

Accurate Human Tissue Modeling and Link Analysis for Implantable Antenna Applications

Thesis Submitted By Soham Ghosh
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1. Title of the Thesis:

Accurate Human Tissue Modeling and Link Analysis for Implantable Antenna Applications

2. Name, Designations & Institution of the Supervisor:

Dr. Bhaskar Gupta
Ex. Vice Chancellor, Jadavpur University,
Former Professor, Department of ETCE,
Jadavpur University, Kolkata- 700032, India

3. List of Publications:

Kindly refer to **Appendix I**

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5. List of Presentations in National/ International Conferences/Workshops:

Kindly refer to **Appendix II**

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I, **Soham Ghosh**, registered on **06.02.2023** do hereby declare that this thesis entitled "**Accurate Human Tissue Modeling and Link Analysis for Implantable Antenna Applications**" contains literature survey and original research work done by the undersigned candidate as part of Doctoral studies.

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Signature of Candidate: *Soham Ghosh*

Date: *04.06.2026*

Certified by Supervisor:

Gupta *04/06/26*

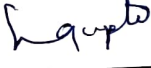
Dr. Bhaskar Gupta
Ex. Vice Chancellor, Jadavpur University,
Former Professor, Department of ETCE
Jadavpur University, Kolkata- 700032, India



Prof. Bhaskar Gupta
Professor (*Retd.*)
Dept. of E. & T.C. Engg.
Jadavpur University
Kolkata

CERTIFICATE FROM THE SUPERVISOR

This is to certify that the thesis entitled "**Accurate Human Tissue Modeling and Link Analysis for Implantable Antenna Applications**" submitted by **Shri Soham Ghosh**, who got his name registered on **06.02.2023** for the award of Ph. D. (Engineering) degree of Jadavpur University is absolutely based upon his own work under the supervision of **Dr. Bhaskar Gupta**, Ex. Vice Chancellor of Jadavpur University, and Former Professor at Dept. of ETCE, Jadavpur University, Kolkata-700032, West Bengal, India and that neither his thesis nor any part of the thesis has been submitted for any degree/diploma or any other academic award anywhere before.


04/06/26
Dr. Bhaskar Gupta
Ex. Vice Chancellor, Jadavpur University,
Former Professor, Department of ETCE
Jadavpur University, Kolkata- 700032, India



Prof. Bhaskar Gupta
Professor (Retd.)
Dept. of E. & T.C. Engg.
Jadavpur University
Kolkata

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Soham Ghosh

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Abstract

Over the last two decades, the popularity of medical bio-telemetry devices has increased interest in therapeutic and diagnostic features for continuous patient monitoring in applications such as temperature, heartbeat, glucose level detection, heart rate management, and so on. Implantable antennas integrated in these medical devices can wirelessly communicate patients' health-related sensed data to the doctor's end, saving them from sudden death or severe condition, thanks to the growing Internet of Medical Things (IoMT) technology.

This thesis presents an examination of a meander-line implantable antenna put within a three-layer human body model, utilizing the Transmission Line Model. The developed antenna is then inserted within an implantable device, together with a printed circuit board, sensor, battery, and other components, and is coated with biocompatible polyvinyl alcohol. For in-vitro assessment of antenna performance, a comparable human body model of a multilayered model is created using the Transmission Matrix Method (TMM). TMM is used to estimate the effective complex permittivities of a three-layer heart model and a seven-layer brain model, with an accuracy of 98.46%.

The study examines the impact of $\pm 20\%$ variations in effective relative permittivity and conductivity on the performance of an implantable antenna system, including return loss (operating frequency and S_{11}), radiation (realized gain, fractional bandwidth), link performance (link margin), and safety measures (SAR for 1g and 10g body models). In the next phase, a tiny dual-band implantable antenna (volume of 21.53 mm^3) is adapted for C-Band 5G biotelemetry in Wireless Capsule Endoscopy (WCE) with higher gain, -10 dB impedance bandwidth, compactness, and particular absorption rate responses than previously published studies. This thesis discusses wireless far-field powering for

a leadless cardiac Transcatheter Pacing (TCP) system using a dual-wideband (1.3 GHz and 2.45 GHz) circular polarized tiny implantable antenna. In terms of performance characteristics, the proposed system outperforms leadless cardiac pacemakers that have been previously reported.

Soham Ghosh
Soham Ghosh

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Chapter 1

Introduction

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1.1 PREFACE

In the last two decades, the popularity of medical bio-telemetry devices has escalated interest in therapeutic and diagnostic functions for continuous patient monitoring in applications like temperature, heartbeat, glucose level detection, heart rate regulation, drug delivery and pain control [1-2]. Implantable antennas embedded in these medical devices can transfer the patients' health-related sensed data to the doctor's end wirelessly to save them from a sudden death or a critical condition with the help of the emerging Internet of Medical Things (IoMT) technology [3]. Erroneous communication link characterization may cause data loss, which leads to the misinterpretation of the patient's health condition and wrong diagnosis. The quality of communication (or link margin), includes the gains of the implantable antenna and the external monitoring device's antenna, operating frequency, transmitted power, path loss, polarization of the antennas, data rate, and signal-to-noise ratio (SNR) [4].

Implantable antenna as a part of Implantable Medical Devices (IMDs) stems from early 1980s when L. S. Taylor [5] designed a S-band microwave syringe to implant in the human brain for hyperthermia of cancer issue. In 1995, Tang *et. al.* [6] introduced Load-Shift Keying (LSK) for efficient data

transmission from implantable antenna over other widely used modulation schemes such as Frequency Modulation (FM), Amplitude Shift Keying (ASK), Phase Shift Keying (PSK) and Frequency Shift Keying (FSK) [7-9]. Now-a-days, the researchers are also using Binary PSK (BPSK), Quadrature PSK (QPSK), Offset QPSK (OQPSK) etc. [10-12] for Wireless Body Area Network (WBAN). Implantable antennas used in Wireless Capsule Endoscopy (WCE) systems, Leadless Cardiac Pacemakers (LCPs), head implant devices for intracranial pressure monitoring, stroke prediction etc., blood glucose monitors etc., improves the communication quality for wireless smart patient monitoring.

1.2 IMPLANTABLE ANTENNAS: WHAT AND WHY

Semantically, “implantable antennas” indicate a class of antennas which can provide a facility of transferring health parameters wirelessly to the external environment from IMDs such as biosensors, pacemaker, glucose monitors, gastric stimulators, defibrillators, etc. [13-18]. The implantable antennas embed in implantable medical devices can wirelessly communicate with external monitoring devices to provide continuous medical records of patient’s health parameters. Wireless Body Area Network (WBAN) can also be utilized for assisting police, fire fighters etc., monitoring pulse rate of sportsmen and providing navigation support within car or at the time of walking [19-21].

A number of frequency bands such as Medical Implant Communication System (MICS), Wireless Medical Telemetry Service (WMTS) and Industrial Scientific and Medical (ISM) band have been allotted to design and operate implantable antennas [22-23]. To design an implantable antenna in above mentioned allotted frequency bands, physical dimensions of the antenna come out to be significantly large. However for proper implantation within human body, the size of the antenna should be reduced at high amount. Different miniaturization techniques such as meandering, spiraling, loading on patch, incorporation of slots

etc., are widely used in previously reported works [24-25]. The critical factors required to consider designing implantable antenna are high lossy body tissues, miniaturization of antenna, low input RF power restriction, Specific Absorption Rate (SAR) etc. [26-27]. Biocompatible encapsulation is performed to reduce corrosion of patch with high-permittivity-human-tissues for patients' safety as well as improvement of communication quality [28-29]. As per some international standard guidelines [30-31], SAR values should be <1.6 W/kg and <2 W/kg for 1-g and 10-g average tissue samples respectively.

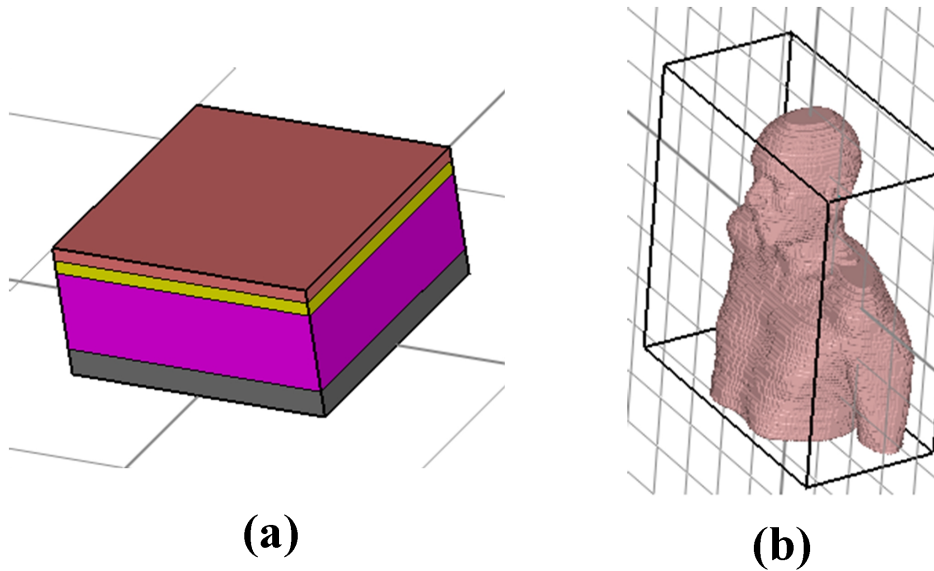


Fig. 1.1: (a) Simple multi-layer Human Body model and (b) Gustav voxel model in CST Microwave Studio

1.3 ANALYSIS EFFORTS

Before surgical implantation of implantable devices within patient's body, proper *in-vitro* testing is required. The testing should be performed within a medium which can provide similar effect as human body. This artificially made

medium is called “Phantom” [32-33]. Different biological phantoms are utilized for different purposes. Dosimetry phantoms are used in radiotherapy. Besides them, phantoms can be utilized in photo-acoustic imaging, breast imaging, image reconstruction etc. In medical imaging, biological phantoms take crucial role [34]. Breast phantom with high fidelity is required for microwave breast tumor and cancer detection system [35].

Generally, full wave techniques are used by scientists to analyze the implantable antenna performances in simulation by placing them within simple human model or within popular voxel models (such as Gustav, Laura, Hugo, Child, Baby, Donna, Emma etc.) as shown in Fig. 1.1. The analysis within realistic voxel model is very time consuming and storage capacity of analysis system is required to be very high; whereas within multilayer model, the time and storage requirements are comparatively very low. However, some amount of error should be considered for model shown in Fig. 1.1 (a) with respect to the model of Fig. 1.1 (b). For practical verification, two approaches can be considered for testing of different implantable systems – (i) *in-vitro* and (ii) *in-vivo*. In *in-vitro* technique, handmade or readymade phantom models are taken as an equivalent model of human body. It is very difficult to develop multilayer model practically. Therefore, homogeneous human phantoms are widely considered for practical analysis. For replicating approximate human body properties, minced pork is widely used for *in-vivo* analysis.

1.4 MOTIVATION OF THE THESIS

Recent advancement of medical bio-telemetry devices is escalating interest in therapeutic and diagnostic functions for continuous patient monitoring. Different types of miniaturized implantable antennas were proposed by many researchers. Proper link establishment between implanted antenna and external

monitoring device is important. Erroneous communication channel may cause data loss which leads to the misinterpretation of the patient's health condition and wrong diagnosis. The quality of communication is coined by the link margin which includes gain of the implantable antenna and the antenna of external monitoring device, operating frequency, transmitted power, path loss related to the distance of communication, polarization of the antennas, data rate, signal-to-noise ratio (SNR) etc. [37].

As wireless communication link of the implantable antenna with monitoring device is established through the human tissue, this communication performance depends on the bioelectric properties of tissue such as relative permittivity and conductivity. Relative permittivities of human tissues are dependent on some factors like age, gender of patients, location of implant, thickness of skin and fat layers, hormonal variations in patients' bodies etc. [38-43]. The relative permittivity of the tissue is dependent on water and oil content, structure of fibers etc. 0.4-0.6 mm thick stratum corneum (a part of the epidermis of the skin layer) contains comparatively lower water content (around 10%) [44]. Hwang et al [45], Boric-Lubecke et al [46] showed that the effective complex permittivity of different body parts is not the same. The generation of tumors in any layer may increase effective complex permittivity by about 20%. The effective dielectric properties can also change due to pathological changes in liver due to cirrhosis, steatosis, hepatocellular carcinoma etc. [47]. Therefore, it is required to study the variation in implantable antenna performances due to possible variations in body-layer parameters in terms of thicknesses, relative permittivities and conductivities of different layers.

In this context, variations in radiation, link and safety performances of implantable antenna in terms of operating frequency, realized gain, fractional bandwidth, link margin, range of communication, SAR profiles for 1g and 10g contiguous tissue models are recorded for considering $\pm 20\%$ possible variations in

different body parameters. Statistical analyses are performed here for analyzing their dependences. Finally an Artificial Neural Network is designed for predicting antenna performance parameters for any certain body parametric values. For practical verification of antenna performances, it is important to develop human body mimicking model. Development of multilayer body model is very difficult to fabricate. Therefore, equivalent homogeneous body model is required to develop. For this, accurate equivalent relative permittivity and conductivity of specific body model should be calculated. Generally trial-and-error method was followed by researchers. In this thesis work, an accurate estimation method of equivalent (mean) relative permittivity and conductivity for any human body models (for e.g. brain model, heart model, generalized three-layer body model etc.) using computation electromagnetics based Transmission Matrix Method (TMM) is proposed.

Now-a-days, 5G communication technology is widely introduced in different fields (like mobile communication, Internet of Things, etc.) due to several advantages such as low latency, fast speed, low-noise data transfer etc. However, 5G facility is not widely introduced in the medical communication to the best of knowledge. Therefore, it will be a major concern of this work. Recently, uses of implantable antennas in different applications such as WCE, LCP etc., are broadly focused by scientists. Therefore, betterment of implantable antenna performances in these biomedical applications is also required for improving the quality of smart patient monitoring.

1.5 NOVEL ASPECTS

Following are the novel aspects aimed at this thesis:

- Analysis of meander-line human body-implantable antenna using Transmission Line Model (TLM), one of the popular analytical techniques. Compare its responses with full wave analysis.
- Development of an estimation technique of effective complex permittivity of human body model to fabricate homogeneous equivalent phantom model accurately for in-vitro measurement of implantable antenna.
- Dependence analyses of the performances of a body-implantable antenna system prototype consisting of antenna, battery, sensor, data management system, etc., encapsulated with a biocompatible material in terms of return loss performances (operating frequency and S_{11}), radiation performances (gain and bandwidth), SAR performances and communication performances (link margin) on effective relative permittivity and effective conductivity variations ($\pm 20\%$ of their mean values at 2.45 GHz).
- Development on different novel antenna systems with several facilities such as wide-bandwidth (to reduce the possibility of detuning due to high lossy nature of human tissues and body movement), circular polarization (to reduce polarization mismatch loss, and its suitability for operating in various body movements) etc., for biomedical applications such as wireless capsule endoscopy, leadless cardiac transcatheter pacing system etc.
- Wireless Power Transfer system with good power transfer efficiency for wireless charging of designed implantable antenna system.
- Introduction of 5G facility in the wireless data transfer from the designed implantable antenna system which is not well-known in this field.

1.6 THESIS OUTLINE

Chapter 1 provides fundamental overview of the operation of implantable antennas in human body model as well as a brief historical background. Next chapters comprise the remaining portion of this dissertation as follows:

Chapter 2 is divided in four parts. The first part deals with a brief review of different types of implantable antennas designed in last decade. Then a brief discussion on modeling techniques of implantable antennas used over the years is presented in next part. The third part of the chapter comprises of the use of implantable antennas in different medical applications such as wireless capsule endoscopy, glucose monitoring, brain implant, wireless power transfer etc. Characterization of equivalent electrical properties of human body phantom in terms of effective relative permittivity and conductivity is very important for phantom development to pursue *in-vitro* testing of implantable devices. The available dielectric measurement techniques of human body phantoms are discussed in last part of Chapter 2.

Chapter 3 presents the analysis of meander-line implantable antenna placed within a three-layer human body model using Transmission Line Model (TLM). Modeling of bended and corner chamfered implantable antenna using this technique has also been performed in this chapter. Chapter 4 deals with the accurate estimation of effective relative permittivity and conductivity of three human body models such brain model, heart model and three-layer body widely used model. Here, TMM is used for calculate properly the effective electrical properties to develop accurate equivalent human body phantom corresponding to multi-layer models for proper *in-vitro* measurement of implantable antennas.

Variations in effective electrical properties of human body can be happened due to variation in age, atmosphere, location of patient, hormonal variations in human body. These variations may lead to vary the radiation and

communication performances of any implanted antenna. The analyses of possible variations in antenna performances due to effective relative permittivity and conductivity are taken up in Chapter 5. The variation analyses of the different performances of an implantable antenna system such as return loss characteristics (operating frequency and S_{11}), radiation characteristics (realized gain, fractional bandwidth), link performances (link margin) and safety measures (SAR for 1g and 10g body models) on $\pm 20\%$ variations in effective relative permittivity and conductivity are performed here.

The Chapter 6 introduces a miniaturized dual-band implantable antenna (volume of 21.53 mm^3) to adapt C-Band 5G biotelemetry in Wireless Capsule Endoscopy (WCE). Some open-ended slots were engraved on patch to tune the antenna properly. The antenna along with the battery, PCB, camera, data-management system was placed within a biocompatible polyvinyl alcohol capsule. The antenna system was fabricated and placed within homogeneous human body model to verify its performance through in-vitro measurements. Proposed system shows the known best performance reported on wireless capsule endoscopic systems. Introduction of 5G facility in WCE biotelemetry is also a novel approach in this application to improve the quality of smart patient monitoring.

Wireless far-field powering to a dual-wideband (1.3 GHz and 2.45 GHz) circular polarized miniaturized implantable antenna designed for biocompatible polyvinyl alcohol-coated leadless cardiac Transcatheter Pacing (TCP) system is discussed in Chapter 7. An external antenna is designed for wireless power transfer at 1.3 GHz and biotelemetry at 2.45 GHz with TCP system. The antennas are fabricated and measured to verify the simulated responses. Proposed system shows the known best performance reported on leadless cardiac pacemakers with respect to the performance parameters.

In the last chapter, conclusions are drawn with brief discussion on future scopes in the domain of performance analyses of implantable antennas.

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Chapter 2

Literature Review

Contents	2.1. Introduction 2.2. Reported Implantable Antennas: Types 2.3. Modeling Techniques of Implantable Antennas 2.4. Applications of Implantable Antennas 2.5. Effective Complex Permittivity Measurement For Accurate Phantom Development 2.6. Conclusion
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2.1 INTRODUCTION

Recent investigations on worldwide medical implants market demonstrate that several kinds of implants such as cardiovascular, neuro-stimulators, spinal, ophthalmic implants are currently on the rise [1-4]. A market analysis report [5] shows that implantable medical devices market size was expected at USD 97.62 billion in 2024 and it is projected to reach around USD 178.15 billion at a compound annual growth rate (CAGR) of 6.2% by 2034. Technological advancements in the medical industry have enhanced the use of medical implants such as breast implants, dental implants, deltoid implants, cochlear implants etc [6]. Over past few decades, a rapid increase is observed in the territory of contemporary medicine due to swift progress with complex electronic devices such as implantable and wearable medical devices, captivating a principal position in medical field.

Through wireless communication, Implantable Medical Devices (IMDs), one of the most widely used biotelemedical systems, may send physiological signals picked up by biosensors and transducers to external monitoring equipment. It is termed as Wireless Body Area Network (WBAN). General architecture of

WBAN is shown in Fig. 2.1. For establishing wireless links between different implantable and wearable sensors (between two implanted sensors, between implanted sensor and link sensor, between in-body sensor and on-body sensor etc.) and link between implanted sensor and exterior monitoring devices, antenna embedded in IMDs are taking vital role.

The communication of implantable antenna is depending on operating frequency of the implant, depth of implantation and design architecture of antenna etc. Previously, for powering pacemakers and data transmission, inductive coupling was used for the implants. However, it is limited by size and quality of the implantable coil and even at low frequencies (typically 1 MHz) where the near-field is large, the implantation depth is limited to just a centimeter [7]. Inverted F, planar loop antennas, spirals, monopoles, dipoles, meanders etc. were designed as implants due to their low profile nature [8-11].

2.2 REPORTED IMPLANTABLE ANTENNAS: TYPES

In the last two decades, the popularity of medical bio-telemetry devices has escalated interest in therapeutic and diagnostic functions for continuous patient monitoring in different purposes like temperature, heartbeat, glucose level detection, heart rate regulation, drug delivery and pain control. Implantable Antennas can be operated at different frequency ranges such as Medical Device Radio band (MedRad) (401-406, 413-419, 426-432, 438-444, 451-457 MHz, Medical Implantable Communication Service (MICS) (402-405 MHz) or Industrial, Scientific and Medical (ISM) (2.45 GHz) band frequencies [12]. Popular applications of implantable devices are – (i) monitoring biophysical parameters such as blood sugar, pressure etc., (ii) pacemakers, glucose sensors, defibrillators, (iii) capsule pills, cochlear and retinal implants, Functional Electrical Stimulators (FES), (iv) hypopnea syndrome diagnosis and chronic obstructive pulmonary disease warning etc. [13-17]. Different types of

implantable antennas were reported by different scientists depending on applications in last decade. Key categories of them are discussed below.

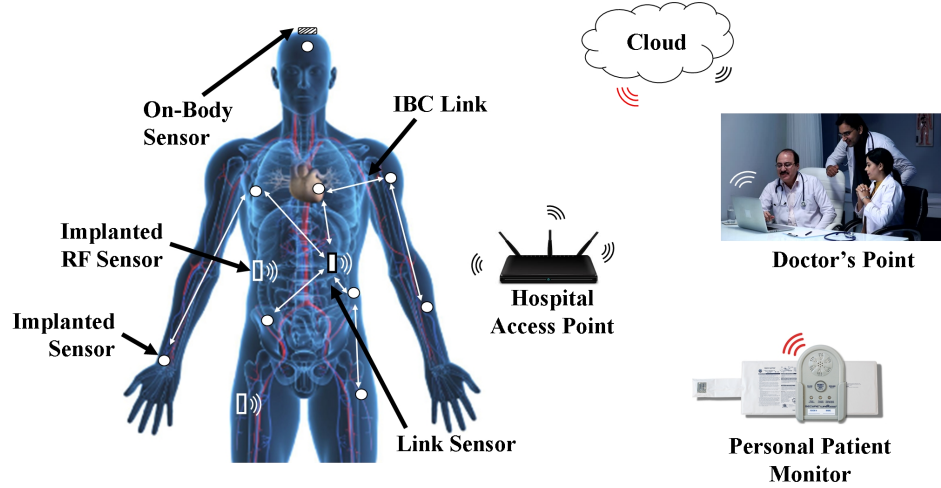


Fig. 2.1: General Architecture of Wireless Body Area Network for Wireless Patient Monitoring

2.2.1: Planar Implantable Antennas:

A capacitive loaded Circularly Polarized (CP) implantable antenna with dimensions $10 \times 10 \times 1.27 \text{ mm}^3$ using Rogers 3010 substrate with dielectric constant (ϵ_r) of 10.2, a thickness of 25 mil and loss tangent ($\tan \delta$) of 0.003 was proposed in [18] as shown in Fig. 2.2 (a). This antenna was simulated and measured within $90 \times 90 \times 25.27 \text{ mm}^3$ skin phantom model ($\epsilon_r = 38$ and conductivity (σ) = 1.44 S/m at 2.45 GHz). The antenna had 10 dB impedance bandwidths of 7.74% and 10.2% in simulation and measurement respectively. The antenna had 1.63 % Axial Ratio (AR) bandwidth for AR < 3dB. In [19], a parylene-C coated differentially fed dual-band linearly polarized implantable antenna (refer to Fig. 2.2 (b)) operating at 402–405 MHz Medical Implant Communication Services (MICS) band and 2.4–2.5 GHz Industrial, Scientific, and Medical (ISM) band within a three-layer human body model (skin-fat-muscle)

was proposed. 20 μm thick parylene-C coated compact dual-band implantable antenna ($10.02 \times 10.02 \times 0.675 \text{ mm}^3$) shown in Fig. 2.2 (c) operating at MICS and 2.45 GHz ISM band with 47.5% and 31.6% respective fractional bandwidths were proposed in [20].

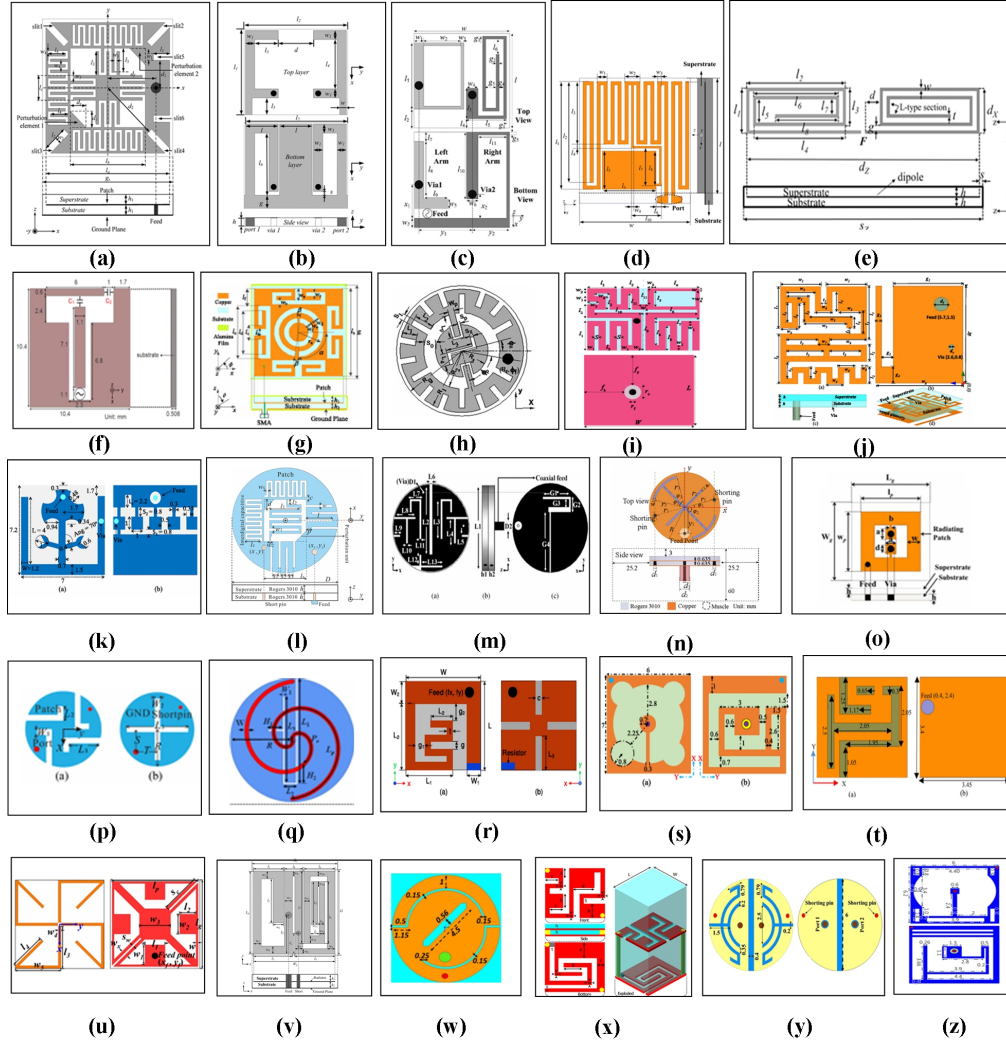


Fig. 2.2: Planer implantable antennas reported in [18]-[44] respectively ©IEEE

A linearly polarized 329 mm^3 meander line implantable dipole antenna within skin phantom operating at 402 MHz band shown in Fig. 2.2 (d) was

proposed in [21]. The antenna radiated over 37.8 % fractional bandwidth with -35 dBi realized gain. In [22], a meander-line based $12.5 \times 12.5 \times 1.27 \text{ mm}^3$ implantable antenna tuned at 402 MHz Med-Radio band within three-layer human body model operating with 6.15 % fractional bandwidth and -32.49 dBi gain was proposed. An L-shape implantable planer dipole antenna (volume 18 mm^3) with linear polarization was designed in [23], which is shown in Fig. 2.2 (e). $10.4 \times 10.4 \times 0.508 \text{ mm}^3$ square-shaped Left Handed CP (LHCP) implantable antenna shown in Fig. 2.2 (f) was designed in [24]. It was tuned at 2.45 GHz with 25.3% 10 dB impedance bandwidth and 10.2% AR bandwidth (for $\text{AR} < 3 \text{ dB}$). Complementary Split Ring Resonator (CSRR) loaded linear polarized miniaturized implantable antenna was designed in [25] to monitor blood glucose level at 2.45 GHz which is shown in Fig. 2.2 (g). A 120.69 mm^3 CP annular ring implantable antenna illustrated in Fig. 2.2 (h), was proposed in [26]. The antenna had 3-dB AR bandwidth of 2.49% and -10 dB return loss bandwidth of 11.4%.

Table 2.1: Comparative Analysis of Reported Planer Implantable Antennas

Ref.	Volume (mm^3)	Freq. (GHz)	Fractional Bandwidth (%)	Realized Gain (dBi)	SAR (1g) for 1W (W/kg)
[18]	127	2.45	7.7	-22	213
[19]	186.3	0.402	49.5	-30.6	-
		2.45	24.1	-22.2	
[20]	67.8	0.402	47.5	-30.5	302.4
		2.45	31.6	-19.5	238.9
[21]	329	0.348	37.8	-35	463
[22]	198.4	0.402	7.26	-32.49	192.59
[23]	18.1	0.402	26.02	-29.4	426.5

[24]	54.9	2.45	25.3	-21.1	356.4
[25]	92	2.45	12.2	-17	-
[26]	120.69	2.45	8.3	-22.7	733.5
[27]	24	0.915	9.84	-28.5	911.56
		2.45	8.57	-22.8	807.34
[28]	21	0.915	8.7	-26.4	380
		1.9	8.2	-23	358
		2.45	7.3	-20.47	363
[29]	10.08	0.928	19.83	-28.74	471
		2.45	8.97	-25.65	313
[30]	120	2.45	20.5	-26.4	712.1
[31]	797.96	0.402	38.1	-34.08	241.5
		2.45	17.6	-15.2	199.5
[32]	91.9	2.45	18.3	-20.3	649
[33]	280.04	2.41	8.5	-22.93	347.08
[34]	63.80	2.25	16.32	-37.36	856.45
[35]	197.4	0.402	23.13	-33.3	241
		1.430	14.13	-21.9	269
		2.450	18.12	-19.6	290
[36]	8	2.4	16.6	-18.3	363
[37]	10.66	0.79- 3.83	217	-20.02	255.52
[38]	4.42	2.45	10.61	-26.54	141
[39]	85.38	2.45	30	-24.7	697.5
[40]	691.52	0.402	12.6	-31.5	318
		2.45	5.7		292
[41]	42.70	1.4	7.2	-29.4	84.8
		2.45	4.2	-30.4	123.7

[42]	0.7	0.915	16.3	-27.4	573
		1.45	8.7	-22.5	530.7
		1.9	5	-21.7	473.8
		2.45	5.4	-21.2	466.7
[43]	7.17	2.45	6.93	-24.4	274.6
[44]	9.44	1.4	134	-27.94	378.1

A 24 mm³ scalp implantable planar dual-band antenna was proposed to radiate at 915 and 2450 MHz with respective realized gains of -28.5 and -22.8 dBi by Shah et. al [27]. A triple-band planer meander-line implantable antenna placed within capsule was proposed by Zada et. al [28]. which is shown in Fig. 2.2 (j). In [29], novel flower-shaped defected grounded (Fig. 2.2 (k)) dual-band implantable antenna was proposed. It radiated with -28.44 and -25.65 dBi gain at 928 MHz and 2.45 GHz respectively. A compact wideband meander-based circular shaped implantable antenna shown in Fig. 2.2 (l) was proposed by Cui et. al [30]. A slotted circular body-implantable antenna operating at 402 MHz and 2.45 GHz was designed in [31]. Another slotted circular implantable antenna shown in Fig. 2.2 (n), was proposed in [32]. Miniaturized 2.45 GHz ISM band-tuned implantable antenna was proposed in [33].

An open-slotted wideband CP scalp implantable antenna was proposed by Zhang et. al [34] for Health Monitor Microsystem which is depicted in Fig. 2.2 (p). The antenna was operated at 2.45 GHz with realized gain of -37.36 dBi and fractional bandwidth of 16.33%. It had 3-dB AR bandwidth of 22%. A spiral implantable antenna printed on Taconic RF-10 substrate was proposed by Trong et. al [35] to radiate at Medical Implant Communications Service (MICS) band (402–405 MHz), Wireless Medical Telemetry Service (WMTS) band (1427–1432MHz) and Industrial, Scientific, and Medical (ISM) band (2400–2480MHz). An ultra-miniature implantable antenna shown in Fig. 2.2 (r) was proposed in [36]

which had fractional bandwidth of 16.6% around 2.45 GHz and gain of -18.3 dBi. An ultra-wideband (217.14%) ultra miniaturized (10.66 mm^3) implantable antenna for Cardiac Leadless Pacemaker was proposed in [37]. An electrically small implantable antenna for scalp and deep tissue biomedical devices was proposed in [38].

Compact broadband CP implantable antenna was designed in [39]. The designed implanted antenna had 10 dB impedance bandwidth of 30% and 3-dB AR bandwidth of 28.7%. A $16.5 \times 16.5 \times 2.54 \text{ mm}^3$ implantable antenna shown in Fig. 2.2 (v) was designed in [40] to operate at MICS and 2.45 GHz ISM bands. For wireless capsule endoscopy, an ultra-miniaturized dual-band planer implantable antenna was proposed in [41]. The proposed antenna radiated at 1.4 and 2.45 GHz with respective gain values of -29.4 and -30.4 dBi. A multiband implantable antenna exhibiting gain values of -27.45, -23.25, -23.14, and -22.48 dBi at 915, 1450, 1900, and 2450 MHz respectively, was presented in [42] for efficient communication in the Internet of Medical Things (IoMT) devices. A full-duplex implantable antenna for capsule endoscopy was proposed by Alshammari et. al [43]. A miniaturized implantable antenna (volume of 9.44 mm^3) designed to achieve ultra-wideband operation for next-generation leadless cardiac pacemakers (LCPs) in [44]. The antenna (Fig. 2.2 (z)) covered different frequency bands from 0.76 to 4.15 GHz, including Industrial, Scientific, and Medical (ISM) (0.869, 0.915, and 2.45 GHz), Wireless Medical Telemetry Service (WMTS) (1.4 GHz), and midfield (1.6 GHz) bands. The results of a comparative study of reported planer antennas are tabulated in Table 2.1.

2.2.2: Conformal Implantable Antennas:

Implantable antennas should be light weight, low profile and robust in nature against the implant internal components. Conformal nature is suitable for such applications and saves internal space for other components to reduce the implant size [45]. Different conformal implantable antennas were proposed in

[45-58] whose geometries are shown in Fig. 2.3. A comparative performance analysis of reported implantable antennas is shown in Table 2.2.

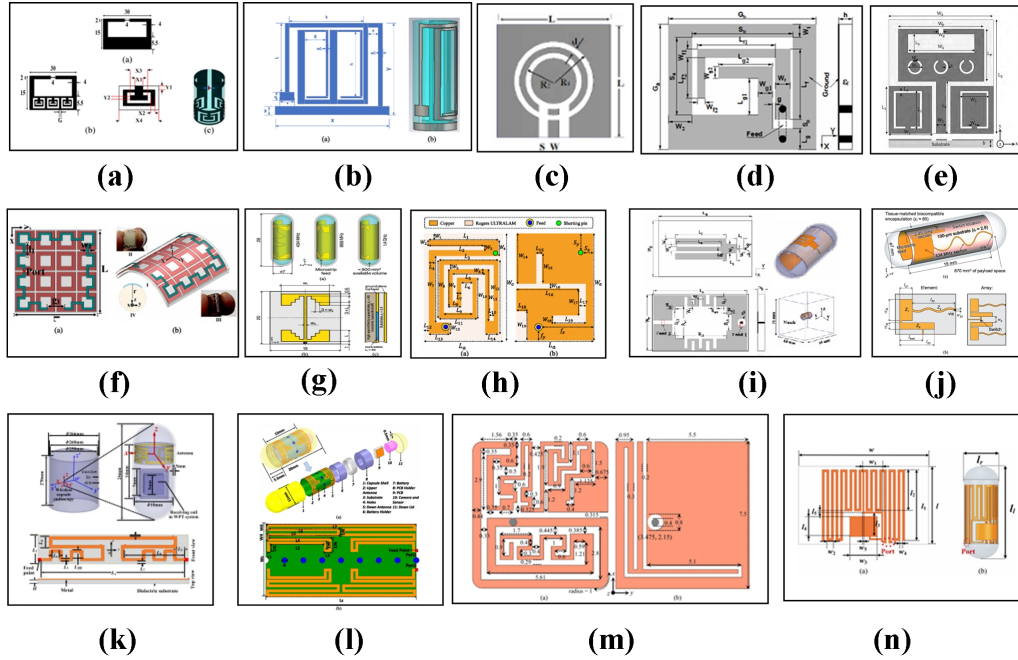


Fig. 2.3: Different conformal implantable antennas reported in [45]-[58] respectively ©IEEE

Table 2.2: Comparative Analysis of Reported Conformal Implantable Antennas

Ref.	Freq. (GHz)	Fractional Bandwidth (%)	Realized Gain (dBi)	SAR (1g) for 1W (W/kg)
[45]	0.403	12.4	-36	189
	2.45	77.53	-19.8	-
[46]	0.402	134.2	-31.5	930
[47]	2.45	57	-9	131
[48]	2.45	31.8	-29.1	368.7
[49]	2.45	24.49	-19.7	-

[50]	0.92	123	-29.33	812
	2.45		-21	680
[51]	0.434	9	-28	34.16
	0.868	10	-16	34.40
	1.4	9	-16.1	36.76
[52]	0.402	38.8	-30.8	289
	0.915	19.6	-19.7	263.5
	1.2	8.1	-18.7	214.9
[53]	4	315	-23.8	331.3
[54]	0.434	10	-28.9	26.7
	2.45	22	-18.6	58.7
[55]	0.430	11.67	-29.7	-
	0.915	9.29	-24.9	278.73
	2.45	19.88	-23.2	-
[56]	0.61-1.51	84.91	-30.4	-
[57]	0.433	9.01	-34.6	131.20
	0.915	8.19	-24.8	216.05
[58]	0.433	65.6	-32.8	-

A Complementary Split Ring Resonator (CSRR) based conformal implantable loop antenna was designed in [45] to radiate MedRadio and ISM bands. A 15 X 15 X 0.79 mm³ flexible implantable capsule antenna for wireless capsule endoscopy shown in Fig. 2.3 (b), was proposed in [46]. A wideband compact slot-based flexible implantable antenna was proposed by Das et. al [47]. In [48], a conformal CP SRR shaped ingestible capsule implantable antenna was designed which is depicted in Fig. 2.3 (d). This antenna was radiating with 31.58% 10dB impedance bandwidth and 13.11% 3dB AR bandwidth. Design and analysis of implantable conformal CPW-fed patch (ICCP) antenna operating at 2.45 GHz with 24.49% fractional bandwidth and -19.6 dBi gain was performed in

[49]. A Reactive Impedance Surface loaded dual band flexible implantable antenna was designed in [50] within skin mimicking model. A conformal implantable antenna printed on biocompatible ceramic material shown in Fig. 2.3 (g) was proposed in [51] for versatile biomedical applications.

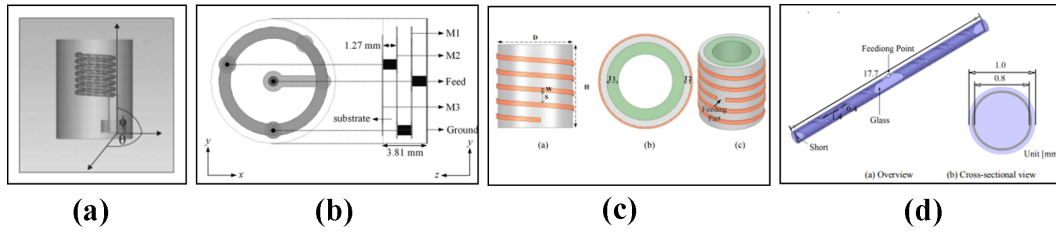
A compact conformal implantable antenna for ingestible capsule endoscope which radiated at five bands, including the MedRadio (401-406 MHz), Industrial, Scientific, and Medical (ISM) (433.1-434.8 MHz, 868-868.6 MHz, 902-928MHz) and midfield (1200 MHz) bands, was proposed in [52]. Dual-polarized ultra-wideband conformal implantable antenna was proposed for wireless capsule endoscopic system in [53]. A dual-band (434 MHz and 2.45 GHz ISM bands) reconfigurable conformal capsule implantable antenna array was proposed by Nikolayev et. al [54]. A 915 MHz tuned loop antenna conforming to the inner wall of the capsule was presented in [55] by adopting the meander and symmetrical structure patch. Conformal Multiple-Input-Multiple-Output (MIMO) ultra-wideband (84.91%) implantable antenna for wireless telemetry in capsule endoscopy was designed in [56]. A 7.5 X 7.15 X 0.75 mm³ meander-line conformal implantable antenna operating at 433 and 915 MHz was proposed in [57]. An ingestible capsule meander-line antenna (Fig. 2.3 (n)) was proposed by Gao et. al [58] for wireless biotelemetry at 400-684 MHz.

2.2.3: Helical Implantable Antennas:

A pseudo-normal-mode helical antenna was designed in [59] as part of deeply implanted wireless sensors by Murphy et. al. The antenna is shown in Fig. 2.4 (a). A multilayer-helical antenna for wireless capsule endoscopy was proposed by Liu et. al [60]. In 2022, a miniaturized (8.2 X 8.2 X 1 mm³) helical implantable antenna operating at MICS band was proposed in [61]. A helical folded dipole implantable antenna for operation at 2.45 GHz was designed in [62]. Table 2.3 shows the comparative analysis of reported helical implantable antennas [59-62].

Table 2.3: Comparative Analysis of Reported Helical Implantable Antennas

Ref.	Freq. (GHz)	Fractional Bandwidth (%)	Realized Gain (dBi)	SAR (1g) for 1W (W/kg)
[59]	0.868	2.32	-	-
[60]	2.45	40	-31	180.7
[61]	0.402	2.49	-60	-
[62]	2.45	35.5	-46.02	-

**Fig. 2.4:** Helical implantable antennas reported in [59]-[62] respectively ©IEEE

2.3 MODELING TECHNIQUES OF IMPLANTABLE ANTENNAS

Implantable Medical Devices (IMDs) are designed by assembling different components such as antenna, battery, biosensor, camera, data management system etc. However, the antenna is an important element for defining robustness of the data reception and transmission between IMDs and external monitoring. Different types of implantable antennas were proposed in last decade which is discussed in Section 2.2. Scientists used different analytical and practical techniques for modeling implantable antennas within human body models. The general process of modeling implantable antennas discussed in Section 2.2 is mentioned in Fig. 2.5 in form of flow chart.

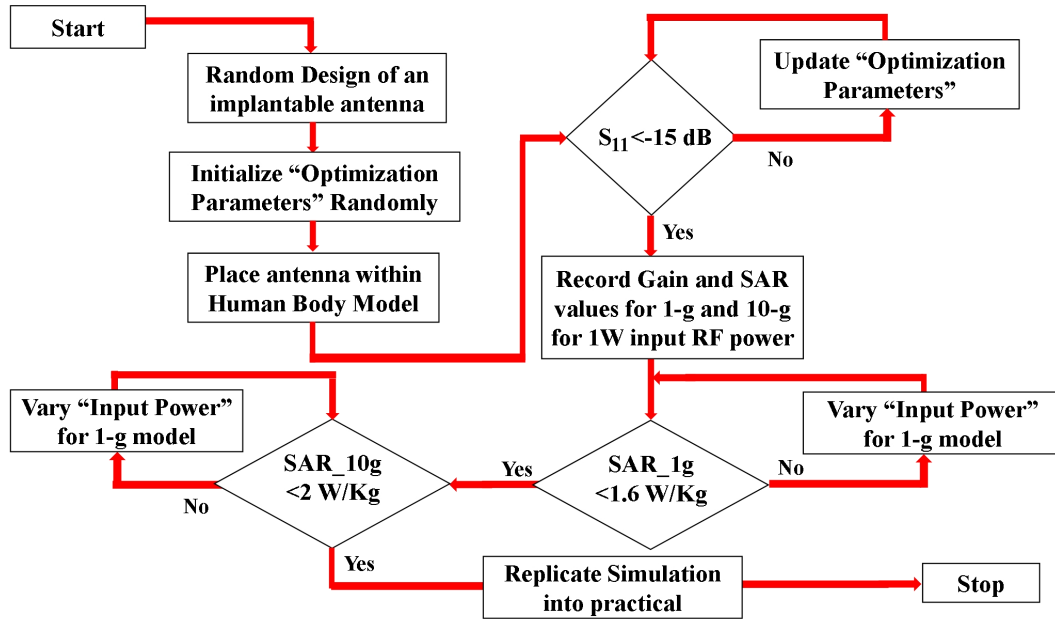


Fig. 2.5: Flow Chart for the numerical design of implantable antenna

Different techniques have been followed by researchers to design the previously reported implantable antennas.

