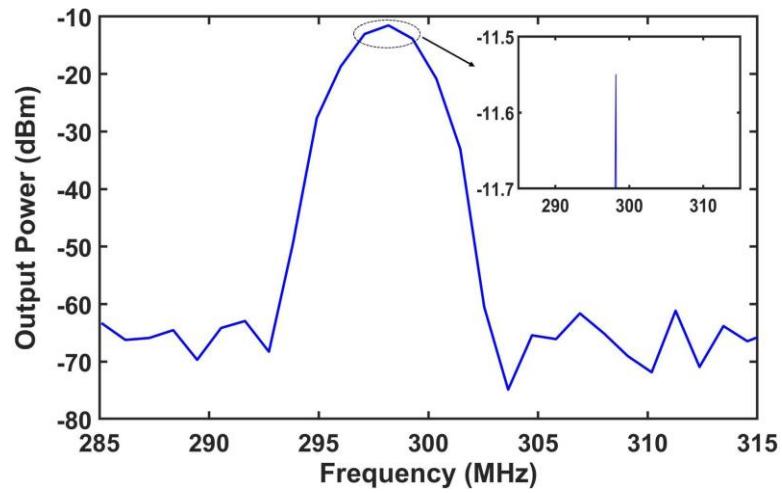


Fig. 5.13. Output power at 298 MHz.

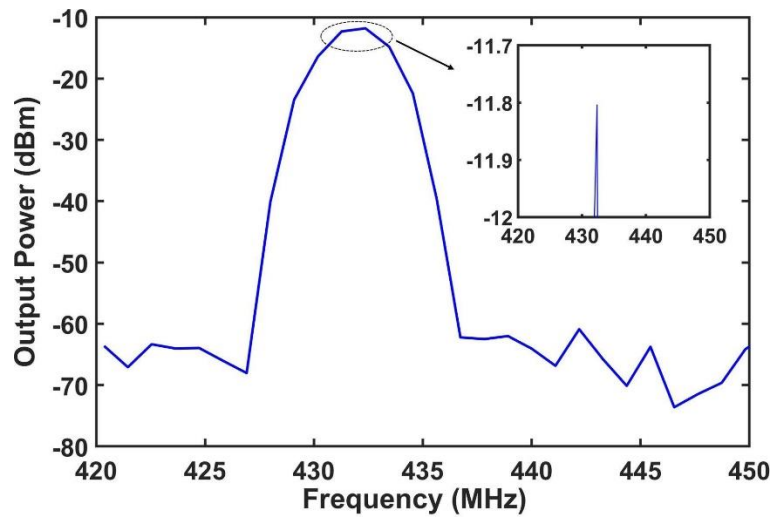


Fig. 5.14. Output power at 432 MHz.

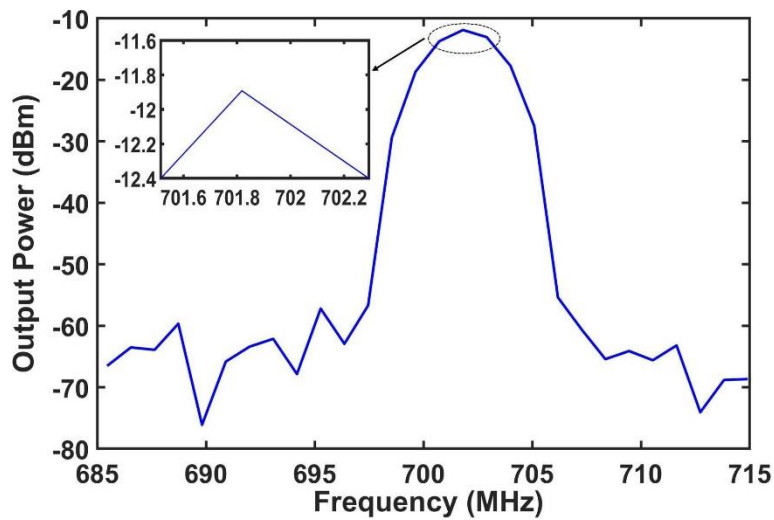


Fig. 5.15. Output power at 702 MHz.

5.4.1. Powering an LED

The frequency, f_{01} , of 300MHz is utilized for wirelessly transferring the power. Keeping this in perspective, an experiment is carried out to demonstrate the power transfer feature of the proposed WIPT system at 300 MHz. Here, an LED is illuminated when power is transferred between TX and RX resonators at a fixed separation distance h . In Fig. 5.16, the measurement setup showing power transfer to an LED consists of Keysight analog signal generator N5181A, TX and RX resonators, full wave bridge rectifier designed using Schottky diode (part number: MMSD701T1G), and red color LED of 5 mm diameter. In this experiment, first an input RF signal at 298 MHz with amplitude of 8 dBm is fed to the TX resonator. Then the signal received at the RX resonator is passed through Schottky diode to achieve RF to DC conversion. The DC output is then used to power the LED. Moreover, it is analyzed that as the separation distance between the TX and RX resonators is increased to a distance greater than 23 mm, the brightness of LED starts decreasing which depicts the fall in the power transfer efficiency. An important thing to note here is that the amplitude of the input RF signal is tuned on the signal generator in order to reach the threshold point of the LED. Moreover, some loss of power takes place during the wireless power transmission due to separation distance and the rectification stage.

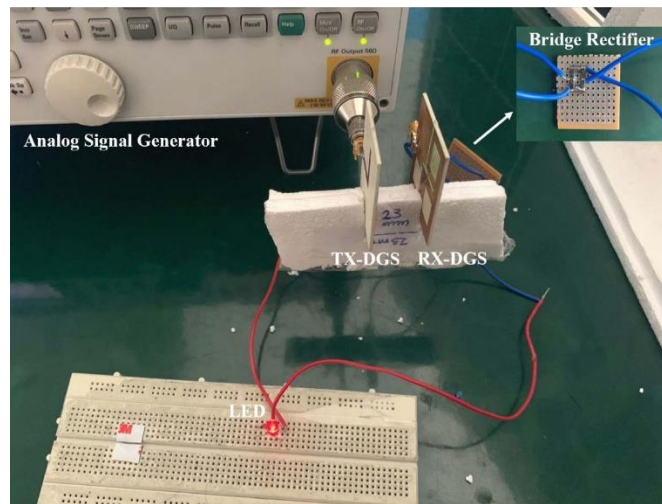


Fig. 5.16. Measurement setup showing power transfer to a LED.

5.4.2. Information Transfer

In Fig. 5.17, the measurement setup to verify data transfer using USRP is shown where higher frequency f_{03} , 702 MHz, is used for the up-link communication and f_{02} , 432 MHz, is utilized for the down-link communication. Here, the measurement setup involves LabView tool, two NI USRP's-2922, and the WIPT system consisting of TX and RX resonators in back-to-back configuration. In the experiment, a string of data "Hello world" is transmitted from the TX-

USRP to RX-USRP at 702 MHz which depicts up-link communication. On the other hand, the same data received at the RX-USRP is re-transmitted towards TX-USRP at 432 MHz which depicts down-link communication. The data transferred in up-link and down-link communication is performed using NI LabView user interface (UI). Furthermore, in Fig. 5.18, the data transfer is validated using performance metrics such as bit error rate (BER) and constellation diagram. In Fig. 5.18(a), the bit error rate plot with respect to separation distance is shown. It can be seen that as the separation distance increases, the bit error rate increases. The BER is equivalent to zero at critical coupling distance i.e. 23mm and approaches towards 0.06 at 33mm distance. Moreover, Figs. 5.18(b)-(c) depicts the constellation plot computed at 23mm and 33mm separation distance. Here, Fig. 5.18(b) represents a circular quadrature amplitude modulation (QAM) constellation plot at 23mm. On the other hand, in Fig. 5.18(c), the constellation plot of the circular QAM appears to be distorted at 33mm. This occurs as a result of low value of BER at 23 mm, whereas, at 33mm, the value of BER is high. Also, the comparison of Fig. 5.18(b) and Fig. 5.18(c) proves the best case scenario at 23 mm and a worst case scenario at 33mm.

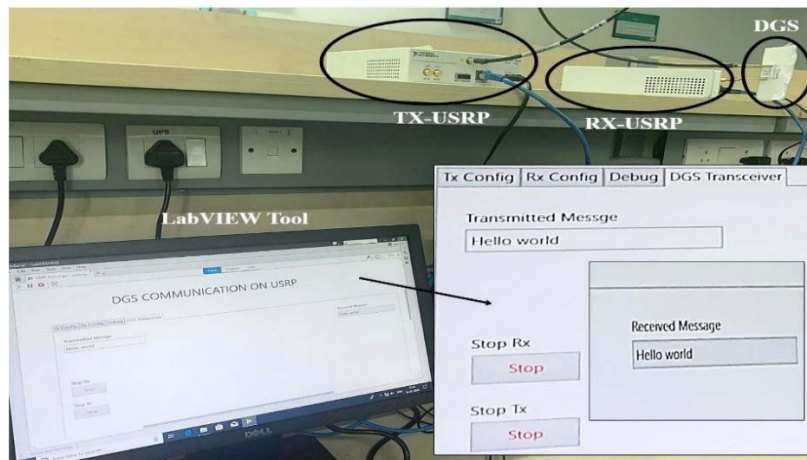
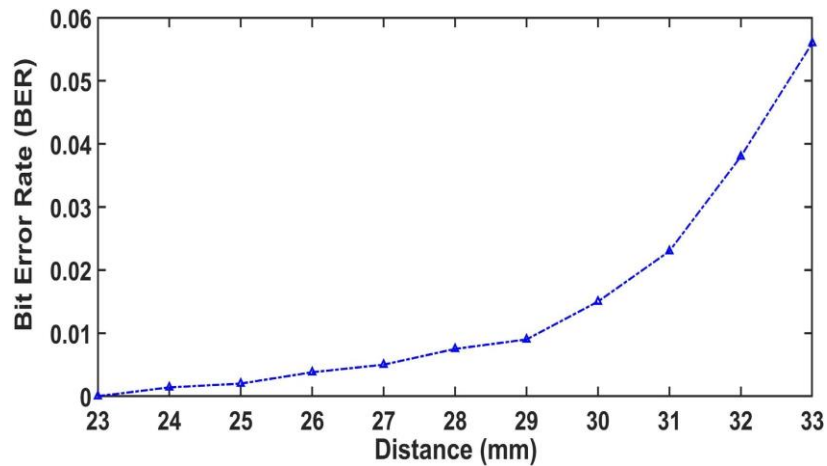


Fig. 5.17. Measurement setup showing data transfer using USRP.



(a)

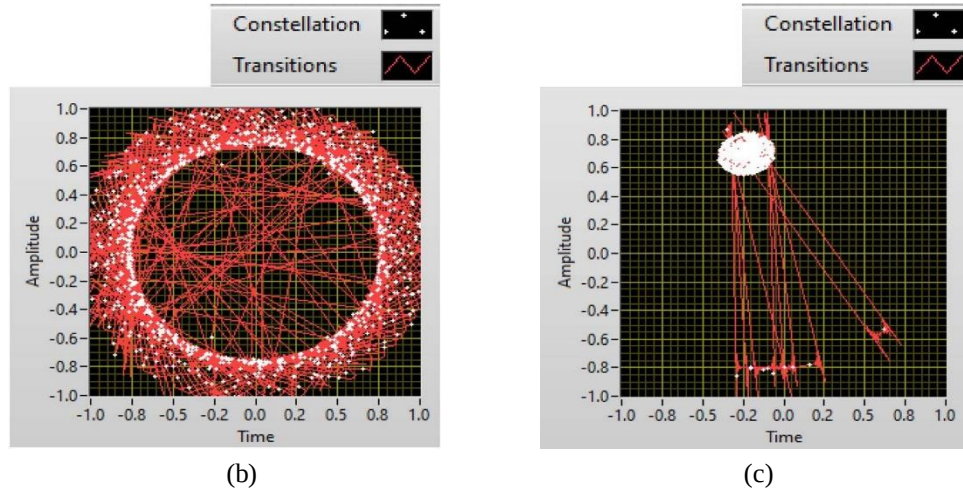


Fig. 5.18. Performance metrics of data transfer. (a) BER plot. (b) Constellation plot computed at 23 mm (best case). (c) Constellation plot computed at 33 mm (worst case).

