



INTEGRATED MICROWAVE SENSOR AND ANTENNA FOR MATERIAL CHARACTERIZATION AND DATA TRANSFER CAPABILITIES

SYAH ALAM

DOCTOR OF PHILOSOPHY

2024



Faculty of Electronic and Computer Technology and Engineering

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**A thesis submitted
in fulfillment of the requirements for the degree of
Doctor of Philosophy**

Faculty of Electronic and Computer Technology and Engineering

UNIVERSITI TEKNIKAL MALAYSIA MELAKA

2024

DECLARATION

I declare that this thesis entitled “Integrated Microwave Sensor and Antenna for Material Characterization and Data Transfer Capabilities “ is the result of my own research except as cited in the references. The thesis has not been accepted for any degree and is not concurrently submitted in candidature of any other degree.

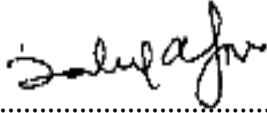
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Date :January 22 , 2024

APPROVAL

I hereby declare that I have read this thesis and in my opinion this thesis is sufficient in terms of scope and quality for the award of Doctor of Philosophy.

Signature  :.....

Supervisor Name : Prof. Dr. Zahriladha Zakaria

Date : January 22, 2024

DEDICATION

my father

A strong and gentle soul who taught me to trust in Allah and believe in hard work.

To my beloved mother

who had passed away on 19th April 2011. May Allah grants her the highest place in Jannah

Insha'Allah.

To my beloved wife

Who has been constant source of support and encouragement during the challenges of

study and life

To the rose of my life, my daughter

To my brothers, sisters, and friends

Thanks for all of your great support and continuous care

ABSTRACT

Microwave sensors have been developed to detect the characteristics of solid and liquid materials. However, there are still limitations where the characteristics of two different types of materials cannot be detected simultaneously. Furthermore, there is another constraint pertaining to the microwave sensor's functionality that only capable for detecting material characteristics, rendering it incompatible with internet-connected devices. For the future, integration microwave sensors with antennas are needed so that they are easily connected to the internet for real-time measurements. This thesis focuses on developing a microwave sensor and antenna so that it can be used independently and simultaneously. This thesis proposes several approaches to produce microwave sensor and antenna using single-port resonators, including dual U-shaped resonators, dual-T-shaped resonators and multicascoded T-shaped resonators. First, dual U-shaped resonators operating at resonant frequencies of 1.20 GHz and 2.10 GHz are proposed for permittivity detection of two different Material Under Test (MUT) independently and simultaneously. Second, a dual T-shaped resonator operating at a frequency of 1.81 GHz and 2.35 GHz is proposed as a microwave sensor and antenna sensor for contact and non-contact detection independently. Finally, a multicascoded T-shaped resonator is proposed which operates at resonant frequencies of 1.20 GHz, 2.01 GHz, and 2.45 GHz for permittivity detection of two different types of Material Under Test (MUT) independently and as antennas for data transfer capabilities. In this thesis, the Material Under Test (MUT) used is a solid material with a known permittivity ranging from 1- 6.15. The permittivity of the Material Under Test (MUT) is obtained using a polynomial equation and compared with the permittivity of the datasheet. Based on measurements and observations of all cases, an average accuracy was obtained in the range of 96% - 99% while the normalized sensitivity was in the range of 0.16% - 1.77%. This work has contribution to the body of knowledge because the proposed sensor has independent characteristics so that it can be used multifunctionally as a microwave sensor and antenna sensor for contact and non-contact detection and also as an antenna for data transfer capability. Therefore, the proposed sensors can be recommended for various applications including biomedical, pharmaceutical and food industries.

PENDERIA GELOMBANG MIKRO DAN ANTENA BERSEPADU BAGI PENCIRIAN BAHAN DAN KEUPAYAAN PEMINDAHAN DATA

ABSTRAK

Penderia gelombang mikro telah dibangunkan untuk mengesan ciri-ciri bahan pepejal dan cecair. Walau bagaimanapun, masih terdapat kekangan di mana pencirian dua jenis bahan yang berbeza tidak dapat dikesan secara serentak. Selain itu, terdapat kekangan lain dari segi kebolehtelapan penderia gelombang mikro yang hanya boleh digunakan untuk mengesan ciri-ciri bahan dan tidak sesuai diintegrasikan dengan peranti yang bersambung ke internet. Pada masa hadapan, penderia gelombang mikro pelbagai fungsi dengan antenna diperlukan supaya ianya mudah disambungkan ke internet untuk pengukuran masa nyata. Tesis ini memberi tumpuan kepada pembangunan penderia gelombang mikro dan antenna yang boleh digunakan secara bebas dan serentak. Tesis ini mencadangkan beberapa pendekatan untuk menghasilkan penderia gelombang mikro dan antenna menggunakan penyalun terminal tunggal, termasuk penyalun berbentuk dua U, penyalun berbentuk dua T dan penyalun berbentuk T pelbagai kaskod. Pertama, dua penyalun berbentuk U yang beroperasi pada frekuensi 1.20 GHz dan 2.10 GHz dicadangkan untuk pengesanan kebolehtelapan dua bahan berbeza secara bebas dan serentak. Kedua, penyalun berbentuk dua T yang beroperasi pada frekuensi 1.81 GHz dan 2.35 GHz dicadangkan sebagai penderia gelombang mikro dan penderia antenna untuk pengesanan sentuhan dan bukan sentuhan secara bebas. Akhir sekali, penyalun berbentuk T pelbagai kaskod dicadangkan yang beroperasi pada frekuensi 1.20 GHz, 2.01 GHz dan 2.45 GHz untuk pengesanan kebolehtelapan dua jenis bahan yang berbeza secara bebas dan sebagai antenna bagi keupayaan pemindahan data. Dalam tesis ini, bahan yang digunakan adalah bahan pepejal dengan kebolehtelapan yang diketahui antara 1- 6.15. Kebolehtelapan bahan diperolehi dari persamaan polinomial dan dibandingkan dengan kebolehtelapan yang didapati dari helaian data. Berdasarkan pengukuran dan pemerhatian semua kes, purata ketepatan diperolehi adalah dalam julat 96% - 99% manakala sensitiviti ternormal adalah dalam julat 0.16% - 1.77%. Kajian ini mempunyai sumbangan kepada ilmu pengetahuan kerana penderia yang dicadangkan mempunyai pencirian bebas supaya ianya boleh digunakan secara pelbagai fungsi untuk pengesanan sentuhan dan bukan sentuhan serta sebagai antenna bagi keupayaan pemindahan data. Oleh itu, penderia yang dicadangkan boleh disyorkan untuk pelbagai aplikasi termasuk industri bioperubatan, farmaseutikal dan makanan.

ACKNOWLEDGEMENT

In the name of Allah, the Most Gracious and the Most Merciful. Alhamdulillah, I praise and thank Allah SWT for His greatness and for giving me the strength and courage to complete this thesis. First and foremost, I would like to express my deepest gratitude to my supervisors Professor Dr. Zahriladha Zakaria, Dr. Noor Azwan Shairi and Professor Indra Surjati (Universitas Trisakti) for their guidance and support throughout my candidature. My exceptionally supportive principal supervisor Professor Dr. Zahriladha Zakaria helped me a lot more than I really deserved during my stay at Universiti Teknikal Malaysia Melaka (UTeM). He advised me with his deep knowledge in the field of electromagnetics, microwave engineering and circuit design. He always welcomed scientific discussion and has given me critical feedback to improve the quality of my research. I would like to gratefully acknowledge his enthusiastic supervision, encouraging attitude, and generously sharing his knowledge. I would like to thank my cosupervisor, Professor Indra Surjati and Dr. Noor Azwan Shairi, who has been of great help, support, and advice.

Special thanks to Universiti Teknikal Malaysia Melaka (UTeM) and Faculty of Industrial Technology (FTI), Universitas Trisakti for the supporting during my PhD study. Besides, I am also grateful to all collaborators who helped me in completing the research project in this thesis, Dr. Eng Teguh Firmansyah (Sultan Ageng Tirtayasa University) and Professor Mudrik Alaydrus (Mercu Buana University).

I would like to present my sincere thankfulness to my beloved father, brother and sister and my deceased mother, who died in April 2011, for their unconditional love, continuous support, encouragement, and prayers during my life and throughout the course of this study. Many thanks to my brothers and sisters for their support and encouragement. Last but not

least, I would like to express my deepest gratitude to my wife for her patience and tolerance over the last two years. I could not be able to finish this work without your support. Thank you for being with me and for your appreciated sacrifices. Thank you, my two dashing boys, Habibie and Hanandra for being in my life. Finally, special thanks to all my colleagues and friends and thanks to everyone who have been directly and indirectly involved in the successful completion of this thesis.

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LIST OF ABBREVIATIONS

<i>AMC</i>	-	Artificial Magnetic Conductor
<i>AS</i>	-	Antenna sensor
<i>AWR</i>	-	Advanced Wireless Revolution
<i>BPF</i>	-	Band Pass Filter
<i>CSRR</i>	-	Complementary Split Ring Resonator
<i>DUT</i>	-	Device Under Test
<i>GPS</i>	-	Global Positioning System
<i>HFSS</i>	-	High Frequency Structure Simulator
<i>HPF</i>	-	High Pass Filter
<i>IDC</i>	-	Interdigital Capacitor
<i>LPF</i>	-	Low Pass Filter
<i>MUT</i>	-	Material Under Test
<i>MS</i>	-	Microwave Sensor
<i>NS</i>	-	Normalized Sensitivity
<i>RF</i>	-	Radio Frequency
<i>RF ID</i>	-	Radio Frequency Identification
<i>S</i>	-	Sensitivity
<i>SRR</i>	-	Split Ring Resonator
<i>TE</i>	-	Transverse Electric
<i>TEM</i>	-	Transverse Electric Magnetic
<i>TM</i>	-	Transverse Magnetic
<i>VNA</i>	-	Vector Network Analyzer
<i>Wi – Fi</i>	-	Wireless Fidelity

LIST OF SYMBOLS

C	-	Capacitance
L	-	Inductance
R	-	Resistance
D	-	Electric Flux
E	-	Electric Field
H	-	Magnetic Field
ϵ_r	-	Dielectric permittivity
ϵ_0	-	Permittivity of Free Space
μ_r	-	Dielectric permeability
μ_0	-	Permeability of Free Space
ϵ'	-	Real Part of Permittivity
ϵ''	-	Imaginary Part of Permittivity
β	-	Propagation Constant
V_p	-	Phase Velocity
λ_g	-	Wavelength
Z_0	-	Characteristic Impedance
n	-	Number of Modes
f_0	-	Resonant Frequency
ω	-	Angular Frequency
h	-	Thickness of substrate
$\tan \delta$	-	Loss Tangent

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LIST OF PUBLICATIONS

The followings are the list of publications related to the work on this thesis:

Journal Publication:

Alam, S., Zakaria, Z., Surjati, I., Shairi, N. A., Alaydrus, M., and Firmansyah, T., 2022. Dual-Band Independent Permittivity Sensor Using Single-Port with a Pair of U-Shaped Structures for Solid Material Detection. *IEEE Sensors Journal*, 22(16), pp. 16111-16119. (Web of Science and Scopus)

Alam, S., Zakaria, Z., Surjati, I., Shairi, N. A., Alaydrus, M., and Firmansyah, T., 2023. Integrated Microwave Sensor and Antenna Sensor Based on Dual T-Shaped Resonator Structures for Contact and Noncontact Characterization of Solid Material. *IEEE Sensors Journal*, 23(12), pp. 13010-13018. (Web of Science and Scopus)

Alam, S., Zakaria, Z., Surjati, I., Shairi, N.A., Alaydrus, M. and Firmansyah, T., 2023. Multifunctional of dual-band permittivity sensors with antenna using multicascode T-shaped resonators for simultaneous measurement of solid materials and data transfer capabilities. *Measurement*, 217, p.113078. (Web of Science and Scopus)

Conference Publication:

Alam, S., Zakaria, Z., Surjati, I., Shairi, N. A., Alaydrus, M., and Firmansyah, T., 2023. Highly Independent Dual-Band Permittivity Sensors for Simultaneous Measurement of Solid Materials. *The 33rd International Conference Radioelektronika (RADIOELEKTRONIKA)*. Pardubice, Czech Republic, 19 – 20 April 2023. IEEE.

Alam, S., Zakaria, Z., Surjati, I., Shairi, N. A., Alaydrus, M., and Firmansyah, T., 2023. Independent Permittivity Sensor Based on Dual T-Shaped Resonator for Solid Material Characterization. *The 2023 IEEE International Symposium on Antennas and Propagation (ISAP2023)*. Kuala Lumpur, Malaysia, 30 October – 02 November 2023. IEEE.

CHAPTER 1

INTRODUCTION

This chapter contains the background of microwave sensors with their characteristics and permittivity of solid materials. Furthermore, it also explains the objective and motivation of the proposed research to integrate the microwave sensor and antenna for data transfer capability.

1.1 Background

Material characterization techniques have been developed in recent years. Various techniques are available for real part measurement of the complex permittivity (dielectric constant). Each technique is useful for certain materials application at certain range of frequencies. The microwave methods of measurements for materials characterization as shown in Figure 1.1 are divided into two types, non-resonant and resonant. The non-resonant methods, over the frequency range can be used to determine the electromagnetic properties, while the resonant methods give more accurate knowledge of dielectric properties over a certain frequency range or a single frequency.

The motivation of this research is driven by the resonant methods with highly accurate, which is better than non-resonant method. Resonant perturbation technique is a well-known and widely used method for material characterization application using microwave resonators (Rammah Ali Alahnomi et al., 2017; Amyrul Azuan Mohd Bahar et al., 2019; Noel et al., 2022). It is mainly used for medium and low loss material characterization using cavity resonators. However, the resonators are normally restricted to

relatively low frequencies since they exhibit high conductor losses and their size become impractically small in the higher frequency ranges.

The resonant perturbation method extracts the permittivity of the material under test (MUT) since the boundary conditions at the cavity walls change before or after material inclusion which is conversely what has been claim by Alahnomi et al. (2016).

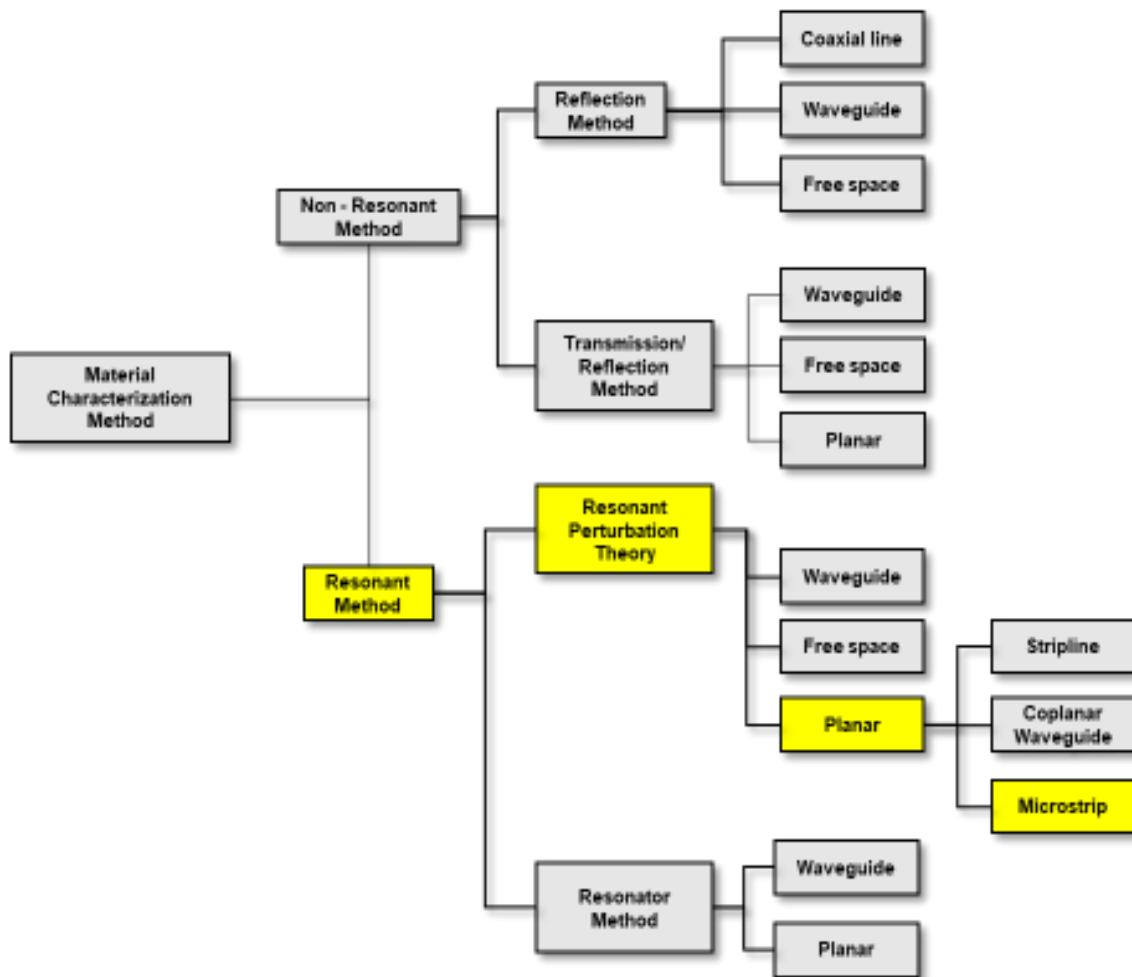


Figure 1.1 Classification of the material characterization methods

A microwave planar sensor serves as a dedicated tool within microwave and radio frequency (RF) engineering, designed to perceive and identify alterations in a range of

physical attributes. These attributes encompass characteristics like dielectric properties, temperature, moisture, and chemical composition. By harnessing microwave technology principles, the sensor engages with its surroundings, generating quantifiable reactions that facilitate the scrutiny and interpretation of various materials and situations. The utility of microwave planar sensors spans across multiple domains as shown in Figure 1.2

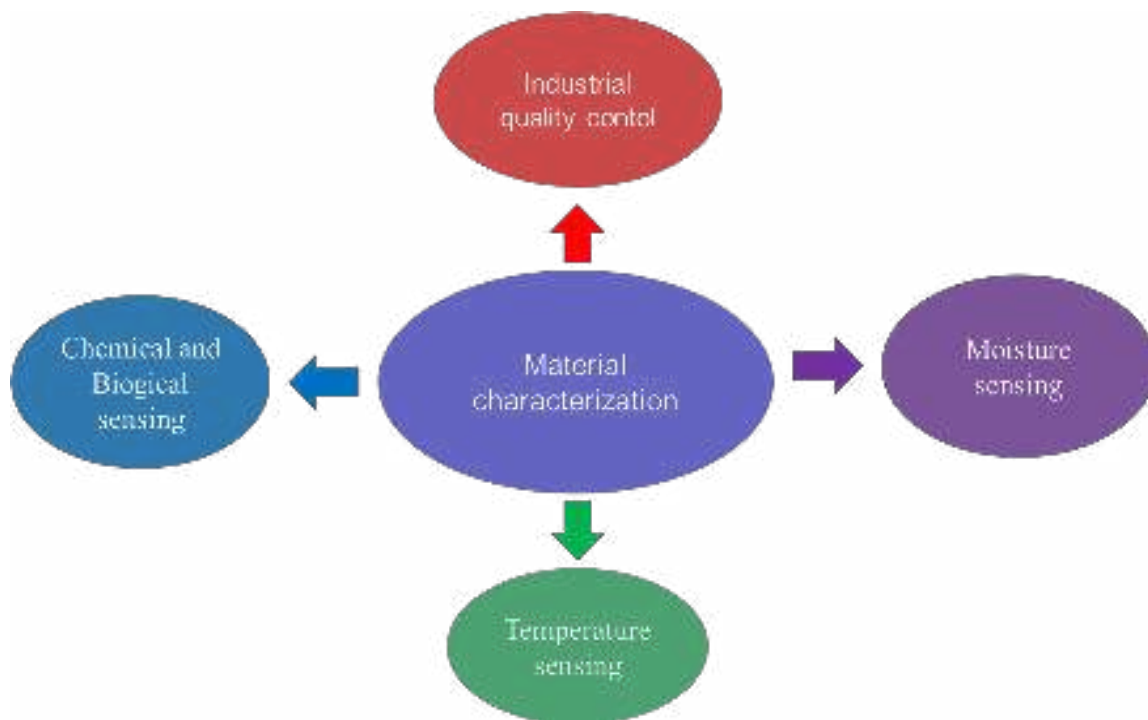


Figure 1.2 Application of material characterization

The significance of exploring material characterization lies in its several applications such as agriculture, where activities like grain drying, enhancing seed germination through seed treatment, and managing insect presence in stored grains utilize high-frequency and microwave electric fields. Moreover, material characterization plays a crucial role in the realm of food processing, ensuring food quality and safety. The dairy industry benefits from

it as well, facilitating a rough assessment of milk constituents such as ionic compounds, fats, carbohydrates, and proteins.

In the context of fruits and vegetables, rapid non-destructive quality assessment becomes imperative to gauge freshness. Material characterization extends its utility to Geoscience, aiding in soil analysis to optimize planting conditions. The domain of bio-sensing benefits from this study, offering insights into tissue and cell properties, thus enhancing understanding of biological processes at the molecular level. Additionally, material characterization contributes to ongoing advancements in radio frequency and microwave hyperthermia, a technique employed in cancer treatment.

1.2 Problem Statement

The development of communication systems lately has been very rapid to support industry 4.0. All devices that support a system can be integrated using the internet of things (Rajab and Cinkelr, 2018). One of the challenges is that a device can be integrated with other devices that can be connected wirelessly (Jabbar et al., 2019). Sensors are devices that are used to detect changes in the characteristics of certain materials in specific environmental conditions (Alahnomi et al., 2021).

One of the material characteristics that can be observed is permittivity. Material characterization is needed to detect the characteristics of materials that can be used for several applications such as the biomedical industry (Omer et al., 2021)(Mohammadi et al., 2019), the food industry (Amar et al., 2020; Solyom et al., 2020; Jahangiri et al., 2022), and the pharmaceutical industry (Kapranov and Kouzaev, 2019)(Amyrul Azuan Mohd Bahar et al., 2019).

Generally, the characterization of solid materials is carried out by observing the permittivity at a certain frequency (Rammah A. Alahnomi et al., 2017)(Muñoz-Enano et al., 2020). Permittivity relates to the ability of a material to store an electric field under certain conditions (Boybay and Ramahi, 2012). Several approaches have been proposed to determine the permittivity of solid materials, for example using planar sensors (Amyrul A. Mohd Bahar et al., 2019)(Al-Behadili et al., 2020), waveguides (Ndoye et al., 2017a)(Lee et al., 2019) and combinations of active electronic components (Abdolrazzaghi et al., 2021)(Abdolrazzaghi and Daneshmand, 2017).

Planar sensors have several advantages including compact design (Karimi, Arsalan and Shamim, 2019), low cost, and high accuracy (Abduljabar, Hamzah and Porch, 2021). In the future, multifunctional sensors are needed to be integrated with the internet-of-things (IoT) and wireless-based devices such as Bluetooth, Z-Wave, Wi-Fi, Zigbee and etc. However, the sensor has a limited function that is to merely detect the permittivity of the material under test (MUT). Therefore, a multifunctional and independent permittivity sensor is needed so that it can be integrated with data transmission capability. Another constraint, there is still a limited number of sensing hotspots, and unable to support for simultaneous measurement. The advantage of simultaneous characterization is that there can be more than one detected material so that it can reduce the time needed to detect a large number of materials. Moreover, material characterization using microwave sensor is carried out by placing the material under test (MUT) in direct contact so that it can reduce the performance of the sensor. This caused the surface of the sensor to potentially be damaged due to oxidation and corrosion. To overcome this limitation, non-contact detection using antenna sensor is proposed as a solution to make the sensor more durable, accurate, and sterile.

This research focuses on producing a permittivity sensor which also functions as an antenna and has independent characteristics so that it can be used for non-contact detection and data transfer capabilities. The proposed sensor uses a planar resonator with one port so that it can also function as an antenna simultaneously. Furthermore, the permittivity is determined by placing the MUT on the surface of the sensing area of the sensor using perturbation theory. The proposed sensor has a different sensing area depending on the resonant frequency and the concentration of the E-field so that the MUT can be placed simultaneously. Finally, microwave sensors for simultaneous measurement and characterization of solid materials for contact and non-contact detection and data transfer capabilities can be proposed for application in the food, biomedical, and pharmaceutical industries to support industry 4.0, which requires the integration between sensors and the internet.

1.3 Research Question

The research questions of this research are as follows:

1. How to design a multifunctional microwave sensor to detect the permittivity of the MUT and as an antenna to transmit electromagnetic waves that can work simultaneously?
2. How the performance and correlation of each resonant frequency when MUTs with different permittivity are placed simultaneously on the proposed microwave sensor?
3. How is the performance of the proposed microwave sensor and antenna when it works simultaneously?

1.4 Research Objective

The overall aim of this research is to propose a microwave sensor that can be used to determine the permittivity of solid materials. The main objectives of the proposed research are:

1. To design and realize microwave sensors which have independent characteristics and multifunctional capabilities.
2. To validate the proposed sensors in the laboratory to determine the permittivity of different solid materials using mathematical models.

1.5 Scope of Research

The scope of the research has the following limitations:

- This research focuses on the design of microwave sensors to detect the permittivity of solid materials. Furthermore, the reference permittivity used refers to the datasheet (RO5880, RO4003C, FR-4, RO3010). Therefore, the permittivity that can be determined is limited to the permittivity range of 1 - 6.15 according to the materials available in the laboratory. In other words, the characterization of other materials is not the focus of this study.
- This research focuses on the structure of the resonator with one input port as a microwave sensor. Therefore, the parameters observed to determine permittivity are the resonant frequency and reflection coefficient (S_{11}). In other words, other parameters were not observed in this study.

- The simulation and design process of the microwave sensor was carried out using HFSS 15.0 while the simulation of the equivalent circuit model used AWR 2009.
- The microwave sensor is realized using an FR4 substrate with a permittivity of 4.3, a thickness of 1.6 mm and a loss of tan of 0.0265. Furthermore, the input connector used is RP-SMA with an impedance of 50 Ohms.
- The proposed microwave sensor is planned to operate at frequencies around 1.20 GHz until 2.45 GHz. The location of the sensing area is determined based on the concentration of the electric field from the resonator at each resonant frequency. Furthermore, the dimensions of the MUT used are 10 mm x 10 mm with a thickness of 1.58 - 1.60 mm.
- Validation and measurements were carried out in the laboratory using a Vector Network Analyzer with an ambient temperature of 25°C - 27°C. Furthermore, the permittivity of the MUT was determined using a numerical model based on a second order polynomial.

1.6 Contribution of thesis

This thesis makes a significant contribution in the microwave sensor area. These contributions can be classified into three categories of developments from microwave sensors. In the first part, a microwave sensor with a dual U-shaped single port resonator is proposed for the detection of two different types of MUT independently. The second part, integration of microwave sensor and antenna sensor is proposed for independent contact and non-contact detection based on dual T-shaped resonators. Finally, a multifunctional

microwave sensor with antenna is proposed for two different types of MUT detection and data transfer capability based on a multicascode T-shaped resonator.

1.6.1 Dual-Band Independent Permittivity Sensor using Single-Port with a Pair of U-Shaped Structures for Solid Material Detection

- This research proposes a single-port dual-band independent microwave sensor for detecting solid materials. The sensor structure consists of a pair U-shaped resonator connected by a power divider and a microstrip feed line with an impedance of 50 Ohms with two different sensing areas. The 1st and 2nd resonators operate at a frequency of 1.20 GHz and 2.10 GHz, respectively. The type of material under test (MUT) is a standard material with a size of 10 mm x 10 mm (RO5880, RO4003C, and FR4). The proposed sensor had a very high accuracy compared with both of using permittivity from datasheet and calibrated frequency from simulation for the 1st and 2nd resonator, respectively. Moreover, the pair of the U-shaped resonators had independent performances and they did not affect each other when the MUT was loaded. Furthermore, it can also operate with single-port measurement, which is a more efficient resource. This novel design has been published in ***IEEE Sensors Journal Vol.22 No.6, August 2022*** under the title of “Dual-Band Independent Permittivity Sensor Using Single-Port with a Pair of U-Shaped Structures for Solid Material Detection”.
- This research has contributed to producing a microwave sensor capable of detecting the permittivity of two different types of MUT simultaneously with high independence. This novel design has been presented in ***2023 33rd International Conference Radioelektronika (RADIOELEKTRONIKA)*** under the title of “Highly

Independent Dual-Band Permittivity Sensors for Simultaneous Measurement of Solid Materials”.

1.6.2 Integrated Microwave Sensor and Antenna Sensor Based on Dual T-Shaped Resonator Structures for Contact and Non-Contact Characterization of Solid Material

- This research proposes the integration of microwave sensors (MS) and antenna sensors (AS) for contact and non-contact characterization of solid materials. Integration between MS and AS were determined based on the reflection coefficient (S_{11}) where S_{11} of ≥ -10 dB was proposed as MS and S_{11} of ≤ -10 dB was proposed as AS. The proposed sensor consists of a dual T-Shaped resonator with single port operating at a resonant frequency of $f_{r1} = 1.81$ GHz for MS and $f_{r2} = 2.34$ GHz for AS using FR-4 substrate with ϵ_r of 4.3, $\tan \delta$ of 0.0265 and thickness of 1.6 mm. Furthermore, the two resonant frequencies had independent characteristics with different sensing areas. MS are proposed for permittivity detection of the Material Under Test (MUT) for contact detection while non-contact detection is proposed by utilizing AS. The procedure for contact detection was carried out by placing the material under test on the surface of the sensor while non-contact detection was carried out by placing the MUT with an air gap between the MUT (d) and the AS was 0.5 - 1.5 mm. The measurement results showed that the average accuracy was 99.30% and 99.31% for the MS and AS for the permittivity range of 1 - 6.15 while the normalized sensitivities of the proposed sensors were 1.15 % and 0.16%, respectively. This novel design has been published in ***IEEE Sensors Journal Vol.23 No 16, July 2023*** under the title of “Integrated Microwave Sensor and Antenna

Sensor Based on Dual T-Shaped Resonator Structures for Contact and Non-Contact Characterization of Solid Material”.

- This research proposes an independent permittivity sensor for solid material characterization. The proposed sensor consists of a dual T-shaped resonator with a single port operating at $f_{r1} = 2.32$ GHz as 1st resonator and $f_{r2} = 1.82$ GHz as 2nd resonator for direct detection which has independent performance with different sensing area so it can be used independently. The permittivity of the MUT is detected directly by placing the Material Under Test (MUT) above the surface of the 1st and the 2nd resonator. Based on measurement result, the proposed sensor can detect four different MUT independently with different sensing area. This research is very useful for the characterization of solid materials independently for several applications such as biomedical, quality control and pharmaceutical industries. This novel design has been presented in **2023 IEEE International Symposium on Antennas And Propagation (ISAP2023)** under the title of “Independent Permittivity Sensor Based on Dual T-Shaped Resonator for Solid Material Characterization”.

1.6.3 Multifunctional of Dual-Band Permittivity Sensors with Antenna using Multicascode T-shaped Resonators for Simultaneous Measurement of Solid Materials and Data Transfer Capabilities

- This paper proposes multifunctional of a dual-band independent permittivity sensor with a WLAN antenna for the simultaneous characterization of solid materials using multicascode T-shaped resonators. The proposed sensor consists of multicascode T-shaped resonators with 50 Ω microstrip lines using a single port operating at three different resonant frequencies namely 2.45 GHz, 2.00 GHz and 1.50 GHz,

respectively. The 1st resonant frequency was used as a WLAN antenna while the 2nd and 3rd resonant frequencies were used as sensors to determine the permittivity of solid materials simultaneously. The proposed sensor had two different sensing areas according to the resonant frequency and concentration of the electric field. Therefore, it can be used to determine permittivity simultaneously. All resonant frequencies of the proposed sensor are independent towards each other so that they did not affect when the material under test (MUT) is loaded simultaneously. Based on the measurement results obtained an average accuracy of 96.81% and 97.54%, while the normalized sensitivity (NS) of the proposed sensor was 0.66% and 1.77% with a permittivity range of 1 – 6.15 from the 2nd and 3rd of resonant frequency, respectively. Furthermore, the higher resonant frequency can be used as the WLAN antenna which produces a directional radiation pattern at a resonant frequency of 2.45 GHz. These results indicate that the sensor and antenna could work simultaneously and did not affect each other significantly. This novel design has been published in ***Measurement Journal Vol.217, August 2023*** under the title of “Multifunctional of Dual-Band Permittivity Sensors with Antenna using Multicascode T-shaped Resonators for Simultaneous Measurement of Solid Materials and Data Transfer Capabilities”.

Therefore, to the author’s knowledge the contribution of the investigations that have been carried out in this thesis is presented for the first time as a solution for integrating sensors and antennas in one device used for material characterization and data transfer capabilities.

1.7 Thesis Outline

This thesis is organized into seven chapters consisting of background, literature review, methodology and three chapters which present the main contributions of independent microwave sensors for solid material characterization, integration of microwave sensors and sensor antennas for contact and non-contact detection and microwave sensors. multifunctional with antenna for data transfer capability. The final chapter presents the conclusions and recommendations for future work of this research. The detailed description for each part of thesis is presented as follows:

- This chapter provides the background of microwave sensors with their characteristics and permittivity of solid materials, on, the importance of material characterization, problem statement, scope and contributions of the work, and the important issues which led to do this research. Furthermore, it also explains the objective and motivation of the proposed research to integrate the microwave sensor and antenna for data transfer capability.
- Chapter 2 provides the literature review that support this research such as: introduction to material characterization, the fundamental concepts of electromagnetic characterization, the measurement techniques for material characterization along with their advantages and disadvantages. The final discussion of this chapter is shown by presenting previous work related to material characterization using microwave sensors, especially for the simultaneous characterization of solid materials both in contact and non-contact as well as multifunctional microwave sensors that can be integrated with other devices.

- Chapter 3 provides the development of proposed microwave sensor based on a single port planar resonator such as introduction and a flowchart of the research project that describes the stages, methods, and design of the proposed microwave sensor. In this research project, the proposed microwave sensor is based on a single-port resonator consisting of three different structures: dual U-shaped resonator, dual T-shaped resonator and multicascoded T-shaped resonator. The discussion in this chapter is continued by determining the design specifications of the proposed sensor, basic principle of resonator, simulation tools, material preparation, accuracy and sensitivity of proposed sensor. The final discussion in this chapter is a measurement setup to validate the performance of the proposed sensor which is represented with accuracy and sensitivity and compared with the existing work.
- Chapter 4 provides the development of dual U-shaped resonator for independent permittivity sensor to characterize two different samples. The discussion in this chapter begins with an introduction to explain the main contribution of the proposed work, design, and simulation of a dual U-shaped resonator, determining the location of sensing areas, scenarios for placing MUTs for independent characterization. The discussion continued with observing the performance of the sensor including the accuracy, sensitivity and correlation of each resonance frequency when characterizing two types of samples with different permittivity simultaneously. The final discussion of this chapter is validating the performance of the proposed sensor through a measurement process and comparing it with existing work to show the novelty of the proposed work.

- Chapter 5, provides the development of integrated microwave sensor (MS) and antenna sensor (AS) based on dual T-shaped resonator for contact and non-contact detection of solid material. The discussion in this chapter begins with an introduction to explain the main contribution of the proposed work, design and simulation of a dual T-shaped resonator, determining the location of sensing areas for contact and non-contact detection, scenarios for placing MUTs for contact and non-contact detection. The discussion continued with observing the performance of the microwave sensor and antenna sensor including the accuracy, sensitivity, and effect of an air gap for non-contact detection using proposed sensor while characterize the MUT. Furthermore, the final discussion of this chapter is validating the performance of the proposed sensor through a measurement process and comparing it with existing work to show the novelty of the proposed work.
- Chapter 6, provides the development of our proposed microwave sensor based on multicascode T-shaped using single port planar resonator. The discussion in this chapter begins with an introduction to explain the main contribution of the proposed work, design, and simulation of multicascode T-shaped resonator, determining the location of sensing areas for microwave sensor, scenarios for placing MUTs for microwave sensor and antenna for data transfer capabilities. The discussion continued with observing the performance of the microwave sensor including the accuracy, sensitivity and while characterize the MUT. In the other hand, the performance of the antenna is indicated by the radiation pattern and data transfer capability using the access point. Furthermore, the final discussion of this chapter is validating the performance of the proposed sensor and antenna through a

measurement process and comparing it with existing work to show the novelty of the proposed work.

- Chapter 7, provides conclusions and suggestions for future work based on the three topics discussed in chapters 4, chapter 5 and chapter 6.

CHAPTER 2

LITERATURE REVIEW

This chapter contains the literature review of this dissertation. The discussion in this chapter begins with an introduction to material characterization and the fundamental concepts of electromagnetic characterization. The following discussion was continued by explaining the measurement techniques for material characterization along with their advantages and disadvantages. The final discussion of this chapter is shown by presenting previous work related to material characterization using microwave sensors, especially for the simultaneous characterization of solid materials both in contact and non-contact as well as multifunctional microwave sensors that can be integrated with other devices.

2.1 Introduction material characterization

Material characterization is a scientific and engineering process used to understand and analyze the properties, behaviour, and composition of materials at the atomic, molecular, and macroscopic levels. It involves a range of techniques and methods aimed at gaining insights into the structure, physical, chemical, mechanical, thermal, and electrical properties of materials. This information is crucial for designing, developing, and using materials in various applications across industries such as manufacturing, electronics, medicine, and construction. The goal of material characterization is to provide a comprehensive understanding of how a material will perform under specific conditions, which helps in making informed decisions about its suitability for a particular purpose.

Solid material characterization based on frequency involves using various techniques to analyze the behaviour and properties of materials as a function of frequency. This approach is particularly useful for understanding how materials respond to different types of energy or interactions at different frequency ranges, which can provide valuable insights into their physical, chemical, and mechanical properties. Materials can have frequency-dependent electromagnetic responses. For example, dielectric spectroscopy studies the electrical response of materials as a function of frequency, revealing information about their polarization and conductivity characteristics. This is important for applications involving insulators, capacitors, and electronic components. Furthermore, dielectric materials exhibit frequency-dependent responses to electric fields. By measuring the dielectric constant and loss tangent over a range of frequencies, we can understand how materials interact with electromagnetic fields at different rates.

Permittivity is a fundamental property of materials that characterizes how they respond to an applied electric field. It measures the ability of a material to allow electric field lines to pass through it and is a critical parameter in understanding the behaviour of dielectric materials within electric fields. Materials with higher permittivity allow more electric field lines to pass through them, while those with lower permittivity resist the penetration of electric field lines. Dielectric materials, which have high permittivity values, are often used to store electrical energy in capacitors by getting polarized when subjected to an electric field. In other words, permittivity is a measure of how a material interacts with electric fields, influencing its electrical behaviour and applications. It's a key parameter in understanding the behaviour of dielectric materials and is vital in various fields of engineering and physics. Moreover, the analysis of materials or the identification of alterations in the electrical

characteristics of substances holds significant implications across various domains, including quality control, biological sensing, underground detection, and the food sector (Zhu et al., 2020)(Jankovic and Radonic, 2017)(Alahnomi et al., 2017). Several methods and techniques are proposed to characterize materials such as cavity mode resonators, transmission line models and microwave planar resonators.

A cavity mode resonator is a confined space or structure designed to trap and resonate electromagnetic waves at specific frequencies (Ndoye et al., 2017a). These resonators can take various shapes, such as rectangular, cylindrical, or spherical cavities, and they have well-defined resonant modes characterized by standing wave patterns of electric and magnetic fields. The material to be characterized is placed inside the resonator cavity. The presence of the material can affect the resonant frequencies and electromagnetic properties of the cavity. By measuring the shift in resonant frequencies, dielectric properties of the material can be extracted, such as its permittivity and loss tangent. Dielectric characterization is essential for understanding how materials respond to electric fields and store electrical energy. In certain cases, cavity mode resonators can also be used to characterize the magnetic properties of materials. Changes in the resonant frequencies can provide insights into the material's magnetic permeability (Liu, Zhang and Huang, 2021). Cavity mode resonators can provide high precision measurements, making them suitable for accurately determining changes in material properties. Cavity resonators can be designed to operate across a wide range of frequencies, from microwave to terahertz and even optical frequencies, allowing for versatile material characterization. Cavity mode resonators are often non-destructive, allowing for repeated measurements on the same sample without altering its properties. However, their design complexity, calibration challenges, and

potential limitations in spatial resolution should be considered when applying these techniques for material characterization. Proper experimental setup, accurate sample preparation, and careful data analysis are essential for obtaining meaningful results.