2.3.1. Particle Swarm Optimization (PSO):

Particle Swarm Optimization (PSO) technique was used by Karacolak et al. [63] to model implantable antenna at MICS (Medical Implant Communication System) and ISM bands. PSO was introduced by Eberhart and Kennedy in 1995. The method of this stochastic optimization algorithm is similar to the swarms of bees as they search for food sources. The main components of PSO are – (i) particles (a swarm of particles which are representing a potential solution of an optimization problem), (ii) position (particle's position is a set of coordinates in a search space of the N-dimensional solution), (iii) velocity (particle's velocity determines its speed and direction) and (iv) fitness (it determines quality of each solution). Position (χ_n) and velocity (v_n) of swarm are initialized randomly. Each

particle remembers its own best fitness ($pbest$) and swarm's best fitness ($gbest$) among all particles. Then the fitness of the particles is evaluated and updated based on below equations by repeating until convergence is reached [63]:

$$\chi_n(t + \Delta t) = \chi_n(t) + \Delta t v_n(t)$$

$$v_n(t + \Delta t) = \xi v_n(t) + \kappa_1 rand() \{pbest - \chi_n(t)\} + \kappa_2 rand() \{gbest - \chi_n(t)\}$$

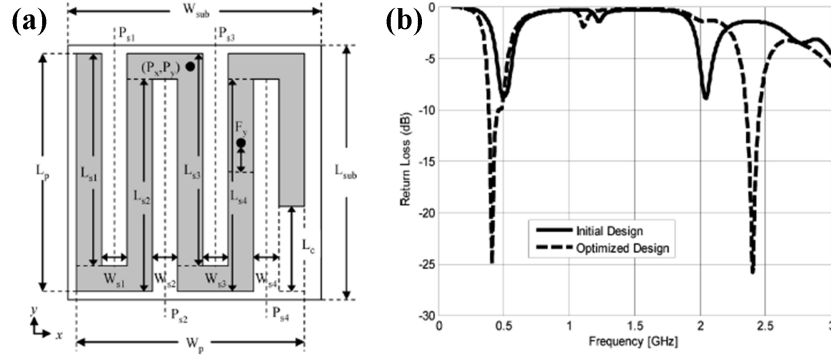


Fig. 2.6: (a) Optimization of implantable antenna using PSO; (b) optimized S_{11} plot by Karacolak et. al [63]. ©IEEE

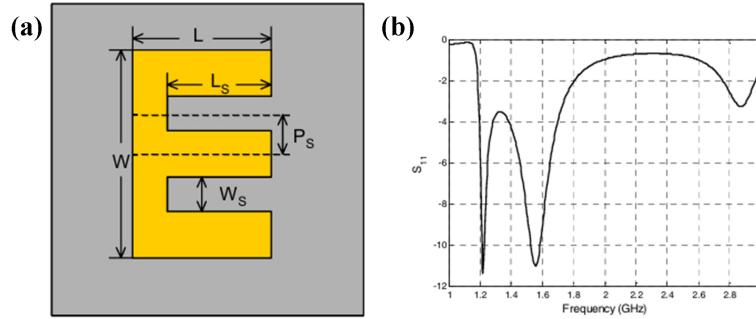


Fig. 2.7: (a) Optimization of implantable antenna using PSO; (b) optimized S_{11} plot by Hood et. al [64]. ©IEEE

Here, ξ = inertial time-varying weight, κ_1 and κ_2 are constants. The function $rand()$ calls a random number generator. The time step Δt is set as 1 ideally. The optimization of implantable antenna using PSO by Karacolak et.al [63] is shown in Fig. 2.6. They considered fitness function to optimize their antenna as $fitness(S_{11}@402\text{ MHz}, S_{11}@2.45\text{ GHz})$. From Fig. 2.6 (b), it is observed

that they were able to tune the antenna architecture properly at 402 MHz and 2.45 GHz properly using PSO algorithm. Hood et. al [64] also used PSO technique to optimize their implantable antenna shown in Fig. 2.7. The antenna was tuned properly at 1.227 GHz and 1.575 GHz by using fitness function: $fitness(S_{11 @ 1.227 \text{ GHz}}, S_{11 @ 1.575 \text{ GHz}})$.

There are several other evolutionary algorithm techniques to design antennas such as Trust Region Framework (TRF), Nelder Mead Simplex (NMS) algorithm, Characteristic Mode Analysis (CMA) evolution, Genetic Algorithm (GA), Interpolated Quasi Newton (IQN), Classical Powell etc. The main disadvantage of TRF algorithm is computational cost. The simulation of body implantable antenna using TRF algorithm makes the computational cost very high [65]. The design and analysis of implantable antenna using NMS and CMA techniques may be disadvantageous because this technique is widely dependent on initial condition and the performance of NMS technique degrades with the increase in number of uncertain parameters [66-67]. GA is also depending on initial condition and model complexity. However, this technique is robust in nature to estimate the solutions of random problems [68].

PSO is a simple optimization technique. However, the main problem related to this technique is that this technique requires apriori knowledge of estimated solution of the problem. From the perspective of antenna application, the antenna responses can be predicted before optimization. Therefore, PSO technique has been preferred to design implantable antennas.

2.3.2. Finite Difference Time Domain (FDTD) method:

Partial Differentiation Method based Finite Difference Time Domain (FDTD) method is the simplest and most widely used popular method of computation electromagnetic (CEM). K. Yee (1996) [69] introduced this technique first for boundary value problem based on differential form of Maxwell's equations. It is very useful for time domain formulations such as wave

(radar pulse) propagation, transient phenomena (switching, lightning) etc. The brief idea of FDTD is given below:

Let us consider an isotropic, non-dispersive, linear and time invariant media. The displacement field ($\vec{D}$) = $\epsilon\vec{E}$ and magnetic flux density ($\vec{B}$) = $\mu\vec{H}$.

Maxwell's equations are,

$$\begin{aligned}\frac{\partial \vec{D}}{\partial t} &= \nabla \times \vec{H} - \vec{J} = \epsilon \frac{\partial \vec{E}}{\partial t} \\ \frac{\partial \vec{B}}{\partial t} &= -\nabla \times \vec{E} = \mu \frac{\partial \vec{H}}{\partial t}\end{aligned}$$

From Taylor's theorem,

$$f\left(a_o + \frac{\Delta a}{2}\right) = f(a_o) + \frac{\Delta a}{2} f'(a)|_{a_o} + \dots \quad (1)$$

and

$$f\left(a_o - \frac{\Delta a}{2}\right) = f(a_o) - \frac{\Delta a}{2} f'(a)|_{a_o} + \dots \quad (2)$$

Then we get from above equations,

$$f'(a)|_{a_o} = \frac{f\left(a_o + \frac{\Delta a}{2}\right) - f\left(a_o - \frac{\Delta a}{2}\right)}{\Delta a} + \varphi(\Delta a^2)$$

The term $\varphi(\Delta a^2)$ is the error term with order 2. In FDTD, the derivatives are replaced by finite differences in space domain. This difference is also known as centered, two point finite difference. If the term Δa is reduced by half amount, the error term $\varphi(\Delta a^2)$ is decreased by a factor of 4. From equation (1), we can write

$$\frac{\Delta a}{2} f'(a)|_{a_o} \approx f\left(a_o + \frac{\Delta a}{2}\right) - f(a_o) + \varphi(\Delta a^2)$$

The above expression is called one-sided difference. Therefore, the FDTD technique is related to Maxwell's equations and Taylor's theorem. Generally, two types of polarizations are there – (i) Transverse Electric (TE) represented by (E_x, E_y, H_z) and (ii) Transverse Magnetic (TM) represented by (H_x, H_y, E_z) and

any practical problem can be represented as the linear superposition between these two polarizations. Let us first consider TE mode for formulating 2D FDTD.

From Maxwell's equations, we can write (considering $\vec{J}=0$)

$$\epsilon_o \frac{\partial E_x}{\partial t} = \frac{\partial H_z}{\partial y} \quad (3)$$

$$\epsilon_o \frac{\partial E_y}{\partial t} = -\frac{\partial H_z}{\partial x} \quad (4)$$

$$\mu_o \frac{\partial H_z}{\partial t} = \frac{\partial E_x}{\partial y} - \frac{\partial E_y}{\partial x} \quad (5)$$

The variables (E_x, E_y, H_z) should be discretized in space and time. A repeating space grid is required here for analysis and one of them is shown in Fig. 2.8(a). The electric field is discretized along the grid lines and magnetic field at the centre of a grid. Using centered two point difference idea, the equations (3-5) can be written as below.

$$\begin{aligned} \epsilon_o \dot{E}_x \left(i + \frac{1}{2}, j \right) &= \frac{H_z \left(i + \frac{1}{2}, j + \frac{1}{2} \right) - H_z \left(i + \frac{1}{2}, j - \frac{1}{2} \right)}{\Delta y} \\ \epsilon_o \dot{E}_y \left(i, j + \frac{1}{2} \right) &= \frac{H_z \left(i - \frac{1}{2}, j + \frac{1}{2} \right) - H_z \left(i + \frac{1}{2}, j + \frac{1}{2} \right)}{\Delta x} \\ \mu_o \dot{H}_z \left(i + \frac{1}{2}, j + \frac{1}{2} \right) &= \left[\frac{E_x \left(i + \frac{1}{2}, j + 1 \right) - E_x \left(i + \frac{1}{2}, j \right)}{\Delta y} \right] \\ &\quad - \left[\frac{E_y \left(i + 1, j + \frac{1}{2} \right) - E_y \left(i, j + \frac{1}{2} \right)}{\Delta x} \right] \end{aligned}$$

A space grid is considered in Fig. 2.8 (b) for TM polarization (H_x, H_y, E_z) . In space, the variables are staggered by half of a grid $(\Delta x, \Delta y)$. In time (Fig. 2.8 (c)), the variable is also staggered by half of grid (Δt) . Equations (1-2) can be written as

$$\begin{aligned} \epsilon_o \left[\frac{\vec{E}^m - \vec{E}^{m-1}}{\Delta t} \right] &= \nabla \times \vec{H}^{m-\frac{1}{2}} \\ \Rightarrow \vec{E}^m &= \vec{E}^{m-1} + \left[\frac{\Delta t}{\epsilon_o} \right] \left[\nabla \times \vec{H}^{m-\frac{1}{2}} \right] \end{aligned} \quad (6)$$

$$\begin{aligned} \mu_o \left[\frac{\vec{H}^{m+\frac{1}{2}} - \vec{H}^{m-\frac{1}{2}}}{\Delta t} \right] &= -\nabla \times \vec{E}^m \\ \Rightarrow \vec{H}^{m+\frac{1}{2}} &= \vec{H}^{m-\frac{1}{2}} - \left[\frac{\Delta t}{\mu_o} \right] \left[\nabla \times \vec{E}^m \right] \end{aligned} \quad (7)$$

Next, FDTD analysis of 3D model (specifically used for antenna design) is performed by considering 3D geometry of Yee's cell [69] (Fig. 2.8 (d)). The electric and magnetic fields using centered two point difference are mentioned below [70].

$$\begin{aligned} E_x^{m+1} \left(i + \frac{1}{2}, j, k \right) &= \left(\frac{\epsilon - \sigma \frac{\Delta t}{2}}{\epsilon + \sigma \frac{\Delta t}{2}} \right) E_x^m \left(i + \frac{1}{2}, j, k \right) \\ &+ \left(\frac{\frac{\Delta t}{\Delta y}}{\epsilon + \sigma \frac{\Delta t}{2}} \right) \left[H_z^{m+\frac{1}{2}} \left(i + \frac{1}{2}, j + \frac{1}{2}, k \right) \right. \\ &\quad \left. - H_z^{m+\frac{1}{2}} \left(i + \frac{1}{2}, j - \frac{1}{2}, k \right) \right] \\ &- \left(\frac{\frac{\Delta t}{\Delta z}}{\epsilon + \sigma \frac{\Delta t}{2}} \right) \left[H_y^{m+\frac{1}{2}} \left(i + \frac{1}{2}, j, k + \frac{1}{2} \right) \right. \\ &\quad \left. - H_y^{m+\frac{1}{2}} \left(i + \frac{1}{2}, j, k - \frac{1}{2} \right) \right] \end{aligned}$$

$$\begin{aligned}
E_y^{m+1}\left(i, j + \frac{1}{2}, k\right) &= \left(\frac{\epsilon - \sigma \frac{\Delta t}{2}}{\epsilon + \sigma \frac{\Delta t}{2}}\right) E_y^m\left(i, j + \frac{1}{2}, k\right) \\
&+ \left(\frac{\frac{\Delta t}{\Delta z}}{\epsilon + \sigma \frac{\Delta t}{2}}\right) \left[H_x^{m+\frac{1}{2}}\left(i, j + \frac{1}{2}, k + \frac{1}{2}\right) \right. \\
&- \left. H_x^{m+\frac{1}{2}}\left(i, j + \frac{1}{2}, k - \frac{1}{2}\right) \right] \\
&- \left(\frac{\frac{\Delta t}{\Delta x}}{\epsilon + \sigma \frac{\Delta t}{2}}\right) \left[H_z^{m+\frac{1}{2}}\left(i + \frac{1}{2}, j + \frac{1}{2}, k\right) \right. \\
&- \left. H_z^{m+\frac{1}{2}}\left(i - \frac{1}{2}, j + \frac{1}{2}, k\right) \right]
\end{aligned}$$

$$\begin{aligned}
E_z^{m+1}\left(i, j, k + \frac{1}{2}\right) &= \left(\frac{\epsilon - \sigma \frac{\Delta t}{2}}{\epsilon + \sigma \frac{\Delta t}{2}}\right) E_z^m\left(i, j, k + \frac{1}{2}\right) \\
&+ \left(\frac{\frac{\Delta t}{\Delta x}}{\epsilon + \sigma \frac{\Delta t}{2}}\right) \left[H_y^{m+\frac{1}{2}}\left(i + \frac{1}{2}, j, k + \frac{1}{2}\right) \right. \\
&- \left. H_y^{m+\frac{1}{2}}\left(i - \frac{1}{2}, j + \frac{1}{2}, k\right) \right] \\
&- \left(\frac{\frac{\Delta t}{\Delta y}}{\epsilon + \sigma \frac{\Delta t}{2}}\right) \left[H_x^{m+\frac{1}{2}}\left(i, j + \frac{1}{2}, k + \frac{1}{2}\right) \right. \\
&- \left. H_x^{m+\frac{1}{2}}\left(i, j - \frac{1}{2}, k + \frac{1}{2}\right) \right]
\end{aligned}$$

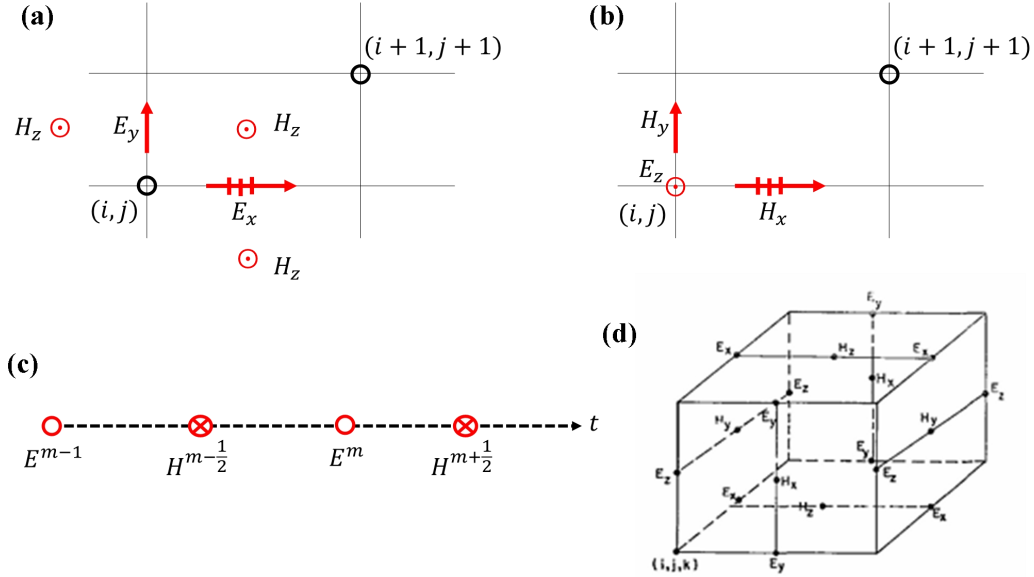


Fig. 2.8: (a) Space grid for TE, (b) space grid for TM, (c) time grid and (d) Yee's cell geometry for FDTD analysis

$$\begin{aligned}
 H_x^{m+\frac{1}{2}}\left(i, j+\frac{1}{2}, k+\frac{1}{2}\right) &= H_x^{m-\frac{1}{2}}\left(i, j+\frac{1}{2}, k+\frac{1}{2}\right) \\
 &\quad - \left(\frac{\Delta t}{\mu \Delta y}\right)\left[E_z^m\left(i, j, k+\frac{1}{2}\right) - E_z^m\left(i, j-1, k+\frac{1}{2}\right)\right] \\
 &\quad + \left(\frac{\Delta t}{\mu \Delta z}\right)\left[E_y^m\left(i, j+\frac{1}{2}, k\right) - E_y^m\left(i, j+\frac{1}{2}, k-1\right)\right]
 \end{aligned}$$

$$\begin{aligned}
 H_y^{m+\frac{1}{2}}\left(i+\frac{1}{2}, j, k+\frac{1}{2}\right) &= H_y^{m-\frac{1}{2}}\left(i+\frac{1}{2}, j, k+\frac{1}{2}\right) \\
 &\quad - \left(\frac{\Delta t}{\mu \Delta z}\right)\left[E_x^m\left(i+\frac{1}{2}, j, k\right) - E_x^m\left(i+\frac{1}{2}, j, k-1\right)\right] \\
 &\quad + \left(\frac{\Delta t}{\mu \Delta x}\right)\left[E_z^m\left(i, j, k+\frac{1}{2}\right) - E_z^m\left(i-1, j, k+\frac{1}{2}\right)\right]
 \end{aligned}$$

$$\begin{aligned}
H_z^{m+\frac{1}{2}}\left(i+\frac{1}{2}, j+\frac{1}{2}, k\right) &= H_z^{m-\frac{1}{2}}\left(i+\frac{1}{2}, j+\frac{1}{2}, k\right) \\
&\quad - \left(\frac{\Delta t}{\mu\Delta x}\right)\left[E_y^m\left(i, j+\frac{1}{2}, k\right) - E_y^m\left(i-1, j+\frac{1}{2}, k\right)\right] \\
&\quad + \left(\frac{\Delta t}{\mu\Delta y}\right)\left[E_x^m\left(i+\frac{1}{2}, j, k\right) - E_x^m\left(i+\frac{1}{2}, j-1, k\right)\right]
\end{aligned}$$

A three-dimensional FDTD analysis of a dual-band implantable antenna was performed by Noroozi et. al [71]. The generalized FDTD method for lossy dielectric media is used for narrowband applications related to biological tissues. However, it is essential to use time domain equations for broadband applications in which the variations in dielectric properties of tissues with frequency should be considered [71]. The problem space considered in [71] is shown in Fig. 2.9 (a). Xia et. al [72] studied an implanted cavity slotted antenna within human arm using FDTD technique at 2.45 GHz. Fig. 2.9 (b) shows the problem space considered in [72].

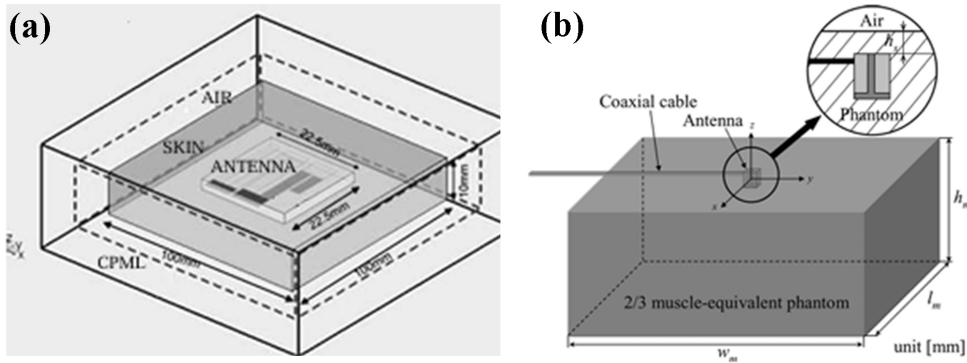


Fig. 2.9: Problem space considered in (a) [71] and (b) [72] for FDTD analysis

2.3.3. Finite Element Method (FEM) method:

The time domain Maxwell's equations are

$$\nabla \times \vec{E} = -\frac{\partial \vec{B}}{\partial t}$$

$$\nabla \times \vec{H} = \frac{\partial \vec{D}}{\partial t} + \vec{J}$$

$$\nabla \cdot \vec{D} = 0$$

$$\nabla \cdot \vec{B} = 0$$

The frequency domain Maxwell's equations are $\left(\frac{\partial}{\partial t} \leftrightarrow -j\omega\right)$

$$\nabla \times \vec{E} = j\omega \vec{B} = j\omega \mu \vec{H}$$

$$\nabla \times \vec{H} = j\omega \vec{D} + \vec{J} = j\omega \epsilon \vec{E} + \vec{J}$$

$$\nabla \cdot \vec{D} = 0$$

$$\nabla \cdot \vec{B} = 0$$

Considering boundary conditions, the Maxwell's equations can be remodeled. One of the important boundary conditions is that the tangential components of both electric and magnetic fields are continuous at the boundary at material interfaces. From frequency domain Maxwell's equations, we get

$$\nabla \times \nabla \times \vec{E} - \omega^2 \epsilon \vec{E} = j\omega \vec{J} \quad (8)$$

Different boundary conditions are there as described below:

- (i) *Dirichlet condition*: It specifies the tangential component of electric field ($\vec{E}$) at the boundary.
- (ii) *Neumann condition*: It tells about the normal component of flux value ($\partial_n \vec{E}$).
- (iii) *Mixed or Cauchy condition*: It is the combination of Dirichlet and Neumann conditions ($\alpha \vec{E} + \partial_n \vec{E}$).

If the system is invariant along (y, z) plane, equation (8) can be written as

$$\partial_x^2 E_x + \omega^2 \epsilon_x E_x = -j\omega J_x$$

Here, Dirichlet boundary condition is considered which specifies as

$$E_x(x_o) = E_x(x_N) = 0$$

There are two approaches for Finite Element Method (FEM) such as,

- (i) Rayleigh-Ritz method (variational)
- (ii) Galerkin's method (weighted residuals)

These two approaches can lead to similar discretization. So, their final forms are quite similar. Here, Galerkin's method is considered for discussing about the FEM technique.

$$\partial_x^2 E_x + \omega^2 \epsilon_x E_x + j\omega J_x = 0$$

Let us take an approximated value of electric field $\tilde{E}_x$ which is approximately equal to E_x . Therefore, the equation can be exactly equal to the zero. It introduces a residual in the above equation. The modified equation is

$$\partial_x^2 \tilde{E}_x + \omega^2 \epsilon_x \tilde{E}_x + j\omega J_x = Res(x)$$

This residual is orthogonal to the function space $\{w_n\}$, where w_n be the weights corresponds to the each of the element. Therefore,

$$\begin{aligned} \int_{x_0}^{x_N} Res(x) w_n(x) dx &= 0 \\ \int_{x_0}^{x_N} [\partial_x^2 \tilde{E}_x + \omega^2 \epsilon_x \tilde{E}_x + j\omega J_x] w_n(x) dx &= 0 \\ \int_{x_0}^{x_N} [\partial_x^2 \tilde{E}_x] w_n(x) dx + \omega^2 \int_{x_0}^{x_N} [\epsilon_x \tilde{E}_x] w_n(x) dx + j\omega \int_{x_0}^{x_N} [J_x] w_n(x) dx &= 0 \\ \int_{x_0}^{x_N} [\partial_x^2 \tilde{E}_x] w_n(x) dx + \omega^2 \int_{x_0}^{x_N} [\epsilon_x \tilde{E}_x] w_n(x) dx &= F_n \end{aligned}$$

Applying the integration by parts,

$$-\int_{x_0}^{x_N} [\partial_x \tilde{E}_x] [\partial_x w_n(x)] dx + \omega^2 \int_{x_0}^{x_N} [\epsilon_x \tilde{E}_x] w_n(x) dx + \left[[\partial_x \tilde{E}_x] w_n(x) \right]_{x_0}^{x_N} = F_n$$

Let $\tilde{E}_x = \sum_{i=0}^N \wp_i \mathcal{F}_i(x)$, $\mathcal{F}_i(x)$ is basis function and $\wp_i$ is expansion coefficients. If $N \rightarrow \infty$, $\tilde{E}_x = E_x$. Practically it is impossible. Putting this expansion in the above equation, we get

$$\sum_{i=0}^N \left\{ \omega^2 \int_{x_0}^{x_N} [\epsilon_x \mathcal{F}_i(x)] w_n(x) dx - \int_{x_0}^{x_N} [\partial_x \mathcal{F}_i(x)] [\partial_x w_n(x)] dx + [\partial_x \mathcal{F}_i(x)] w_n(x) \Big|_{x_0}^{x_N} \right\} \wp_i = F_n$$

As per Galerkin's method, $\mathcal{F}_i(x) = w_i(x)$. Therefore,

$$\sum_{i=0}^N \left\{ \omega^2 \int_{x_0}^{x_N} [\epsilon_x \mathcal{F}_i(x)] \mathcal{F}_i(x) dx - \int_{x_0}^{x_N} [\partial_x \mathcal{F}_i(x)] [\partial_x \mathcal{F}_i(x)] dx + [\partial_x \mathcal{F}_i(x)] \mathcal{F}_i(x) \Big|_{x_0}^{x_N} \right\} \wp_i = F_n$$

In matrix form, we can write,

$$\{\omega^2 [M_{ij}] - [S_{ij}] + [B_{ij}]\} \wp = F$$

Where $[M_{ij}]$ is material matrix, $[S_{ij}]$ is stiffness matrix and $[B_{ij}]$ is matrix related to boundary conditions. This is the brief idea of FEM. Different implantable antennas were designed using FEM based simulators such as HFSS (High Frequency Structure Simulator), CST Studio Suite etc. The antennas proposed in [18-64] were designed using FEM technique. FEM can handle complex geometries like 3D human anatomical and canonical models, boundary, materials etc. more efficiently than FDTD with better accuracy. However, the cost of simulation and time complexity are more in FEM than FDTD. FDTD and FEM are popular for simulating implantable antenna prototypes due to their ability of handling complex geometries of human tissue model and recording antenna performances within human body models for time-varying electromagnetic fields.

Another popular technique of antenna design is Method of Moment (MoM) technique. This method is simpler than FDTD and FEM techniques. However, arbitrary basis functions are taken to estimate antenna performances. The

accuracy and convergence of MoM technique depends on the estimation of basis function. For the analysis of implantable antenna within human body model, it is very difficult to consider any basis function arbitrarily. In spite of having simulation complexity in FDTD and FEM techniques, these techniques provide better responses of implantable antennas than MoM techniques. Therefore, FDTD and FEM techniques were widely used for implantable antennas than MoM technique.

2.4 APPLICATION OF IMPLANTABLE ANTENNAS

In last decades, several implantable antennas were reported for different purposes. The reported implantable antennas can be classified into two major application areas [73] – (i) detection applications and (ii) communication applications. The antennas designed for detection purposes can perform sensing and imaging for detection of several diseases. However, antennas designed for communication purposes are employed for providing wireless communication between sensors, implants, medical devices etc [73].

There are three types of communication techniques used in medical applications. If any wearable antenna placed on clothing or adhered to skin communicates with any other device placed on the same platform, then the communication is coined as *on-body communication*. When the wireless communication is performed between wearable antenna and external devices, then the mode of communication is known as *off-body communication*. When any implantable antenna communicates with external devices, then this type of communication is called *in-body communication* [73-74]. One of the popular in-body implantable medical devices is pacemaker. If any medical device is placed inside human body by swallowing like pills, then this implantable medical device is called ingestible device [75]. For detecting skin, brain and breast cancers, off-

body communication is widely used. However, for imaging or acquiring information related to internal organs or detecting blood related information such as glucose level, pressure level etc., cardiovascular diseases, in-body communication is necessary. Implantable communication antennas are used for different applications such as remote patient monitoring, Wireless Body Area Network (WBAN), telemedicine etc. [73-78]. To characterize the implantable antennas practically, two types of testing are performed – in-vitro and in-vivo testing. *In-vitro* testing of implantable antennas is performed inside human body-mimicking phantom model (head, eye, ears, breast, heart etc.). However, testing implantable antennas within real human or animal body part (especially minced pork) is termed as *in-vivo* testing.

Frequency Band Name	Frequency Range	Applications
Ultra-Low Frequency (ULF) and Extremely Low Frequency (ELF)	Below 1 Hz. 1 Hz to 300 Hz	Electroencephalography (EEG) and Magnetocardiography (MCG)
Very Low Frequency (VLF)	3 Hz to 30 Hz	Electrocardiography (ECG)
Low Frequency (LF)	30 Hz to 300 Hz	Electromyography (EMG)
Medium Frequency (MF)	300 Hz to 3 MHz	Diathermy and Medical RF ablation
High Frequency (HF)	3 MHz to 30 MHz	Medical Imaging
Ultra-High Frequency (UHF)	300 MHz to 3 GHz	Magnetic Resonance Imaging (MRI)
Microwave Frequency	1 GHz to 30 GHz	Hyperthermia
Radio Frequency (RF)	30 MHz to 300 GHz	Wireless Medical devices and Biotelemetry

Fig. 2.10: Frequencies related for several biomedical applications [73]

2.4.1. Bandwidth requirements for Biomedical Applications:

Depending on the application of the implantable antennas, the bandwidth requirements can be varied. Chishti et. al [73] tabulated the required frequencies related to different biomedical applications which is shown in Fig. 2.10. Some of them are discussed below [73]:

- a) Medical Imaging: To capture proper images using MRI and ultrasound, high bandwidth (order of kHz) is required [73].
- b) ECG and EEG: The ECG and EEG signals are generally low-frequency and low bandwidth (few Hz) signals [73].
- c) Hyperthermia: High frequency range is required to generate and deliver energy for treatment of hyperthermia. The required bandwidth for this purpose is about several MHz [73].
- d) Biomedical Communication: It depends upon the transmitting data. For low-data transmission, it requires few kHz bandwidths. For high-definition data transmission, several MHz bandwidths are required [73-78].

2.4.2. Medical Applications of Implantable Antennas:

In this section, a detailed review of implantable antennas for different medical applications is presented.

- 1) Cardiac Leadless Pacemaker: As per recent estimation, about 3 million people worldwide use pacemakers. The urgency of this device is rising as cardiac arrest alone tells upon more than 25 millions worldwide [79]. The traditional cardiac pacemakers have certain drawbacks due to device-related complications [80]. Besides, the lead of the pacemaker may interact with Magnetic Resonance Imaging (MRI) systems which cause tissue heating [81]. Cardiac Leadless Pacemaker can overcome these problems of traditional cardiac pacemaker. A leadless pacemaker system has many advantages, such as less invasiveness, high

efficiency (no system connection, MRI compatibility), cost-effectiveness etc. [81-82].

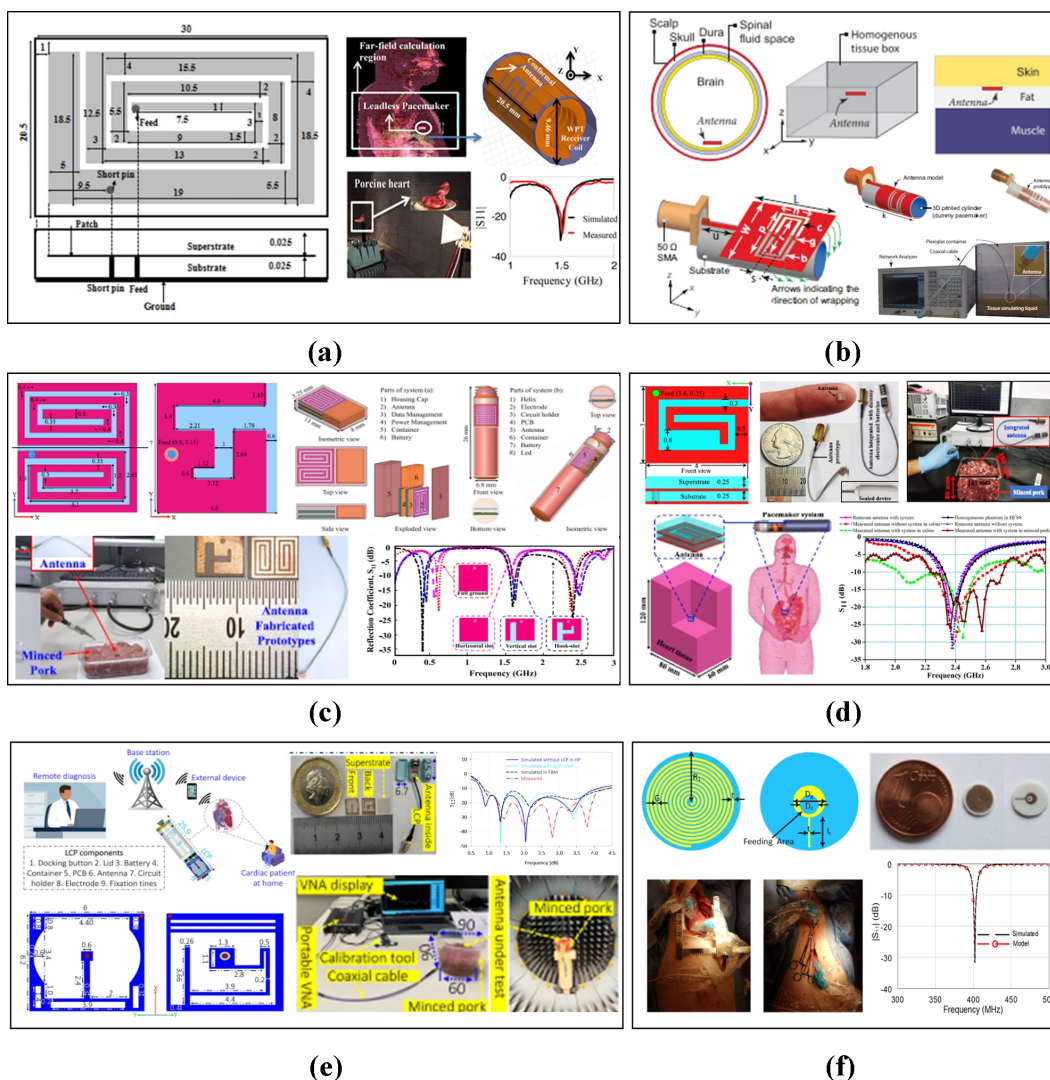


Fig. 2.11: Some implantable antennas for Leadless Cardiac Pacemaker applications reported in [81-84], [44] and [85] respectively

The authors of [81] designed a 30.75 mm³ conformal spiral type implantable antenna with a diameter of 9.46 mm and a length of 20.5 mm at MICS band (shown in Fig. 2.11 (a)) for Cardiac Leadless Pacemaker (CLP). The antenna wrapped around pacemaker surface was designed within a homogeneous

heart phantom model ($\epsilon_r=66$, $\sigma = 0.6$ S/m at 402 MHz) and in-vivo testing was performed within porcine heart. The antenna was operated with around -32 dBi gain within porcine heart. Another right angled Split Ring Resonator based conformal (diameter of 6.58 mm and a length of 20 mm) implantable antenna operating with -35 dBi gain at 2.45 GHz was reported in [83].

A triple-band implantable antenna shown in Fig. 2.11(c), was presented in [84]. The antenna was operated at MICS (402-405 MHz), ISM band (433.1-434.8 MHz and 2.4-2.48 GHz) and midfield (1.52-1.69 GHz) with respective gain values of -30.5 dBi, -18.2 dBi and -22.6 dBi. The bandwidths at 402 MHz, 1600 MHz and 2450 MHz were 148, 171 and 219 MHz respectively. A miniaturized implantable antenna shown in Fig. 2.11 (d) was designed for leadless cardiac pacemaker at 2.4 GHz with -25.8 dBi gain in [82]. Alghamdi et. al [44] reported a compact broadband implantable antenna (refer to Fig. 2.11 (e)) for leadless pacemaker with fractional bandwidth of 138% and realized gain of -27.94 dBi at 1.4 GHz. A circular spiral implantable antenna with volume of 22.20 mm³ was designed in [85] which is shown in Fig. 2.11(f).

2) Wireless Capsule Endoscopy: The attractiveness of Wireless Capsule Endoscopy (WCE) is increasing because of the ability to observe diseases using physiological signals and real-time biological images transferred from swallowed or implanted ingestible medical devices which transmit patients' health parameters (blood sugar, vitamin level, generation of tumors, etc.) to external diagnostic devices while passing through the Gastrointestinal (GI) tract [30]. The WCE system can capture images with a high-resolution inside the GI tract while traversing along it [58]. A WCE generally comprises of an antenna sensor for wireless transmission of sensed health-parameters, camera, battery, Printed Circuit Board (PCB) and data management system. The capsule system has typical dimensions of 26 X 11 mm² [41], [86]. Implantable antennas must overcome several obstacles, including small size, high-speed communication,

human safety and a reliable communication link [87]. Different implantable antennas were reported to achieve these challenges for WCE. A capacitor-loaded meander-slotted wideband implantable antenna (Fig. 2.12 (a)) was described in [30]. This antenna was radiating over fractional bandwidth (FBW) of 20.5% (2.17-2.69 GHz) around 2.45 GHz with -26.4 dBi gain. However, the size of the antenna was large ($\pi \times 102 \times 0.635 \text{ mm}^3$). At 433 MHz, a conformal meander-line capsule antenna shown in Fig. 2.12 (b) with an FBW of 65.59% and a gain of -32.8 dBi was demonstrated in [58].

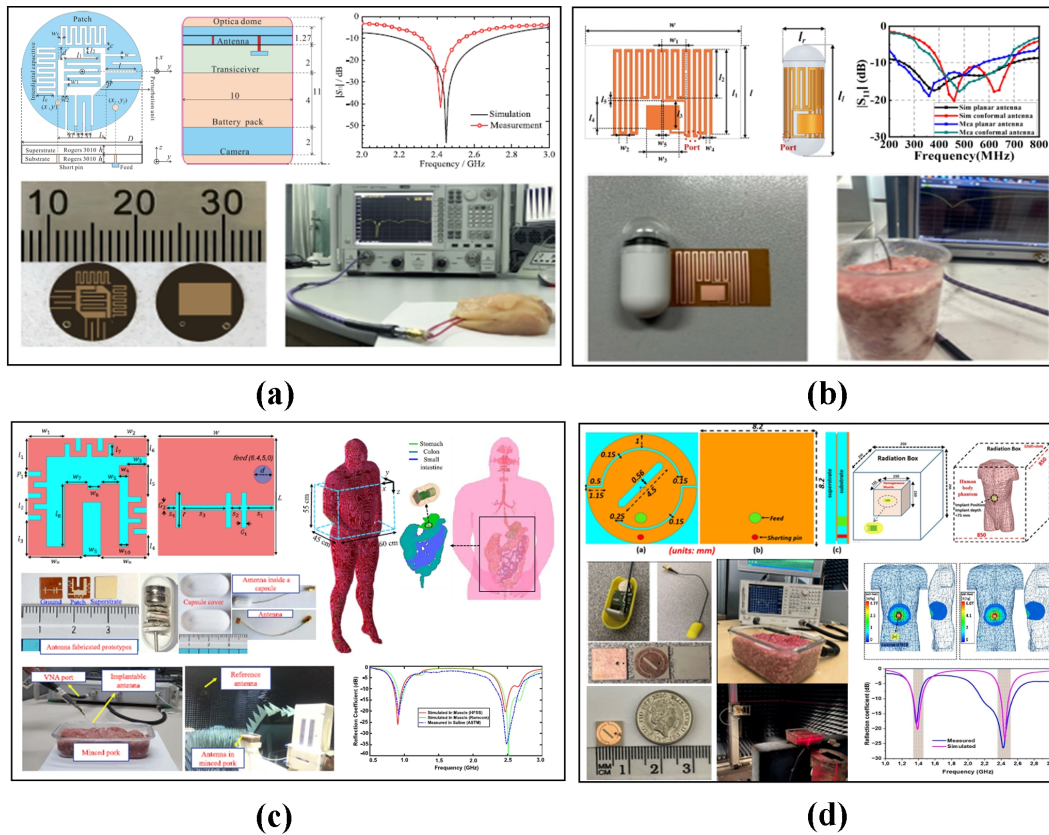


Fig. 2.12: Some reported implantable antennas for Wireless Capsule Endoscopy applications in [30], [58], [86] and [41] respectively

Hayat et. al [86] proposed an ultra-miniaturized (volume of 2.11 mm^3) dual-band implantable antenna [86]. Despite being small, the bandwidth of this

antenna was narrow (FBWs of 13.63% and 6.28% around 915 MHz and 2.45 GHz, respectively) in nature. This capsule implantable antenna can communicate up to 5 m and 3.8 m at 915 MHz and 2.45 GHz, respectively, at 100 kbps data rate. In the studies above, the miniaturization of the antenna for WCE inevitably leads to a reduction in bandwidth. Dual-band antennas provide the advantage of high-speed data transmission and reception at two different frequencies. In [41], a dual-band slotted circular implanted antenna (Fig. 2.12 (d)) with a 42.7 mm^3 volume that operates at 1.4 GHz and 2.45 GHz was proposed for wireless capsule endoscopy. The antenna radiated at 1.4 GHz and 2.45 GHz with FBW of 7.2% and 4.2% respectively and realized gain of -29.4 dBi and -30.4 dBi. The bandwidth of this design is narrow which leads to reduced speed of data transmission, increase signal interference and sensitivity to channel noise etc.

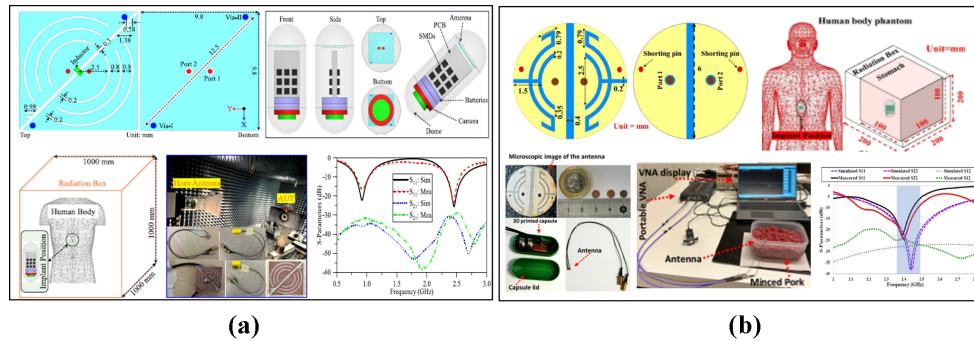


Fig. 2.13: Some reported MIMO implantable antennas for Wireless Capsule Endoscopy applications presented in [87], [43] respectively

There are some other reported implantable antennas for wireless capsule endoscopy which has been mentioned below. In [87], a compact dual-band Multiple Input Multiple Output (MIMO) antenna was designed for wireless capsule endoscopy. The size of the antenna was 28.81 mm^3 . The antenna had -29.8 dBi and -24.6 dBi gains at two operating frequencies 915 MHz and 2450 MHz respectively. This antenna is shown in Fig. 2.13 (a). A in-band compact full-duplex implantable antenna shown in Fig. 2.13 (b), was designed in [43] for

wireless capsule endoscopy. The antenna had two elements with two input ports. It was operated at 2.45 GHz with realized gain of -24.4 dBi.

3) Glucose Monitoring: Continuous glucose monitoring is very important for diabetic patients. Subcutaneous real-time glucose sensors can monitor the glucose level in blood continuously [25]. Embedding implantable antenna with these sensors can enable the facility of wireless communication with external devices to transmit recorded physiological information. Real-time glucose monitoring by subcutaneous sensor was depicted by Hassan et. al [88] which is shown in Fig. 2.14. Different implantable antennas were designed for this purposes which have been discussed below.

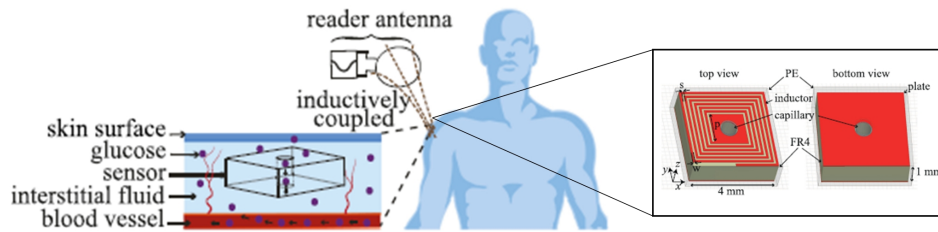


Fig. 2.14: Concept of Real-time Glucose Monitoring by Subcutaneous sensor illustrated by Hassan et. al [88].

A complementary SRR loaded miniaturized ($8.5 \times 8.5 \times 1.27 \text{ mm}^3$) C-shaped slotted circularly polarized implantable antenna was reported in [25] for wideband (12.2%) real-time glucose monitoring. The antenna shown in Fig. 2.15 (a), was placed within a three-layer human body model for simulation and pork for practical verification. An ultra-wideband CP implantable antenna shown in Fig. 2.15 (b) was designed in [89] for continuous glucose monitoring. The antenna had realized gain of -24.64 dBi and -20.04 dBi at 1.9 GHz and 2.45 GHz respectively.

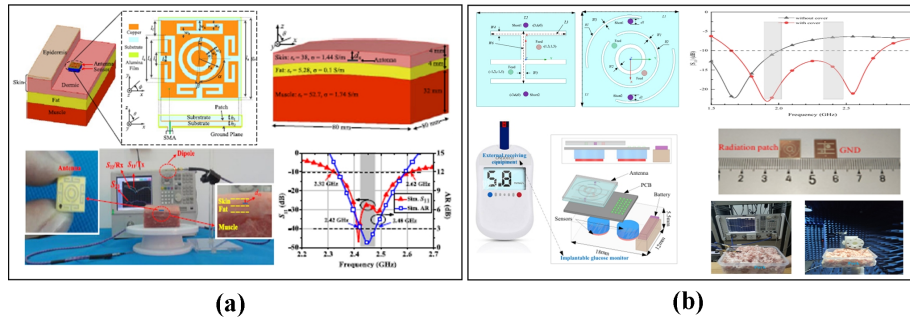


Fig. 2.15: Reported Implantable Antennas for Real-time Glucose Monitoring in [25] and [89] respectively.

4) Head Implant: The head-implantable antennas take crucial role to monitor intracranial pressure, manage disorders caused by traumatic injuries and strokes [90] etc. Intracranial Pressure (ICP) is exerted by Cerebrospinal Fluid (CSF) inside skull. Usually, this pressure ranges from 7 to 15 mmHg in a supine adult. Accumulation of excess fluid or swelling may increase ICP which causes severe headaches and brain damage [91].

In [27], the authors designed a 24 mm³ dual band implantable antenna system shown in Fig. 2.16 (a) for ICP monitoring. The antenna system was placed within an artificial human head model. The antenna operated at 915 MHz and 2.45 GHz with respective gains of -28.5 dBi and -22.8 dBi. Three compact implantable antennas (Capacitively Loaded Loop (CLL), Complementary SRR loop (CSRL), and Inter-digital Capacitor Loop (IDCL)) for CSF monitoring were proposed by Manoufali et. al [92]. A 10 X10 mm² meander-line brain implantable antenna printed on 0.5 mm thick Taconic RF 35 substrate was proposed in [93] for brain-machine interface at 2.45 GHz with realized gain of -25 dBi.