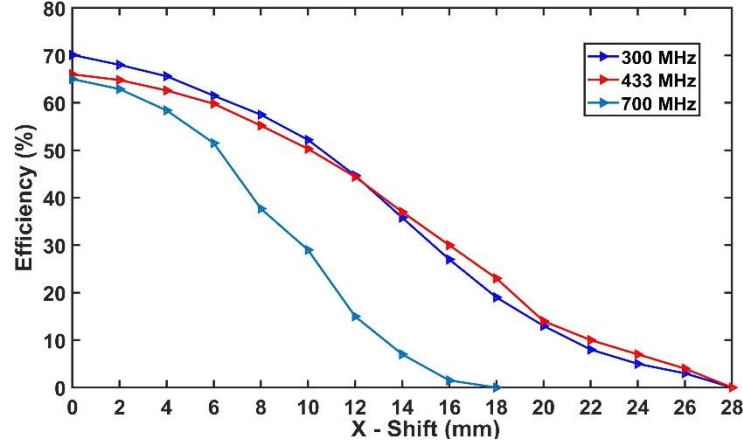
Finally, the proposed WIPT system can be assessed based on a metric, called Figure of Merit (FoM), using expression (4.6). This metric apparently relates resonant frequency, size, power/information transmission distance, and system efficiency (η_{WIPT}). In ideal scenario, the FoM should be as high as possible. However, in practical situations this is regulated by many factors such as size, operating frequency, and the transmission distance. For the proposed system, operating at several hundred MHz range at multiple frequencies, the computed FoMs are 0.4, 0.38, and 0.37 at 300 MHz, 433 MHz, and 700 MHz respectively. This is a very promising outcome for the first time reported tri-band WIPT system capable of power transmission and full-duplex information transmission. The Table 5.3 compares the performance metrics of the proposed system with other systems operating around the same design frequencies. It is apparent that the proposed system significantly enhances the state-of-the-art designs in terms of efficiency, FoM, compactness, and transmission range. Furthermore, the proposed system is capable of information transfer as well and this is a very exciting outcome towards achieving wirelessly powering and information transfer in the futuristic applications.

Table 5.3: Performance comparison of proposed WIPT system with the state-of-the-art designs

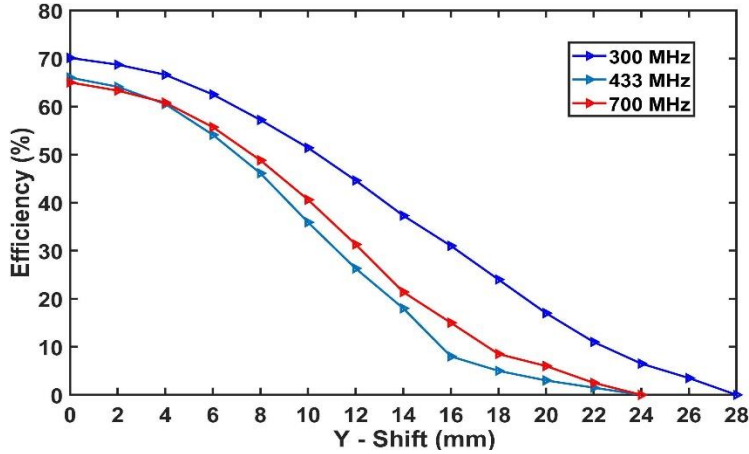
System	f_0 (MHz)	Size (mm ²)	h (mm)	η (%)	FoM
[9]	1 and 20	TX 40x40 RX 20x20	7	30 and 30	0.14 and 0.19
[24]	300 and 700	30x15	16	71 and 72	0.76 and 0.77
[64]	0.2 and 6.78	125x89	25	78 and 70.6	0.17 and 0.19
This work	300, 433 and 700	43x38	23	70.1, 66 and 65	0.4, 0.38 and 0.37

5.5. Measurement Challenges

In this sub-section, two different types of challenges experienced during the measurement of the proposed WIPT system are shown. The first challenge talks about the misalignment of the RX resonator in lateral and angular shift types. On the other hand, the second challenge aims at power transfer at four different frequency bands concurrently.



(a)



(b)

Fig. 5.19. Effect of lateral misalignment. (a) Shift in the X -direction. (b) Shift in the Y -direction.

The WIPT system can be greatly impacted by lateral and angular misalignments between the TX and RX. In order to understand this impact, the TX and RX are laterally shifted along the horizontal (X -axis) direction and then along the vertical (Y -axis) direction. The effect of lateral shift on efficiency is studied and the obtained results are shown in Figs. 5.19(a) and (b). For Fig. 5.19(a), the result is recorded when RX is shifted between 0 mm and 28mm with a step size of 2 mm. At 0 mm, the RX is in line of sight and is perfectly aligned with the TX, whereas 28 mm represents 65 % of misalignment in the positive X -direction. It is apparent in Fig. 5.19(a)

that the efficiency at the three frequencies is maximum at no misalignment (i.e. misalignment of 0 mm) and degrades with the increase in misalignment. For 300 MHz and 433 MHz, the efficiency degrades to zero percent at 28 mm, whereas, for 700 MHz, similar scenario occurs at 18 mm. The results of misalignment along the Y-axis, shown in Fig. 5.19(b), conveys that maximum and minimum efficiencies, at 300 MHz, follows the similar pattern. On the other hand, at 433 MHz and 700 MHz, the maximum efficiencies remain similar while the minimum efficiencies, at 0 % converge at 24 mm. It is therefore imperative to note that a perfect alignment between the RX and TX must be achieved for obtaining the optimal efficiency. Subsequently, in Fig. 5.20, the angular misalignment for various angular positions between 0° and 360° of TX and RX are studied. The efficiency is maximum at 0° and 360° angular shift. For 300 MHz and 433 MHz, the efficiency recorded is lowest at 180° angular shift of the receiver position. In addition, it can be seen that the efficiency curve below 180° follows a decreasing pattern whereas, above 180°, the plot depicts an increasing pattern. Furthermore, at 700 MHz, the efficiency drops to 0 % at 90° and 270° angular positions. A slight increase in the efficiency by 10 % is observed at 180°. Finally, a rapid rise is observed after crossing 270° until it reaches the maximum value at 360°. This gives a clear idea about the order of alignments between RX and TX for achieving decent performance.

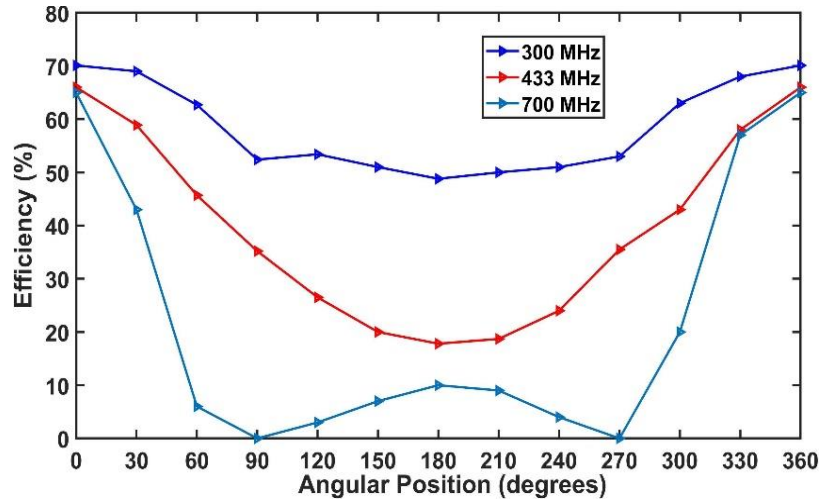


Fig. 5.20. Effect of angular misalignment of RX on efficiency.

Furthermore, the challenge while adding the fourth frequency band to the proposed WIPT system design is discussed. As an example, the fourth frequency band is taken to be 900 MHz. This is obtained by adding an excitation slot and rectangular loop to the already designed tri-band resonator, as shown in Fig. 5.21, employed in the WIPT system. Furthermore, the excitation slot is mounted by external capacitor C_{P4} to obtain the desired resonance at fourth band. The combination of inductance of each rectangular loop and individual capacitor mounted in the excitation slot gives rise to filter characteristics at four frequency bands i.e. 300 MHz, 433

MHz, 700 MHz and 900 MHz. However, during the design of eventual WIPT system, the impedance matching using open stub technique results in poor matching performance at 900 MHz frequency band as shown in the EM simulated S-parameter plot in Fig. 5.22. In addition, the value of reflection coefficient $|S_{11}|$ becomes greater than -10 dB.

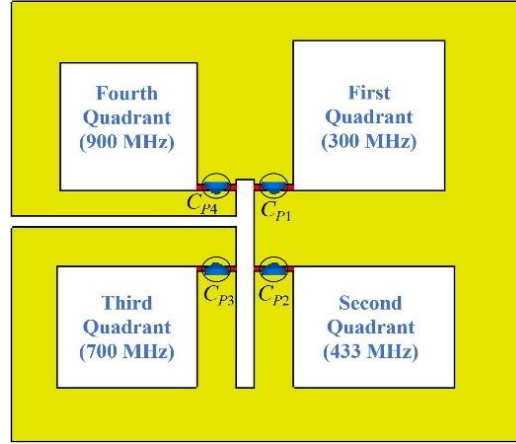


Fig. 5.21. Geometry of quad-band DGS resonator.

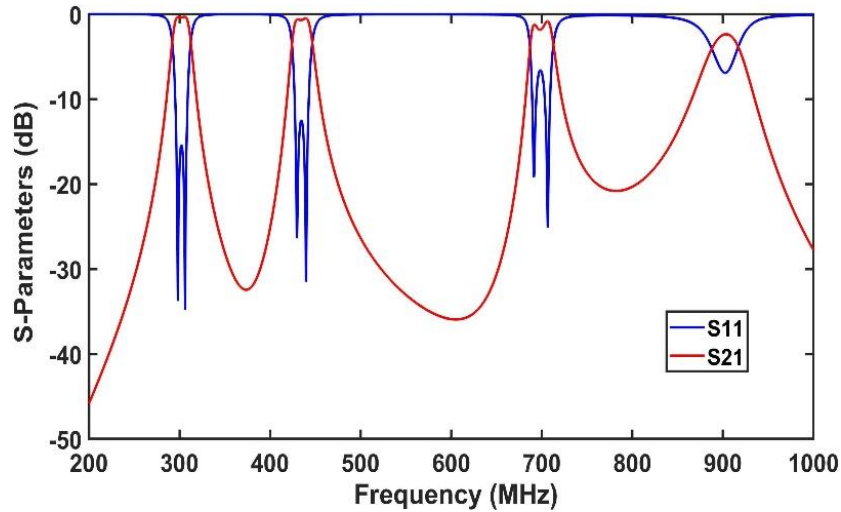


Fig. 5.22. EM simulated S-parameter plot of the quad-band WIPT system.

To overcome the challenge of poor matching performance, the proposed quad-band WIPT system design can be revamped by re-arranging each resonator in a quadrant. The re-arrangement needs to be performed in a pattern so that a common stub runs over the excitation slot of each resonator. Most importantly, the required stub length for each resonator operating at a particular frequency should be maintained to obtain perfect match. However, it may face challenges during measurement as the alignments between TX and RX will be critical, losses will be more prominent with the increase in bands and hence will result in poorer system efficiency, etc.