Furthermore, transmission line model is an equivalent electrical circuit model that represents a transmission line's behaviour, typically consisting of series and parallel combinations of resistors, inductors, and capacitors. It provides a simplified way to describe the propagation of electromagnetic waves along a transmission line. Transmission line model resonators are circuits designed based on the transmission line concept to create resonant structures that interact with electromagnetic waves at specific frequencies (Ebrahimi, Scott and Ghorbani, 2018b). These resonators can be simple or complex, such as rectangular or circular shapes etched on a printed circuit board or integrated on a substrate. The material to be characterized is placed within or near the transmission line model resonator structure. The presence of the material alters the resonant properties of the circuit due to changes in its dielectric constant and loss tangent. The introduction of the material affects the resonant frequencies and impedance characteristics of the transmission line model resonator. These shifts are related to changes in the effective permittivity of the material. By measuring the changes in resonant frequencies or impedance, researchers can extract information about the dielectric properties of the material, such as its permittivity and loss tangent at microwave frequencies (Ebrahimi, Scott and Ghorbani, 2018b). In some cases, transmission line model resonators can also be designed to characterize the magnetic properties of materials by considering the impact of the material on the resonator's inductance (Ebrahimi, Scott and Ghorbani, 2018a). However, considerations related to sample alignment, calibration, substrate effects, and potential simplifications should be

considered when using these resonators for material characterization. Proper experimental setup, careful data analysis, and awareness of potential limitations are essential for obtaining meaningful and accurate results.

Microwave planar resonators are flat, two-dimensional structures that can be fabricated on substrates using microfabrication techniques. They are designed to support specific resonant modes, often at microwave frequencies. The material to be characterized is placed within or near the planar resonator structure. The presence of the material affects the resonant properties of the structure, leading to shifts in resonant frequencies and impedance characteristics. By measuring the changes in resonant frequencies or impedance, researchers can extract information about the material's dielectric properties, including permittivity, loss tangent, and polarization behaviour at microwave frequencies (Li et al., 2022). In some cases, microwave planar resonators can be designed to study the magnetic properties of materials by considering their influence on the resonator's inductance. In some cases, microwave planar resonators can be designed to study the magnetic properties of materials by considering their influence on the resonator's inductance (Saadat-Safa et al., 2019). Microwave planar resonators are specifically designed for microwave frequency ranges, typically in the gigahertz (GHz) range. This makes them suitable for applications in radar, wireless communication, and microwave devices (Zarifi and Daneshmand, 2016). The planar geometry of these resonators allows for easy integration with other microwave components, making them versatile tools in microwave engineering. Microwave planar resonators can offer high sensitivity to changes in material properties, allowing for accurate measurements of dielectric and magnetic responses. Planar resonators are often compact and portable, making them suitable for laboratory and field measurements. Moreover, the planar

geometry of these resonators allows for easy integration with other microwave components, enabling the development of microwave devices. However, accurate measurements require precise sample placement and alignment within the resonator structure, which can be challenging for irregularly shaped or non-uniform materials. Moreover, the choice of substrate material and fabrication process can influence the resonator's behaviour and changes in temperature, humidity, or electromagnetic interference in the environment can affect measurements and may require controlled conditions.

2.2 Fundamental Concepts of Electromagnetic Characterization

Electromagnetic characterization refers to the study and analysis of the electromagnetic properties of materials and structures. Understanding these properties is crucial in various fields, including telecommunications, electronics, materials science, and electromagnetic compatibility. Electromagnetic waves are oscillating electric and magnetic fields that propagate through space. Different materials interact with electromagnetic waves in unique ways. Dielectric constant (permittivity) and magnetic permeability are key material properties affecting wave propagation through a medium.

2.2.1 Dielectric properties material

Dielectric properties are fundamental characteristics of materials related to their behaviour in electric fields. These properties are crucial in understanding how materials respond to and interact with electric fields, making them essential for various applications in electronics, telecommunications, energy storage, and more. The fundamental theory of dielectric properties revolves around concepts such as polarization, electric dipoles, and the

interaction of materials with electric fields. Electric polarization is the process by which positive and negative charges within a material become displaced in response to an applied electric field. This displacement leads to the creation of electric dipoles—pairs of positive and negative charges separated by a small distance. The degree of polarization depends on the material's intrinsic properties and the strength of the applied electric field.

The dielectric constant (ϵ_r) is a measure of a material's ability to store electrical energy in an electric field. It quantifies the ratio of the electric displacement produced within a material to the electric field applied. A high dielectric constant indicates that a material can be strongly polarized by an electric field, making it useful for applications like capacitors. Dielectric loss ($\tan \delta$) represents the energy dissipated as heat within a dielectric material when subjected to an oscillating electric field. It is associated with the imaginary part of the complex dielectric constant (ϵ^*). A low dielectric loss is desirable in applications where energy efficiency is crucial. Both the electric field and electric flux density are related and they can be expressed by:

$$D = \epsilon_o E + P \quad (2.1)$$

Where ϵ_o is the free space permittivity of material. When the applied materials are linear and isotropic, the dipole moment P is in the same direction as E ($P=E$):

$$\begin{aligned} D &= \epsilon_o E + P \\ &= \epsilon_o (1 + X_e) \\ &= \epsilon_o \epsilon_r E \\ &= \epsilon_o E \end{aligned} \quad (2.2)$$

The electric behaviour of a materials with low conductivity is characterized by complex parameter known as permittivity (ϵ):

$$\varepsilon = \varepsilon' - j\varepsilon'' \quad (2.3)$$

The permittivity relies on the interaction between a quantified substance and an applied electric field. This interaction is founded on energy storage and energy dissipation phenomena. The segment of energy exchange between the field and the material characterized by minimal loss is termed energy storage, whereas the absorption of electromagnetic energy by the material is identified as energy dissipation. The real component (ε') of the material pertains to energy storage, while the imaginary component (ε'') pertains to energy dissipation.

The units for permittivity are in Farads per meter (**F/m**), however, in microwave electronics are often dimensionless quantities relative permittivity (ε_r).

$$\begin{aligned} \varepsilon_r &= \frac{\varepsilon}{\varepsilon_0} \\ \varepsilon_r &= \frac{\varepsilon' - j\varepsilon''}{\varepsilon_0} \\ \varepsilon_r &= \varepsilon_r' - j\varepsilon_r'' \end{aligned} \quad (2.4)$$

Where $\varepsilon_0 = 8.854 \times 10^{12}$ **F/m** is the permittivity of the free space, ε_r' is the relative dielectric permittivity and ε_r'' is the loss factor (Chen et al., 2004).

2.2.2 Electromagnetic resonant modes

Electromagnetic resonant modes are specific patterns of electromagnetic field distributions that form within confined or resonant structures when they are excited by an external electromagnetic source. These modes represent specific ways in which electromagnetic energy can be stored and oscillate within a defined space. Understanding electromagnetic resonant modes is essential in various applications, including optics, antennas, cavity resonators, and waveguides.

Resonant modes occur when the natural frequencies of an electromagnetic structure match the frequency of the incident electromagnetic wave. At these frequencies, the electromagnetic energy becomes trapped within the structure, causing it to oscillate with high amplitudes. Electromagnetic resonant modes can occur in a variety of geometries, such as open and closed cavities, waveguides, and antennas. Each geometry supports specific types of resonant modes with distinct field patterns.

In cavity resonators, such as microwave cavities, electromagnetic waves bounce back and forth between reflective surfaces, creating resonant modes. These cavities are used in applications like microwave ovens, particle accelerators, and lasers. Resonant modes also play a crucial role in antennas, where they determine the antenna's radiation pattern and efficiency. Antennas are designed to resonate at specific frequencies to transmit or receive electromagnetic signals effectively.

Electromagnetic resonant modes in planar resonators are specific patterns of electromagnetic fields that form within flat or planar structures when they are excited by external electromagnetic sources. Planar resonators are two-dimensional structures that can support a variety of resonant modes, each with distinct electromagnetic field distributions. These modes are important in applications such as microwave engineering, antennas, integrated circuits, and optical devices. Planar resonators are flat structures with well-defined boundaries. These structures can be etched onto or fabricated within a planar substrate, often using technologies like microfabrication or printed circuit board techniques. Planar resonators have specific resonant frequencies at which they can efficiently store and exchange electromagnetic energy. These frequencies are determined by the dimensions and geometry of the resonator. Different planar resonator geometries support various resonant

modes, each characterized by a unique pattern of electric and magnetic field distributions. These modes include fundamental (or lowest-order) resonant modes as well as higher-order modes. Planar resonators can be tuned by adjusting their dimensions or material properties. This tunability is useful for designing filters, oscillators, and sensors with specific frequency responses. Planar resonators are conducive to integration within larger systems, making them suitable for compact and high-density circuit designs. Planar resonators offer advantages such as ease of fabrication, compatibility with semiconductor processes, and the potential for integration with other planar components. However, planar resonators may face challenges related to radiation losses, substrate effects, and material properties that could limit their performance in certain applications.

The planar microstrip device stands out as one of the most widely utilized resonant techniques, owing to its simplified manufacturing process, compact size, and its compatibility for integration with both active and passive microwave components. Figure 2.1(a) illustrates the configuration of a microstrip line, encompassing a ground plane and a conductor with width ($\mathbf{w}$), separated by a dielectric substrate. Notably, a substrate with a high dielectric constant concentrates the field in its proximity. This phenomenon is depicted in Figure 2.1(b) through field sketch lines. In scenarios where the dielectric constant is absent ($\epsilon_r = 1$), a two-wire line featuring a flat strip conductor situated above a ground plane within a uniform medium (Air) is established. Under these conditions, a straightforward TEM transmission line materializes, characterized by a propagation constant ($\beta = k_0$) and a phase velocity ($\mathbf{v_p} = \mathbf{c}$).

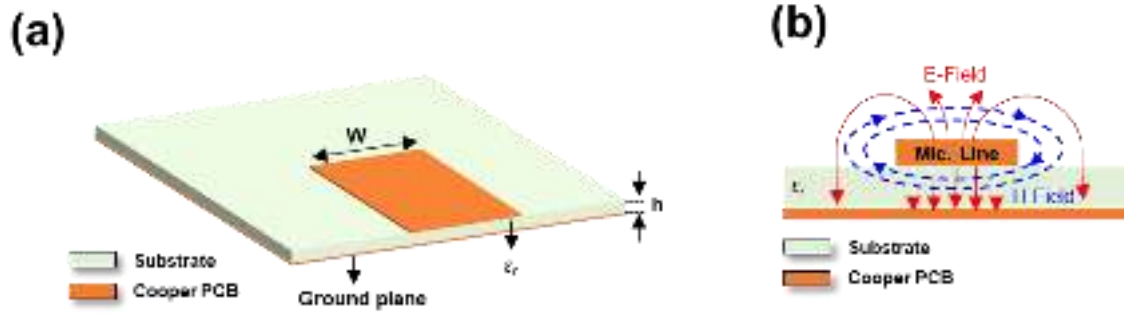


Figure 2.1 (a) structure of microstrip line, (b) electric and magnetic field of microstrip line

The microstrip structure typically encompasses a greater concentration of field lines within the dielectric area situated between the ground plane and the width conductor, as well as within the upper air region above the substrate, albeit to a lesser extent. This disparity implies that microstrip fields are not confined solely within a uniform dielectric region, unlike the fields observed in stripline configurations. Consequently, microstrip cannot support the propagation of TEM (Transverse Electro-Magnetic) waves. The phase velocity of TEM fields in the dielectric region is influenced by the factor $c/\sqrt{\epsilon_r}$, whereas the phase velocity of the TEM wave in the air region remains c . This disparity renders the establishment of a phase-matching condition between the dielectric-air interface unattainable.

The wave undergoing propagation along a microstrip line deviates from the conventional TM or TE waveforms, as well as from a pure TEM wave. Consequently, the precise field configuration of the microstrip line encompasses a hybrid TM-TE nature, commonly referred to as a quasi-TEM wave. Analyzing the propagation of waves in the microstrip structure necessitates the application of advanced analytical methods and rigorous

numerical simulations. However, a preliminary approach involves employing static or quasi-static solutions to compute the distributed capacitance, subsequently facilitating the derivation of phase velocity, characteristic impedance, and propagation constant. Consequently, the propagation constant and phase velocity can be determined as follows:

$$\beta = k_o \sqrt{\epsilon_r} \quad (2.5)$$

$$v_p = \frac{c}{\sqrt{\epsilon_e}} \quad (2.6)$$

Where β is the propagation constant and v_p is the phase velocity.

For further exploration into the fundamental concept of effective dielectric permittivity, Figure 2.2 illustrates microstrip lines utilizing the quasi-static approach. When air functions as the filling medium as depicted in Figure 2.2(a) and possesses a dielectric constant of ($\epsilon_r = 1$), the microstrip line becomes capable of sustaining a TEM wave, with the phase velocity (v_p) being equivalent to the speed of light (c). Conversely, if the transmission system incorporates a dielectric material characterized by a permittivity greater than unity ($\epsilon_r > 1$), as shown in Figure 2.2(b), the microstrip remains supportive of a TEM wave and its phase velocity will become:

$$v_p = \frac{c}{\sqrt{\epsilon_r}} \quad (2.7)$$

Moreover, when introducing a dielectric material characterized by a dielectric permittivity (ϵ_r) that is only partially integrated into the transmission system, as depicted in Figure 2.2(c) and Figure 2.2(d), the notion of an effective dielectric constant (ϵ_{eff}) is brought forth.

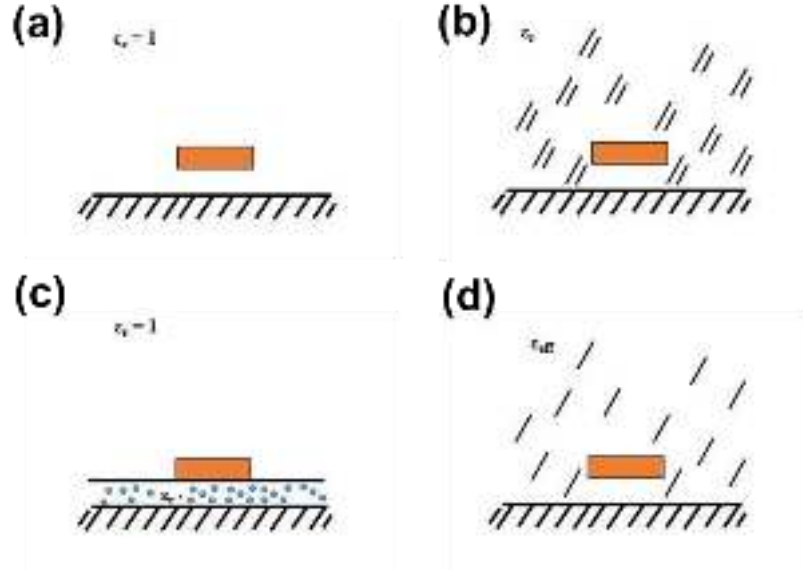


Figure 2.2 (a) Microstrip filled completely with air, (b) Microstrip filled completely with ϵ_r , (c) Microstrip filled partially with ϵ_r , (d) microstrip filled completely with ϵ_{eff}

This concept serves to compute essential transmission line parameters, including phase velocity (v_p), wavelength (λ_g), and characteristic impedance (Z_o):

$$\lambda_g = \frac{\lambda_o}{\sqrt{\epsilon_{eff}}} \quad (2.8)$$

$$v_p = \frac{c}{\sqrt{\epsilon_{eff}}} \quad (2.9)$$

$$Z_o = \frac{Z_{oc}}{\sqrt{\epsilon_{eff}}} = \frac{1}{v_p C_1} \quad (2.10)$$

2.2.3 Resonant perturbation theory

Resonant Perturbation Theory is a mathematical and theoretical framework used in quantum mechanics to analyze and understand the behaviour of quantum systems when subjected to external forces or perturbations. This theory is particularly relevant when dealing with systems that have discrete energy levels and are subjected to time-dependent

perturbations, such as electromagnetic fields or potential fluctuations. Resonant Perturbation Theory provides a systematic approach to studying how these perturbations affect the energy levels and transitions of quantum systems, especially when the perturbation frequency is close to the energy difference between quantum states.

Perturbation theory is a solution to determine changes in energy due to the presence of small disturbances such as electric or magnetic fields. In the microwave sensor, the perturbation theory can be used to determine the permittivity of the Material Under Test (MUT). The electric and magnetic fields of the resonator are disturbed by placing the MUT on top of the resonator. The interaction of the MUT with the E-field and H-field will cause changes to the electric and magnetic fields of the resonator. The changes in the resonant frequency and the properties of the MUT are related and described in the following equation (Alahnomi et al., 2017):

$$\frac{\Delta f_r}{f_r} = \frac{\int_v (\Delta \epsilon E_1 \cdot E_0 + \Delta \mu H_1 \cdot H_0) dv}{\int_v (\epsilon_0 |E_0|^2 + \mu_0 |H_0|^2) dv} \quad (2.11)$$

where f_r is the resonant frequency, while Δf_r is the shift of the resonant frequency before and after the MUT is loaded. $\Delta \epsilon$ and $\Delta \mu$ represent changes in permittivity and permeability while perturbation volume is represented by v . The distribution of electric and magnetic fields without perturbation and with perturbation are shown by E_0 , H_0 , and E_1 , H_1 , respectively. Furthermore, the change in the permeability ($\Delta \mu$) may be assumed to zero for the dielectric material as discussed in (Ansari et al., 2015). However, the stored energies of both electric field and magnetic field in the resonant structure have to be equal to each other at resonance frequency. Therefore, the expression in (2.11) can be simplified as (Boybay and Ramahi, 2012; Ansari et al., 2015):

$$\frac{\Delta f_r}{f_r} = \frac{f_{vs}(\Delta \epsilon \cdot E_0 \cdot E_1) dv}{f_v(\Delta \epsilon \cdot |E_0|^2) dv} \quad (2.12)$$

Referring to (2.12), it is clear that the change in permittivity of the MUT is in line with the changes in the resonance frequency. The MUT placed in the sensing area will perturbate the E-field. The interaction of the MUT to the resonator can be assumed as a capacitive load that causes the resonant frequency to shift.

2.3 Measurement techniques for material characterization

Material characterization involves the measurement and analysis of various properties of materials, and electromagnetic materials are no exception. The dielectric constant, or permittivity, is a key parameter describing how a material responds to an electric field. Common techniques for measuring dielectric constant include capacitance measurements, resonant cavity methods, and transmission line methods.

2.3.1 Non resonant methods

Non-resonant methods for characterizing materials involve techniques that do not rely on the excitation of specific resonant frequencies within the material being studied. These methods are used to determine various properties of materials, such as dielectric constant, conductivity, permeability, and other electromagnetic parameters, over a broad frequency range without focusing on particular resonances. Unlike resonant methods, which target specific frequencies where the material exhibits resonant behaviour, non-resonant methods aim to gather information across a wide spectrum of frequencies. This allows for a comprehensive understanding of the material's electromagnetic properties, including how they vary with frequency.

Non-resonant techniques require a transmission line to guide the wave towards the material being examined, followed by the acquisition of both reflected and transmitted energies. Options for the transmission line encompass metallic and dielectric waveguides, free space, planar transmission lines, and coaxial cables. Non-resonant methods predominantly encompass reflection and transmission approaches. In the context of reflection methods, material properties are ascertained by analyzing the phase and amplitude measurements of signal reflections.

Non-resonant methods require a means of directing the electromagnetic energy toward a material, and then collecting what is reflected from the material, and/or what is transmitted through the material as shown in Figure 2.4. In principle, all types of transmission lines can be used to carry the wave for non-resonant methods, such as coaxial line, hollow metallic waveguide, dielectric waveguide, planar transmission line, and free space.

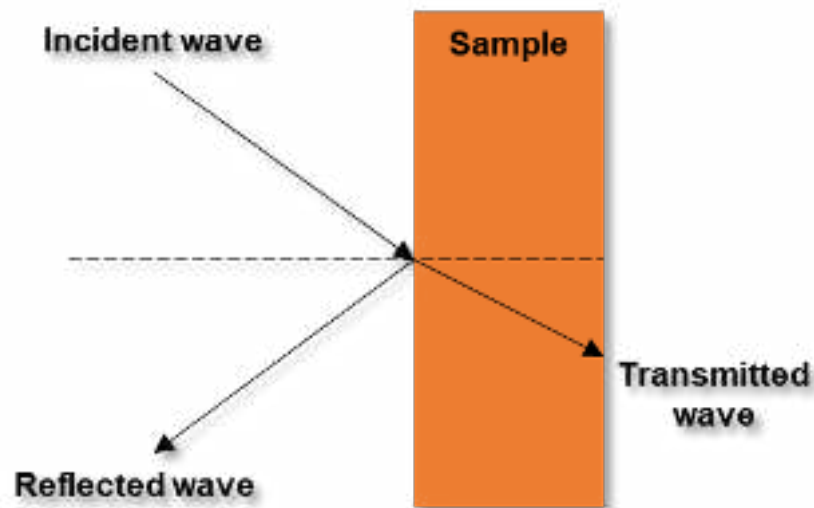


Figure 2.3 Boundary condition for material characterization using a non-resonant method (Ghione and Pirola, 2017)

2.3.1.1 Reflection methods

Reflection techniques involve sending electromagnetic waves towards a specimen being examined. The characteristics of the material under scrutiny are inferred from the reflection coefficient at a specified reference plane. Typically, such methods are capable of measuring just one attribute – either permittivity or permeability.

Two types of reflections are often used in materials property characterization: open-circuit reflection and short-circuit reflection, and the corresponding methods are called open-reflection method and shorted reflection method. As coaxial lines can cover broad frequency bands, coaxial lines are often used in developing measurement fixtures for reflection methods.

In a reflection method, the measurement fixture made from a transmission line is usually called measurement probe or sensor. In order to increase the measurement accuracy and sensitivity, or to satisfy special measurement requirements, the measurement probes are often specially designed. Figure 2.4 shows an open-circuited flanged coaxial dielectric probe, and the sample under test directly contacts the open end of the coaxial line. The impedances at the two sides of the interface are different, so there is a reflection when electromagnetic wave propagates through the interface, and the reflectivity is determined by the impedances of the media at the two sides of the interface. As the impedance of the side occupied by the sample is related to the electromagnetic properties of the sample, from the reflectivity at the interface, the properties of the sample can be obtained.

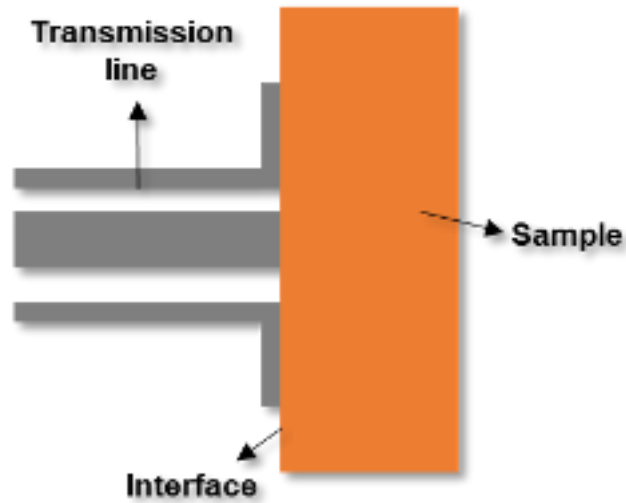


Figure 2.4 Open-circuited reflection (Ghione and Pirola, 2017)

As the impedance of the space filled with the sample is related to the permittivity and permeability of the sample, in principle, both the permittivity and permeability can be obtained provided sufficient independent reflection measurements are made. However, in most cases, only one independent measurement is made, so only one materials property parameter, either permittivity or permeability, can be obtained.

Furthermore, for shorted-circuit reflection, a piece of sample is inserted in a segment of shorted transmission line, with the end face of the sample located at a distance Δl from the short. We assume that only the prime mode exists in the transmission line, the sample is homogeneous and isotropic, and there exists only transverse electric field in the transmission line. Figure 2.5 shows a general case of short-circuited reflection.

For the configuration shown in Figure 2.5, standing waves are built in the region between the sample and the short circuit and between the calibration plane and the sample front face. Depending on the sample length and the other lengths, at certain frequencies permittivity can be measured accurately, and at other frequencies permeability can be

measured accurately. Meanwhile, according to the parameters to be measured, we put the samples at suitable positions.

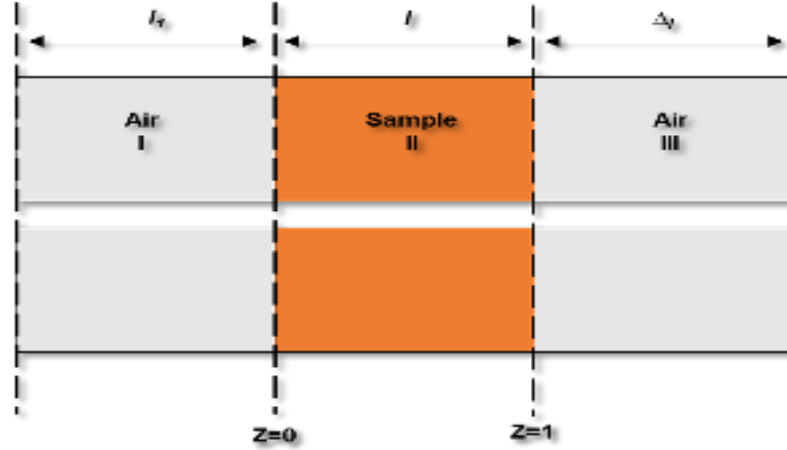


Figure 2.5 Short-circuited reflection (Ghione and Pirola 2017)

The position of the short termination is a low electric field and high magnetic field region, and the position $\lambda/4$ away from the short termination is a high electric field and low magnetic field region. For permittivity measurements the sample should be moved away from the short termination, while for permeability measurements the sample should be moved near to the short termination.

2.3.1.2 Transmission/reflection methods

When determining the effective permittivity of composite materials, it is essential that the dimensions of the sample are significantly greater than the dimensions of the inclusions. When dealing with composites that contain inclusions of a size comparable to the wavelength of the microwave signal, such as fiber composites, conventional techniques like coaxial line and waveguide methods are unsuitable. In such scenarios, free-space methods become the preferred choice. The subsequent sections outline the fundamental

criteria for conducting free-space measurements and subsequently explore three instances of free-space techniques. In order to attain precise measurement outcomes through the utilization of the free-space technique, a number of conditions must be met. These primarily encompass the far-field condition, dimensions of the sample, and the ambient measurement surroundings. In free-space measurements, the fulfilment of the far-field criterion necessitates that the distance **(d)** separating the antenna and the sample guarantees the incident wave from the antenna to the sample can be treated as a planar wave(Ghione and Pirola, 2017).

$$d > \frac{2D^2}{\lambda} \quad (2.13)$$

where λ is the wavelength of the operating electromagnetic wave and D is the largest dimension of the antenna aperture. For an antenna with circular aperture, D is the diameter of the aperture, and for an antenna with rectangular aperture, D is the diagonal length of the rectangular aperture. When the far-field requirement is fulfilled, free space can be taken as a uniform transmission line, and most of the measurement schemes discussed above for coaxial reflection measurements can be realized by free space.

In the measurement of permittivity and permeability of planar samples using free-space method, if the sample size is much smaller than the wavelength, the responses of the sample to electromagnetic waves are similar to those of a particle object. To achieve convincing results, the size of the sample should be larger than the wavelength of the electromagnetic wave. To further minimize the effects of the scatterings from the sample boundary, the sample size should be twice larger than the wavelength. In a free-space transmission structure, as the electromagnetic wave is not limited by a sharp boundary, the measurement results may be affected by the environments. At lower frequencies, the effects

of environments are more serious. To minimize the effects of the environments, it is recommended to conduct free-space measurements in an anechoic chamber.

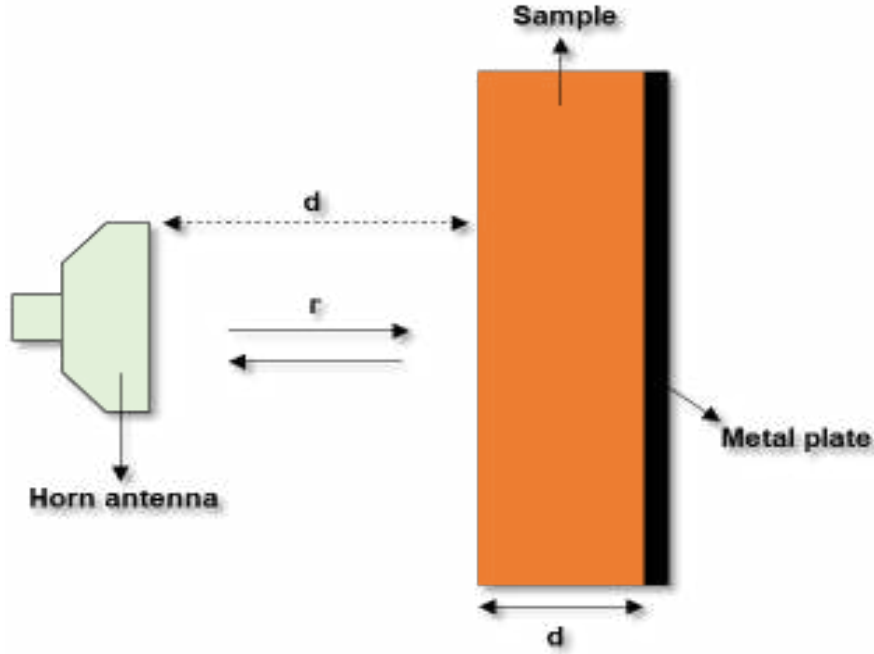


Figure 2.6 Free-space reflection method (Ghione and Pirola 2017)

Figure 2.6 shows a typical setup for short-circuited free-space reflection measurement. The sample backed by a metal plate is placed in front of an antenna with a distance satisfying the far-field requirement. The complex reflectivity S_{11} at the interface between the free space and the sample is given by:

$$S_{11} = \frac{jZ \tan(\beta d) - 1}{jZ \tan(\beta d) + 1} \quad (2.14)$$

Where Z is the wave impedance of material under test normalized to the wave impedance of material under test. For nonmagnetic materials, the normalized wave impedance is given by free space and β is the phase constant in the material under test. For nonmagnetic materials, the normalized wave impedance is given by:

$$Z = \frac{1}{\sqrt{\epsilon_r}} \quad (2.15)$$

and the phase constant β is given by:

$$\beta = \frac{2\pi}{\lambda} \sqrt{\epsilon_r} \quad (2.16)$$

Where λ is the free-space wavelength and d is the thickness of the sample. Therefore, the dielectric permittivity of the sample can be obtained from the complex reflectivity.

2.3.2 Resonant methods

This method is often called dielectric resonator method. It can be used to measure the permittivity of dielectric materials and the surface resistance of conducting materials. Though resonant methods can only measure materials properties at single or several discrete frequencies, resonant methods are widely used because of their high accuracy and sensitivity. Resonant methods include resonator methods and resonant-perturbation methods. In a resonator method, the sample under test serves as a resonator or a key part of a resonator, and the properties of the sample are determined from the resonant properties of the resonator. In a resonant-perturbation method, the sample is introduced into a resonant structure, and the properties of the sample are calculated from the change of resonant properties of the resonant structure caused by the introduction of the sample (Martín et al., 2022).

In actual measurement fixtures for materials property characterization, resonators are often shielded by conductors. In the measurement of dielectric materials, the surface impedance of the shielding conductors, usually metals, is assumed to be known. If the properties of the dielectric material are known, from the resonant properties of the resonator,

the surface impedance of the shielding conductors can be obtained. In the final part of this chapter, we discuss the measurement of surface impedance of high-conductivity materials using dielectric resonator methods.

2.3.2.1 Dielectric resonant methods

Dielectric resonator methods are widely used in the characterization of low-loss dielectric materials. In the design of dielectric resonator methods for the characterization of the dielectric properties of materials, the structures of both the dielectric material to be studied, and the metal shield should be designed. Different dielectric resonator methods have different structures of metal shield and dielectric sample, and thus have different algorithms for the calculation of materials properties. The configuration often used in the dielectric resonator method is shown in Figure 2.7.

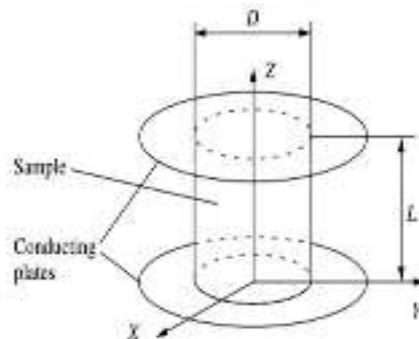


Figure 2.7 A cylindrical dielectric sandwiched between two conducting plates (Chen et al., 2004)

2.3.2.2 Resonant perturbation methods

In a resonant perturbation method, the sample is inserted into a resonator, and the properties of the sample are calculated from the changes of the resonant frequency and the quality factor of the resonator caused by the sample. In a dielectric resonator method, the

sample under test resonates in the measurement circuit, and the dielectric properties of the sample are deduced from its resonant frequency and quality factor. In a resonant perturbation method, the sample under test is introduced to a resonator, and the electromagnetic properties of the sample are deduced from the change of the resonant frequency and quality factor of the resonator. Owing to its high accuracy and sensitivity, and its flexibility in sample preparation, resonant perturbation methods are widely used for low-loss bulk samples, powders, small-size samples, and samples of irregular shapes.

Generally, there are three types of cavity perturbations: cavity-shape perturbation, wall-impedance perturbation, and material perturbation. Cavity-shape perturbation can be achieved by pulling out or pushing in part of the cavity wall. Usually, cavity-shape perturbations change the resonant frequency of the cavity and the stored energy in the cavity, but do not change the energy dissipation in the cavity. The cavity-shape perturbation is often used to retune the resonant frequency of a resonant cavity. Wall-impedance perturbation can be achieved by replacing part of the cavity wall, keeping the shape of the cavity unchanged. The wall-impedance perturbation is often used for measuring the surface impedance of conductors.

In material perturbation, a material is introduced into a cavity, and the resonant frequency and quality factor of the cavity are thus changed. The complex permittivity or complex permeability of the material can be determined from the changes of the resonant frequency and the quality factor of the cavity due to the introduction of the sample. In the following discussion, “cavity-perturbation method” refers to “material-perturbation method.”

2.3.2.3 Resonant planar circuit methods

Similar to their coaxial and waveguide counterparts, planar-circuit resonant methods include resonator methods and resonant-perturbation methods. Usually, resonant methods have higher accuracy and sensitivity than non-resonant ones, but resonant methods can only provide results at a single frequency or several discrete frequencies. In a resonator method, the sample under test is used as a substrate of a planar resonator. The dielectric constant of the sample is calculated from the resonant frequency of the resonator, and the loss tangent of the material is calculated from the quality factor of the resonator. Figure 2.8 shows a straight ribbon microstrip resonator. The measurement frequency is mainly determined by length of the microstrip, and the conductivity of the strip and ground conductors affects the sensitivity of loss tangent measurement.

In a resonant-perturbation method, part of the substrate of a planar resonator is replaced by the sample under study. From the changes of the resonant frequency and quality factor, the properties of the sample can be derived (Abdulnour et al., 1995). Figure 2.8 shows a measurement fixture for resonant-perturbation method. The position A-A is close to the open end of the resonator, so the electric field dominates at the position A-A. If a sample is placed at position A-A, its dielectric properties can be measured. The position B-B is near the center of the resonator, so the magnetic field dominates at the position B-B. If a sample is placed at the position B-B, its magnetic properties can be measured. Usually, calibration with a standard sample is needed to obtain absolute values of materials properties.

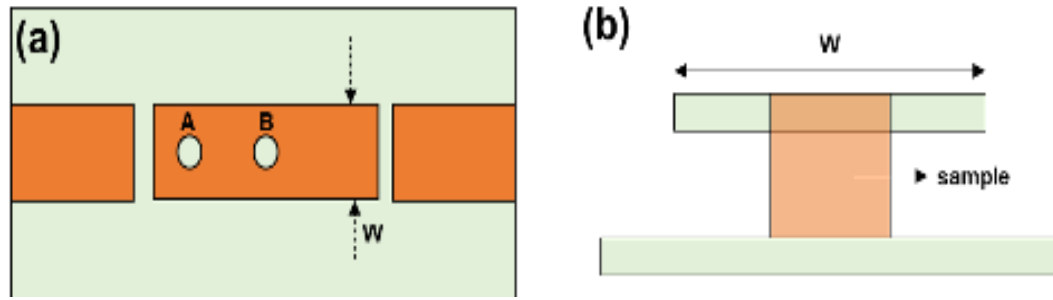


Figure 2.8 Perturbation to the substrate of a microstrip resonator. (a) Top view and (b) cross-section view at positions A-A and B-B (Ghione and Pirola, 2017)

2.4 Analysis summary for material characterization

The selection of techniques is typically influenced by factors such as desired measurement precision, frequency of interest, anticipated permittivity value (ϵ_r), material characteristics (such as isotropic or homogeneous), material form (like powder, liquid, sheet, or solid), sample size limitations, contact or non-contact methods, potential for damage, non-destructive or destructive approaches, and temperature conditions. As previously mentioned, the choice of dielectric measurement technique is closely tied to various factors, and this decision is specific to the material being tested. Each material's electrical characterization varies based on its unique dielectric properties.

Furthermore, the behavior of dielectric materials can change with temperature, size, and testing frequency. As the frequency increases, the dielectric constant tends to decrease, leading to a frequency-dependent relationship. Consequently, the dielectric constant may exhibit differences from high-frequency measurements when materials are tested at lower frequencies. Comparative information about measurement techniques, along with their respective advantages and disadvantages, is presented in Table 2.1 and Table 2.2.

Table 2.1 Advantages and disadvantages of measurement technique of material characterization (Muñoz-Enano *et al.*, 2020; Alahnomi *et al.*, 2021)

Measurement techniques	Advantages	Disadvantages
Transmission and reflection line	- Direct measurement of transmission. - Broad Frequency Range. - Support for Non-Destructive measurement. - High Precision. - Suitable for dielectric measurements.	- Require precise sample placement and alignment within the transmission line. - Calibration complexity. - Substrate effects. - Limited spatial resolution.
Resonant method	- High sensitivity. - Direct measurement. - Frequency selectivity. - Suitable for dielectric and magnetic characterization. - Can be applied to a wide range of materials, including solid, liquid, powder, and composite materials.	- Complex design and fabrication - Require precise sample placement and alignment within the resonator's geometry. - Environmental effects. - Substrate effects. - Limited spatial resolution.
Free space	- Non-contact measurements. - Accommodate relatively large sample sizes, making them suitable for bulk materials or samples that cannot be easily placed within resonant cavities. - Broad frequency range. - Dielectric and reflectivity measurements. - High Spatial Resolution.	- Complex setup. - Environmental interference. - Limited control over Propagation - Signal attenuation. - Sample shape and alignment. - Limited applications for thin films.

Table 2.2 Extracting the Dielectric Properties of Different Materials at Various Measurement Methods (Muñoz-Enano *et al.*, 2020; Alahnomi *et al.*, 2021)

Measurement techniques	Materials	Dielectric properties	Parameters
Transmission and reflection line	- Solid - Liquid - Biomaterial - Powder material	μ_r and ϵ_r	S_{11} and S_{21}

Measurement techniques	- Materials	Dielectric properties	Parameters
Resonant method	- Solid - Liquid - Biomaterial - Powder material - Metals and Dielectrics with Temperature-Dependent Properties	μ_r and ϵ_r	S_{11} and S_{21}
Free space	- Solid - Liquid - Powder material - Materials with Temperature-Dependent Properties - Materials with Reflective Properties	μ_r and ϵ_r	Frequency and Q-Factor

2.5 Recent development of microwave planar sensor

A microwave planar sensor is a specialized device used in microwave and radio frequency (RF) engineering for sensing and detecting changes in various physical parameters, such as dielectric properties, temperature, humidity, and chemical composition. It utilizes the principles of microwave technology to interact with the environment and produce measurable responses, enabling the monitoring and analysis of different substances and conditions. Applications of microwave planar sensors include:

1. Industrial quality control: planar sensors are used in various industrial processes for quality control and monitoring. For instance, they can detect permittivity, permeability and thickness of materials, monitor the curing of composite materials, and assess the properties of liquids and solids.