For establishing near-field wireless power transfer mechanism for brain implant, another meander-line based brain-implantable antenna was designed in [94]. A compact self-triplexing slotted three-port head-implanted antenna device was designed in [95] for triple band (0.403 GHz, 0.915 GHz and 1.41 GHz)

applications radiating with respective fractional bandwidths of 48.4%, 29.8% and 22.1 % and respective realized gains of -35.7 dBi, -31.3 dBi and -28.6 dBi. A similar type of quadruplexing brain implantable antenna was designed in [96]. The antenna provided peak gain values of -36.24 dBi, -34.78 dBi, -31.39 dBi, and -30.10 dBi at 168 MHz, 433 MHz, 915 MHz and 1400 MHz, respectively.

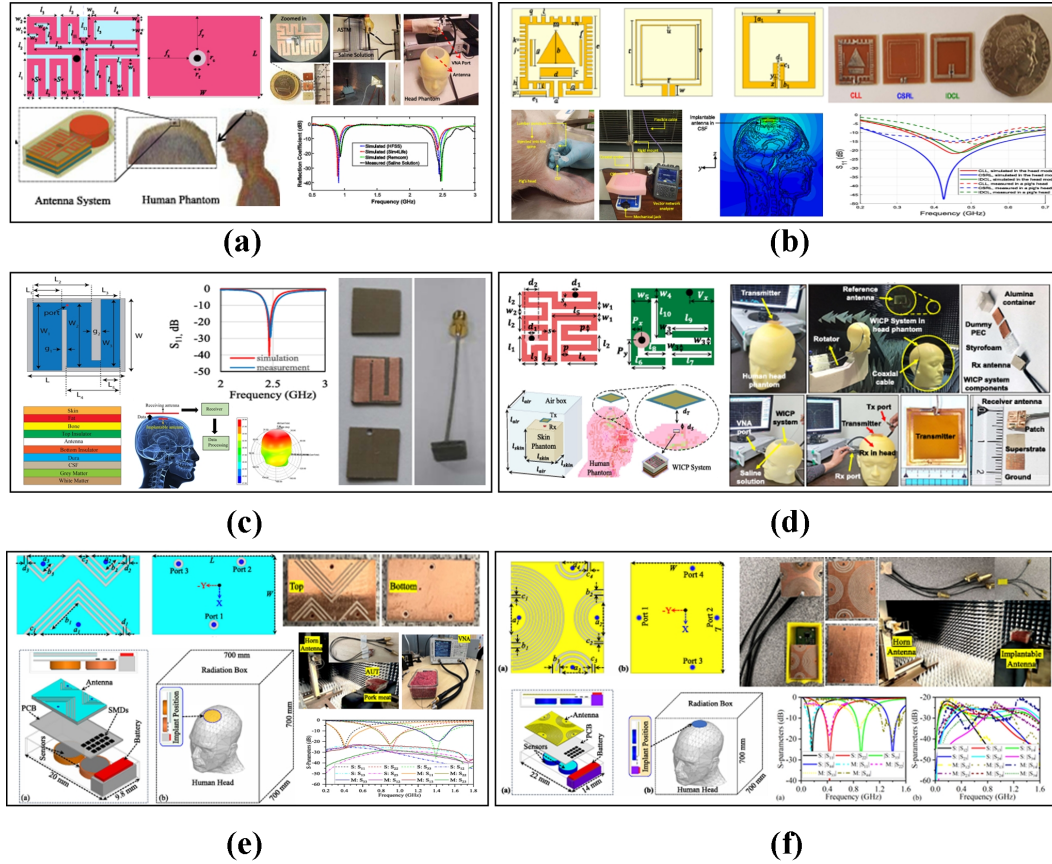


Fig. 2.16: Different brain-implantable antennas reported in [27], [92-96] respectively

Along with above mentioned applications of disease detection, implantable antennas are also used for sensing fat content, blood pressure, tumor development, lung cancer and other chronic diseases [73, 76-78].

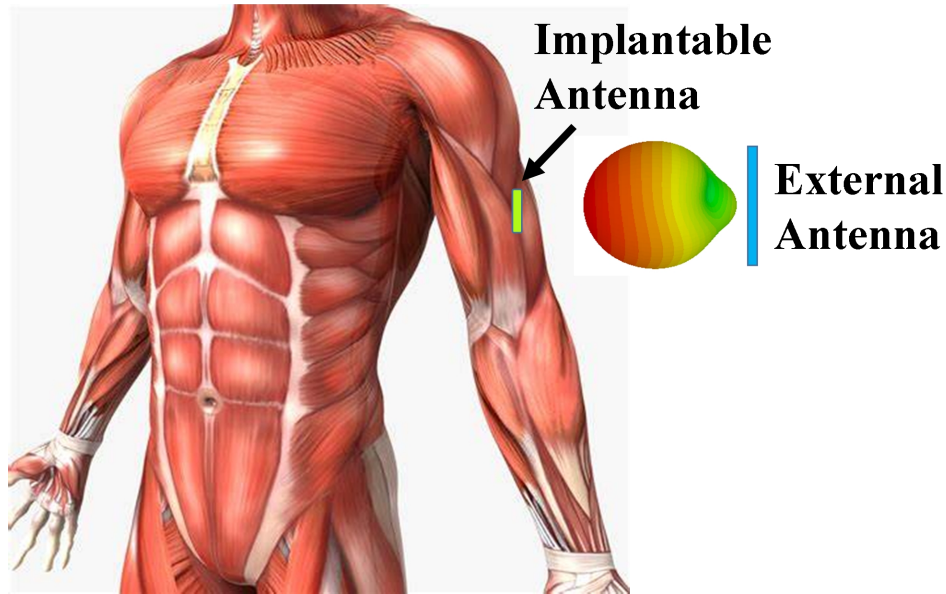


Fig. 2.17: Schematic diagram of Wireless Power Transfer to Implantable Antenna from External Antenna

2.4.3. Wireless Power Transfer to Implantable Antenna Systems:

Generally, batteries used to power the implantable medical devices, have limited lifetime. It is a major problem related to implants. Periodic surgery is required to replace these batteries to operate the devices properly. It may compromise safety for the patients. Wireless Power Transfer (WPT) can improve the battery issue and safety issue of the patient [98]. The WPT techniques can be decomposed into two types – (i) electromagnetic WPT and (ii) non-electromagnetic WPT [99].

Due to low charging efficiency, low power transfer efficiency and sensitivity to the misalignment between transmitter and receiver, non-electromagnetic WPT [100-102] is not suitable for charging batteries embedded in implantable medical devices [99, 103]. Electromagnetic WPT is classified into three types – (i) near field, (ii) mid-field and (iii) far-field electromagnetic WPT [99]. The schematic diagram of WPT to implantable antenna using electromagnetic WPT is shown in Fig. 2.17.

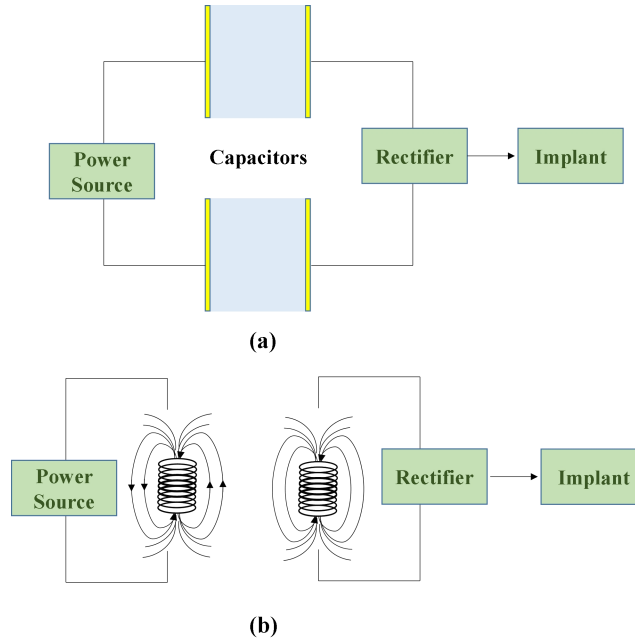


Fig. 2.18: Reactive Near-field WPT due to (a) capacitive coupling and (b) inductive coupling

(a) Reactive Near-field WPT: The field region immediately surrounding the antenna is known as reactive near-field. The range of this field region is typically considered as one wavelength from the antenna [104]. Inductive and capacitive effects are strong in this region. Capacitive coupling was used in [105-111] for transferring power to the implantable medical devices. The WPT using capacitive coupling to a skin-implant was performed in [108].

In capacitive coupling, wireless power transfer to the implantable devices is performed using two capacitors as shown in Fig. 2.18 (a). However, this technique is not applicable for deeply implanted devices because Power Transfer Efficiency (PTE) drops drastically for this system. The term PTE in percentage (%) is calculated by the equation [99]

$$PTE (\%) = \frac{|S_{21}|^2}{1 - |S_{11}|^2} \times 100$$

Here S_{21} is the transmission coefficient and S_{11} is the reflection coefficient. Generally $|S_{11}| \approx 0$. Therefore the above equation is approximated as

$$PTE (\%) = |S_{21}|^2 \times 100$$

Table 2.4: Comparative Analysis of Reported Capacitor Coupling WPT System

Ref.	Frequency (MHz)	Transfer Distance (mm)	Delivered Power (mW)	PTE (%)
[105]	1	3	12	36
[106]	6.78	75	10	0.45
[107]	434	5	100	35
[108]	308/434	5	70	47.5/46.2

Table 2.5: Comparative Analysis of Reported Inductive Coupling WPT System

Ref.	Frequency (MHz)	Transfer Distance (mm)	Delivered Power (mW)	PTE (%)
[109]	1180	6.6	0.0555	0.0019
[110]	403	6	92	42.4
[111]	27.12-40.68	2.57	-	94

The inductive coupling is also a reactive WPT technique for short-distance application. Here, two coils are placed in close proximity so that coupling takes place as shown in Fig. 2.18 (b). Due to application of power to the primary coil, the current will induce in the secondary coil due to mutual coupling. This induced current is used to provide power to the implantable devices. Table 2.4 and 2.5 shows comparative analyses for previously reported capacitor coupling and inductive coupling WPT works respectively.

(b) Radiative Mid-Field and Far-Field WPT: The limitations of above mentioned reactive WPT can be overcome by using radiative mid-field and far-field wireless power transfer techniques. These techniques are capable of transferring power to deeply implanted medical devices [112-113]. The approximate ranges of antenna's mid field and far-field regions are indicated by below equation [99, 114].

$$0.62 \sqrt{\frac{d^2}{\lambda_g}} < \text{Mid - Field} < \frac{2d^2}{\lambda_g} < \text{Far - Field}$$

Here, d is the largest dimension of antenna and λ_g is the guided wavelength. By convention, MHz to sub GHz frequency range is considered to design transmitting antenna for WPT to implantable devices. In mid-field WPT, the achieved PTE (%) value is better than far-field WPT [99].

Table 2.6: Comparative Analysis of Reported Mid-field WPT System

Ref.	Frequency (GHz)	Transfer Distance (mm)	Received Power (mW)	PTE (%)
[114]	1.47	50	6.7	0.67
[115]	1.5	55	5.6	0.56
[94]	1.9	20	-	0.257
[116]	1.7/2.1	40/10	-	0.5/3.38
[117]	2.45	50	-	1.45

Different mid-field and far-field WPT systems were previously reported in literature [114-122]. Table 2.6 shows the comparative analysis of different mid-field WPT systems. Table 2.7 depicts the comparison between different far-field

WPT systems. From these two tables, it is observed that mean PTE value is about 1.5% for mid-field WPT systems and about 0.22% for far-field WPT systems.

Table 2.7: Comparative Analysis of Reported Far-field WPT System

Ref.	Frequency (GHz)	Transfer Distance (mm)	Received Power (mW)	PTE (%)
[118]	0.915	400	0.00514	0.001626
[119]	2.4	63	-	1.22
[120]	2.45	500	3.98×10^{-3}	0.00741
[121]	2.45/5.8	80	-	0.0125/0.079
[122]	0.915	300	0.02512	0.00794

2.5 EFFECTIVE COMPLEX PERMITTIVITY MEASUREMENT FOR ACCURATE PHANTOM DEVELOPMENT

Before implanting sensors and antennas within patients' bodies, in vitro testing within artificial human body equivalent "Phantom" model for radiotherapy, photo-acoustic imaging, breast imaging, image reconstruction etc. is required [123-124]. It is a very challenging to construct multi-layered phantom due to complexity, cost and possible formation of unwarranted air gaps between two layers or mixing chemicals of two layers while one layer is placed on another layer in hot semi-solid form. However, this problem can be resolved by using its equivalent homogeneous phantom. There is no possibility of generating air gaps inside homogeneous phantom; cost and complexity of designing homogeneous phantom are comparatively lower than multilayered model.

In multilayered phantom, while electromagnetic (EM) wave propagates through different layers, internal reflections occur in each interface between two layers. When homogeneous phantom is used, these internal reflections cannot be

accounted for. Therefore, it is important to estimate equivalent complex permittivity of multilayered phantom such a way that amount of overall transmission and reflection of EM waves for equivalent phantom should be same as those for multilayered model.

Resonant-cavity method and waveguide method are widely utilized for characterizing complex permittivity of different materials [125-126]. In resonant cavity method, the complex permittivity can be obtained by cavity perturbation from the changes in quality factor and resonant frequency [127]. This method provides high accuracy for the low loss materials with low permittivity [128]. Human tissues are highly lossy with high dielectric constant (in most of the cases); therefore, resonant cavity method is not suitable in this case.

The waveguide method is categorized into two types: resonant [129] and non-resonant [130] methods. Proper machining is required here for high accuracy. The material under test (MUT) should be kept within a metallic chamber for these two cases. These methods show low accuracy while the sample length is the multiple of $1/2$ wavelength.

Free-space measurement method [131-132] is more efficient than previous methods because direct contact between MUT (kept within sample holder) and active components of the setup is not required. However, cost of the sample holder is high as well as the fact that the sample should be large and flat for obtaining high accuracy [133-134]. Diffraction effect at the edge of MUT has been observed here which leads to another disadvantage of this method.

The most widely used method of dielectric constant measurement is open ended coaxial probe technique [135-136], where an open ended coaxial probe is placed within/on the material and Vector Network Analyzer (VNA) is utilized for recording reflection coefficient to obtain permittivity and loss tangent measurement. This technique cannot provide proper result while some air gap is present between coaxial probe and MUT.

Nicolson-Ross-Weir (NRW) method [137], NIST iterative method [138], Short Circuit Line (SCL) method [139] etc. are some other mathematical methods. For these methods, the MUT is kept within a waveguide and dielectric constant of the MUT is calculated by scattering parameters (S_{11} , S_{12} , S_{21} and S_{22}) [140]. NRW method is fast and non-iterative method; however this method is not appropriate for materials with low loss and large size [141]. NIST iterative method shows better accuracy than NRW method but NIST is applicable for real relative permittivity measurement only [142]. Therefore, these techniques cannot be used for effective complex permittivity measurement of human body model.

There are two approaches to calculate effective complex permittivity of multilayered model in the view of EM wave interaction – based on scattering parameters (related to power density) and based on propagation constant [143]. However for any implantable antenna, both propagation constant and scattering parameters are important for communication channel characterization.

2.6 CONCLUSION

This chapter presents a review of modeling techniques of different types of implantable antennas reported for different biomedical applications. Scientists used particle swarm optimization along with FDTD and FEM simulators for designing their antennas within human body model. These techniques may predict the antenna responses efficiently. However, high time complexity and huge RAM usage are the major drawbacks of these methods. In previously reported works, planar, conformal and helical type implantable antennas were designed at different Federal Communications Commission (FCC) approved medical frequencies such as MICS band (402-405 MHz), MedRadio band (401-406 MHz), WMTS band (1427-1432 MHz), ISM band (2.4-2.48 GHz) etc. The engineers were widely focused on biotelemetry applications for leadless pacemaker,

wireless capsule endoscopy, blood sugar, pressure monitoring, head implants, wireless power transfer etc.

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Chapter 3

Transmission Line Model for Implantable Antennas

Contents	3.1. Introduction 3.2. Problem Formulation 3.3. Transmission Line Model of Right Angle Meander-Line Implantable Antenna 3.4. TLM for Miniaturized Meander-Line Implantable Antenna 3.5. TLM for Arbitrary Bended Meander-line Implantable Antenna 3.6. Comparison with Previously Reported Works 3.7. Conclusion
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3.1 INTRODUCTION

The antenna of an implantable device plays a crucial role in providing communication with the external environment to transfer recorded health-related data such as blood sugar level, pressure level, intraocular pressure, pulse rate, etc. for diagnosis [1-3]. Implantable antenna has to deal with some design considerations such as operating frequency band, miniaturized structure for proper implantation, Specific Absorption Rate (SAR) to ensure tissue safety, etc. According to some standard guidelines [4-5], maximum SAR averaged over a 1 g cubical shaped contiguous tissue should be less than 1.6 W/kg and should be less than 2 W/kg for 10 g tissue model.

In previous works, different miniaturized implantable antennas have been proposed. Meandering, waffling, spiraling etc. are popular techniques for antenna miniaturization by enhancing its effective current path length and thereby ensuring compactness of the implantable antenna [6]. In [7-9], shorting wall has

been utilized for miniaturization by 50%. The resonant frequency can also be reduced by placing thick superstrate on patch.

Superstrate reduces effective permittivity by insulating the implantable antenna from highly lossy human tissue which has high permittivity too. Superstrate with higher permittivity value can also reduce the phase velocity of the electromagnetic wave propagating through the material. This results in lowering the resonant frequency by shortening the effective wavelength [10]. Comparative analysis between serpentine and spiral radiators have been studied in [11] in terms of the effect of length, feed location, substrate and superstrate thicknesses and their materials.

Different techniques have been followed by researchers to design previously reported implantable antennas. Particle swarm optimization technique was used by Macon et al. [12] to model implantable antenna at MICS (Medical Implant Communication System) and ISM bands. Combination of Finite Element and Boundary Integral techniques has been used by Yu et al. [13] for designing conformal antennas. Full wave analyses technique based FDTD (Finite Difference Time Domain) method was utilized in [11]. The authors of [20-22] have used the FDTD technique for designing meander-line and right-angled bend spiral implantable antennas. Besides, most of the researchers have designed implantable antennas with the Finite Element Method (FEM) microwave simulators using trial-and-error method, which may increase the time complexity and storage usage of RAM. In this chapter, we intend to use transmission line model for designing and optimizing meander-lined implantable antenna at 2.45 GHz. This technique can be used for other shapes (spiral, Hilbert line etc.) also.

3.2 PROBLEM FORMULATION

The first part of this chapter deals with right-angle bend meander-line body-implantable antenna as shown in Fig. 3.1 (a). Reflection coefficient and

radiation characteristics of this implantable antenna are calculated. Then, the model is extended to arbitrary bends (with acute and obtuse angles) as shown in Fig. 3.1 (b). Finally transmission line model of corner-chamfered meander-line implantable antenna (refer to Fig. 3.1 (c)) is formulated.

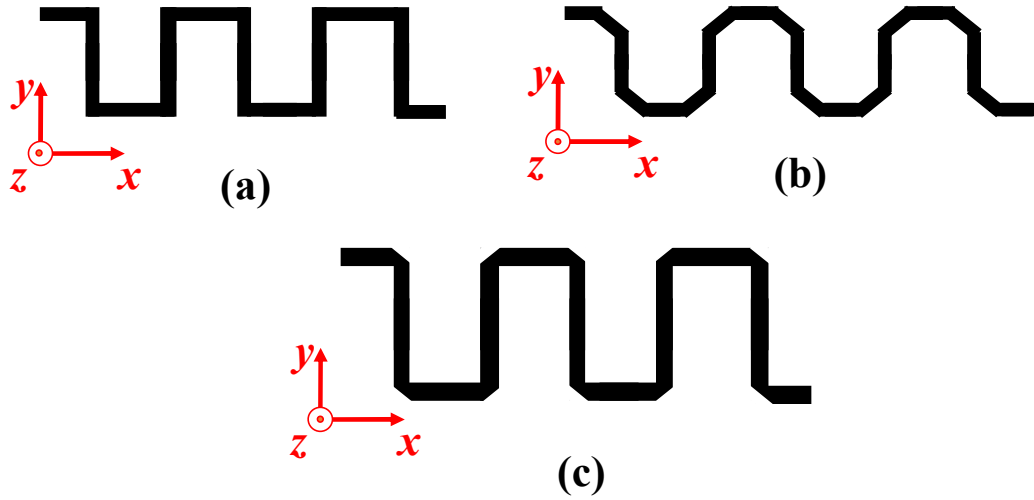


Fig. 3.1: (a) Right angle bended, (b) arbitrary bended and (c) corner chamfered meander-line implantable antenna geometries

3.3 TRANSMISSION LINE MODEL OF RIGHT ANGLE MEANDER-LINE IMPLANTABLE ANTENNA

3.3.1: Design of Implanted Antenna within Body Model:

Figs. 3.2 (a) and (b) show top and bottom views of the preliminary implantable antenna structure respectively. This is a simple right angled bend-meander line antenna. This antenna has been chosen to build up primary knowledge of transmission line analysis for an implantable antenna. Here, copper is used as the conductor parts such as radiator and ground part. Arlon AD 430 ($\epsilon_r = 4.3$ and $\tan \delta = 0.003$, thickness = 0.762 mm) is utilized here for both

superstrate and substrate. We are considering that our antenna is placed within muscle layer of three layered simplified human body model shown in Fig. 3.2 (c). This model consists of skin, fat and muscle (outside to inside) and their electrical properties are taken for 2.45 GHz ISM band based on Gabriel's database [14-15] as mentioned in Table 3.1. As the communication link between implantable antenna and an external device is in the broadside direction of the muscle implantable antenna and bone is not present in that direction, bone is not considered here for characterization of the communication link. Some tissues like blood, blood vessels, nerve etc. are neglected because of their insignificant effect in terms of electrical properties with respect to other layers.

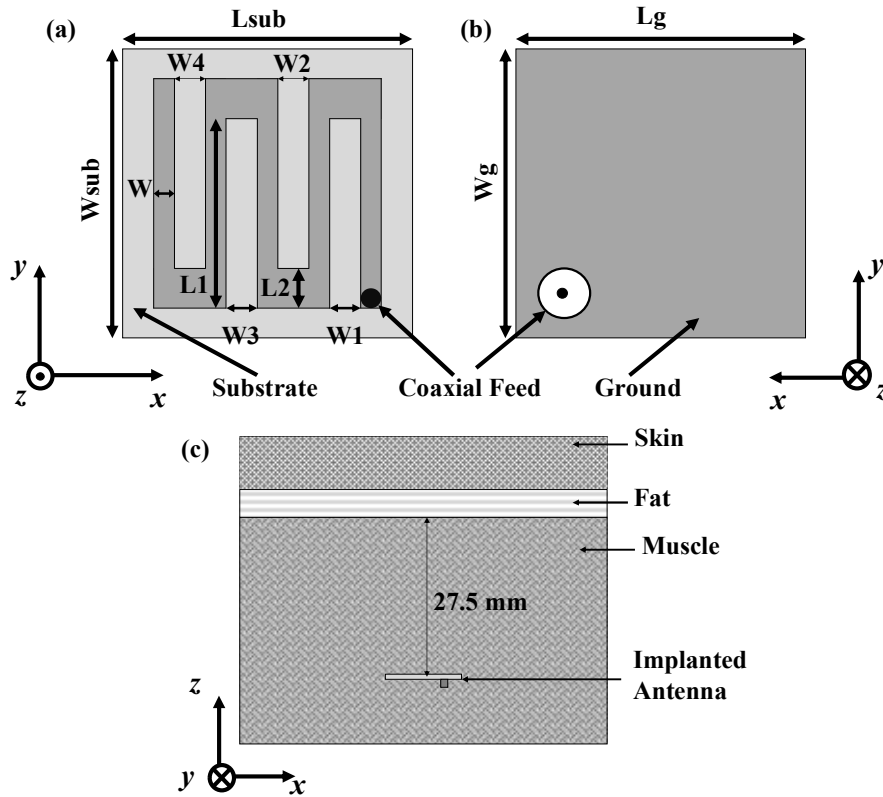


Fig. 3.2: Top and (b) bottom views of antenna where dimensions are in mm: $L_{sub} = L_g = 24.21$, $W_{sub} = W_g = 25.55$, $W1 = W2 = W3 = W4 = 2.7$, $L1 = 17$, $L2 = 3$ and $W = 2$; (c) side view of the implantable antenna within human phantom.

Table 3.1: Electrical Properties of body layers at 2.45 GHz

Layer	Relative Permittivity	Conductivity (S/m)
Skin	38	1.46
Fat	10.8	0.268
Muscle	52.7	1.78

The electrical nature of the muscle layer is complex and its relative permittivity can be written as,

$$\epsilon_r^* = \epsilon_r' - j\epsilon_r'' \quad (1)$$

where $\epsilon_r' = 52.7$ and $\epsilon_r'' = \epsilon_r' \times \tan \delta = 14.23$. Therefore, magnitude of ϵ_r^* is approximately 54.58. The wavelength in free space is $\lambda_o = c/f$, where $c = 3 \times 10^8$ m/s. So, For 2.45 GHz λ_o is obtained as 0.1224 m. Now, wavelength in muscle will be,

$$\lambda_m = \frac{\lambda_o}{\sqrt{|\epsilon_r^*|}} \quad (2)$$

Therefore, the value of λ_m is 0.0203m. The value of implantation depth is taken as 27.5 mm so that the interface between muscle and fat layers will be at the far-field region of implanted antenna. Thicknesses of skin and fat layers are taken as 4 mm and 2 mm respectively.

3.3.2: Transmission Line Model:

3.3.2.1: *Antenna in Free Space*

At first, the transmission line model of meander-line antenna is developed by considering it in free space. Meander-line structure is subdivided into 17 segments as shown in Fig. 3.3 (a). Each segment is considered as a transmission line and lumped circuit element values for each segment are calculated using

TLM. The even numbered segments are bend regions and odd numbered sections are divided into two types of line segments – x-directed segments (3, 7, 11 and 15) and y-directed segments (1, 5, 9 and 17). Each x or y- directed line segment can be considered as a transmission line and is equivalent to a lumped T network as shown in Fig. 3.3 (b). The lumped equivalent circuits for the right angled bends can also be considered as T network as shown in Fig. 3.3 (b). The values of lumped inductance and capacitance L_n and C_n respectively for the n^{th} segment (when $n = \text{odd}$ for this case) can be written using [16] as,

$$L_n = \frac{Z_o l_n \sqrt{\epsilon_{eff}}}{2c} \quad (3)$$

$$C_n = \frac{l_n \sqrt{\epsilon_{eff}}}{Z_o c} \quad (4)$$

where Z_o is the characteristic impedance of line segment, l_n is the length of this segment, c is the velocity of light and ϵ_{eff} is the effective permittivity.

In 1978, Garg et al. [17] have given empirical formulas for shunt capacitance and series inductance of the TLM equivalent circuit for right angled bend. Here we have used the expressions given by [17] for even segments which are bend sections, as

$$L_n = 100h \left\{ 4 \sqrt{\frac{W_n}{h}} - 4.21 \right\} (nH) \quad (5)$$

$$C_n = W_n \left\{ (9.5\epsilon_r + 1.25) \frac{W_n}{h} + 5.2\epsilon_r + 7 \right\} (pF) \quad (6)$$

, For $\frac{W_n}{h} > 1$

$$C_n = W_n \left\{ \frac{(14\epsilon_r + 12.5) \frac{W_n}{h} - 1.83\epsilon_r + 2.25}{\sqrt{\frac{W_n}{h}}} + \frac{0.02\epsilon_r}{\frac{W_n}{h}} \right\} (pF) \quad (7)$$

, For $\frac{W_n}{h} < 1$

According to [17], accuracy of eq. (5) is around 3% for $0.5 \leq \frac{w_n}{h} \leq 2$ whereas accuracy of (6-7) is 5% for $0.1 \leq \frac{w_n}{h} \leq 5$. Using the equations (5-7), lumped T network elements for all segments are calculated as follows.

- For sections 1 and 17, $L = 2.16$ nH and $C = 2.49$ pF.
- For x-directed sections i.e., for sections 3,7,11 and 15, $L = 0.53$ nH and $C = 0.53$ pF.
- For y- directed sections 5, 9 and 13, $L = 1.78$ nH and $C = 2.05$ pF.
- For bend sections, $L = 0.28$ nH and $C = 0.59$ pF.

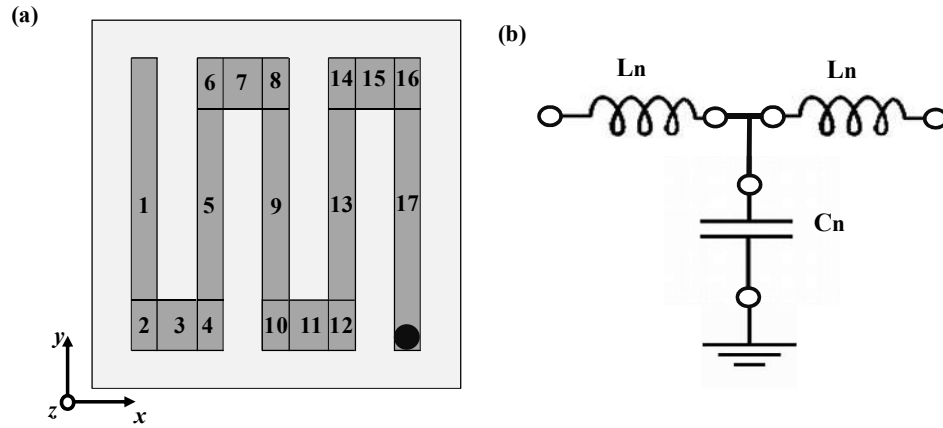


Fig. 3.3: (a) Segmentation of meander-lined radiator. (b) T network model for n^{th} segment.

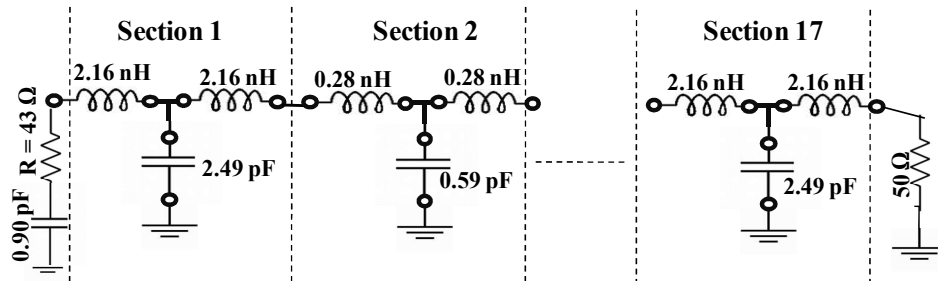


Fig. 3.4: Lumped circuit representation of the antenna in free space

A variable capacitance (C_{open}) is added at the open end of the meander-line structure and a resistance (R) is connected in series with it to represent the ohmic loss. The equivalent circuit for the antenna structure (shown in Fig. 3.4) is designed and simulated. The value of C_{open} is obtained as 0.90 pF. Initially the value of R is arbitrarily chosen as 43 Ω . The return loss response of the circuit shown in Fig. 3.4 is shown in Fig. 3.5 (a). The return loss value obtained from this plot is 14.82 dB at 3.22 GHz. Further, this antenna is simulated using CST Microwave Studio 2019 software. The simulated return loss (Fig. 3.5 (a)) of this antenna is observed as 27.06 dB at 3.27 GHz. From this plot, it can be concluded that the estimated circuit model using TLM provides accurate equivalent response of the original antenna structure.

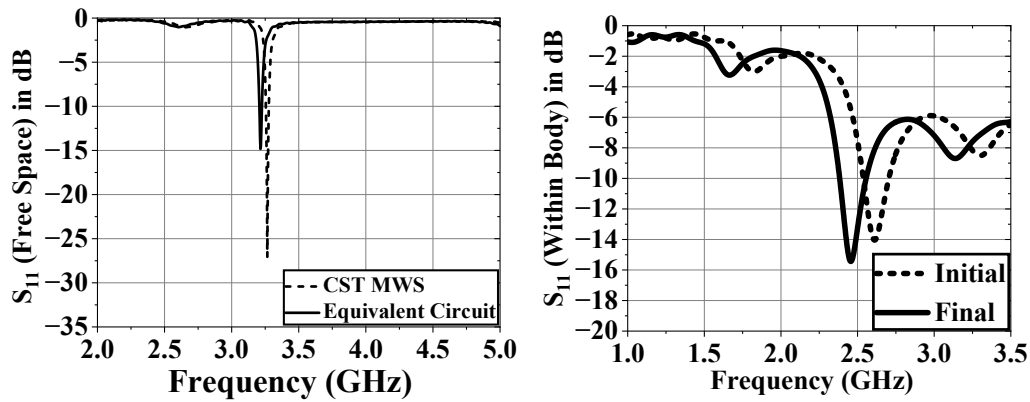


Fig. 3.5: Reflection coefficient responses for (a) antenna in free space and (b) antenna within three-layered human body

3.3.2.2: Antenna within Human Body Model

Next, the antenna is placed within three-layered human body model and the system shown in Fig. 3.2 (c) is simulated in CST. The return loss value of this system is 14.04 dB at its operating frequency 2.61 GHz (please see Fig. 3.5 (b)-‘Initial’). A RLC type network is cascaded with lumped circuit model in between

section 17 as equivalence of human body model and the termination (please refer to Fig. 3.6) and tuned to match return loss response with the curve shown in Fig. 3.5 (b-‘initial’) so that the RLC network may provide equivalent effect of the three-layer body model. Therefore, the values of R, C and L values are obtained as $31\ \Omega$, $1.72\ \text{pF}$ and $2.88\ \text{nH}$ respectively. The selection of a T-network (consisting of two series arms and one shunt arm) is not arbitrary. It represents a lumped-element approximation of a short section of a lossy transmission line or a series of material boundaries. Series elements model the ohmic losses and magnetic energy storage (inductance) associated with the physical path the current takes when transitioning from the antenna through the superficial tissue layers. Shunt element models the transverse displacement currents (capacitance) and dielectric losses (conductance) of the tissues acting as a lossy dielectric medium under the antenna.

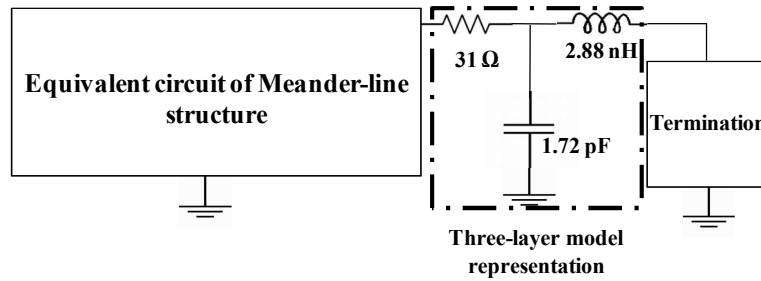


Fig. 3.6: Equivalent circuit representation of antenna inside body model

In this work, we have taken T-type RLC network as equivalent of human body. If the depth of implantation or body layer properties changes, the values of R, L and C will also be changed due to change in effective complex permittivity of communication channel between the implanted antenna and external device. According to the S_{11} plot of implantable antenna within human body model, the antenna is not tuned at 2.45 GHz. Therefore, the antenna architecture should be optimized properly. The lumped circuit parameters related to antenna dimensions are optimized to tune the return loss response at 2.45 GHz keeping values of RLC

network elements fixed as it is desired to optimize antenna structure without changing any property of human body model for this step. Table 3.2 contains the optimization range and optimized values of different circuit elements using TLM.

Table 3.2: Optimization Parameters

Parameter Name	Range	Optimized value
Resister at open end	20.79-100 Ω	89.7 Ω
Inductor at Section 3	0.13-0.38 nH	0.15 nH
Capacitor at Section 3	0.27-0.79 pF	0.66 pF
Inductor at Section 7	0.13-0.38 nH	0.22 nH
Capacitor at Section 7	0.27-0.79 pF	0.42 pF
Inductor at Section 11	0.12-0.36 nH	0.22 nH
Capacitor at Section 11	0.27-0.79 pF	0.63 pF
Inductor at Section 15	0.07-0.40 nH	0.13 nH
Capacitor at Section 15	0.01-0.45 pF	0.29 pF

Using Table 3.2, some modifications in physical dimensions of meander-line antenna structure are performed. At initial stage, all the values of W1, W2, W3 and W4 were taken as 2.7 mm. After optimization their values are obtained as 1.5 mm, 2.49 mm, 2.61 mm and 2.61 mm respectively. The return loss curve of optimized antenna structure using TLM within human body as simulated in CST is shown in Fig. 3.5 (b) which shows that this system is operable at 2.45 GHz with return loss value of 15.44 dB.

3.4 TLM FOR MINIATURIZED MEANDER-LINE IMPLANTABLE ANTENNA

In previous section, simple meander line implantable antenna at 2.45 GHz has been designed using TLM. Area of the radiator of the designed implantable

antenna is 408.2 mm^2 . This antenna is too large to place it within human body; it should be further miniaturized. If number of meanders is increased, the antenna size will be small due to increment of current path length. In this stage, our objective is to reduce the radiator size by taking respective lengths of vertical lines as 8 mm and length of right angled bends as 0.6 mm (Please see Fig. 3.7(a)).

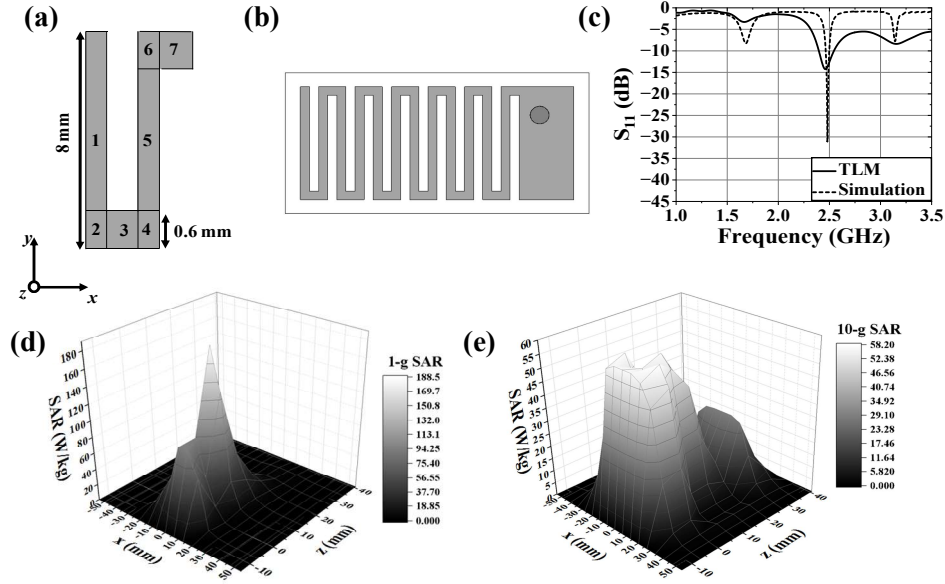


Fig. 3.7: (a) Single meander line structure for miniaturization; (b) Final structure of miniaturized meander lined antenna; (c) return loss responses of TLM and simulation; Simulated 3D SAR distribution due to 1W average input power for (d) 1-g and (e) 10-g contiguous tissue model at 2.45 GHz when SAR calibration line is positioned at $y = 0 \text{ mm}$.

The TLM equivalent circuit element values of the meander structure shown in Fig. 3.7(a) are calculated using equations (5-7). Cascading six similar sub-circuits, the $<10 \text{ dB}$ return loss is obtained at 2.45 GHz as shown in Fig. 3.7(c). Therefore, six meander line segments are cascaded to obtain final antenna structure (See Fig. 3.7(b)). The final result is a six meander lined antenna (please refer Fig. 3.7 (b)) which has return loss value of 31.03 dB at 2.48 GHz (see Fig.

3.7(c)). The area of the radiator is 144 mm^2 . Therefore, this antenna is miniaturized enough compared to the previous antenna.

Quality factor is an important parameter to characterize the performance of implantable antenna within human body in terms of efficiency. The quality (Q) factor is defined as [18],

$$Q = \max\left(\frac{2\omega W_E}{P_{rad}}, \frac{2\omega W_H}{P_{rad}}\right) \quad (8)$$

where ω is angular frequency, W_E is average energy stored due to electric field and W_H is average energy stored due to magnetic field. P_{rad} is the total power radiated from antenna. There are basically two types of Q factors depending upon the implantable antenna environment – unloaded and loaded Q factors. Loaded quality factor relates to the antenna's efficiency while it is placed within human body and unloaded quality factor determines the antenna efficiency in free space. Loaded Q factor takes into account of dielectric properties of the surrounding tissue because they have influence on implanted antenna performances such as resonant frequency, gain etc. The quality factor can also be represented as,

$$Q = \frac{f_r}{BW} \quad (9)$$

where f_r is resonant frequency and $BW = -10 \text{ dB}$ bandwidth obtained from reflection coefficient plot. In this work, resonant frequency of the antenna is 3.26 GHz and its bandwidth is 60 MHz. Therefore unloaded quality factor (Q_u) of this antenna is obtained as 54.33. The bandwidth of the antenna after implantation is reduced to 50 MHz around 2.45 GHz. Thus, loaded quality factor (Q_l) is obtained as 49. Therefore, $\frac{Q_l}{Q_u} = 0.9 < 1$. As the value of this ratio is less than 1, it implies higher loss of energy due to lossy tissues. In spite of having lossy surrounding tissues, this ratio is closer to 1 which implies that the antenna in Fig.

3.7(b) provides good performances within human body. The respective peak values of SAR for our design within three-layer body model are 220.26 W/kg and 59.73 W/kg for 1-g and 10-g contiguous tissue models for application of 1 Watt or 30 dBm average input power. The maximum average input power is chosen as 7 mW and 24 mW for over 1g and 10g of body model respectively to keep the SAR within standard prescribed limits. Figs. 3.7 (d) and (e) present simulated 3D 1g SAR distribution and 10-g SAR distribution at 2.45 GHz for the designed implantable antenna within muscle tissue layer respectively for 1 Watt average input power application. A comparison between TLM technique and FEM technique in terms of time and memory requirement is shown in Table 3.3. Around 85.65% memory requirement and 93.33% time consumption are reduced by using TLM instead of FEM technique.

Table 3.3: Comparison between TLM and FEM Techniques

Technique Name	Time Taken	Memory Taken
TLM (using MATLAB)	52s	726 kb
FEM (using CST)	13 min	4.94 Mb

3.5 TLM FOR ARBITRARY BENDED MEANDER-LINE IMPLANTABLE ANTENNA

In this section, Transmission Line Model (TLM) of arbitrary bended meander-line implantable antenna is discussing here. While right-angle bend is considered, the bend section (PP') is considered as a rectangular segment (referring to Fig. 3.8 (a-b)). However, for arbitrary angle bend (shown in Fig. 3.8 (c)), the bend section is not taken as rectangular segment. The non-right angle bends can be either acute angle bend (refer to Fig. 3.9 (a)) or obtuse angle bend

(refer to Fig. 3.9 (b)). For non-right angle bend, the empirical formulas (5-7) of equivalent circuit model are to be modified.