5.6. Conclusion

In this chapter, a systematic scheme to measure and characterize a WIPT system has been proposed. At first, a new compact tri-band WIPT system design using rectangular shaped defects arranged in the form of tree type structure has been proposed. This entailed reports of theoretical foundations which were then subsequently validated through EM simulation, circuit simulation, and measurement results. The reported DGS based WIPT system exhibits respective measured efficiency of 70.1%, 66%, and 65% at 300, 433, and 700 MHz at a separation distance of 23 mm. The proposed system demonstrates the ability to transfer power at 300 MHz and information transfer in duplex-mode at 433 MHz for down-link communication, and 700 MHz for up-link communication. The experimental setup for power transfer is explained by illuminating an LED, whereas for information transfer, a string of data is transferred over two different frequencies to validate up-link and downlink communication. Moreover, for the first time reported tri-band system, the designed system possesses high degree of compactness with excellent FoM and thus substantially advances the existing state-of-the-art. Finally, the proposed work also provides insights about the challenges associated with the measurement related to misalignment between the TX and RX and the corner cases to achieve the best performance.

Chapter 6

Single Input Multiple Output (SIMO) and Multi Input Multiple Output (MIMO) DGS-WPT System

This chapter is based on the following research paper:

- S. Verma, D. Rano, and M. S. Hashmi “Dual-Band SIMO and MIMO Based Defected Ground Structure for Wireless Power Transfer Applications,” *IEEE Transactions on Microwave Theory and Techniques*. (Under Review)

This chapter presents a dual-band SIMO- and MIMO- WPT system using triangular shaped defected ground structure (DGS). In SIMO-WPT system, a single transmitter operating at 300 MHz and 700 MHz, inputs power to two different receivers. On the other hand, in MIMO-WPT system, two inputs of the dual-band transmitter operating at 300 MHz and 700 MHz transfers power to two individual DGS receivers. The power transfer between the transmitter section and the receiver section of both the WPT systems are based on the principle of resonant inductive coupling. Moreover, the proposed SIMO and MIMO WPT systems are shown for simultaneous power transfer at two operating frequencies.

The next section elaborates the design and modelling of the transmitter (TX) and receiver (RX-1 and RX-2) DGS resonator of SIMO WPT system. The technique to develop a SIMO DGS-WPT system and its equivalent circuit are shown in Section 6.2. Section 6.3 presents the realization of MIMO DGS-WPT system and the development of its equivalent circuit. The comparison of both SIMO and MIMO WPT systems are explained in section 6.4. Subsequently, experimental evaluation of the proposed designs are included in section 6.5.

6.1. SIMO DGS Resonator

In Fig. 6.1, a dual-band defected ground structure (DGS) based transmitter (TX) and receiver (RX) resonator are realized separately to study the band-stop filter (BSF) characteristics of single input-multiple output (SIMO) WPT system. The DGS resonators are designed using Rogers 3003 substrate having $\epsilon_r = 3.3$, substrate thickness $h_{sub} = 0.76$ mm, and metal thickness t

$= 17.5 \mu\text{m}$. In TX-DGS resonator, the top plane consists of 50Ω microstrip feed line (MSL) of length L_F and width W_F , and is connected to lumped ports P_1 and P_2 as shown in Fig. 6.1(a). The bottom plane, of overall length L_{sub} and width W_{sub} , shown in Fig. 6.1(b), consists of two cascaded triangular shaped defects of height H_{T1} and H_{T2} , and base B_{T1} and B_{T2} respectively. Both the defects are connected to individual excitation slots of length L_S and width W_S . Furthermore, these excitation slots are mounted by compensation capacitor C_{P1} and C_{P2} to achieve distinct resonance frequencies at 300 MHz and 700 MHz. Here, both the excitation slots are located exactly below the microstrip feed line. On the other hand, in RX-DGS resonator as shown in the Fig. 6.2, two single band triangular shaped DGS resonators operating separately at 300 MHz and 700 MHz are designed. The top plane, of overall length L_{sub2} and width W_{sub2} , shown in Fig. 6.2(a), is similar for both the RX-DGS resonators operating at 300 MHz and 700 MHz. In addition, it consists of 50Ω microstrip feed line of length L_{F2} and width W_{F2} . Moreover, the bottom plane structures shown in Fig. 6.2(b)-(c) are identical in size except the height H_{T3} and H_{T4} of the triangular shaped defects, and the values of the compensation capacitors C_{P3} and C_{P4} . Subsequently, EM simulation is performed using CST Microwave Studio on TX-DGS resonator. Fig. 6.3 depicts EM simulated S-parameter plot of the TX-DGS resonator showing dual-band bandstop response at 300 MHz and 700 MHz respectively. Moreover, a single band S-parameter response at 300 MHz and 700 MHz can also be obtained for two separate RX-DGS resonators. The design parameters of dual-band triangular shaped TX- and RX-DGS resonators are listed in Table 6.1.

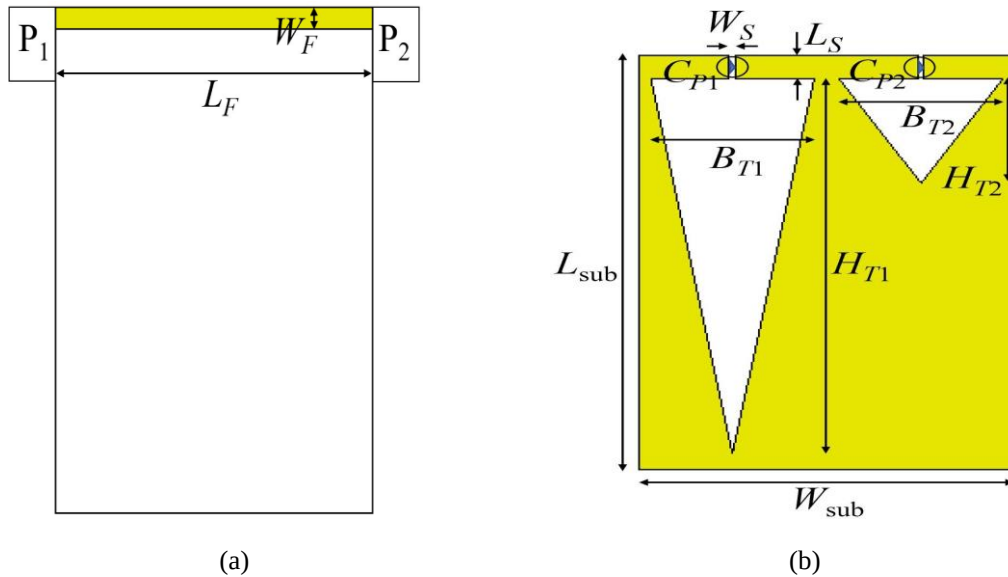


Fig. 6.1. SIMO TX-DGS resonator. (a) Top plane. (b) Bottom plane.

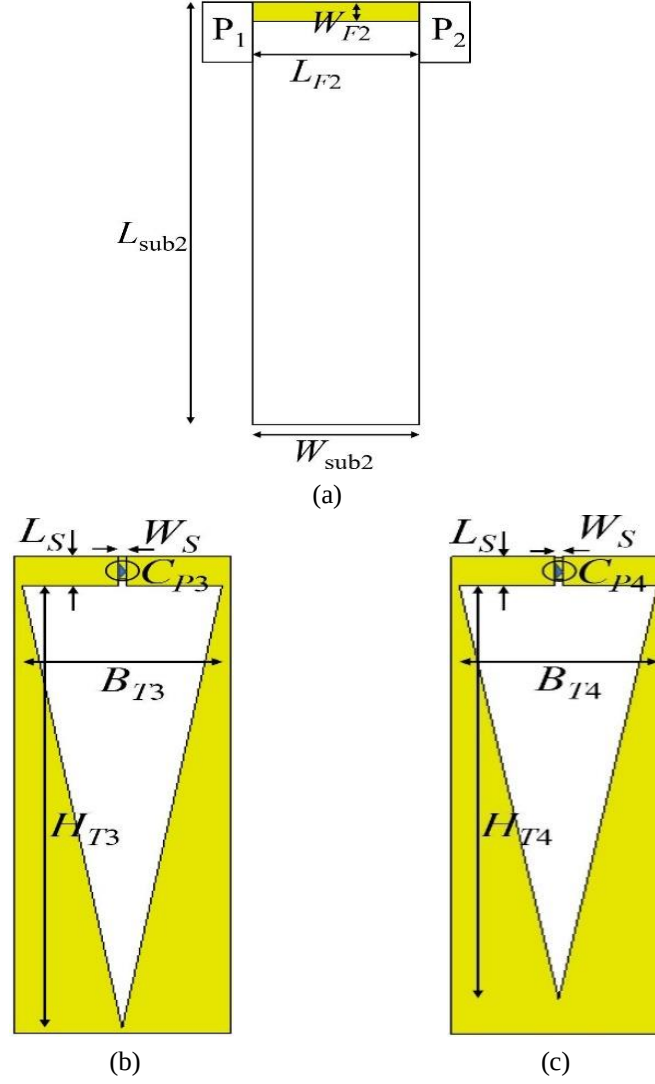


Fig. 6.2. SIMO RX-DGS resonator. (a) Top plane. (b) Bottom plane of DGS resonator operating at 300 MHz. (c) Bottom plane of DGS resonator operating at 700 MHz.

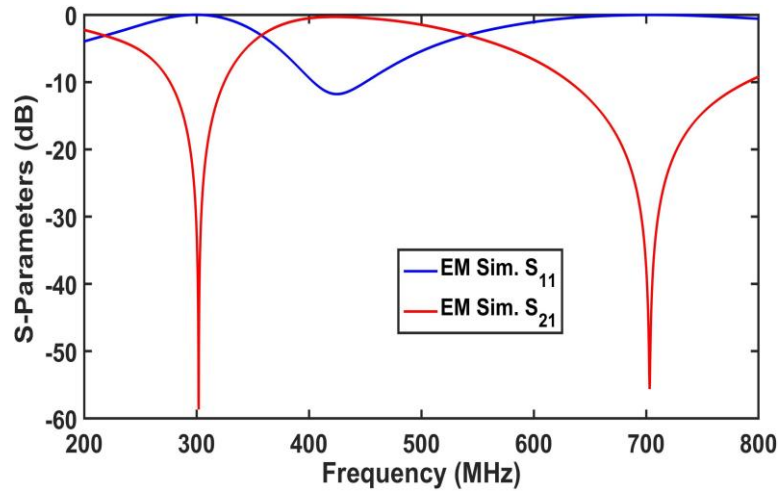


Fig. 6.3. EM simulated S-parameter plot of TX-DGS resonator operating at 300 MHz and 700 MHz.

Table 6.1: Design Parameters of Dual-Band SIMO DGS Resonator.