2. Moisture sensing: microwave sensors can measure moisture levels in materials, such as soil, food products, and building materials. Changes in moisture content affect the dielectric properties of the material, which can be detected by the sensor.
3. Temperature sensing: variations in temperature can also affect dielectric properties. Microwave planar sensors can be used to measure temperature changes in industrial processes, environmental monitoring, and medical applications.
4. Chemical sensing: some planar sensors are designed to detect changes in chemical composition. For example, changes in the dielectric properties of a material exposed to certain gases can be indicative of chemical reactions or gas concentrations.
5. Biological sensing: microwave planar sensors can be used in biomedical applications for detecting biological molecules, pathogens, or changes in tissue properties. This has potential applications in medical diagnostics and healthcare.

In Figure 2.9, a dual-mode magnetic-LC (MLC) resonators-based microwave sensor with ultra-high sensitivity and ultra-low estimation error is proposed (Zhang, Ruan and Cao, 2022). By using MLC resonators loaded on an open-ended flared microstrip patch, a dual-mode resonance of the proposed sensor can be observed. The dual-mode microwave sensor is designed and measured to operate at 12.6 GHz and 16.2 GHz. When the proposed dual-mode microwave sensor is submerged in the oils to be tested, the changes of resonance frequency and peak attenuation are utilized for sample identification and dielectric characterization. The complex permittivity of oils under test are extracted by building an approximate model in High Frequency Structure Simulator (HFSS).

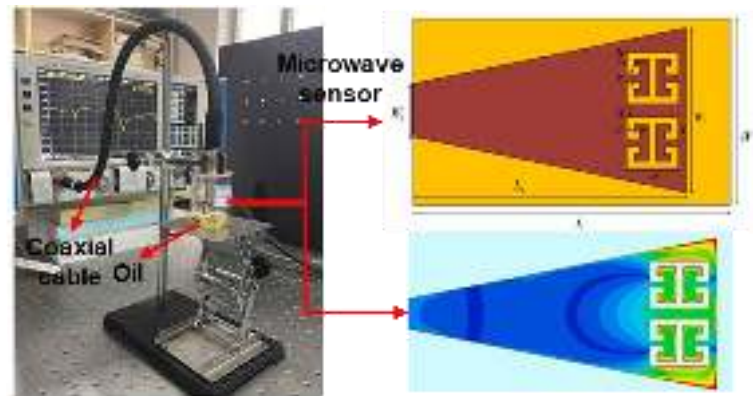


Figure 2.9 Microwave sensor for oil characterization(Zhang, Ruan and Cao, 2022)

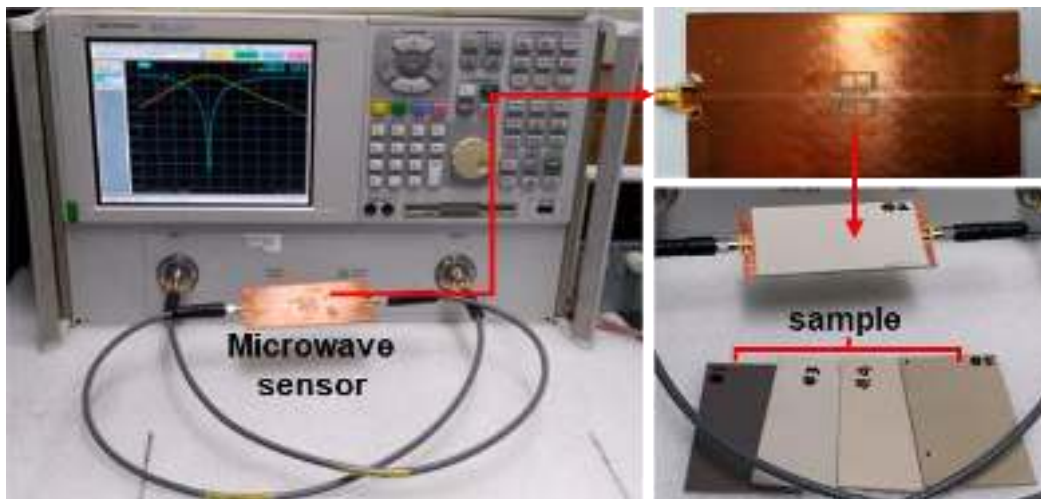


Figure 2.10 Microwave sensor for solid material characterization (Yeo and Lee, 2019a)

Furthermore, Figure 2.10 shows microwave sensor based on an interdigital-capacitor-shaped defected ground structure (IDCS-DGS) in a microstrip transmission line for the dielectric characterization of planar solid materials. The proposed IDCS-DGS was designed by modifying the straight ridge structure of an H-shaped aperture and worked at resonant frequency of 1.5 GHz. The length of the Material Under Test (MUT) was slightly reduced to 90 mm because there existed two protruding parts in the SMA connectors for soldering in the two ports of the fabricated sensors.

2.6 Microwave sensor for material characterization

Permittivity is one of the key parameters to determine the characteristics of materials based on their electrical properties. In other words, permittivity is the ability of a material to transmit or permit an electric field (Chakyar et al., 2017). Therefore, determining the permittivity characteristics of the material is necessary for microwave engineering. Furthermore, the characterization of materials based on permittivity is essential for several applications, including the biomedical, food industry, and pharmaceutical industry. Generally, the permittivity of the Material Under Test (MUT) can be observed and determined by using a microwave sensor such as liquid (Lee et al., 2019; Zidane et al., 2021), solid (Rammah A. Alahnomi et al., 2017; Kiani et al., 2018), and temperature (Ndoye et al., 2017b; Abduljabar, Hamzah and Porch, 2018). The characteristics and permittivity of the material can be determined and measured using a microwave sensor with several methods such as a planar resonator (Acevedo-Osorio, Reyes-Vera and Lobato-Morales, 2020), cavity (Caleffo and Correra, 2018), waveguide (Kapranov and Kouzaev, 2019).

Planar resonators have the advantages of compact design, low production costs, and easy fabrication (Muñoz-Enano et al., 2020; Gulsu et al., 2021). In order to increase the accuracy and sensitivity of the planar resonator, several previous studies have proposed a microwave sensor with high sensitivity and accuracy using a split ring resonator (SRR) (Saadat-Safa et al., 2019), complementary split-ring resonator (CSRR) (Yang et al., 2016), interdigital structure (Yeo and Lee, 2019a; Liu, Zhang and Huang, 2021) substrate integrated waveguide (Lobato-Morales et al., 2014; Ndoye et al., 2017a) and microstrip (Sanders, Yao and Huang, 2015).

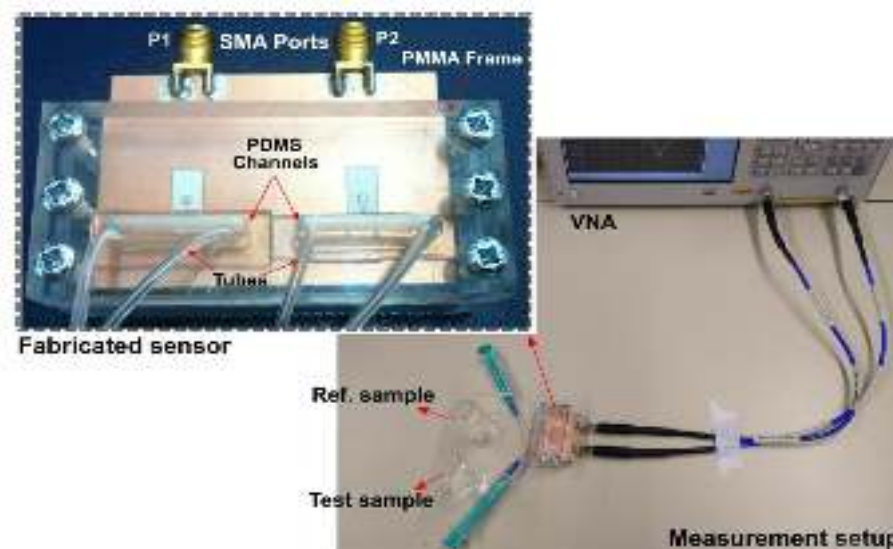


Figure 2.11 Differential microwave sensor for characterization of Glycerol-Water solutions(Ebrahimi et al., 2020)

Generally, Material Under Test (MUT) characterized by microwave sensors has limitations in terms of number and detection time for a large number of quantities. To detect the permittivity and its characteristics, the MUT is placed on the top of the surface of the resonator, also known as the perturbation method. This will cause the resonant frequency of the resonator to shift according to the permittivity of the MUT.

In (Kiani, Rezaei and Navaei, 2020), a dual-mode resonator based on SRR was proposed to detect two different types of liquids. Dual-mode resonators operating at frequencies of 5.7 and 7.8 GHz successfully detect the permittivity of ethanol, glucose, and distilled water. A dual-mode resonator operating at the frequencies of 0.5 GHz and 1.03 GHz has been described to eliminate the interfering effect of RH on the humidity for chemical sensing. Dual-mode resonator microfluidic sensors have been investigated for simultaneous detection of temperature and various liquids as shown in Figure 2.11. Another study was conducted to detect the permittivity of solid MUT using dual-band sensors with CSRRs

(Yang et al., 2016)(Lee and Yang, 2014), SRR (Al-Behadili et al., 2020; Ebrahimi et al., 2020), complementary bisymmetric-SRR (Haq et al., 2019), magnetic LC (Ebrahimi, Scott and Ghorbani, 2020a), and microstrip line ring resonator (Lim, Kim and Hong, 2018).

2.7 Microwave sensor for contact and non-contact detection

Generally, MS are used to detect permittivity (Kiani, Rezaei and Navaei, 2020)(Bagci, Gulsu and Akaoglu, 2022) by placing the material under test on the surface of the sensing area of the sensor. Several previous studies have proposed MS to characterize solid and liquid materials using Complementary Split Ring Resonator (CSRR) (Rahman et al., 2020)(Chen, Xu and Yao, 2018), Split Ring Resonator (SRR) (Xu et al., 2018)(Chakyar et al., 2017), planar inductor-capacitor (LC) resonators (Ebrahimi, Scott and Ghorbani, 2018b), IDC (Gugliandolo et al., 2021; Wang et al., 2021), microstrip line ring resonator (Lim, Kim and Hong, 2018), and Substrate Integrated Waveguide (SIW) (Ndoye et al., 2017b; Soltan, Sadeghzadeh and Mohammad-Ali-Nezhad, 2022). However, the structure of the previous sensors still uses a two-port configuration so it has disadvantages such as being wasteful of resources and cannot be connected directly to wireless devices. Furthermore, a single port configuration MS has also been proposed to detect the permittivity of solid materials using slot-loaded microstrip (Yeo and Lee, 2019b) and aperture coupling (Behdani et al., 2020). The advantages of MS with a single port are having independent characteristics, a simpler structure, more economical resources, and having a hybrid function as an antenna so that it can be used to transmit data using a wireless network. Generally, material characterization using MS is carried out by placing the MUT in direct contact so that it can

reduce the performance of the sensor. This caused the surface of the sensor to potentially be damaged due to oxidation and corrosion. To overcome this limitation, non-contact detection is proposed as a solution to make the sensor more durable, accurate, and sterile.

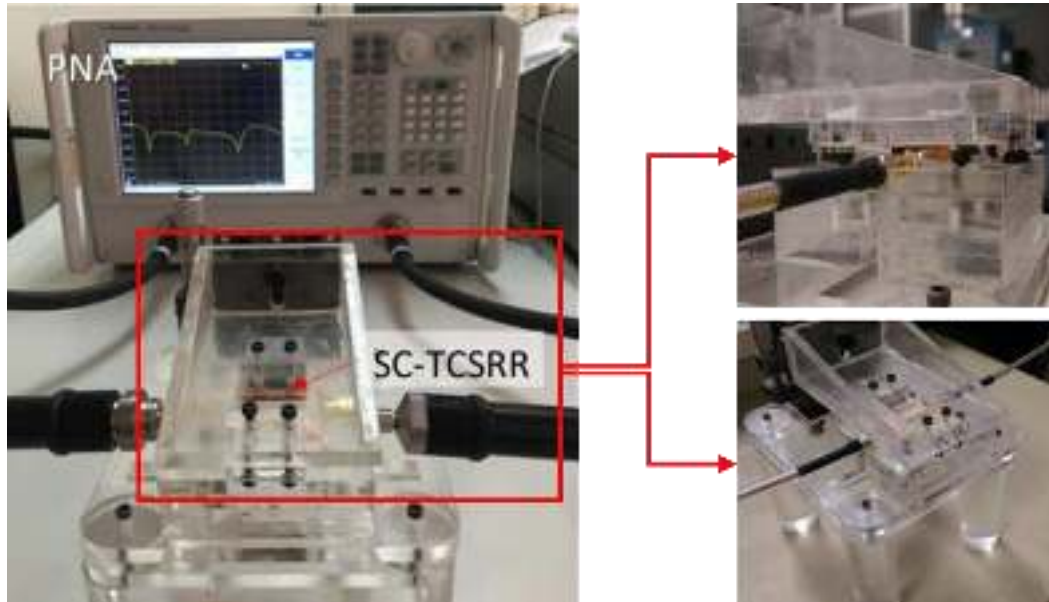


Figure 2.12 Non-contact measurement of complex permittivity and thickness by using planar resonators(Yang et al., 2016)

Non-contact detection using MS by utilizing the electric and magnetic fields from the resonator has been proposed to characterize several materials (Huang, Farahanipad and Singh, 2017; Yao, Tong and Cui, 2020; Li et al., 2022). Previous research conducted by Yang et al. (2016) proposed non-contact detection for complex permittivity and thickness of solid materials using a single-compounded triple complementary resonator split ring resonator with a maximum gap of 0.2 mm as shown in Figure 2.12. Furthermore, the research proposed by Ali et al. (2021) described non-contact detection for permittivity and thickness detection of MUT for solid materials using an interdigital capacitor (IDC) resonator with a gap of 0.3 mm. However, the gap between the MUT and the sensor was still limited and

reduced the performance and sensitivity of the sensor. Another study by Wu and Wang (2023) proposed non-contact detection with an antenna based on an AMC array for permittivity detection of water and ethanol in oil. Furthermore, the research presented by Li et al. (2022) proposed a microwave active AS for sensing liquid properties. However, the measurement process required the antenna interrogator to detect the permittivity of the MUT.

2.8 Multifunctional microwave sensor

Generally, the characterization of solid materials is carried out by observing the permittivity at a certain frequency (Rammah A. Alahnomi et al., 2017; Muñoz-Enano et al., 2020). Permittivity relates to the ability of a material to store an electric field under certain conditions (Boybay and Ramahi, 2012). Several approaches have been proposed to determine the permittivity of solid materials, for example using planar sensors (Amyrul A. Mohd Bahar et al., 2019)(Al-Behadili et al., 2020), waveguides (Ndoye et al., 2017a; Lee et al., 2019) and combinations of active electronic components (Abdolrazzaghi and Daneshmand, 2017; Abdolrazzaghi et al., 2021). Planar sensors have several advantages including compact design (Karimi, Arsalan and Shamim, 2019), low cost, and high accuracy (Abduljabar, Hamzah and Porph, 2021). However, the sensor has a limited function that is to merely detect the permittivity of the material under test (MUT). In the future, multifunctional sensors are needed to be integrated with the internet-of-things and wireless-based devices and another application (Aghoutane et al., 2022; Najumunnisa et al., 2022). Another limitation, there is still a limited number of materials that can be characterized simultaneously. The advantage of simultaneous characterization is that there can be more

than one detected material so that it can reduce the time needed to detect a large number of materials.

The previously proposed permittivity sensor (Armghan et al., 2021; Wang et al., 2021) used a resonator with two ports so that it was not possible to function as an antenna simultaneously. Several previous studies have provided simultaneous material characterization solutions to determine the permittivity and thickness using microstrip line ring resonators (Lim, Kim and Hong, 2018), interdigital capacitors (Yeo and Lee, 2019a), and microstrip SRR structures (Lee and Yang, 2014).

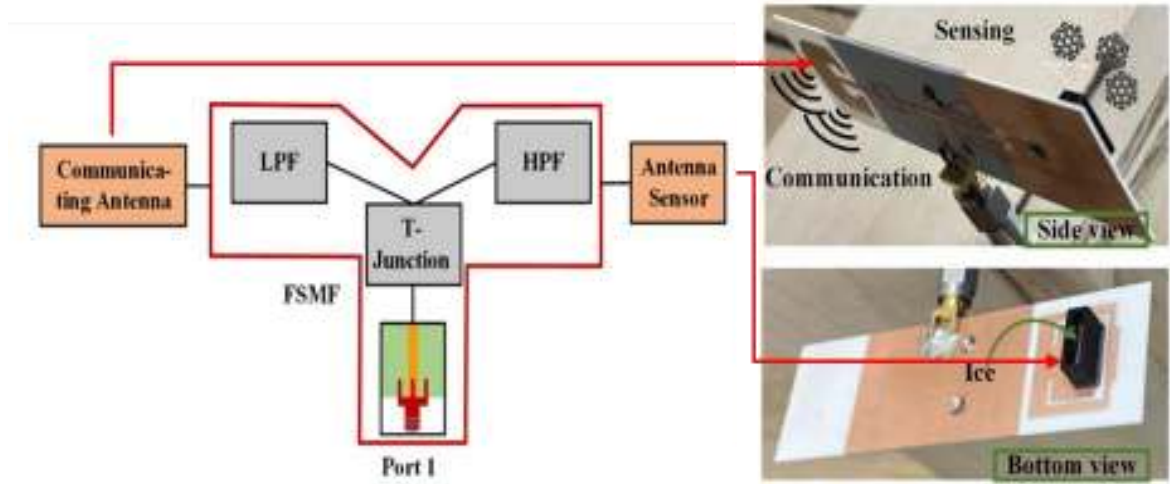


Figure 2.13 Sensing and communication plane of the dual-functional antenna system (Alam, Cheffena and Rajo-Iglesias, 2022)

Furthermore, the detection of the permittivity of solid materials was also proposed using LC resonators (Ebrahimi, Scott and Ghorbani, 2018b) and CSRR (Ansari, Jha and Akhtar, 2015) using a dual resonant frequency approach in a single sensor structure to differentiate the permittivity of the material under test and the reference material (Piekarz et al., 2017). However, the previously proposed permittivity sensor still used a resonator with two ports so it was not possible to function as an antenna and a sensor simultaneously. To

produce a permittivity sensor that can also function as an antenna, a single port resonator can be used as a solution. Previous research presented in Yeo and Lee (2019b) proposed the antenna as a permittivity sensor for solid materials operating at a resonant frequency of 2.5 GHz. However, the research only produced a resonator that worked at a single frequency and had one sensing area, so it was not possible to determine the permittivity of two different types of MUTs simultaneously.

In addition, research conducted by Behdani et al. (2020) proposed a hybrid function of a dual-band resonator with an antenna where the highest resonant frequency was used as a sensor to detect the permittivity of solid materials and low frequency was used as an antenna for local transmission data. However, the previously proposed sensor only had one sensing area location so it could not be used to determine the permittivity of two types of MUTs simultaneously. Furthermore, the sensor and antenna were coupled to each other so that the structure was more complex, and the fabrication was complicated. Besides, the antenna and sensor were not independent of each other. Therefore, to produce a sensor that can determine permittivity simultaneously, it is necessary to have a sensor structure that operates at several different resonant frequencies, has different sensing areas, and has independent characteristics.

Based on literature review, typical application demanding wireless readout systems would be for a simultaneous communication-and-sensing unit for chip-less RFID sensor tags measuring temperature, humidity, and other characteristics, and communicating with a reader, at the same time (Yeo and Lee, 2019b). Moreover, in Alam and Cheffena (2022) integrated microwave antenna/sensor is proposed for sensin and communication applications as shown in Figure 2.13. The dual-functional ability of the proposed system is achieved by

integrating a two-port microwave sensor and a Wi-Fi antenna with a novel frequency-selective multipath filter (FSMF). The FSMF ensures efficient system operation by not affecting the operational bandwidth of the communicating antenna in the presence of different sensed materials under test (MUTs) on top of the sensor. Furthermore, previous studies presented by Alam, Cheffena and Rajo-Iglesias (2022) proposed dual-functional communication and sensing antenna system for real time detection of ice accretion in a wireless communications system (e.g., remote GPS antennas, base stations, radome of an aircraft, mobile communication systems, radio broadcasting antennas, etc.). The fixed band communicating antenna of the proposed dual-functional system is proposed for Wi-Fi application at 2.45 GHz, and the narrowband antenna sensor is used for ice/water detection.

2.9 Research position

Based on the literature review, recent developments of microwave sensors can be identified based on several criteria including resonant frequency, type of sample, structure of resonator, number of ports, number of sensing areas, independent performance, detection techniques and also multifunctionality as other devices.

The microwave passive planar resonators can be designed in any selective operating frequency for material characterizations as reported in literature where they are operating at different frequencies of 1.16 GHz (Abdolrazzaghi et al., 2021), 2.00 GHz (Ebrahimi, Scott and Ghorbani, 2018b), 3.50 GHz and 8.71 GHz (Haq et al., 2019), 5.35 GHz and 7.99 GHz (Armghan et al., 2021), 9.54 GHz and 12.34 GHz (Behdani et al., 2020). Table 2.3 shows the selection of the resonant frequency for microwave sensors based on previous studies.

Table 2.3 Resonant frequency of microwave sensor based on previous studies

Reference	Resonant frequency	Material
(Abdolrazzaghi et al., 2021)	1.16 GHz	Liquid
(Ebrahimi, Scott and Ghorbani, 2018b)	2.00 GHz	Solid
(Haq et al., 2019)	3.50 GHz 8.71 GHz	Solid
(Armghan et al., 2021)	5.35 GHz 7.99 GHz	Solid
(Behdani et al., 2020)	9.54 GHz 12.34 GHz	Solid

Table 2.3 shows that the resonance frequencies used for microwave sensors are generally in the range 1 - 12 GHz with the types of materials detected including solid and liquid materials. Furthermore, previous research has described several microwave sensors with various sensing area locations based on the topology and structure of the resonator used as the sensor. Table 2.4 shows the topology and location of sensing areas from microwave sensors based on previous research.

Table 2.4 Topology of microwave sensor based on previous studies

Reference	Topology	Number of port	Sensing area location	Function
(Abdolrazzaghi et al., 2021)	Split Ring Resonator (SRR)	2	Gap of resonator	Microwave sensor
(Ebrahimi, Scott and Ghorbani, 2018b)	LC resonator	2	Ground plane of resonator	Microwave sensor
(Haq et al., 2019)	Complementary Metamaterial Resonators	2	Gap of resonator	Microwave sensor
(Armghan et al., 2021)	Complementary Symmetric Split-Ring Resonators	2	Gap of resonator	Microwave sensor
(Behdani et al., 2020)	Multilayer resonator	1	In the middle of resonator	Microwave sensor and antenna

Several previous studies have proposed a microwave sensor for the detection of solid materials with various dimensions and permittivity of MUT as shown in Table 2.4. Table 2.4 shows that generally the sample is placed in the gap of the SRR (Abdolrazzaghi et al., 2021) and CSRR (Haq et al., 2019) (Armghan et al., 2021), whereas for the LC resonator it is placed on the ground plane and for the multilayer resonator it is placed in the middle of the resonator. Moreover, the dimension and permittivity range of samples from previous worked are shown in Table 2.5.

Table 2.5 Dimension and range of permittivity of MUT based on previous studies

Reference	Resonant frequency	Material	Dimension of MUT	Range of permittivity
(Ebrahimi, Scott and Ghorbani, 2018b)	2.00 GHz	Solid	(10 x 10 x 1.6) mm ³	2.2 – 10.2
(Haq et al., 2019)	3.50 GHz 8.71 GHz	Solid	(10 x 10 x 1) mm ³	1 – 16.5
(Armghan et al., 2021)	5.35 GHz 7.99 GHz	Solid	(20 x 15 x 1) mm ³	1 – 12
(Behdani et al., 2020)	9.54 GHz 12.34 GHz	Solid	(5 x 5 x 0.508) mm ³	1 – 12.85

Based on Table 2.5, the dimensions of the solid material used in previous research are symmetrical with a range of 5 mm - 20 mm with a permittivity range of 1 - 12. Furthermore, previous research proposed a microwave sensor using SRR (Ebrahimi, Scott and Ghorbani, 2018a) , CSRR (Ansari, Jha and Akhtar, 2015), CBSRR (Haq et al., 2019), SSRR with spur line (Rammah Ali Alahnomi et al., 2017), CSRRs (Armghan et al., 2021) and interdigital capacitors (Wang et al., 2021) to characterize solid materials in the permittivity range 1 - 12.9. The sample is placed in the gap of the resonator which is the sensing area with the highest electric field. However, the sensor only detect single materials

so it cannot be used to independently detect two different types of samples. In addition, the structure of the proposed sensor uses two ports, making it impossible to function as an antenna. Furthermore, other studies have proposed a microwave sensor using a magnetic LC resonator (Ebrahimi, Scott and Ghorbani, 2020a) and microstrip line ring resonator (Lim, Kim and Hong, 2018) to characterize solid materials in the range permittivity of 1 - 10.7. The proposed sensor operates in multiband. However, each sensing area based on resonance frequency is not independent from each other, so it cannot be used to characterize two different types of material simultaneously.

Another study describes a microwave sensor for independent characterization of liquid materials and solid materials using a dual active SRR (Abdolrazzaghi and Daneshmand, 2017) and LC resonator (Ebrahimi, Scott and Ghorbani, 2018b) with a two different sensing area. However, its structure uses two ports and dedicated for microwave sensor only so it cannot be proved as an antenna. In addition, the sample is placed directly so that it can be corrosive and damage the surface of the resonator. On the other hand, a microwave sensor with non-contact detection based on SC-TCSRR (Yang et al., 2016) is proposed to detect the permittivity and thickness of solid materials. The sample is placed in the sensing area of the sensor and separated by 2 mm air gap. However, this study is dedicated only as non-contact sensor and cannot be proved as an antenna.

Furthermore, a multifunctional microwave sensor as an antenna has been described to characterize ice materials (Alam, Cheffena and Rajo-Iglesias, 2022). The sensor and antenna are separated by using FMSF where the low frequency is used as the antenna and the high frequency is used as the sensor. Moreover, the previous work (Behdani et al., 2020), proposed multilayer resonator for microwave sensor and antenna for data transmission.

However, the limitation of this work is that the sensor only has one sensing area location so it cannot be used to characterize two different types of material simultaneously.

Overall, the research position of microwave sensors in terms of the type of sample, permittivity range, independent characteristic, multifunctional performance and design complexity is shown in Table 2.6.

Table 2.6 Research position based on previous works

Ref/Year	Method	Freq. (GHz)	Material	Range of ϵ_r	Num. of Sensing Area /Port	Independent Performance	Simultaneous Measurement	Integrated with Antenna	Contact and non-contact	Design complexity
(Armghan et al., 2021)	CSRRs	5.35 7.99	Solid	1 – 11.9	1 / 2	No	No	No	No	Moderate
(Haq et al., 2019)	CBS-SRR	4.50 6.79	Solid	1 – 12.9	1 / 2	No	No	No	No	Moderate
(Al-Behadili et al., 2020)	SRR	1.24 2.08	Solid	1 – 6	1 / 2	No	No	No	No	Moderate
(Ebrahimi, Scott and Ghorbani, 2020a)	Magnetic LC	1.62 2.38	Solid	1 – 10.7	1 / 2	No	No	No	No	Moderate
(Lim, Kim and Hong, 2018)	Microstrip Line Ring Resonator	4.89 9.81	Solid	1 – 3.48	1 / 2	No	No	No	No	Moderate
(Rammah Ali Alahnomi et al., 2017)	SSRR with spur lines	2.20	Solid	1 – 4.4	1 / 2	No	No	No	No	Moderate
(Ansari, Jha and Akhtar, 2015)	CSRR	2.65	Solid	1 – 3	1 / 2	No	No	No	No	Moderate
(Ebrahimi, Scott and Ghorbani, 2018a)	SRR	2.10	Solid	1 – 10.7	1 / 2	No	No	No	No	High
(Ebrahimi and Ghorbani, 2020)	CSRR	2.48	Liquid	1 – 80	1 / 2	No	No	No	No	Moderate

Ref/Year	Method	Freq. (GHz)	Material	Range of ϵ_r	Num. of Sensing Area /Port	Independent Performance	Simultaneous Measurement	Integrated with Antenna	Contact and non-contact	Design complexity
(Ebrahimi, Scott and Ghorbani, 2018b)	LC Resonator	2.41	Solid	1 – 10.2	2 / 2	No	No	No	No	High
(Behdani et al., 2020)	Aperture coupling	9.54 12.30	Solid	1 – 12.85	1 / 1	No	No	Yes	No	High
(Abdolrazzaghi and Daneshmand, 2017)	Dual Active SRR	1.03 1.49	Liquid	1 – 20	2 / 2	Yes	No	No	No	High
(Yeo and Lee, 2019b)	Slot loaded	2.50	Solid	1 – 10.2	1 / 1	No	No	No	No	Moderate
(Wang et al., 2021)	Interdigital capacitors	1.50	Solid	1 – 10.2	1 / 2	No	No	No	No	High
(Piekarz et al., 2017)	Couple line section	1.50	Solid	1 – 10.2	1 / 2	No	No	No	No	High
(Yang et al., 2016)	SC-TCSRR	1.40 2.60 4.30	Solid	1 – 3.65	1 / 2	No	Yes	No	Yes	High
(Alam, Cheffena and Rajo-Iglesias, 2022)	FSMF	2.45 4.90	Solid	1 – 6.15	1 / 1	Yes	No	Yes	No	High

Based on Table 2.3, previous studies of multiband microwave sensors showed high dependency performances of each band. Therefore, the development of resonators with different and independent sensing areas is needed to detect the permittivity of the MUT independently. It can also be a solution to detect permittivity of different MUT with high accuracy and reduce the detection time for large quantities of MUT. Furthermore, the existing sensor is only used to determine the permittivity of solid and liquid materials. Thus, a multifunction sensor with an antenna is needed as a solution so that the sensor can be integrated with devices connected to the internet. Moreover, material characterization using MS is carried out by placing the MUT in direct contact so that it can reduce the performance of the sensor. This caused the surface of the sensor to potentially be damaged due to oxidation and corrosion. To overcome this limitation, non-contact detection is needed as a solution to make the sensor more durable, accurate, and sterile.

2.10 Concluding remarks

This chapter has comprehensively explained the fundamentals and theory for material characterization using microwave sensors. Several techniques and methods have been proposed, including resonant and non-resonant such as waveguide, planar resonator and transmission line model. The characterization of the material can be determined using the perturbation theory whereby the introduced sample changes the characteristics of the electric and magnetic fields of the resonator. Furthermore, a microwave sensor with a planar resonator is recommended because it has several advantages such as low cost, high accuracy and high capability to be integrated with other devices. Based on a literature review,

microwave sensors are generally only used to characterize one type of sample due to the limited location of sensing areas. In addition, the functionality is also limited only as a sensor so that it cannot be used for other needs such as antennas or filters. Therefore, multifunctional sensors are needed to integrate sensors with other devices such as Wi-Fi, Zigbee and internet-of-things (IoT) so that they support measurements in real time conditions.

CHAPTER 3

METHODOLOGY

This chapter presents the development of our proposed microwave sensor based on a single port planar resonator. The discussion in this chapter begins with an introduction and a flowchart of the research project that describes the stages, methods, and design of the proposed microwave sensor. In this research project, the proposed microwave sensor is based on a single-port resonator consisting of three different structures: dual U-shaped resonator, dual T-shaped resonator and multicascoded T-shaped resonator. The discussion in this chapter is continued by determining the design specifications of the proposed sensor, basic principle of resonator, simulation tools, material preparation, accuracy and sensitivity of proposed sensor. The final discussion in this chapter is a measurement setup to validate the performance of the proposed sensor which is represented with accuracy and sensitivity and compared with the existing work.

3.1 Introduction

In this study, the planar method was used to detect the permittivity of the MUT by placing the sample on the surface of the resonator which has a high concentration of electrical fields. The sample used is a solid material whose permittivity is known with predetermined dimensions. The parameters observed were the resonant frequency and reflection coefficient of the resonator before and after the MUT was loaded. The focus of this research is to produce a microwave sensor that has independent characteristics with different sensing area locations so that it can simultaneously detect the permittivity of two

different types of samples. Furthermore, the independent characteristics are used to integrate sensors and antennas for contact and non-contact detection and are also used as antennas for data transfer capabilities.

3.2 Flowchart of this project

The development of the proposed microwave sensor begins with finding research gaps by conducting literature studies related to microwave sensors for the detection of solid and liquid materials. The observed performance of the sensor includes the sensitivity, accuracy and application of the microwave sensor. Furthermore, modelling, simulation and optimization of the proposed sensor is carried out using HFSS 15.0. The main target of this research is to produce a microwave sensor that has independent characteristics so that it can be used to detect two types of MUT with different permittivities. The proposed sensor structure uses a resonator with a single port so that it is possible to function as an antenna. In this study, the proposed sensor was designed using a substrate type FR-4 with a permittivity of 4.3, a loss tan of 0.0265 and a thickness (h) of 1.6 mm. The proposed sensor has multiband characteristics with different sensing areas so that it has the possibility to simultaneously detect two types of MUTs with different permittivity. Sensing areas are determined based on the resonant frequency and the surface of the resonator which has the highest electric field concentration. The permittivity of the MUT is determined by observing the shift in frequency when the MUT is placed on the surface of the sensing area which is modelled using a polynomial equation. In this study, four types of solid materials were proposed, including RO5880, RO4003C, FR-4 and RO3010 with a permittivity range of 1 – 6.15. The dimensions of the MUT are adjusted based on the placement and location of the

sensing area. Furthermore, validation is carried out by carrying out the measurement process in the laboratory by comparing the permittivity based on the calculation results with the permittivity from the datasheet. The performance of the sensor is indicated by the sensitivity and accuracy when detecting the permittivity of the MUT. The flowchart of this study is shown in detail in Figure 3.1.

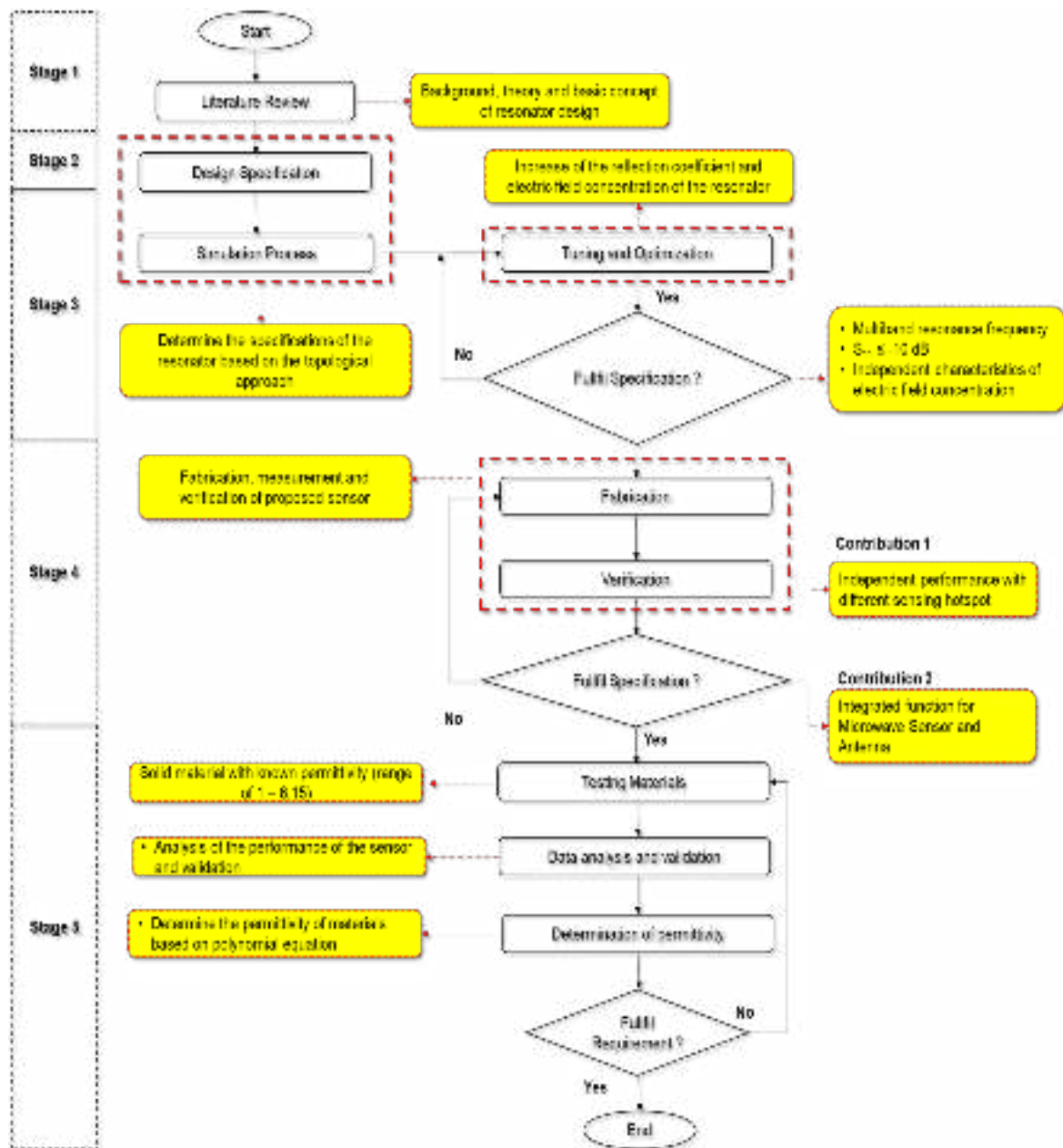


Figure 3.1 Flowchart of the project

In general, this research has two main contributions, including:

- (i) Produce a microwave sensor that has independent characteristics for simultaneous detection of permittivity of solid materials.
- (ii) Produce an integrated microwave sensor and antenna for contact and non-contact detection and data transfer capabilities.

Overall, this research consists of 5 stages including literature review, specification and design of sensor, simulation process, fabrication and verification and in the final stage is data analysis and validation as illustrates in Figure 3.2.