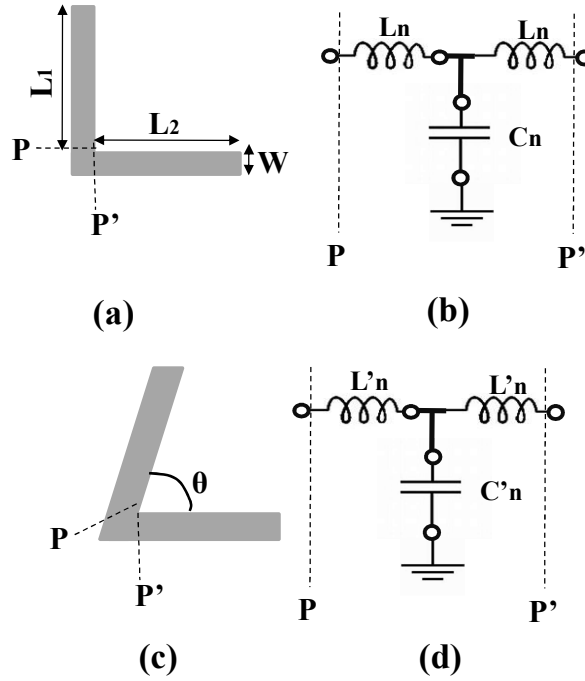


Fig. 3.8: Right angle bend: (a) geometry and (b) lumped circuit model; Arbitrary angle bend: (c) geometry and (d) lumped circuit model

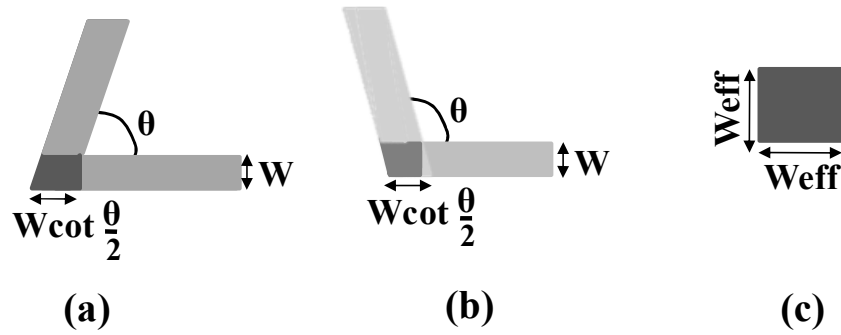


Fig. 3.9: Arbitrary bends: (a) acute and (b) obtuse angle; (c) effective rectangular representation of arbitrary bend section

Here, an equivalent rectangular segment with effective width W_{eff} is considered (shown in Fig. 3.9 (c)) in place of a non-right angle bend in such a way that area of bend section for non-right angle bend is equal to the area of the effective rectangular bend. If the angle of the bend is θ , the effective width (W_{eff}) will be,

$$W_{eff} = W \sqrt{\cot \frac{\theta}{2}} \quad (10)$$

The inductance and capacitance are dependent on width (W_n) and height of substrate (h). The expressions (5-7) are modified by replacing the term W_n by W_{eff} and introducing an empirical term k .

$$L'_n = 100h \left\{ 4 \sqrt{\frac{W_{eff}}{h}} - 4.21 \right\} (nH) \quad (11)$$

$$C'_n = W_{eff} \left\{ (9.5\epsilon_r + 1.25) \frac{W_{eff}}{h} + (5.2 + k)\epsilon_r + 7 \right\} (pF) \quad (12)$$

, For $\frac{W_{eff}}{h} > 1$

$$C'_n = W_{eff} \left\{ \frac{(14\epsilon_r + 12.5) \frac{W_{eff}}{h} - (1.83 + k)\epsilon_r + 2.25}{\sqrt{\frac{W_{eff}}{h}}} + \frac{0.02\epsilon_r}{\frac{W_{eff}}{h}} \right\} (pF) \quad (13)$$

, For $\frac{W_{eff}}{h} < 1$

Expression of k is mentioned below using curve fitting [20],

$$k = \begin{cases} 25.21 - 25.21 \left(\frac{\theta - \frac{\pi}{4}}{\frac{\pi}{4}} \right), & \frac{W_{eff}}{h} < 1 \text{ and } 0 < \theta < 90^\circ \\ 4.55 \left(\frac{\theta - \frac{\pi}{2}}{\frac{\pi}{2}} \right), & \frac{W_{eff}}{h} < 1 \text{ and } 90^\circ < \theta < 180^\circ \\ 25.21 \cot(1.145\theta - 0.2281), & \frac{W_{eff}}{h} > 1 \end{cases}$$

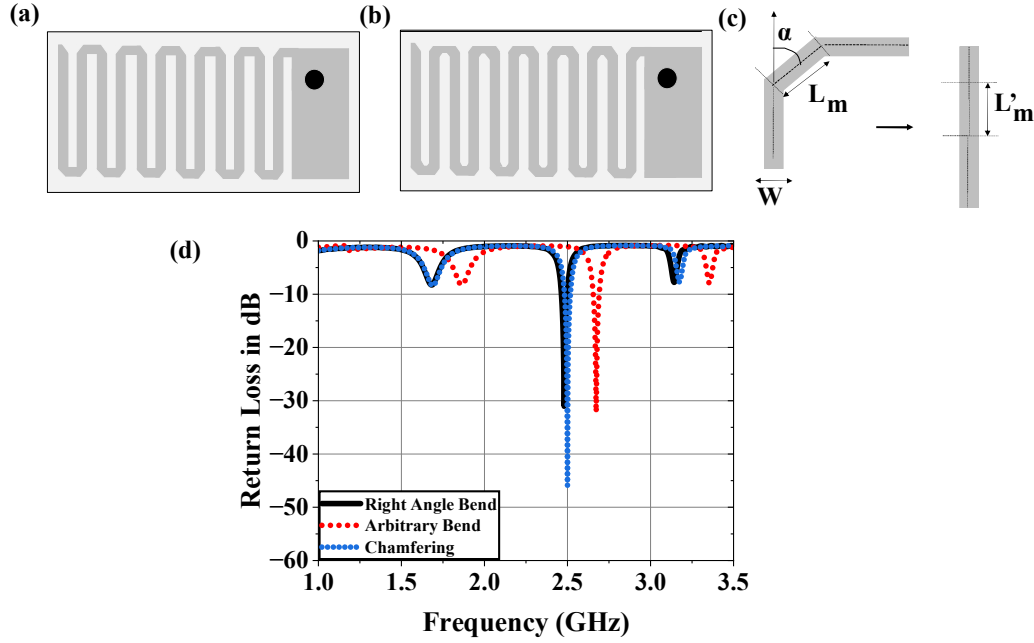


Fig. 3.10: Bending effects: (a) Chamfering; (b) Tilted Bends; (c) Analysis of tilted bend sections using TLM; and (d) Comparison of Return loss plots of non-right angle bend with right-angle bended implantable antenna

Chamfering of the bends reduces the stray radiated emission to minimize losses and unwanted reactance [20]. The effect of chamfering is observed here. The bend sections of the antenna are chamfered as shown in Fig. 3.10 (a) and consequent return loss plot is shown in Fig. 3.10 (d). The return loss value is increased to 45.87 dB at 2.49 GHz. The return loss value is improved due to reduction of reactance part. The realized gain is slightly reduced to -35.48 dBi at 2.49 GHz. Next, effect of bend sections tilted at arbitrary angle α (refer to Fig. 3.10 (b)) is studied and resulting return loss response is illustrated in Fig. 3.10 (d). For analysis, the bend section is replaced by a rectangular bend section as shown in Fig. 3.10 (c). The width of the m^{th} effective rectangular bend is taken same as width of the tilted bend and the length (L'_m) is calculated as

$$L'_m = L_m \cos \alpha \quad (14)$$

As a result, the operating wavelength is changed. This may lead to change in inductive and capacitive effect (based on Equations 3-4). This antenna is simulated by considering $\alpha = 30^\circ$. The frequency of operation is shifted to 2.67 GHz with return loss value 32.14 dB and realized gain of -36.31 dBi.

3.6 COMPARISON WITH PREVIOUSLY REPORTED WORKS

As per our survey, TLM is used to design and analyze any implantable antenna for the first time by us. Therefore, it is not possible to compare this work with similar works. Scientists have followed different optimization techniques and trial and error method for designing this kind of antennas. It is time consuming and the RAM usage is very high because the simulating software needs to simulate the whole system (implanted antenna with body model) several times. In [13, 21-23], FDTD technique is used by authors.

Table 3.4: Comparison with literature [21]

Number of cells in Ref.	Values
[21]	3182656
This Work	2314130

Number of cells taken to analyze meander-line antenna implanted within homogeneous human body model is 3182656 which takes huge time to obtain results of the system if the time step increases. In this work, TLM equivalent circuit is used. Hence, much less time is taken for analyze and optimize the antenna structure within human body model. Table 3.4 shows a comparison with

[21] in terms of cells in mesh. From this table, it can be observed that greatly reduced number of cells is used in this work which can reduce time consumption and RAM usage. The reason behind any resonant frequency shift or impedance mismatch or loss due to any part of antenna can be easily identified by looking into change in values of equivalent circuit element which is enough for analysis of the implantable antenna. Along with that, we are representing human body effects here by simple RLC network. That's why; we are able to reduce simulation complexity of 3D human body model by only observing the lumped element values in 2D plane. From this point of view, our proposed technique is advantageous over the present methods. Furthermore, representation of human body effect with a simple RLC network is also an important outcome of this work.

While the Transmission Line Model (TLM) offers exceptional computational efficiency, its accuracy degrades significantly for complex geometries and higher frequencies because it reduces a three-dimensional electromagnetic problem to a one-dimensional quasi-TEM approximation. At higher frequencies, the model fails as the substrate thickness or strip width approaches a significant fraction of the wavelength, triggering frequency-dependent dispersion, surface waves, and higher-order modes that empirical lumped elements cannot capture. Similarly, in complex or tightly packed geometries like non-uniform meanders, the localized evanescent fields of adjacent bends physically overlap and generate strong parasitic near-field coupling; because the empirical formulas from literature assume isolated discontinuities, they completely miss this mutual interaction. Consequently, when the structural dimensions or tight trace spacing violate the assumptions of isolated, uniform segments, full-wave numerical methods (like FEM or FDTD) become strictly necessary to capture the true radiation and impedance characteristics.

3.7 CONCLUSION

The transmission line modeling of meander line implantable antenna for ISM band is demonstrated in this chapter. Equivalent lumped circuit has been desired for the meander line antenna. The effect of human body layers on the performance of implanted antenna has been replicated by cascading simple T networks of resistor, inductor and capacitor. Miniaturization of the meander-lined implantable antenna has also been achieved using transmission line modeling technique. The analyses of chamfered and tilted bend sections of any implanted meander-line antenna using TLM are also performed here. The TLM approach in designing and analysis of an implantable antenna is a novel approach with several advantages which has not been performed yet as per our knowledge. Around 85.65% memory requirement and 93.33% time consumption are reduced by using TLM instead of FEM technique.

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Chapter 4

Mean Complex Permittivity Estimation of Human Body Model

Contents	4.1. Introduction 4.2. Problem Formulation 4.3. Background of Complex Permittivity Measurement 4.4. General Concept of Transmission Matrix Method 4.5. Proposed Complex permittivity measurement method of N-layer Model using TMM 4.6. Complex permittivity estimation and validation of Different body models at different conditions 4.7. Comparison of TMM with realistic voxel body model simulation 4.8. Verification of Frequency Independency 4.9. Accuracy Analysis 4.10. Comparison with Previously Reported Works 4.11. Conclusion
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4.1 INTRODUCTION

Implantable antennas and biosensors in implantable medical devices can sense and send continuously monitored health parameters such as glucose level, heart rate, brain related parameters etc. wirelessly to external monitoring devices for smart patient monitoring [1-3]. For radiotherapy, photo-acoustic imaging, breast imaging, image reconstruction, and other applications, in vitro testing using an artificial human body comparable "phantom" model is necessary prior to implanting sensors and antennas within patients' bodies [4-5].

Building a multi-layered phantom is extremely difficult because of its complexity, expense, and potential for creating unnecessary air gaps between two layers or combining chemicals from two layers when one layer is placed on top of another in a heated, semi-solid state. However, by employing its equivalent

homogeneous phantom, this issue can be solved. Since there is no chance of creating air holes inside a homogeneous phantom, its design is less complicated and less expensive than a multilayered model. Internal reflections take place at every interface between two layers in a multilayered phantom, when electromagnetic (EM) waves travel through many layers. It is impossible to account for these internal reflections when homogeneous phantom is utilized. Thus, it is crucial to estimate the multilayered phantom's equivalent complex permittivity in a way that ensures total transmission and reflection of EM waves for the equivalent phantom and the multilayered model are equal. If the entire system is replaced with a single, homogeneous layer of the same total thickness, that new model must possess a specific complex permittivity that yields the exact same scattering parameters as the multi-layered model.

A true average (like a volume-weighted arithmetic mean) assumes that every part of the medium contributes equally to the wave behavior, regardless of position or wave direction. In electromagnetics, this is rarely true due to two main factors: (i) if the electric field is parallel to the layer boundaries, the fields add differently than if the E-field is perpendicular to them. A simple average completely ignores field orientation; (ii) waves decay as they travel through lossy media. The first layer encounters the strongest field and contributes far more to the reflection property than a deeper layer. A true average treats a buried layer and a surface layer identically, which fails to capture reality. Therefore, it is explicitly a phenomenological value—it is the value that makes the external wave behavior of the simplified model match the complex reality.

The Federal Communications Commission (FCC) has authorized a number of frequency bands for use in biomedical applications, including the MICS band (402-405 MHz), MedRadio band (401-406 MHz), WMTS band (1427-1432 MHz), and ISM band (2.4-2.48 GHz) [6]. To test implantable antennas for various purposes, such as breast cancer detection, brain-machine

interface, retinal, and cardiovascular applications, several kinds of phantom models are needed.

4.2 PROBLEM FORMULATION

Transmission Matrix Method (TMM) is used here to estimate the effective complex permittivity of multilayered human body models in order to create homogeneous phantoms. A N-layer generalized multilayer model's complex permittivity is first calculated using TMM, taking into consideration the reflection and transmission coefficients in the junctions between any two layers caused by an impinged electromagnetic wave with a peak value of " E_o " that propagates from the outside to the multilayer model. Using CST Microwave Studio software, a homogeneous phantom with estimated complex permittivity is constructed. Then, in compliance with practical specifications, an implanted antenna is constructed inside this phantom. The effectiveness of complex permittivity prediction using the proposed method is observed by comparing the responses in the related multilayered model with the simulated radiation responses of the homogeneous phantom-implanted antenna in terms of S_{11} , gain, and operating frequency.

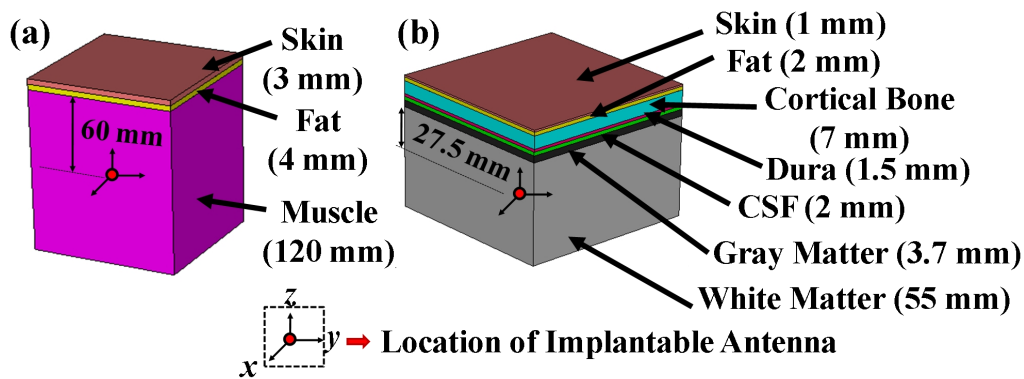


Fig. 4.1: Human body models considered for analysis-(a) three-layer heart model and (c) seven-layer brain model

Using the proposed TMM technique, the effective complex permittivity of the two well-known and often used human body models, depicted in Fig. 4.1(a)-(b) [7-8], is computed for the 402 MHz and 2.45 GHz frequency bands. To determine the correctness of the proposed approach, responses from implanted antennas in multilayer models and the matching expected homogeneous model are recorded.

Here, we further examine the effects of $\pm 20\%$ differences in each layer's thickness, relative permittivity, and conductivity on the prediction of mean effective complex permittivity. In order to validate the simulated reactions, phantom models are practically created by combining deionized water, Diethylene Glycol Monoethyl Ether (DGME), and NaCl in various ratios.

4.3 BACKGROUND OF COMPLEX PERMITTIVITY MEASUREMENT

The waveguide and resonant-cavity methods are frequently used to characterize the complicated permittivity of various materials [9–10]. Cavity perturbation from changes in quality factor and resonant frequency can yield the complex permittivity in the resonant cavity approach [11]. For low permittivity and low loss materials, this approach offers great accuracy [12]. Since human tissues typically have a high dielectric constant and are highly lossy, the resonant cavity approach is not appropriate in this situation. Waveguide techniques come in two varieties: resonant [13] and non-resonant [14]. Here, precise machining is necessary for high accuracy. For these two situations, the material under test (MUT) ought to be stored in a metallic chamber. When the sample length is a multiple of half a wavelength, these techniques exhibit poor accuracy. Since direct contact with MUT (maintained within the sample holder) and active components of the setup is not necessary, the free-space measurement method [15–16] is more effective than earlier approaches. However, in order to achieve high accuracy, the sample must be large and flat, and the sample holder is

expensive [17–18]. Another drawback of this approach is the diffraction effect that has been seen here at the MUT edge. The open ended coaxial probe technique is a commonly used method for measuring dielectric constants [19–20]. It involves placing an open ended coaxial probe inside or on top of the material and using a Vector Network Analyzer (VNA) to record the reflection coefficient in order to measure permittivity and loss tangent. When there is an air gap between the coaxial probe and the MUT, this approach cannot produce the desired results. Other mathematical techniques include the Nicolson-Ross-Weir (NRW) approach [21], the NIST iterative method [22], the Short Circuit Line (SCL) method [23], and others. With these techniques, the MUT is contained within a waveguide, and the scattering parameters (S_{11} , S_{12} , S_{21} , and S_{22}) are used to determine the MUT's dielectric constant [24]. Although the NRW approach is quick and non-iterative, it is not suitable for materials that are big and have low loss [25]. Although the NIST iterative technique is more accurate than the NRW method, it is only useful for measuring real relative permittivity [26].

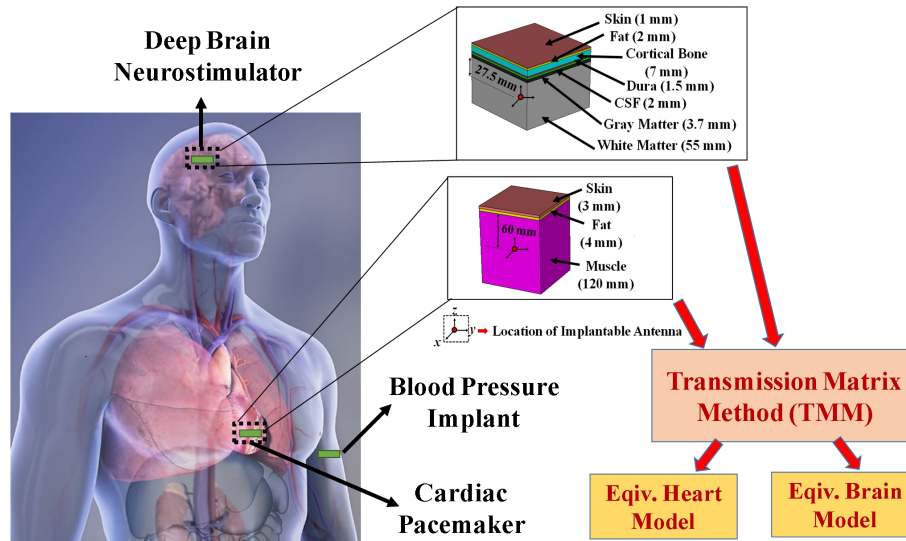


Fig. 4.2: Overview of the Work

As a result, these methods are ineffective for measuring the complicated permittivity of human body models. In the context of electromagnetic wave interaction, the effective complex permittivity of a multilayered model can be computed using either the propagation constant or the scattering parameters, which are connected to the power density [27]. However, for the purpose of characterizing a communication channel, both the scattering parameters and the propagation constant are crucial for any implantable antenna. Both the reflection coefficient and the propagation constant can be related in a transmission matrix for a multilayered model. As a result, TMM is an excellent alternative for achieving this goal. Fig. 4.2 provides an overview of this effort.

4.4 GENERAL CONCEPT OF TRANSMISSION MATRIX METHOD

Transmission Matrix Method (TMM) is applicable for both Transverse-Electric (TE) and Transverse-Magnetic (TM) incident waves illuminating a multilayer model (refer to Fig. 4.3) [28]. A transmission matrix consists of two matrices— amplitude transmission matrix $[A]$ and discontinuity transfer matrix $[D]$. The amplitude transmission matrix is related to the amount of attenuation (α) and phase shift (β) occurring in a particular layer, when any EM wave propagates through it. The discontinuity transfer matrix at an interface between two consecutive layers is related to the reflection (I) and transmission (T) coefficients. The amplitude transmission matrix and discontinuity transfer matrix for l^{th} and $(l+1)^{\text{th}}$ layers of a multilayer dielectric medium (shown in Fig. 4.3) are expressed respectively as shown in equations (1) and (2),

$$[A]_{(l+1)} = \begin{bmatrix} e^{-\gamma_{l+1}(Z_{l+1}-Z_l)} & 0 \\ 0 & e^{\gamma_{l+1}(Z_{l+1}-Z_l)} \end{bmatrix} \quad (1)$$

and

$$[D]_{(l+1)l} = \frac{1}{1 - \Gamma_{(l+1)l}} \begin{bmatrix} 1 & -\Gamma_{(l+1)l} \\ -\Gamma_{(l+1)l} & 1 \end{bmatrix} \quad (2)$$

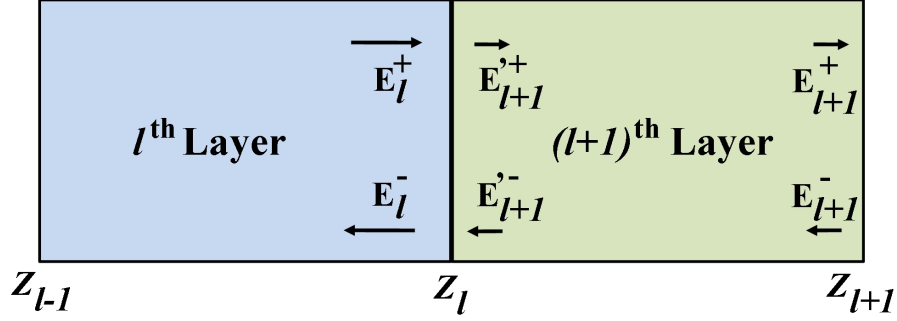


Fig. 4.3: Propagation of electric field from l^{th} layer to $(l+1)^{\text{th}}$ layer

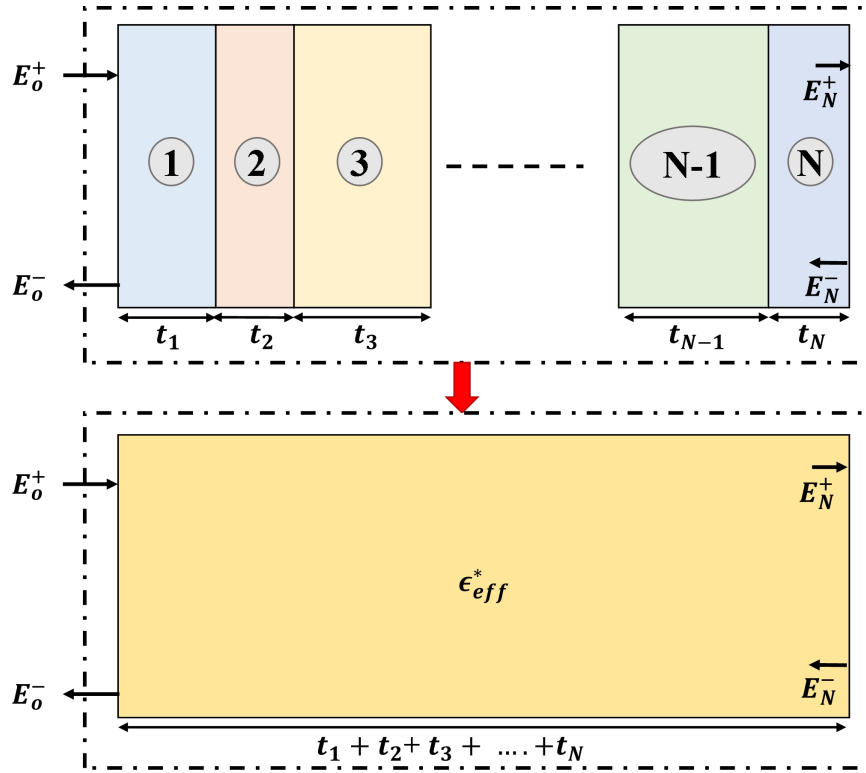


Fig. 4.4: Equivalent model of an N-layer generalized multilayer

Here γ_{l+1} is the propagation constant of layer $(l+1)$, $\Gamma_{(l+1)l}$ is reflection coefficient at the boundary between l^{th} and $(l+1)^{\text{th}}$ layers, $(Z_l - Z_{l-1})$ and $(Z_{l+1} -$

Z_l) are respective thicknesses of l^{th} and $(l+1)^{\text{th}}$ layers. From Fig. 4.3, these two matrices are expressed as,

$$\begin{bmatrix} E_{l+1}^+ \\ E_{l+1}^- \end{bmatrix} = [A]_{l+1} \begin{bmatrix} E_{l+1}'^+ \\ E_{l+1}'^- \end{bmatrix} \text{ and } \begin{bmatrix} E_{l+1}'^+ \\ E_{l+1}'^- \end{bmatrix} = [D]_{(l+1)l} \begin{bmatrix} E_l^+ \\ E_l^- \end{bmatrix}$$

Therefore,

$$\begin{bmatrix} E_{(l+1)}^+ \\ E_{(l+1)}^- \end{bmatrix} = [T]_{(l+1)l} \begin{bmatrix} E_l^+ \\ E_l^- \end{bmatrix} = [A]_{(l+1)} [D]_{(l+1)l} \begin{bmatrix} E_l^+ \\ E_l^- \end{bmatrix} \quad (3)$$

Here $[T]_{(l+1)l} = [A]_{(l+1)} [D]_{(l+1)l}$ is the transmission matrix between l^{th} and $(l+1)^{\text{th}}$ layers.

4.5 PROPOSED COMPLEX PERMITTIVITY MEASUREMENT METHOD OF N-LAYER MODEL USING TMM

A N-layer model shown in Fig. 4.4 is considered here to estimate effective complex permittivity (ϵ_{eff}^*) for the whole model using TMM. Let us consider, the thicknesses of layers 1, 2, 3 N as $t_1, t_2, t_3, \dots, t_N$ respectively. An EM wave with peak value ' E_o^+ ' is impinged into N-layer model. Internal reflections happen in each junction between consecutive two layers. Let, internal reflection coefficient between i^{th} and $(i+1)^{\text{th}}$ layers be $\Gamma_{(i+1)i}$. Propagating through the dielectric layers, the wave is attenuated and phase shifted due to presence of dielectric loss factor. Let, the peak value of EM wave at impinged position and terminated section of i^{th} layer be ' $E_i'^+$ ' and ' E_i^+ ' respectively. Then equation of EM wave in i^{th} layer is

$$E_i^+ = E_i'^+ e^{-\alpha_i t_i} \cos(\omega t - \beta t_i)$$

It is important to fix ϵ_{eff}^* based on not only relative permittivities and conductivities of all layers but also internal reflections and attenuations due to all layers taken into account. Therefore, Transmission Matrix Method (TMM) is

considered here for estimating complex permittivity of human body model as mentioned in Section 4.4. The proposed method for complex permittivity estimation of an N-layer generalized model is illustrated in Fig. 4.5 as flowchart and mentioned below.

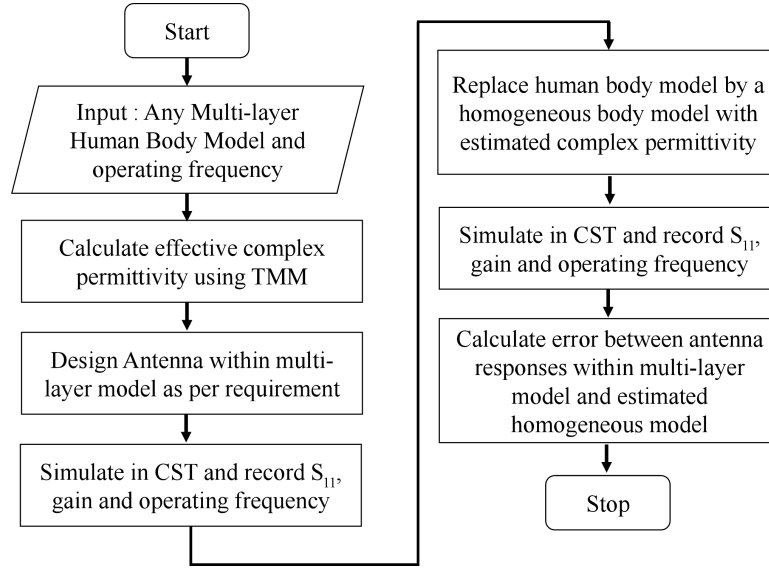


Fig. 4.5: Flow diagram of complex permittivity estimation and validation

Step 1: No of interfaces = No of discontinuity matrices = N; as there is a discontinuity between Layer-0 and Layer-1 too. No of amplitude matrices = No. of layers = N.

Step 2: Calculate propagation constant (γ_i) for each layer using formulas 4-6, [11].

$$\alpha_i = \omega \sqrt{\frac{\mu_i \epsilon_i}{2} \left[\sqrt{1 + \left(\frac{\sigma_i}{\omega \epsilon_i} \right)^2} - 1 \right]} \quad (4)$$

$$\beta_i = \omega \sqrt{\frac{\mu_i \epsilon_i}{2} \left[\sqrt{1 + \left(\frac{\sigma_i}{\omega \epsilon_i} \right)^2} + 1 \right]} \quad (5)$$

$$\gamma_i = \alpha_i + j\beta_i \quad (6)$$

where, μ_i , ϵ_i and σ_i are permeability, permittivity and conductivity of i^{th} layer respectively [11].

Step 3: Calculate $[A]_i$ matrix for each layer.

Step 4: Calculate the reflection coefficients $\Gamma_{(i+1)i}$ between two consecutive layers using below formula [12]

$$\Gamma_{(i+1)i} = \frac{\eta_{(i+1)} - \eta_i}{\eta_{(i+1)} + \eta_i} \quad (7)$$

where, $\eta_{(i+1)}$ and η_i are intrinsic impedances of i^{th} and $(i+1)^{\text{th}}$ layers. The intrinsic impedance of i^{th} layer is defined as [12],

$$\eta_i = \frac{j\omega\mu_i}{\gamma_i} \quad (8)$$

Step 5: Calculate $[D]_{(i+1)i}$ matrices for each interface.

Step 6: Calculate Transmission Matrix for N-layer dielectric model as

$$\begin{aligned} [T]_{1..N} &= [D]_{10} \cdot [A]_1 \cdot [D]_{21} \cdot [A]_2 \dots [D]_{(N-1)N} \cdot [A]_N \\ &= \begin{bmatrix} t_{11} & t_{12} \\ t_{21} & t_{22} \end{bmatrix} \end{aligned} \quad (9)$$

Step 7: Let effective complex permittivity of equivalent model (Fig. 4.3) be $\epsilon_{eff}^* = f(\epsilon_{eff}, \sigma_{eff})$ [13], where ϵ_{eff} is effective relative permittivity and σ_{eff} is effective conductivity. Calculate effective propagation constant γ_{eff} using equations (4-6) by

$$(\epsilon_i, \sigma_i, \mu_i) \rightarrow (\epsilon_{eff}, \sigma_{eff}, \mu_{eff}).$$

$$\text{Step 8: Calculate } [A]_{eff} = \begin{bmatrix} e^{-\gamma_{eff}(Z_{eff})} & 0 \\ 0 & e^{\gamma_{eff}(Z_{eff})} \end{bmatrix}$$

where $Z_{eff} = \sum_{i=1}^N t_i$.

Step 9: Calculate reflection coefficient between Layer-0 and effective homogeneous model ($\Gamma_{(eff)0}$) using equations (8-9) by considering $(\gamma_i, \mu_i) \rightarrow (\gamma_{eff}, \mu_{eff})$.

Step 10: Calculate

$$[D]_{eff,0} = \frac{1}{1-\Gamma_{(eff)0}} \begin{bmatrix} 1 & -\Gamma_{(eff)0} \\ -\Gamma_{(eff)0} & 1 \end{bmatrix}.$$

$$[T]_{eff} = [D]_{eff,0} \cdot [A]_{eff} \quad (10)$$

Step 12: Solve below equations obtained by substituting $[T]_{1..N} = [T]_{eff}$.

$$t_{11} = \frac{e^{-\gamma_{eff}(Z_{eff})}}{1 - \Gamma_{(eff)0}} \quad (11)$$

$$t_{22} = \frac{e^{\gamma_{eff}(Z_{eff})}}{1 - \Gamma_{(eff)0}} \quad (12)$$

Obtain α_{eff} and β_{eff} by solving above two equations.

Step 13: Calculate effective permittivity and conductivity from below equations –

$$\epsilon_{eff} = \frac{(\alpha^2 - \beta^2)}{\omega^2 \mu_{eff}} \quad (13)$$

$$\sigma_{eff} = \omega \epsilon_{eff} \sqrt{\left(\left(\frac{(\alpha^2 + \beta^2)}{\omega^2 \mu_{eff} \epsilon_{eff}} \right)^2 - 1 \right)} \quad (14)$$

4.6. COMPLEX PERMITTIVITY ESTIMATION AND VALIDATION OF DIFFERENT BODY MODELS AT DIFFERENT CONDITIONS

The proposed method of complex permittivity estimation using TMM is mentioned in Section 4.4. In this section, two popular multilayered human body models (refer to Fig. 4.1) are considered to estimate their effective complex permittivities using proposed technique at two widely used medical frequencies – 402 MHz and 2.45 GHz [1-8]. Dielectric properties of each layer as shown in Fig. 4.1 are mentioned in Table 4.1 [30-31].

Table 4.1: Dielectric Properties of Different Body Layers

Model No.	Tissue Name	402 MHz		2.45 GHz	
		ϵ_r	σ	ϵ_r	σ
Model 1 [16,18]	Skin	46.7	0.69	37.88	1.44
	Fat	5.58	0.04	5.28	0.1
	Muscle	57.1	0.8	54.81	2.21
Model 2 [17,18]	Skin	46.7	0.69	37.88	1.44
	Fat	5.58	0.04	5.28	0.1
	Cortical Bone	13.1	0.0915	11.4	0.394
	Dura	46.6	0.828	42	1.67
	CSF	71	2.25	66.2	3.46
	Gray Matter	57.4	0.738	48.9	1.81
	White Matter	42	0.637	36.2	1.21

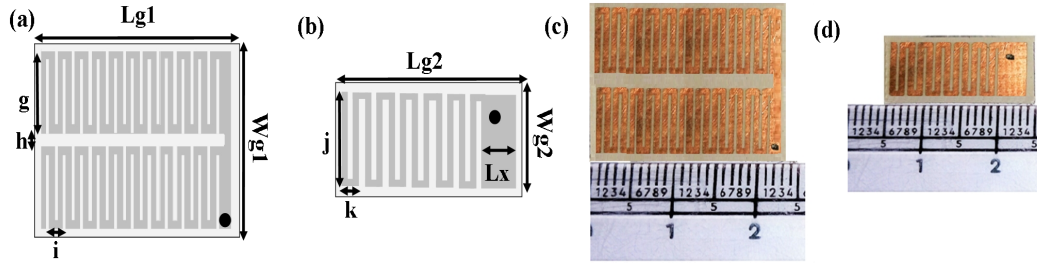


Fig. 4.6: Simulated implantable antennas at (a) 402 MHz band and (b) 2.45 GHz band where $Lg1 = 23$ mm, $Wg1 = 16.4$ mm, $g = 7.2$ mm, $h = 1$ mm, $i = 1.5$ mm, $Lg2 = 20$ mm, $Wg2 = 10$ mm, $j = 6.8$ mm, $k = 0.6$ mm, $Lx = 3.6$ mm; Fabricated prototype of the antennas tuned at (c) 402 MHz band and (d) 2.45 GHz band

4.6.1. Three-layer Heart Model

A three-layer heart model shown in Fig. 4.1 (a) is considered as Model 1 which was used by Sharma et al. [7] for in-vitro analysis of a cardiac pacemaker system.

1) At 402 MHz MICS band:

A double meander-line implantable antenna shown in Fig. 4.6(a) is designed to operate within the three-layer heart model. The antenna design procedure is widely discussed in Ghosh et. al [36]. Here, the design steps are briefly discussed. At first, a simple $20\text{mm} \times 20\text{mm}$ rectangular patch is taken. For ensuring biocompatibility, polyvinyl alcohol (PVA) coating is used to encapsulate the antenna. The antenna within human body model is tuned at 3.18 GHz. Next, a single meander-line cut is added on the patch to increase the current path length. Due to this, the resonant frequency is tuned to 2.81 GHz. For further decrement, six meanders are incorporated (refer to Fig. 4.6(b)). This antenna is tuned to 2.45 GHz with return loss value of 25.35 dB. This design is fixed for verifying the responses at 2.45 GHz in the next subsection. For further decrement, twelve meanders are placed on patch. The modified antenna architecture has return loss

value of 15.69 dB at 901 MHz. In the following step, similar 12-meander section is added with previously designed 12-meander section using a rectangular strip.

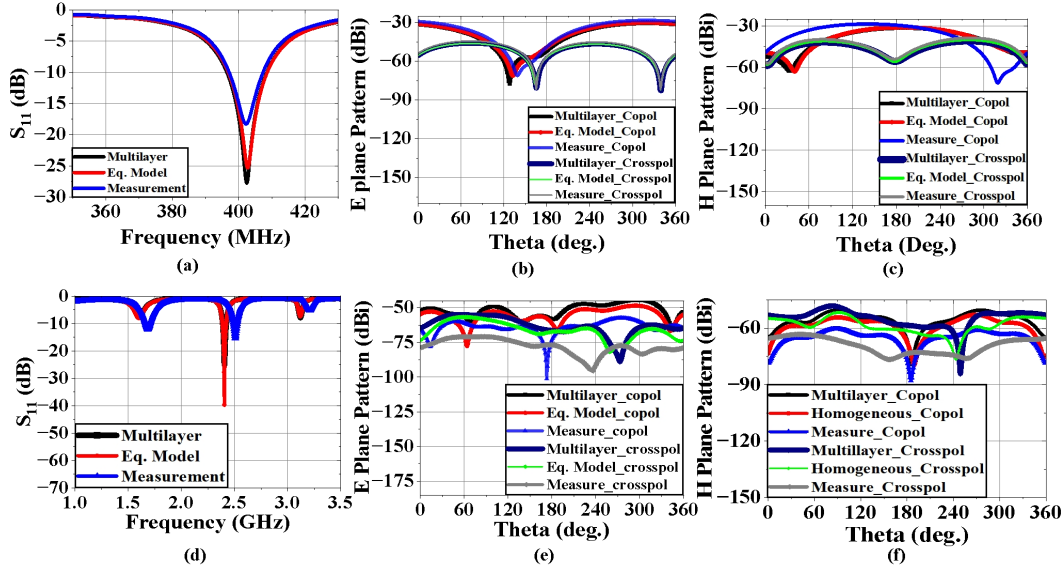


Fig. 4.7: Comparison of return loss and far field (E and H plane) radiation patterns of implanted antenna within multilayer, estimated equivalent model and practical phantom models at (a-c) 402 MHz and (d-f) 2.45 GHz for Model 1 (Heart Model)

After this modification, the return loss value of this antenna is 26.54 dB at 401.9 MHz. Therefore, this design is finalized to verify the operation at 402 MHz. The realized gain of this antenna is -31.7 dBi at the operating frequency (shown in Fig. 4.7(a-c)). Next, effective complex permittivity of the heart model is calculated using proposed method. The respective effective relative permittivity and conductivity of the homogeneous equivalent heart model are obtained as 57.1 and 0.75 S/m. Complex permittivity of this model is $(57.1-j33.75) \epsilon_0$. The three-layer heart model shown in Fig. 4.1(a), is now replaced by homogeneous heart model with estimated effective complex permittivity. The return loss and realized gain of the antenna within this homogeneous model are 25.34 dB and -31.2 dBi respectively at 401.85 MHz as shown in Fig. 4.7(a-c).

2) At 2.45 GHz ISM band:

A miniaturized meander-line implantable antenna as shown in Fig. 4.6(b) is designed to operate within the three-layer heart model. The return loss and realized gain of this antenna are 25.35 dB and -44.1 dBi respectively at operating frequency 2.45 as illustrated in Fig. 4.7 (e-g). The calculated effective relative permittivity and conductivity of the homogeneous equivalent heart model are obtained at 2.45 GHz as 54 and 2.2 S/m. Complex permittivity of this model is $(54-j16.16) \epsilon_0$. In simulation, the return loss and realized gain of the antenna within estimated homogeneous heart model are 25.21 dB and -44.5 dBi respectively (refer to Fig. 4.7(e-g)).

3) Effect of Variations in body parameters on ϵ_{eff}^* :

Kiourti et. al [30] mentioned that $\pm 20\%$ variations in human body electrical parameters can be possible with respect to mean values due to several conditions such as age, gender, location of patient. In [35], the reason behind the consideration of this range of variability is widely discussed. Therefore, effects of possible variations in skin and fat thicknesses, implantation depth in muscle layer, relative permittivities in skin, fat and muscle layers and conductivities in skin, fat and muscle layers on effective complex permittivity variations are studied below:

a) *Effect of $\pm 20\%$ thickness variations in skin and fat:*

The respective ranges of thicknesses of skin and fat layers are considered to be (2.2 mm-3.8 mm) and (3.2 mm-4.8 mm). Using TMM, the calculated complex permittivity of homogeneous heart model is not changed due to these variations for both 402 MHz and 2.45 GHz. In simulation, it is also observed

that variations in skin and fat thicknesses cannot affect significantly the antenna performances in terms of operating frequency, return loss and realized gain.

b) *Effect of $\pm 20\%$ implantation depth variation:*

The effective complex permittivity of heart model is also calculated for variable implantation depth ranging between 42-78 mm ($\pm 20\%$ of mean depth 60mm). After calculation, it is observed that relative permittivity is fixed at 57.1 and 54 for 402 MHz and 2.45 GHz respectively. For both the cases, the effective conductivity linearly varies up to 60 mm depth with a slope of 0.02×10^{-3} S/mm². Then it gets fixed at its mean value.

c) *Effect of $\pm 20\%$ relative permittivity variations:*

Dependence of effective complex permittivity on $\pm 20\%$ variations in relative permittivity of each layer at 402 MHz and 2.45 GHz is analyzed here. The effective complex permittivity value is not changed due to variation in skin and fat layers. The reason is that the muscle layer dominates over skin and fat layers due to high dielectric constant and large implantation depth. The effective relative permittivity is unchanged due to these variations. The effective conductivity is varied only over $\pm 2.27\%$ of mean conductivity value which is insignificant from the viewpoint of practical measurement.

d) *Effect of $\pm 20\%$ conductivity variations:*

Considering $\pm 20\%$ variations in conductivities of skin and fat layers, effective complex permittivity of heart model remains same at its mean condition. Almost linear relationship between effective complex permittivity and conductivity of muscle layer is observed for both 402 MHz and 2.45 GHz with

$\frac{d\epsilon_{eff}}{d\sigma_m} = 0.78$ slope. Around $\pm 20.15\%$ variation is expected in effective complex permittivity due to possible variation in muscle conductivity irrespective of frequencies.

4) Fabrication and Measurement:

The mean heart models for 402 MHz and 2.45 GHz are practically fabricated for practical verification of complex permittivity estimation. The antennas shown in Fig. 4.6(a-b) are fabricated using photolithography and their prototypes are depicted in Fig. 4.6(c-d). In this work, different human phantom models are developed by mixing deionized (DI) water, Diethylene Glycol Monoethyl Ether (DGME) and sodium chloride (NaCl) in different proportions. The DI water is prepared in the Laboratory of Jadavpur University, India. DGME and NaCl are purchased from Sigma Aldrich, India. To prepare heart phantom for 402 MHz, 51.05% (v/v) DGME is mixed with 108 ml DI water at 45 °C, 800 rpm for 30 minutes with Remi 2MLH magnetic stirrer. The dielectric measurement of this mixture is performed using 85070E Agilent dielectric kit which is calibrated using 100 ml DI.

The developed solution has respective relative permittivity and conductivity of 56.87 and 0.82 S/m. The required respective relative permittivity and conductivity to mimic heart phantom at 402 MHz are 57.1 and 0.75 S/m. The tolerance values for these values are considered as $\pm 10\%$ which is sufficient for mimicking. The error margins for relative permittivity and conductivity are 0.4% and 8.5% respectively which are within the tolerance band. The dielectric measurement setup of phantom and in vitro testing of antenna within fabricated phantom are shown in Fig. 4.8. Next, to mimic heart model at 2.45 GHz, 47.6% (v/v) DGME is mixed with 105ml DI water in similar way. From dielectric measurement, the respective relative permittivity and conductivity are obtained as

54.82 and 1.78 S/m. The required respective relative permittivity and conductivity to mimic heart phantom at 2.45 GHz are 54 and 2.2 S/m. The error margins of relative permittivity and conductivity are obtained as 1.49% and 19.09% respectively. To increase conductivity of the mixture, 0.2g NaCl is thoroughly with the mixture. It leads to the increase of conductivity to 2.31 S/m with the error margin of 5%.

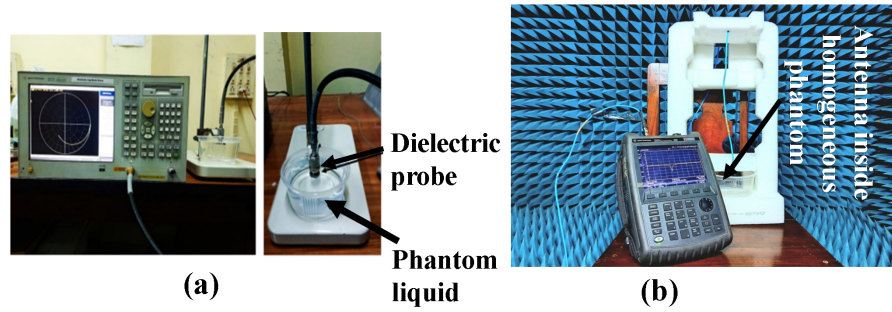


Fig. 4.8: (a) Dielectric measurement of the phantom and (b) antenna testing within phantom

Fabricated antenna shown in Fig. 4.6(c) is encapsulated by biocompatible PVA and placed within heart phantom has return loss and realized gain at measured operating frequency 401.2 MHz are 17.35 dB and -32.79 dBi respectively. The equivalent heart model at 2.5 GHz is fabricated by mixing 109 ml DI water, 51 ml DGME and 0.32 g NaCl so that the mixture has relative permittivity and conductivity of 53.89 and 2.07 S/m respectively. The fabricated antenna shown in Fig. 4.6(c) is placed at the phantom and its responses are recorded. The antenna within designed phantom has return loss value of 15.19 dB with realized gain -42.87 dBi at 2.51 GHz.

4.6.2. Seven-layer Brain Model

Next, the seven-layer brain model shown in Fig. 4.1 (b) is considered. The antennas illustrated in Fig. 4.6 (a) and (b) are placed within white matter of the brain model considering their suitable conditions in terms of frequency range.

1) At 402 MHz MICS band:

The antenna shown in Fig. 4.6 (a) is implanted in the brain model and simulated in CST. The return loss value of this antenna is 25.27 dB at 402 MHz and realized gain of -28.67 dBi at the operating frequency (shown in Fig. 4.9(a-c)). Next, effective complex permittivity of the seven layer brain model is calculated using the proposed method. The respective effective relative permittivity and conductivity of the homogeneous equivalent heart model are obtained as 57.4 and 1.81 S/m.

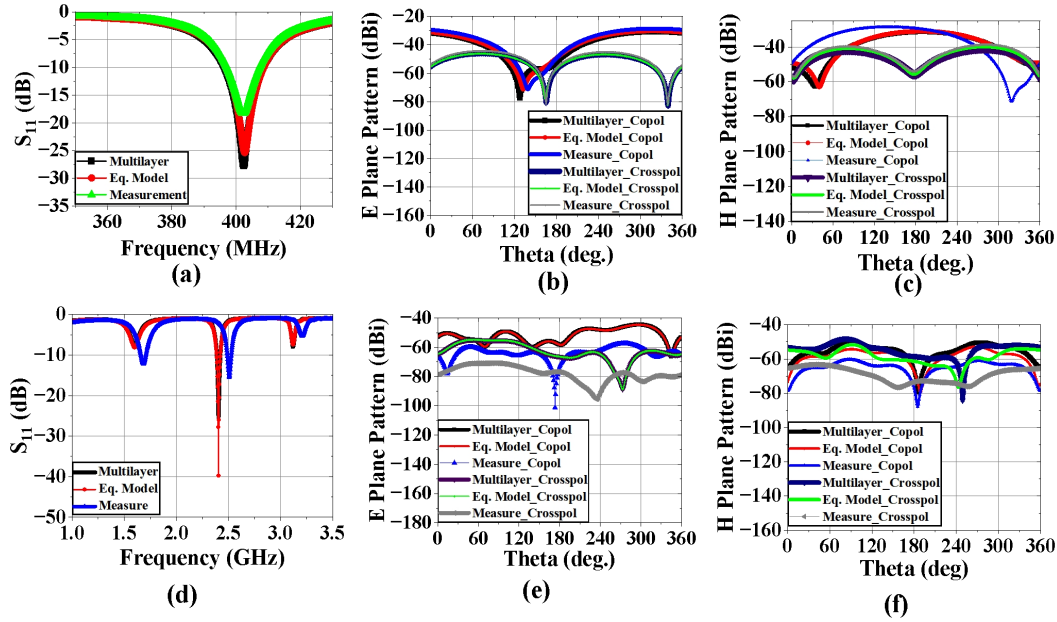


Fig. 4.9: Comparison of return loss and far field (E and H plane) radiation patterns of implanted antenna within multilayer, estimated equivalent model and practical phantom models at (a-c) 402 MHz and (d-f) 2.45 GHz for Model 2 (Brain model)

Complex permittivity of this model is $(57.4-j81.45) \epsilon_0$. The seven-layer brain model shown in Fig. 4.1(b), is now replaced by homogeneous brain model with estimated effective complex permittivity. The return loss and realized gain of

the antenna within this homogeneous brain model are 24.34 dB and -29.34 dBi respectively at 402.75 MHz as shown in Fig. 4.9(a-c). To develop brain phantom at 402 MHz, 48.15% (v/v) DGME is mixed with 108 ml DI water at 45 °C, 800 rpm for 30 minutes. The mixture has respective relative permittivity and conductivity of 57.87 and 1.82 S/m with respective error margins of 0.81% and 0.54% which represents effectively mimic of the homogeneous brain model at 402 MHz. The measured return loss and realized gain within this model are 18.65 dB and 34.55 dBi at 400.78 MHz.