Design Parameters	Values	Design Parameters	Values
L_F (mm)	30	L_{sub2} (mm)	41
$W_F = W_{F2}$ (mm)	1.9	$W_{sub2} = L_{F2}$ (mm)	13
L_{sub} (mm)	43	H_{T3} (mm)	38
W_{sub} (mm)	30	$B_{T3} = B_{T4}$ (mm)	12
H_{T1} (mm)	38.9	H_{T4} (mm)	35.5
$B_{T1} = B_{T2}$ (mm)	13	C_{P1} (pF)	7.1
H_{T2} (mm)	10.8	C_{P2} (pF)	3.2
L_S (mm)	2.5	C_{P3} (pF)	6.7
W_S (mm)	0.5	C_{P4} (pF)	1.2

6.2. SIMO DGS-WPT System Design

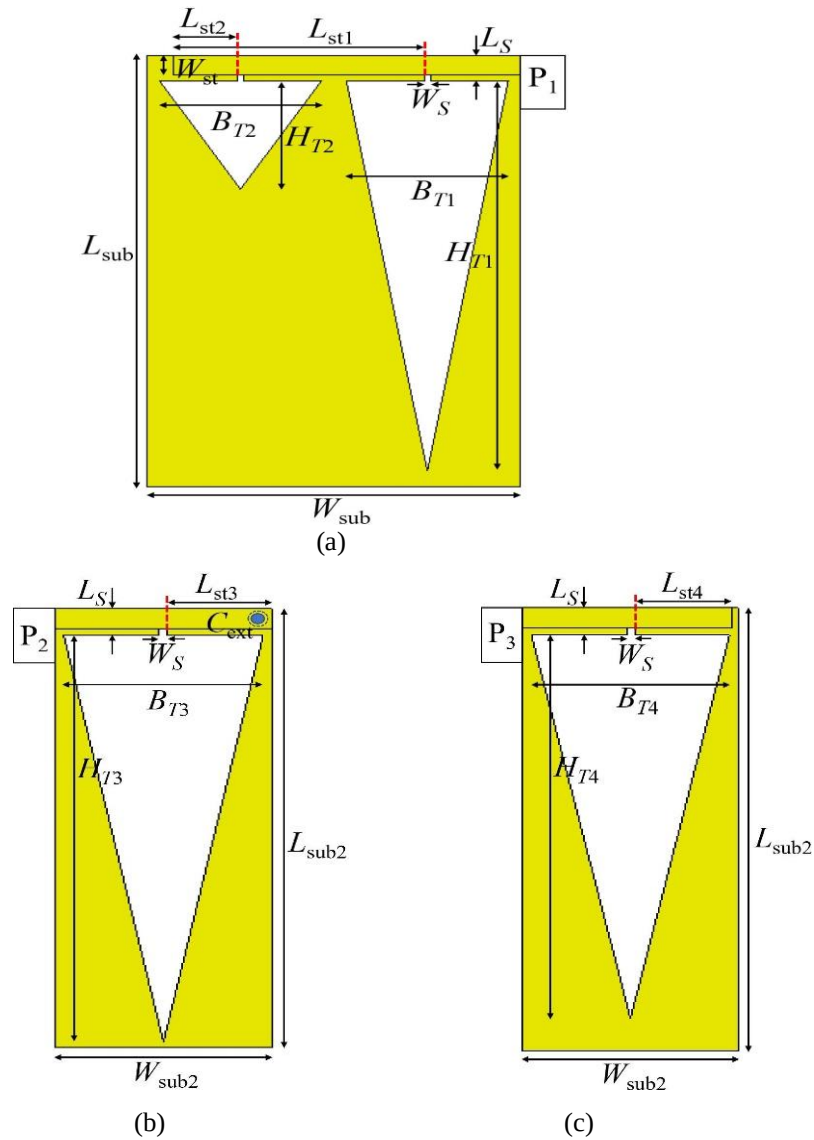


Fig. 6.4. Top and bottom plane of modified SIMO DGS resonator. (a) TX-DGS resonator operating at 300 MHz and 700 MHz. (b) RX-1 DGS resonator operating at 300 MHz. (c) RX-2 DGS resonator operating at 700 MHz.

To achieve WPT between a single TX-DGS resonator and two separate RX-1 and RX-2 DGS resonators, the TX- and both the RX- DGS resonators are modified with one end of the microstrip feed line (MSL) terminal connected with a $50\ \Omega$ port and the other end as open circuit. In Fig. 6.4(a), the top plane of the modified TX-DGS resonator is shown which consists of open stub at the top plane for impedance matching. The length of open stub at the two operating frequencies are different i.e. at $f_{01} = 300\text{ MHz}$, $L_{st1} = 20.15\text{ mm}$ and at $f_{02} = 700\text{ MHz}$, $L_{st2} = 5.15\text{ mm}$. On the other hand, in Figs. 6.4(b)-(c), the modified RX-1 and RX-2 DGS resonators operating separately at 300 MHz and 700 MHz are shown. The top plane of the RX-1 DGS resonator operating at 300 MHz consists of an open stub of length $L_{st3} = 6.25\text{ mm}$, while the RX-2 DGS resonator operating at 700 MHz consists of a stub of length $L_{st4} = 5.85\text{ mm}$. The stub length of each resonator is measured from the reference plane represented by dashes. Most importantly, the reference plane is positioned exactly above the excitation slot to obtain near perfect match. Moreover, the stub width W_{st} of TX-DGS resonator is similar for both the RX-1 and RX-2 DGS resonator. Here, the RX-1 DGS resonator consists of an external capacitor $C_{ext} = 1.1\text{ pF}$ shorted between the top plane and the bottom plane. This is required if the stub length does not provide enough capacitance for perfect matching. Furthermore, it can be seen that the bottom plane of the TX- and both the RX- DGS resonators are similar to the bottom plane of the original resonators as described in the previous section. In Fig. 6.5, these resonators are placed back-to-back separated by a distance of $h = 18\text{ mm}$. The separation distance transforms the bandstop characteristics of the TX- and RX-1 and RX-2 DGS resonator into bandpass characteristics. The value of separation distance is identified using EM simulation in critical coupling region. In critical coupling region, the power is transferred between the TX and RX-1 and RX-2 DGS resonator at maximum efficiency. This is due to the fact that input impedance is equivalent to $50\ \Omega$ at the operating frequency. Also, a gap of $d = 2\text{ mm}$ is maintained between the RX-1 and RX-2 DGS resonator to satisfy the SIMO topology.

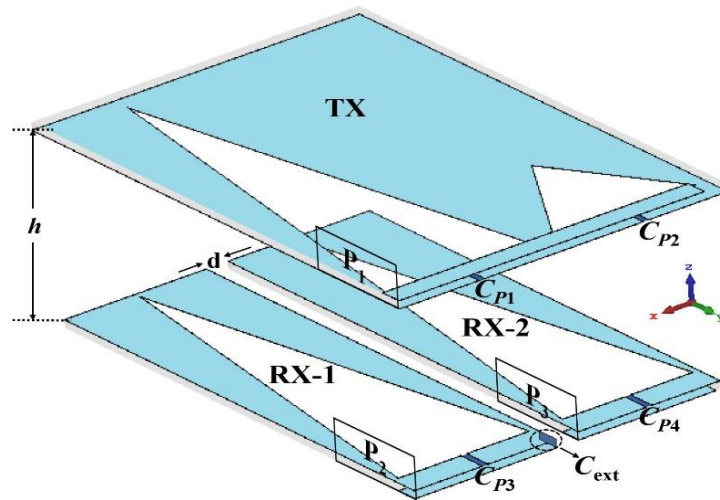


Fig. 6.5. Back-to-back configuration of SIMO based DGS-WPT system.

6.2.1. Equivalent Circuit

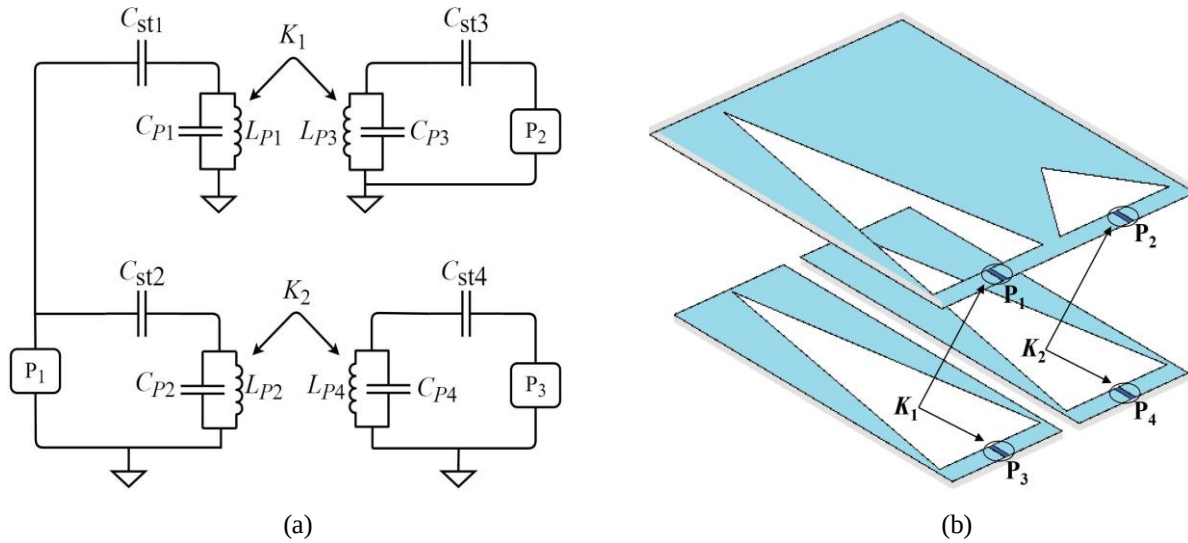


Fig. 6.6. Coupled SIMO DGS-WPT system. (a) Equivalent circuit. (b) Design parameters extraction using CST tool.

The proposed WPT system is modelled using parallel LC circuits and stub capacitance C_{st} as shown in Fig. 6.6(a). In each LC circuit, a triangular shaped defect represents inductance (L) and compensation capacitor loaded in the excitation slot depicts capacitance (C). Both L and C combine to show resonance at the operating frequency. To obtain the inductance of each resonator using the expression in (6.1) [66], first, the imaginary part of impedance of each resonator is computed. This is done by eliminating the MSL from the top plane and substituting lumped port in the excitation slot of each resonator during the EM simulation as shown in Fig. 6.6(b). The inductance of the modified TX- and RX-1 and RX-2 DGS resonators are mutually coupled which is defined by coupling coefficients K_1 and K_2 . These coupling coefficients between the TX- and RX-1 and RX-2 DGS resonators are related to mutual coupling L_{m1} and L_{m2} by (6.2) [28]. In this equivalent circuit, the capacitance C_{st} across the stub length of individual resonator is computed using the expression shown in (6.3) [28], where Z_0 is the characteristic impedance of the stub, β is the phase shift constant. In the modified TX-DGS resonator, $C_{st1} = 1.9$ pF at 300 MHz, and $C_{st2} = 0.55$ pF at 700 MHz, while in the modified RX-DGS resonators, $C_{st3} = 1.82$ pF at 300 MHz, and $C_{st4} = 0.61$ pF at 700 MHz. The four stub capacitance values C_{st1} , C_{st2} , C_{st3} , and C_{st4} are responsible for impedance matching at the two operating frequencies. These capacitances are connected between the LC circuit and lumped ports P_1 , P_2 and P_3 . Subsequently, the equivalent circuit is simulated using advanced design system (ADS) tool to obtain the S-parameter plot. The S-parameter plot obtained from both the circuit simulation and EM simulation compare favorably, as can be seen in Fig. 6.7(a), and thus the reliability of the developed circuit model is validated. The optimized design parameters of the SIMO DGS-WPT system are listed in Table 6.2. Moreover, Fig. 6.7(b) shows the cross-coupling between RX-1

and RX-2 modified DGS resonators. It can be seen that the cross-coupling parameter ($S_{32} < -20$ dB) shows perfect isolation at the two operating frequencies.

$$L_{P1} = \frac{\text{imag}(Z_{11})}{2\pi f_{01}}, L_{P3} = \frac{\text{imag}(Z_{33})}{2\pi f_{01}} \quad (6.1)$$

$$L_{P2} = \frac{\text{imag}(Z_{22})}{2\pi f_{02}}, L_{P4} = \frac{\text{imag}(Z_{44})}{2\pi f_{02}}$$

$$K_1 = \frac{L_{m1}}{L_{P1}} = \frac{\text{imag}(Z_{13})}{2\pi f_{01} L_{P1}}, K_2 = \frac{L_{m2}}{L_{P2}} = \frac{\text{imag}(Z_{24})}{2\pi f_{02} L_{P2}} \quad (6.2)$$

$$C_{st1} = \frac{\tan(\beta L_{st1})}{2\pi f_{01} Z_0}, C_{st3} = \frac{\tan(\beta L_{st3})}{2\pi f_{01} Z_0} \quad (6.3)$$

$$C_{st2} = \frac{\tan(\beta L_{st2})}{2\pi f_{02} Z_0}, C_{st4} = \frac{\tan(\beta L_{st4})}{2\pi f_{02} Z_0}$$