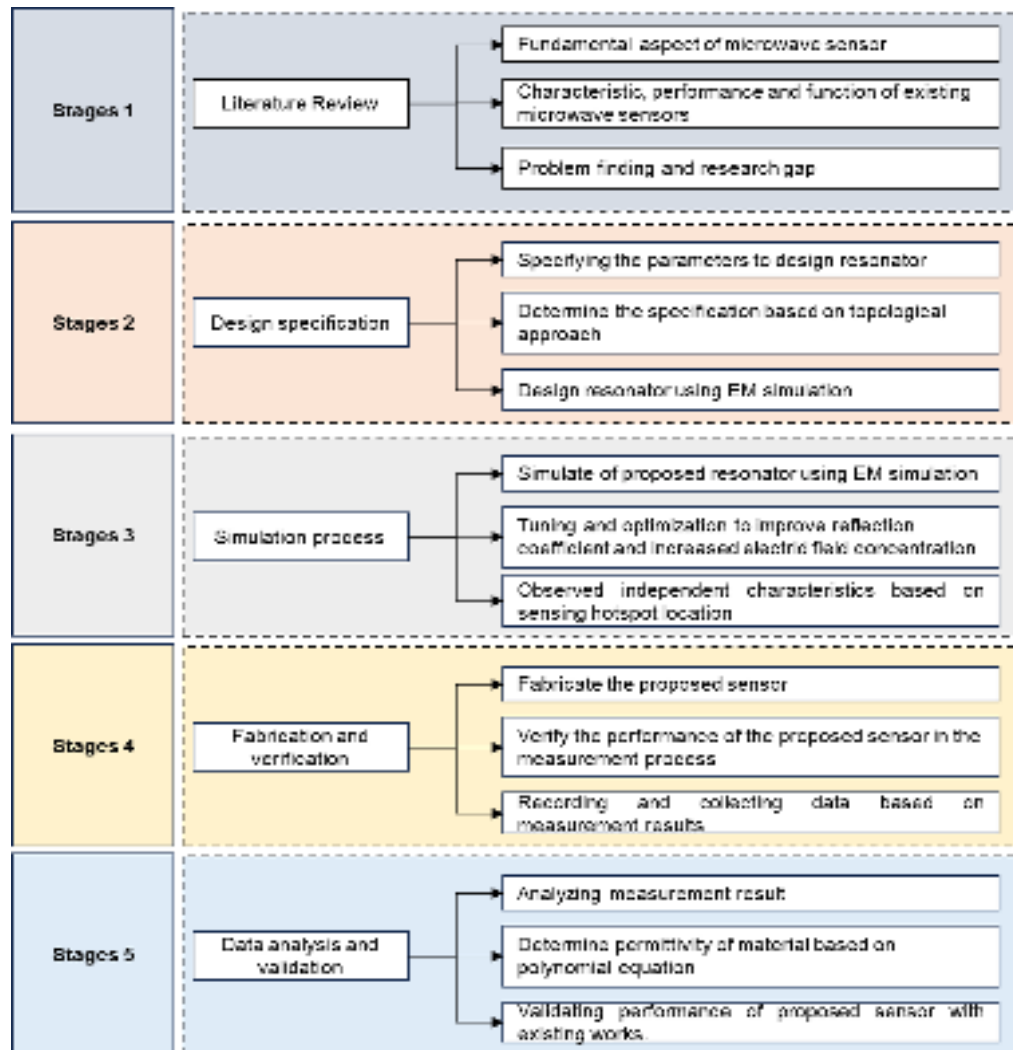


Figure 3.2 Five phase of this project

Stage 1: Literature Review

The first stage of this research project is to explore previous research related to microwave sensors for permittivity detection of solid and liquid materials. The literature review is carried out by observing the performance of existing sensors, especially those with independent characteristics and for simultaneous measurement purposes. Furthermore, observations were also made on the function of the existing sensors. Generally, microwave sensors are only used to determine the characteristics of solid and liquid materials such as permittivity, thickness, and conductivity. Thus, a multifunctional microwave sensor is needed so that it can be integrated with other internet-based devices.

Stage 2: Design specification of sensors

The specifications of the proposed sensor are determined based on the resonant frequency and topology of the resonator to be used. In this study, the resonant frequency used is in the range of 1 – 3 GHz and the proposed resonator topology uses a single port so that it can also function as an antenna. Furthermore, the developed sensor consists of three different topologies, namely dual U-shaped resonator, dual T-shaped resonator and multiscode T-shape resonator. All resonators operate in multiband and have independent characteristics with different sensing areas.

Furthermore, the resonator is connected to the RP-SMA connector via a microstrip line with an impedance of 50 Ω . The parameters observed were the resonant frequency and reflection coefficient before and after the material was placed on the sensing area of the sensor. The Institute of Electrical and Electronics Engineers defined the field sensor as an electrically small device such as passive devices that is used for measuring electric or

magnetic fields with a minimum of perturbation to field being measured (IEEE Standards Association, 2013).

In this study, the resonator was proposed to operate with frequency range of 1 – 3 GHz with single port and impedance of 50 Ω . The overall design specification of proposed resonator is shown in Table 3.1.

Table 3.1 Design specifications of proposed microwave sensor

Specification	Value
Resonant frequency (f)	1 – 3 GHz
Impedance (Z)	50 Ω
Number of ports	1
Accuracy	$\geq 90\%$
Normalized Sensitivity (NS)	0.5%

Stage 3 : Simulation and design of the sensor

Simulation and design of the sensor is carried out using HFSS 15.0 while the equivalent circuit model is using AWR 2009. In the simulation process, the concentration of electric and magnetic fields is observed to determine the sensing area. The location of the resonator that has the highest electric field concentration is used as a sensing area to place the MUT. Furthermore, the surface of the resonator with the highest magnetic field concentration can be recommended as an antenna.

The sensing region of the resonator is an area that has a high concentration of E-field. The electric field will be related to the capacitance while the magnetic field is related to the inductance of the resonator. The illustrations of the E-field and H-field on the microstrip line are shown in Figure 3.3 (Rammah Ali Alahnomi et al., 2017). Figure 3.3 shows that the electric field in the microstrip line was directed outwards (up and down). Meanwhile, the

magnetic field was around the conductor of the resonator. However, the resonator would have a capacitive behavior when the electric field was higher than the magnetic field and vice versa. The electric field stored in the resonator could be used to detect the MUT permittivity. When the MUT is placed on the resonator, the electric field will change so that the resonant frequency shifts (Haq et al., 2019). It can be used for the detection and characterization of materials.

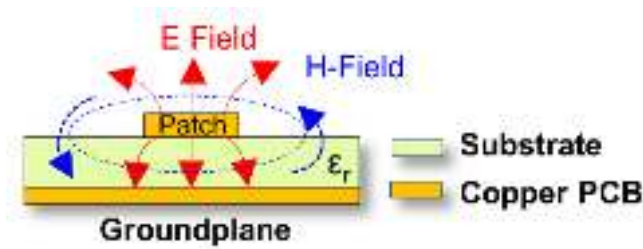


Figure 3.3 Electrical and magnetic field from microstrip line

The performance of the microwave sensor is observed by placing the MUT at the location of the sensing area of the resonator, respectively. The dimensions of the MUT are determined based on the potential location of the sensing area and its availability.

In this project, the dimensions of the MUT used are 10 mm x 10 mm, thickness of 1.60 mm with a permittivity range of 1 – 6.15. To determine the independent characteristics of the sensor, simulations are carried out by placing two types of MUTs which have different permittivity simultaneously in the sensing area of resonator based on resonant frequency, respectively.

Stage 4 : Fabrication and verification

The fabrication of the proposed sensor was carried out using FR-4 substrate with the specifications as already explained in section 3.2. The configuration of the sensor consists

of the top layer of the substrate which functions as a resonator for sensing area location and the bottom layer as the ground plane. To connect the sensor to the Vector Network Analyzer (VNA), an RP-SMA connector with an impedance of 50 ohms is used as the input port as shown in Figure 3.4. The measurement process is carried out by placing the MUT on the sensing area of the proposed sensor.

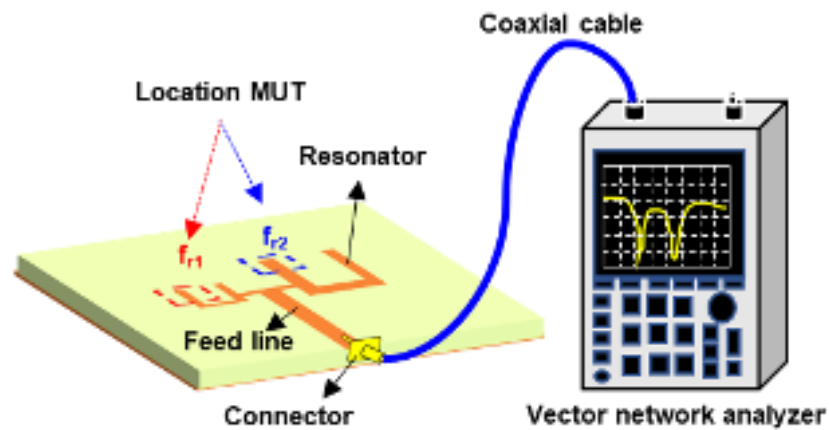


Figure 3.4 Measurement setup of proposed microwave sensor

The type of material under the test used is a solid material with known permittivity. The specifications of the MUT used are shown in Table 3.2.

Table 3.2 Specification and types of MUT

Type of MUT	Parameter		
	Permittivity (ϵ_r)	Thickness (h)	$\tan \delta$
Vacuum	1	-	0
RO5880	2.2	1.6 mm	0.0009
RO4003	3.55	1.6 mm	0.0027
FR4	4.3	1.6 mm	0.0265
RO3006	6.15	1.6 mm	0.0025

Verification of the sensor is observed by placing the MUT at the location of the sensing area. Based on the perturbation theory, the resonant frequency of the sensor will shift when the MUT is placed in line with the change in permittivity. To observe the independent characteristics of the proposed sensor, several scenarios are proposed as follows:

- (1) MUTs having different permittivity are placed on the sensing area of the proposed sensor alternately.
- (2) MUTs having different permittivity are placed on the sensing area of the proposed sensor simultaneously.

Furthermore, the surface of the resonator with the highest magnetic field concentration was proposed as an antenna whose performance was verified by measuring the radiation pattern in the anechoic chamber.

Stage 5: Data analysis and validation

Data analysis was carried out by observing the shift in the resonant frequency when the MUT was placed on the sensing area of the sensor, respectively. The independent performance is observed by observing the correlation of each resonant frequency when the MUTs are placed together. The sensitivity of the sensor is shown by comparing the resonant frequency of the sensor before and after the MUT is placed on the sensing area, respectively.

The validation of the sensor is done by comparing the permittivity obtained from the measurement results with the reference permittivity from the datasheet. Therefore, to get accurate and precise results, the measurement process is repeated 3 times. Moreover, the average value of the measurement process is used as the final result.

Furthermore, to demonstrate the novelty and contribution of the proposed research, a fair comparison was made with the existing sensors based on the literature review. The main contribution of this research is to produce a microwave sensor which has independent characteristics and can be used for simultaneous measurements.

3.3 Design of U-shaped resonator

In this research, the U-shaped resonator is introduced as a microwave sensor. The model development of the U-shaped resonator and the location of the sensing hotspot is described as follows.

3.3.1 Basic concept of U-shaped resonator

U-shaped resonator is a type of microwave resonant structure that is widely used in microwave and radio frequency (RF) circuits. Resonators are devices designed to exhibit a specific resonance frequency, allowing them to selectively transmit or filter signals at that frequency while attenuating others. The U-shape resonator consists of a conductive metal trace or strip formed into the shape of a "U" on a substrate, usually made of high-dielectric constant material like alumina (Al_2O_3) or printed circuit board (PCB) material.

The U-shaped resonator operates based on the principle of electromagnetic resonance. As the current flows along the arms, it generates magnetic fields around them. The two arms of the U-shaped resonator act as open-ended transmission lines. Furthermore, when the signal reflects at the open ends, it combines with the incoming signal, leading to constructive interference. This constructive interference results in the formation of standing waves within the U-shaped structure. This standing wave pattern leads to a strong

accumulation of electromagnetic energy within the U-shaped structure. The resonant frequency of the U-shaped resonator is primarily determined by its physical dimensions, including the lengths of the two arms and the distance between them. Altering these dimensions allows for tuning the resonant frequency to a desired value. At the resonant frequency, the standing waves create regions of high electric and magnetic field strengths between the arms of the U-shaped resonator. As a result, electromagnetic energy becomes trapped and accumulated within the resonator.

The U-shaped resonator's physical dimensions, especially the lengths of the arms and the gap between them, determine the resonant frequency. Consequently, the U-shape resonator exhibits high impedance at the resonant frequency, effectively blocking or attenuating signals that do not match this frequency. This selectivity allows the U-shaped resonator to act as a filter, passing signals at or close to the resonant frequency while rejecting others. The U-shaped resonator operates in a transverse magnetic (TM) mode, specifically TM₀₁₀ mode, where the magnetic field lines are primarily oriented perpendicular to the plane of the U-shaped structure. This mode of operation enables the formation of standing waves and efficient energy storage. Furthermore, the structure of the U-shaped resonator is illustrated in Figure 3.5 (a) and Figure 3.5 (b).

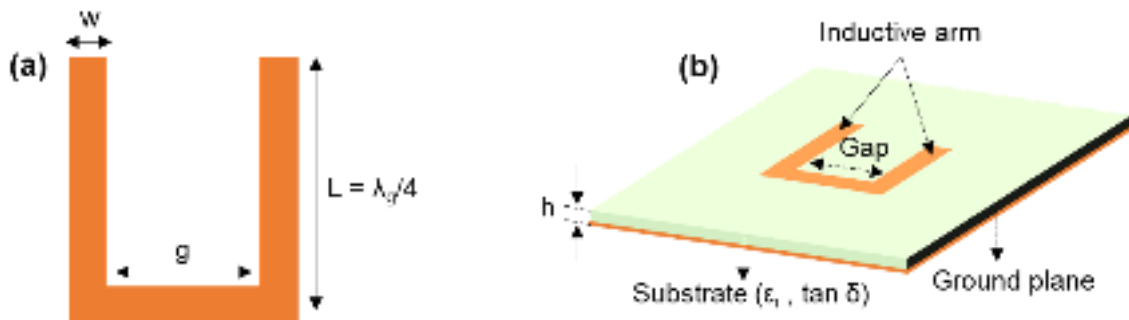


Figure 3.5 Structure of U-shape resonator; (a) design of U-shaped resonator, (b) configuration of U-shaped resonator in perspective view

Figure 3.5 (a) shows the structure of a U-shaped resonator that consists of two inductive metal arms illustrated by **(L)** while the gap between the two arms **(w)** is represented as capacitive. The strip width of the U-shaped resonator is represented by **(g)**. The U-shaped resonator with two inductive arms is on the top layer of the substrate while the bottom layer functions as a ground plane as shown in Figure 3.5 (b). Based on the quarter-wavelength resonant principle, the length of the U-shaped arms **(L)** was calculated according to the resonant frequency and the dielectric constant of the substrate material. Therefore, for different materials, the resonant frequency would shift even if the geometric parameters of the resonator were the same.

Generally, the dimensions of the resonator are determined based on the resonant frequency. Furthermore, the resonant frequency is related to the free space wavelength (λ) and guided wavelength (λ_g) which can be determined using equation 3.1 and equation 3.2.

$$\lambda = \frac{c}{f} \quad (3.1)$$

$$\lambda_g = \frac{\lambda}{\varepsilon_{eff}} \quad (3.2)$$

where the free-space wavelength (λ) is the wavelength of an electromagnetic wave in a vacuum, c is the speed of light in free space ($3 \times 10^8 \text{ m/s}$), f is the resonant frequency, λ_g is the effective wavelength of an electromagnetic wave propagating through a guiding medium, such as a transmission line or waveguide and ε_{eff} is the effective dielectric constant.

The length of U-shape resonator is determined according to equation (3.3) (Tu et al., 2022), where each resonator resonates independently at its quarter-wavelength frequency ($\lambda_g/4$). The effective dielectric constant is calculated by equation (3.4) (Zhang et al., 2023), and it is determined by the dielectric constant and thickness of the substrate as well as the width of the transmission line.

$$f_n = \frac{nc}{4(L+\Delta L)\sqrt{\epsilon_{eff}}} \quad (\text{for } n = 1, 2, 3, \dots \text{ and so on}) \quad (3.3)$$

$$\epsilon_{eff} = \frac{\epsilon_r+1}{2} + \frac{\epsilon_r-1}{2} \left(\frac{1}{\sqrt{1+\frac{12h}{w}}} \right) \quad (3.4)$$

$$Z_0 = \frac{60}{\sqrt{\epsilon_{eff}}} \ln \left(\frac{8h}{w} + \frac{w}{4h} \right) \quad (3.5)$$

$$\Delta L = 0.412h \frac{(\epsilon_{eff}+0.3) \left(\frac{w}{h} + 0.264 \right)}{(\epsilon_{eff}-0.258) \left(\frac{w}{h} - 0.8 \right)} \quad (3.6)$$

where f_n is the resonant frequency, n is the integer order of resonance, ϵ_{eff} is the effective dielectric constant, L is the length of the stub, ΔL is the extended equivalent length due to fringing field, c is the speed of light in free space ($3 \times 10^8 \text{ m/s}$), ϵ_r is the dielectric constant of substrate, h is the thickness of the substrate, and w is the width of the transmission line.

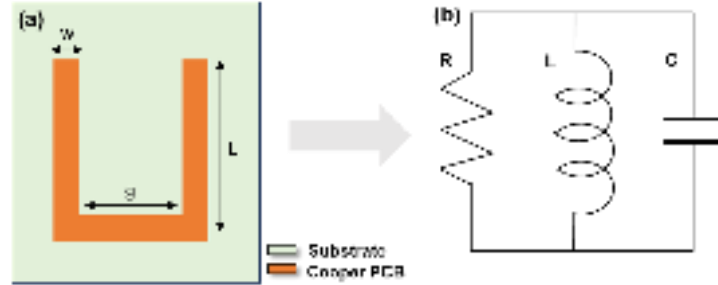


Figure 3.6 U-shaped resonator structure and equivalent circuit

Furthermore, the U-shaped resonator can be modeled with an equivalent circuit as shown in Figure 3.6. The structure of a U-shaped resonator that consists of two inductive metal arms illustrated by (L) while the gap between the two arms (g) is represented as capacitive and resistor represented $\tan \delta$ of the substrate. The strip width of the U-shaped resonator is represented by (w). The equivalent circuit of a U-shaped resonator using parallel RLC can be modeled based on Boutejdar et al. (2011).

Furthermore the capacitance and inductance of resonator can be determined using the following equation (Boutejdar, 2011):

$$jX_{LC} = \frac{j\omega L \left(\frac{1}{j\omega C} \right)}{j\omega L + \frac{1}{j\omega C}} = \frac{j\omega L}{1 - \omega^2 LC} \quad (3.7)$$

$$\omega_0 = \frac{1}{\sqrt{LC}} \quad (3.8)$$

$$X_{LC} = \frac{1}{\omega_0 C \left(\frac{\omega_0}{\omega} - \frac{\omega}{\omega_0} \right)} \quad (3.9)$$

$$L = \frac{Z_0}{\omega_c} \rightarrow \omega_c L = X_L \quad (3.10)$$

$$C = \frac{1}{\omega_0 X_{LC} \left(\frac{\omega_0}{\omega} - \frac{\omega}{\omega_0} \right)} \quad (3.11)$$

where Z_0 is the characteristic impedance, ω_c is the cut off frequency, ω_g is the guided angular frequency, X_{LC} is the total reactance of the resonator, L is the inductance and C is the capacitance. It can be considered to be simple LC circuit with a response frequency f_0 , using the following equation (Zhou et al., 2013; Ansari, Jha and Akhtar, 2015):

$$f_0 = \frac{1}{2\pi\sqrt{LC}} \quad (3.12)$$

where f_0 indicates the resonant frequency of the resonator and π is 3.14159

3.3.2 Development model of dual U-shaped resonator

The development of a dual U-shaped resonator begins with determining the dimensions and parameters of the resonator using the equations described in equations 3.3 until equation 3.6. Furthermore, the two elements of the U-shaped resonator are connected using stub based on microstrip line with an impedance of 100 Ohm and 50 Ohm which functions as a voltage divider and impedance matching between the impedance of the transmission line and the impedance of the resonator. The structure of the dual U-shaped

resonator is shown in Figure 3.7 (a) while Figure 3.7 (b) shows the transmission lines model for the whole resonator.

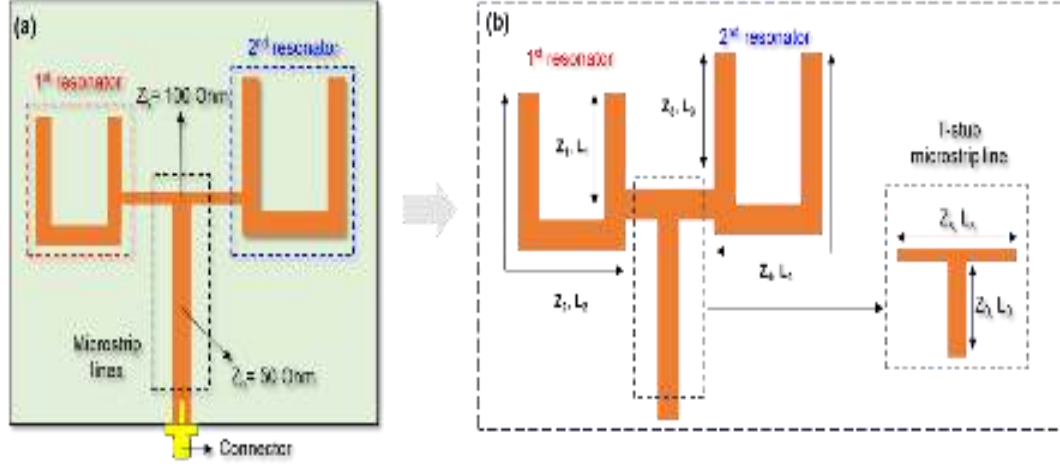


Figure 3.7 Design of dual U-shaped resonator; (a) structure of dual U-shaped resonator, (b) transmission model of dual U-shaped resonator

The impedance of the resonator can be determined using the transmission line approximation (loss line) as shown in equation 3.12 while for β can be determined using equation 3.13 as follows:

$$Z_{in} = Z_0 \left[\frac{Z_L + jZ_0 \tan(\beta L)}{Z_0 + jZ_L \tan(\beta L)} \right] \quad (3.12)$$

$$\beta = \frac{\omega}{\sqrt{LC}} \quad (3.13)$$

while Z_{in} represents input impedance, Z_0 is impedance of resonator, Z_L is the impedance at the output or load end of the resonant structure or transmission line, β refers to the propagation constant of the electromagnetic waves traveling along the transmission line and L represents the length of transmission line.

Moreover, impedance constant (B) and width (W_z) of microstrip line with impedance of 50 Ω can be determined using equation 3.14 and equation 3.15 as follows:

$$B = \frac{377\pi}{2Z_0\sqrt{\epsilon_r}} \quad (3.14)$$

$$W_z = \frac{2}{\pi} \left[B - 1 - \ln(2B - 1) + \frac{\epsilon_r - 1}{2\epsilon_r} \left\{ \ln(B - 1) + 0.39 - \frac{0.61}{\epsilon_r} \right\} \right] \quad (3.15)$$

In the case of an open condition, the transmission line is effectively disconnected or open at its far end. When an open circuit is present at the termination, the input impedance (Z_{in}) will be purely reactive. It means there will be no real power transfer between the source and the load. Moreover, for a lossless transmission line, the input impedance in the open condition will be purely reactive and equal to the characteristic impedance (Z_0) of the transmission line as shown in equation 3.16. Furthermore, for short condition the input impedance (Z_{in}) in the short condition will be purely resistive and equal to the negative of the characteristic impedance (Z_0) of the transmission line as shown in equation 3.17.

$$Z_{in}^{open} = jZ_0 \tan(\beta L) \quad (3.16)$$

$$Z_{in}^{short} = -jZ_0 \cot(\beta L) \quad (3.17)$$

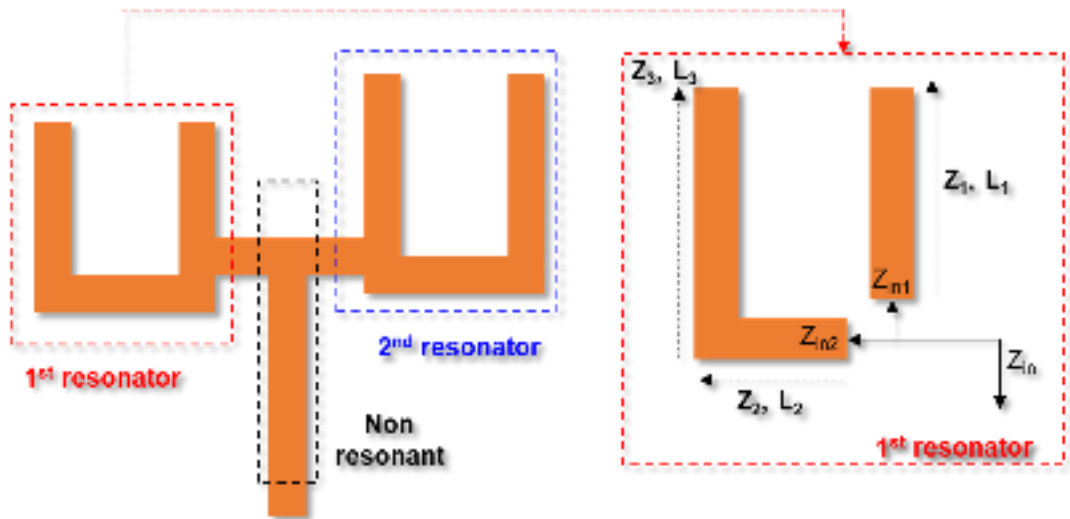


Figure 3.8 Development model of dual U-shaped resonator based on transmission line theory for 1st resonator

Figure 3.8 shows that dual U-shaped resonators can be modeled based on the transmission line approach where each resonator is connected to $\mathbf{Z}_{in}$ using non-resonant T-shaped stubs. It should be noted, according to transmission line theory, $\mathbf{Z}$ represents width of transmission line which correlates with impedance while $\mathbf{L}$ represents the length of the transmission line which correlates with the resonant frequency. Furthermore, the 1st resonator is modeled with two inputs from the transmission line which are represented by $\mathbf{Z}_{in1}$ and $\mathbf{Z}_{in2}$ where the conditions are open in. The value of $\mathbf{Z}_{in1}$ and $\mathbf{Z}_{in2}$ can be determined using equation 3.19 and 3.20 as follows:

$$\mathbf{Z}_{in} = \mathbf{Z}_{in1} + \mathbf{Z}_{in2} \quad (3.18)$$

$$\mathbf{Z}_{in1} = -j\mathbf{Z}_1 \cot(\theta_1) \quad (3.19)$$

$$\mathbf{Z}_{in2} = -j\mathbf{Z}_2 \cot(\theta_2 + \theta_3) \quad (3.20)$$

where $\mathbf{Z}_{in1}$ represent input impedance of 1st transmission line while $\mathbf{Z}_{in2}$ represent input impedance of 2nd and 3rd transmission lines, $\mathbf{Z}_{in}$ represents total input impedance for 1st resonator and θ_1 , θ_2 and θ_3 refers to the electrical length or phase shift of a 1st, 2nd and 3rd transmission line of 1st resonator. The resonant frequencies of resonator can be extracted from admittance condition $Y_{in} = 0$ or impedance condition $Z_{in} = \infty$. Furthermore, admittance of resonator can be determined using equations 3.21 to 3.23 as follows.

$$Y_0 = Y_0 \frac{Y_L + jY_0 \tan \theta}{Y_0 + jY_L \tan \theta} \quad (3.21)$$

$$Y_{in} = Y_0 \frac{Y_L + jY_0 \left(L^{2\pi/\lambda_g} \right)}{Y_0 + jY_L \left(L^{2\pi/\lambda_g} \right)} \quad (3.22)$$

$$\lambda_g = \frac{c}{f\sqrt{\epsilon_{eff}}} \quad (3.23)$$

Based on equation 3.21 to 3.23, the admittance and resonant frequency of the 1st resonator based on transmission line approximation can be determined using equation 3.24 to 3.29 as follows.

$$Y_{in1} = -\frac{Y_1}{j \tan \theta_1} \quad (3.24)$$

$$Y_{in1} = \frac{j}{Z_1} (\tan \theta_1) \quad (3.25)$$

$$f_{r1} = \frac{nc}{2(L_1\sqrt{\epsilon_{eff}})} \quad (3.26)$$

$$Y_{in2} = -\frac{Y_2}{j \tan \theta_2 + \theta_3} \quad (3.27)$$

$$Y_{in2} = \frac{j}{Z_2} \tan(\theta_2 + \theta_3) \quad (3.28)$$

$$f_{r2} = \frac{nc}{2(L_2+L_3)\sqrt{\epsilon_{eff}}} \quad (3.29)$$

Furthermore, the 2nd resonator can also be modeled based on the transmission line approach using the same principle as the 1st resonator as shown in Figure 3.9.

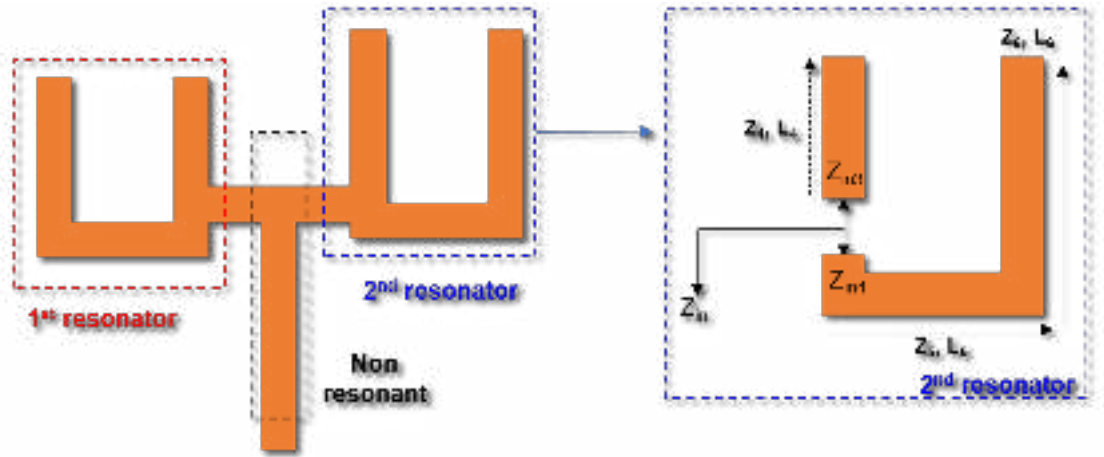


Figure 3.9 Development model of dual U-shaped resonator based on transmission line theory for 2nd resonator

Based on Figure 3.9, using transmission line theory, Z_{in3} and Z_{in4} can be determined using equations 3.30 and 3.32 as follows.

$$Z_{in} = Z_{in3} + Z_{in4} \quad (3.30)$$

$$Z_{in3} = -jZ_3 \cot(\theta_3) \quad (3.31)$$

$$Z_{in2} = -jZ_4 \cot(\theta_4 + \theta_5) \quad (3.32)$$

Referring to equations 3.24 to 3.26, the admittance and resonant frequency of the 2nd resonator can be determined using equations 3.33 to 3.38.

$$Y_{in3} = -\frac{Y_3}{j \tan \theta_3} \quad (3.33)$$

$$Y_{in3} = \frac{j}{Z_3} (\tan \theta_3) \quad (3.34)$$

$$f_{r3} = \frac{nc}{2(L_3)\sqrt{\epsilon_{eff}}} \quad (3.35)$$

$$Y_{in4} = -\frac{Y_2}{j \tan (\theta_2 + \theta_3)} \quad (3.36)$$

$$Y_{in4} = \frac{j}{Z_4} \tan(\theta_4 + \theta_5) \quad (3.37)$$

$$f_{r4} = \frac{nc}{2(L_4 + L_5)\sqrt{\epsilon_{eff}}} \quad (3.38)$$

Furthermore, dual U-shaped resonators also can be modelled using equivalent circuits based on R, L and C as shown in Figure 3.10. The sensor consists of dual U-shaped resonators connected by a microstrip line ($Z_{L1} = 50 \text{ Ohm}$) and a power divider ($Z_{L2} = Z_{L3} = 100 \text{ Ohm}$) for the two resonators. The 1st resonator was indicated by R_{r1} , C_{r2} and L_{r2} while the 2nd resonator was indicated by R_{r2} , C_{r2} and L_{r2} . To prevent a short circuit between port 1 and the two resonators, C_{g1} and C_{g2} are added to the model. Furthermore, the power divider is modeled using a transmission line with an impedance of Z_{L2} and Z_{L3} which is also used as impedance matching between the impedance of the connector that connected with port 1 (Z_0) and the impedance of the resonators.

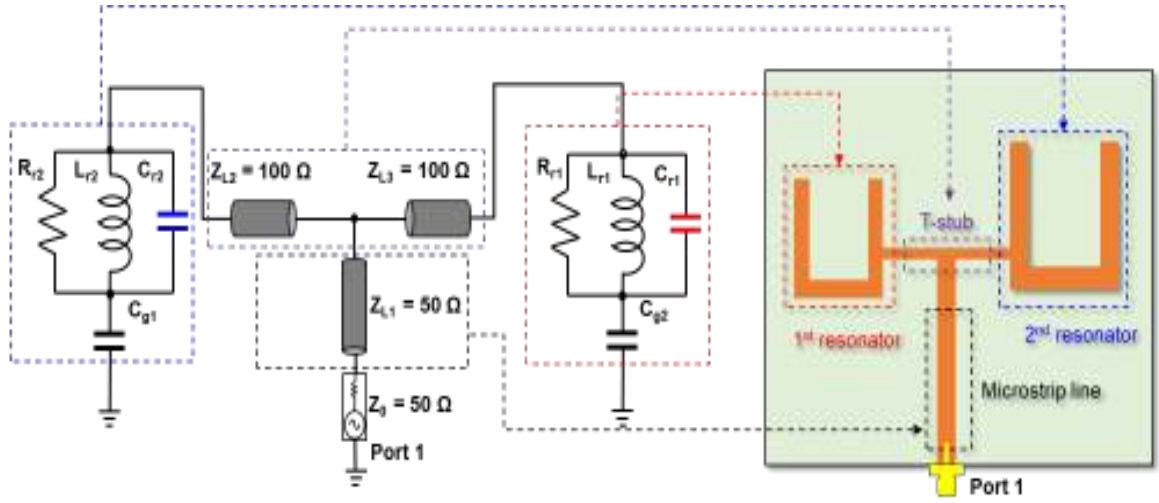


Figure 3.10 Equivalent circuit model of dual U-shaped resonators

The resonant frequency of the resonator can be determined from the values of L_r and C_r while R_r is used to control the value of the reflection coefficient using equation 3.37 as follows (Al-Behadili et al., 2020):

$$f_o = \frac{1}{2\pi \sqrt{L_r C_r}} \quad (3.39)$$

where f_o indicates the resonant frequency of the resonator while L_r and C_r are used as the resonator circuit.

3.3.3 Potential location of sensing area and independent performance of dual U-shaped resonator

The location of the sensing area from the resonator is determined based on the concentration of the electric field at a certain resonant frequency. In this study, the proposed sensor is planned to operate at $f_{r1} = 1.29$ GHz and $f_{r2} = 2.01$ GHz. The location of the sensing area of the U-shape resonator is on the arm of the letter U as shown in Figure 3.11.

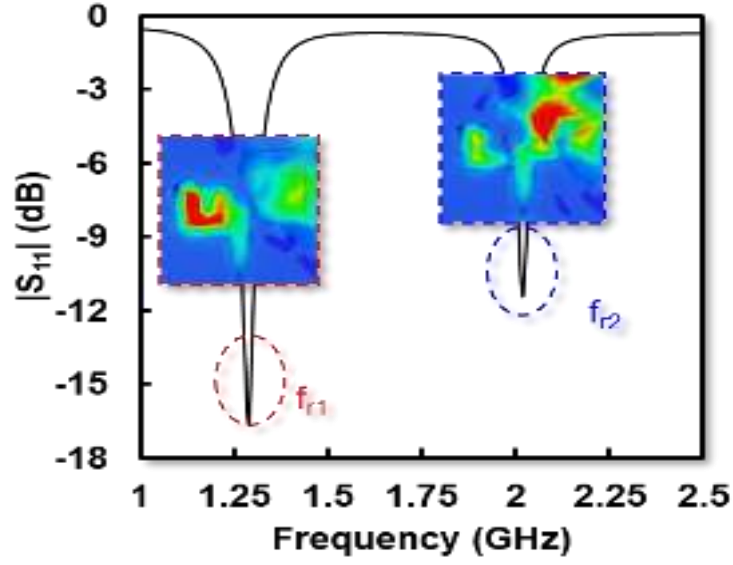


Figure 3.11 Simulation of S_{11} , resonant frequency and sensing area location of dual U-shaped resonator

Figure 3.11 shows that the highest electric field concentrations are obtained when $f_{r1} = 1.29$ GHz and $f_{r2} = 2.01$ GHz with a reflection coefficient $(S_{11}) \leq -10$ dB. This shows that the sensing area represents the resonant frequency of each resonator and is at a different location. This finding can be used to detect the permittivity of two types of MUT simultaneously on each resonator that resonates at different frequencies. Therefore, further observations can be made to observe the independent performance of the sensors by placing MUTs at potential sensing area locations simultaneously.

Furthermore, the independent performance of the dual U-shape resonator can be observed by placing the MUT at a predetermined sensing area location. Furthermore, the independent performance of the sensor is determined by several stages shown in Figure 3.12.

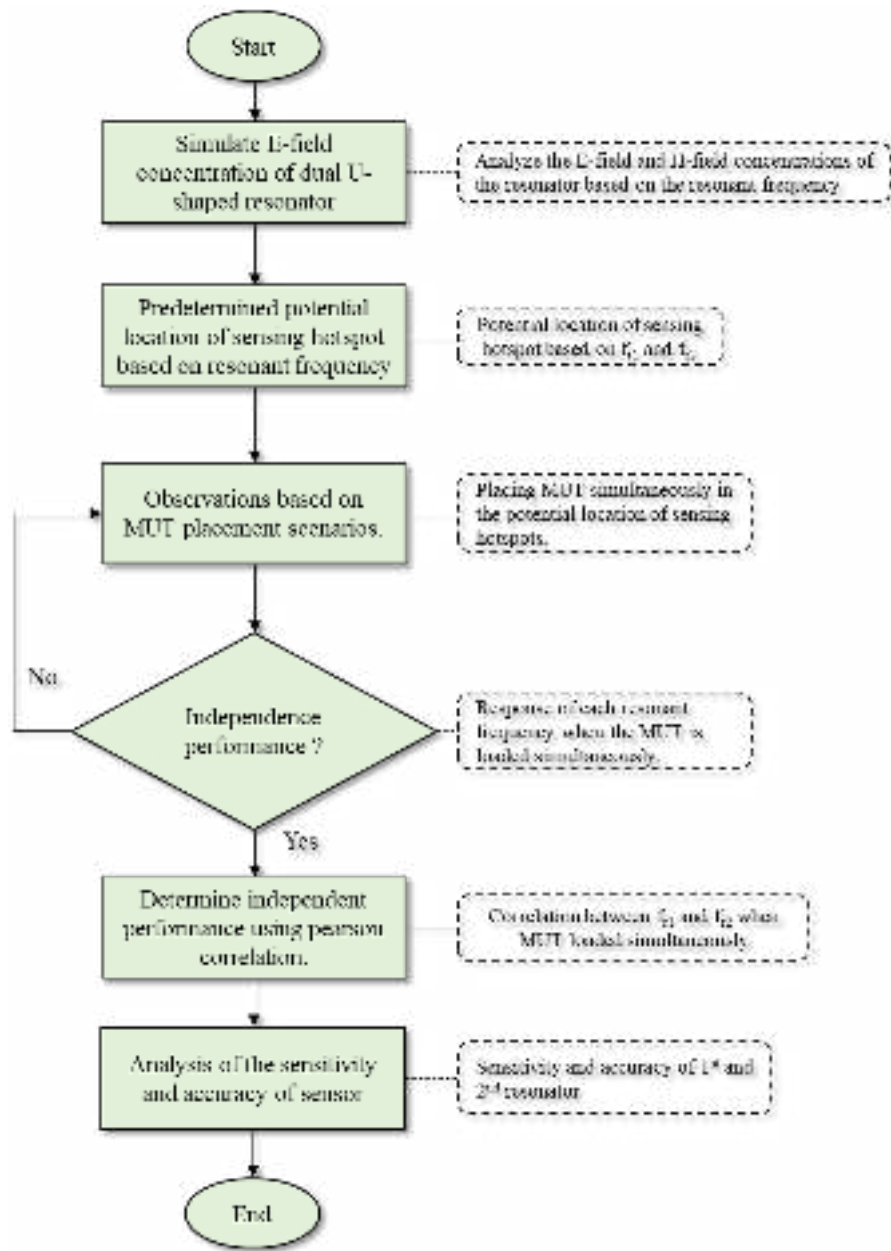


Figure 3.12 Stage to determine the independent performance

In general, each resonant frequency of the resonator will shift to a low frequency when the MUT with a different permittivity is loaded based on the potential location of the sensing area. Based on Figure 3.12, the independent performance of the sensor is determined based on the response of each resonant frequency when the MUT is loaded at a potential

location of a sensing area. To demonstrate the independent performance of the sensors, the following scenario is proposed:

- (i) Sensing area based on f_{r1} in vacuum condition (unloaded) while sensing area based on f_{r2} is loaded with MUT.
- (ii) Sensing area based on f_{r2} in a vacuum condition (unloaded) while sensing area based on f_{r1} is loaded with MUT.