2) At 2.45 GHz ISM band:

Then, the antenna shown in Fig. 4.6 (b) is implanted in brain model and it is simulated in CST. The simulated return loss is 38.78 dB at 2.49 GHz with -33.6 dBi realized gain. The proposed method provides the effective relative permittivity and conductivity values of the brain model are 48.9 and 1.78 S/m. Complex permittivity of this model is $(48.9-j13.07) \epsilon_0$. Thereafter, this brain model is replaced by a single-layer equivalent model with predicted electrical parameters. The simulated equivalent model shows the return loss value of 37.82 dB at 2.49 GHz with realized gain -33.07 dBi.

The equivalent model is fabricated for practical validation by mixing 49% (v/v) DGME with 100 ml DI water and 0.2 g NaCl. The mixture has respective relative permittivity and conductivity of 47.69 and 1.76 S/m with respective error margins of 2.4% and 1.1% which represents effectively mimic of the homogeneous brain model at 2.45 GHz. The fabricated antenna is placed within fabricated phantom and it radiates with respective realized gain and return loss values -36.37 dBi and 16.74 dB at measured operating frequency 2.52 GHz.

Table 4.2: Comparisons between TMM-based Predictions and Voxel model Simulations

Model	Freq. (GHz)	Return Loss (dB)			Realized gain (dBi)		
		Voxel model	Equiv. Model	Error (%)	Voxel model	Equiv. Model	Error (%)
Heart	0.402	25.89	25.34	0.69	-32.45	-31.2	3.57
	2.45	25.43	25.21	0.47	-43.91	-44.5	1.58
Brain	0.402	25.19	24.34	0.92	-29.63	-29.34	1.03
	2.45	36.27	37.82	1.21	-33.21	-33.07	0.83

4.7. COMPARISON OF TMM WITH REALISTIC VOXEL BODY MODEL SIMULATION

This section compares the performance of implantable antennas in Gustav CST voxel heart and brain models with the projected equivalent heart and brain models in order to determine error margin. The realistic voxel heart and brain models used to record the implanted antenna's performance parameters at 402 MHz and 2.45 GHz are displayed in Fig. 4.10. Main comparisons between full multilayer model simulations and forecasts based on TMM are shown in Table 4.2.

4.8. VERIFICATION OF FREQUENCY INDEPENDENCY

The effectiveness of the proposed approach to determine complex permittivity of several human body models is investigated in the preceding

sections at two widely used medical frequencies: 402 MHz and 2.45 GHz. The relative permittivities and conductivities of the heart and brain models at various frequencies are calculated using the proposed approach and responses of implantable antenna performances within multilayer and corresponding predicted homogeneous equivalent model are reported in Table 4.3 in order to confirm the robustness of this method at lower (<100 MHz) or higher (>3 GHz) frequencies. The table shows that the proposed approach may calculate effective complex permittivity for various frequencies with low error margins (average error margin is $\sim 2.5\%$).

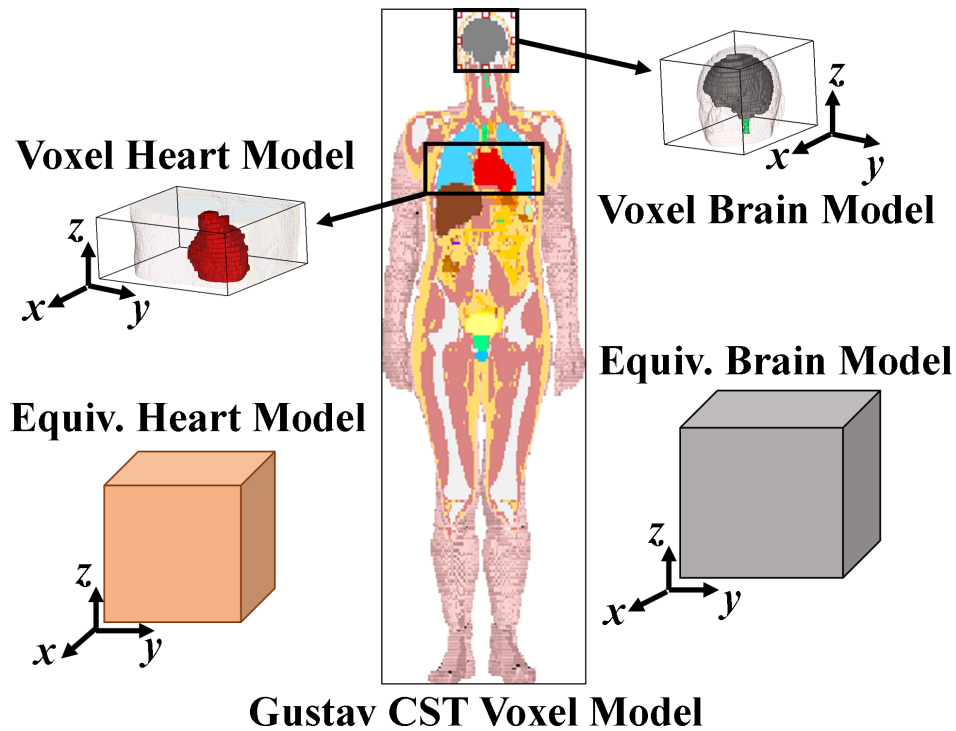


Fig. 4.10: Comparison of Implantable antenna performances within Gustav CST heart and brain models with performances within predicted equivalent heart and brain models

Table 4.3: Frequency Dependency Analysis of the Proposed Method

Model No.	Freq.	Return Loss (dB)		Realized gain (dBi)	
		Multi Layer	Equiv. Model	Multi Layer	Equiv. Model
1	80 MHz	32.45	35.6	-42.98	-44.27
	100 MHz	30.47	30.98	-47.31	-46.59
	3 GHz	18.79	17.42	-31.67	-33.45
	5.8 GHz	16.22	15.87	-28.33	-29.68
2	80 MHz	31.57	32.21	-41.73	-41.27
	100 MHz	27.15	26.42	-45.06	-45.50
	3 GHz	20.36	19.83	-36.98	-37.08
	5.8 GHz	20.47	20.32	-32.56	-33.93

4.9. ACCURACY ANALYSIS

Human body models are multilayered in nature. To overcome practical difficulties during multilayer human body phantoms as mentioned in Section 4.1, accurate homogeneous body models should be fabricated for in vitro measurement of implantable devices. Therefore, accuracy analysis of effective complex permittivity prediction using TMM method is important which is performed in this section. The accuracy of predicted effective complex permittivity of the human models at 402 MHz MICS band and 2.45 GHz ISM band are analyzed in Table 4.4 in terms of return loss and realized gain of implantable antenna. In this work, the correlation coefficient technique is used to

estimate the accuracy of prediction. The correlation coefficient ($\Re$) can be defined as,

$$\Re = \frac{cov(g, h)}{\sigma_g \sigma_h}$$

Where,

$$cov(g, h) = \frac{\sum gh}{n} - \frac{\sum g}{n} \frac{\sum h}{n}$$

$$\sigma_g = \sqrt{\frac{\sum g^2}{n} - \left(\frac{\sum g}{n}\right)^2}$$

$$\sigma_h = \sqrt{\frac{\sum h^2}{n} - \left(\frac{\sum h}{n}\right)^2}$$

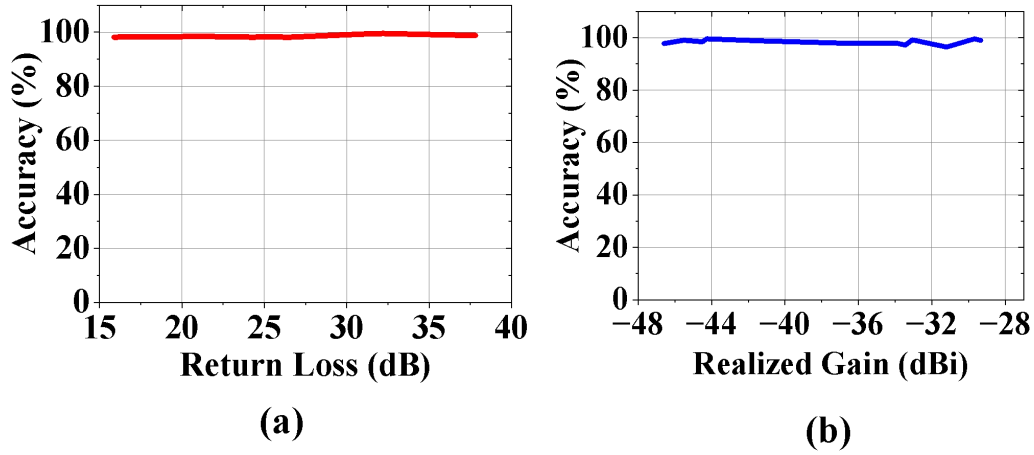


Fig. 4.11: Accuracy plots with respect to return loss and realized gain values

Two datasets are prepared for accuracy analyses using correlation coefficient method. In first dataset, 25 sets of the values of error in % (h) vs return

loss in dB (g) are prepared for these two body models at different operating frequencies ranging between 80 MHz and 5.8 GHz. The accuracy related to this dataset is obtained as 98.57%. In similar way, accuracy related to realized gain (dBi) is calculated as 98.38%. The accuracy plots of return loss and realized gain are illustrated in Fig. 4.11. The average accuracy of the proposed technique is 98.46% with respect to multilayer body model.

Table 4.4: Accuracy Analysis

Model No.	Freq. Band	Return Loss (dB)			Realized gain (dBi)		
		Multi Layer	Equiv. Model	Error (%)	Multi Layer	Equiv. Model	Error (%)
1	MICS	26.54	25.34	1.29	-31.7	-31.2	1.22
	ISM	25.35	25.21	1.57	-44.1	-44.5	0.87
2	MICS	25.27	24.34	1.22	-28.67	-29.34	2.59
	ISM	38.78	37.82	2.07	-33.6	-33.07	1.53

4.10. COMPARISON WITH PREVIOUSLY REPORTED WORKS

There is very less number of works in which complex permittivity of human body model has been calculated. In some literature, accuracy has not been reported. However, this work is compared with two reported works [32-34]. The comparative analysis of the proposed technique with these works is shown in Table 4.5. The authors in [34] proposed a technique based on modified open ended coaxial probe to measure the complex permittivity of skin models efficiently with around 95% accuracy based on only reflectometry technique. From this table, we can conclude that the proposed technique can estimate

complex permittivity of different human models required for different applications more accurately than other reported works for the in vitro testing of any implantable antenna in different medical frequency bands.

Table 4.5: Comparison of the Accuracy of Estimation of the Proposed Technique with Previously Reported Works

Reference	Technique followed	Accuracy (%)
[32]	Open ended coaxial probe	<98.2
[33]	Artificial Neural Network	95
[34]	Modified Open ended Coaxial Probe	~95
This Work	Transmission Matrix Method	98.46

4.11. CONCLUSION

In this work, computational electromagnetics-based accurate effective complex permittivity estimation of a human body models for in vitro measurement of implantable antenna is proposed using Transmission Matrix Method (TMM). A TMM based complex permittivity estimation method for human body model is proposed to overcome the problems of the existing techniques. To analyze the accuracy of this method, two meander-line based implantable antennas operating at 402 MHz MICS and 2.45 GHz ISM bands respectively are designed. These antennas are placed firstly within multilayered popular heart and brain models and then placed within their equivalent models

predicted by the proposed method. The implantable systems are simulated in CST. Finally the predicted homogeneous models are developed by mixing DI water, DGME and NaCl with different ratios to replicate the simulated models practically. The results show that the proposed complex permittivity estimation method can achieve 98.46 % average accuracy. This method is independent of the number of tissue layers, operating frequency, structure of implantable antenna etc.

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Chapter 5

Implantable Antenna System within Equivalent Human Body Model

Contents	5.1. Introduction 5.2. Problem Formulation 5.3. Implantable Antenna System Design 5.4. Estimation of Effective Complex Permittivity of Three-Layer Body Model 5.5. Performance of Implantable Antenna within Equivalent Body Phantom 5.6. Dependence Analyses and Discussion 5.7. Experimental Validation 5.8. Novelty Analysis 5.9. Conclusion
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5.1 INTRODUCTION

In Chapter 3, a corner-chamfered meander-line implantable antenna has been designed within a three-layer human body model. In Chapter 4, the practical problems related to the fabrication of multilayer human body has been discussed. Therefore, the designed implantable antenna is implanted within equivalent homogeneous human body phantom. Before surgical implantation of implantable devices within patient's body, proper in-vitro testing is required. The testing should be performed within a medium which can provide similar effect as human body. This artificially made medium is called "Phantom" [1-2]. Different biological phantoms are utilized for different purposes. Dosimetry phantoms are used in radiotherapy. Besides them, phantoms can be utilized in photo-acoustic imaging, breast imaging, image reconstruction etc. In medical imaging, biological phantoms take crucial role [3]. Breast phantom with high fidelity is required for

microwave breast tumor and cancer detection system [4]. To test implantable pacemakers, cardiovascular devices, defibrillators, infusion systems, implantable lenses and other implantable telemetry devices, accurate biological phantom development is very important.

The electrical properties of human body are dependent on different factors such as location, gender, age of patients etc. Effective permittivity and conductivity are influenced by the electrical and physical property variations of different tissue layers. Thickness of skin layer is different for different parts of the body. The thinnest skin is found in eyelids whereas the thickest skin is present in palms and soles due to presence of stratum lucidum [5]. The thickness of skin also depends on gender of patient. Male skin is thicker than female due to higher concentrations of collagen and testosterone [6-7]. Vitamin C serums help to enhance collagen to thicken the skin whereas, UV light exposure, tobacco, aging, excessive sugar intake reduce collagen amount [8]. Aging is one of the most important reasons of fat thickness variation because it is associated with redistribution of subcutaneous adipose tissue [9]. Non-healthy diet containing fatty foods can increase fat thickness. Different diseases like Madelung's disease (fatty tumors), Gottron's syndrome etc. appear due to fat layer thickness variation [10-11]. The variations of skin and fat thicknesses can change effective permittivity and conductivity of human body. The water content, oil content, blood content, structure of tissue and fibers etc. can also take crucial role to determine the effective complex permittivity. 0.4-0.6 mm thick Stratum Corneum (a part of epidermis of skin layer) contains comparatively lower water content (around 10%) [12]. According to Gabriel's database [13], relative permittivity and conductivity of wet skin is larger than dry skin due to higher water content in wet skin than dry skin. Dizziness, dry mouth, weakness, less frequent darker urination etc. are some symptoms of dehydration of skin which may vary implanted antenna performances due to variation in effective permittivity of human body. Skin redness, irritation, skin diseases like eczema, psoriasis etc. are related to dry

skin (Xeroderma) [14]. Hwang et al [15], Boric-Lubecke et al [16] showed that the effective complex permittivity of different body parts is not same. Therefore, it is not recommended to study implantable antenna performance within phantom model with fixed relative permittivity and conductivity. The generation of tumor in any layer may increase effective complex permittivity by about 20%. The effective dielectric properties can also change due to pathological change in liver due to cirrhotic, steatosis, hepatocellular carcinoma etc. [17]. Therefore, performance of implanted antenna can vary with several factors – (i) location of implantation, (ii) skin and fat thickness of different patients, (iii) individual physical and electrical properties of body layers, (iv) generation of diseases after implantation, (v) hormonal effect in patients etc. due to effective permittivity and conductivity variation.

From measurement point of view, there is a possibility of changes in effective permittivity and conductivity of the homogeneous phantom during in-vitro testing of implantable antenna. The phantom model can be developed by mixing several chemicals such as deionized water, NaCl, sucrose, DGME (Diethylene Glycol Monoethyl Ether), Agarose etc. During measurement of implantable antenna, evaporation can happen which leads to reduce effective relative permittivity and conductivity for reduction of water content in phantom liquid. Bacterial activities can also vary effective complex permittivity of semi-solid phantom. Therefore, dependence of implantable antenna performance on variations in effective electrical permittivity of phantom is required to investigate which has not been done yet to the best of our knowledge. Therefore in this chapter, dependence of implantable antenna performance on effective relative permittivity and conductivity of equivalent human body phantom model is investigated.

This chapter is decomposed into three parts. The calculation of effective relative permittivity and conductivity of the three-layer human body model is performed in the first part of this chapter. Next, implantable antenna kept within a

implantable medical device is put into the homogeneous model and its performances are recorded. In final part, the variations of antenna system are recorded and analyzed for $\pm 20\%$ possible variations in effective relative permittivity and conductivity.

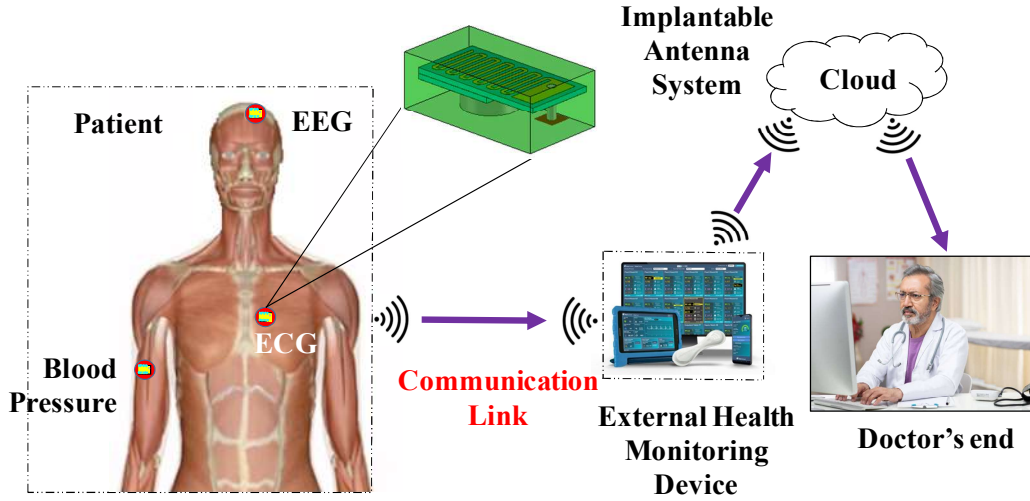


Fig. 5.1: Overview of a wireless communication link between an implantable antenna system and an external health monitoring device

5.2 PROBLEM FORMULATION

For proper sensitivity analysis, it is important to design a prototype of an implantable system by placing the designed corner-chamfered meander-line implantable antenna within a biocompatible coating along with sensor, battery, camera, Printed Circuit Board (PCB) etc. Therefore, an implantable medical device has been designed and placed within homogeneous human body model. Next, the system responses (S_{11} , operating frequency, realized gain, fractional bandwidth, Specific Absorption Rate (SAR) for 1g and 10g tissue model, link margin and communication range) are recorded for possible variations in effective

relative permittivity and conductivity of human body model. The overview of the wireless patient monitoring using implantable antenna system is shown in Fig. 5.1.

5.3 IMPLANTABLE ANTENNA SYSTEM DESIGN

For proper analysis of the implantable antenna performances, a prototype of implantable medical device (shown in Fig. 5.2) containing the corner-chamfered antenna, camera, PCB and sensor encapsulated by 0.5 mm thick biocompatible coating made of 10 g gelatin and 15 ml deionized (DI) water which has relative permittivity of 41.57 and loss tangent of 0.39 at 2.45 GHz, was taken [18]. A coaxial probe with 0.63 mm probe radius providing input impedance of 50Ω was used for exciting the antenna system. Different views of the antenna system are shown in Fig. 5.2.

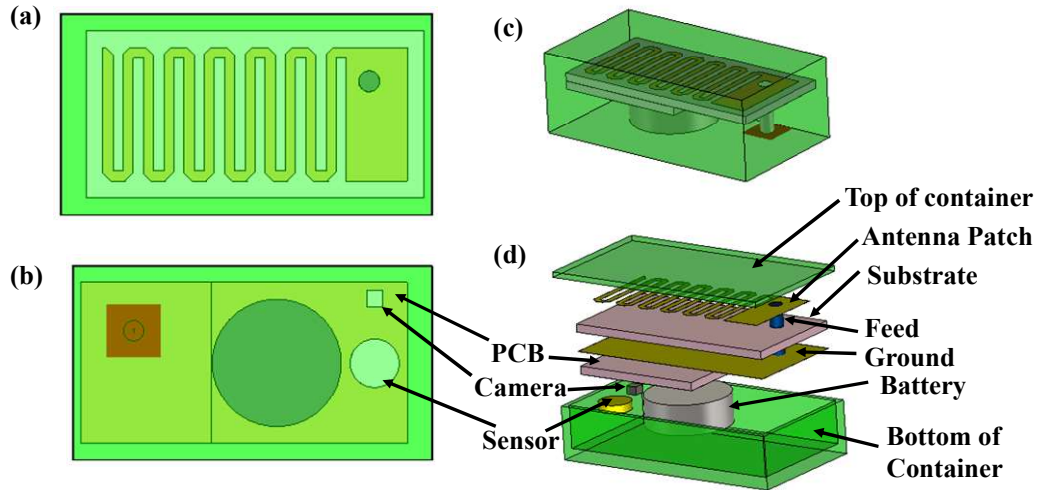


Fig. 5.2: Detailed architecture of Implantable device: (a) Top view, (b) Bottom view, (c) Isometric view and (d) Exploded view

5.4 ESTIMATION OF EFFECTIVE COMPLEX PERMITTIVITY OF THREE-LAYER BODY MODEL

In Chapter 4, the effective complex permittivity estimation using Transmission Matrix Method (TMM) for any N-layer model is discussed. In this section, the effective complex permittivity of the three-layer body model shown in Fig. 3.2 (c) is calculated using similar process.

5.4.1. Properties of Human Tissues and Free Space

In this section, different properties of three human tissues are studied. According to Gabriel's database [19-22], relative permittivities of skin, fat and muscle are 38, 10.8 and 52.7 respectively and conductivities are 1.46, 0.268 and 1.95 S/m respectively at 2.45 GHz. Attenuation constants (α), phase-shift constants (β) and intrinsic impedances (η) are calculated using equations (1), (2) and (3) respectively [23] and tabulated in Table 5.1 by substituting relative permittivity and conductivity of each layer in these equations for linear frequency 2.45 GHz.

$$\alpha = \omega \sqrt{\frac{\mu\epsilon}{2} \left[\sqrt{1 + \left(\frac{\sigma}{\omega\epsilon}\right)^2} - 1 \right]} \quad (1)$$

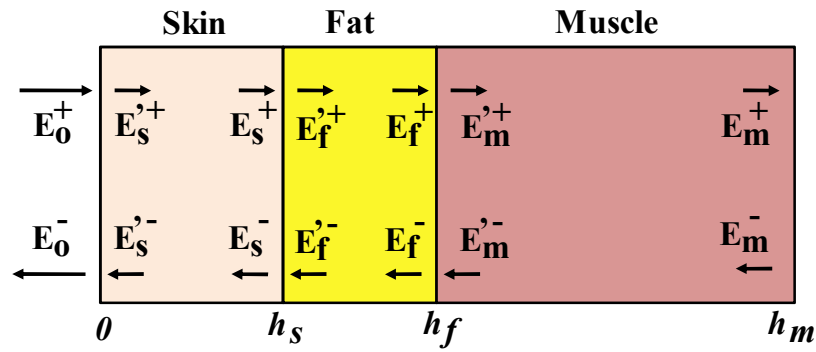
$$\beta = \omega \sqrt{\frac{\mu\epsilon}{2} \left[\sqrt{1 + \left(\frac{\sigma}{\omega\epsilon}\right)^2} + 1 \right]} \quad (2)$$

$$\eta = \frac{j\omega\mu}{\gamma} \quad (3)$$

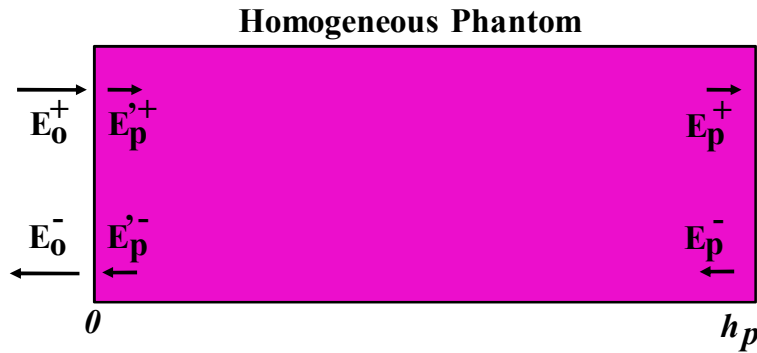
Here, ω is angular frequency, ϵ is permittivity, σ is conductivity and μ is permeability and $\gamma = \alpha + j\beta$ is propagation constant. For free space, relative permittivity, conductivity and permeability are 1, 0 and 1 respectively. Attenuation and phase constant and intrinsic impedance of free space is mentioned in Table 5.1 [23].

Table 5.1: Properties of Skin, Fat, Muscle and Free Space at 2.45 GHz

Medium	Attenuation constant (Np/m)	Phase constant (rad/m)	Intrinsic Impedance (Ω)
Skin	44.73	319.48	$59.41 + j8.32$
Fat	16.63	169.33	$113 + j11.12$
Muscle	45.57	375.28	$50.8 + j6.17$
Free Space	0	8.17	377.14



(a)



(b)

Fig. 5.3: Electric field distribution in (a) three layered body model and (b) homogeneous phantom

5.4.2. TMM for Three-layer Human Body Model

A three-layered body model is shown in Fig. 5.3 (a). The thicknesses of skin (h_s), fat (h_f) and muscle (h_m) are 4 mm, 2 mm and 20 mm respectively (along +z axis). An electric field propagating through free space ($\vec{E}_1$) is assumed to be incident on skin layer with peak amplitude E_0 .

1) Transmission Matrix at the interface between air and skin

Reflection coefficient (Γ) at boundary between air and skin can be calculated as,

$$\Gamma_{s0} = \frac{\eta_s - \eta_0}{\eta_s + \eta_0} = -0.73 + j0.03 \quad (4)$$

Therefore,

$$\frac{1}{1 - \Gamma_{s0}} = 0.58 + j0.01. \quad (5)$$

Propagation constant of skin (γ_s) = $44.73 + j319.48 \text{ m}^{-1}$.

Amplitude transmission matrix for skin layer is,

$$[A]_{skin} = \begin{bmatrix} 0.24 - j0.8 & 0 \\ 0 & 0.35 + j1.15 \end{bmatrix}$$

Discontinuity transfer matrix for the interface between air and skin layer is,

$$[D]_{skin,air} = \begin{bmatrix} 1 & 0.73 - j0.03 \\ 0.73 - j0.03 & 1 \end{bmatrix} \times (0.58 + j0.01)$$

Thereafter, transmission matrix for air and skin can be expressed in equation (6) as,

$$[T]_{skin,air} = [A]_{skin} \times [D]_{skin,air}$$

$$[T]_{skin,air} = \begin{bmatrix} 0.14 - j0.46 & 0.09 - j0.34 \\ 0.16 + j0.48 & 0.19 + j0.67 \end{bmatrix} \quad (6)$$

Therefore, we can write,

$$\begin{bmatrix} E_s^+ \\ E_s^- \end{bmatrix} = \begin{bmatrix} 0.14 - j0.46 & 0.09 - j0.34 \\ 0.16 + j0.48 & 0.19 + j0.67 \end{bmatrix} \begin{bmatrix} E_0^+ \\ E_0^- \end{bmatrix}$$

where E_0^+ , E_0^- , E_s^+ and E_s^- are respective peak values of transmitted electric field in air, reflected electric field from air and skin interface, transmitted electric field in skin and the reflected electric field from skin and fat interface.

2) Transmission Matrix at the interface between skin and fat

Reflection coefficient (Γ) at boundary between skin and fat can be calculated as,

$$\Gamma_{fs} = \frac{\eta_f - \eta_s}{\eta_f + \eta_s} = 0.31 - j0.02 \quad (7)$$

Propagation constant of fat is $\gamma_f = 16.63 + j169.33 \text{ m}^{-1}$.

Amplitude transmission matrix for fat layer,

$$[A]_{fat} = \begin{bmatrix} 0.91 - j0.32 & 0 \\ 0 & 0.97 + j0.34 \end{bmatrix}$$

Discontinuity transfer matrix for the interface between skin and fat layer is

$$[D]_{fat,skin} = \begin{bmatrix} 1 & -0.31 + j0.02 \\ -0.31 + j0.02 & 1 \end{bmatrix} \times (1.45 - j.04)$$

Then, transmission matrix for fat and skin can be expressed

as, $[T]_{fat,skin} = [A]_{fat} \times [D]_{fat,skin}$

$$[T]_{fat,skin} = \begin{bmatrix} 1.31 - j0.5 & -0.39 + j0.18 \\ -0.45 - j0.11 & 1.42 + j0.45 \end{bmatrix} \quad (8)$$

From Fig. 5.3 (a), it can be written as,

$$\begin{bmatrix} E_f^+ \\ E_f^- \end{bmatrix} = \begin{bmatrix} 1.31 - j0.5 & -0.39 + j0.18 \\ -0.45 - j0.11 & 1.42 + j0.45 \end{bmatrix} \begin{bmatrix} E_s^+ \\ E_s^- \end{bmatrix}$$

where E_s^+ , E_s^- , E_f^+ and E_f^- are respective peak values of transmitted electric field in skin, reflected electric field from fat and skin interface, transmitted electric field in fat and the reflected electric field from muscle and fat interface.

3) Transmission Matrix at the interface between fat and muscle

Reflection coefficient (Γ) at boundary between fat and muscle is calculated as,

$$\Gamma_{mf} = \frac{\eta_m - \eta_f}{\eta_m + \eta_f} = -0.38 + j0.01 \quad (9)$$

Propagation constant of muscle (γ_m) = $45.57 + j375.28 \text{ m}^{-1}$.

Amplitude transmission matrix for muscle layer,

$$[A]_{muscle} = \begin{bmatrix} 0.14 - j0.38 & 0 \\ 0 & 0.84 + j2.34 \end{bmatrix}$$

Discontinuity transfer matrix for the interface between fat and muscle layer is

$$[D]_{muscle,fat} = \begin{bmatrix} 1 & 0.38 - j0.01 \\ 0.38 - j0.01 & 1 \end{bmatrix} \times (0.72 + j0.01).$$

Therefore, transmission matrix for muscle and fat layers is

$$[T]_{muscle,fat} = [A]_{muscle} \times [D]_{muscle,fat}$$

$$[T]_{muscle,fat} = \begin{bmatrix} 0.10 - j0.27 & 0.04 - j0.10 \\ 0.23 + j0.63 & 0.58 + j1.69 \end{bmatrix} \quad (10)$$

From Fig. 5.3(a), it can be written as

$$\begin{bmatrix} E_m^+ \\ E_m^- \end{bmatrix} = \begin{bmatrix} 0.10 - j0.27 & 0.04 - j0.10 \\ 0.23 + j0.63 & 0.58 + j1.69 \end{bmatrix} \begin{bmatrix} E_f^+ \\ E_f^- \end{bmatrix}$$

where E_m^+ , E_m^- , E_f^+ and E_f^- are respective peak values of transmitted electric field in muscle, reflected electric field from muscle and superstrate interface,

transmitted electric field in fat and the reflected electric field from muscle and fat interface.

4) Transmission Matrix at the interface between fat and muscle

Overall transmission matrix for the 3-layer human body model can be written as

$$[T]_{muscle,air} = [T]_{muscle,fat} \times [T]_{fat,skin} \times [T]_{skin,air}$$

$$\begin{bmatrix} E_m^+ \\ E_m^- \end{bmatrix} = \begin{bmatrix} 0.11 + j0.06 & 0.59 + j0.01 \\ 0.23 + j0.63 & 0.58 + j1.64 \end{bmatrix} \begin{bmatrix} E_0^+ \\ E_0^- \end{bmatrix} \quad (11)$$

5) Equivalent Complex Permittivity Estimation

Three-layer body model is now replaced by a homogeneous phantom (Fig. 5.3 (b)) with relative permittivity, conductivity, attenuation constant, phase constant and intrinsic impedance of ϵ_r , σ , α , β and η respectively. E_0^+ , E_0^- , E_p^+ and E_p^- are respective peak values of transmitting electric field in air, reflecting electric field from phantom to air, transmitting electric field in phantom and the reflecting electric field from phantom and superstrate interface. The reflection coefficient for the interface between air and phantom is considered as Γ . Then, the transmission matrix can be written as

$$\begin{bmatrix} E_p^+ \\ E_p^- \end{bmatrix} = [T]_{phantom,air} \begin{bmatrix} E_0^+ \\ E_0^- \end{bmatrix}$$

$$\text{So, } \begin{bmatrix} E_p^+ \\ E_p^- \end{bmatrix} = \begin{bmatrix} \frac{e^{-\gamma h}}{1-\Gamma} & \frac{-(\Gamma)e^{-\gamma h}}{1-\Gamma} \\ \frac{-(\Gamma)e^{\gamma h}}{1-\Gamma} & \frac{e^{\gamma h}}{1-\Gamma} \end{bmatrix} \begin{bmatrix} E_0^+ \\ E_0^- \end{bmatrix}$$

Here, thickness of phantom is h_p and propagation constant is $\gamma = \alpha + j\beta$. Comparing the transmission matrix of the homogeneous phantom with the transmission matrix of three-layer body model, it can be written as

$$\frac{e^{-\gamma h_p}}{1-\Gamma} = 0.11 + j0.06 \quad (12)$$

$$\frac{e^{\gamma h_p}}{1-\Gamma} = -1.39 - j0.80 \quad (13)$$

By solving these equations, the final effective relative permittivity and effective conductivity of the homogeneous equivalent phantom are obtained as 52.7 and 1.9 S/m.

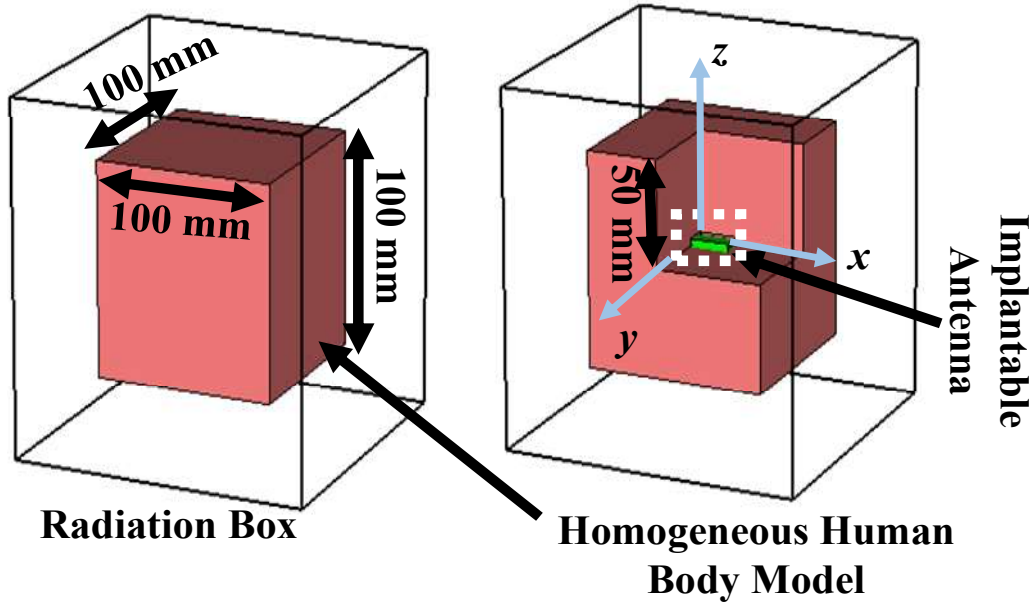


Fig. 5.4: Homogeneous human body model consider to simulate implantable antenna system

5.5 PERFORMANCE OF IMPLANTABLE ANTENNA WITHIN EQUIVALENT BODY PHANTOM

The corner-chamfered designed implantable antenna is encapsulated by biocompatible gelatin along with the prototypes of camera, PCB, sensor, battery to make a prototype of an Implantable Medical Device (IMD) as shown in Fig. 5.2. The IMD is placed within $100 \times 100 \times 100 \text{ mm}^3$ estimated equivalent cubical Homogeneous Human Body Phantom (HHBP) as shown in Fig. 5.4. HHBP is enclosed by an air box which is considered as radiation boundary of the implanted system. Numerical simulations are performed using Finite Element Method (FEM) method using CST MWS 2019 commercial software. Here, a Gaussian source is fed to the implanted antenna and frequency response is measured at 2.5 GHz. After simulating the antenna system within HHBP, the S_{11} is observed as -42.18 dB (Fig. 5.5). The antenna system is radiating with realized gain of -38.92 dBi (Fig. 5.6) and fractional bandwidth of 5.67 % at 2.5 GHz.

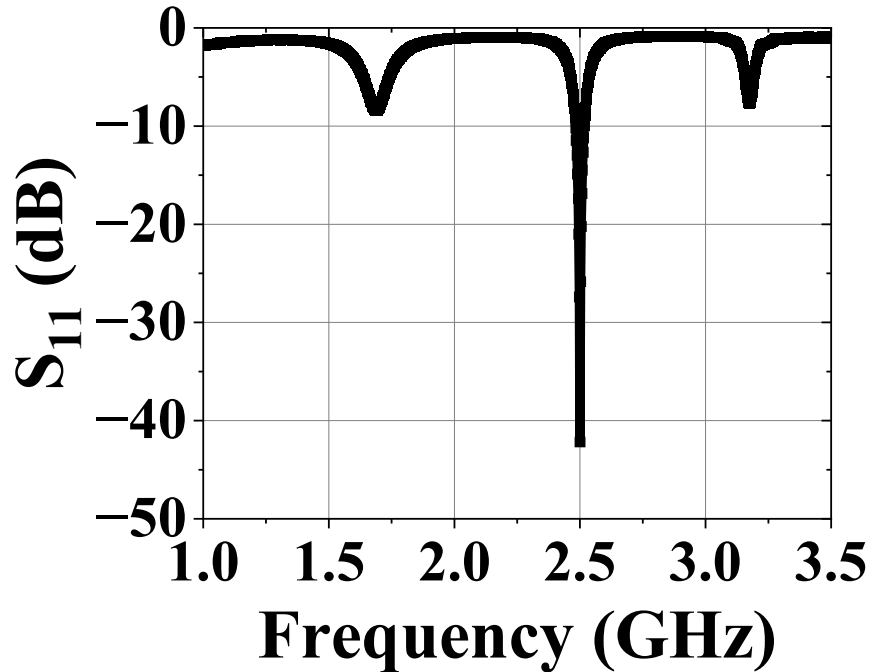


Fig. 5.5: S_{11} plot of implantable antenna system within HHBP

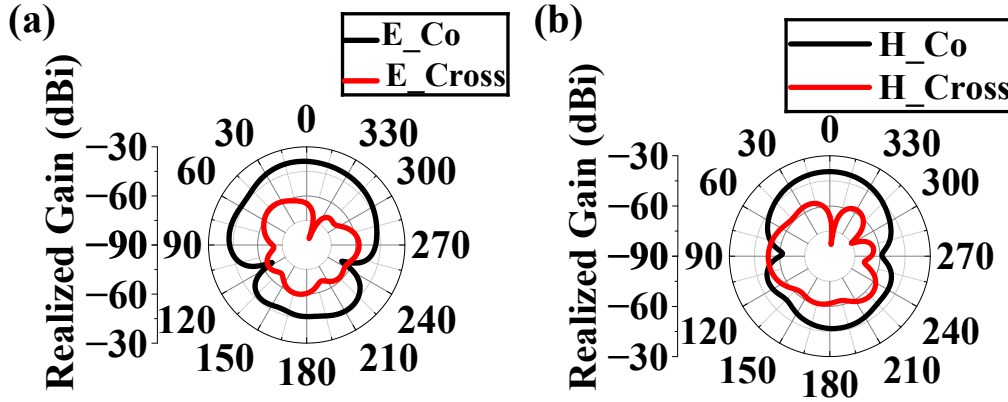


Fig. 5.6: Co and Cross polar plots of implantable antenna system within HHBP for (a) E- plane and (b) H-plane at 2.5 GHz

The safety performance of the implanted antenna was evaluated in terms of compliance with international safety guidelines, the IEEE Std. C95.1-2019 [24] guideline. Initially, 1 W average input power was applied to the implanted antenna to observe the SAR distributions for 1 g and 10 g contiguous models. The maximum SAR for the 1 g and the 10 g should be less than 1.6 W/kg and 2 W/kg, respectively, as per the standard guidelines [24]. The peak SAR values for the 1 g and 10 g cubical tissues were obtained as 327.49 W/kg and 69.47 W/kg respectively. The maximum input powers for 1 g and 10 g of body tissues were 8.24 mW and 27.98 mW respectively, for maintaining the SAR values within the standard SAR limits. For implantable antenna system, it is important to perform the link analysis to ensure the proper noiseless data transmission through high-lossy human body to the external monitoring devices.

For proper establishment of medical communication with external monitoring devices to conduct continuous patient monitoring, the link margin ($\mathcal{L}$) should be calculated using (1). Based on the SAR profile, the input power of the implantable antenna will be $P_{TX} = 8.24 \text{ mW} = 9.16 \text{ dBm}$.

$$\mathcal{L} (dB) = (\mathcal{R})_{Link}(dB) - (\mathcal{R})_{Required}(dB) \quad (14)$$

where

$$(\mathcal{R})_{Link}(dB) = \wp_{Tx} + \mathcal{G}_{Tx} + \mathcal{G}_{Rx} - 20 \log \left(\frac{4\pi d}{\lambda} \right)$$

$$(\mathcal{R})_{Required}(dB) = \frac{\mathcal{W}_b}{\mathcal{N}_0} + 10 \log_{10} \mathcal{B}_r - \mathcal{G}_C + \mathcal{G}_D$$

Table 5.2: Link Budget Analysis Parameters

	Parameters	Variable	Values
Transmitter	Frequency	f_r	2.5 GHz
	Transmitted Power	$\wp_{Tx}$	9.16 dBm
	Tx Antenna gain	$\mathcal{G}_{Tx}$	-38.92dBi
Receiver	Receiving Antenna gain	$\mathcal{G}_{Rx}$	4.95 dBi
	Temperature	T_O	293 K
	Boltzmann Constant	$\mathcal{K}$	1.38×10^{-23}
	Noise Power Density	$\mathcal{N}_0$	199.95 dB/Hz
Signal Quality	Distance	d	1-15m
	Ideal-BPSK	$\frac{\mathcal{W}_b}{\mathcal{N}_0}$	9.6 dB
	Coding gain	$\mathcal{G}_C$	0
	Fixing deterioration	$\mathcal{G}_D$	2.5 dB

The expression $(\mathcal{R})_{Link}$ represents the ratio of the Rx power received by patch antenna at any distance (d) to noise power density of body implanted chamfered meander lined antenna sensor. $(\mathcal{R})_{Required}$ denotes the ratio of required carrier power to noise power at Rx end. A simple patch antenna is designed as a receiver which is operating at 2.5 GHz with realized gain of 4.95 dBi. The Receiver is kept at variable distance from human body model as shown in Fig. 5.7.

In this work, Binary Phase Shift Keying (BPSK) modulation with normalized signal to noise ratio $\left(\frac{W_b}{N_o}\right)$ 9.6 dB is considered and $\mathcal{L} (dB)$ of the system for bit rates of 7 kbps, 100 kbps and 1 Mbps are calculated using parameters mentioned in Table 5.2. From Fig. 5.8, it is observed that the implantable antenna sensor is able to communicate at the distance of more than 15 m (~37m) for bit rate of 7 kbps. If the value of B_r is increased to 100 kbps, the transfer range is reduced to 10 m. For further increment of data rate to 1 Mbps, the implantable antenna can communicate within 3.5 m.