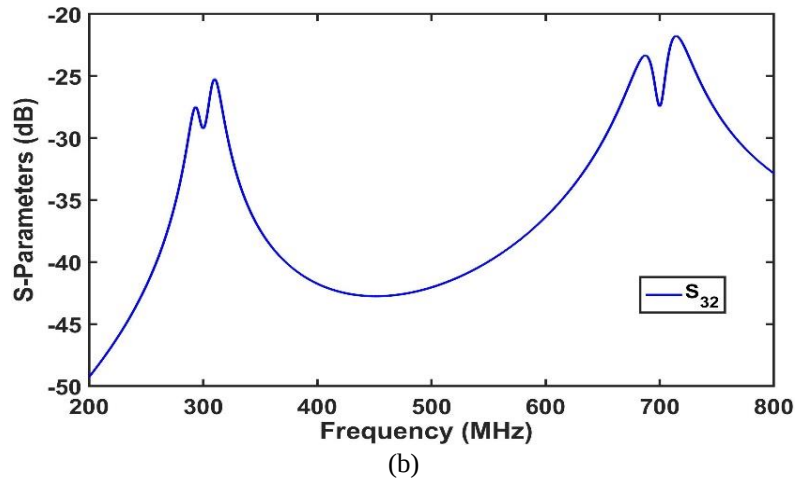
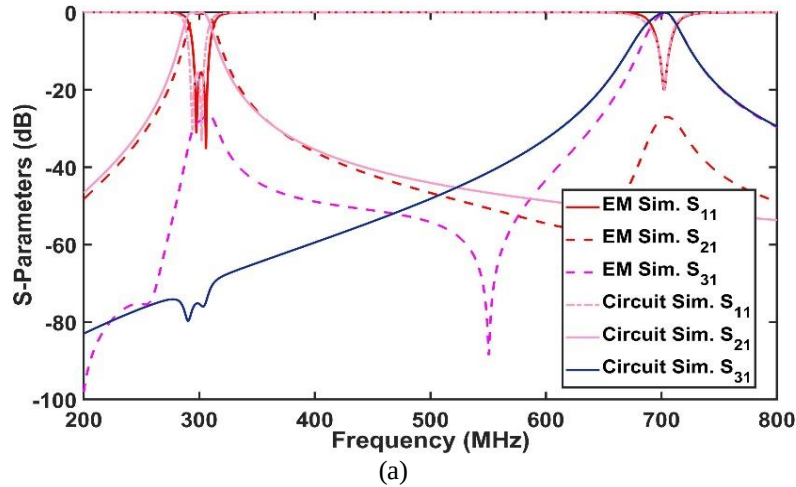


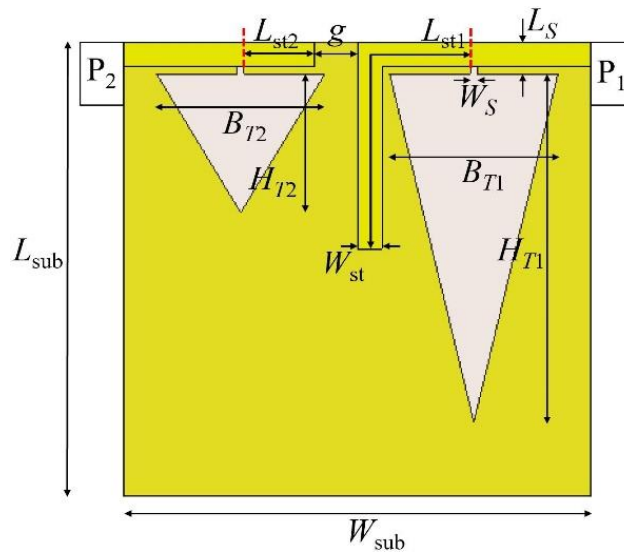
Fig. 6.7. EM simulation and circuit simulation S-parameter plot of SIMO DGS-WPT system. (a) Comparison of EM simulation and circuit simulation S-parameters. (b) Cross coupling between RX-1 and RX-2 modified DGS resonators.

Table 6.2: Design Parameters of Dual-Band SIMO and MIMO DGS-WPT System.

Design Parameters	SIMO	MIMO	Design Parameters	SIMO	MIMO	Design Parameter	SIMO	MIMO
L_{sub} (mm)	43	36	W_{sub2} (mm)	13	16	C_{p2} (pF)	2.7	2.7
W_{sub} (mm)	30	36	L_{st3} (mm)	6.25	7.75	C_{p3} (pF)	5.1	6.2
L_{st1} (mm)	20.15	23.25	H_{T3} (mm)	38	29.6	C_{ext} (pF)	1.1	1.3
L_{st2} (mm)	5.15	5.45	B_{T3} (mm)	12	14	C_{p4} (pF)	0.6	0.8
W_{st} (mm)	1.9	1.9	L_{st4} (mm)	5.85	7.45	C_{st1} (pF)	1.9	2.3
L_s (mm)	2.5	2.5	H_{T4} (mm)	35.5	27.5	C_{st2} (pF)	0.55	0.49
W_s (mm)	0.5	0.5	B_{T4} (mm)	12	14	C_{st3} (pF)	1.82	2.3
H_{T1} (mm)	38.9	27.6	L_{p1} (nH)	39	28.3	C_{st4} (pF)	0.61	0.76
B_{T1} (mm)	13	13	L_{p2} (nH)	16	15.8	L_{m1} (nH)	2.08	1.76
H_{T2} (mm)	10.8	11	L_{p3} (nH)	41.5	34.1	L_{m2} (nH)	0.91	1
B_{T2} (mm)	13	13	L_{p4} (nH)	44	35.16	η_1 (%) at 300 MHz	79.4	88.5
L_{sub2} (mm)	41	33	C_{p1} (pF)	5.5	7.8	η_2 (%) at 700 MHz	79.5	88.3

6.3. MIMO DGS-WPT System Design

In MIMO DGS-WPT system, Fig. 6.8 represents the top and bottom plane of the two input TX-DGS resonator and two separate RX-1 and RX-2 DGS resonators. The top plane of the two input TX-DGS resonator and RX-1 and RX-2 DGS resonator, as shown in Figs. 6.8(a)-(c), consists of independent open stubs of length $L_{st1} = 23.25$ mm at 300MHz and $L_{st2} = 5.45$ mm at 700Mhz, $L_{st3} = 7.75$ mm at 300MHz and $L_{st4} = 7.45$ mm at 700 MHz. Here, the two stubs of length L_{st1} and L_{st2} are separated by a gap g which gives rise to gap capacitance C_g . The width W_{st} of each stub holds a similar value of 1.9 mm. Moreover, it can be seen that the bottom plane of the TX-, RX-1 and RX-2 DGS resonators follows a similar defect pattern as the bottom plane of the modified SIMO DGS resonator. However, the dimensions and the values of the compensation capacitors are different.



(a)

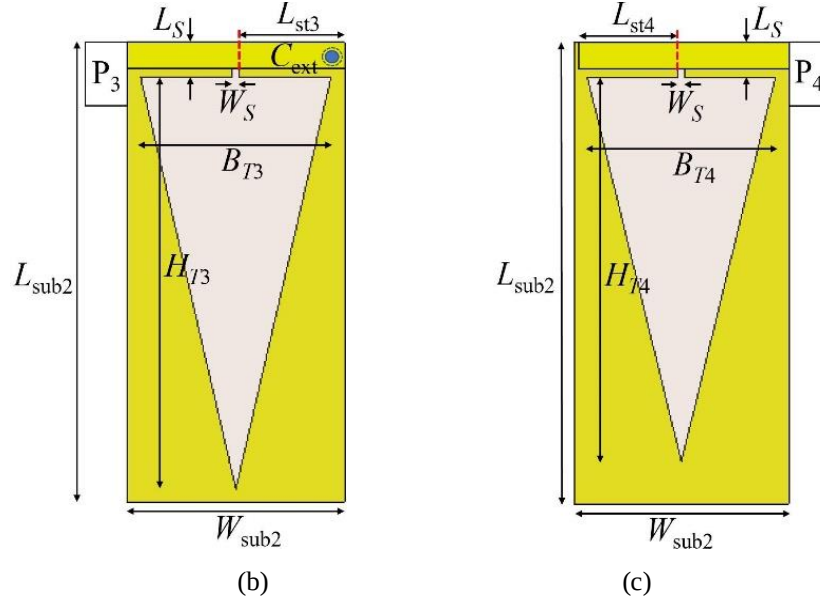


Fig. 6.8. Top and bottom plane of MIMO DGS resonator. (a) Two input TX-DGS resonator operating at 300 MHz and 700 MHz. (b) RX-1 DGS resonator operating at 300 MHz. (c) RX-2 DGS resonator operating at 700 MHz.

Furthermore, the two input TX-DGS resonator are coupled with RX-1 and RX-2 receiver DGS resonators in a back-to-back configuration at a separation distance h as shown in the Fig. 6.9. Both the RX-1 and RX-2 DGS resonators are separated by a similar gap d as shown in the SIMO DGS-WPT system. Moreover, the TX-, RX-1 and RX-2 DGS resonators are terminated by ports P_1 and P_2 at the TX side and ports P_3 and P_4 at the RX side.

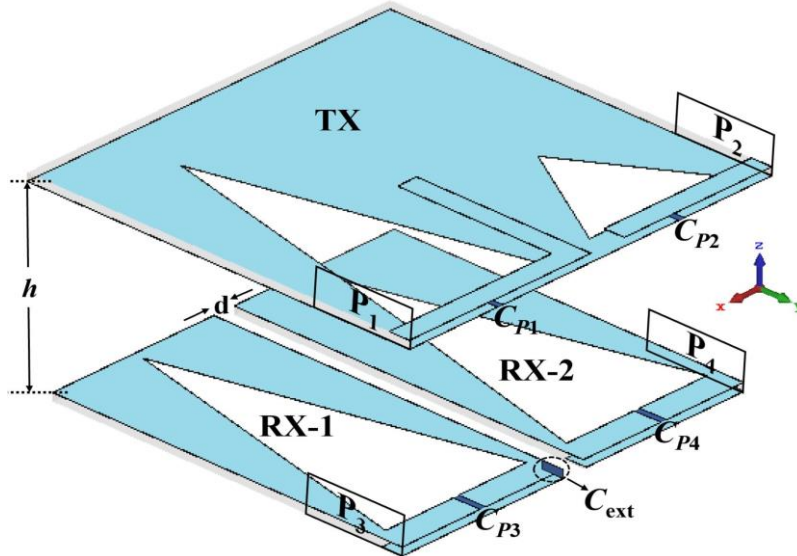


Fig. 6.9. Back-to-back configuration of MIMO based DGS-WPT system.

6.3.1. MIMO WPT System Equivalent Circuit

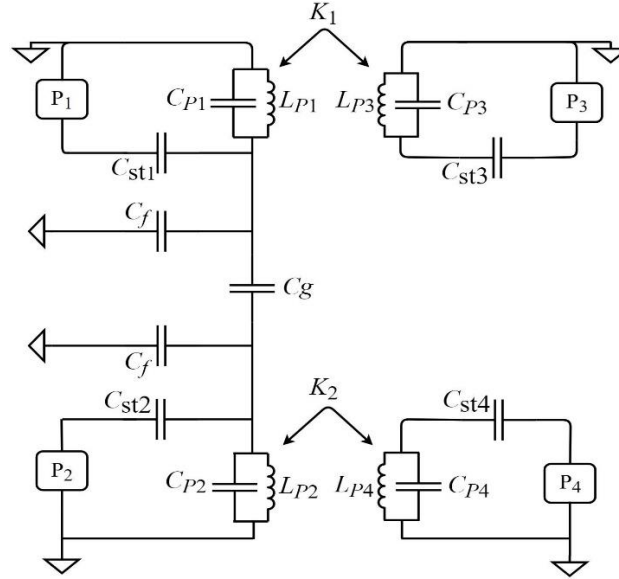


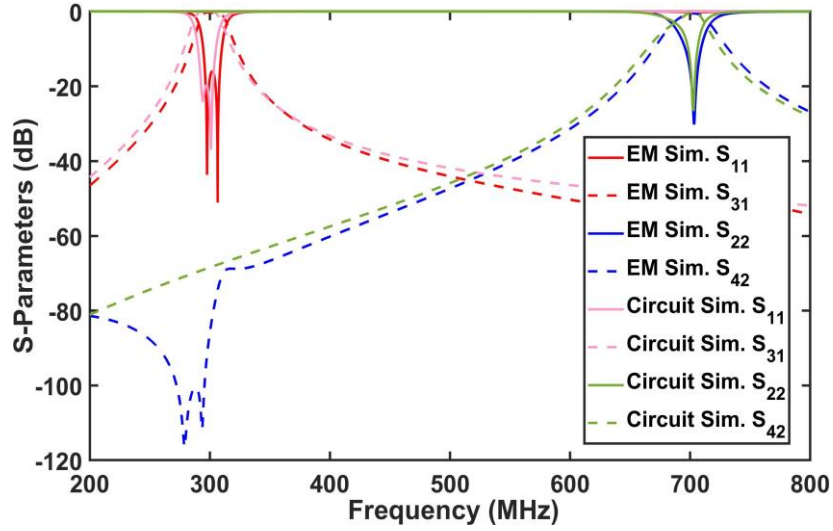
Fig. 6.10. Equivalent circuit of MIMO based DGS-WPT system.