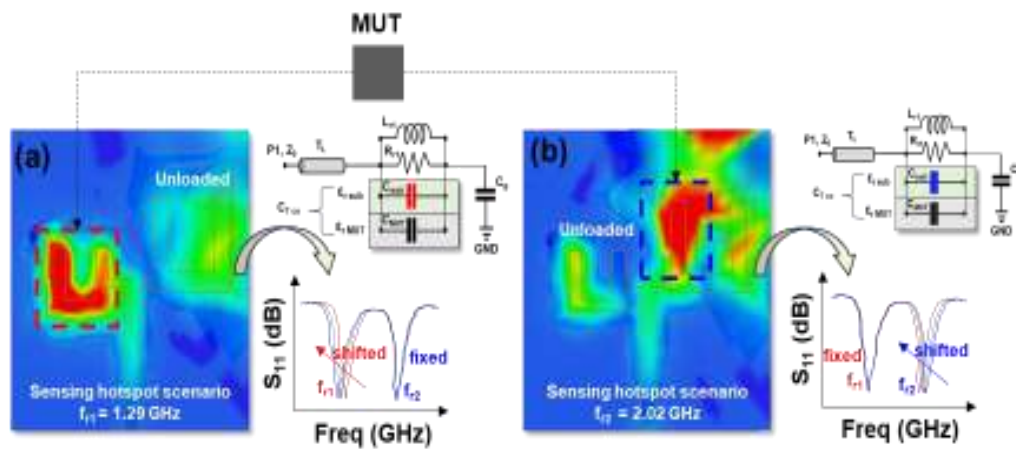


Figure 3.13 Placement scenario of MUT; (a) scenario (i), (b) scenario (ii)

Figure 3.13 illustrates scenario (i) and scenario (ii) for the placement of the MUT on the proposed sensor. The response of each resonant frequency is also described to observe the effect of MUT loading at each location of the sensing area. Scenario (i) shows that f_{r1} shifts to low frequency when MUT is placed at one of the potential locations of the sensing area of the 1st resonator, while the 2nd resonator is in an unloaded condition with f_{r2} is fixed as shown in Figure 3.13 (a). The same findings are also obtained for scenario (ii) as shown in Figure 3.13 (b). Furthermore, the resonant frequency after the sensor is loaded by the MUT (f_{loaded}) can be determined using equation 3.40 as follows:

$$f_{loaded} = \frac{1}{2\pi \sqrt{L_r (C_r + C_{MUT})}} \quad (3.40)$$

where f_{loaded} is the resonant frequency after the MUT is loaded, L_r and C_r are the inductance and capacitance of the resonator while C_{MUT} is the capacitance of the material under test. Applying a dielectric of MUT to the resonators modifies the value of C_r capacitor, which in turn shifts and produce different resonant frequency. This frequency shift is measured and used for characterization of the material under test.

Furthermore, the correlation of f_{r1} and f_{r2} can be analyzed to determine the independence of the proposed sensor based on the potential location of the sensing area when the MUT is loaded simultaneously. In this study, an approximation with the Pearson correlation is proposed to determine the correlation coefficient of the proposed sensor using equation 3.41 as follows:

$$r = \frac{\sum_{n=1} (x_i - \bar{x})(y_i - \bar{y})}{\sqrt{\sum_{n=1} (x_i - \bar{x})^2 \sum_{n=1} (y_i - \bar{y})^2}} \quad (3.41)$$

where r is the correlation coefficient, x_i represents the permittivity of MUT A (located in sensing hotspot of 1st resonator) and MUT B (located in sensing hotspot of 2nd resonator) while y_i represents the resonant frequency of each resonator. $\bar{x}$ and $\bar{y}$ represent the average value of the MUT permittivity and resonant frequency respectively. To determine the level of correlation, the correlation coefficient scale was used which is shown in Table 3.6 (Erdem, Ceyhan and Varli, 2014).

Table 3.3 Scale of correlation coefficient

Scale of Correlation Coefficient (r)	Value
$0 < r \leq 0.19$	Very Low Correlation
$0.2 < r \leq 0.39$	Low Correlation
$0.4 < r \leq 0.59$	Moderate Correlation
$0.6 < r \leq 0.79$	High Correlation
$0.8 < r \leq 1$	Very High Correlation

3.4 Design of T-shaped resonator

T-shaped resonator is typically fabricated on a dielectric substrate, such as a printed circuit board (PCB) or a high-dielectric constant material. It consists of a metallic trace formed into the shape of a "T." The horizontal part of the "T" forms the top of the resonator, while the vertical part extends down from the centre of the top, resembling the stem of the "T." The metallic trace acts as a conductor to carry the electrical signal. The T-shaped resonator operates based on the principle of electromagnetic resonance. When excited with an input signal at its resonant frequency, the resonator generates standing waves and accumulates electromagnetic energy within its structure. This accumulation of energy at the resonant frequency leads to efficient energy storage and influences the behaviour of the signals passing through the resonator. The resonant frequency of the T-shaped resonator is determined by its physical dimensions, such as the length and width of the top part and the length of the vertical stem. By adjusting these dimensions, the resonant frequency can be adjusted to a desired value. T-shaped resonators are designed to have high impedance at their resonant frequency. This characteristic allows them to act as frequency-selective components or filters. They can efficiently filter and pass signals at or near their resonant frequency while attenuating signals at other frequencies. This filtering capability is essential in various communication and electronic systems.

The electric field (E-field) in a T-shaped resonator refers to the distribution of electric charges and the associated electric force within the resonator structure. When the resonator is excited with an input signal at its resonant frequency, an oscillating electric current flows through the resonator's metallic trace. The electric field lines originate from the charged regions and form standing waves within the T-shaped structure. At the resonant frequency,

the electric field exhibits its maximum amplitude, resulting in regions of high electric field strength between the arms of the "T." These regions store and accumulate electromagnetic energy as the electric charges oscillate. The magnetic field (H-field) in a T-shaped resonator corresponds to the distribution of magnetic lines of force resulting from the flow of electric current through the metallic trace. The magnetic field lines form loops around the conductor, creating a magnetic flux that interacts with the electric field and contributes to the overall resonance of the structure. As with the electric field, the magnetic field also exhibits standing wave patterns within the resonator. At the resonant frequency, the magnetic field reaches its maximum amplitude, and regions of high magnetic field strength form between the arms of the "T." These regions store and accumulate electromagnetic energy as the magnetic lines of force interact with the electric field.

The T-resonator is a configuration of microstrip planar which consists of sample open-end T-pattern of transmission lines stub and feedlines. The quarter-half wavelength is developed in the T-resonator for reducing the radiation, improving the accuracy of measurement, and minimizing the circuit size. The open-ended stub is selected as a quarter wavelength at certain operating frequencies which resonates at odd modes with a 50 Ohms characteristic impedance for the microstrip line. The stub length can be approximately modelled with the fundamental equation for a resonator of quarter-wave (Lätti et al., 2007):

$$L = \frac{nc}{4f\sqrt{\epsilon_{eff}}} \quad (3.42)$$

where n is the order of the resonance ($n = 1, 3, 5, \dots$), c is the light speed [m/s], f is the resonance frequency [GHz] and ϵ_{eff} is the effective dielectric constant. If the length of the ribbon is l , the relationship between the resonant frequency and the effective dielectric constant ϵ_{eff} of the microstrip can be calculated from the resonant frequency f as follows:

$$\varepsilon_{eff} = \left(\frac{nc}{2lf} \right)^2 \quad (3.43)$$

The gaps and the coupling to the rest of the lines add a loading that can be represented by an incremental change in the length of the resonant section. Therefore, the effective length of the resonator can be determined as follows:

$$\varepsilon_{eff} = \left(\frac{nc}{2(l+l_c)f} \right)^2 \quad (3.44)$$

$$f = \left(\frac{nc}{2(l+l_c)\sqrt{\varepsilon_{eff}}} \right) \quad (3.45)$$

where l_c is the contribution of the capacitance of the two open ends. It is clear that to obtain accurate value of dielectric constant of the substrate, the effect of l_c should be taken into consideration. Effects of the open-end stub and T-junction have been considered and optimized for accurately obtaining the operating frequencies. Furthermore, the structure of T-shaped resonator which includes an open-ended transmission stub and feed lines as shown in Figure 3.14 (a). The T-shaped resonator can be modeled with an equivalent circuit using a combination of resistor, capacitor, and inductor as shown in Figure 3.14 (b).

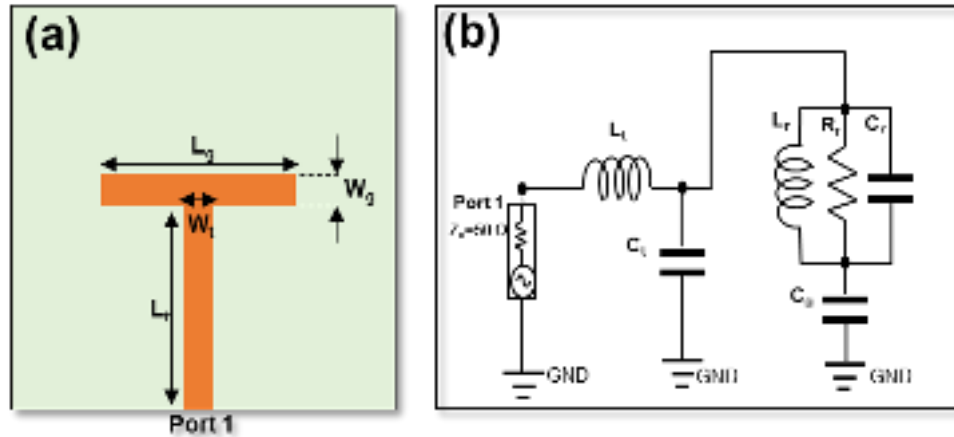


Figure 3.14 (a) Physical structure of T-shaped resonators, (b) equivalent circuit of T-shaped resonators

Figure 3.14 (a) shows the structure of a T-shaped resonator consisting of microstrip lines represented by W_t and L_t and open-ended transmission stubs represented by W_g and L_g (Wu and Tu, 2022). The equivalent circuit model in Figure 3.14 (b) shows that the microstrip line is represented by inductors and capacitors L_t and C_t which function as impedance-matching circuits between the resonator and port 1 while the open-ended transmission stub is represented by a combination of resistors, inductors and capacitors R_r , L_r and C_r which are all connected to ground. The width of the open-ended transmission stub of W_g represents capacitive while the length of L_g represents inductive, and the resistor R_r represents $\tan \delta$ of the substrate.

3.4.1 Development model of Dual T-shaped resonator for contact and non-contact detection

In this study, a dual T-shape resonator was developed as a microwave sensor and antenna sensor for permittivity detection of MUTs both in contact and non-contact. The resonator is planned to operate at two different resonant frequencies, namely $f_{r1} = 2.35$ GHz and $f_{r2} = 1.50$ GHz. The high frequency is used as an antenna sensor for non-contact detection while the low frequency is used as a microwave sensor for contact detection. The stages of developing the model of the resonator are shown in Figure 3.15.

Figure 3.15 (a) shows the first stage where the 1st resonator which consists of a T shape resonator which is directly connected with microstrip lines and connectors which operates at $f_{r1} = 2.35$ GHz. The second stage is shown in Figure 3.15 (b) where the 2nd resonator consisting of microstrip lines and open-ended stubs is connected to the 1st resonator which operates at $f_{r2} = 1.57$ GHz.

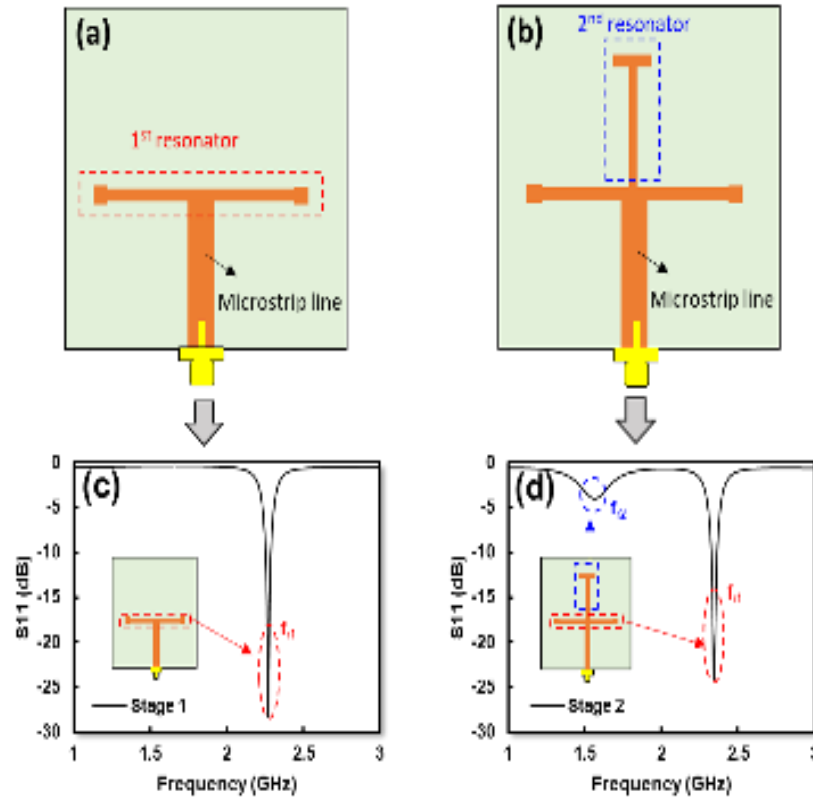


Figure 3.15 Development stages of the dual T-shaped resonator; (a) 1st stages, (b) 2nd stages

Furthermore, the potential location of sensing areas for contact and non-contact detection is determined through observation of the electric and magnetic field concentrations of the resonator based on the resonant frequency, respectively. Figure 3.16 shows the steps for determining the function of the resonator as a microwave sensor and antenna sensor for contact and non-contact detection.

Figure 3.16 shows the stages for determining contact and non-contact detection from the proposed sensor. Contact and non-contact detection is determined based on the characteristics of the resonator which is represented by the concentration of the electric field, magnetic field and reflection coefficient based on the resonant frequency, respectively.

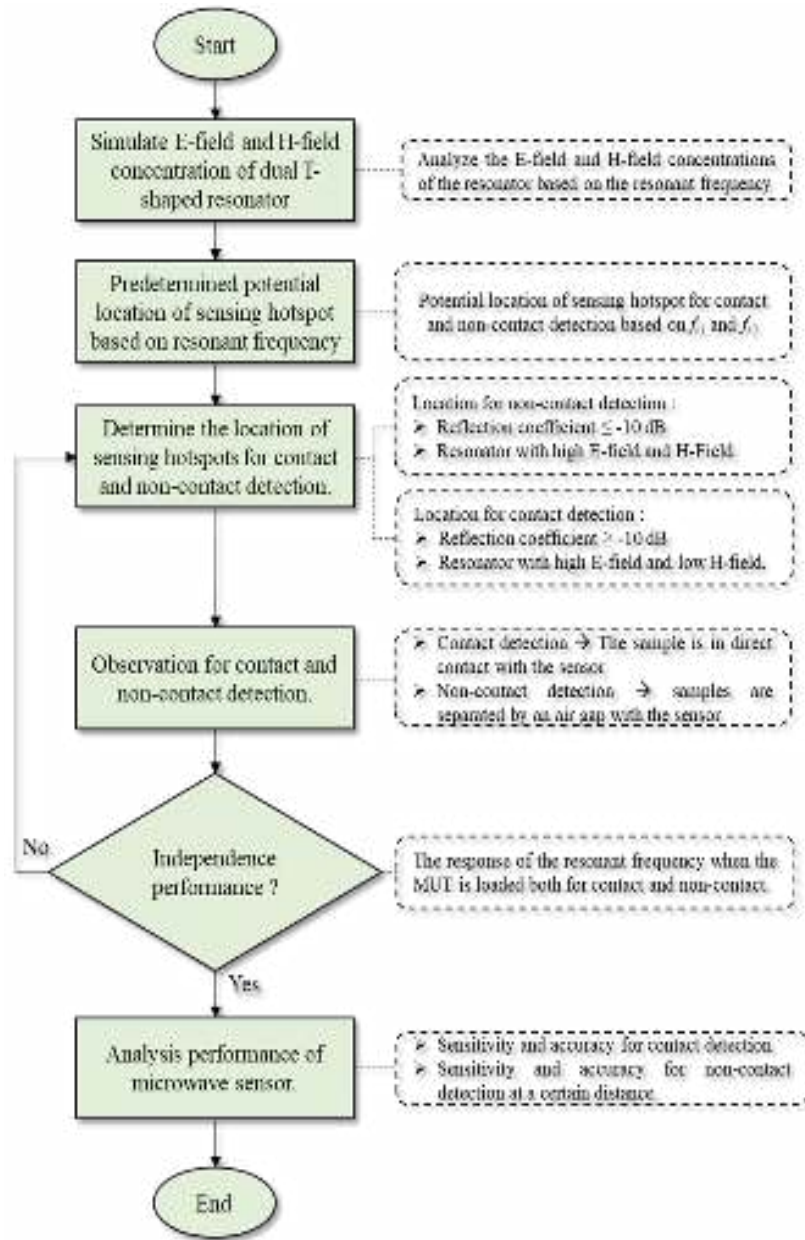


Figure 3.16 Stages for determining contact and non-contact detection of proposed sensor

Resonators with reflection coefficients (S_{11}) ≤ -10 dB are proposed as antenna sensors for non-contact detection while contact detection utilizes resonators with reflection coefficients (S_{11}) ≥ -10 dB. Furthermore, the sensing area is determined based on the concentration of the electric field (E-field) and magnetic field (H-field) of each resonator

based on the resonance frequency, respectively. The surface of the resonator which had a high concentration of E-field and a low concentration of H-field was used as an microwave sensor for contact detection. On the other hand, non-contact detection utilized the surface of the resonator which has the same location as E and H-fields. Generally, resonators that function as antennas transmit electromagnetic waves where the E and H-fields were in the same location, therefore it can be utilized for antenna sensor. Analysis and observations of the electric and magnetic field concentrations of the dual T-shaped resonator will be presented in the next chapter.

Furthermore, for contact detection, the sample is placed in direct contact with the sensor while for non-contact detection, the sensor and sample are separated by an air gap with a certain distance. It should be noted, the gap between the sensor and the sample will affect accuracy and sensitivity. Therefore, further observations are needed to observe the effect of the gap between the sensor and the sample on the performance of the antenna sensor. Figure 3.17 illustrates scenarios of contact and non-contact detection using the proposed sensor.

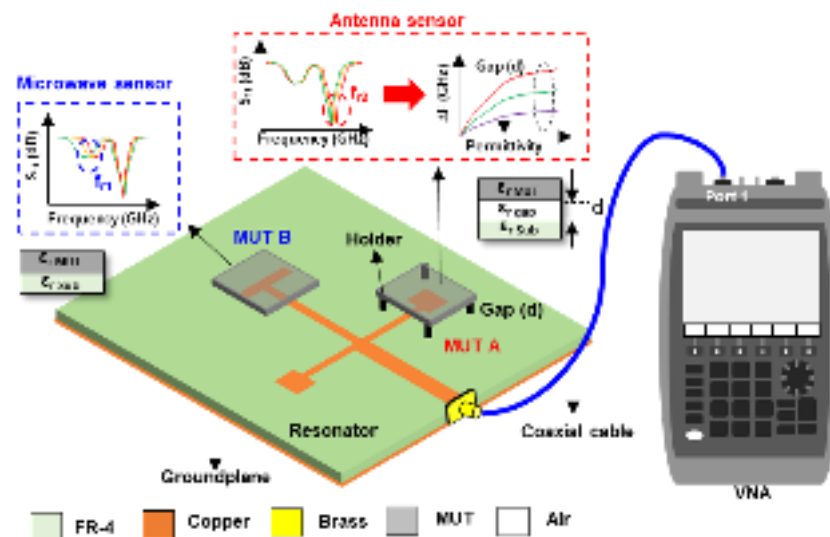


Figure 3.17 Scenario for contact and non-contact detection using dual T-shaped resonator

Figure 3.17 shows that samples for contact and non-contact detection are placed in different locations according to the characteristics of each resonator. The response of each resonant frequency when the sample is placed on the sensing area shows the independent performance of the sensor. Furthermore, the dual T-shaped resonator can also be modelled with an equivalent circuit using R , L and C as shown in Figure 3.18.

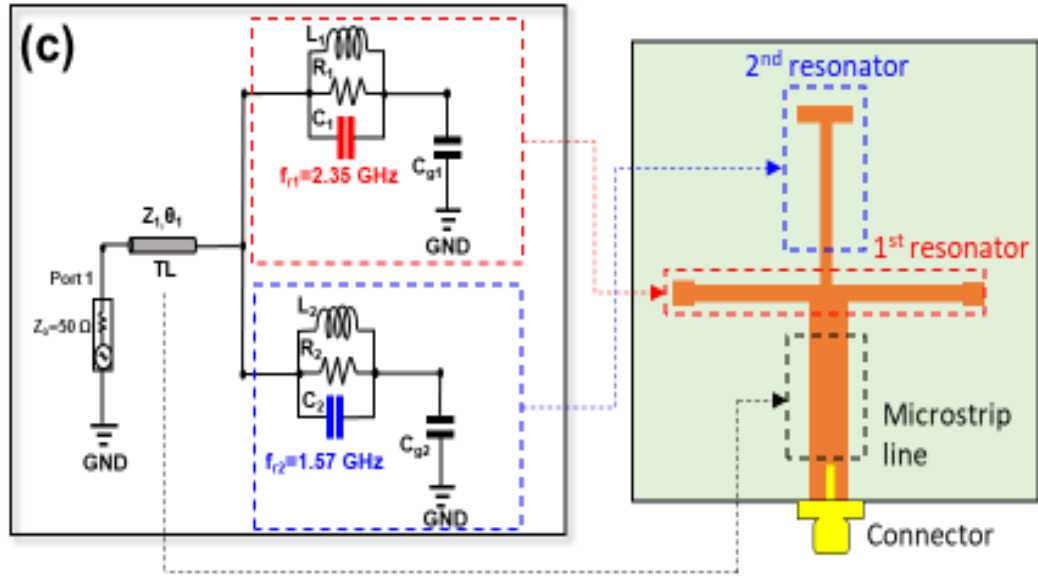


Figure 3.18 Equivalent circuit model of dual T-shaped resonator

The equivalent circuit model using R as resistors, L as inductors, and C as capacitors of the proposed resonator is shown in Figure 3.18 where the 1st resonator is represented by R_1 , R_2 and C_1 while the 2nd resonator is represented by R_2 , L_2 and C_2 . To control the impedance of the resonator, the transmission line was modelled and represented by Z_1 , θ_1 which was connected to port 1 with an impedance of 50 Ω . Furthermore, C_{g1} and C_{g2} were proposed to prevent a short circuit between the resonator and the connector.

3.4.2 Development model of muticascode T-shaped resonator for multifunctional microwave sensor and antenna

A cascode configuration is a common technique used in electronic circuits, especially in amplifiers and other high-frequency applications. In a cascode configuration, two or more resonators are connected in series, with the output of one resonator connected to the input. By using a multistage cascode configuration, the resonator can achieve enhanced performance in terms of gain, bandwidth, and linearity compared to a simple single-stage resonator. Each cascode stage contributes to the overall gain and bandwidth, allowing for better frequency selectivity and performance at the resonant frequency. The specific design and implementation of a multistage cascode resonator can vary depending on the desired resonant frequency, the application, and other design requirements.

In this study, multicascode T-shaped resonator is proposed as a microwave sensor that functions to detect two types of MUT with different permittivities and as an antenna for data transfer capability. The proposed sensor consists of three T-shape resonators configured in series with a multicascode approach operating at $f_{r1} = 2.45$ GHz, $f_{r2} = 2.01$ GHz and $f_{r3} = 1.25$ GHz. The structure of multicascode T-shaped resonator is shown in Figure 3.19.

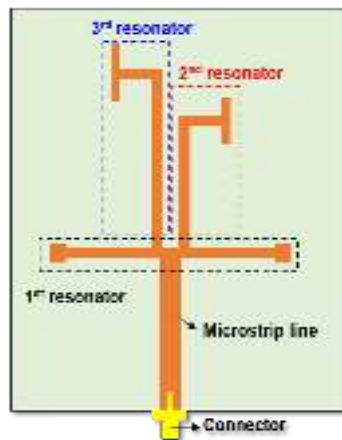


Figure 3.19 Structure of multicascode T-shaped resonator

Figure 3.19 shows that the resonator is arranged with a multicascode configuration and is connected to one input port. The resonator operates at three different frequencies based on the length of the arms of the T-shaped resonator. Furthermore, the development stages of the multicascode T-shaped resonator are shown in Figure 3.20.

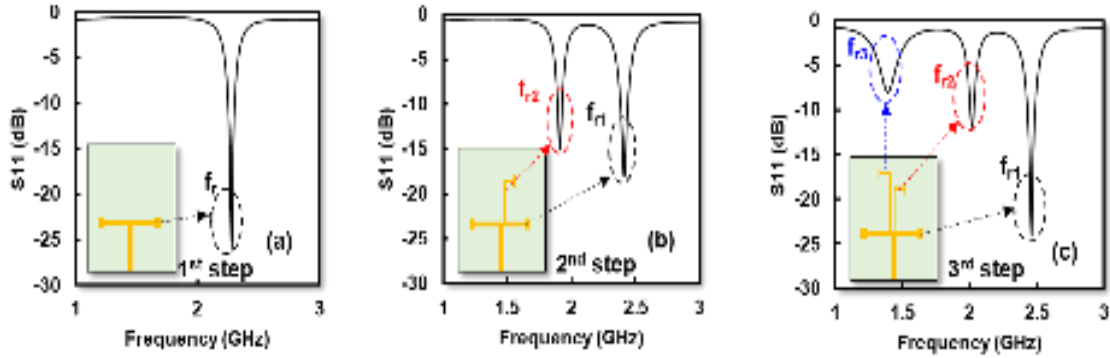


Figure 3.20 Development model of multicascode T-shaped resonator; (a) 1st step, (b) 2nd step, (c) 3rd step

Based on Figure 3.20, the proposed sensor consists of three T-shaped resonators connected via microstrip lines and a connector with an impedance of 50 Ω . The development model of the resonator is shown in Figure 3.20(a), Figure 3.20(b), and Figure 3.20(c). The 1st step model shows a T-shaped resonator design which was connected directly to the microstrip line and connector with an impedance of 50 ohms that operates at only one resonant frequency of 2.46 GHz. In the 2nd step model, a resonator which has an identical shape to the 1st resonator was added to generate a new resonant frequency of 2.01 GHz. Finally, in the 3rd step model, the last resonator was added which functions to generate a new resonant frequency of 1.47 GHz.

Furthermore, to determine the functionality of the proposed resonator as a microwave sensor and antenna, several stages are proposed as shown in Figure 3.21.

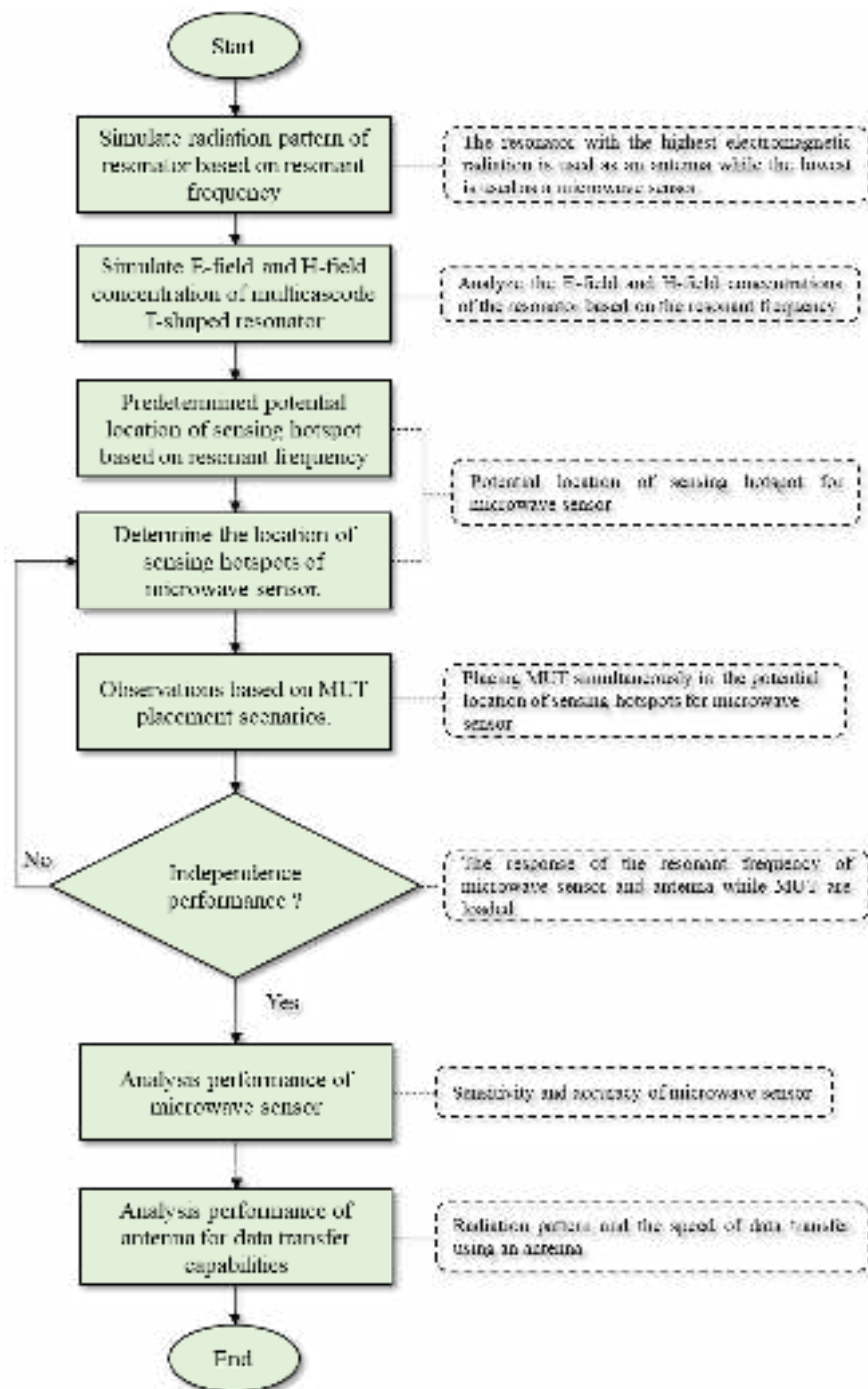


Figure 3.21 Stages to determine the functionality of the resonator as a microwave sensor and antenna

Figure 3.21 shows that the functionality of the resonator as a microwave sensor and antenna is determined based on the resonant frequency and its characteristics. Resonators

that produce high electromagnetic radiation can be used as antennas while those with the lowest radiation are used as microwave sensors. The resonator which produces high electromagnetic radiation can be used as an antenna while the lowest one is used as a microwave sensor. The analysis and observation of the concentration of electric and magnetic fields from the multicascode T-shaped resonator will be presented in the next chapter. Furthermore, the scenario of placing the MUT A and MUT B on the microwave sensor and the functionality of the resonator as an antenna is shown in Figure 3.22.

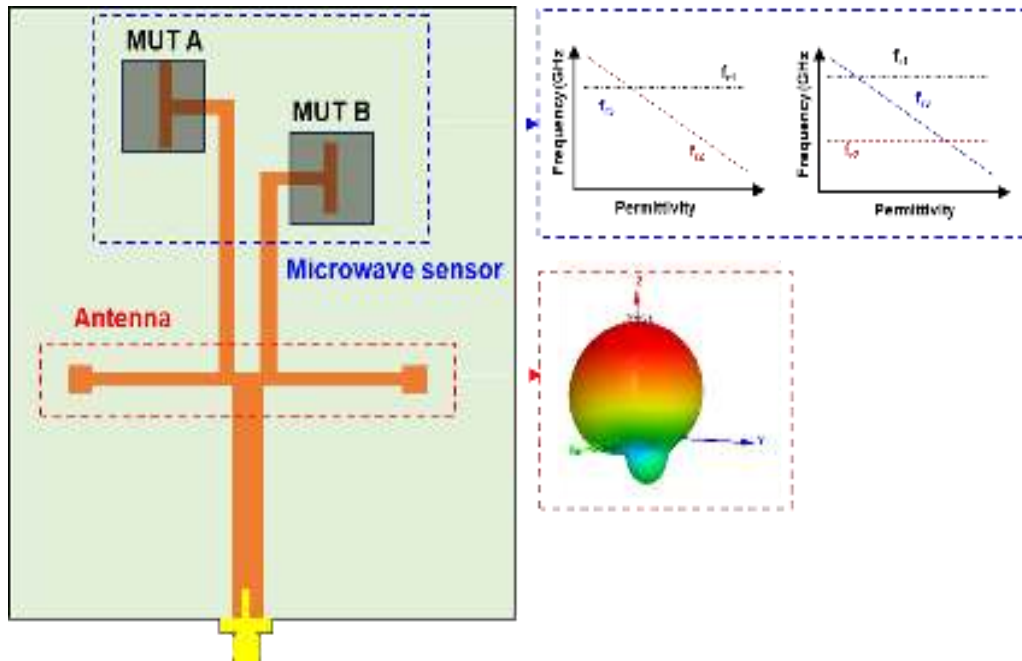


Figure 3.22 MUT placement scenarios on microwave sensors and the functionality of the resonator as an antenna

The multicascode T-shaped resonator can also be modelled using an equivalent circuit based on R, L and C. Figure 3.23 shows the equivalent circuit of a multicascode T-shaped resonator. Furthermore, the development equivalent circuit model of the proposed resonator shown in Figure 3.23 consists of a combination of resistor, capacitor, and inductor.

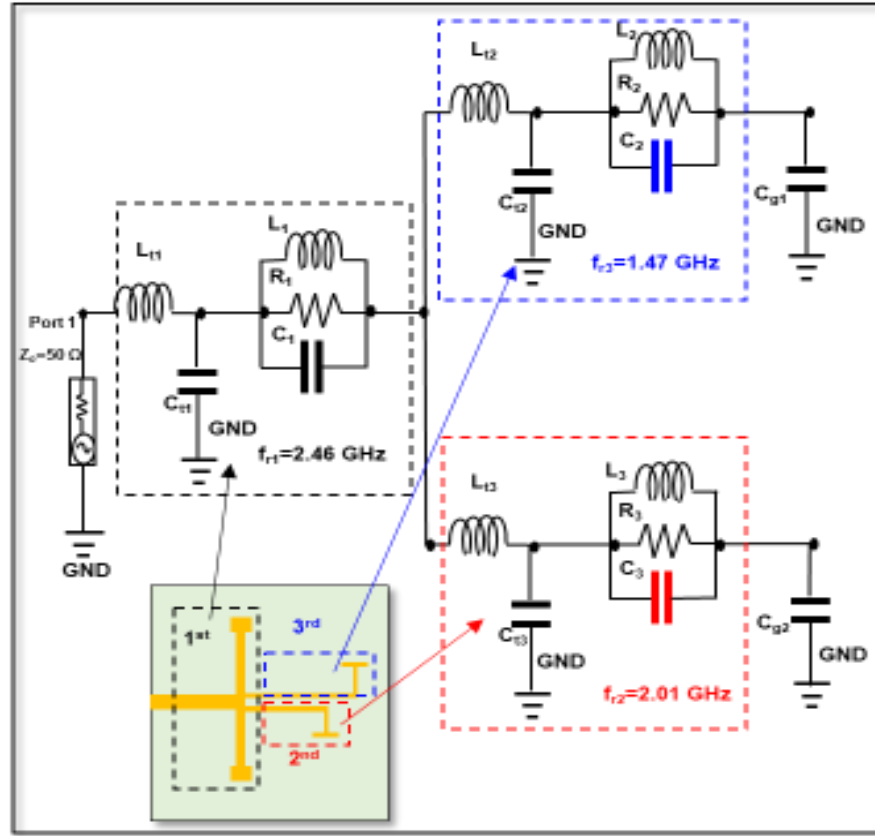


Figure 3.23 Equivalent circuit model of multicascode T-shaped resonator

The 1st resonator represented operates at a resonant frequency of 2.46 GHz by R_1 , L_1 and C_1 while C_{11} and L_{11} are functioned as impedance matching circuits for high frequencies. Furthermore, the 2nd and 3rd resonators operated at resonant frequencies of 2.01 GHz and 1.47 GHz are represented by R_3 , C_3 , L_3 and R_2 , C_2 , L_2 while C_{13} , L_{13} , C_{12} and L_{12} functioned as impedance matching circuits for low frequencies, respectively. Furthermore, C_{g1} and C_{g2} are added to show that the proposed resonator is an open circuit where C_{g1} and C_{g2} show a gap between the resonator and ground plane. The resonant frequency after the sensor is loaded by the MUT (f_{loaded}) can be determined using equation 3.40.

3.5 Simulation tools

Simulation of the resonator was carried out based on finite element method (FEM) and equivalent circuit model (EQC) using HFSS 15.0 and AWR 2009. High-Frequency Structure Simulator (HFSS), and it is a widely used electromagnetic simulation software developed by Ansys, Inc. HFSS is primarily employed for the design, analysis, and optimization of high-frequency electronic components and systems, particularly in the radio frequency (RF), microwave, and millimetre-wave frequency ranges.

HFSS is an electromagnetic simulation tool that uses numerical methods, such as the Finite Element Method (FEM) or the Finite Difference Time Domain (FDTD) method, to solve Maxwell's equations. These equations describe how electromagnetic fields behave and interact with different structures and materials. After creating the 3D model, users define the necessary simulation settings, such as frequency range, boundary conditions, material properties, excitation sources, and desired results (e.g., S-parameters, radiation patterns, impedance, etc.).

HFSS employs advanced numerical algorithms and adaptive meshing techniques to accurately simulate the electromagnetic behaviour of the designed components and structures. This solver technology ensures high accuracy and efficiency, especially for complex geometries and high-frequency applications. Designing a resonator using HFSS (High-Frequency Structure Simulator) typically involves several stages to ensure the desired performance and accuracy. Figure 3.24 show the key stages in the resonator design process using HFSS.

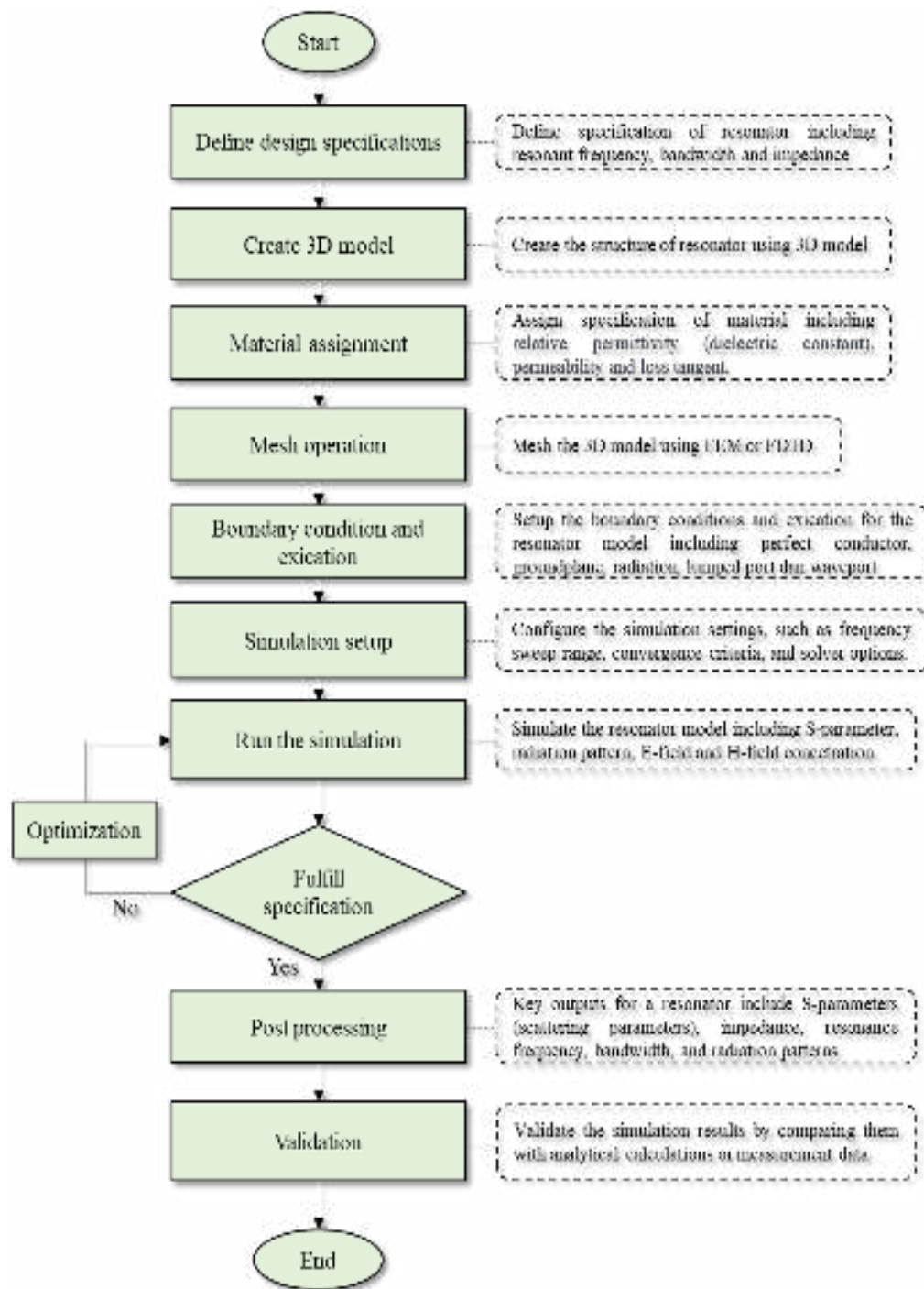


Figure 3.24 The key stages in designing resonator using HFSS 15.0

Furthermore, equivalent circuit of the resonator is modeled using Applied Wave Research (AWR) Design Environment developed by National Instruments (NI). AWR

Design Environment is a comprehensive suite of software tools designed for the design, simulation, and optimization of RF and microwave circuits and systems. It is popular in industries such as telecommunications, aerospace, defense, and consumer electronics. AWR supports circuit simulation using various numerical methods, including linear and nonlinear analysis, harmonic balance, time-domain simulation, and more.