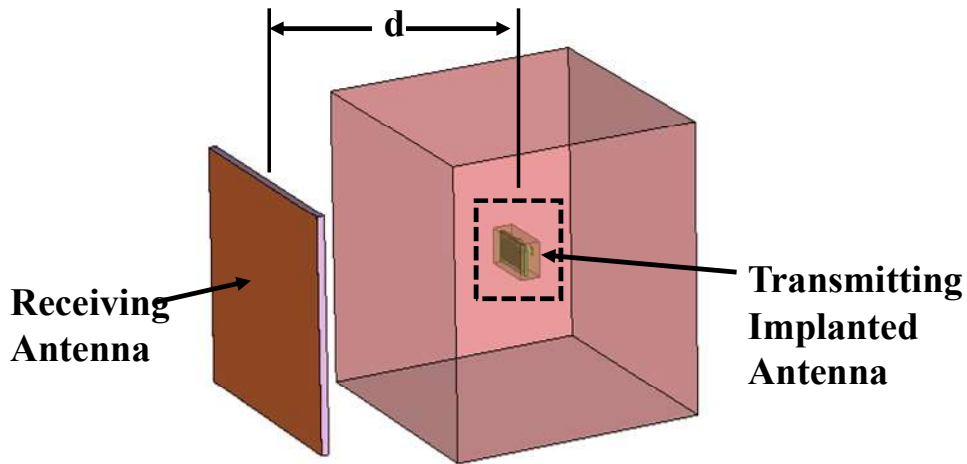


Fig. 5.7: Simulation Setup for Link Margin Analysis

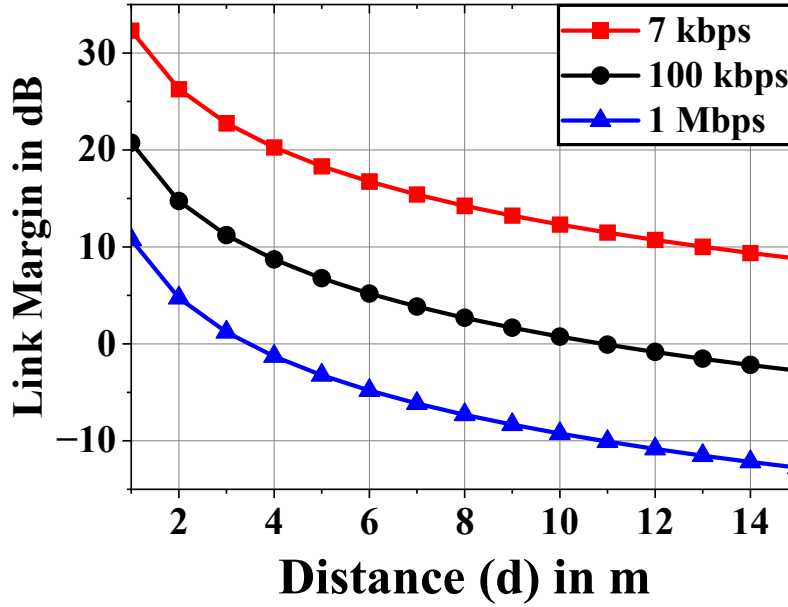


Fig. 5.8: Link Margin vs Distance Plots for data rates of 7 kbps, 100 kbps and 1 Mbps

Noise temperature is one of the important parameters of any system. Noise temperature is the measurement of equivalent level of available noise power introduced by a component or source. Let us consider the effective noise temperature of the system, which is T_{sys} . Therefore, the modified expression of the generalized link margin is given in (15).

$$\begin{aligned}
 L = & P_{Tx} + G_{Tx} + G_{Rx} - 20 * \log \left(\frac{4\pi d}{\lambda} \right) - \left(\frac{E_b}{N_o} \right) - 10 * \log(D_r) + G_C \\
 & - G_D - 10 * \log(T_{sys}) \quad (15)
 \end{aligned}$$

To characterize the wireless communication link between the implantable antenna and an external device, carrier link margin (CLM) and data link margin (DLM) were required to learn. CLM is the difference between the achieved carrier-to-noise power and the required carrier-to-noise power in human body communication, and DLM is the difference between the data signal-to-noise power

achieved and the data signal-to-noise power required. The CLM and DLM are calculated using (15) for 14 K noise temperature and 100 Kbps data rate and plotted in Fig. 5.9 for variable distance (d) between the receiver and transmitting implanted antenna.

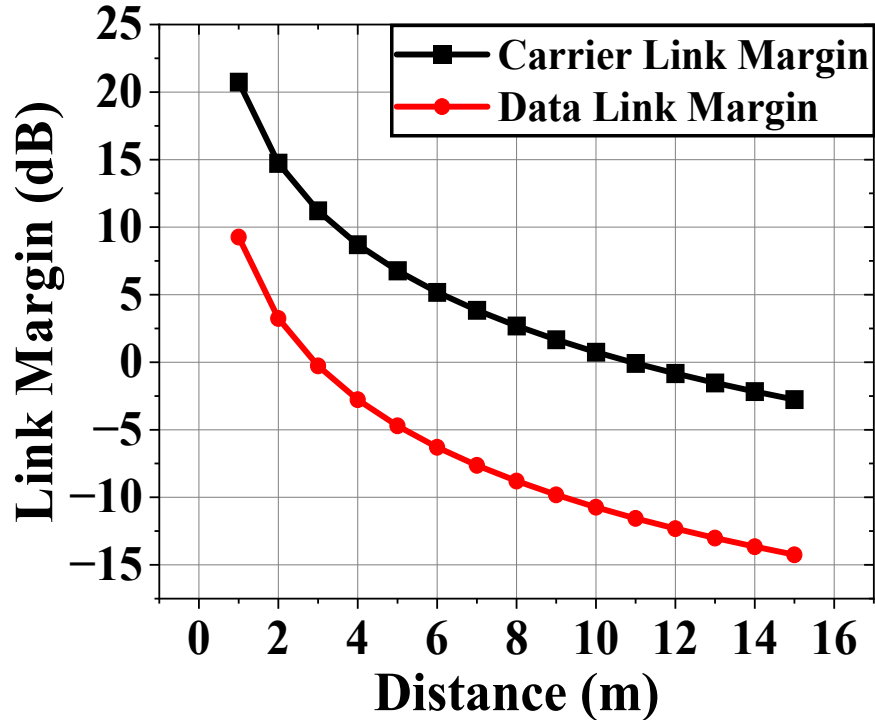


Fig. 5.9: Carrier-Link-Margin (CLM) and Data-Link-Margin (DLM) vs transfer distance (d) plot of the implanted system for noise temperature of 14K

5.6 DEPENDENCE ANALYSES AND DISCUSSION

It is enough to study the possible variation in link performance by considering $\pm 20\%$ variations in mean bioelectric properties of the overall human body [25-27]. In this section, dependence analyses of radiation (S_{11} , realized gain, operating frequency and fractional bandwidth), safety issues (SAR for 1g and

SAR for 10g tissues for 1Watt power) and communication (CLM and DLM for 100 kbps at 6 m distance) performances are studied for three cases: (i) variation in ϵ_{eff} by $\pm 20\%$, (ii) variation in σ_{eff} by $\pm 20\%$ and simultaneous variation in ϵ_{eff} and σ_{eff} by $\pm 20\%$. For each case, the mass density of the homogeneous body model is taken as 1000 Kg/m^3 .

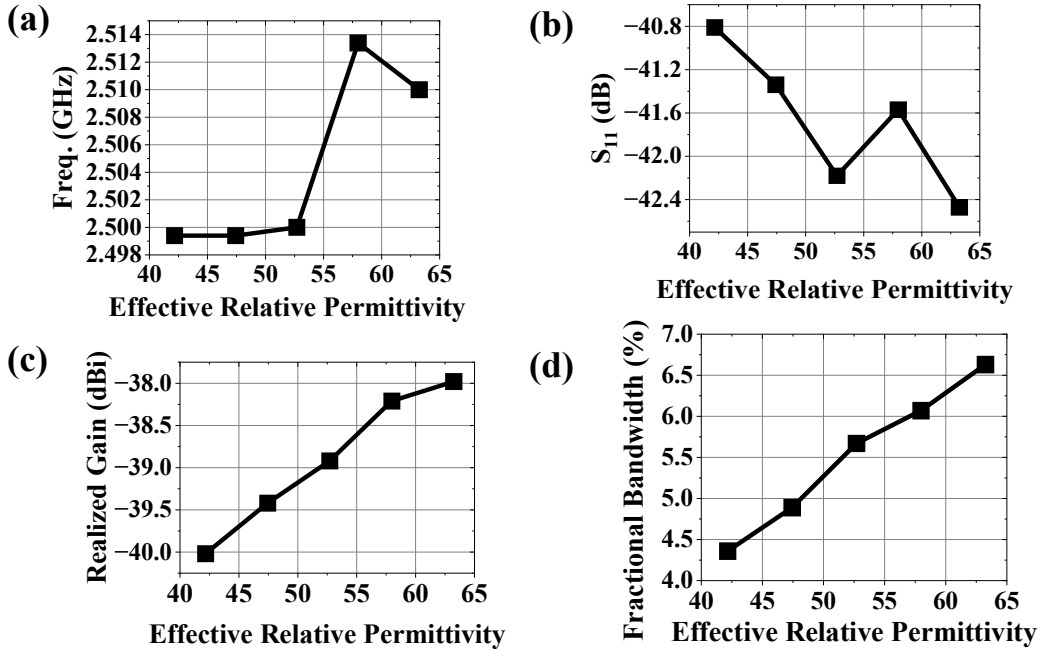


Fig. 5.10: Radiation performances in terms of Operating Frequency, S_{11} , Realized gain and Fractional Bandwidth for $\epsilon_{eff} \pm 20\%$

5.6.1. Variation in effective relative permittivity by $\pm 20\%$

1) Radiation Performance

The radiation performance of implantable antenna system is evaluated in terms of operating frequency, reflection coefficient, realized gain and fractional bandwidth. Their dependences on $\pm 20\%$ possible variation in effective relative

permittivity of human body are plotted in Fig. 5.8. From Fig. 5.10, it is observed that maximum -0.02%, -3.28%, -22.38% and -23.1% variations are computed in operating frequency, reflection coefficient, realized gain and fractional bandwidth respectively when effective relative permittivity is decreased by 20%. Similarly, respective maximum variations of 0.54%, 0.69%, 24.17% and 16.93% are observed for operating frequency, reflection coefficient, realized gain and fractional bandwidth values when effective relative permittivity is increased by 20%.

2) Safety Issues

The safety performance of implantable antenna system is evaluated in terms of SAR for 1g and SAR for 10g contiguous tissue models and plotted in Fig. 5.11 (a) and (b) respectively. From Fig. 5.11, it is observed that maximum -68.19% and -46.06% variations are computed in SAR for 1g and 10g respectively when effective relative permittivity is decreased by 20%. Similarly, respective maximum variations of -34.31% and 26.69% are observed for SAR due to 1g and 10g tissues when effective relative permittivity is increased by 20%.

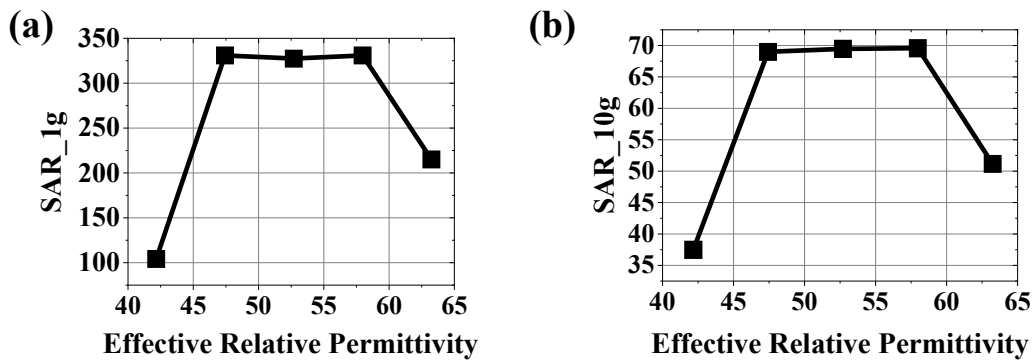


Fig. 5.11: Safety performances in terms of SAR of 1g and 10g tissue for $\epsilon_{eff} \pm 20\%$

3) Communication Performance

The communication performance of the implantable antenna system is computed in terms of CLM and DLM performances and depicted in Fig. 5.12 (a) and (b) respectively. From Fig. 5.12, it is observed that maximum -11.89% and 14.09% variations are obtained in CLM and DLM respectively when effective relative permittivity is decreased by 20%. Respective maximum variations of -79.81% and -52.63% are observed for CLM and DLM when effective relative permittivity is increased by 20%.

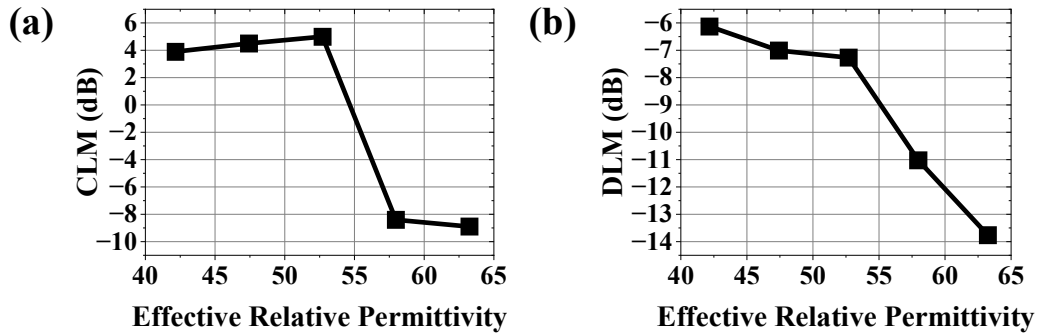


Fig. 5.12: Communication performances in terms CLM and DLM for $\epsilon_{eff} \pm 20\%$

5.6.2. Variation in effective conductivity by $\pm 20\%$

1) Radiation Performance

The dependences of radiation performances on $\pm 20\%$ possible variation in effective conductivity of human body are plotted in Fig. 5.13. From Fig. 5.13, it is observed that maximum -0.4%, -0.88%, -22.38% and -3.53% variations are computed in operating frequency, reflection coefficient, realized gain and fractional bandwidth respectively when effective conductivity is decreased by 20%. Similarly, respective maximum variations of 0.01%, 0.78%, 22.57% and 1.41% are observed for operating frequency, reflection coefficient, realized gain and fractional bandwidth values when effective conductivity is increased by 20%.

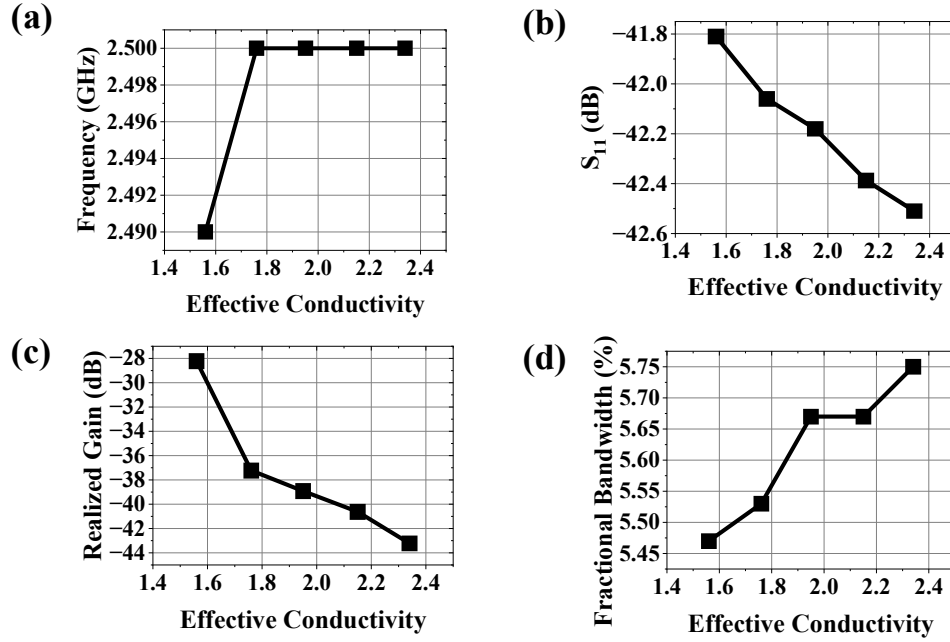


Fig. 5.13: Radiation performances in terms of Operating Frequency, S_{11} , Realized gain and Fractional Bandwidth for $\sigma_{eff} \pm 20\%$

2) Safety Issues

The safety performance of implantable antenna system is evaluated for $\pm 20\%$ variations in effective conductivity and plotted in Fig. 5.14 (a) and (b) respectively. From Fig. 5.14, it is observed that maximum -28.26% and -45.33% variations are computed in SAR for 1g and 10g respectively when effective conductivity is decreased by 20%. Similarly, respective maximum variations of 4.08% and -27.04% are observed for SAR due to 1g and 10g tissues when effective conductivity is increased by 20%.

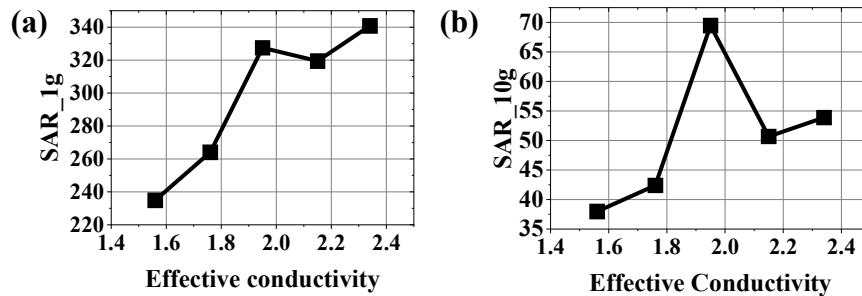


Fig. 5.14: Safety performances in terms of SAR of 1g and 10g tissue for $\sigma_{eff} \pm 20\%$

3) Communication Performance

The communication performance of the implantable antenna system is computed in terms of CLM and DLM performances for variable effective conductivity and depicted in Fig. 5.15 (a) and (b) respectively. From Fig. 5.15, it is observed that maximum -12.49% and 13.67% variations are obtained in CLM and DLM respectively when effective conductivity is decreased by 20%. Respective maximum variations of -69.21% and -55.83% are observed for CLM and DLM when effective conductivity is increased by 20%.

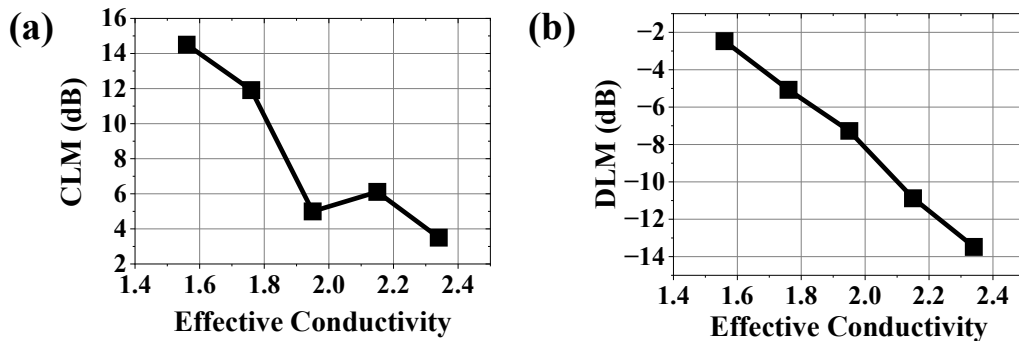


Fig. 5.15: Communication performances in terms CLM and DLM for $\sigma_{eff} \pm 20\%$

5.6.3. Simultaneous Variations in both effective relative permittivity and effective conductivity by $\pm 20\%$

1) Radiation Performance

The dependences of radiation performances on $\pm 20\%$ possible variations in both effective relative permittivity and conductivity of the human body are plotted in Fig. 5.16. Surface plots of the recorded operating frequency, S_{11} , realized gain and fractional bandwidth performed here. From these plots, it is observed that the maximum variations in operating frequency, S_{11} , realized gain and fractional bandwidth from mean condition are 0.4%, 25%, 86.29% and 29.21% respectively for this scenario.

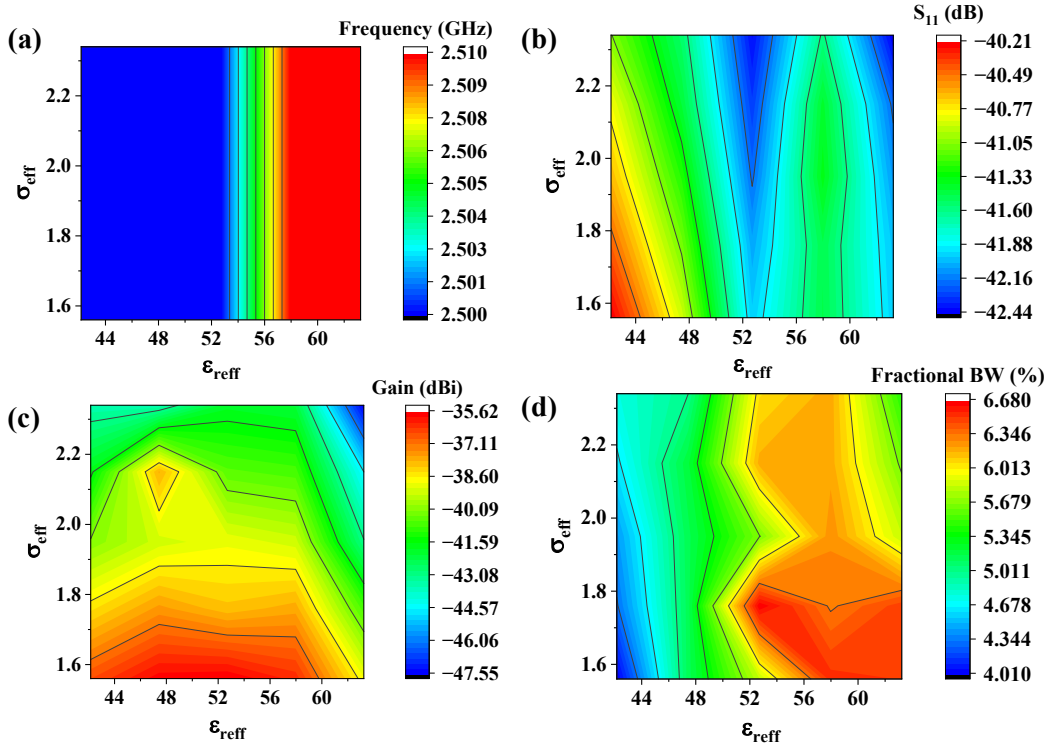


Fig. 5.16: Radiation performances in terms of Operating Frequency, S_{11} , Realized gain and Fractional Bandwidth for both $\epsilon_{\text{reff}} \pm 20\%$ and $\sigma_{\text{eff}} \pm 20\%$

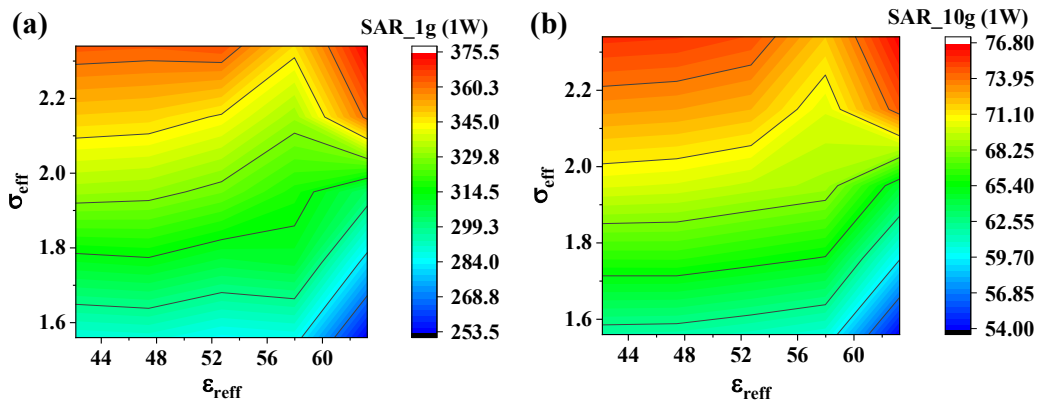


Fig. 5.17: Safety performances in terms of SAR of 1g and 10g tissue for both $\epsilon_{\text{reff}} \pm 20\%$ and $\sigma_{\text{eff}} \pm 20\%$

2) Safety Issues

The safety performance of implantable antenna system is evaluated for $\pm 20\%$ simultaneous variations in effective relative permittivity and effective conductivity and plotted in Fig. 5.17 (a) and (b) respectively. From Fig. 5.17, it is observed that maximum 15.29% and 10.42% variations are computed in SAR for 1g and 10g respectively.

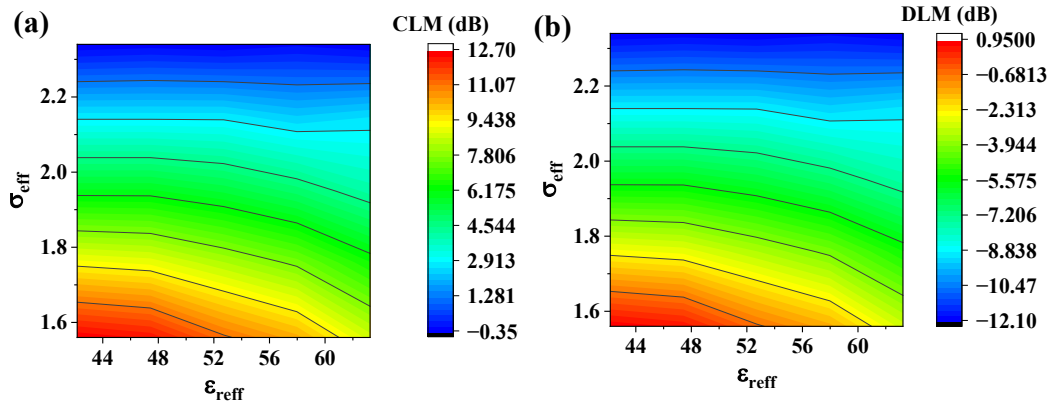


Fig. 5.18: Communication performances in terms CLM and DLM for both $\epsilon_{\text{reff}} \pm 20\%$ and $\sigma_{\text{eff}} \pm 20\%$

3) Communication Performance

The communication performance of the implantable antenna system is computed in terms of CLM and DLM performances at 6 m distance from implanted antenna at a data rate of 100 kbps for variable effective relative permittivity and effective conductivity and depicted in Fig. 5.18 (a) and (b) respectively. From Fig. 5.18, it is observed that maximum 78.69% and 15.47% variations are obtained in CLM and DLM respectively. From a clinical standpoint, these variations represent the natural diversity in human tissue properties due to differences in age, hydration, body mass index (BMI), or disease states. A drop in link margin from 15 dB to 5 dB means the device remains operational, as any margin above 0 dB theoretically maintains the link. However,

a 5 dB margin drastically reduces the system's resilience against patient movement, body orientation changes, and environmental noise. In a clinical deployment, this significantly increases the risk of data loss, necessitating robust error-correction layers and data retransmission protocols to ensure patient safety and continuous monitoring.

5.7 EXPERIMENTAL VALIDATION

In this section, in-vitro performance analysis of simulated implantable antenna system is described. The physical prototypes of implantable antenna and receiving antenna were fabricated. The fabricated prototype of implantable antenna is shown in Fig. 5.19 (a). Small battery ($\pi \times 3.95^2 \times 3.6 \text{ mm}^3$), prototype of PCB (Arlon AD 430 covered with copper in both sides), prototypes of sensor and camera (Arlon AD 430) were taken to replicate the simulated antenna system as shown in Fig. 5.19 (a). To excite the antenna, 50Ω coaxial cable was connected using soldering (Fig. 5.19 (b, c)) and other end was connected with one of the two ports of vector network analyzer (VNA) for S-parameter measurement as illustrated in Fig. 5.19 (d). Top and bottom views of implantable antenna system before coating are shown in Fig. 5.19 (b) and (c) respectively.

The whole system was covered with biocompatible gelatin cover made of 15g DI water and 10g gelatin which is depicted in Fig. 5.19 (e). For in-vitro measurement, homogeneous semi-solid body phantom model was developed by 66.3% DI water, 31.6% DGME, 0.4% NaCl and 1.7% Agar agar powder. The dielectric properties in terms of relative permittivity and conductivity of the phantom were measured using Agilent 85070E dielectric kit. The developed phantom had effective relative permittivity and conductivity of 51.75 and 1.91 S/m respectively at 2.5 GHz. The designed antenna system was then implanted within semi-solid human body phantom at 50 mm depth from top surface as

shown in Fig. 5.19 (d) to measure the reflection coefficient of the implanted antenna. The comparison between simulated and measured S_{11} responses was plotted in Fig. 5.19 (f). The measured S_{11} was -22.97 dB at 2.43 GHz. The antenna system had measured realized gain of -32.68 dBi at its operating frequency. The radiation pattern measurement setup of implantable antenna is shown in Fig. 5.20. The simulated and measured E and H plane plots of the designed antenna within homogeneous human body model are illustrated in Fig. 5.21.

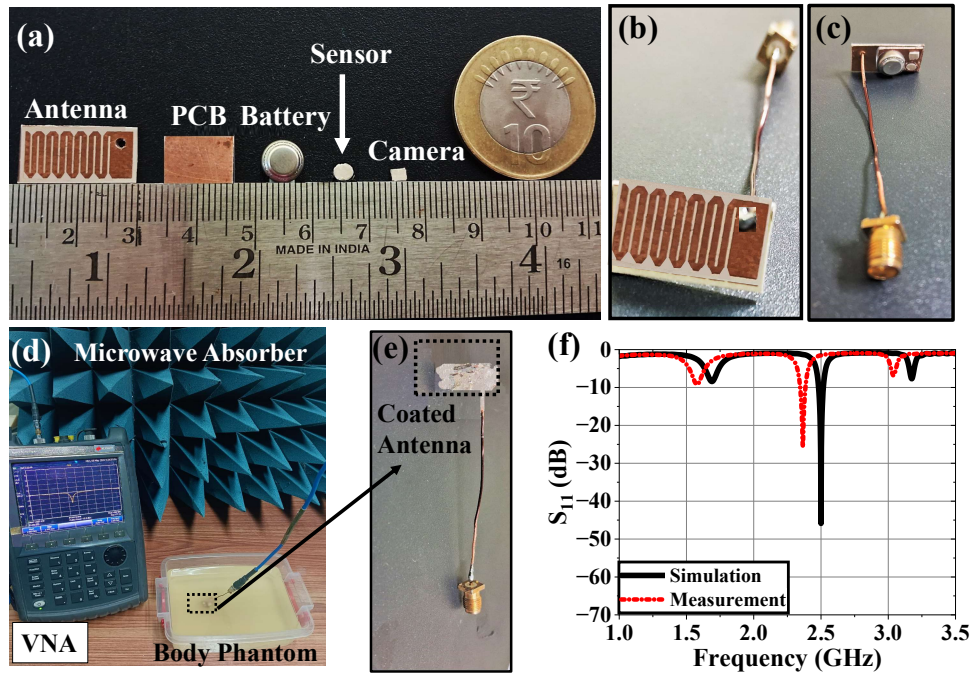


Fig. 5.19: (a) Fabricated prototypes of all components of Implantable antenna system, (b) top and (b) bottom views of implantable antenna system without cover, (d) reflection coefficient measurement setup of implantable antenna, (e) top view of antenna system with capsule and (f) comparison between simulated and measurement

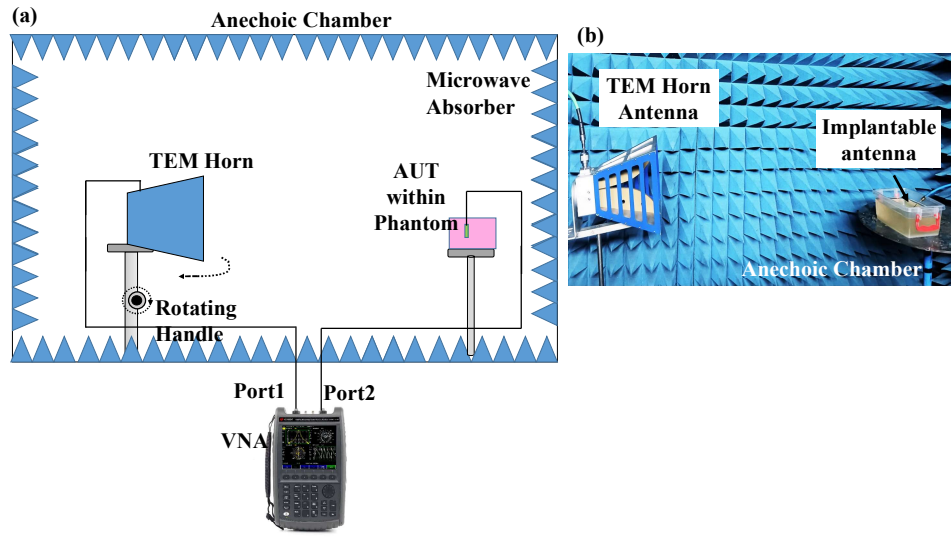
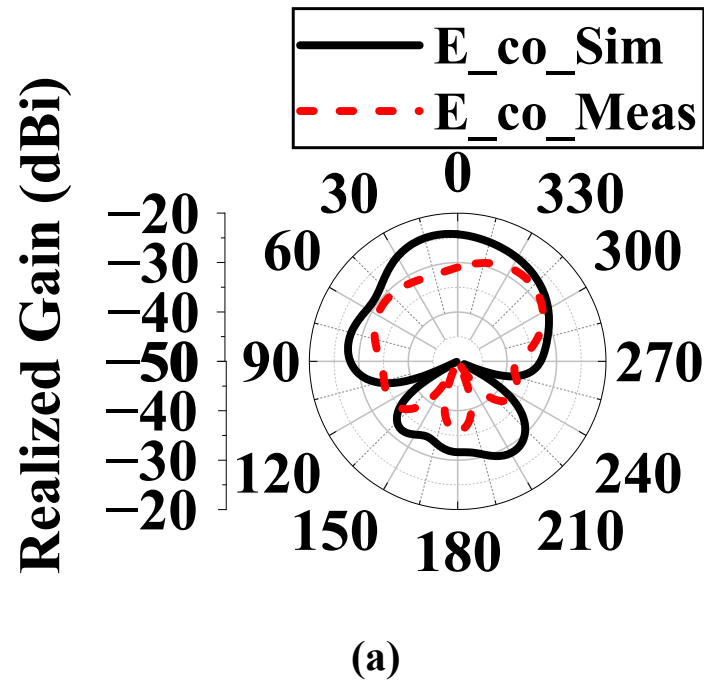
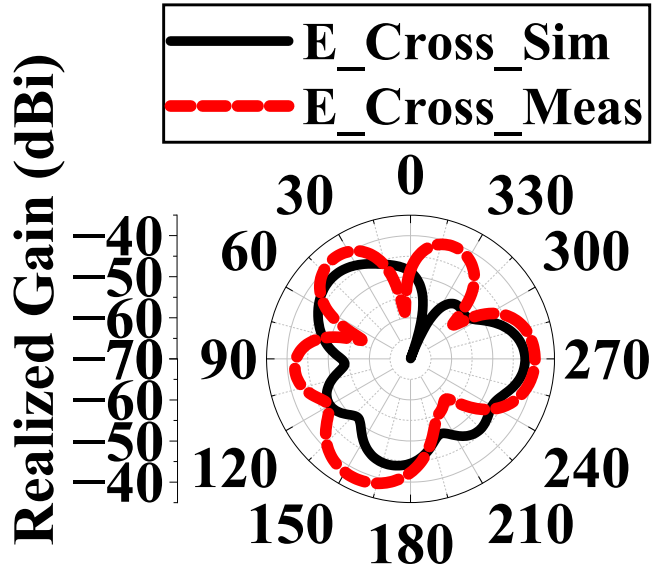
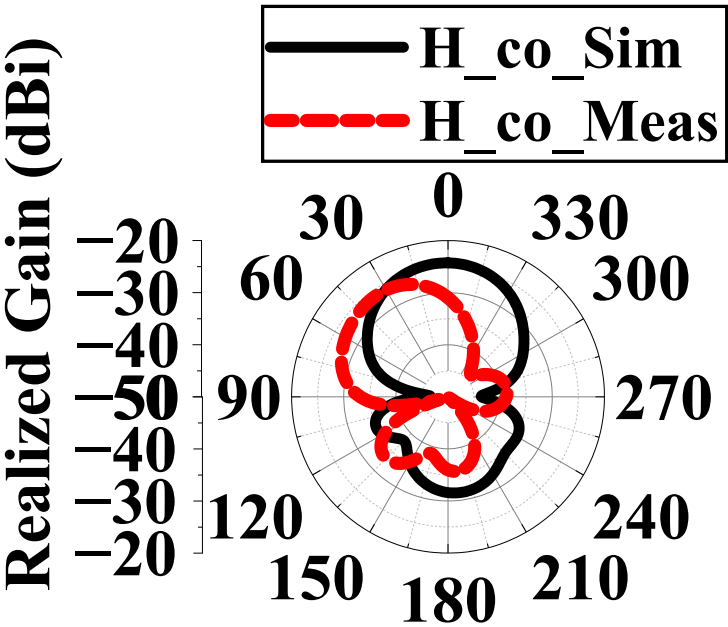


Fig. 5.20: Radiation Pattern measurement setup





(b)



(c)

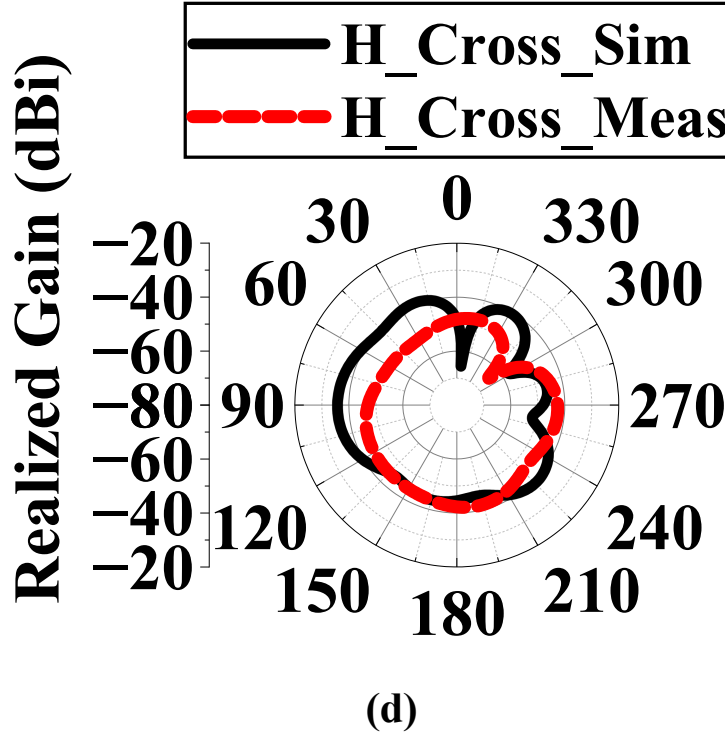


Fig. 5.21: Co and cross polar radiation patterns at (a-b) E plane and (c-d) H plane obtained in simulation and measurement

5.8 NOVELTY ANALYSIS

The novelty of this work compared with previous works reported in [28-34] is highlighted in Table 5.3. The relevance of this work is mentioned below.

1) Table 5.3 concludes that this is the first work where the CLM and DLM of muscle-implanted antenna with an external monitoring device are performed.

2) The SAR value for 1 g tissue for 1W input power is also good compared to others for ensuring the patient's safety after implantation.

3) Radiation and reflection performances in terms of S_{11} and gain within the body model are better than many other implantable antennas.

4) The authors in [32] designed an implantable antenna which could communicate up to 13m variation. In [34], the effect of phantom size on the gain

and the efficiency of the spiral-implanted antenna is discussed. The possibility of variations in S_{11} and frequency of an implanted antenna due to variation in relative permittivity of the body was observed in [35]. In this work, a comprehensive approach has been taken considering the impact of both human body electrical parameters, such as effective relative permittivity and conductivity, on the link margins (CLM, DLM), operating frequency, realized gain, data rate, channel loss, range of communication, bandwidth, and SAR performance of the implanted antenna.

Table 5.3: Comparison with Similar Works

Ref.	Freq. (GHz)	S_{11} (dB)	Gain (dBi)	SAR _{1g} (1W) (W/Kg)	CLM (dB)	DLM (dB)	Variation in parameters Tested
[28]	2.4	-25	-24.7	697.5	No	No	No
[29]	0.4	-20	-18.9	No	No	No	No
[30]	0.915, 2.45	-25, -40	-30.47, -24.71	658, 589	No	No	No
[31]	0.915, 2.45	-20, -37	-36, -30.1	333	No	No	No
[32]	0.915	-20	-23.23	270.3	No	No	Sensing Performance
[33]	0.915, 2.45	-19, -15	-26.30, -20.9	306.19, 252.36	No	No	No
[34]	0.402, 2.45	No	-37, -24.5	No	No	No	Gain, efficiency
[35]	2.45	-11	No	No	No	No	S_{11} and frequency
This Work	2.5	-42.2	-38.92	327.49	20.73 (d = 1m, Ts =13K) (First)	9.28 (d = 1m, Ts =13K) (First)	S_{11}, frequency, Gain, BW, CLM, DLM, bandwidth and SAR performance (First)

5.9 CONCLUSION

In this chapter, designed corner chamfered meander-line implantable antenna printed on Arlon AD 430 substrate is placed within homogeneous human body model equivalent to the three-layer (skin-fat-muscle) model using proposed Transmission Matrix Method (TMM) to overcome the problems related to multilayer phantom development. The simulated radiation (realized gain, return loss, operating frequency and fractional bandwidth) and communication (CLM and DLM) performances and safety issues (SAR profiles) are recorded and experimentally validated. The variation analyses of these performance parameters due to $\pm 20\%$ variations in mean relative permittivity and conductivity are also performed here.

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Chapter 6

Ingestible Medical Device Design for Wireless Capsule Endoscopy

Contents	6.1. Introduction 6.2. Problem Formulation 6.3. Ingestible Antenna System Design 6.4. Simulated Responses within Gustav Model 6.5. Experimental Setup and Measurement 6.6. Link Budget Analysis 6.7. Novelty of this work 6.8. Conclusion
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6.1. INTRODUCTION

The attractiveness of Wireless Capsule Endoscopy (WCE) is interesting because of its enabling diseases observation using physiological signals and real-time biological images transferred from swallowed or implanted ingestible medical (IIM) devices. IIM sensors transmit patients' health parameters (gastrointestinal imaging and tumor detection) to external diagnostic devices while passing through gastrointestinal (GI) tract [1]. The WCE system can capture images with a high-resolution inside GI tract while traversing along it [2]. A WCE, generally comprises an antenna for wireless transmission of sensed health-parameters, camera, battery, PCB, data management system etc., has typical dimensions of $26 \times 11 \text{ mm}^2$ [3-4]. Ingestible antennas must overcome some obstacles, like small size, high-speed communication, human safety, and a reliable communication link [5].

Various ingestible antennas have been documented to meet these WCE problems [1–8]. For WCE systems, single band ingestible antennas were proposed

in the works [1-2], [7-8]. Although the reported antennas in [2], [7-8] were not well-minimized in terms of size, the radiation performances of the WCE antennas in [1-2] and [8] showed acceptable gain and bandwidth responses. Due to changes in the tissue milieu of the GI tract, detuning may result from limited bandwidth in [7] and [9].

In [4-6], dual-band communication for WCE applications was accomplished by adding open-ended slots to the ground and radiator planes. The benefit of high-speed data transmission and reception at two distinct frequencies is offered by dual-band antennas. The antennas' gains are impacted in some way by the ground's inclusion of slots to achieve dual-band nature. Growing gain performances had an impact on the Impedance Bandwidth (IBW) performances in the works [4] and [6]. Reducing the WCE antenna system's strength also decreases the communication range and increases noise in data transfer from the GI tract to the outside world. In order to communicate over long distances while traveling through the GI tract, a small, broadband implanted antenna with a high gain is therefore needed.

6.2. PROBLEM FORMULATION

In this chapter, an open-slotted miniaturized ($\pi \times 3^2 \times 0.762 \text{ mm}^3$) ingestible antenna system is designed for WCE applications at 2.45 GHz (Industrial, Scientific, and Medical (ISM) band) and 4.55 GHz (n79 5G New Radio C- band). The 4.55 GHz belonging to the sub-6 GHz is favored because it has fewer propagation losses than higher frequencies, enabling wider coverage. In addition, it provides enough bandwidth for low-latency, high-speed data transport [9]. Due to having the 5G NR capabilities, the proposed system has the ability to increase network capacity and improve the speed of communication with an ultra-low latency transmission [10]. The overview of this work is shown in Fig. 6.1.

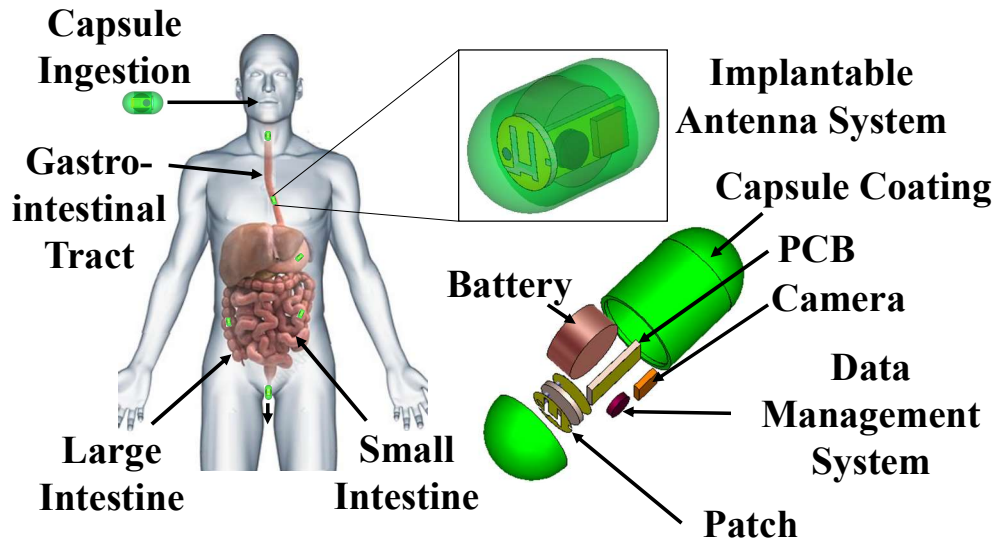


Fig. 6.1: Overview of ingestible antenna system for Wireless Capsule Endoscopy

6.3. INGESTIBLE ANTENNA SYSTEM DESIGN

6.3.1. Structural Design

An open-ended slotted circular antenna [Fig. 6.2 (a)] is designed to implement a compact dual-band ingestible antenna for wireless capsule endoscopy. The antenna is designed over a 0.762 mm Taconic CER-10 substrate ($\epsilon_r = 10$, $\tan \delta = 0.0035$) measuring $\pi \times 3^2 \times 0.762 \text{ mm}^3$. A full ground is present at the bottom of the substrate [Fig. 6.2 (b)]. The design parameters of the antenna are mentioned in Table 6.1.

A coaxial probe with 0.63 mm probe radius (d_p) providing an input impedance of 50Ω is used for exciting the antenna system. The designed antenna, along with the camera, battery, data management system, and printed circuit board (PCB), is placed within a biocompatible Polyvinyl Alcohol (PVA) capsule. Fig. 6.2 (c) illustrates an exploded view of the designed implantable medical capsule device.