The Fig. 6.10 depicts the equivalent circuit of MIMO DGS-WPT system. The RF input to the TX-DGS resonator is achieved through two ports P_1 and P_2 . Both these ports are connected to two different LC (L_{P1} , C_{P1} and L_{P2} , C_{P2}) circuits through independent stub capacitance C_{st1} and C_{st2} . In addition, the gap capacitance C_g between the two independent stub capacitance C_{st1} and C_{st2} are computed using the equation in (6.4). The fringing capacitance C_f rising from the edges of the open ended stub line connected to P_1 and P_2 is computed using (6.5). The values of both the gap and fringing capacitance are obtained by extracting Y-parameters using EM simulation at either of the two resonance frequencies. In the simulation model, the ground plane is eliminated and the two open stub lines connected to two ports remains unchanged. Furthermore, both the receiver DGS resonators RX-1 and RX-2 pick up the magnetic field using their respective LC circuit. The current flowing through the LC (L_{P3} , C_{P3} and L_{P4} , C_{P4}) circuits of both the RX-DGS resonators are received by ports P_3 and P_4 via stub capacitance C_{st3} and C_{st4} . The values of inductance, coupling coefficients and stub capacitance of TX- and RX-1 and RX-2 DGS resonators are computed using the similar equations in (6.1), (6.2) and (6.3). The design parameters and the values of the MIMO equivalent circuit are described in Table 6.2. Subsequently, the equivalent circuit is simulated in the ADS environment to obtain the S-parameter plot. A comparison of both the EM simulation and circuit simulation results, as shown in Fig. 6.11(a), verify good agreement with each other. Moreover, Fig. 6.11(b) shows the cross coupling between the two inputs of TX-DGS resonator and RX-1 and RX-2 DGS resonator. The cross coupling between the two inputs of the TX-DGS resonator is represented by S_{21} whereas, the cross coupling between RX-1 and RX-2 DGS resonators is given by S_{43} . Both

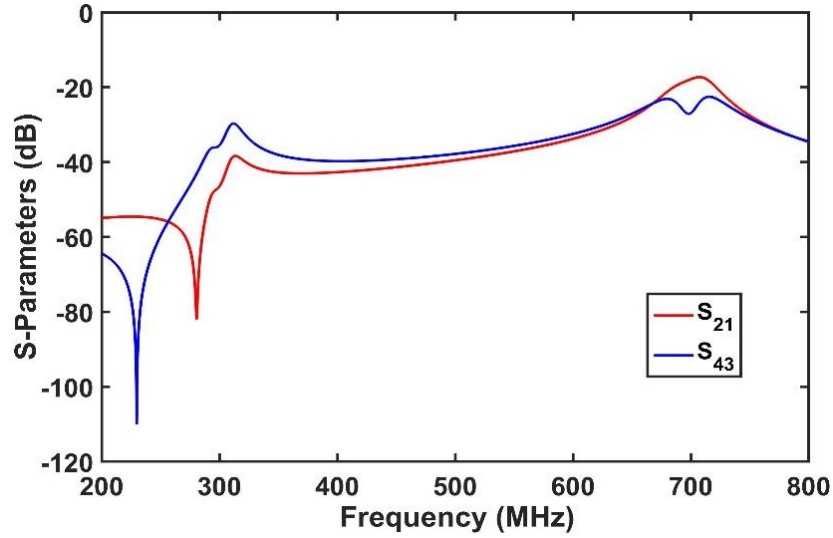
the coupling parameters S_{21} and S_{43} are lower than -20 dB which depict negligible leakage at both the frequencies of interest.

$$C_g = \frac{-\text{imag}(Y_{21})}{2\pi f_0} \quad (6.4)$$

$$C_f = \frac{\text{imag}(Y_{11} + Y_{21})}{2\pi f_0} \quad (6.5)$$



(a)



(b)

Fig. 6.11. EM simulation and circuit simulation S-parameter plot of MIMO DGS-WPT system. (a) Comparison of EM simulation and circuit simulation S-parameters. (b) Cross coupling between two inputs of TX-DGS resonator and RX-1 and RX-2 DGS resonator.

6.4. Comparison of SIMO and MIMO DGS-WPT System

In this section, a comparison among SIMO and MIMO topology of DGS based WPT system is discussed. In SIMO DGS-WPT system, the area of the TX resonator (1290 mm^2) and RX-1 and RX-2 resonator (533 mm^2) is almost similar to the area of the TX resonator (1296 mm^2) and RX-1 and RX-2 resonator (528 mm^2) of the MIMO DGS-WPT system. However, it can be seen in Fig. 6.12 that the efficiency of MIMO based DGS-WPT system at two frequencies ($\eta_1 = 88.5 \%$ at $f_{01} = 300 \text{ MHz}$, $\eta_2 = 88.3 \%$ at $f_{02} = 700 \text{ MHz}$) is greater than the efficiency of SIMO DGS-WPT system at similar frequencies ($\eta_1 = 79.4 \%$ at $f_{01} = 300 \text{ MHz}$, $\eta_2 = 79.5 \%$ at $f_{02} = 700 \text{ MHz}$). In SIMO WPT system, this occurs as a result of common open stub line in the TX resonator which does not provide perfect impedance matching at the two design frequencies. However, to obtain perfect matching, firstly, the area of the transmitter resonator is increased. Then the triangular defect and excitation slot operating at f_{01} and f_{02} is tuned according to the required stub length for perfect match. Furthermore, in MIMO WPT system, two independent open stubs in the TX resonator provides perfect matching at the two frequencies. Most importantly, the SIMO WPT system can achieve identical efficiency figures as MIMO WPT system. However, the transmitter area of the SIMO TX resonator will exceed the MIMO TX resonator.

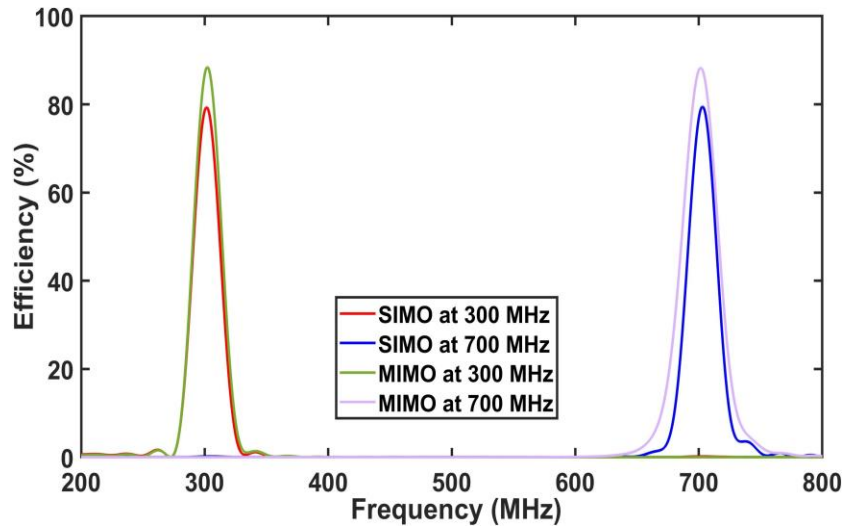


Fig. 6.12. Efficiency comparison between SIMO and MIMO DGS-WPT system.

6.5. Measurements and Results

This section includes the thorough measurement and characterization of the developed SIMO and MIMO DGS-WPT system. The fabricated prototype shown in Fig. 6.13(a) and Fig. 6.13(b) depicts the top and bottom planes of the proposed SIMO and MIMO DGS resonator. In addition, the measurement setup shows the TX and RX-1 and RX-2 DGS resonator of both the

SIMO and MIMO DGS-WPT system are separated by a plastic foam to maintain a distance of $h = 18$ mm for optimal efficiency. Furthermore, the S-parameter plot of the WPT systems are extracted using Agilent vector network analyzer (VNA, N9926A) operating over the frequency range of 200-800 MHz. The measured S-parameter plot of SIMO DGS-WPT system, in Fig. 6.14(a), shows the comparison with EM simulated S-parameter plot. The measured S-parameter plot depicts resonance characteristics at 300 MHz with $|S_{21}| = -1.63$ dB, while at 700 MHz with $|S_{31}| = -1.55$ dB. Both the frequencies at 300 MHz and 700 MHz show $|S_{11}|$ with approximate value of -20 dB. On the other hand, in Fig. 6.14(b), the comparison of measured and EM simulated S-parameter plot of MIMO DGS-WPT system is shown. The measured S-parameter plot depicts resonance characteristics at 300 MHz with $|S_{31}| = -1.24$ dB, while at 700 MHz with $|S_{42}| = -1.35$ dB. The return loss coefficient $|S_{11}|$ and $|S_{22}|$ are approximately equivalent to -25 dB. The measured results of both the WPT systems shows slight variation in the resonance frequency from the operating frequency of 300 MHz and 700 MHz. This occurs due to the tolerances of the compensation capacitors and external capacitor of both the SIMO and MIMO DGS-WPT system. In addition, other factors such as the inset of plastic foam which adds on to the permittivity, and the in-house fabrication introduces losses such as soldering loss, connector loss etc. also introduce some anomaly. To overcome the shift in the frequency, capacitor values with low tolerances and frequency tuning should be used.