AWR also offers integrated electromagnetic (EM) simulation capabilities through its AXIEM solver. This allows designers to analyze planar structures, such as microstrip and stripline circuits, antennas, and RF packages, while considering the effects of parasitics and electromagnetic coupling. The software provides schematic capture tools that enable engineers to create circuit schematics using drag-and-drop components. Additionally, designers can perform layout and physical design of circuits, ensuring they meet the manufacturing requirements.

An equivalent circuit is a simplified representation of the device or component, which allows to observe its electrical characteristics and performance in different circuits and systems. The resonant frequency of the resonator corresponds to the frequency at which it exhibits maximum response or energy storage. The equivalent circuit captures this resonance behavior using passive elements like inductors, capacitors, and resistors. The equivalent circuit of a resonator typically consists of an inductor and a capacitor, which form a resonant LC circuit. The inductor represents the device's magnetic energy storage, while the capacitor represents the electrical energy storage. In some cases, a resistor (***R***) might also be included to account for the device's energy losses. Figure 3.25 shows key stages to design an equivalent circuit using AWR:

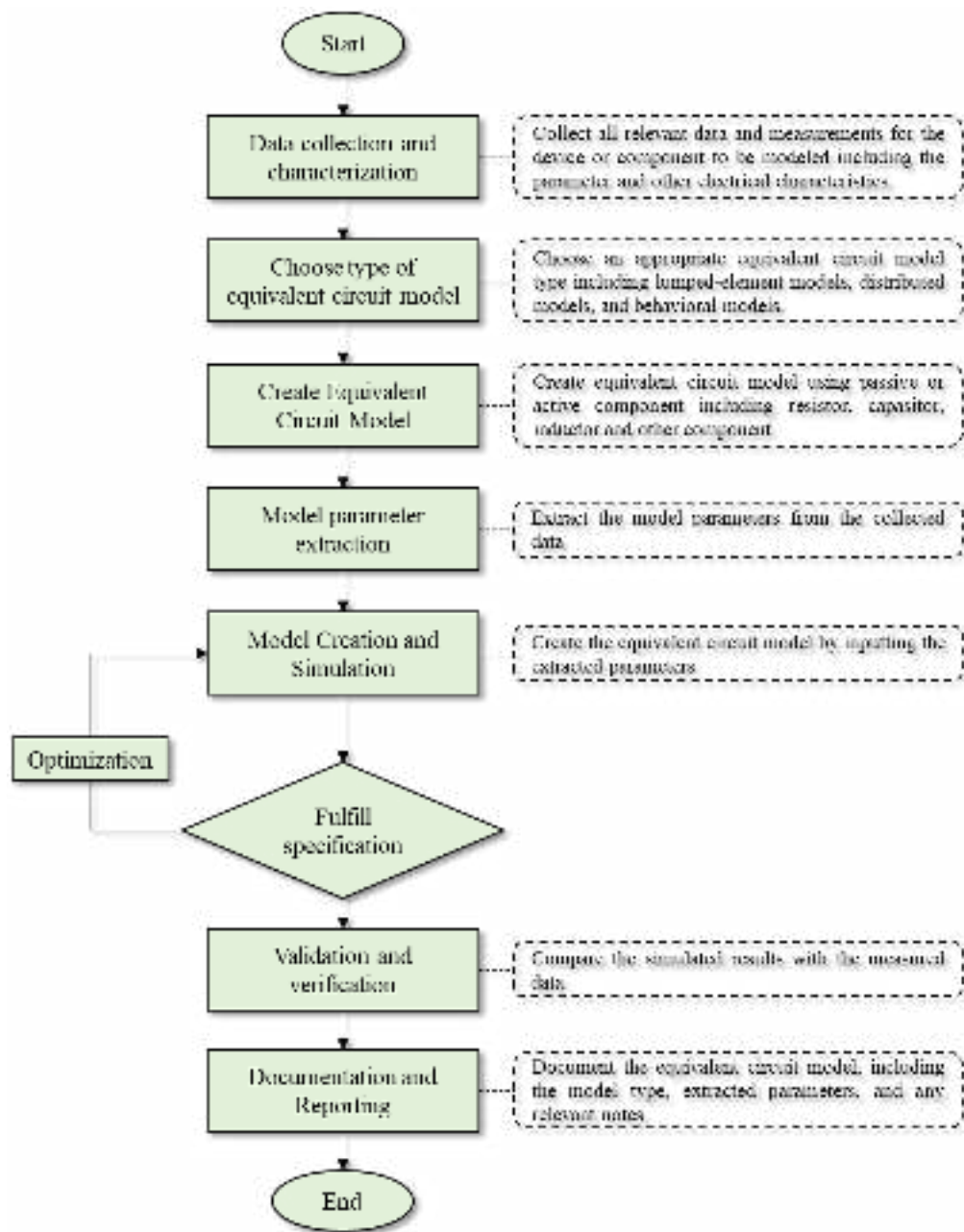


Figure 3.25 The key stages in designing equivalent circuit model using AWR

3.6 Sample preparation

Sample preparation for microwave sensors refers to the process of preparing the samples or specimens that will be analyzed or tested using microwave sensors. Microwave

sensors are devices that utilize electromagnetic waves in the microwave frequency range to detect, measure, and analyze various parameters of the sample. Proper sample preparation is essential to ensure accurate and reliable measurements by the microwave sensor. The characteristics of materials refer to the properties and attributes that define their behaviour and performance in various applications. These characteristics determine how a material responds to external stimuli, such as mechanical forces, temperature changes, electric and magnetic fields, and chemical interactions. Furthermore, Figure 3.26 shows the stages of sample preparation.



Figure 3.26 The stages of sample preparation

In this study, the sample used is a solid material with a known permittivity based on the datasheet. The material characteristic that will be determined using a microwave sensor is permittivity. Permittivity is a characteristic of a material's that shows ability to permit the formation of an electric field within it when subjected to an external electric field. It quantifies how the polarization of the material affects the electric field and how much electric charge can be stored within the material. In the other words, permittivity (ϵ) is a fundamental property of solid materials that describes how they respond to an applied electric field. It is a crucial parameter in understanding the behaviour of materials in electrical and electromagnetic applications. The permittivity of a material determines how much electric flux is allowed to pass through it and how it interacts with an external electric field. Therefore, samples must be prepared according to the characteristics and specifications of the microwave sensor. The dimensions and placement of the sample on the sensor are shown in Figure 3.27 (a) and Figure 3.27 (b).

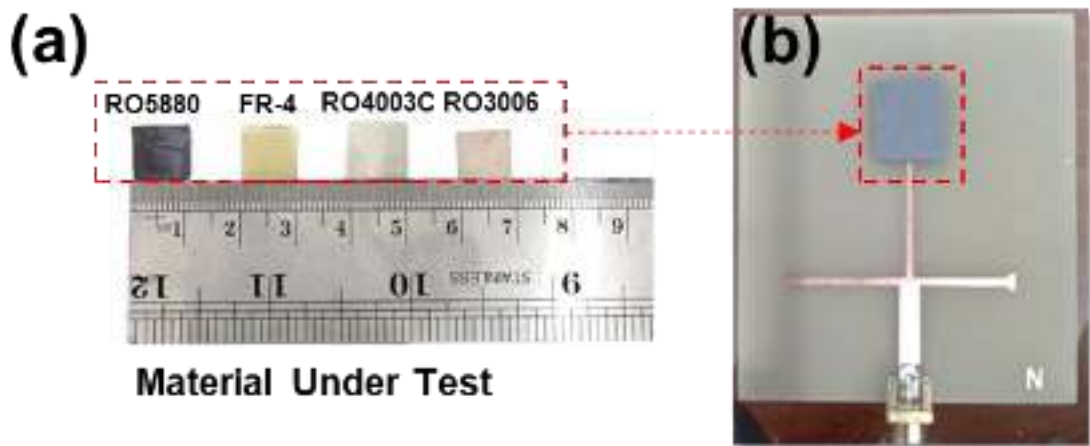


Figure 3.27 (a) Dimension of samples, (b) placement of the samples on the sensor

3.7 Accuracy of microwave sensor using polynomial curve fitting mode

Determining permittivity using a polynomial equation involves fitting experimental data to a polynomial model that relates the material's dielectric constant (permittivity) to the frequency of an applied electric field. The polynomial equation is a mathematical function expressed as a sum of powers of the frequency. By measuring the material's response at different frequencies and fitting the data to the polynomial equation, the permittivity can be extracted. The degree of the polynomial determines the complexity and accuracy of the fit. To determine the relative permittivity of the tested sample, an empirical equation can be modelled based on the relationship between the frequency (f_o) and the permittivity (ϵ_r) using curve fitting method which is 2nd order polynomial as indicated in equation 3.46 as follows:

$$\epsilon_r = af_o^2 + bf_o + C \quad (3.46)$$

where the coefficients **a**, **b**, **c** can be found by using standard materials under test (MUT). Furthermore, the accuracy of the sensor is determined by comparing the permittivity reference (ϵ_r **reference**) from the datasheet and the permittivity extracted from the measurement process (ϵ_r **calculated**) based on equation 3.45 as follows:

$$Accuracy = \frac{\epsilon_r \text{ calculated}}{\epsilon_r \text{ reference}} \times 100\% \quad (3.47)$$

3.8 Sensitivity and normalized sensitivity of microwave sensor

The sensitivity of a microwave sensor refers to its ability to detect and measure small changes in the parameter it is designed to sense. In the context of microwave sensors, this parameter could be physical quantities such as distance, displacement, temperature, humidity, pressure, or other properties that affect the sensor's behaviour or response. High

sensitivity is desirable as it allows the sensor to detect even subtle changes in the measured parameter, making it more accurate and reliable in various applications.

The interaction of the MUT to the resonator can be assumed as a capacitive load that causes the resonant frequency to shift. The ratio between the frequency shift and the change in the permittivity of the MUT is expressed as sensitivity. The sensitivity of the sensor (**S**) can be calculated using the equation 3.48 as follows (Wang *et al.*, 2021):

$$\Delta f = f_{unloaded} - f_{loaded} \quad (3.48)$$

$$S = \frac{\Delta f}{\Delta \epsilon_r} = \frac{(f_{unloaded} - f_{loaded})}{\epsilon_r (MUT) - \epsilon_r (Reference)} \quad (3.49)$$

where Δf represents the difference of the resonant frequency before (**$f_{unloaded}$**) and after the MUT is loaded (**f_{loaded}**) while $\Delta \epsilon_r$ is the difference in permittivity between the reference and the measured samples. Generally, the reference permittivity used is a vacuum with $\epsilon_r = 1$ as described in (Ebrahimi, Scott and Ghorbani, 2018a). The normalized sensitivity (**NS**) can also be calculated using the equation 3.48 as follows:

$$NS = \frac{1}{\Delta \epsilon_r} \left(\frac{f_{unloaded} - f_{loaded}}{f_{unloaded}} \right) \% \quad (3.50)$$

3.9 Measurement setup

Measurements processes were made using the Vector Network Analyzer (VNA) a frequency range of 1 GHz - 2.5 GHz and a sweep frequency of 0.01 GHz to validate the performance of the proposed sensor. The proposed sensor is connected to a VNA using a coaxial cable with an impedance of 50 ohms. The measurement process of proposed sensor using a VNA involves the following steps as shown in Figure 3.28.

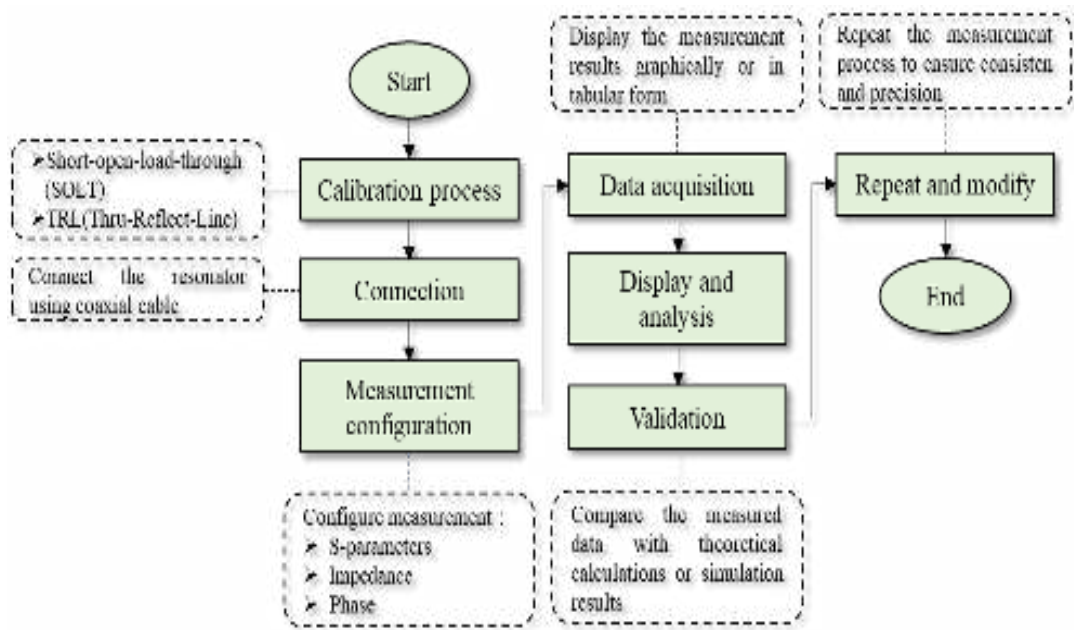


Figure 3.28 Measurement process of microwave sensor

Based on Figure 3.28, the first step of the measurement process is the calibration of the vector network analyser based on the frequency range to be used. Calibration is the essential first step to ensure accurate measurements. Network Analyzers typically come with different calibration methods, such as short-open-load-through (SOLT) or TRL (Thru-Reflect-Line) calibration. These calibrations compensate for the systematic errors introduced by the test setup, cables, connectors, and other components. The second step is to connect the resonator with the VNA using a coaxial cable with an impedance of 50 Ohms. The third step is to configure the Network Analyzer for the specific measurement to perform on the microwave sensor. This could involve selecting the appropriate measurement mode including S-parameters, impedance, frequency range, sweep frequency, and data resolution. The next stage is to acquire data from measurement results and validate them with simulation results or theoretical calculations. The results of the measurement process may be different from the simulation or calculation results. Therefore, repeated measurements are needed to

get accurate and precise results. Figure 3.29 shows the measurement setup of the resonator using VNA.

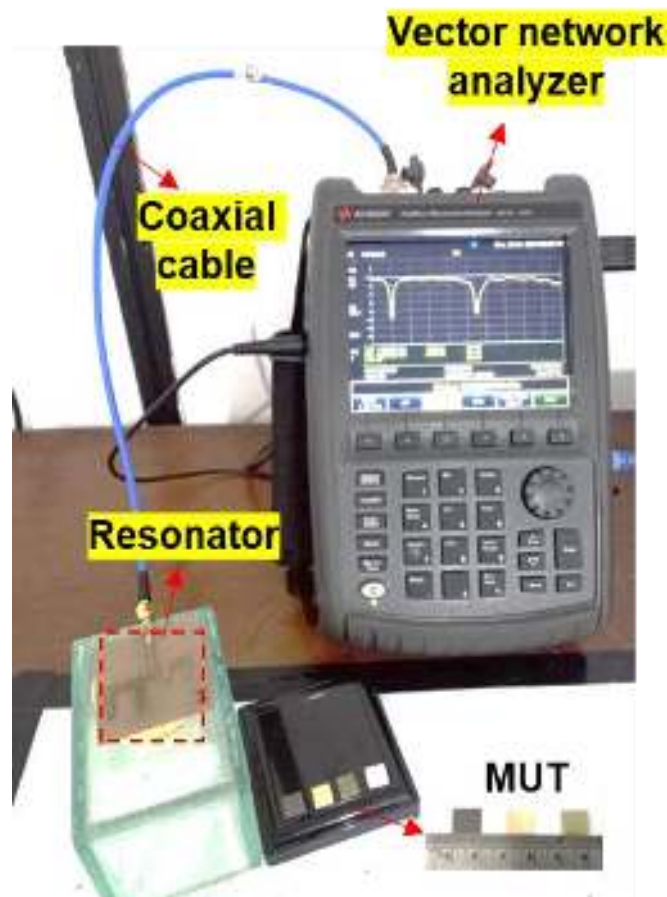


Figure 3.29 Measurement setup of microwave sensor

Furthermore, to measure the performance of the resonator as an antenna, radiation pattern measurements were carried out in the anechoic chamber. The radiation pattern is a graphical representation of how an antenna, or a radiating element distributes its radiated power in three-dimensional space. It shows the relative strength of the radiated electromagnetic waves in different directions from the antenna. The measurement process of radiation pattern involves the following steps as shown in Figure 3.30.

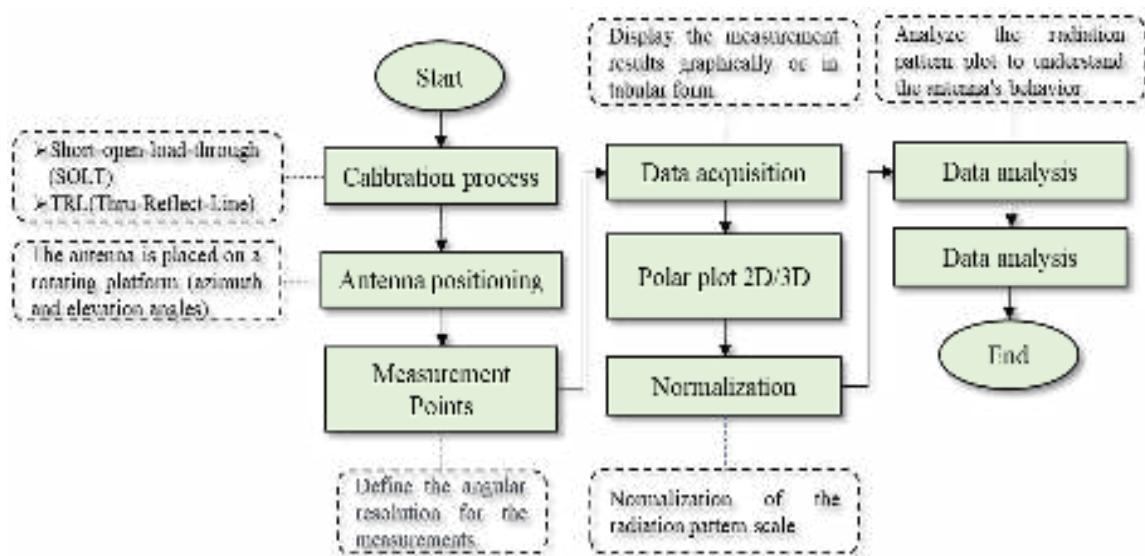


Figure 3.30 Measurement process of radiation pattern

Figure 3.30 shows the stages in measuring the radiation pattern of antenna. Before starting the measurement, set up the antenna and the measurement system in an open area or an anechoic chamber, which minimizes reflections from surrounding surfaces. Calibrate the measurement system to eliminate any systematic errors, such as cable losses and amplifier gain. The next step is positioning the antenna and the measurement system according to the measurement requirements. The antenna is placed on a rotating platform (azimuth and elevation angles) to cover the entire space around it with an angle $0^\circ - 360^\circ$. Furthermore, the measurement system records the received signal strength or power level. This can be done using a receiving antenna, a probe, or a scanning system that moves around the antenna under test. The data is typically collected for both azimuth and elevation angles. After data acquisition, the collected data is used to generate a radiation pattern plot. For antennas with rotational symmetry, a polar plot is commonly used, representing the radiation pattern in a 2D plane. For antennas with more complex radiation patterns, a 3D plot is used to visualize the radiation pattern in all directions. Radiation pattern measurements are often normalized

to the peak value of the main lobe (the strongest lobe) to provide a relative representation of the antenna's performance. Normalization allows easy comparison between different antennas and provides a clear picture of their directional characteristics. Final step is to analyze the radiation pattern plot to understand the antenna's behaviour. Key features to examine include the main lobe, side lobes, beamwidth, front-to-back ratio, and sidelobe levels. These characteristics are crucial in determining how the antenna radiates its energy and how it interacts with the surrounding environment. Figure 3.31 shows the measurement setup for measuring the radiation pattern in the anechoic chamber.

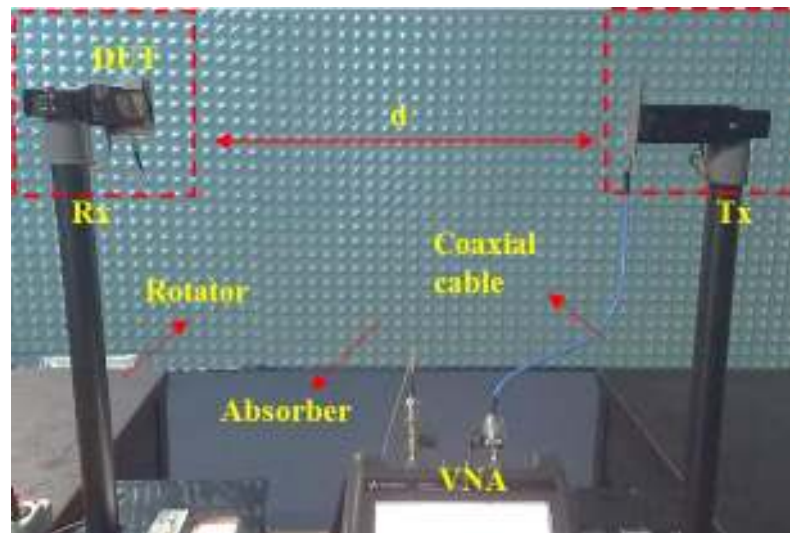


Figure 3.31 Measurement setup of radiation pattern in anechoic chamber

3.10 Concluding remarks

In this chapter, the stages and development of a single port microwave sensor with independent characteristics for permittivity detection of solid materials and as an antenna for data transfer capability have been comprehensively described. The characteristics of the

resonator are determined by observing the reflection coefficient, impedance and radiation pattern based on simulations and measurements.

First, a dual U-shape resonator was introduced for permittivity detection of solid materials independently for two types of samples with different permittivities. The resonator was proposed to operate at two resonant frequencies $f_{r1} = 1.29$ GHz and $f_{r2} = 2.02$ GHz with different sensing area locations so that they can be used for permittivity detection of two different types of MUT simultaneously. Furthermore, the independent performance of the sensor is determined using Pearson's correlation. However, the sample is still placed directly on the surface of the sensor so that it can be corrosive and damage the surface of the sensor.

Second, a dual T-shape resonator was introduced as a microwave sensor and antenna sensor for permittivity detection of two different types of MUTs by contact and non-contact. The sensor is proposed to operate at two resonant frequencies $f_{r1} = 2.32$ GHz and $f_{r2} = 1.81$ GHz with different sensing area locations that have independent characteristics. Determination of the functionality of the microwave sensor and antenna sensor is based on the characteristics of the radiation pattern and reflection coefficient of each resonator. The resonator with the best reflection coefficient and radiation pattern is proposed as an antenna sensor for non-contact detection while the other is used as a microwave sensor for contact detection. Non-contact detection is proposed by placing the sample on the sensing area surface which is separated by an air gap (d).

Furthermore, a multicascode T-shaped resonator is proposed as a multifunctional microwave sensor which is used for permittivity detection of two types of samples simultaneously and as an antenna for data transfer capability. The sensor is proposed to operate at three different frequencies $f_{r1} = 2.45$ GHz, $f_{r2} = 2.00$ GHz and $f_{r3} = 1.50$ GHz and

has different sensing area locations. The high frequency (f_1) is used as an antenna while the low frequency (f_2 and f_3) is used as a microwave sensor. Determination of resonator functionality as a microwave sensor and antenna is based on the characteristics and performance of the radiation pattern where the resonator with the best radiation pattern is used as an antenna while the others are used as microwave sensors.

The performance of the sensor is determined based on the response of the resonant frequency when the sample is placed simultaneously on each sensing area of the sensor while the performance of the antenna is shown by observing the radiation pattern and functioning of the antenna for data transfer capability. The sample used is a solid material with a known permittivity in the range 1 - 6.15. The dimensions of the sample are determined based on the availability of potential sensing area locations from the sensor. The permittivity of the sample extracted using a polynomial equation is compared with the reference permittivity of the datasheet to determine the accuracy of the sensor. The findings from the simulation and measurement results will be discussed in the next chapter.

CHAPTER 4

INDEPENDENT PERMITTIVITY SENSOR BASED ON DUAL U-SHAPED RESONATOR

This chapter presents the development of dual U-shaped resonator for independent permittivity sensor for characterize two different sample. The discussion in this chapter begins with an introduction to explain the main contribution of the proposed work, design and simulation of a dual U-shaped resonator, determining the location of sensing areas, scenarios for placing MUTs for independent characterization. The discussion continued with observing the performance of the sensor including the accuracy, sensitivity and correlation of each resonance frequency when characterizing two types of samples with different permittivity simultaneously. The final discussion of this chapter is validating the performance of the proposed sensor through a measurement process and comparing it with existing work to show the novelty of the proposed work.

4.1 Introduction

In this chapter, the planar method was used to detect the permittivity of the MUT by placing the sample on the surface of the resonator which has a high concentration of electrical fields. The sample used is a solid material whose permittivity is known with predetermined dimensions. The parameters observed were the resonant frequency and reflection coefficient of the resonator before and after the MUT was loaded. The focus of this research is to produce a microwave sensor that has independent characteristics with

different sensing area locations so that it can simultaneously detect the permittivity of two different types of samples. Furthermore, the independent characteristics are used to integrate sensors and antennas for contact and non-contact detection and are also used as antennas for data transfer capabilities which is explained in chapter 5 and chapter 6.

4.2 Design and simulation of dual U-shaped resonator

The design and simulation of the U-shaped resonator was carried out using HFSS 15.0 while the equivalent circuit used AWR 2009. The stages and simulation process are explained in the next sub-chapter.

4.2.1 Dual U-shaped resonator with single port

The dual U-shaped resonator consists of two pairs of U-shaped resonators connected to a transmission line. The resonator is proposed to work at $f_{r1} = 1.21$ GHz and $f_{r2} = 2.09$ GHz. The structure and simulation result of the dual U-shape resonator is shown in Figure 4.1.

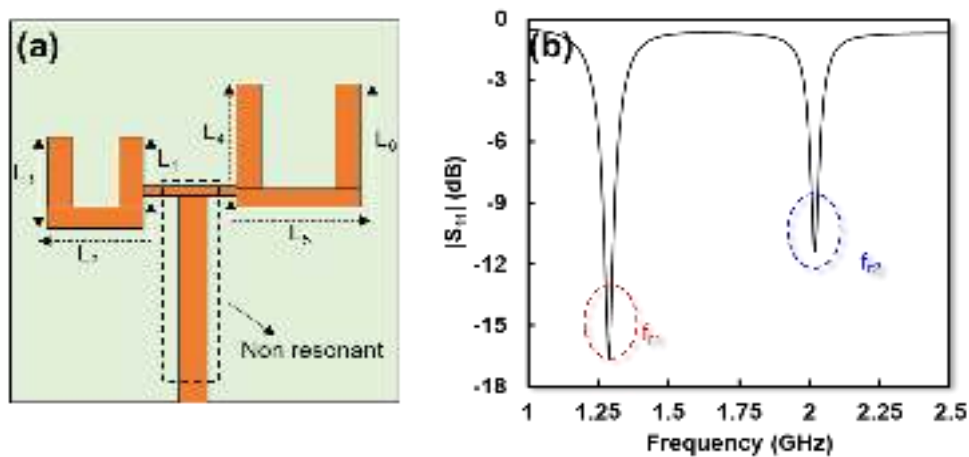


Figure 4.1 (a) Design of dual U-shaped resonator, (b) simulation result of S_{11} of dual U-shaped resonator

Figure 4.1 (a) shows the structure of a U-shaped dual resonator represented by the length (L) of each resonator. The 1st resonator is represented by L_1 , L_2 and L_3 while the 2nd resonator is represented by L_4 , L_5 and L_6 . The two resonators are connected by a T-stub which functions as a non-resonant power divider. Furthermore, Figure 4.1 (b) shows that the proposed resonator operates at two resonance frequencies namely $f_{r1}=1.29$ GHz and $f_{r2}=2.01$ GHz with S_{11} of -16.67 dB and -11.04 dB, respectively.

The electric field concentrations of both resonators are simulated using HFSS 15.0 based on their respective resonant frequencies. The simulation results of the electric field concentration of the dual U-shaped resonator at f_{r1} and f_{r2} are shown in Figure 4.2.

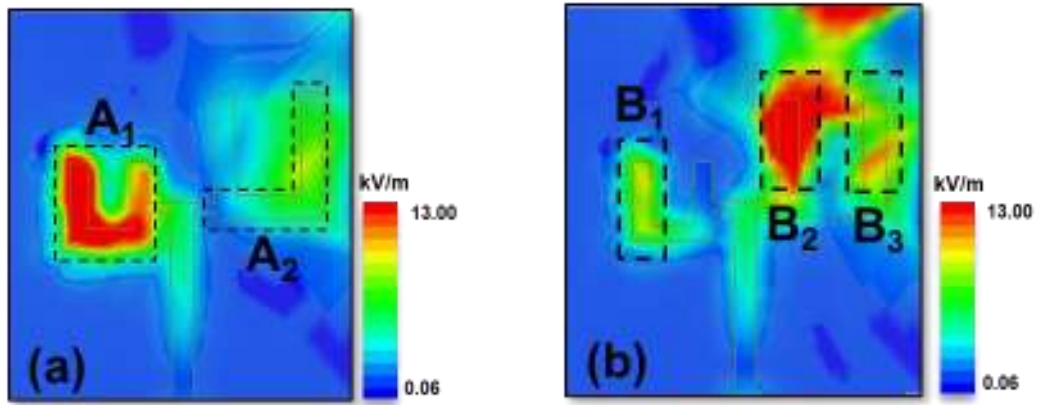


Figure 4.2 (a) Design of dual U-shaped resonator, (b) simulation result of S_{11} of dual U-shaped resonator

Figure 4.2 shows that the electric field concentrations of the resonators are at different locations based on their respective resonant frequencies. Figure 4.2 (a) shows that the concentration of the electric field for $f_{r1}=1.29$ GHz is in two different locations referring to A_1 and A_2 while for $f_{r2}=2.01$ GHz it is in three different locations referring to B_1 , B_2 and B_3 as shown in Figure 4.2 (b). These findings indicate that the two resonators work together

to generate the resonant frequency of the resonator. Therefore, the resonant frequency of the resonator can be determined based on the transmission line model.

The resonant frequency of an electromagnetic resonator is related to the electric field strength within the capacitive element. The electric field within the capacitor is directly proportional to the voltage across its plates and inversely proportional to the distance between them. Therefore, the resonant frequency of the dual U-shape resonator can be determined based on the length of the arms of each resonator which has a high electric field concentration, referring to Figure 4.2 (a) and Figure 4.2 (b). The transmission line model of the two resonators based on the resonator arm with the highest electric field concentration is shown in Figure 4.3 (a) and Figure 4.3 (b).

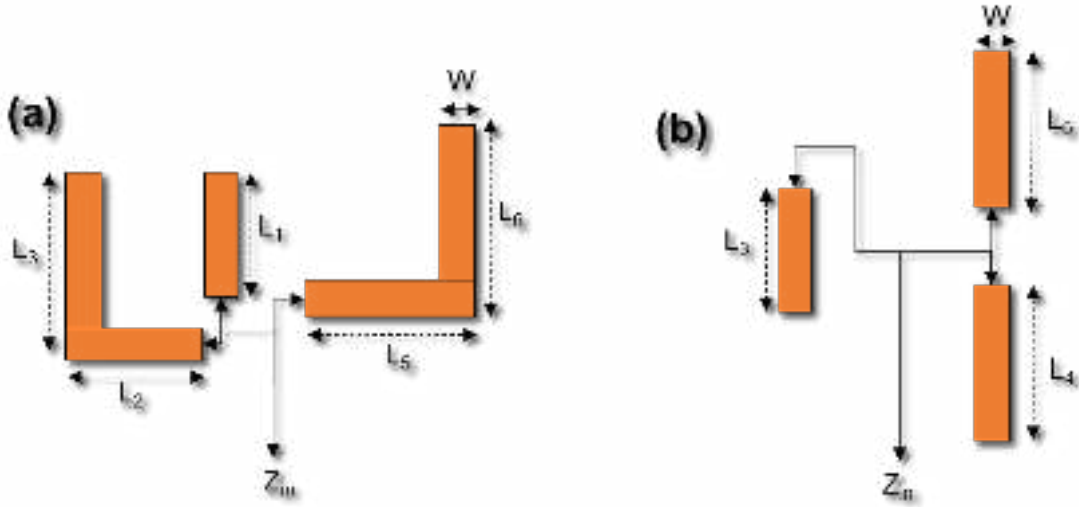


Figure 4.3 Transmission line model of proposed resonator; (a) 1st resonator, (b) 2nd resonator

Based on equation 3.14 and equation 3.15, the width of microstrip line (W_z) with impedance of 50Ω of the resonator can be determined as follows:

$$B = \frac{377\pi}{2Z_0\sqrt{\epsilon_r}} = \frac{(377)(3.14)}{2(50)\sqrt{4.3}} = 5.72$$

$$W_z = \frac{2}{\pi} \left[B - 1 - \ln(2B - 1) + \frac{\epsilon_r - 1}{2\epsilon_r} \left\{ \ln(B - 1) + 0.39 - \frac{0.61}{\epsilon_r} \right\} \right]$$

$$W_z = \frac{2}{3.14} \left[(5.72) - 1 - \ln(10.44 - 1) + \frac{4.3 - 1}{2(4.3)} \left\{ \ln(5.72 - 1) + 0.39 - \frac{0.61}{4.3} \right\} \right]$$

$$= 3.1 \text{ mm}$$

Based on the calculation, the width of the microstrip line (**W_z**) is 3.1 mm with an impedance (**Z₀**) of 50 Ohm. The width and length of the arm of the resonator is obtained from the optimization process using EM simulation software which represents the resonant frequency of the resonator as shown in Table 4.1.

Table 4.1 Optimized dimensions of the arm of the proposed resonator

Parameter	Dimension (mm)
L₁	10
L₂	11
L₃	10
L₄	15
L₅	15
L₆	15
w	2.5

Furthermore, the resonant frequency of each resonator can be determined based on the equation 3.4, equation 3.29 and 3.38 as follows:

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left[1 + 12 \frac{1.6}{2.5} \right]^{-\frac{1}{2}}$$

$$\epsilon_{eff} = \frac{4.3 + 1}{2} + \frac{4.3 - 1}{2} \left[1 + 12 \frac{1.6}{2.5} \right]^{-\frac{1}{2}} = 3.32$$

$$f_{r1} = \frac{1 (3 \times 10^8)}{2(L_1 + L_2 + L_3 + L_5 + L_6) \sqrt{\epsilon_{eff}}}$$

$$f_{r1} = \frac{1 (3 \times 10^8)}{2(10 + 11 + 10 + 15 + 15) \sqrt{3.32}} = 1.35 \text{ GHz}$$

$$f_{r2} = \frac{1 (3 \times 10^8)}{2(L_3 + L_6 + L_4)\sqrt{3.28}}$$

$$f_{r2} = \frac{1 (3 \times 10^8)}{2(10 + 15 + 15)\sqrt{3.32}} = 2.06 \text{ GHz}$$

To validate the calculations, a simulation with EM simulation software is proposed to observe the resonant frequency and performance of the resonator. The resonant frequency of each resonator can be modelled and simulated using EM simulation based on the arm of the resonator with the highest electric field concentration refer to Figure 4.3 (a) and Figure 4.3 (b). Figure 4.4 shows the simulation results which represent the resonant frequency of the proposed resonator.

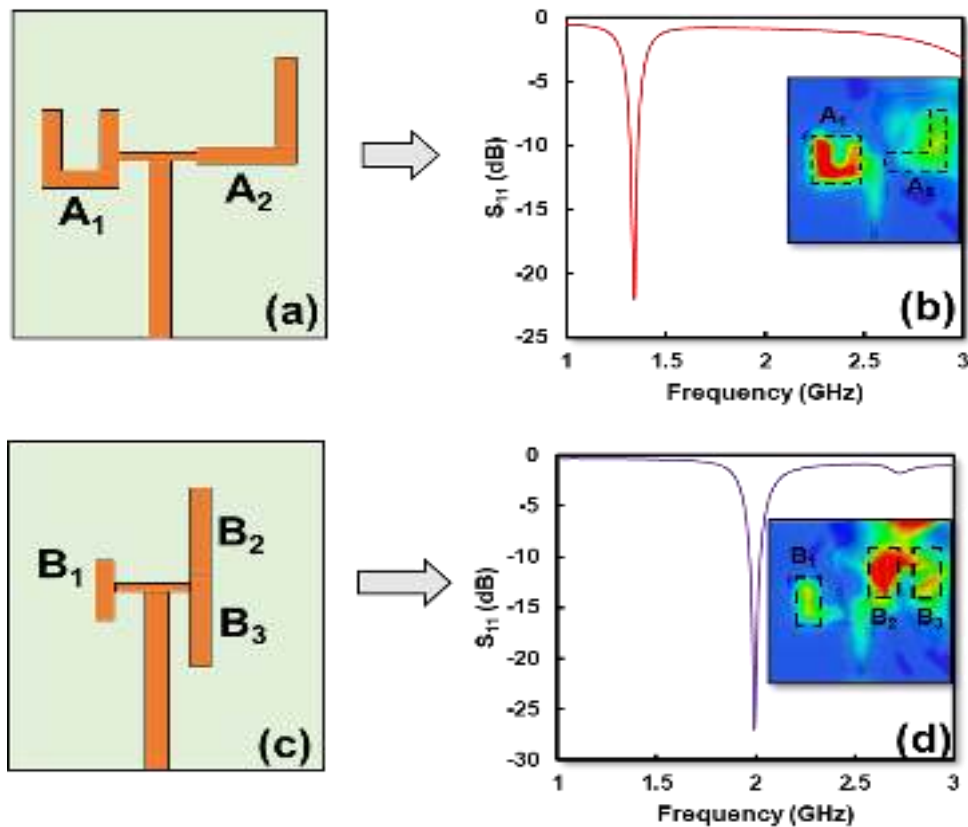


Figure 4.4 Simulation result of S_{11} using the transmission line model; (a) 1st resonator GHz, (b) 2nd resonator

Figure 4.4 (a) shows the simulation results of the first resonator with A_1 and A_2 operating at a resonant frequency of 1.34 GHz while the second resonator with B_1 , B_2 and B_3 operating at a resonant frequency of 2.01 GHz as shown in Figure 4.4 (b). Furthermore, this result shows that the simulation results using the EM simulation software are in line with the calculation where the resonator operates at a dual frequency of around 1.3 GHz and 2 GHz. These findings indicate that the transmission line approach can be used to determine the resonant frequency of a dual U-shape resonator.