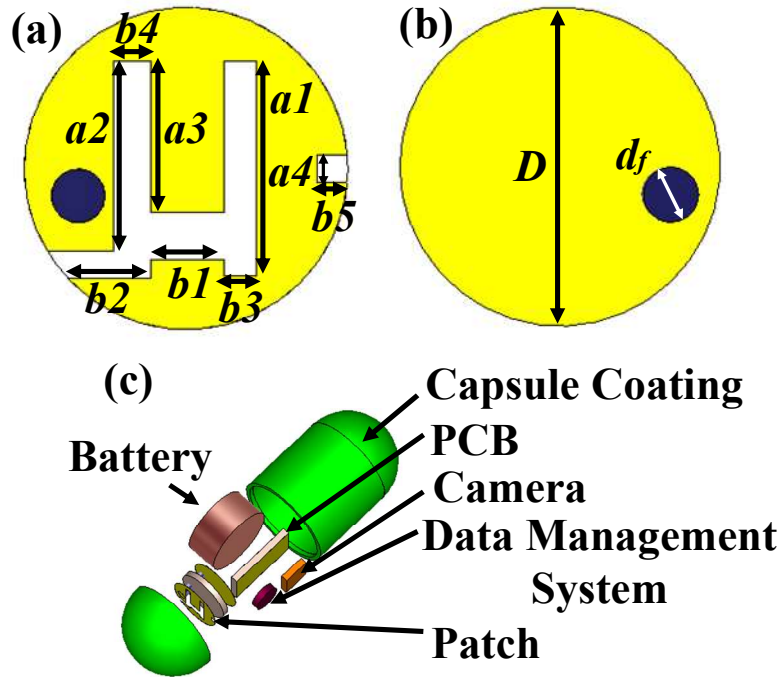


Fig. 6.2: Different views of proposed system: (a) Top, (b) Bottom and (c) Exploded views

Table 6.1: Design Parameters of the WCE Ingestible Antenna

Parameters	Values	Parameters	Values
D	6 mm	$b1$	1.35 mm
$a1$	4 mm	$b2$	1.54 mm
$a2$	3.5 mm	$b3$	0.6 mm
$a3$	2.8 mm	$b4$	0.7 mm
$a4$	0.5 mm	$b5$	0.6 mm

6.3.2. Simulation Environment

The designed system is implanted within a $100 \times 100 \times 100 \text{ mm}^3$ cubical mean Homogeneous Body Model (HBM) at the center as shown in Fig. 6.3. The

relative permittivity and conductivity of the HBM are set to be frequency-dependent as considered in [8].

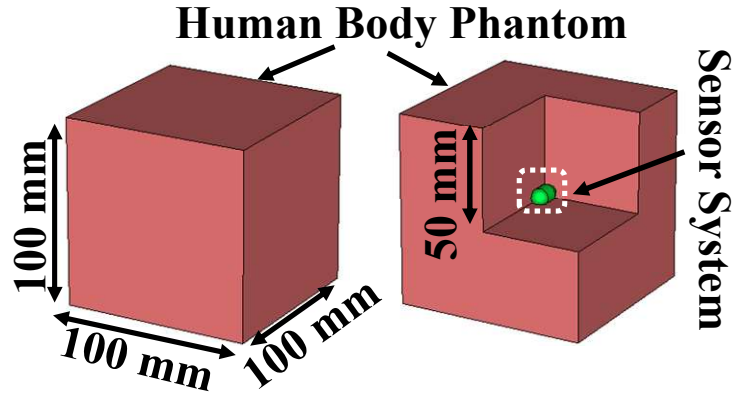


Fig. 6.3: Homogeneous simulation setup of ingestible antenna system

6.3.3. Design Evolution

In Step 1, a circular patch ($\pi \times 3^2 \times 0.762 \text{ mm}^3$) is adopted [Fig. 6.4 (a)]. The $|S_{11}| > -6 \text{ dB}$ is achieved by this implanted antenna around 4.5 GHz. However, the objective is to tune the antenna at both 2.45 GHz ISM band and 4.5 GHz n79 5G band. Therefore, an L-type open slot is cut on radiator in Step 2 [Fig. 6.4 (b)] and corresponding reflection coefficient plot is shown in Fig. 6.5. Due to incorporation of the slot, the antenna is tuned to 2.88 GHz with S_{11} value of -28.59 dB. For a further improvement, a horizontal slot is added on radiator in the following step.

In Step 3, resonant frequency is shifted to 2.66 GHz with a reflection coefficient of -20.17 dB. The S_{11} curve also showed a substandard but essential match around 4.03 GHz ($S_{11} = -8.23 \text{ dB}$) [Fig. 6.5]. A vertical slot ($0.6 \times 4 \text{ mm}^2$) is then added in Step 4 [Fig. 6.4 (d)]. The operating frequency is shifted from 2.66 GHz to 2.04 GHz. The S_{11} is observed as -19.57 dB. An improvement in S_{11} response towards -10 dB at second resonance is observed in Fig. 6.5.

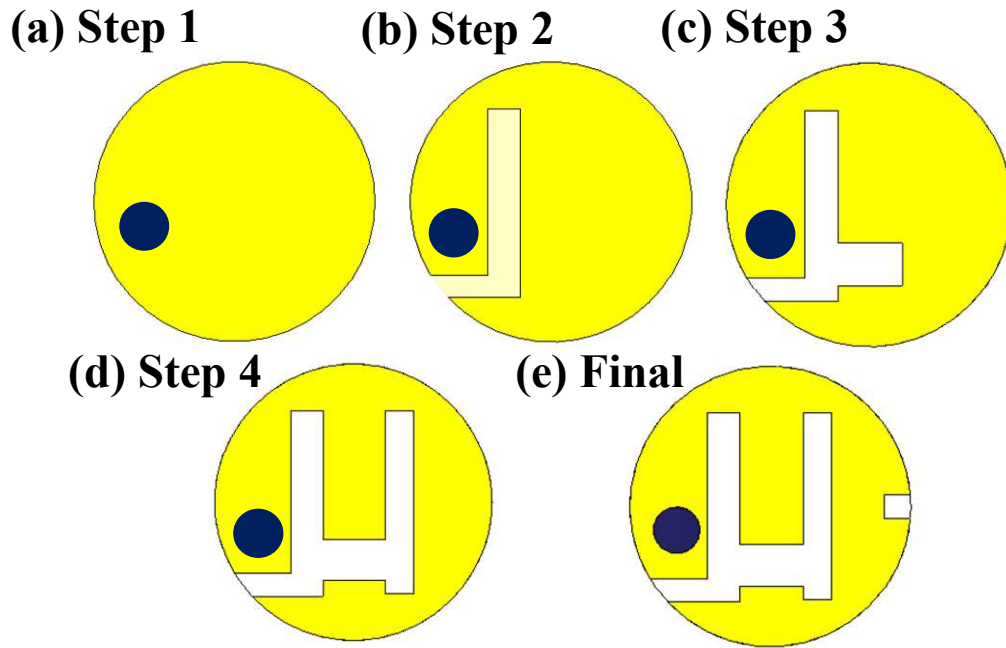


Fig. 6.4: Design Evolution Steps of the Antenna Part

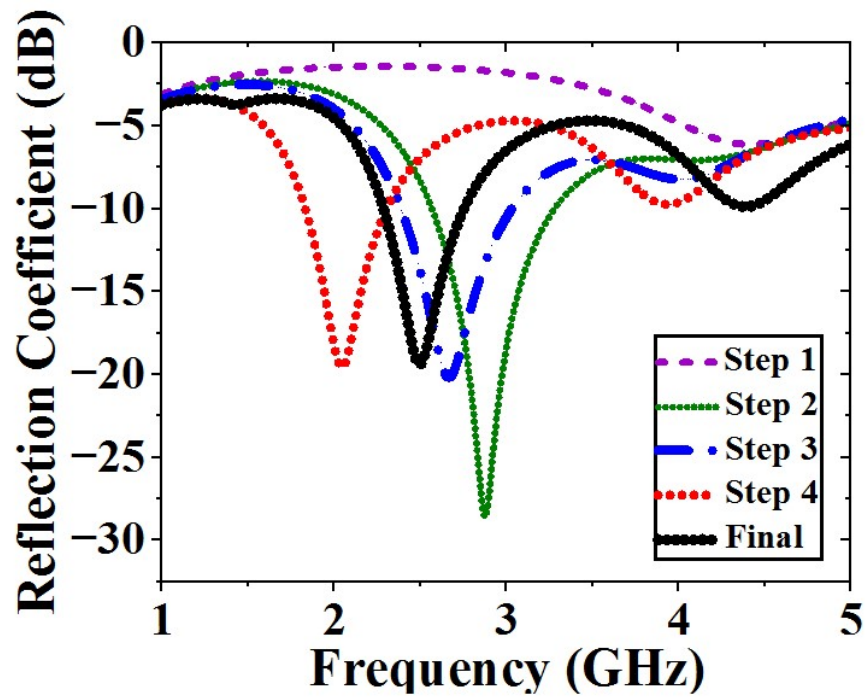


Fig. 6.5: Reflection coefficient plots of all steps

Finally, another open-ended slot (0.5 mm X 0.6 mm) [Fig. 6.4 (e)] is incorporated on patch. Due to this incorporation, the antenna within human body is able to communicate in two frequencies – 2.45 GHz and 4.55 GHz with reflection coefficients of -20.61 dB and -12.48 dB respectively. Therefore, the objective is achieved in Step 5. This final antenna is radiating with fractional bandwidths of 20.42% (2.23 – 2.75 GHz) and 13.85% (4.31 – 4.92 GHz) at 2.45 GHz and 4.55 GHz within phantom, respectively.

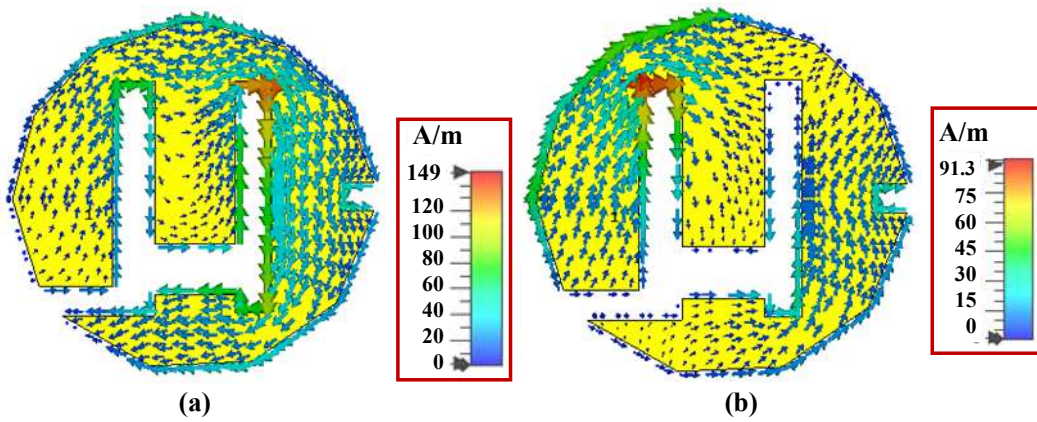


Fig. 6.6: Surface current distribution at (a) 2.45 GHz and (b) 4.55 GHz

6.3.4. Simulated Responses of Final Antenna

Fig. 6.6 illustrates the surface current distributions of the proposed antenna within HBM at 2.45 GHz and 4.55 GHz. The currents are highly dense at the edges of the open slots. The antenna radiates with the simulated realized gains of -24.08 dBi and -26.11 dBi at 2.45 GHz and 4.55 GHz respectively. Updated IEEE C95.1-2019 [11-12] limits the peak 1-g and 10-g SAR values below 1.6 and 2 W/kg respectively (mass density (ρ) = 1000 kg/m³). The coaxial feeding method is applied here. 1-g SAR is the peak spatial-average SAR computed over a cube of 1 g of tissue. SAR information at 2.45 and 4.55 GHz are tabulated in Table 6.2.

Table 6.2: SAR Profile of WCE system within HBM

Freq. (GHz)	SAR (W/Kg)		Max. allowable input power (mW)	
	1-g	10-g	1-g	10-g
2.45	220.26	59.73	7	24
4.55	321.52	68.38	5.29	11.75

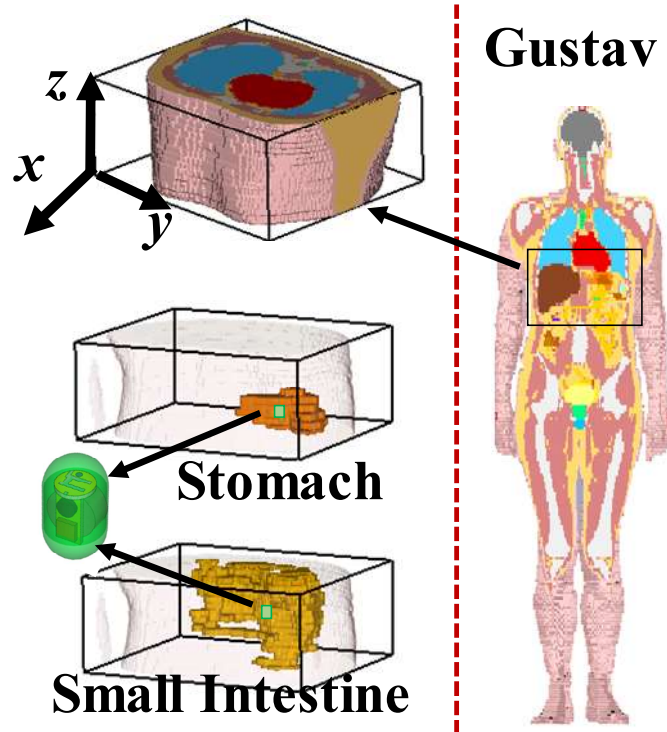


Fig. 6.7: Simulation setup in CST Gustav Model

6.4. SIMULATED RESPONSES WITHIN GUSTAV MODEL

The robustness of the proposed WCE system is assessed using the CST Gustav phantom model to confirm its stability in variable environment as shown in Fig. 6.7. The device performances within the CST Gustav model (Stomach and Small intestine) and the HBM are tabulated in Table 6.3. A comparison of simulated 1-g SAR responses of the implantable antenna for within HBM and

Gustav model is performed in Fig. 6.8. According to this table, the antenna retains consistent gain, IBW, SAR, and good impedance matching at both frequencies with the change of implanted environment.

Table 6.3: Performance within Different Simulation Environments

Model	Freq (GHz)	S ₁₁ (dB)	Gain (dBi)	IBW (%)	SAR_1g (W/Kg)
HBM	2.45	2.45	-24.08	20.42	220.26
	4.55	4.55	-26.11	13.85	321.52
Stomach	2.41	-28.49	-25.97	19.23	273.65
	4.46	-10.69	-27.35	11.57	347.86
Small Intestine	2.49	-15.76	-23.65	18.75	259.65
	4.52	-11.17	-26.28	9.66	369.72

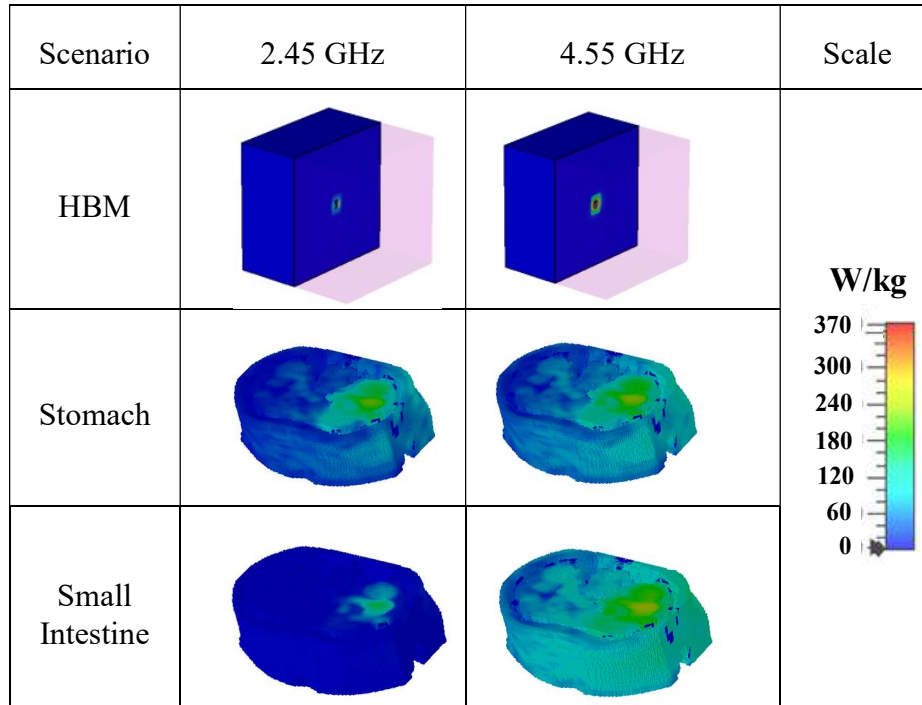


Fig. 6.8: Simulated 1-g SAR distributions within HBM and Gustav model

6.5. EXPERIMENTAL SETUP AND MEASUREMENT

The physical prototype of the ingestible antenna is fabricated. The fabricated prototype of the implantable antenna is shown in Fig. 6.9. A small lithium battery, prototype of PCB, prototypes of a data-management system and camera (Arlon AD 430) are taken to replicate the simulated antenna system. The system is placed within a polyvinyl alcohol (PVA) capsule.

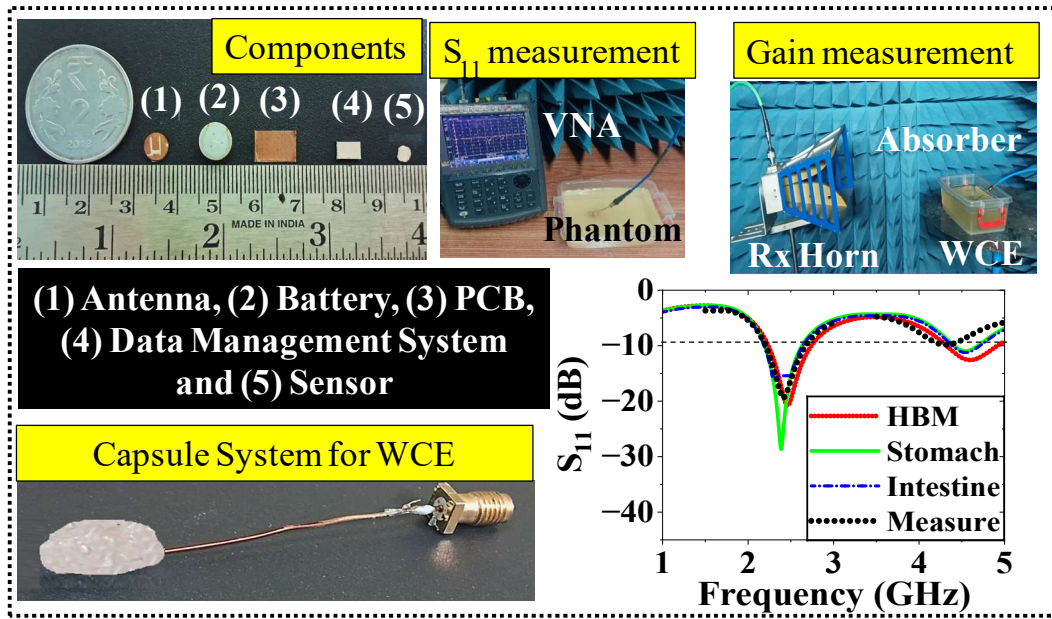


Fig. 6.9: The measurement setups for S_{11} and gain measurement of the fabricated WCE prototype and comparison of simulated S_{11} in different tissue environments.

For the purpose of practically measuring the system at 2.45 GHz and 4.55 GHz, two homogeneous phantoms are created based on FCC recommendation [12] where the composite human body comprising of all different tissues is considered. 66% DI water, 31.5% DGME, 0.5% NaCl, and 2% Agarose powder are thoroughly mixed to produce the body phantom at 2.45 GHz. To generate phantom at 4.55 GHz, a mixture of 52% DI water, 45.5% DGME, 0.5% NaCl, and 2% Agarose powder is made. The comparison between simulated and

measured S_{11} responses is plotted in Fig. 6.9. The measured S_{11} were -18.99 dB at 2.39 GHz and -10.05 dB at 4.28 GHz.

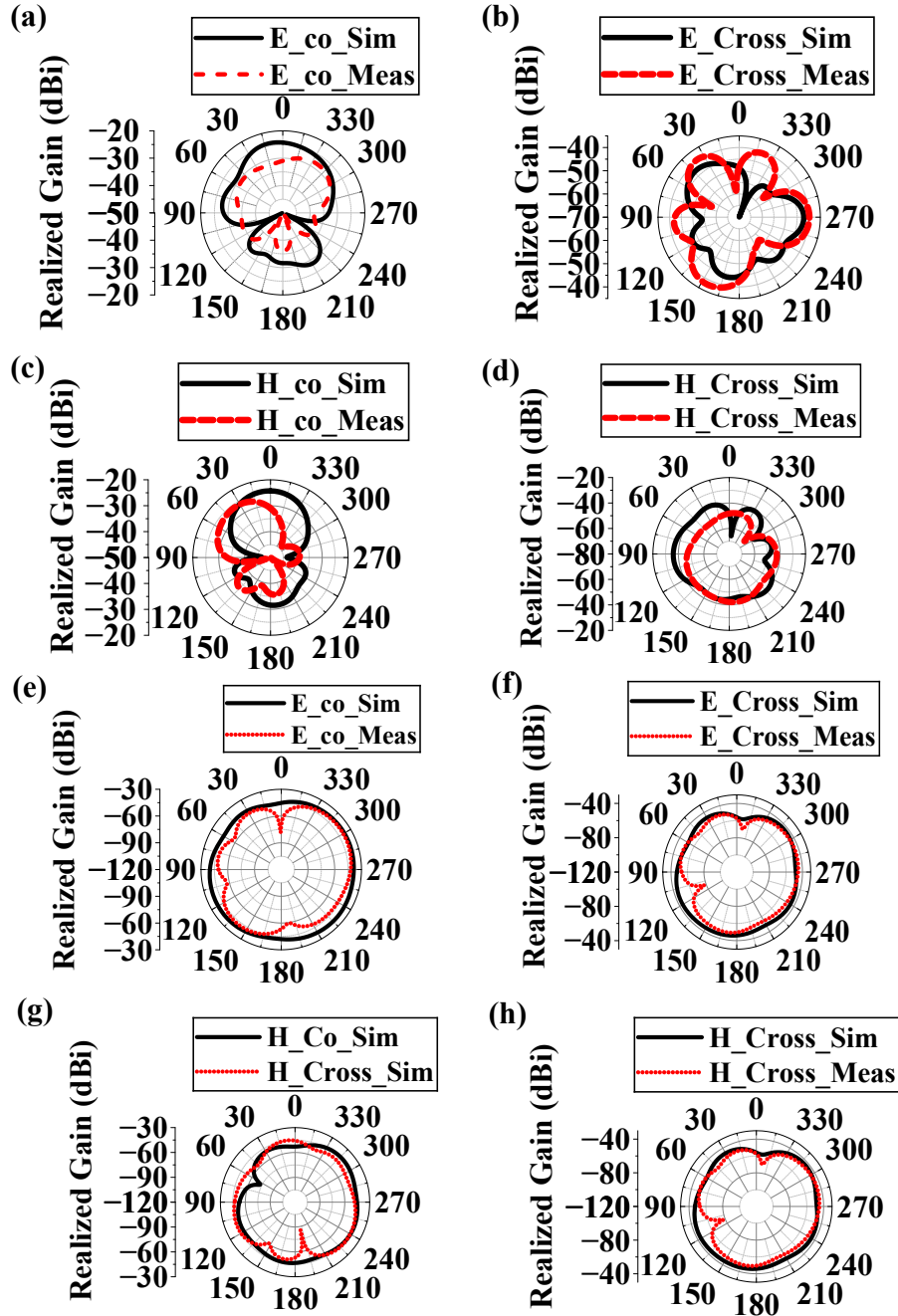


Fig. 6.10: Simulated and measured far-field radiation co and cross polar patterns at 2.45 GHz (a-d) and 4.55 GHz (e-h)

The measured fractional bandwidths are 20.08% and 10.69% around measured operating frequencies 2.39 GHz and 4.28 GHz respectively. Simulated and measured co- and cross-polar radiation patterns of the proposed implantable antenna system at 2.45 GHz and 4.55 GHz for E and H planes are plotted in Fig. 6.10. Measured realized gains of the fabricated system within semisolid phantom are -29.28 dBi and -32.47 dBi at respective measured operating frequencies. The high gain of the proposed antenna WCE promises to enhance the signal-to-noise ratio (SNR) of the transceiver frontend sensitivity with an ultra-low-noise receiver subsystem active device process technology [13-14].

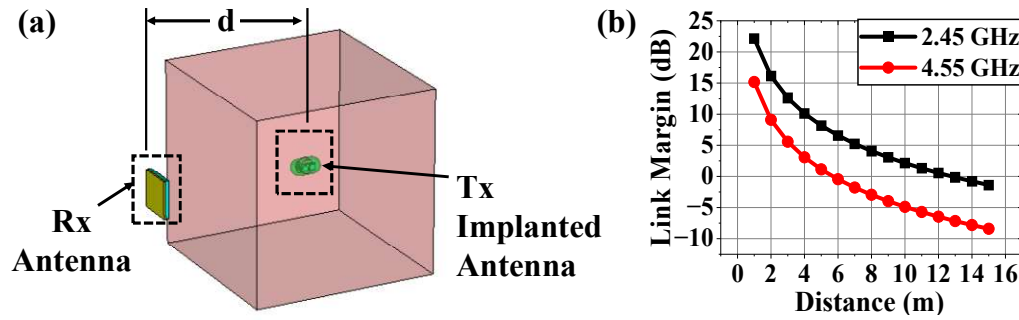


Fig. 6.11: (a) Simulation setup for link margin analysis and (b) link margin vs distance plots at 2.45 GHz and 4.55 GHz

6.6. LINK BUDGET ANALYSIS

Generally, the data sent from WCE are in form of videos and images. For link budget analysis, a wideband patch antenna placed outside human body at variable distance (d) ranging between 1 to 15 m, is taken as receiver and the proposed antenna system is considered as transmitter. The Rx antenna has realized gains of 4.95 dBi and 2.09 dBi at 2.45 GHz and 4.55 GHz respectively. The link parameters (Tx power, polarization, bit error rate, coding gain, fixation deterioration etc.) and analysis process are considered for this analysis as

described in [15]. Fig. 6.11 (b) shows the link margin versus distance plots for 2.45 GHz and 4.55 GHz of the system shown in Fig. 6.11 (a).

At 78 Mbps data rate, the WCE system can communicate up to 12 and 6 m distances at 2.45 GHz and 4.55 GHz respectively. Lateral (along x- and y- axes) and angular misalignments (0° - 270°) of the WCE system are investigated using the S_{21} (transmission coefficient) responses in Fig. 6.12 for understanding its robustness in practical deployment scenarios as performed in [16]. Ingestible antennas for WCE typically have bit error rates (BER) between 10^{-2} and 10^{-5} [17]. In this work, expected bit error rate is around 1×10^{-5} by performing the process mentioned in [18].

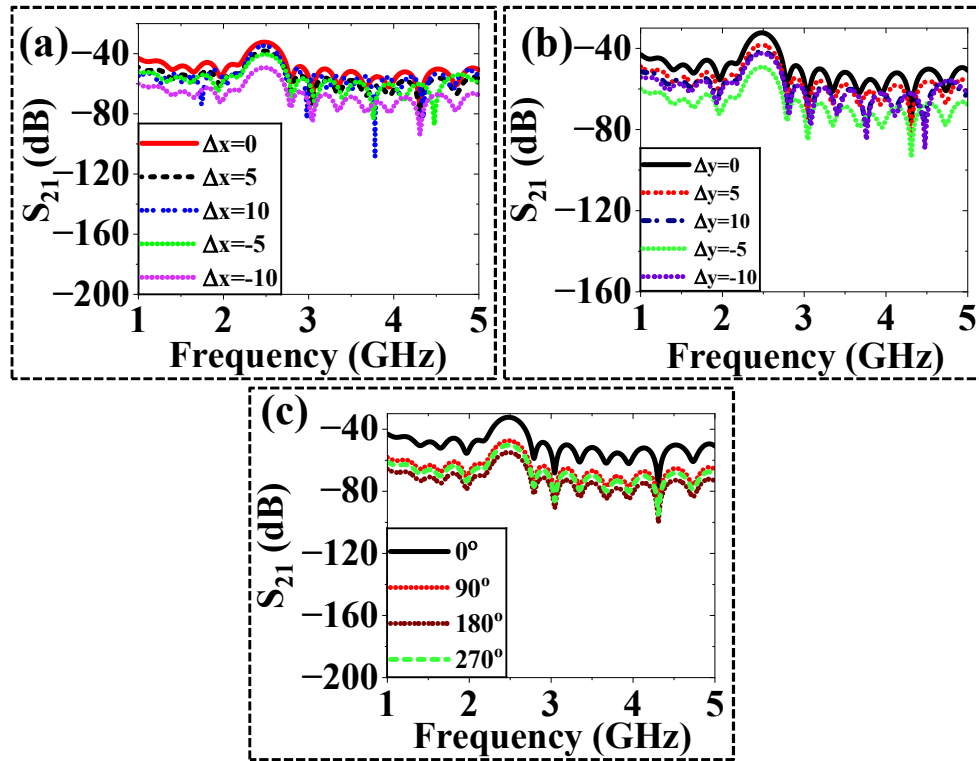


Fig. 6.12: S_{21} responses for (a) lateral misalignment in x-axis, (b) lateral misalignment in y-axis and (c) angular misalignment of WCE system

6.7. NOVELTY OF THIS WORK

The novelty of this work compared with the previous works reported in [1], [2], [4], [5], [6], [7] and [8] is highlighted in Table 6.4. The significance and timeliness of this work are summarized below.

Table 6.4: Comparison with Previously Reported WCE Antennas

Ref.	Freq (GHz)	Gain (dBi)	IBW (%)	Size (mm ³)	SAR _{1g} (1W) (W/Kg)
[1]	2.45	-26.4	20.5	49.85	712.1
[2]	0.433	-32.8	65.59	168.4	No
[4]	0.915 2.45	-28.2 -24.5	13.63 6.28	2.11	420.3 233.2
[5]	0.915 2.45	-29.8 -24.6	23.6 12.14	28.81	369.2 247.4
[6]	1.4 2.45	-29.4 -30.4	7.2 4.2	42.495	84.8 123.7
[7]	2.45	-31.9	6.53	100.48	137.9
[8]	2.45	-24.6	23.57	62.95	290.68
This Work	2.45 4.55	-24.08 -26.11	20.42 13.85	21.53	220.26 321.52

(i) Table 6.4 concludes that the proposed antenna system provided better gain than other reported antennas for WCE at 2.45 GHz. The proposed work's IBW (%) value is also better than the antenna systems reported in [4], [5], [6], [7] and [8].

(ii) From Table 6.4, compactness of the proposed antenna system is better than other reported antennas (except [4]).

(iii) SAR performance of the proposed antenna also ensures better safety than other works except [6] and [7].

(iv) This antenna has a 5G communication facility for WCE applications which has not been presented in the reported works [1-8].

6.8. CONCLUSION

In this chapter, a compact dual-band novel ingestible antenna for 5G biotelemetry in the WCE system has been presented. The proposed antenna achieves two frequency bands: 2.45 GHz ISM band with FBW of 20.42% and 4.55 GHz n79 5G NR band with FBW of 13.85%. Open-ended slots are incorporated on the radiator part of the sensor to attain compactness of the antenna (volume of 21.53 mm^3) to make it suitable for WCE. Biocompatible PVA is used to design a capsule to encapsulate the antenna system. The sensor system is placed within a homogeneous human body equivalent phantom for in-vitro testing. This antenna is communicating upto 12 m and 6 m distances at 2.45 GHz and 4.55 GHz, respectively. The reported findings promise to improve the visualization and image quality of the colonic and small bowel mucosa; enable capsule endoscopy-driven therapeutics; enhance gastric capsules development; and accommodate histological examination of the mucosa to aid monitoring and diagnosis of various conditions. the system can dynamically revert to the 2.45 GHz ISM link as a robust fallback if the 4.55 GHz 5G link fails or drops below acceptable signal-to-noise thresholds. This configuration represents a classic dual-band trade-off: the 4.55 GHz band provides ultra-high data rates over short distances, ideal for real-time, high-definition video streaming when the external receiver is placed close to the patient. Conversely, the 2.45 GHz band suffers less attenuation in lossy body tissues, expanding the communication range and maintaining connectivity during patient movement at the expense of bandwidth.

To manage this trade-off, a dual-band transceiver can switch to 2.45 GHz for critical telemetry, lower-resolution imagery, or retransmission commands, ensuring the system remains continuously operational.

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Chapter 7

Implantable Antenna for Cardiac Transcatheter Pacing System

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7.1. INTRODUCTION

With the major advancement in the design of cardiac pacemakers after its successful implantation in 1958, concerning miniaturization, multi-chamber pacing, innovative packaging capability, low power electronics, inclusion of antenna for wireless communication etc., demand of Leadless Cardiac Pacemakers (LCP) is increasing for patients suffering congestive heart failure, hypertrophic cardiomyopathy etc [1-2]. Leadless cardiac pacemakers eliminate the need for transvenous leads and a subcutaneous surgical pocket, which significantly reduces long-term hardware-related complications like lead fractures or pocket infections. Because the self-contained capsule is implanted minimally invasively via a catheter directly into the right ventricle, patients do not experience the typical upper-chest surgical scar or physical bulge associated with traditional generators. This direct intra-cardiac positioning minimizes device-

related discomfort and removes post-operative upper-extremity mobility restrictions, allowing a faster return to normal daily activities. Furthermore, clinical data shows a substantial decrease in overall complication rates—often up to 60 percent fewer issues compared to conventional systems—making them a highly reliable alternative for eligible patients. Finally, this design drastically lowers the specific risk of systemic device infection, providing a safer long-term profile for individuals with higher vulnerability to pathogens.

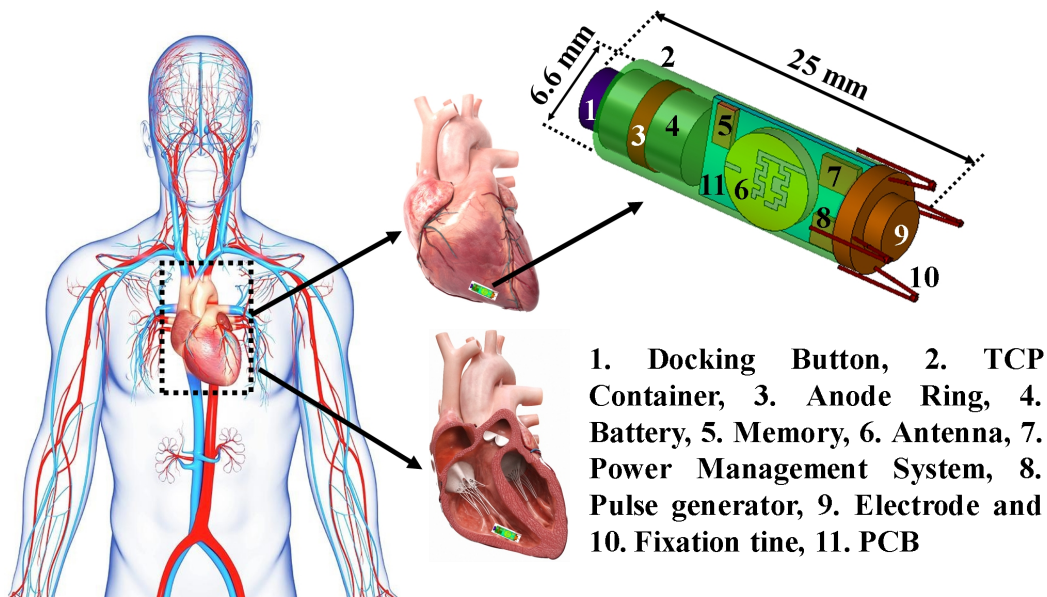


Fig. 7.1: Schematic of Micra's TCP system within human heart

Generally two types of LCP systems are implanted within patient's heart, such as Micra's Transcatheter Pacing (TCP) system and Nanostim [3]. Around 93% size reduction, reduced invasiveness, cost effectiveness, introduction of latest technologies (such as Internet of Medical Things (IoMT), Wireless Power Transfer (WPT) to increase battery life and reduce surgical problems), smart communication technologies etc., highlights the utilization of Micra's TCP system [3-4]. Several antennas have been developed in last decade for LCP

application to improve the quality of wireless patient monitoring from the outside of the body, which can also be termed as “*my care link*” in [1]. Biocompatible polyamide encapsulated conformal spiral antenna with size of $\pi \times 4.73^2 \times 20.5 \text{ mm}^3$ operating at 400 MHz Medical Implant Communication Service (MICS) band was designed by Das *et. al* [5]. The typical size of Micra’s TCP should be $\pi \times 3.35^2 \times 25.9 \text{ mm}^3$ and size of Nanostim LCP should be $\pi \times 3^2 \times 42 \text{ mm}^3$ [1]. The overall size of the antenna system designed by Das *et. al*. [5] was $20.5 \times 30 \times 0.025 \text{ mm}^3$. Therefore, the size of the system developed in [5] was quite large for LCP.

A complementary Split Ring Resonator based conformal antenna of size $\pi \times 3.29^2 \times 25.9 \text{ mm}^3$ for Nanostim LCP application at 2.45 GHz Industrial, Scientific and Medical (ISM) band was proposed by Asif *et. al* [6]. The designed antenna had realized gain of -35 dBi. A triple band (402 MHz, 1.6 GHz and 2.45 GHz) ground-slotted spiral antenna was reported by Shah *et. al* [7] which had realized gains of -30.5, -22.6 and -18.2 dBi respectively. The overall size of the system is $\pi \times 3.4^2 \times 26 \text{ mm}^3$. Zada *et. al*. [8] designed a rectangular spiral antenna LCP system with size $\pi \times 2.95^2 \times 26 \text{ mm}^3$ at 2.45 GHz operated with realized gain of -25.9 dBi.

Sharma *et. al*. [1] proposed an ultra-miniaturized (volume of 2 mm^3) antenna for implantable cardiac pacemaker operating at three frequencies 915 MHz, 2.45 GHz and 5.8 GHz with respective gain values -34.1, -29.9 and -21.8 dBi. In [9], a spiral type implantable antenna system for LCP with overall size $\pi \times 6^2 \times 35 \text{ mm}^3$ was designed to operate at 402 MHz. However, the size of the overall system is not recommended for LCP application. Recently, another LCP system with size $\pi \times 3.35^2 \times 25.9 \text{ mm}^3$ was proposed by Alghamdi *et. al* [2].

7.2. PROBLEM FORMULATION

In this work, a miniaturized dual band circularly polarized antenna is presented for Micra's TCP application. The overview of this work is shown in Fig. 7.1 [10-12]. The implantable antenna is designed inside biocompatible polyvinyl alcohol (PVA) ($\epsilon_r = 9.8$) with $\phi = 6.6$ mm and $l = 25$ mm as per the dimensions of Micra's TCP system. The TCP system comprises of fixation tines, anode ring, battery, antenna, PCB, electrode, pulse generator, sensor, power management system for WPT and docking button as shown in Fig. 7.1. The proposed system inside TCP system is placed within three-layer heart model as considered in [1].

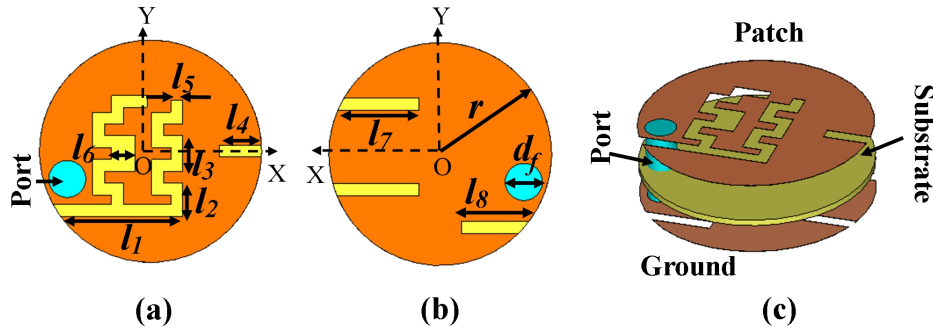


Fig. 7.2: Geometry of the proposed antenna for TCP system: (a) top view, (b) bottom view and (d) exploded view where $l_1 = 3.3$, $l_2 = 0.9$, $l_3 = 0.95$, $l_4 = 1.15$, $l_5 = 0.3$, $l_6 = 0.6$, $l_7 = 2.2$, $l_8 = 1.8$, $r = 3$ and $d_f = 0.63$ (all units are in mm).

7.3. DESIGN METHODOLOGY OF TCP SYSTEM

7.3.1. Geometry of TCP System and Simulation Setup

The top, bottom and exploded views of the proposed antenna printed on Taconic CER-10 substrate ($\epsilon_r = 10$, $\tan \delta = 0.0035$) substrate are shown in Fig. 7.2 and the design parameters are mentioned in figure caption. The proposed

antenna is placed with fixation tines, anode ring, battery, antenna, PCB, electrode, pulse generator, sensor, power management system for WPT and docking button within polyvinyl alcohol (PVA) encapsulation. The designed system is placed inside a three-layer heart model as shown in Fig. 7.3.

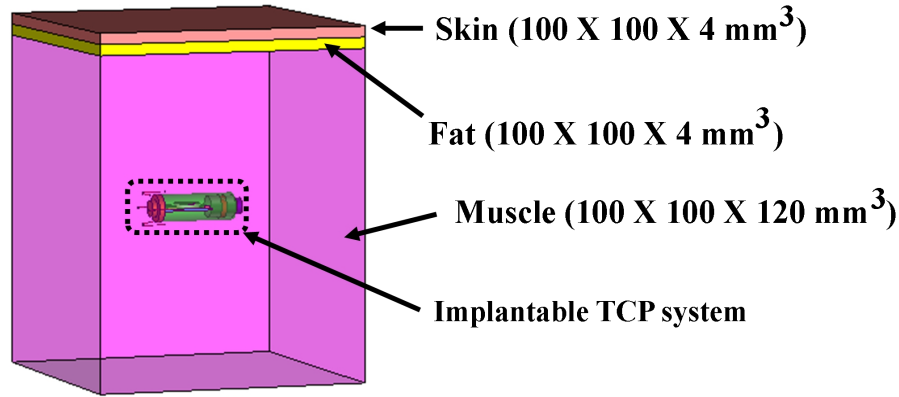


Fig. 7.3: Simulation setup of proposed TCP system

7.3.2. Design Evolution

The evolution steps of the proposed antenna is depicted in Fig. 7.4 and corresponding simulated S_{11} and Axial Ratio (AR) vs frequency plots are shown in Fig. 7.5. In the first step, a simple circular patch antenna with the diameter of 6 mm on 0.254 mm thick substrate is taken and placed within heart model after inserting within TCP container. This antenna shows poor impedance matching within frequency range of 0.2-3.5 GHz. The 3 dB AR value of this antenna is observed at frequency range of 3.3-3.56 GHz.

A meander-line open slot is incorporated in the next step; the antenna is tuned at 1.81 GHz with -10 dB IBW of 23.76%. The $AR > 3$ dB is observed at this frequency for this step. In Step 3, another meander slot is added on patch and

connected with previous slot. A dual band nature is achieved at this step in Fig. 7.5 (a). The antenna designed in Step 3 is radiating at 1.43 GHz and 2.62 GHz with IBW of 15.38% and 11.45% respectively. The 3 dB ARBW of 5.2% is obtained around 2.62 GHz due to addition of the rectangular open slot.

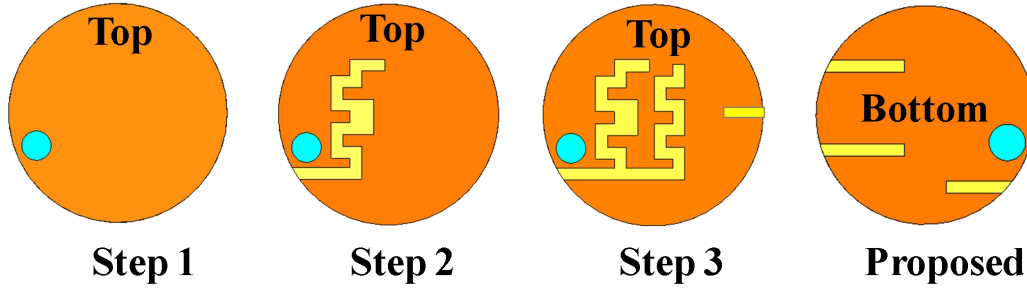


Fig. 7.4: Evolution of Antenna for cardiac TCP system

To reduce the operating frequency from 2.62 GHz to 2.45 GHz and improve CP response [13], three- rectangular open-ended slots are etched out in ground plane. Due to this, the operating frequencies are shifted to 1.3 GHz and 2.45 GHz. The S_{11} values at these frequencies are -15.35 dB and -27.38 dB respectively. The -10 dB IBWs of 25.58% (1.15-1.48 GHz) and 22.03% (2.27-2.67 GHz) and 3 dB ARBW of 21.71% (1.1-1.37 GHz) and 25.51% (2.39-3.02 GHz) are observed at 1.3 GHz and 2.45 GHz respectively. Surface current distribution over TCP antenna at 1.3 GHz and 2.45 GHz are displayed in Fig. 7.6 (a) and (b) respectively. The maximum realized gains at these frequencies are -24.6 dBi and -20.8 dBi respectively. Fig. 7.7 shows simulated E and H plane radiation patterns of final antenna at 1.3 GHz and 2.45 GHz. From Fig. 7.6, it is concluded that the main polarization of implantable TCP antenna is LHCP at both frequencies. The antenna architecture considered in Step 4 is finalized to propose for TCP system.