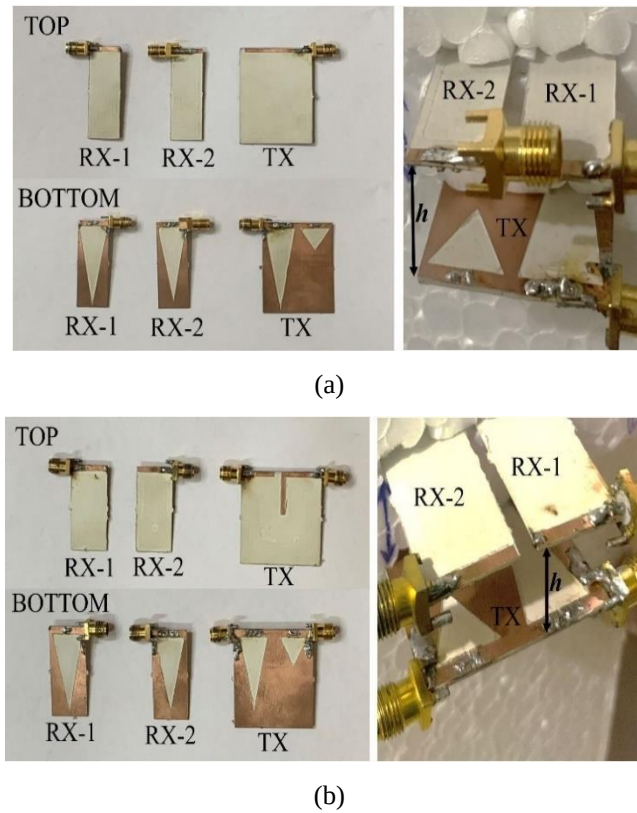
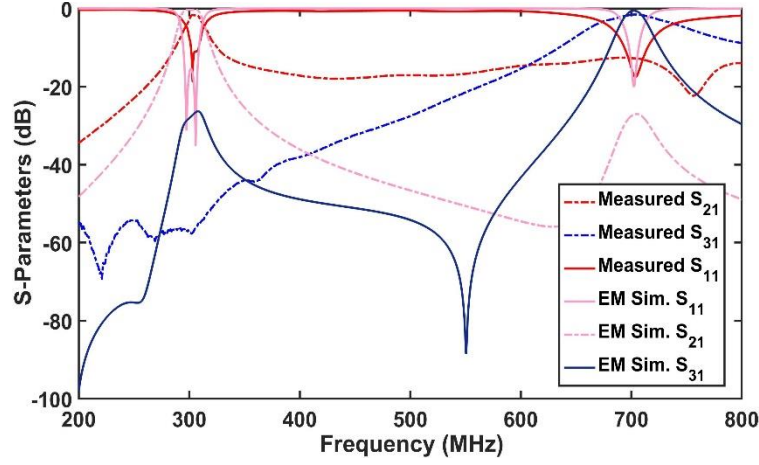
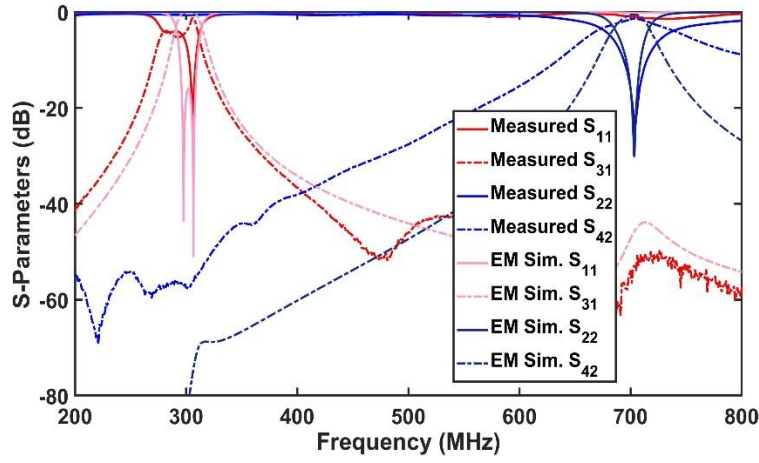


Fig. 6.13. Measurement setup of the fabricated WPT system using (a) SIMO DGS resonator. (b) MIMO DGS resonator.



(a)



(b)

Fig. 6.14. Comparison of EM simulated and measured S-parameter plot. (a) SIMO DGS-WPT system. (b) MIMO DGS-WPT system.

$$\begin{aligned}
 \eta_{SIMO1} &= \frac{|S_{21}|^2}{1-|S_{11}|^2}, \quad \eta_{SIMO2} = \frac{|S_{31}|^2}{1-|S_{11}|^2} \\
 \eta_{MIMO1} &= \frac{|S_{31}|^2}{1-|S_{11}|^2}, \quad \eta_{MIMO2} = \frac{|S_{42}|^2}{1-|S_{22}|^2}
 \end{aligned} \tag{6.6}$$

The measured efficiency of SIMO and MIMO DGS-WPT system at 300 MHz and 700 MHz is computed using the equation in (6.6). The efficiency (η_{SIMO1}) of SIMO DGS-WPT system at 300 MHz is recorded at 68.7 %, while, at 700 MHz, the efficiency (η_{SIMO2}) is 70 %. However, in MIMO DGS-WPT system, the efficiency (η_{MIMO1}) at 300 MHz is 75.1 %, while at 700 MHz, the efficiency (η_{MIMO2}) is 73.2 %.

6.6. Conclusion

In this chapter, a new dual-band DGS-WPT system based on single input-multiple output (SIMO) and multi input-multiple output (MIMO) topologies has been proposed. First, the theoretical foundations are developed which are subsequently validated through EM simulation, circuit simulation, and measurement results. The reported DGS-WPT system based on SIMO topology exhibits respective measured efficiency of 68.7 % and 70 % at 300 MHz and 700 MHz, whereas the DGS-WPT system based on MIMO topology measures efficiency of 75.1 % and 73.2 % at 300 MHz and 700 MHz respectively. Both the systems achieve different efficiency characteristics in an almost similar area of TX- and RX-1 and RX-2 DGS resonators separated by a distance of $h = 18\text{mm}$.

Chapter 7

Conclusion and Future Work

7.1 Conclusion

In the thesis, we have designed and proposed compact DGS resonators to be utilized for compact energy efficient RF wireless power transfer systems. The resonators possess high quality factor in order to achieve highly efficient coupled WPT system. Moreover, the resonators are tuned to transfer power wirelessly at single, dual and triple band frequency range. The dual and triple band resonators have the advantage to support simultaneous wireless information and power transfer. Subsequently, SIMO and MIMO wireless topology have been implemented by utilizing DGS resonators to enable power transfer between single transmitter and multiple receivers and multiple transmitter and multiple receivers. Furthermore, a summary of each of the proposed works has been described below:

a) First is the design of three different types of compact DGS resonators for applications such as IoT, biomedical implantable devices, and wireless power where miniature-sized circuits play a critical role.

b) Second is the design of two different types of compact highly efficient resonant inductive WPT systems by utilizing two DGS resonators placed in a back-to-back configuration at a critical coupling distance h .

c) Thirdly, a new technique for impedance matching using symmetric coupled lines for resonant inductive WPT system is proposed. This technique provides an enhanced WPT system efficiency and robust operation. In addition, it helps in achieving high FoM in comparison to the state-of-the-art designs.

d) Fourthly, a systematic scheme to measure and characterize a WIPT (wireless information and power transfer) system has been proposed. This system utilizes three frequencies 300, 433, and 700 MHz to obtain WIPT operation. Here, the 300 MHz frequency band is used for power transfer only, whereas the 433 and 700 MHz frequency range is used for downlink and uplink of data transfer. Furthermore, the proposed work also provides insights about the challenges associated with the measurement related to misalignment between the TX and RX and the corner cases to achieve the best performance.

e) Finally, a new dual-band DGS-WPT system based on single input-multiple output (SIMO) and multi input-multiple output (MIMO) topologies has been proposed. These types of WPT systems are used to power portable electronic gadgets. Furthermore, a comparison of

SIMO and MIMO WPT systems, based on efficiency, with similar circuit areas of the TX- and RX- DGS resonators, is made to evaluate the effectiveness of the proposed design.

7.2 Future Work

The future work includes the following aspects:

- 1) To implement a printed resonator with edge coupling/side coupling to achieve power transfer at single and dual frequency bands. The utilization of printed resonator will prevent the etching in the ground plane and can be useful in on body communication.
- 2) To design and model dual band resonant coupled wireless power transfer system using coupled line impedance matching technique.
- 3) To implement a quad band resonant coupled wireless power transfer system using DGS resonator for implantable telemetry systems.
- 4) To develop a technique which can improve the efficiency of WPT system during lateral and angular misalignment of the receiver position. The misalignment results in the degradation of overall WPT system efficiency when the receiver is not in line of sight with the transmitter.
- 5) To work on static and dynamic WPT system for applications involving electric vehicle charging, drones, etc.
- 6) To design the components of a WPT system such as rectenna to deliver power in the far-field region.

References

- [1] O. Jonah and S. V Georgakopoulos, "Wireless power transfer in concrete via strongly coupled magnetic resonance," *IEEE Trans. Antennas Propag.*, vol. 61, no. 3, pp. 1378–1384, 2013.
- [2] M. Falavarjani, M. Shahabadi and J. Rashed Mohassel, "Design and implementation of compact WPT system using printed spiral resonators," *Electron. Lett.*, vol. 50, no. 2, pp. 110–111, Jan. 2014.
- [3] S. Stoecklin, A. Yousaf, T. Volk, and L. Reindl, "Efficient Wireless Powering of Biomedical Sensor Systems for Multichannel Brain Implants," *IEEE Transactions on Instrumentation and Measurement*, vol. 65, no. 4, pp. 754-764, April 2016.
- [4] R. F. Xue, K. W. Cheng and M. Je, "High-efficiency wireless power transfer for biomedical implants by optimal resonant load transformation," *IEEE Trans. Circuits Syst. I*, vol. 60, no. 4, pp. 867–874, 2013.
- [5] S. H. Kang, V. T. Nguyen and C. W. Jung, "Analysis of MR-WPT using planar textile resonators for wearable applications," *IET Microw Anten. Propag.*, vol. 10, no. 14, pp. 1541–1546, 2016.
- [6] A. Buffi, A. Michel, P. Nepa and B. Tellini, "RSSI Measurements for RFID Tag Classification in Smart Storage Systems," *IEEE Transactions on Instrumentation and Measurement*, vol. 67, no. 4, pp. 894-904, April 2018.
- [7] K. R. Brinker, M. Vaccaro and R. Zoughi, "Application-Adaptable Chipless RFID Tag: Design Methodology, Metrics, and Measurements," *IEEE Transactions on Instrumentation and Measurement*, vol. 69, no. 6, pp. 3882-3895, June 2020.
- [8] R. Correia, N. B. D. Carvalho, G. Fukuda, A. Miyaji and S. Kawasaki, "Backscatter Wireless Sensor Network with WPT Capabilities," *International Microwave Symposium*, pp. 1–4, 2015.
- [9] G. Wang et al., "A dual band wireless power and data telemetry for retinal prosthesis," *Int. Conf. IEEE Eng. Med. Biol. Soc.*, pp. 4392–4395, 2006.
- [10] M. Kung and K. Lin, "Dual-band coil module with repeaters for diverse wireless power transfer applications," *IEEE Trans. Microw. Theory Tech.*, vol. 66, no. 1, pp. 332–345, 2018.
- [11] Z. Ling, F. Hu, L. Wang, J. Yu and X. Liu, "Point-to-Point Wireless Information and Power Transfer in WBAN With Energy Harvesting," *IEEE Access*, vol. 5, pp. 8620-8628, 2017.
- [12] S. Aldhaher, P. D. Mitcheson, J. M. Arteaga, G. Kkelis, and D. C. Yates, "Light-weight wireless power transfer for mid-air charging of drones," *European Conference on Antennas and Propagation (EUCAP)*, pp. 336- 340, 2017.

- [13] J. Huh, S. W. Lee, W. Y. Lee, G. H. Cho and C. T. Rim, "Narrow-Width Inductive Power Transfer System for Online Electrical Vehicles," *IEEE Transactions on Power Electronics*, vol. 26, no. 12, pp. 3666-3679, Dec. 2011.
- [14] S. Li and C. C. Mi, "Wireless Power Transfer for Electric Vehicle Applications," *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 3, no. 1, pp. 4-17, March 2015, doi: 10.1109/JESTPE.2014.2319453.
- [15] Z. Yalong, H. Xueliang, Z. Jiaming and T. Linlin, "Design of wireless power supply system for the portable mobile device," in *Proc. IWS*, pp. 1-4, Apr. 2013.
- [16] H. H. Lee, S. H. Kang, C. W. Jung, "3D-spatial efficiency optimisation of MR-WPT using a reconfigurable resonator-array for laptop applications," *IET Microw. Antennas Propag.*, vol. 11, no. 11, pp. 1594-1602, 2017.
- [17] J. Dai and D. C. Ludois, "A survey of wireless power transfer and a critical comparison of inductive and capacitive coupling for small gap applications," in *IEEE Trans. Power Electron.*, vol. 30, no. 11, pp. 6017-6029, 2015.
- [18] Z. Zhang, H. Pang, A. Georgiadis and C. Cecati, "Wireless power transfer—An overview," *IEEE Trans. Ind. Electron*, vol. 66, no. 2, pp. 1044-1058, Feb. 2019.
- [19] S. Hekal, A. Allam, A. B. Abdel-Rahman and R. K. Pokharel, *Compact Size Wireless Power Transfer Using Defected Ground Structures*. Singapore: Springer, 2019.
- [20] F. Jolani, Y. Yu and Z. Chen, "A Planar Magnetically Coupled Resonant Wireless Power Transfer System Using Printed Spiral Coils," *IEEE Antennas and Wireless Propagation Letters*, vol. 13, pp. 1648-1651, 2014, doi: 10.1109/LAWP.2014.2349481.
- [21] S. Hekal and A. B. Abdel-Rahman, "New compact design for short range wireless power transmission at 1 GHz using H-slot resonators," *Proc. EUCAP*, Lisbon, Portugal, pp. 12-17, Apr. 2015.
- [22] S. Hekal, A. B. Abdel-Rahman, H. Jia, A. Allam, R. K. Pokharel and H. Kanaya, "Strong resonant coupling for short-range wireless power transfer applications using defected ground structures," *IEEE Wireless Power Transf. Conf. (WPTC)*, pp. 1-4, 2015.
- [23] S. Hekal et al., "A novel technique for compact size wireless power transfer applications using defected ground structures," *IEEE Trans. Microw. Theory Techn.*, vol. 65, no. 2, pp. 591-599, Feb. 2017.
- [24] F. Tahar, A. Barakat, R. Saad, K. Yoshitomi and R. K. Pokharel, "Dual-Band Defected Ground Structures Wireless Power Transfer System With Independent External and Inter-Resonator Coupling," *IEEE Transactions on Circuits and Systems II: Express Briefs*, vol. 64, no. 12, pp. 1372-1376, Dec. 2017.
- [25] K. Dautov, M. Hashmi, G. Nauryzbayev and N. Nasimuddin, "Recent Advancements in Defected Ground Structure-Based Near-Field Wireless Power Transfer Systems," *IEEE Access*, vol. 8, pp. 81298-81309, 2020.