4.2.2 Structure of dual U-shaped resonator with single port

The proposed sensor was designed based on FR4 substrate with dielectric constant (ϵ_r) of 4.3, loss $\tan(\delta)$ of 0.0265, and thickness (h) of 1.6 mm.

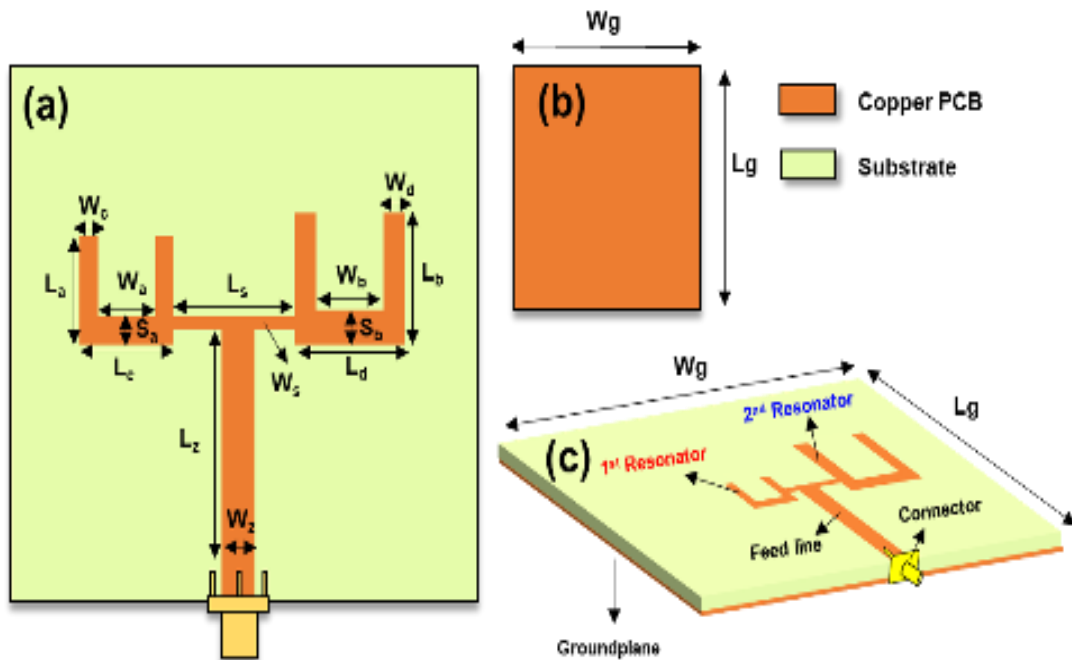


Figure 4.5 Structure of dual U-shaped resonator; (a) from the front view (b) from the back view (c) from perspective view

The sensor consists of two U-shaped resonators connected by a microstrip line ($Z_{L1} = 50 \text{ Ohm}$) and a power divider ($Z_{L2} = Z_{L3} = 100 \text{ Ohm}$) for the two resonators. The dimensions of the proposed sensor are 50 mm x 50 mm x 1.6 mm with the design structure, shown in Figure 4.5 (a), Figure 4.5 (b). In addition, both are connected to port 1, which is an SMA connector with an impedance of $Z_0 = 50 \text{ ohms}$ using a microstrip feed line as shown in Figure 4.5 (c). The overall optimized dimensions of the proposed sensor with dual U-shaped resonators are shown in Table 4.2.

Table 4.2: Dimension of dual U-shaped resonators

Parameter	Value (mm)	Parameter	Value (mm)
W_g	50 mm	L_g	50 mm
W_a	6 mm	L_a	10 mm
W_b	9 mm	L_b	15 mm
W_c	2.5 mm	L_c	11 mm
W_d	3 mm	L_d	15 mm
W_z	3 mm	L_z	24 mm
W_s	1 mm	L_s	9 mm
S_a	2 mm	S_b	3 mm

Based on simulation result using HFSS 15.0, each resonator operated at a different resonant frequency where the 1st resonator operates at $f_{r1} = 1.290 \text{ GHz}$, and the 2nd resonator operates at $f_{r2} = 2.020 \text{ GHz}$. Furthermore, Figure 4.6 shows the circuit model with values of the impedance and components of R_r , L_r and C_r of the proposed resonator using AWR 2009. Based on Figure 4.6 (a), the 1st resonator was indicated by $\mathbf{R_{r1}} = 8.45 \text{ } \Omega$, $\mathbf{C_{r1}} = 0.99 \text{ pF}$ and $\mathbf{L_{r1}} = 19.90 \text{ nH}$ while the 2nd resonator was indicated by $\mathbf{R_{r2}} = 9.175 \text{ } \Omega$, $\mathbf{C_{r2}} = 2.71 \text{ pF}$ and $\mathbf{L_{r2}} = 88 \text{ nH}$. To prevent a short circuit between port 1 and the two resonators, $\mathbf{C_{g1}} = 0.987 \text{ pF}$ and $\mathbf{C_{g2}} = 2.71 \text{ pF}$ are added to the model. Furthermore, the power divider is modeled using a transmission line with an impedance of $\mathbf{Z_{L2}}$ and $\mathbf{Z_{L3}}$ which is also used as impedance

matching between the impedance of the connector that connected with port 1 (Z_0) and the impedance of the resonators.

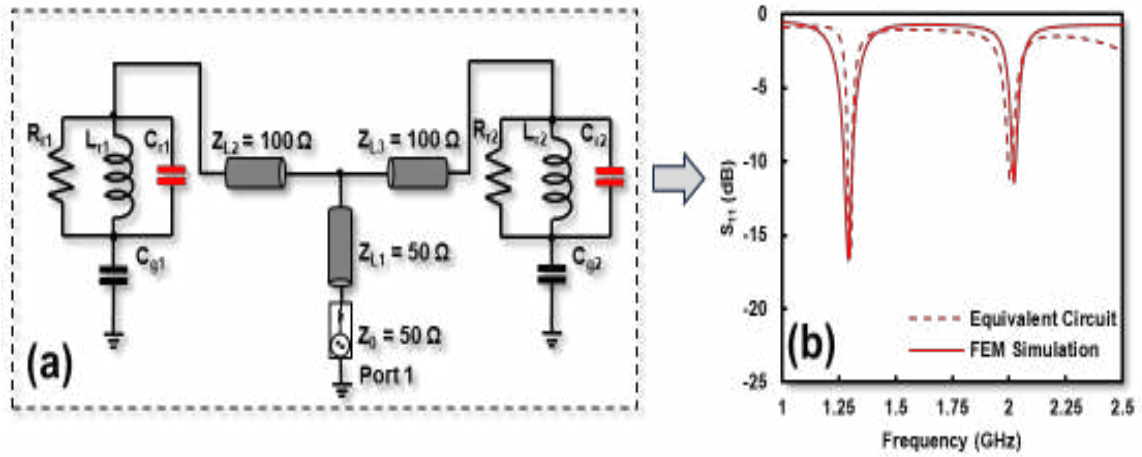


Figure 4.6 Equivalent circuit model of dual U-shaped resonators, (b) Simulation result of EQC vs FEM

Based on simulation using the equivalent circuit model, the resonators operate at $f_{r1} = 1.29$ GHz and $f_{r2} = 2.04$ GHz, respectively. These findings indicate that the equivalent circuit can be used to represent the proposed resonator and operates at frequencies around 1.2 and 2 GHz.

Furthermore, Figure 4.6 (b) shows a comparison of the simulation based on Finite Element Model (FEM) and equivalent circuit model. It can be clearly seen that the experimental results obtained show good agreement for element model and full-wave EM simulation.

4.3 Potential location of sensing area

The sensing region of the resonator is an area that has a high concentration of E-field. The electric field will be related to the capacitance while the magnetic field is related to the

inductance of the resonator. In order to observe and determine the E-field and H-field of the U-shaped resonator, a simulation was performed using HFSS 15.0. The resonant frequency used was $f_{r1} = 1.290$ GHz and $f_{r2} = 2.020$ GHz for each resonator as shown in Figure 4.7.

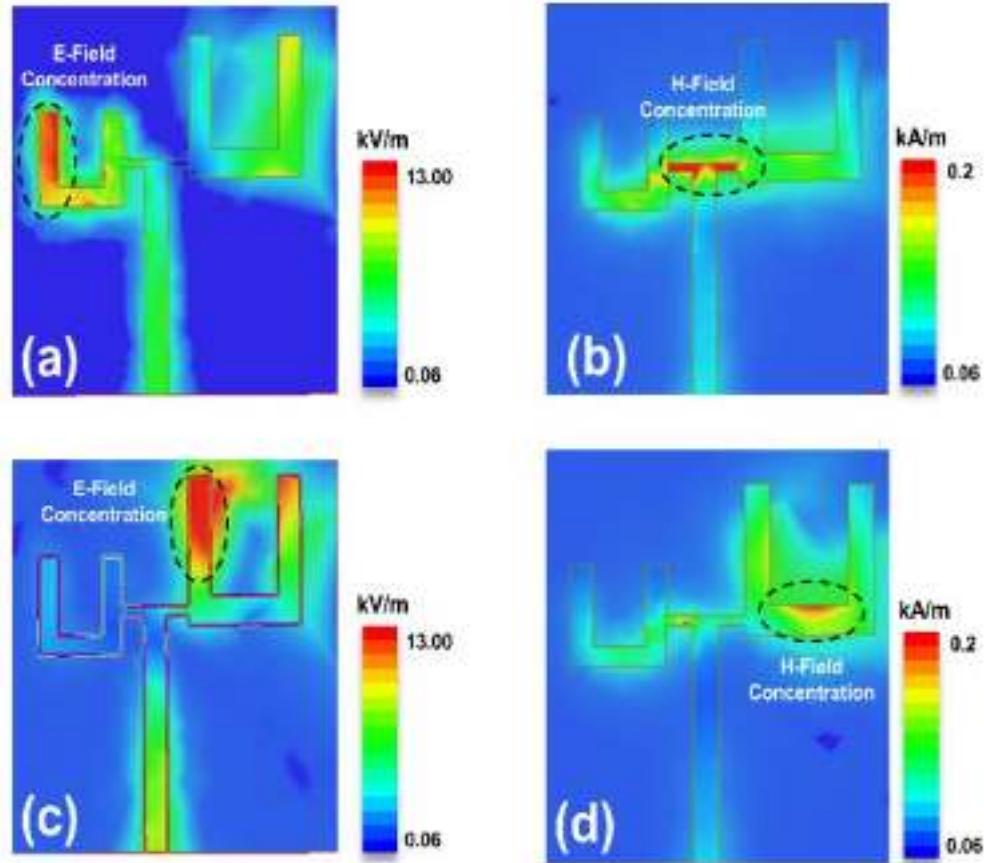


Figure 4.7 Simulation of E-field and H-field at $f_{r1} = 1.290$ GHz and $f_{r2} = 2.020$ GHz (a) E-field at f_{r1} (b) H-field at f_{r1} (c) E-field at f_{r2} (d) H-field at f_{r2}

Figure 4.7 shows that each resonator had different E-Field and H-field areas in f_{r1} and f_{r2} . The E-field was stored in the left arm of the 1st and 2nd resonators at $f_{r1} = 1.290$ GHz and $f_{r2} = 2.020$ GHz, respectively, shown in Figure 4.7 (a) and Figure 4.7 (c). However, the H-fields of both resonators had vanished and existed in different areas shown in Figure 4.7(b) and Figure 4.7(d).

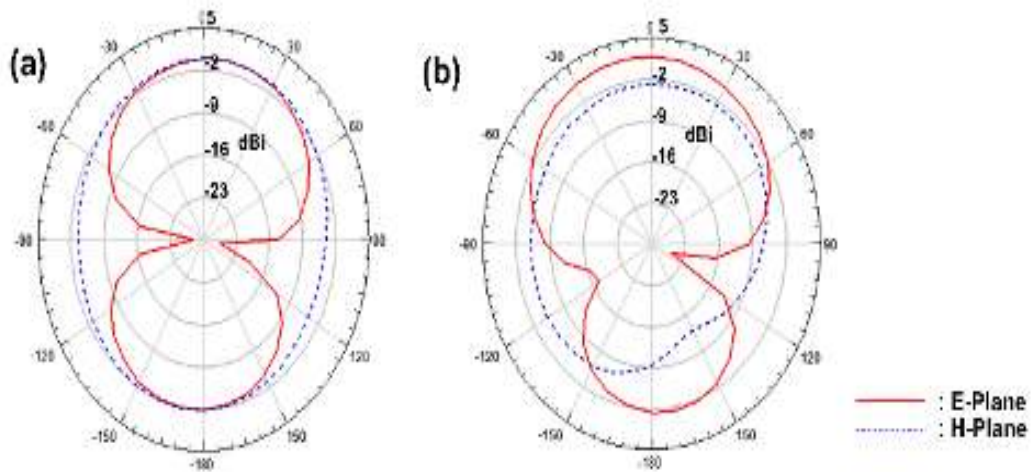


Figure 4.8 Simulation of E-field and H-field at $f_{r1} = 1.290$ GHz and $f_{r2} = 2.020$ GHz (a) E-field at f_{r1} (b) H-field at f_{r1} (c) E-field at f_{r2} (d) H-field at f_{r2}

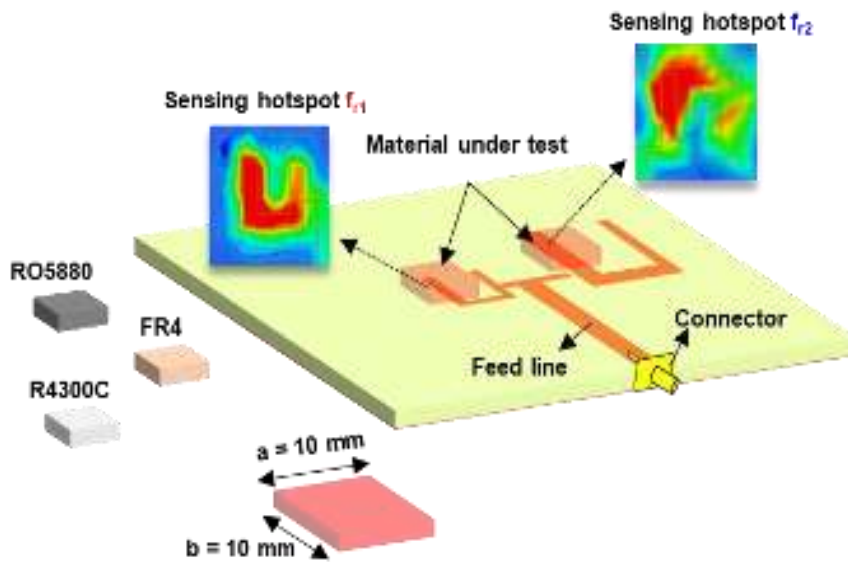


Figure 4.9 Placement material under test in 1st and 2nd resonator

Furthermore, the radiation pattern of the resonator at both resonant frequencies is shown in Figure 4.8 (a) and Figure 4.8 (b) where $f_{r1}=1.290$ GHz and $f_{r2}=2.020$ GHz with power input power level is 1 Watt. This shows that both resonators radiate waves at the resonant frequency in terms of the E-plane (red marker) and the H-plane (blue marker). The

sensing area with a strong E-Field of the 1st resonator with $f_{r1} = 1.290$ GHz is around on the left edge, while the 2nd resonator with $f_{r2} = 2.020$ GHz was around on the top edge of each U-shaped resonator as mentioned in Figure 4.7. Therefore, to detect permittivity independently, MUTs can be placed on sensing areas located on the left arm of the U-shaped resonator as shown in Figure 4.9.

4.4 Simulation result from scenario of MUT placement

A simulation using HFSS was carried out by placing the MUT in the sensing area area of each resonator in order to show the performance of the proposed sensor. The resonant frequency and reflection coefficient of the resonator were observed before and after loading the MUT. Simulation processes were carried out under two different scenarios:

- (1) The MUT was placed on the sensing area of the 1st resonator while the 2nd resonator was unloaded.
- (2) The MUT was placed on the sensing area of the 2nd resonator while the 1st resonator was unloaded.

Figure 4.10 (a) shows the resonant frequency response of the 1st resonator (f_{r1}) when the MUT was loaded with a permittivity of 1 – 10, while the 2nd resonator was unloaded as mentioned in scenario (1). The dimensions of the MUT used were 10 mm x 10 mm with a thickness of 1.6 mm. The simulation results show that the resonant frequency of the 1st resonator (f_{r1}) moved to the lower frequency in line with the change in permittivity of the MUT from 1.300 GHz to 1.220 GHz as shown in Figure 4.10 (b). The high permittivity of the MUT caused the resonant frequency of the resonator to shift. In other words, the resonant frequency shifted to a lower one. However, the resonant frequency of the 2nd resonator (f_{r2})

shifted but not significantly between 2.010 GHz and 2.030 GHz. Furthermore, the simulation and frequency response during scenario (2) are shown in Figure 4.11 (a) and Figure 4.11 (b).

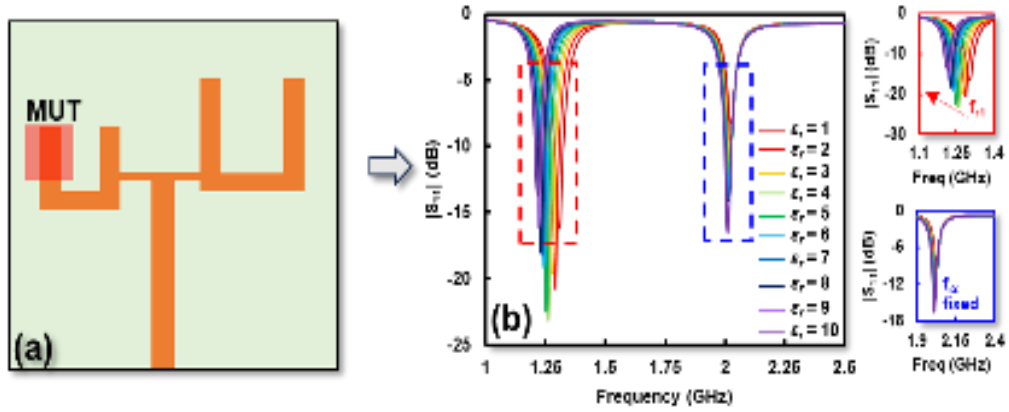


Figure 4.10 (a) Placement MUT for scenario (1), (b) response of resonant frequency of proposed resonator during scenario (1)

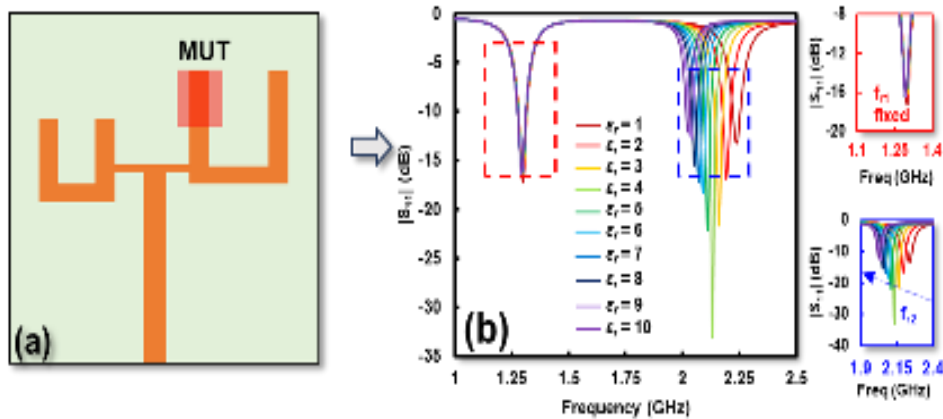


Figure 4.11 (a) Placement MUT for scenario (2), (b) response of resonant frequency of proposed resonator during scenario (2)

Figure 4.11 (a) shows that the MUT is placed at the sensing area of the 2nd resonator while the 1st resonator is unloaded. Moreover, Figure 4.11 (b) shows a simulation of the 2nd resonator when the MUT was loaded with a permittivity of 1 to 10 while the 1st resonator was unloaded (vacuum). Based on simulation, f_2 shifted from 2.240 GHz to 2.020 GHz

respectively when the MUT was loaded on the 2nd resonator, while for f_{r1} , it was fixed at 1.290 GHz. It is clearly seen that the resonant frequency of the 2nd resonator (f_{r2}) shifted to the lower frequency in accordance with the increase in the permittivity of the MUT, while for the frequency of the 1st resonator (f_{r1}) was fixed. The correlation between f_{r1} and f_{r2} when the MUT loaded during scenario (1) and scenario (2) is shown in Figure 4.12 (a) and Figure 4.12 (b).

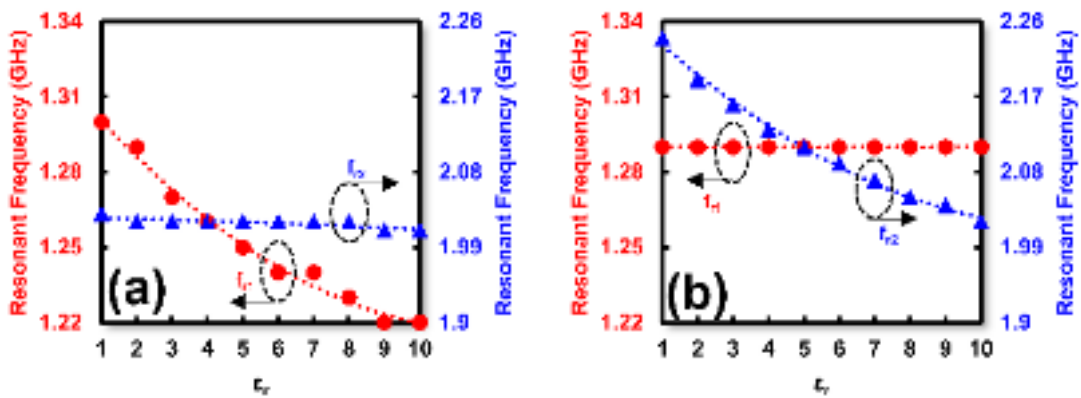


Figure 4.12 Correlation of f_{r1} and f_{r2} ; (a) during scenario (1), (b) during scenario (2)

The Δf of the resonator could be determined by comparing the difference between the resonant frequency of the lowest and highest permittivity of the MUT based on equation 3.48. The simulation results show that the 1st and 2nd resonators had Δf of 0.08 GHz and 0.22 GHz, respectively. The Δf of the 2nd resonator was higher than the 1st resonator due to the characteristics of high resonance frequency, which was more sensitive to small variations in the permittivity of the MUT, as described in (Armghan *et al.*, 2021). Furthermore, Figure 4.13 shows the effect of $\tan \delta$ on the reflection coefficient of the 1st and 2nd resonator, respectively. It is clearly seen that the reflection coefficient of the resonator became lower in line with the increase of the $\tan \delta$ of the MUT.

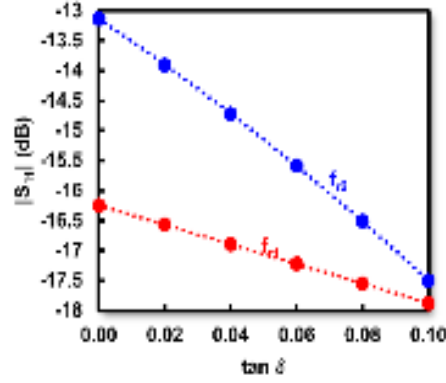


Figure 4.13 Correlation of f_{r1} and f_{r2} with the variations of $\tan \delta$

The change in the reflection coefficient on f_{r2} was higher from -13 dB to -17.5 dB compared with f_{r1} , which was -16.3 dB to -18 dB. It should be mentioned that these MUTs are assumed low loss such that by increasing the loss tangent from 0 to 0.1 (Haq *et al.*, 2019). The loss tangent of the MUTs influences the bandwidth or the quality factor of the S_{11} characteristics of the resonators (Armghan *et al.*, 2021).

4.5 Verification of dual U-shaped resonator

Fabriaction and verification of the resonator shape was carried out in the laboratory using a Vector Network Analyzer (VNA) with an ambient temperature of 25°. The stages and measurement process are explained in the next sub-chapter.

4.5.1 Fabrication and measurement of dual U-shaped resonator

Measurements were made using the Keysight Field Fox Network Analyzer (NA) N9915A with a frequency range of 1 GHz - 2.5 GHz and a sweep frequency of 0.01 GHz to validate the performance of the proposed sensor. The proposed sensor is connected to a

Network Analyzer using a coaxial cable with an impedance of 50 ohms. Measurements were carried out in a closed room with a temperature of 25° C with 3-fold measurement in one cycle to ensure short - term repeatability. The repeated measurements give an exact outcome of the average data values ($\bar{x}$) and shows consistent results for both resonators, respectively.

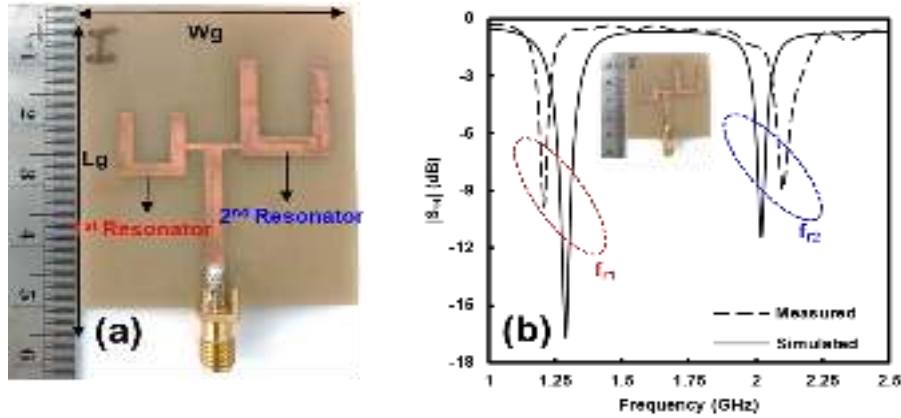


Figure 4.14 (a) Fabricated of proposed resonator, (b) comparison from simulation and measurement result of proposed resonator

Figure 4.14(a) shows fabricated of proposed resonator while a comparison of the measurement and simulation results of the proposed sensor is shown in Figure 4.14 (b). It is clearly seen that the measurement results from the resonator had the same characteristics as the simulation results, namely dual-band. However, there were still frequency shifts of 3.7% and 3.8% for each resonant frequency in the resonator between the simulation results and measurements, this was due to the variation of permittivity in the substrate, i.e range of ϵ_r is 4.25 – 4.35 and also the error during fabrication process. Based on the measurement results obtained $f_{r1} = 1.200$ GHz and $f_{r2} = 2.100$ GHz for each resonator.

An illustration of the measurement setup of the proposed sensor is shown in Figure 4.15 (a) while an overview of the direct measurement process is shown in Figure 4.15 (b). In this research, measurements were carried out under two different scenarios:

- (1) The MUT was placed on the sensing area of the 1st resonator while the 2nd resonator was unloaded.
- (2) The MUT was placed on the sensing area of the 2nd resonator while the 1st resonator was unloaded.

Furthermore, the resonant frequency of each resonator was observed using a Network Analyzer after the MUT was loaded on each resonator. Based on the theory, the resonant frequency would shift after the MUT was loaded, as described in (Rammah Ali Alahnomi *et al.*, 2017).

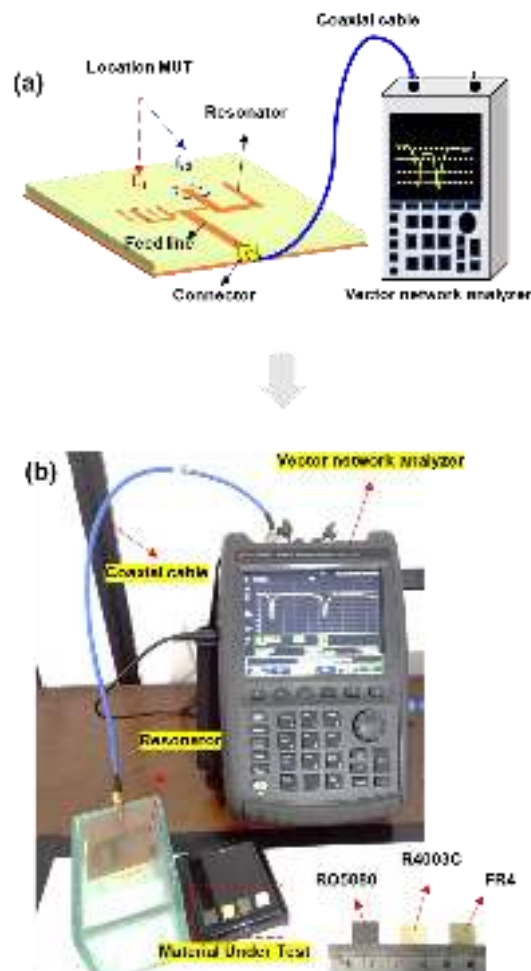


Figure 4.15 (a) Measurement setup (b) photograph of measurement process

In this study, three types of standard materials with known permittivity used as MUT are RO5880, RO4003C, and FR4. RO5880 had a permittivity of 2.2, loss tan ($\tan \delta$) of 0.0009, and a thickness of 1.57 mm, for RO4003C the permittivity was 3.38, loss tan ($\tan \delta$) was 0.0028, and a thickness of 1.57 mm, while for FR4 the permittivity was 4.3, loss tan ($\tan \delta$) was 0.0265 and a thickness of 1.6 mm. The dimensions of the MUT loaded on the resonator were 10 mm x 10 mm x 1.6 mm and placed at the sensing area of each resonator independently as mentioned in Figure 4.10 and Figure 4.11.

Figure 4.16 (a) shows the measurement results of the resonant frequency in scenario (1) where the MUT was loaded above the 1st resonator and 2nd resonator is unloaded.

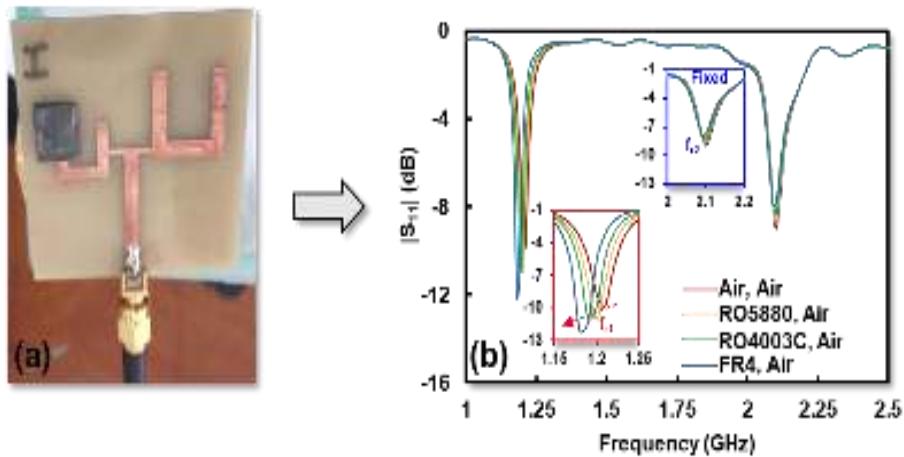


Figure 4.16 (a) Measurement based on scenario (1), (b) response of resonant frequency of proposed resonator during scenario (1)

The resonant frequency of the 1st resonator shifted to the lower frequency in accordance with the increase in the permittivity of the MUT while the frequency of the 2nd resonator was fixed. The resonant frequency of the 1st resonator (f_{r1}) shifted from 1.200 GHz to 1.190 GHz for RO5880 and 1.180 GHz for RO4003C while for FR4 to 1.170 GHz as shown in Figure 4.16 (b). On the other hand, Figure 4.17 (a) shows the results of the measurement of the resonant frequency in condition (2) where the MUT was loaded on the

2nd resonator.

Similar to scenario (1), the resonant frequency of the 2nd resonator (f_{r2}) changes in accordance with the increase of the permittivity of the MUT while the frequency of the 1st resonator was fixed. The resonant frequency of the 2nd resonator $f_{r2} = 2.100$ GHz shifted to 2.070 GHz for RO5880 and 2.050 GHz for RO4003C while for FR4 shifted to 2.020 GHz as shown in Figure 4.17 (b).

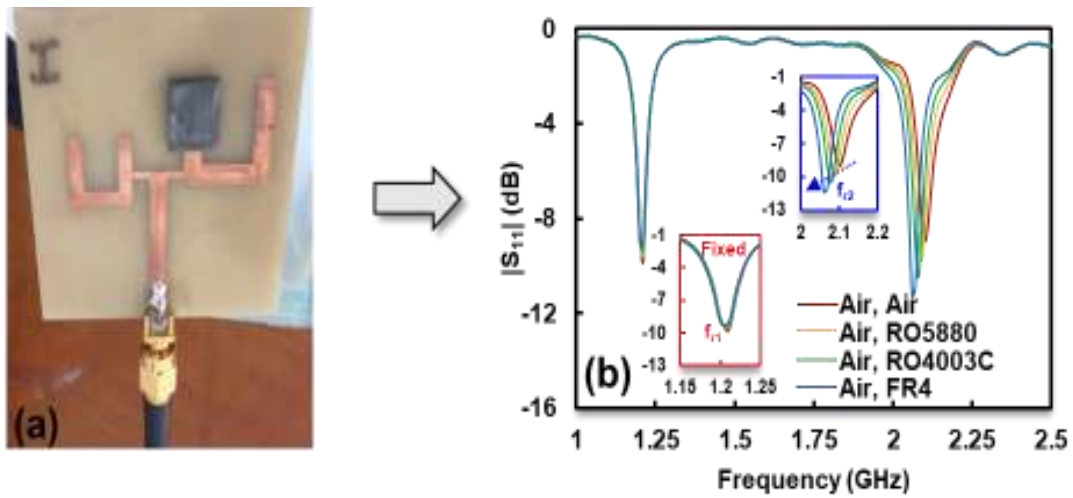


Figure 4.17 (a) Measurement based on scenario (2), (b) response of resonant frequency of proposed resonator during scenario (2)

The measurement results show that the behavior of the two resonators had a good agreement with the simulation. The resonant frequencies of the two resonators f_{r1} and f_{r2} were independent and did not affect each other when the MUT was loaded. Furthermore, both resonators were also sensitive to small variations in the permittivity of the MUT. Figure 4.18 (a) and Figure 4.18 (b) shows the correlation between f_{r1} and f_{r2} during scenario (1) and scenario (2).

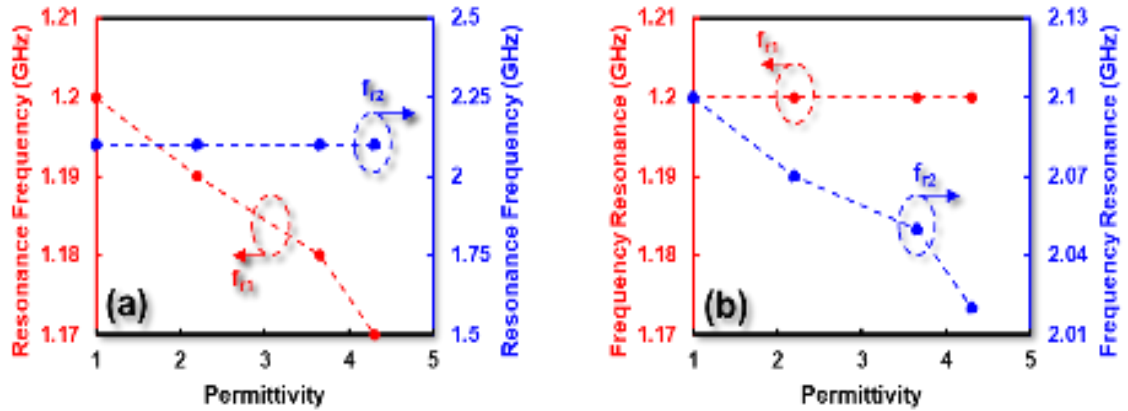


Figure 4.18 Correlation of f_{r1} and f_{r2} ; (a) during scenario (1), (b) during scenario (2)

Based on Figure 4.18 (a) and Figure 4.18 (b), the resonant frequency of the resonator during scenario (1) and scenario (2) is shown in Table 4.2.

Table 4.2 Resonant frequency of proposed resonator based on scenario (1) and scenario (2)

MUT	Permittivity	Scenario			
		Scenario (1)		Scenario (2)	
		f_{r1} (GHz)	f_{r2} (GHz)	f_{r1} (GHz)	f_{r2} (GHz)
Vacuum	1	1.20	2.10	1.20	2.10
RO5880	2.2	1.19	2.10	1.20	2.07
RO4003C	3.65	1.18	2.10	1.20	2.05
FR4	4.3	1.17	2.10	1.20	2.02

4.5.2 Accuracy and sensitivity of dual U-shaped resonator

In order to calibrate the sensor and developing mathematical model for measuring the permittivity, MUT with known permittivity were used as calibration samples to obtain the relation between the permittivity and resonance frequency (Ebrahimi, Scott and Ghorbani, 2020b). Three measurements were taken for each calibration sample, and the average values ($\bar{x}$) were calculated and used in curve fitting method. Each standard MUT

was placed on the maximum electric field and the result was measured and recorded after the calibration was made (Rammah Ali Alahnomi *et al.*, 2017). Repeatability measurement of proposed resonator during scenario (1) and scenario (2) are shown in Figure 4.19 (a) and Figure 4.19 (b)

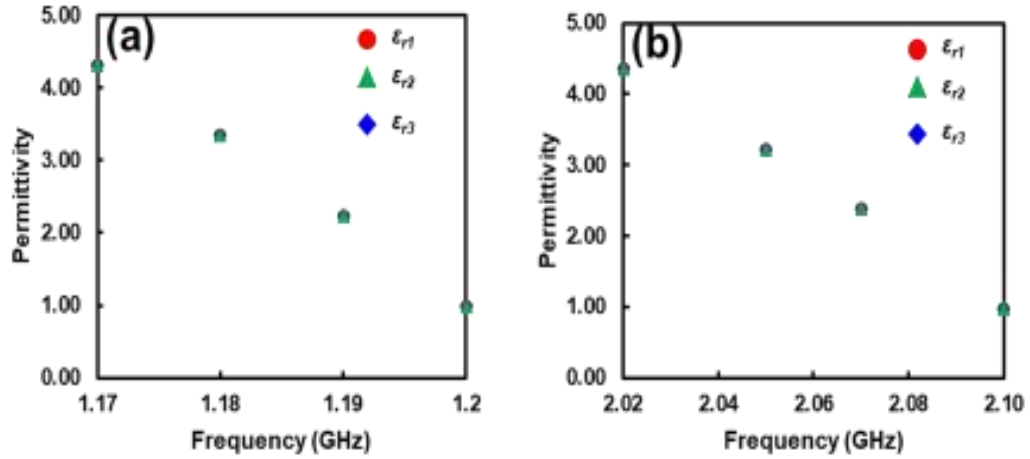


Figure 4.19 Repeatability measurement of proposed resonator; (a) during scenario (1), (b) during scenario (2)

Figure 4.19 (a) and Figure 4.19 (b) shows that the repeated measurements give an exact outcome of the average data values ($\bar{x}$) and the variations of the minimal results available show the error bars extremely small. The average value of the resonant frequency based on repeated measurements is shown in Table 4.4 and Table 4.5.

Table 4.4: Resonant frequency from repeatability measurement during scenario (1)

MUT	ϵ_r	f_{r1}'	f_{r1}''	f_{r1}'''	Average ($\bar{x}$)
Vacuum	1	1.20	1.20	1.20	1.20
RO5880	2.2	1.19	1.19	1.19	1.19
RO4003C	3.38	1.18	1.18	1.18	1.18
FR4	4.3	1.17	1.17	1.17	1.17

Table 4.5: Resonant frequency from repeatability measurement during scenario (2)

MUT	ϵ_r	f_{r2}'	f_{r2}''	f_{r2}'''	Average ($\bar{x}$)
Vacuum	1	2.10	2.10	2.10	2.10
RO5880	2.2	2.07	2.07	2.07	2.07
RO4003C	3.38	2.05	2.05	2.05	2.05
FR4	4.3	2.02	2.02	2.02	2.02

Furthermore, in order to verify the measurement results, a mathematical model approach was used using a curve fitting model. Figure 4.20 (a) and Figure 4.20 (b) illustrate a comparison of the fitted 2nd order polynomial of the permittivity of the MUT.