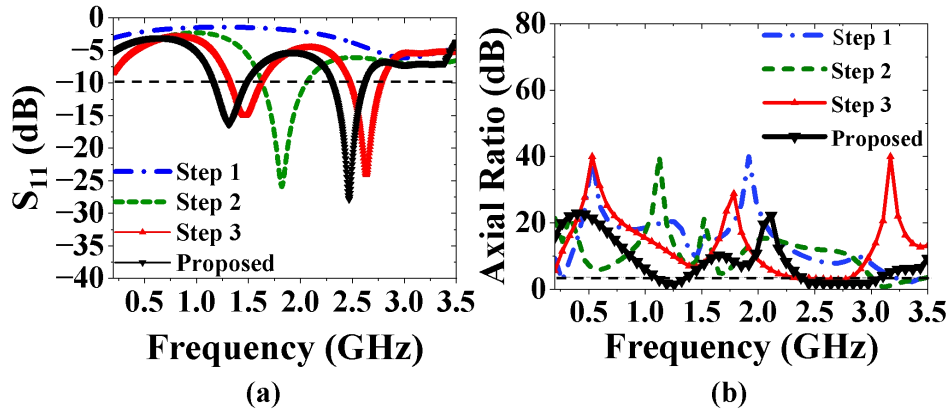


Fig. 7.5: Simulated (a) S_{11} (dB) and (b) Axial Ratio (dB) versus Frequency plots of all steps

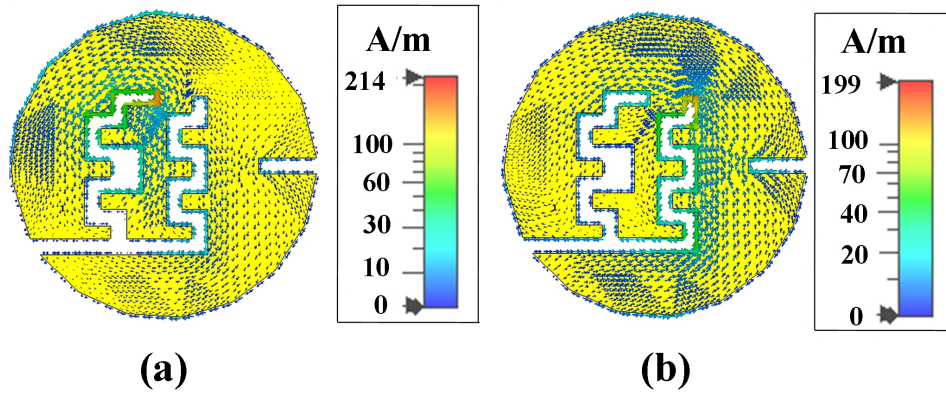


Fig. 7.6: Surface current distributions of TCP antenna at 1.3 and 2.45 GHz

7.3.3. SAR Profile of the proposed TCP System

For ensuring patient safety in presence of proposed system, SAR is evaluated as per human safety guidelines of IEEE C95.1-2019 [14]. At first, 1W RF power is supplied to the implanted antenna and SAR values for 1g and 10g tissues are noted at 1.3 GHz and 2.45 GHz and maximum allowable input powers in mW are obtained. Table 7.1 shows the SAR profile of the proposed TCP system to validate its safety before use. From this table, it is concluded that the proposed TCP system will not be a life-threatening heat source to hamper the

tissue of heart.

Table 7.1: SAR Profile of proposed TCP antenna system

Frequency (GHz)	SAR in W/kg for 1W input		Maximum safe power to operate (mW)	
	1g	10g	1g	10g
1.3	237.49	62.4	6.52	25.48
2.45	219.67	58.1	7.94	29.62

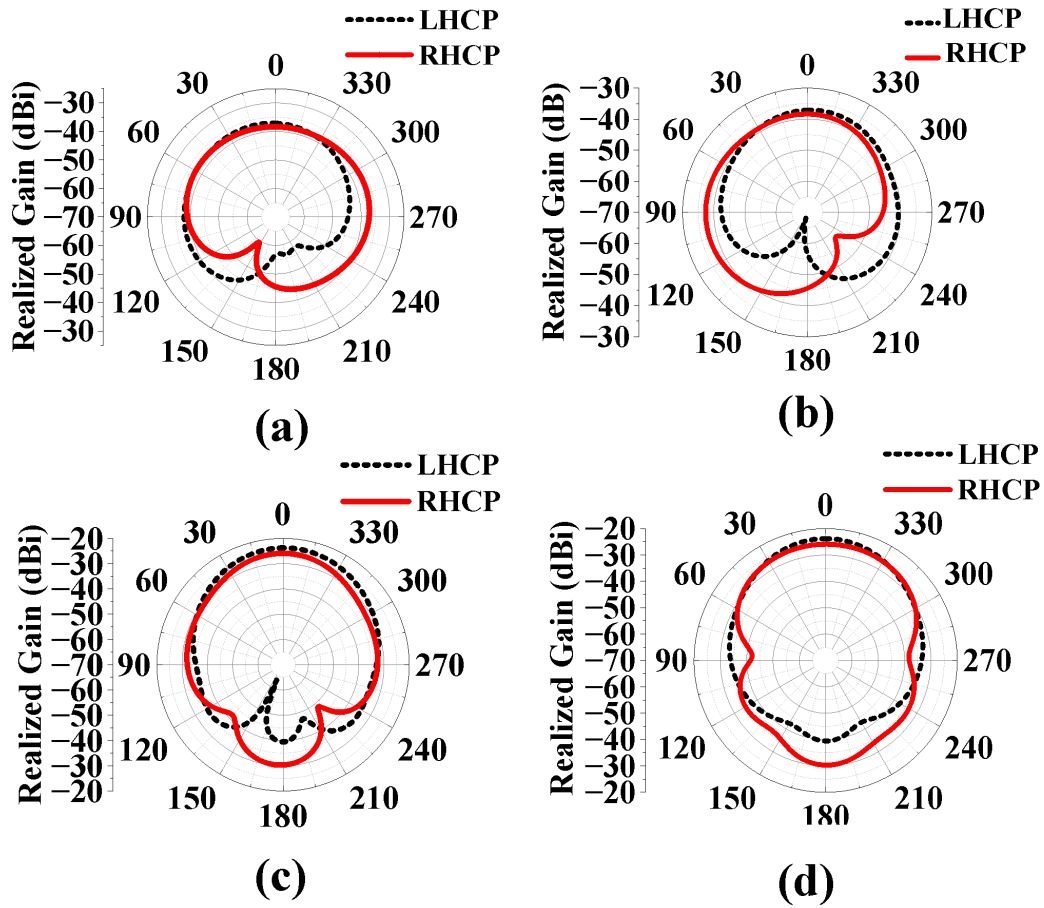


Fig. 7.7: Radiation patterns of the TCP antenna at (a) E and (b) H planes for 1.3 GHz and at (c) E and (d) H planes for 2.45 GHz

7.4. DESIGN OF EXTERNAL ANTENNA

Battery replacement is required for any cardiac pacemakers after completion of its limited lifespan which leads to encumbrance condition to the patients due to periodic invasive surgeries [15]. Wireless Power Transfer (WPT) is a way to reduce this problem. An external CP corner-chamfered patch antenna is designed (please see Fig. 7.8 (a)) on 30 mil thick Arlon AD 430 substrate ($\epsilon_r = 4.3$ and $\tan \delta = 0.003$). The external antenna is placed at different distances d (satisfying far-field condition [16]) from the skin layer of human body as shown in Fig. 7.8 (b).

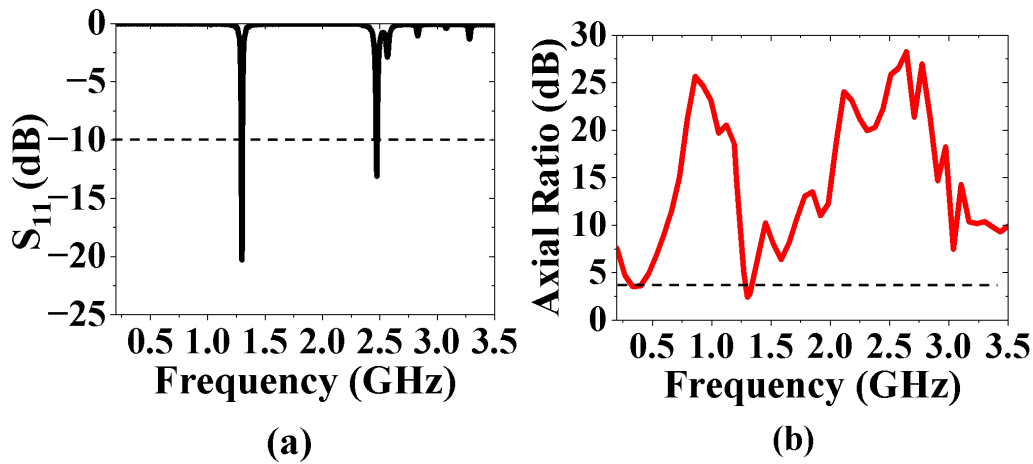


Fig. 7.8: (a) Return loss and (b) axial ratio of External Antenna

The external antenna has two operating frequencies – 1.3 GHz and 2.47 GHz with respective return losses of 20.29 dB and 13.79 dB. The axial ratios at 1.3 GHz and 2.47 GHz are 2.69 dB and 22.24 dB respectively which imply that this antenna has circular and linear polarizations at 1.3 GHz and 2.47 GHz respectively. The E and H plane radiation patterns of this antenna at 1.3 GHz and 2.47 GHz are plotted in Fig. 7.9. The realized gains of this antenna at 1.3 GHz and 2.47 GHz are 2.39 dBi and 3.58 dBi respectively. Fig. 9 validates that this antenna has LHCP at 1.3 GHz.

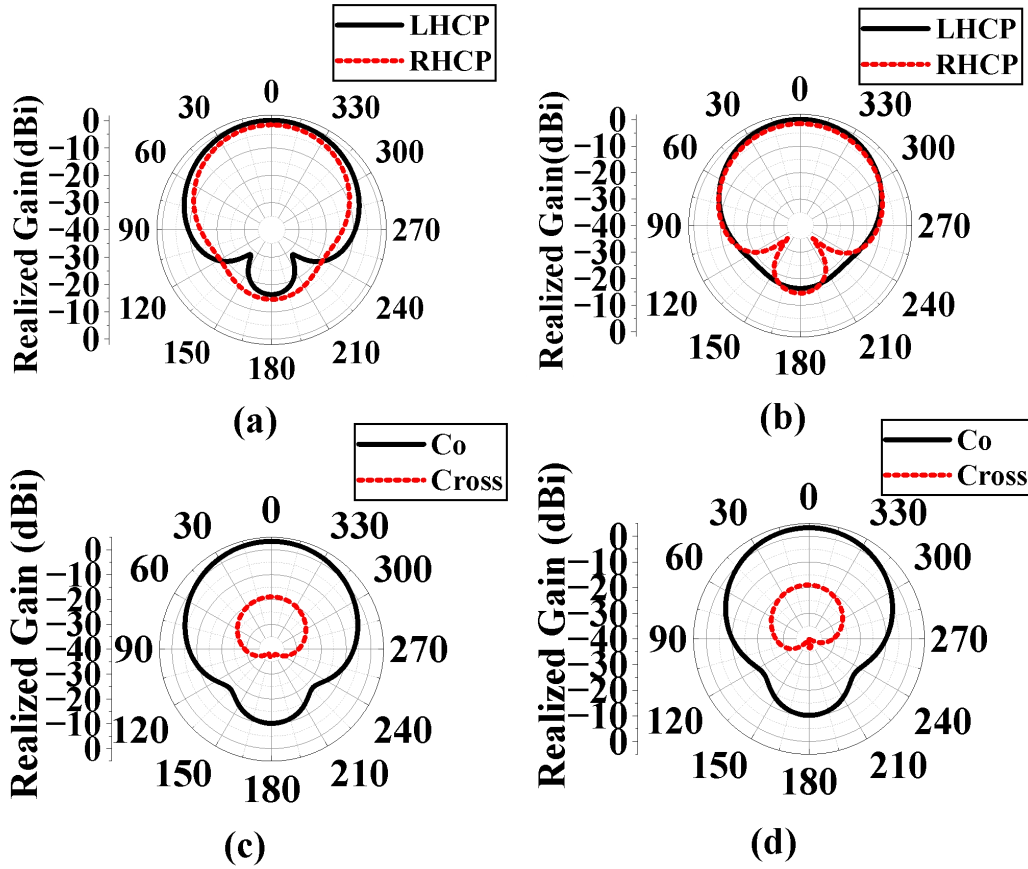


Fig. 7.9: Realized gain plots of the external antenna at (a) E and (b) H planes for 1.3 GHz and (c) E and (d) H planes for 2.47 GHz

7.5. FABRICATION AND MEASUREMENT

7.5.1. TCP System

Fig. 7.10 (a) shows the fabricated prototypes of the TCP components including implantable antenna. The TCP components are encapsulated using biocompatible polyvinyl alcohol (PVA) as illustrated in Fig. 7.10 (b). Before placing the proposed TCP system within fabricated homogeneous human body phantom

model, this antenna system is simulated within realistic Gustav voxel model as shown in Fig. 7.10 (c) and recorded the reflection and radiation performances.

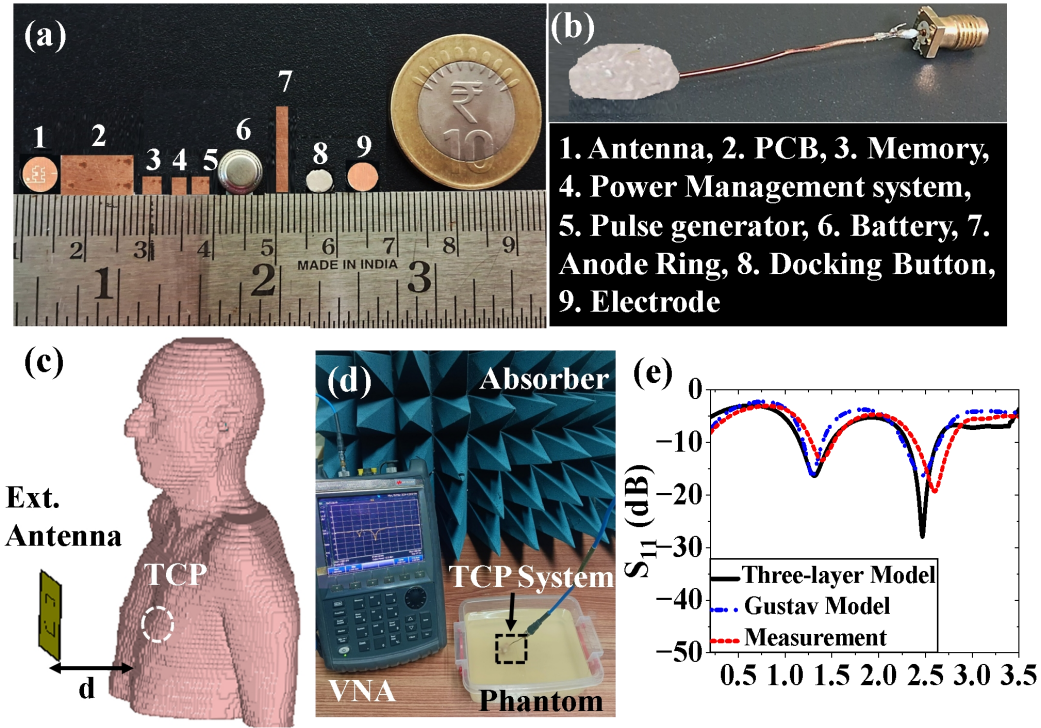


Fig. 7.10: (a) Fabricated prototypes of components for TCP system, (b) PVA encapsulated TCP system for practical verification; (c) realistic simulation of TCP system; (d) measurement setup of return loss characteristics of TCP system within homogeneous phantom; (e) comparison plot of S_{11} responses obtained from simulation and measurement

The in-vitro testing of TCP system performance is carried out within 100 X 100 X 100 mm³ homogeneous body phantom ($\epsilon_r = 52.7$, $\sigma = 1.95$ S/m) developed by mixing 63% deionized water, 27% DGME, 8% Agar agar powder and 2% sodium chloride (NaCl) in every 100 ml solution (Please refer to Fig. 7.10

(d)). The return loss responses of TCP systems obtained from simulation in three-layer heart model, voxel model and measurement are compared in Fig. 7.10 (e).

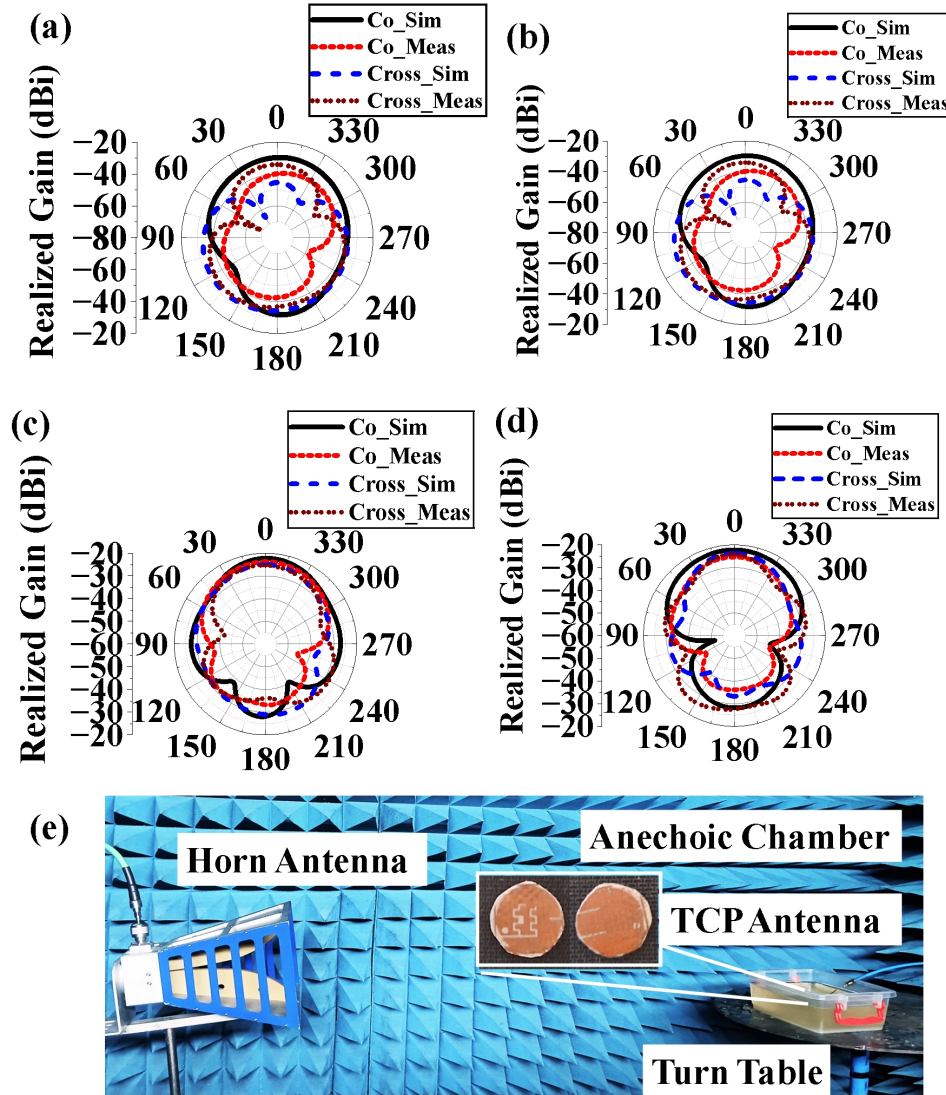


Fig. 7.11: Simulated and measured radiation patterns at 1.3 GHz for (a) E and (b) H planes and at 2.45 GHz for (c) E and (d) H planes; (e) Far-field measurement setup in anechoic chamber

The return loss values obtained by simulating TCP system within Gustav voxel model are 15.94 dB and 16.29 dB at respective operating frequencies 1.27 GHz and 2.49 GHz. The fabricated TCP system is placed within phantom. The measured return loss values of the implanted system are 13.15 dB and 18.78 dB at measured operating frequencies 1.38 GHz and 2.57 GHz respectively. The measured -10 dB IBWs at 1.27 GHz and 2.49 GHz are 17.39% (1.22-1.46 GHz) and 14.96 % (2.34-2.72 GHz) respectively. The measured 3 dB ARBW values of the TCP system are 13.38% and (1.3-1.47 GHz) and 8.84% (2.46-2.68 GHz). Fig. 7.11 (a-d) shows the simulated and measured radiation patterns in both E and H-planes within anechoic chamber as shown in Fig. 7.11(e).

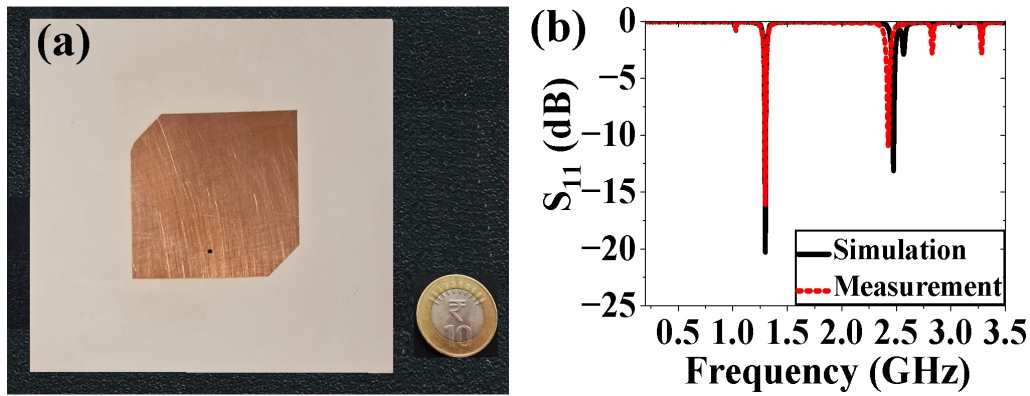


Fig. 7.12: (a) Fabricated prototype of Monitoring Antenna, (b) simulated and measured S_{11} plots of this antenna

7.5.2. External Antenna

The fabrication prototype of the external corner-chamfered dual band monitoring antenna is illustrated in Fig. 7.12 (a). The simulated and measured S_{11} responses of this antenna are plotted in Fig. 7.12 (b). The measured return losses of the fabricated antenna are 16.20 dB and 10.91 dB at measured operating

frequencies 1.29 GHz and 2.42 GHz respectively. The simulated and measured radiation patterns for E and H planes at 1.3 GHz and 2.47 GHz are plotted in Fig. 7.13. The fabricated antenna has realized gains of 2.11 dBi and 3.16 dBi at 1.29 GHz and 2.42 GHz respectively. Due to measurement in semi-anechoic chamber, losses due to cables and connectors, variation in radiation patterns for both TCP antenna and external antenna has been observed.

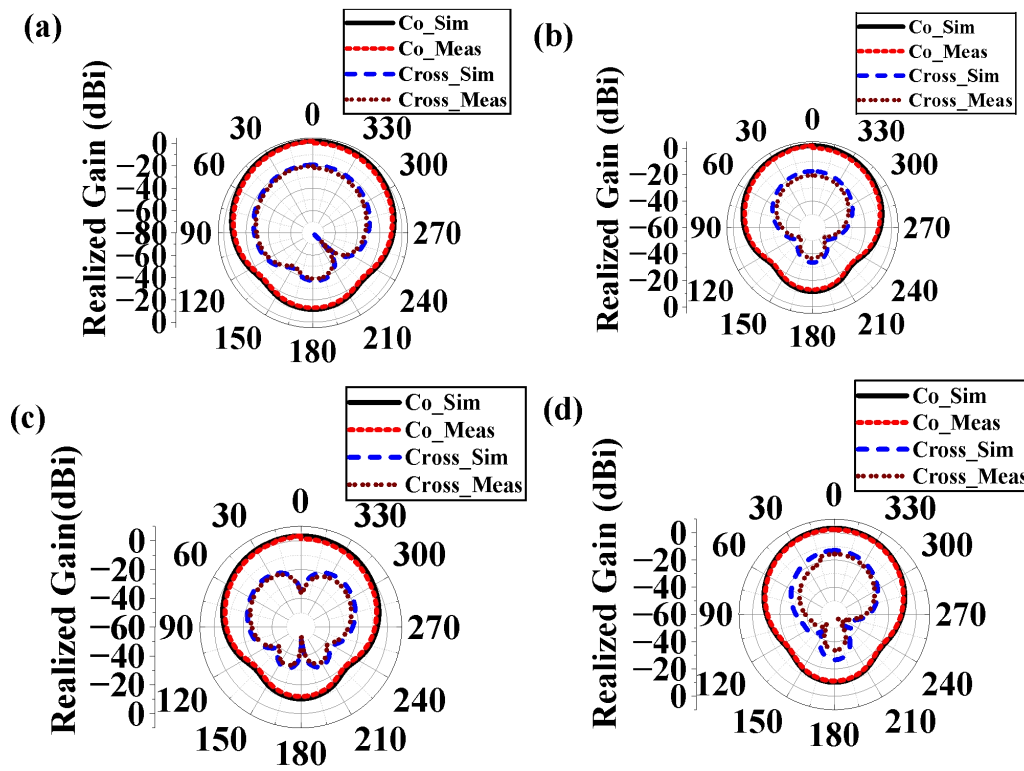


Fig. 7.13: Simulated and measured radiation patterns at 1.3 GHz for (a) E and (b) H planes and at 2.47 GHz for (c) E and (d) H planes for external antenna

7.6. WPT SYSTEM FOR FARFIELD POWERING AT 1.3 GHZ

The arrangement setup of the Wireless Power Transfer (WPT) link is illustrated in Figs. 7.8(b) and 7.10 (c) by utilizing circular polarized Tx external

antenna and Rx TCP system at 1.3 GHz. The transfer distance between these two antennas is considered as d . The power transfer efficient $\eta = |S_{21}|^2$, where S_{21} is the transmission coefficient, displays the performance of the WPT system [17]. Therefore, the S_{21} values for different transfer distances ranging between 60 to 200 mm for the simulation setups shown in Figs. 7.8 (b) and 7.10 (c) are recorded. In practical, the Tx antenna is placed at the transfer distances considered in simulation from the phantom container. The comparison between simulated and measured S_{21} responses verses distance values is plotted in Fig. 7.14 (a).

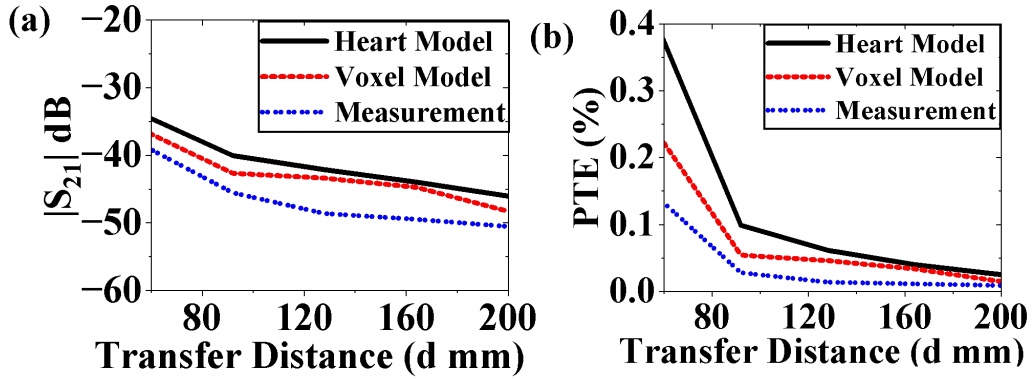


Fig. 7.14: (a) Variation in S_{21} with transfer distances and (b) PTE (%) vs transfer distances in simulation and measurement

The calculated Power Transfer Efficiency (PTE) values corresponding to the recorded S_{21} values are plotted in Fig. 7.14 (b). The S- parameter approach for WPT system performance is widely used for availability of Vector Network Analyzer (VNA). From Fig. 7.14, it is seen that the magnitude of S_{21} as well as the PTE (%) are reduced with the increase in transfer distance between Tx and Rx elements due to incorporation of path loss factor $(20 \log_{10}(\frac{4\pi d}{\lambda}))$. It is also detected from the Fig. 7.14 (d) that at 60 mm distance (d), high S_{21} value (-33.89 dB) and maximum PTE (0.38%) are observed in simulation.

In measurement, the maximum S_{21} and PTE values are obtained as -40.09 dB and 0.13% at the same distance respectively. These deviations in simulation and measurement results are seen due to misalignment and fabrication tolerances during in-vitro testing. In Table 7.2, a comparison of the efficiency of the presented WPT system with some previously reported WPT systems is performed.

Table 7.2: Comparison with Reported WPT systems

Ref.	Frequency (GHz)	PTE (%)
[18]	0.433	0.015
[19]	0.915	0.01
[20]	0.915	0.001
[21]	2.45	0.007
This Work	1.3	0.38

7.7. LINK BUDGET ANALYSIS AT 2.45 GHZ

Link budget analysis is important to characterize the wireless communication link between implanted TCP system and external monitoring device to obtain patient's heart related data and LCP operation from implanted system. In this condition, implantable TCP system is performing as transmitter and external device is receiver. The transmitting power should be <7.94 mW (8.99 dBm) at 2.45 GHz as per safety SAR performance obtained from Table 7.1. The link parameters considered in this work is tabulated in Table 7.3. Impedance mismatch losses of both antennas are calculated using below formula:

$$e_{imp} = -10 \log(1 - |\Gamma|^2) \text{ (dB)}$$

$$|\Gamma| = 10^{(S_{11} \text{ (dB)})/20}$$

The TCP antenna has CP (circular polarization) and the external antenna has LP (Linear Polarization) at 2.45 GHz. Therefore, a polarization mismatch will be found. The polarization mismatch loss can be calculated as,

$$e_p = \frac{1}{2} + \left\{ \left(\frac{r_{Tx}^2 - 1}{r_{Tx}^2 + 1} \right) \cos 2\phi \right\}$$

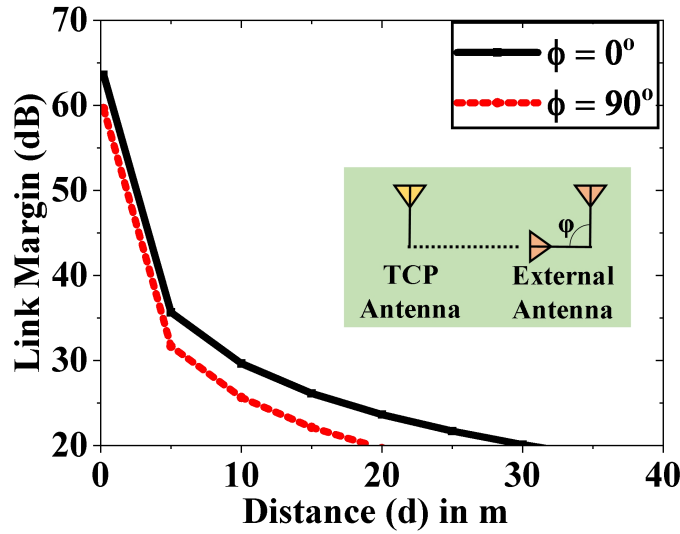


Fig. 7.15: Calculated Link Margin (dB) vs distance plot at 2.45 GHz

Here, ϕ is the angle between Tx and Rx antennas. Link margin (LM) calculation for transferring biological data up to 40 m at 1 Mbps [1], is performed using equations mentioned in [22]. Sharma *et. al* [1] mentioned that 20 dB link margin is sufficient for TCP application. The link margin (dB) verses transfer distance (d) plot at 2.45 GHz for 1 Mbps data rate due to $\phi = 0^\circ = 180^\circ$ and $90^\circ = 270^\circ$ are plotted in Fig. 7.15. From this plot, it is observed that the proposed TCP can transfer biological data effectively at 2.45 GHz up to the distances of 30 m and 19 m for $\phi = 0^\circ = 180^\circ$ and $90^\circ = 270^\circ$ respectively whether, the designed TCP system by Sharma *et. al* [1] could communicate up to 21 m distance at similar conditions. It shows the novelty of this work.

Table 7.3: Link Budget Parameters

Transmitter		
Frequency (GHz)	f_r	2.45
Transmitting Antenna Gain (dBi)	G_{Tx}	-20.8
Transmitting Power (dBm)	P_{Tx}	-16
Polarization	P	CP
Impedance Mismatch Loss (dB)	ML_{Tx}	0.0087
Linear Axial Ratio	r_{Tx}	1.58
Receiver		
Receiving Antenna Gain (dBi)	G_{Rx}	3.16
Polarization	P	LP
Temperature (K)	T	293
Boltzmann Constant	k	1.38×10^{-23}
Noise Power Density (dBm/Hz)	N_o	-203.9
Impedance Mismatch Loss (dB)	ML_{Rx}	0.36
Signal Quality		
Data rate (Mbps)	B_r	1
Ideal BPSK	E_b/N_o	9.6
Fixing Deterioration	G_d	2.5
Coding Gain	G_c	0

7.8. NOVELTY ANALYSIS

In this chapter, a miniaturized dual band circularly polarized antenna is presented for Micra's TCP application. The overview of this work is shown in Fig. 7.1. Compared with reported LCPs tabulated in Table 7.4, the proposed antenna system has comparative gain and better IBW and ARBW (%) values than reported

LCPs. Specific Absorption Rate (SAR) of the proposed system is also better than other reported LCPs which ensures the better patient safety. From this table, it is also observed that the volume of the designed antenna is smaller than other reported antennas except [8].

Table 7.4: Comparison with Previously Reported Leadless Pacemaker System

Ref.	Freq (GHz)	Size (mm ³)	IBW (%)	ARBW (%)	Gain (dBi)	SAR _{1g} (1W) (W/Kg)	WPT
[5]	0.402	30.75	3.73	No	-35.2	513.7	No
[6]	2.45	71.31	12.2	No	-35	No	No
[7]	0.402 1.6 2.45	17.15	36.82 10.69 8.93	No	-30.5 -22.6 -18.2	588 491 305	No
[8]	2.45	6	21.8	No	-25.9	270.28	No
[10]	0.915 1.4 2.45	10.66	238.39	No	-28.7 -20 -16.7	228.95 255.52 222.52	No
[9]	0.402	22.22	6.97	No	-30.3	No	No
[11]	0.61 1.65 2.45	32.3	33.33 12.5 13.3	No 4.9 5.4	-41.7 -38.4 -28.9	860 819 807	No
[2]	0.915 1.4 2.45	9.44	138	No	-32.4 -27.9 -19.8	262.8 378.1 556.9	No
[12]	0.915 1.4 2.45	8.33	152.7	No	-31.3 25.8 -21.9	320.4 332.8 464.1	No
This Work	1.3 2.45	7.18	25.58 22.03	21.71 25.51	-24.6 -20.8	237.49 219.67	Yes (PTE = 0.38%)

Wireless Power Transfer (WPT) facility with a Power Transfer Efficiency (PTE) of 0.38% is also introduced here for wireless far-field charging of the pacemaker battery to reduce the need of invasive surgery. Novelty of the proposed work is analyzed in Table 7.4. The proposed TCP system can transfer biological data up to 30 m distance at 2.45 GHz Industrial, Scientific and Medical

(ISM) band for 1 Mbps data rate which is better than the TCP system proposed by Sharma et. al [1].

7.9. CONCLUSION

In this communication, a miniaturized dual wideband circularly polarized differential open-slotted circular shaped antenna is designed for a cardiac leadless TCP system. The proposed TCP antenna has wide bandwidth and circular polarization at its both operating frequencies 1.3 GHz and 2.45 GHz respectively. Its better realized gains of -24.6 dBi and -20.8 dBi, wider IBWs of 25.58 % and 22.03 %, wider ARBW of 21.71% and 25.51 %, better SAR values of 237.49 W/kg and 219.67 W/kg for 1 g tissue samples at 1 W input power at its respective operating frequencies 1.3 GHz and 2.45 GHz than previously reported LCP antennas shows the novelty of the proposed TCP system. An external dual band antenna is designed here for Wireless Power transfer to TCP system to reduce the invasive surgical complexities at 1.3 GHz and wireless reception of biological data from TCP system at 2.45 GHz Industrial, Scientific and Medical Band. The PTE of the WPT system is 0.38 % which is better than many other reported WPT system. The designed TCP can send the heart-data up to 30 m at 2.45 GHz. Given the low measured Power Transfer Efficiency (PTE) of 0.13% at a 60 mm distance, this Wireless Power Transfer (WPT) system is strategically intended for trickle charging to supplement and extend the primary battery's life, rather than fully charging it from empty.

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Chapter 8

Conclusion and Future Scope

Contents	8.1. Overall Summary of the Thesis 8.2. Principal Contributions 8.3. Future Scope
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8.1. OVERALL SUMMARY OF THE THESIS

The thesis work aims to connect the analysis of implantable antenna systems for various healthcare applications, including wireless power transfer (WPT) for wirelessly charging implant batteries, cardiac transcatheter pacing (TCP) systems, and wireless capsule endoscopy (WCE), or also accurate tissue modeling for appropriate in-vitro testing of implantable antennas within homogeneous phantoms.

Implantable antenna performance is typically designed and analyzed using full wave approaches in human body models. These methods cost time, RAM and resources, yet they can provide us reliable results. Therefore, Chapter 3 analyzes meander-line implantable antenna performance instead of full wave techniques using a Transmission Line Model (TLM), one of the popular analytical techniques for antenna design. It also aims to determine how much memory and time reduction may be possible using TLM while taking accuracy of results into consideration.

Chapter 4 focuses on estimating the equivalent complex permittivity of various human body models, such as the brain and the heart in order to accurately

test implantable antennas in vitro using the Transmission Matrix Method (TMM) at various medical bands (402 MHz MICS band and 2.45 GHz ISM band). Within its multi-layered models, the suggested complex permittivity estimation approach can get an average accuracy of 98.46% with regard to implanted antenna performances. The amount of tissue layers, operating frequency, implanted antenna structure, etc. are all unaffected by this technique.

Our suggested analytical method is then used to evaluate the equivalent relative permittivity and conductivity of the three-layer human body model examined in Chapter 3. In practice, implantable antennas are integrated inside implanted devices rather than being directly placed into biological tissues. Consequently, Chapter 5 designs an implanted antenna system. The calculated equivalent human body model is used to install the implantable medical device, which has a corner-chamfered antenna, camera, PCB, and sensor enclosed by a 0.5 mm thick biocompatible coating composed of 10 g gelatin and 15 g deionized (DI) water. The implanted antenna system's performance metrics are noted. Simulated results for radiation (realized gain, return loss, operational frequency, and fractional bandwidth), communication (CLM and DLM), and safety concerns (SAR profiles) are recorded and experimentally validated. Chapter 5 additionally conducts the variation studies of these performance characteristics resulting from $\pm 20\%$ deviation from mean relative permittivity and conductivity.

Wireless Capsule Endoscopy (WCE), one of the most widely used implantable antenna applications in the healthcare sector, typically consists of an antenna for wireless transmission of observed health parameters, a camera, a battery, a PCB, a data management system, etc. Its usual dimensions are 26 X 11 mm². The construction of a miniature implantable antenna system for Wireless Capsule Endoscopy (WCE) applications are covered in Chapter 6. To ensure biocompatibility, the antenna, battery, PCB, camera, and data management system are all enclosed in a polyvinyl alcohol capsule. 5G technology has gained popularity in the recent period of high speed, low latency, and large capacity

connectivity. This work, which is shown in Chapter 6, successfully introduces 5G functionality into a WCE application. These findings have the potential to improve the colonic and small bowel mucosa's visualization and image quality, facilitate therapeutics driven by capsule endoscopy, improve the development of gastric capsules, and allow for histological examination of the mucosa to help monitor and diagnose a variety of conditions.

Patients with hypertrophic cardiomyopathy, congestive heart failure, and other conditions are increasingly using Leadless Cardiac Pacemakers, or LCPs, in the contemporary medical field. Two kinds of LCP systems, such as Nanostim and Micra's Transcatheter Pacing (TCP) device, are typically implanted in patients' hearts. For Micra's TCP application, a miniature dual band circularly polarized antenna is shown. According to the dimensions of Micra's TCP system, the implanted antenna is made of biocompatible Poly-Vinyl Alcohol (PVA). Fixation tines, an anode ring, battery, antenna, PCB, electrode, pulse generator, sensor, power management system for WPT and docking button are all parts of the TCP system. The proposed system is positioned inside the three-layer heart model of the TCP system. The TCP antenna operates at 1.3 GHz and 2.45 GHz, respectively, and has a broad bandwidth and circular polarization. In order to reduce the complexity of invasive surgery, an external dual band antenna is proposed here for wireless power transfer to the TCP system at 1.3 GHz and for wireless biological data receiving from the TCP system at 2.45 GHz in the Industrial, Scientific, and Medical Band.

8.2. PRINCIPAL CONTRIBUTIONS

- In Chapter 3, the meander line implantable antenna for the ISM band is analyzed using Transmission Line Modeling (TLM). To the best of our knowledge, the TLM technique is a novel approach with significant advantages that has not yet been used in the design and

analysis of an implanted antenna. By adopting TLM instead of the FEM technique, the memory need is lowered by about 85.65% and the time consumption is cut by 93.33%.

- In comparison to the current methods, the equivalent complex permittivity estimation method for various human body models at various medical frequencies suggested in Chapter 4 exhibits higher estimation accuracy.
- In place of the commonly used generalized link margin, the concepts of Carrier-Link Margin (CLM) and Data-Link Margin (DLM) are presented for accurate link characterization in the use of implantable medical devices. In comparison to other options, the SAR value for 1 g of tissue at 1W input power is likewise good for guaranteeing the patient's safety following implantation. In comparison to numerous other documented implanted antennas, the implanted Antenna System proposed in Chapter 5 exhibits superior radiation and reflection performances in terms of S_{11} and gain within the body model.
- In Chapter 6, the implanted antenna for the WCE application, which is housed in a biocompatible capsule, communicates with fractional bandwidths/gains of 13.85% / -26.11 dBi at 2.45 GHz and 4.55 GHz in a human body phantom, respectively. Compared to other documented antennas for WCE at 2.45 GHz, the suggested antenna system offers superior compactness, gain, and fractional bandwidth. The 5G NR N79 band, which includes the 4.55 GHz (4.31 GHz–4.93 GHz) resonance frequency of the planned WCE system, has the potential to boost network capacity and enhance communication speed with ultra-low latency transmission.
- A miniature dual band circularly polarized antenna for Micra's TCP application is introduced in Chapter 7. The suggested antenna system outperforms reported LCPs in terms of IBW and ARBW (%) and SAR

values, and it has a comparable gain. In order to minimize the necessity for invasive surgery, which has not yet been extensively researched for this application, a Wireless Power Transfer (WPT) facility with a Power Transfer Efficiency (PTE) of 0.38% is also presented here for wireless far-field charging of the pacemaker battery. The suggested TCP system outperforms the existing TCP system in that it can transmit biological data at a rate of 1 Mbps across a distance of 30 meters at 2.45 GHz in the Industrial, Scientific, and Medical (ISM) band.

8.3. FUTURE SCOPE

Over the course of the investigation, a number of distinct open and prospective areas have emerged. Below is a summary of these:

- The transmission line model of a meander-line implantable antenna is described in the thesis work. Further research can also be done on the use of the transmission line model for different implanted antenna architectures.
- Implantable antenna system prototypes are used here for various medicinal purposes. Nevertheless, incorrect connections are made between the parts that are put within the capsule.
- WCE and TCP are two medical application areas that are the focus of this effort in order to design new implantable antennas. Other well-known biomedical uses that will be prioritized in the near future include retinal prosthesis, tumor detection with microwave imaging,

deep brain stimulation (DBS), intracranial pressure monitoring (ICP), etc.

- Further research may be conducted on the communication performance and potential polarization mismatch of a WCE system as it passes through various GI tract bends.

Appendix I

LIST OF JOURNAL PUBLICATIONS:

- [1] **Soham Ghosh** and B. Gupta, "Accurate Equivalent Human Body Phantom Dielectric Design for In Vitro Measurement of Implantable Antennas for Biotelemetry," *IEEE Transactions on Dielectrics and Electrical Insulation*, vol. 33, no. 2, pp. 956-965, April 2026, doi: 10.1109/TDEI.2025.3632236.
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- [3] **Soham Ghosh**, Sunday C. Ekpo, Fanuel Elias, Stephen Alabi and Bhaskar Gupta, "Effects of Dielectric Properties of Human Body on Communication Link Margins and Specific Absorption Rate of Implanted Antenna System," *Sensors (MDPI)*, vol. 25, no. 11, June 2025 doi: 10.3390/s25113498.
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- [10] **Soham Ghosh** and Bhaskar Gupta, "Analysis of muscle implanted antenna performance with the variation of Implantation Depth," *Frequenz*, vol. 78, no. 9-10, April 2024, pp. 467-478. doi:10.1515/freq-2023-0421.

LIST OF CONFERENCE PUBLICATIONS:

- [1] **Soham Ghosh** and Bhaskar Gupta, "Meanderline-stub based Human Body Implantable Antenna Design for Biomedical Applications," *4th International Conference on Computer, Communication, Control & Information Technology (C3IT)*, IEEE, Hooghly, India, 2024, doi: 10.1109/C3IT60531.2024.10829409.
- [2] Sanjana Chatterjee, **Soham Ghosh** and Bhaskar Gupta, "Madelung's Disease Detection using Wireless Communication Between Implanted Antenna and External Receiving Antenna," *4th International Conference on Computer, Communication, Control & Information Technology (C3IT)*, IEEE, Hooghly, India, 2024, doi: 10.1109/C3IT60531.2024.10829454.
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(CODEC), Kolkata, India, December 2023, pp. 1-2, doi: 10.1109/CODEC60112.2023.10465705.

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[6] **Soham Ghosh**, Ardhendu Kundu and Bhaskar Gupta, "Performance Analysis of Muscle Implanted Antenna with the Variation of Effective Permittivity and Conductivity of Human Body," *2023 International Conference on Microwave, Optical, and Communication Engineering (ICMOCE)*, Bhubaneswar, India, May 2023, pp. 1-4, doi: 10.1109/ICMOCE57812.2023.10166875.

Appendix II

LIST OF PRESENTATIONS IN NATIONAL/ INTERNATIONAL CONFERENCES/WORKSHOPS:

[1] Sanjana Chatterjee, **Soham Ghosh**, Piyali Basak and Bhaskar Gupta, "Novel Vitamin D3 Delivery Technique Loaded in Starch Nanoparticle using Implantable Antenna System," 3rd International Conference on Recent Trends in Materials Science & Devices, Haryana, 24-26 March 2025. (Presented in Conference)

[2] **Soham Ghosh** and Bhaskar Gupta, "Meanderline-stub based Human Body Implantable Antenna Design for Biomedical Applications," 4th International Conference on Computer, Communication, Control & Information Technology (C3IT), IEEE, Hooghly, India, 2024, doi: 10.1109/C3IT60531.2024.10829409.

[3] Sanjana Chatterjee, **Soham Ghosh** and Bhaskar Gupta, "Madelung's Disease Detection using Wireless Communication Between Implanted Antenna and External Receiving Antenna," 4th International Conference on Computer, Communication, Control & Information Technology (C3IT), IEEE, Hooghly, India, 2024, doi: 10.1109/C3IT60531.2024.10829454.

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[5] Suman Dey, **Soham Ghosh**, Mishor Biswas and Bhaskar Gupta, "Semi-Circular Ring slotted Circular Patch Antenna for 5G Applications," 2023 8th International Conference on Computers and Devices for Communication (CODEC), Kolkata, India, December 2023, pp. 1-2, doi: 10.1109/CODEC60112.2023.10465705.

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Engineering & their Applications (IC2E3), Srinagar Garhwal, India, September 2023, pp. 1-4, doi: 10.1109/IC2E357697.2023.10262464.

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Soham Ghosh
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