- [26] A. Barakat, K. Yoshitomi and R. K. Pokharel "Design Approach for Efficient Wireless Power Transfer Systems during Lateral Misalignment," *IEEE Trans. Microw. Theory Techn.*, vol. 66, no. 9, pp. 4170-4177, Sep. 2018.
- [27] S. Verma, D. Rano, M. Hashmi and V. Bohara, "A High Q Dual E-Shaped Defected Ground Structure for Wireless Power Transfer Applications," *Asia-Pacific Microwave Conference (APMC)*, pp. 1435-1437, 2018.
- [28] S. Verma, D. Rano and M. Hashmi, "A novel dual band defected ground structure for short range wireless power transfer applications," in *Proc. IEEE Wireless Power Transf. Conf. (WPTC)*, London, U.K., pp. 1-4, Jun. 2019.
- [29] S. Verma, D. Rano and M. Hashmi, "Enhancing the performance of defected ground structure type near-field radiofrequency WPT system by coupled-line impedance matching," *IET Microwaves, Antennas & Propagation*, vol. 14, no. 12, pp. 1431-1439, Oct. 2020.
- [30] S. Verma, D. Rano, S. Malhotra and M. S. Hashmi, "Measurements and Characterization of a Newly Developed Novel Miniature WIPT System," *IEEE Transactions on Instrumentation and Measurement*, vol. 70, no. 2004211, pp. 1-11, 2021.
- [31] D. Guha, Y. M. M. Antar, *Microstrip and Printed Antennas: New Trends, Techniques and Applications*, Dec. 2010.
- [32] M. S. Alam, N. Misran, B. Yatim and M. T. Islam, "Development of Electromagnetic Band Gap Structures in the Perspective of Microstrip Antenna Design," *International Journal of Antennas and Propagation*, vol. 2013, pp. 1-22, 2013.
- [33] A. Abdolkhani, "Fundamentals of Inductively Coupled Wireless Power Transfer Systems," *Wireless Power Transfer - Fundamentals and Technologies*. London, United Kingdom: IntechOpen, 2016.
- [34] C. Park, S. Lee, G. H. Cho and C.T. Rim, "Innovative 5-m-off-distance inductive power transfer systems with optimally shaped dipole coils," *IEEE Trans. Power Electron.*, vol. 30, pp. 817-827, Feb. 2015.
- [35] Z. H. Ye, Y. Sun, X. Dai, C. S. Tang, Z. H. Wang and Y. G. Su, "Energy efficiency analysis of U-coil wireless power transfer system" *IEEE Trans. Power Electron.*, vol. 31, pp. 4809-4817, Jul. 2016.
- [36] T.P. Duong and J. W. Lee, "Experimental results of high-efficiency resonant coupling wireless power transfer using a variable coupling method," *IEEE Microw. Wirel. Compon. Lett.*, vol. 21, pp. 442-444, 2011.
- [37] B. L. Cannon, J. F. Hoburg, D. D. Stancil and S. C. Goldstein, "Magnetic resonant coupling as a potential means for wireless power transfer to multiple small receivers," *IEEE Trans. Power Electron.*, vol. 24, pp. 1819-1825, 2009.

- [38] R. Shadid, S. Noghianian and A. Nejadpak, "A literature survey of wireless power transfer," *IEEE International Conference on Electro Information Technology (EIT)*, pp. 782–787, May 2016.
- [39] A. Christ, M. Douglas, J. Nadakuduti and N. Kuster, "Assessing Human Exposure to Electromagnetic Fields From Wireless Power Transmission Systems," in *Proceedings of the IEEE*, vol. 101, no. 6, pp. 1482-1493, June 2013.
- [40] L. Tan, X. Huang, H. Huang, Y. Zou and H. Li, "Transfer efficiency optimal control of magnetic resonance coupled system of wireless power transfer based on frequency control," *Sci. China Technol. Sci.*, vol. 54, pp. 1428–1434, 2011.
- [41] A. M. Jawad, R. Nordin, S. K. Gharghan, H. M. Jawad, and M. Ismail, "Opportunities and Challenges for Near-Field Wireless Power Transfer: A Review," *Energies*, vol. 10, no. 7, pp. 1-28, Jul. 2017.
- [42] S. Kim, B. Bae, S. Kong, D. H. Jung, J. J. Kim and J. Kim, "Design, implementation and measurement of board-to-board wireless power transfer (WPT) for low voltage applications," *IEEE 22nd Conf. Electr. Perform. Electron. Packag. Syst. EPEPS 2013*, pp. 91–95, 2013.
- [43] J. Lee, Y. S. Lim, W. J. Yang and S. O. Lim, "Wireless power transfer system adaptive to change in coil separation," *IEEE Trans. Antennas Propag.*, vol. 62, no. 2, pp. 889–897, Feb. 2014.
- [44] M. E. Celik and E. Aydin, "An Efficient Inductive Coil Link Design for Wireless Power Transfer to Visual Prostheses," *Acta Physica Polonica*, vol. 132, no. 3, pp. 535-537, 2017.
- [45] D. H. Werner and S. Ganguly, "An Overview of Fractal Antenna Engineering Research," *IEEE Antennas and Propagation Magazine*, vol. 45, no. 1, pp. 38, Feb. 2003.
- [46] Z. Wei, Y. W. Ming, F. Yin-Yin, H. C. Kang, Z. X. Yu and M. Ge, "Design of the Hilbert Fractal Defected Ground Structure Low-pass Filter," *IEEE Int. Conf. Communication Problem-Solving (ICCP)*, vol., no., pp. 1, Dec. 2014.
- [47] A. Kumar, A. Sawant and M. V. Kartikeyan, "Investigation of Fractal DGS Microwave Filters," *IEEE National Conference Communications (NCC)*, vol., no., pp. 1, Feb. 2013.
- [48] J. Beneat and P. Jarry, "Miniaturized RF/microwave filters using fractals," *International IEEE MTT-S Microwave Symposium Digest (MTT)*, vol., no., pp. 1-3, June 2012.
- [49] A. B. Abdel-Rahman, A. K. Verma, A. Boutejdar and A. S. Omar, "Control of bandstop response of Hi-Lo microstrip low-pass filter using slot in ground plane," *IEEE Trans. Microwave Theory and Tech.*, vol. 52, no. 3, pp. 1008-1013, March 2004.
- [50] D. J. Woo, T. K. Lee, J. W. Lee, C. S. Pyo and W. K. Choi, "Novel U-slot and V-slot DGSs for bandstop filter with improved Q factor," *IEEE Trans. Microwave Theory and Tech.*, vol. 54, no. 6, pp. 2840-2847, June 2006.
- [51] D. Ahn, J. S. Park, C. S. Kim, J. Kim, Y. Qian and T. Itoh, "A Design of the Low-Pass Filter Using the Novel Microstrip Defected Ground Structure," *IEEE Trans. on Microwave Theory and Tech.*, vol. 49, no. 1, Jan. 2001.

- [52] H. W. Liu, Z. F. Li and X. W. Sun, "A novel fractal defected ground structure and its application to the low-pass filter," *Microw. Opt. Technol. Lett.*, vol. 39, no. 6, pp. 453–456, Oct. 2003.
- [53] Z. Ma, K. Kikuchi, Y. Kobayashi, T. Anada and O. Hagiwara, "Novel microstrip dual-band bandstop filter with controllable dual-stopband response," in *Proc. Asia-Pacific Microw. Conf.*, pp. 1177-1180, Dec. 2007.
- [54] X. Hu, Q. Zhang and S. He, "Dual-band-rejection filter based on split ring resonator (SRR) and complimentary SRR," *Microw. Opt. Technol. Lett.*, vol. 51, no. 10, Oct. 2009.
- [55] V. K. Velidi and S. Sanyal, "Compact planar dual-wideband bandstop filters with cross coupling and open-ended stepped impedance resonators," *ETRI Journal*, vol. 32, no.1, Feb. 2010.
- [56] S. Y. Huang and Y. H. Lee, "A compact E-shaped patterned ground structure and its applications to tunable bandstop resonator," *IEEE Trans. Microw. Theory Tech.*, vol. 57, Iss. 3, pp. 657–666, Mar. 2009.
- [57] M. R. Saad et al., "Analysis of near field wireless power transfer using bow-tie defected ground structure," *IEEE Asia Pacific Microwave Conference (APMC)*, pp. 495-498, Nov. 2017.
- [58] S. Hekal et al., "Compact wireless power transfer system using defected ground bandstop filters," *IEEE Microw. Wireless Compon. Lett.*, vol. 26, no. 10, pp. 849–851, Oct. 2016.
- [59] A. Barakat, K. Yoshitomi and R. K. Pokharel, "Design and Implementation of Dual-Mode Inductors for Dual-Band Wireless Power Transfer Systems," *IEEE Transactions on Circuits and Systems II: Express Briefs*, vol., pp. 1-5, Nov. 2018.
- [60] T. Jensen et al., "Coupled transmission lines as impedance transformer," *IEEE Trans. Microw. Theory Tech.*, 55, (12), pp. 2957–2965, 2007.
- [61] F. W. Grover, *Inductance Calculations: Working Formulas and Tables*. New York, NY, USA: Courier Corporation, 1946.
- [62] I. Bahl, M. Bozzi and R. Garg, *Microstrip Lines and Slotlines, Third Edition*, Artech, 2013.
- [63] S. Malhotra and M. Hashmi, "Near-Field WPT using Defected Ground Structures for UHF RFID Applications," *IEEE International Conference on RFID Technology and Applications (RFID-TA)*, pp. 16-21, 2019.
- [64] D. Ahn and P. P. Mercier, "Wireless power transfer with concurrent 200-kHz and 6.78-MHz operation in a single-transmitter device," *IEEE Trans. Power Electron.*, vol. 31, no. 7, pp. 5018–5029, Jul. 2016.
- [65] D. H. Weon and S. Mohammadi, "High performance 3-D helical RF transformers," in *IEEE MTT-S Int. Microw. Symp. Dig.*, pp. 1897–1900, Jun. 2007.
- [66] R. Sharaf, A. B. Abdel-Rahman, A. S. Abd El-Hameed, A. Barakat, S. Hekal and A. Allam, "A New Compact Dual-Band Wireless Power Transfer System Using Interlaced Resonators," *IEEE Microwave and Wireless Components Letters*, vol. 29, no. 7, pp. 498-500, July 2019.