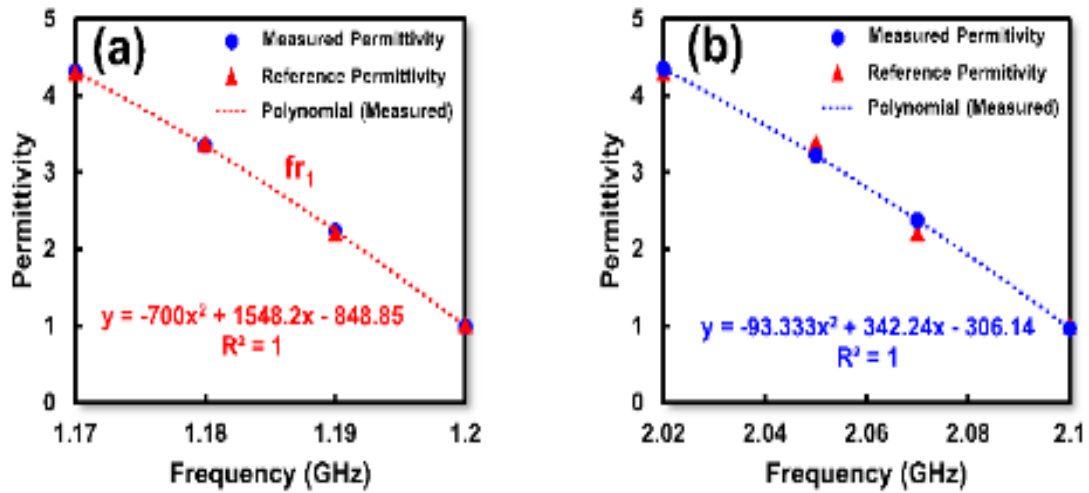


Figure 4.20 The fitting curve of MUT from measurement compared with reference (a) during scenario (1) (b) during scenario (2)

The fitted equation is expressed to extract materials with unknown permittivity compared with permittivity from datasheet based on equation 3.46. The red marker indicates the measurement result-based permittivity. Meanwhile, the blue marker indicates the reference permittivity in the standard material. Figure 4.20 (a) and Figure 4.20 (b) show the measurement results of the 1st and 2nd resonator in excellent agreement with the reference permittivity of the MUT for both resonant frequencies. Finally, the permittivity measurement

for the 1st and 2nd resonator can be expressed in a mathematical model as follows:

$$\varepsilon_r (f_{r1}) = -700f^2 + 1548.2f - 848.85 \quad (4.1)$$

$$\varepsilon_r (f_{r2}) = -93.333f^2 + 342.24f - 306.14 \quad (4.2)$$

Based on the following polynomial equation, the permittivity of the MUT during scenario (1) and scenario (2) can be determined as shown in Table 4.3 and Table 4.4. Furthermore, accuracy is obtained based on equation 3.47 by comparing the permittivity of the calculation results with the permittivity of the datasheet.

Table 4.3 Permittivity, error and accuracy of proposed sensor during scenario (1)

MUT	Permittivity(ε_r)		Error (%)	Accuracy %
	Reference	Calculated		
Vacuum	1	0.99	1.00	99.00
RO5880	2.2	2.24	1.73	98.27
RO4003C	3.65	3.35	1.01	98.99
FR4	4.3	4.31	0.33	99.67

Table 4.4 Permittivity, error and accuracy of proposed sensor during scenario (2)

MUT	Permittivity(ε_r)		Error (%)	Accuracy %
	Reference	Calculated		
Vacuum	1	0.97	3.45	96.55
RO5880	2.2	2.37	7.92	92.08
RO4003C	3.65	3.22	4.73	95.27
FR4	4.3	4.35	1.14	98.86

The Δf , sensitivity (S) and normalize sensitivity (NS) of both resonators can be determined based on equation 3.48, equation 3.49 and equation 3.50. Table 4.5 and Table 4.6 show the calculation of the normalized sensitivity and sensitivity of the two proposed scenarios.

Table 4.5 Δf sensitivity and normalize sensitivity of proposed sensor during scenario (1)

MUT	Δf (GHz)	S (%)	NS (%)
Vacuum	0	-	-
RO5880	0.01	8.33	0.69
RO4003C	0.02	8.40	0.70
FR4	0.03	9.09	0.76

Table 4.6 Δf sensitivity and normalize sensitivity of proposed sensor during scenario (2)

MUT	Δf (GHz)	S (%)	NS (%)
Vacuum	0	-	-
RO5880	0.03	25.00	1.98
RO4003C	0.05	21.01	1.60
FR4	0.08	24.24	1.15

Based on calculation from Table 4.8 and Table 4.9, the maximum sensitivity of proposed sensor was $9.09 \text{ MHz}/\Delta\epsilon_r$ and $25.00 \text{ MHz}/\Delta\epsilon_r$ while the maximum normalized sensitivities (NS) of f_{r1} and f_{r2} were 0.76 and 1.98, respectively, Furthermore, the sensitivity of the 1st resonator and 2nd resonator is shown in Figure 4.21 (a) and Figure 4.21 (b).

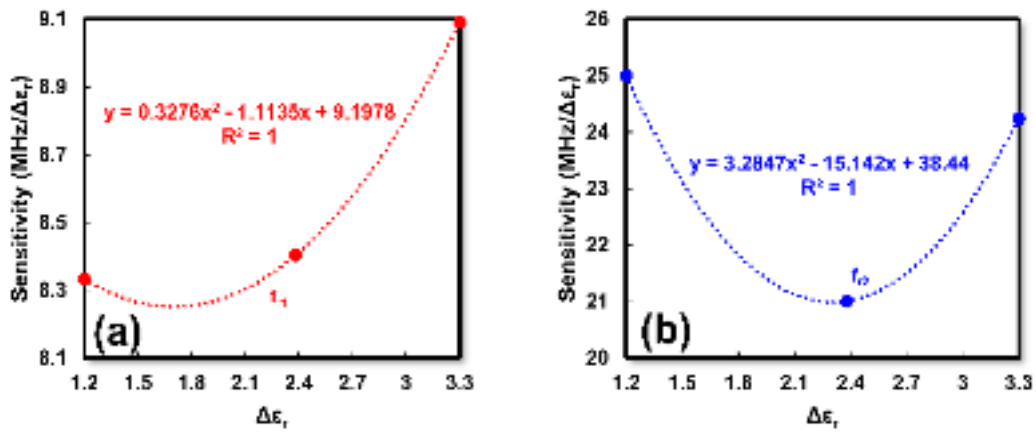


Figure 4.21 Sensitivity of proposed sensor (a) 1st resonator (b) 2nd resonator

The measurement results were in good agreement with the simulation results where The sensitivity of the 2nd resonator was higher than the 1st resonator, which is in line with

the previous study mentioned in (Haq *et al.*, 2019). It is clearly seen that f_{r1} and f_{r2} had an average accuracy of 99.02% and 96.44%, while the normalized sensitivity of the two resonators had an average of 0.72% and 1.58%, respectively. In addition, the proposed sensor has independent characteristics with two different sensing area locations so that it can be used to characterize two types of samples with different permittivities. In order to determine the independent performance of the proposed resonator, several experiments were carried out and explained in the next sub-chapter.

4.6 Independent performance of dual U-shaped resonator

The independent performance of the two resonators was observed by placing the MUT on both resonators simultaneously based on potential location sensing area referring to Figure 4.2. In this experiment, six potential locations of sensing area of proposed sensors are used to place the MUT simultaneously as shown in Figure 4.22 (a) and Figure 4.22 (b).

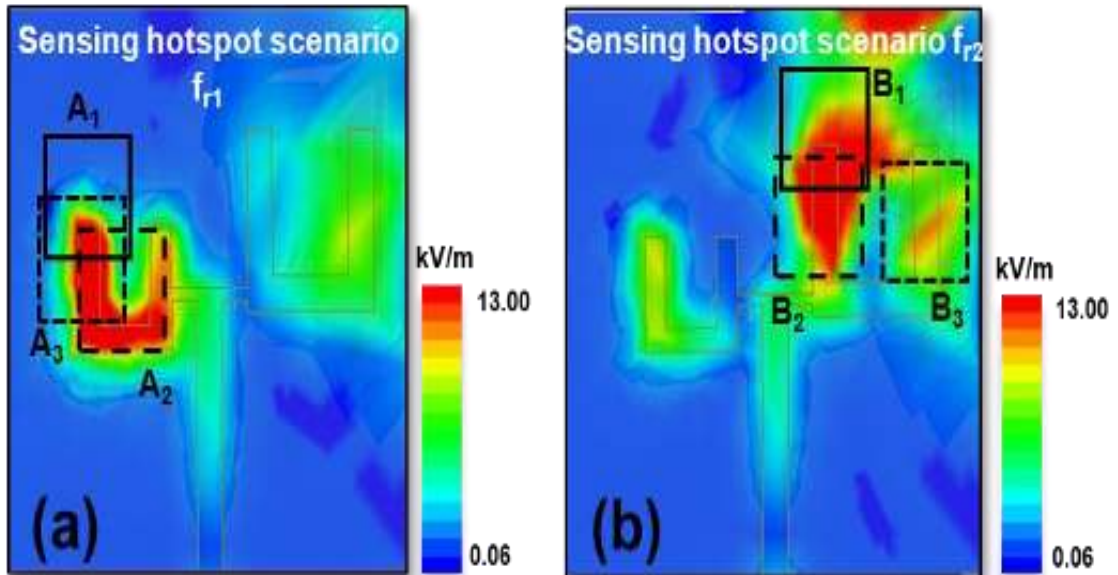


Figure 4.22 (a) sensing area scenario of $f_{r1} = 1.29$ GHz (b) sensing area scenario of $f_{r2} = 2.02$ GHz

To achieve independent sensing areas, a scenario for the placement of Materials Under Test (MUT) is proposed around the sensing area. This scenario includes the placement of MUTs in positions A_1 , A_2 , and A_3 for the 1st resonator, and B_1 , B_2 , and B_3 for the 2nd resonator as shown in Figure 4.22 (a) and Figure 4.22 (b). Using the same measurement setup as described in sub-chapter 4.3, nine MUT placement conditions are proposed to determine the independent performance of the proposed sensor as shown in Table 4.7.

Table 4.7 Placement of MUT based on nine conditions

Conditions	Location of MUT	
	MUT A	MUT B
C_1	A_1	B_1
C_2	A_1	B_2
C_3	A_1	B_3
C_4	A_2	B_1
C_5	A_2	B_2
C_6	A_2	B_3
C_7	A_3	B_1
C_8	A_3	B_2
C_9	A_3	B_3

Measurement and verification are carried out by placing the MUT on both resonators simultaneously with the following conditions:

- (1) The MUT on the 2nd resonator is fixed while the 1st resonator is placed with different of MUT.
- (2) the MUT is placed on the 1st resonator fixed while the 2nd resonator is placed differently.

Illustrations of MUT placement from scenarios (1) and (2) are shown in Figure 4.23 (a) and Figure 4.23 (b). In this experiment, RO5880 was used as the fixed material for both proposed conditions. The independent performance was observed from the response of f_{r1}

and f_{r2} when the MUTs were placed simultaneously on each sensing area of the proposed resonator.

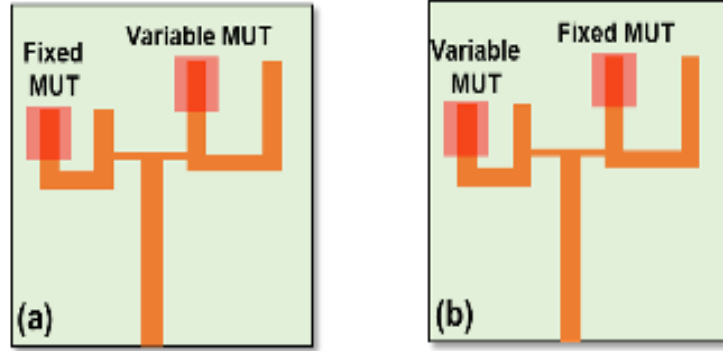


Figure 4.23 Illustration of MUT placement; (a) based on scenario (1), (b) based on scenario (2)

Furthermore, the scenario observed in this experiment is only on S_6 and S_8 while for complete data will be shown in the appendix. The measurement process of proposed sensor based on C_6 and C_8 is shown in Figure 4.24 (a) and Figure 4.24 (b).

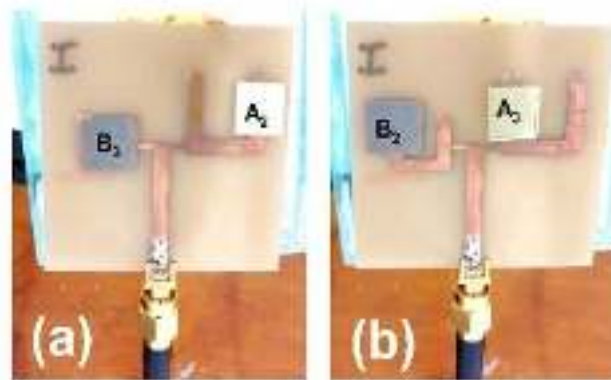


Figure 4.24 Placement scenario of MUT based on conditions; (a) C_6 , (b) C_8

The measurement results of C_6 with scenario (1) and (2) are shown in Figure 4.25 (a) and Figure 4.25 (b). In scenario (1), f_{r1} and f_{r2} are independent and not correlated with each other as shown in Figure 4.25 (a). However, in scenario (2) f_{r1} and f_{r2} affect each other as

shown Figure 4.25 (b), which is indicated by a shift in the resonance frequency simultaneously when the MUT is loaded.

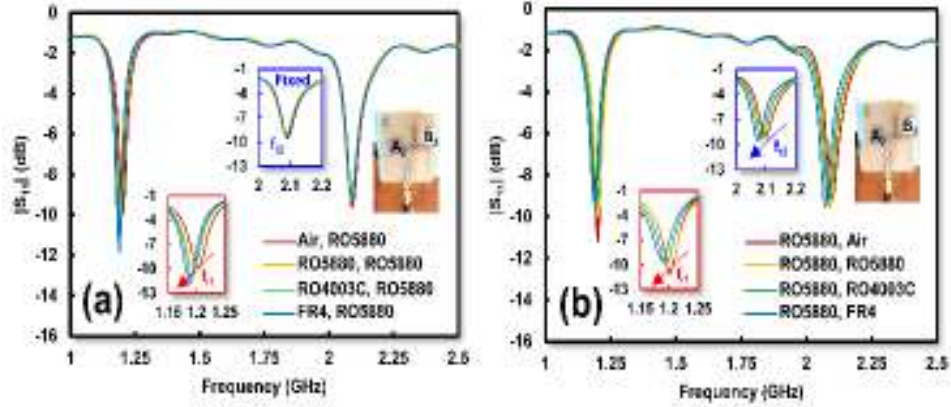


Figure 4.25 Measurement result of 1st and 2nd resonator from C₆; (a) response of f_{r1} and f_{r2} during scenario (1), (b) response of f_{r1} and f_{r2} during scenario (2)

Furthermore, Figure 4.26 (a) and Figure 4.26 (b) show the measurement results of C₈ with scenario (1) and (2). It can be clearly seen that f_{r1} and f_{r2} have high independence for both conditions while do not affect each other when the MUT is loaded simultaneously. Therefore, S_8 was used to determine the correlation between the two resonant frequencies and the MUT.

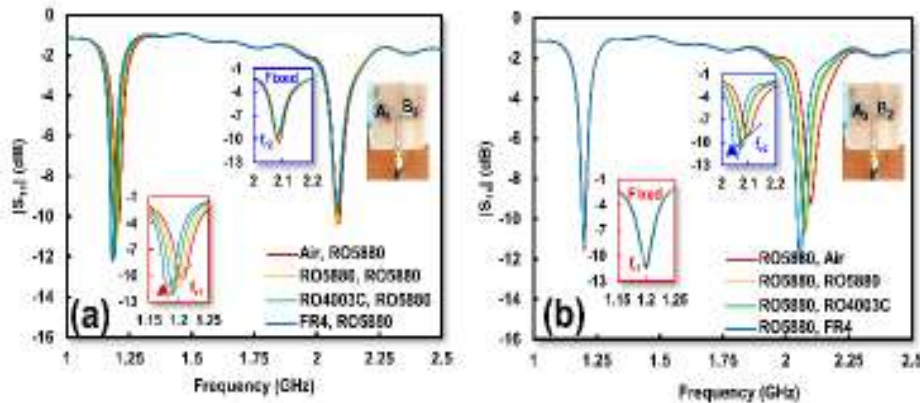


Figure 4.26 Measurement result of 1st and 2nd resonator from C₈; (a) response of f_{r1} and f_{r2} during scenario (1), (b) response of f_{r1} and f_{r2} during scenario (2)

In order to determine the correlation of the 1st and 2nd resonators, Pearson correlation model was used as an approximation. Partial correlation was used to observe the relationship between f_{r1} and f_{r2} as shown in Figure 4.25.

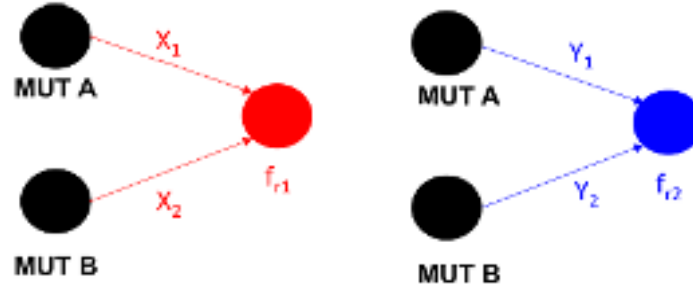


Figure 4.27 Partial correlation model for f_{r1} and f_{r2}

where X_1 and X_2 is the correlation between MUT A and B with f_{r1} while Y_1 and Y_2 represents the correlation between MUT A and B with f_{r2} when loaded simultaneously. The correlation between the MUT and the two resonant frequencies can be determined based on equation 3.41. To determine the level of correlation, the correlation coefficient scale was used which is mentioned in Table 3.6. Furthermore, Table 4.8 shows the correlation coefficient based on proposed scenario dan conditions.

Table 4.8 Correlation coefficient of proposed sensor

Condition	Correlation coefficient			
	X_1	X_2	Y_1	Y_2
C_1	0.94	0.07	0	0.92
C_2	0.94	0.14	0	0.40
C_3	0.49	0.10	0.59	0.86
C_4	0.90	0.03	0.001	0.90
C_5	0.90	0.08	0.001	0.91
C_6	0.72	0	0.56	0.99
C_7	0.97	0.07	0.06	0.88
C_8	0.97	0.01	0.076	0.93
C_9	0.85	0.08	0.42	0.90

It is clearly seen that the correlation of the two resonators is strongly influenced by the placement scenario of the MUT as shown in Table IV. The best conditions are obtained on the C₈ where the correlation of MUT A with f_{r1} is very high with $X_1 = 0.97$ while the correlation of MUT B with f_{r1} is very low with $X_2 = 0.01$. Furthermore, the correlation of MUT A with f_{r2} is also very low represented by $Y_1 = 0.076$ while the correlation of f_{r2} with MUT B is very high and represented by $Y_2 = 0.97$.

These results indicate that the 1st resonator and 2nd resonator have high independence indicated by a very low correlation (r) when both resonators are loaded MUT simultaneously. In this paper, nine conditions have been applied to determine the location of the MUT and S₈ is proposed to be the location of the MUT for the two resonators that have the highest independence.

4.7 Comparison with existing works

Based on the measurements and demonstrated simulation results, it is possible to measure permittivity of different MUT using a pair of U-Shaped resonators independently with different sensing areas. In addition, the proposed sensor also had an excellent accuracy both of using permittivity from datasheet and calibrated frequency from simulation. Table V shows the comparison between the previous sensors with the proposed sensor. The parameters for comparison include models, range of permittivity, MUT, accuracy, number of sensing areas, number of ports, and independence performance. In (Armghan et al., 2021)-(Haq et al., 2019), the researchers proposed dual-band microwave sensors to measure the permittivity of the material using CSRRs, complementary bisymmetric-SRR and SRR. However, the previous sensor only had one sensing area to place

the MUT, thus it could not be used to perform independent measurements. Whilst (Ebrahimi, Scott and Ghorbani, 2020a)-(Ebrahimi, Scott and Ghorbani, 2018a) proposed a microwave sensor using magnetic LC, microstrip line ring resonator SSRR with spur lines and SRR with dual-port to measure the permittivity of the material. Another literature review, proposes reflective microwave sensors (Ebrahimi, Scott and Ghorbani, 2018b, 2020b; Behdani et al., 2020) to detect permittivity of solid and liquid materials. However, the existing microwave sensor could not be used to measure the permittivity of the MUT independently due the high dependency performances of each band and using dual-port connected concurrently resulting in wastage of resources. Therefore, the proposed sensor depicted in Table V exhibits high accuracy of permittivity detection of MUTs for both resonances of the resonator. Furthermore, the proposed sensor can detect the permittivity of two different types of MUTs independently with different sensing areas.

This study proposes a new approach to detect the permittivity of the MUT independently with different sensing areas. The U-shaped technique is proposed to generate different sensing areas to place two types of MUT independently. Furthermore, the proposed sensor only uses single-port to make it simpler and easier to fabricate. Furthermore, nine condition for the placement of MUTs are proposed to determine the sensing area area that has high independence. The independence of the two resonators was observed and determined using the Pearson correlation when the MUT was loaded on the resonator simultaneously. This sensor has been used to characterize standard materials with known permittivity (Air, RO5880, RO4003C and FR4). From the measurement results, the correlation between MUT and each resonant frequency is very low which is represented by $X_2 = 0.010$ and $Y_1 = 0.076$ in scenario 8 (S_8). This shows that the proposed dual band sensor

has high independence and does not affect each other when the MUT is loaded simultaneously. Therefore, the proposed sensor can be an ideal solution for applications in the food, quality control, biomedical and pharmaceutical industries.

Table 4.9 Comparison of proposed sensor based on previous works

Reference	Model	Range of permittivity	Material Under Test (MUT)	Resonance frequency (GHz)		Accuracy (%)		NS (%)		Num. Of sensing area	Num. of ports	High independence performance
				f_{r1}	f_{r2}	f_{r1}	f_{r2}	f_{r1}	f_{r2}			
(Armghan et al., 2021)	CSRRs	1 – 11.9	Solid	5.35	7.99	94.24	97.76	3.40	3.53	1	2	No
(Haq et al., 2019)	CBS-SRR	1 – 12.9	Solid	4.50	6.79	99.03	98.07	3.59	3.12	1	2	No
(Al-Behadili et al., 2020)	SRR	1 – 6	Solid	1.24	2.08	98.18	96.97	4.10	3.93	1	2	No
(Ebrahimi, Scott and Ghorbani, 2020a)	Magnetic LC	1 – 10.7	Solid	1.62	2.38	95.2	-	3.04	4.57	1	2	No
(Lim, Kim and Hong, 2018)	Microstrip Line Ring Resonator	1 – 3.48	Solid	4.89	9.81	92.5	-	2.59	2.68	1	2	No
(Rammah Ali Alahnomi et al., 2017)	SSRR with spurlines	1 – 4.4	Solid	2.22	-	98.36	-	1.34	-	1	2	No
(Ansari, Jha and Akhtar, 2015)	CSRR	1 – 3	Solid	2.65	-	97.71	-	5.38	-	1	2	No
(Ebrahimi, Scott and Ghorbani, 2018a)	SRR	1 – 10.7	Solid	2.10	-	98.24	-	3.11	-	1	2	No

Reference	Model	Range of permittivity	Material Under Test (MUT)	Resonance frequency (GHz)		Accuracy (%)		NS (%)		Num. Of sensing area	Num. of ports	High independence performance
				f_{r1}	f_{r2}	f_{r1}	f_{r1}	f_{r2}	f_{r1}			
(Ebrahimi, Scott and Ghorbani, 2020b)	CSRR	1 – 80	Liquid	2.48	-	99.75	-	-	-	1	1	No
(Ebrahimi, Scott and Ghorbani, 2018b)	LC Resonator	1 – 10.2	Solid	2.41	-	99.90	-	3.28	-	2	2	No
(Behdani et al., 2020)	Aperture coupling	1 – 12.85	Solid	9.54	12.30	-	92.3	0.64	-	1	1	No
(Abdolrazzaghi and Daneshmand, 2017)	Dual Active SRR	1 – 20	Liquid	1.03	1.49	-	-	-	-	2	2	No
This work	Dual U-Shaped Resonator	1 – 4.3	Solid	1.21	2.10	99.02	96.44	0.76	1.15	2	1	Yes

4.8 Concluding remarks

In this chapter, a dual-band independent permittivity sensor using single-port with a pair of U-shaped structures for solid material detection was successfully presented. The sensor included a pair of U-shaped resonators operating in the resonant frequencies of $f_{r1} = 1.200$ GHz and $f_{r2} = 2.100$ GHz using FR4 substrate with single-port. The U-shaped technique is proposed to generate different sensing areas to place two types of MUT independently. The proposed sensor had a very high accuracy compared with permittivity from datasheet. A brief comparison of the proposed work with existing sensors reported from the literature shows that the proposed sensor had the advantage of measuring the permittivity of the MUT independently with very high accuracy and two different sensing areas. Therefore, the proposed sensor can be an ideal solution for applications in the food industry, quality control and biomedical.

However, the radiation of proposed resonator is still low, so it is necessary to optimize it so that it can transmit electromagnetic waves. In addition, the sample being characterized is still placed directly on the surface of the sensor so that it can be corrosive and damage the surface of the resonator. Therefore, a resonator that has high radiation characteristics is needed so that the sensor can be integrated with the antenna and can be used to characterize materials and transmit electromagnetic waves simultaneously.

CHAPTER 5

INTEGRATED MICROWAVE SENSOR AND ANTENNA SENSOR BASED ON DUAL T-SHAPED RESONATOR

In the previous chapter 4, dual-band independent permittivity sensor using single-port with a pair of U-shaped structures for solid material detection was successfully presented. However, the radiation of proposed resonator is still low, so it is necessary to optimize it so that it can transmit electromagnetic waves. In addition, the sample being characterized is still placed directly on the surface of the sensor so that it can be corrosive and damage the surface of the resonator. Therefore, a resonator that has high radiation characteristics is needed so that the sensor can be integrated with the antenna and can be used to characterize materials and transmit electromagnetic waves simultaneously. This chapter presents the development of integrated microwave sensor (MS) and antenna sensor (AS) based on dual T-shaped resonator for contact and non-contact detection of solid material. The discussion in this chapter begins with an introduction to explain the main contribution of the proposed work, design and simulation of a dual T-shaped resonator, determining the location of sensing areas for contact and non-contact detection, scenarios for placing MUTs for contact and non-contact detection.

The discussion continued with observing the performance of the microwave sensor and antenna sensor including the accuracy, sensitivity and effect of an air gap for non-contact detection using proposed sensor while characterize the MUT. Furthermore, the final discussion of this chapter is validating the performance of the proposed sensor through a

measurement process and comparing it with existing work to show the novelty of the proposed work.

5.1 Introduction

This chapter describes of integrating contact and non-contact detection for the characterization of solid materials based on MS and AS. The sensors are proposed using dual T-Shaped resonators with a single port that operate at two different resonant frequencies at $f_{r1} = 1.81$ GHz and $f_{r2} = 2.35$ GHz for MS and AS, respectively. Furthermore, the proposed sensor has independent characteristics and different sensing area locations for contact and non-contact detection so that the characterization of the MUT can be carried out separately where the low frequency (f_{r1}) is used as MS for non-contact detection while high frequency (f_{r2}) is used as AS. The proposed sensor has been successfully simulated, modelled using EM simulation, and validated by the measurement process. The range of permittivity detected is 1 - 6.15 for MS and AS, respectively. Furthermore, the sensitivity of the AS with non-contact detection can be evaluated based on the air gap (d) of 0.5 - 1.5 mm.

5.2 Design and simulation of dual T-shaped resonator

The design and simulation of the dual T-shaped resonator was carried out using HFSS 15.0 while the equivalent circuit used AWR 2009. The stages and simulation process are explained in the next sub-chapter.

5.2.1 Dual T-shaped resonator with single port

The proposed resonator consists of two T-shape resonators connected directly to a microstrip line with an impedance of $50\ \Omega$. The dual T-shaped resonator is designed to operate at $f_{r1} = 2.35\text{ GHz}$ and $f_{r2} = 1.57\text{ GHz}$. The development stages of the dual T-shaped resonator have been described in Figure 3.15 in chapter 3. The development consists of two stages which indicate the resonant frequency of each resonator. It should be noted, the resonant frequency of each resonator is determined based on the length of the arm of the T-shape resonator. Furthermore, the resonant frequency of each resonator can be determined based on the transmission line model.

Based on the similar approach as the dual U-shaped resonator described in chapter 4, the resonant frequency can be determined based on the length of the arm of the resonator which has a high concentration of electric field. The E-field simulation results of the dual T-shape resonator at $f_{r1} = 2.35\text{ GHz}$ and $f_{r2} = 1.57\text{ GHz}$ are shown in Figure 5.1 (a) and Figure 5.1 (b)

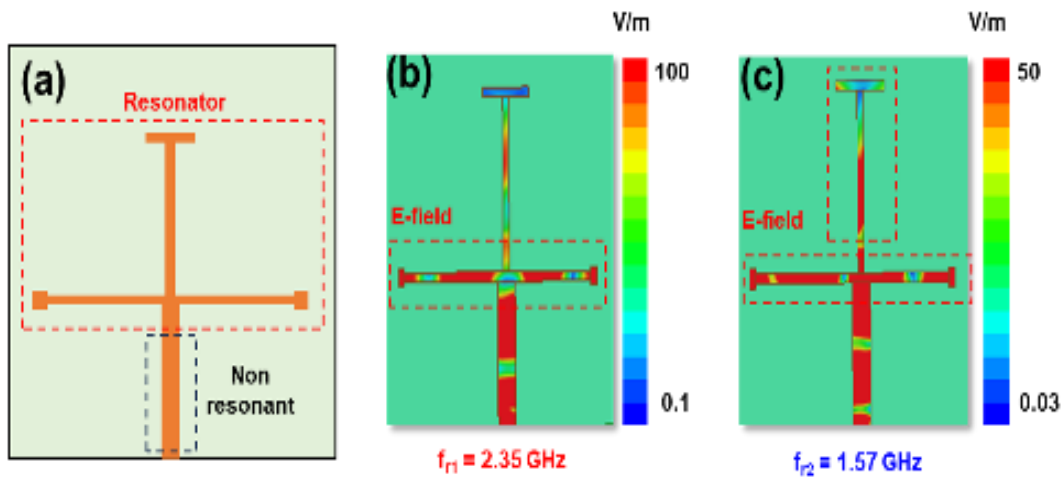


Figure 5.1 Structure and simulation of E-field of dual T-shaped resonator; (a) structure of dual T-shaped resonator, (b) E-field at $f_{r1} = 2.35\text{ GHz}$, (c) E-field at $f_{r2} = 1.57\text{ GHz}$

Figure 5.1 (a) shows the structure of a dual T-shape resonator consisting of a 1st and 2nd resonator connected by a microstrip line with an impedance of 50 Ω . The highest concentration of the electric field of the resonator at $f_{r1} = 2.35$ GHz is around the arms of the 1st resonator as shown in Figure 5.1 (b) while for $f_{r2} = 1.57$ it is around both arms of the 1st and 2nd resonator as shown in Figure 5.1 (c). These findings indicate that the two resonators work together to generate a second resonance frequency. Therefore, the resonant frequency of the resonator can be determined using a transmission line model based on the arm of the resonator which has a high electric field concentration at each resonant frequency as shown in Figure 5.2.

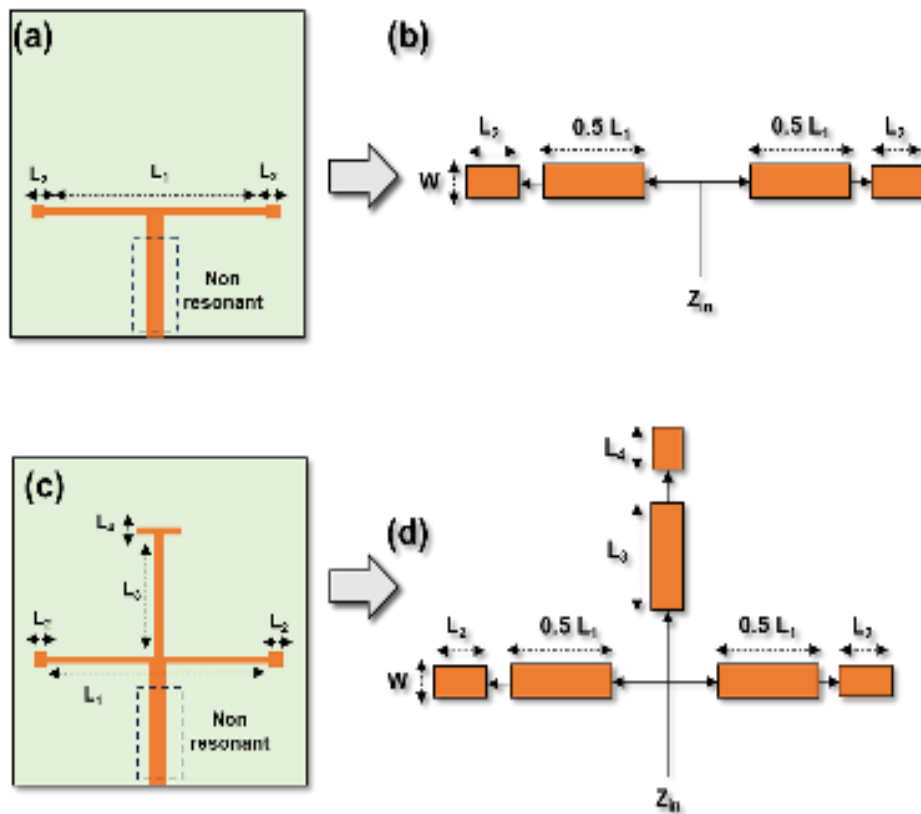


Figure 5.2 Structure and simulation of E-field of dual T-shaped resonator; (a) structure of dual T-shaped resonator, (b) E-field at $f_{r1} = 2.35$ GHz, (c) E-field at $f_{r2} = 1.57$ GHz

Figure 5.2 (a) and Figure 5.2 (c) show the structure of the resonator in the 1st stage and 2nd stage which is represented by L_1 , L_2 , L_3 and L_4 which are connected to Z_{in} . The dual T-shape resonator is connected to a non-resonant microstrip line with an impedance of 50 Ω . The transmission line model of the 1st and 2nd stages is shown in Figure 5.1 (b) and Figure 5.1 (d).

Based on equation 3.14 and equation 3.15, the width of microstrip line (W) with impedance of 50 Ω can be determined as follows:

$$B = \frac{377\pi}{2Z_0\sqrt{\epsilon_r}} = \frac{(377)(3.14)}{2(50)\sqrt{4.3}} = 5.72$$

$$W = \frac{2}{\pi} \left[B - 1 - \ln(2B - 1) + \frac{\epsilon_r - 1}{2\epsilon_r} \left\{ \ln(B - 1) + 0.39 - \frac{0.61}{\epsilon_r} \right\} \right]$$

$$W = \frac{2}{3.14} \left[(5.72) - 1 - \ln(10.44 - 1) + \frac{4.3 - 1}{2(4.3)} \left\{ \ln(5.72 - 1) + 0.39 - \frac{0.61}{4.3} \right\} \right]$$

$$= 3.1 \text{ mm}$$

Based on the calculation, the width of the microstrip line (W) is 3.1 mm with an impedance (Z_0) of 50 Ohm. The length of the arm of the resonator is obtained from the optimization process using EM simulation software which represents the resonant frequency of the resonator as shown in Table 5.1.

Table 5.1 Optimize dimensions of the arm of the proposed resonator

Parameter	Dimension (mm)
L_1	32
L_2	1
L_3	18
L_4	1
w	1

Furthermore, the resonant frequency of each resonator can be determined based on the equation 3.4 and equation 3.45 as follows:

$$\varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left[1 + 12 \frac{1.6}{1} \right]^{-\frac{1}{2}}$$

$$\varepsilon_{eff} = \frac{4.3+1}{2} + \frac{4.3-1}{2} \left[1 + 12 \frac{1.6}{3.1} \right]^{-\frac{1}{2}} = 3.017$$

$$f_{r1} = \frac{nc}{2(L_1+2L_2)\sqrt{\varepsilon_{eff}}}$$

$$f_{r1} = \frac{(1)(3 \times 10^8)}{2(32+2)\sqrt{3.017}} = 2.54 \text{ GHz}$$

$$f_{r2} = \frac{nc}{2(L_1+2L_2+L_3+L_4)\sqrt{\varepsilon_{eff}}}$$

$$f_{r2} = \frac{(1)(3 \times 10^8)}{2(32+2+18+1)\sqrt{3.017}} = 1.62 \text{ GHz}$$

To validate the calculations, a simulation with EM simulation software is proposed to observe the resonant frequency and performance of the resonator.

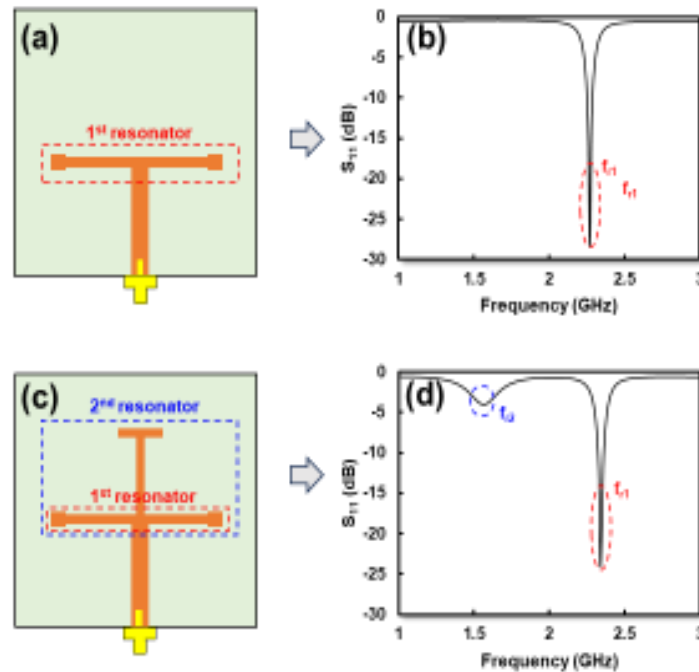


Figure 5.3 Simulation result of S_{11} using the transmission line model; (a) 1st resonator, (b) 2nd resonator

The resonant frequency of each resonator can be modelled and simulated using EM simulation based on the arm of the resonator with the highest electric field concentration as shown in Figure 5.3 (a) and Figure 5.3 (c). Moreover, Figure 5.3 (b) shows the simulation results of the first resonator with L_1 and L_2 operating at a resonant frequency of 2.44 GHz while the second resonator with L_1 , L_2 , L_3 and L_4 operating at a resonant frequency of 1.57 GHz as shown in Figure 5.3 (d).

This result shows that the simulation results using the EM simulation software are in line with the calculation where the resonator operates at a dual frequency of around 1.6 GHz and 2.5 GHz. These findings indicate that the transmission line approach can be used to determine the resonant frequency of a dual U-shape resonator.

5.2.2 Structure of dual T-shaped resonator with single port

The proposed sensor was designed using an FR-4 substrate with a permittivity of 4.3, a loss tan of 0.0265, and a thickness (h) of 1.6 mm. The resonator consists of two T-shape resonators which represent their respective resonant frequencies as described in sub-chapter 5.2. The structure and dimensions of the proposed sensor are shown in Figure 5.4(a) and Figure 5.4 (b) using a dual T-shaped resonator connected directly to the SMA connector via a microstrip line with an impedance of 50 Ω . In this study, the resonator is designed to operate at $f_{r1} = 2.35$ GHz and $f_{r2} = 1.57$ GHz. The overall dimensions of the proposed sensor are shown in Table 5.2.

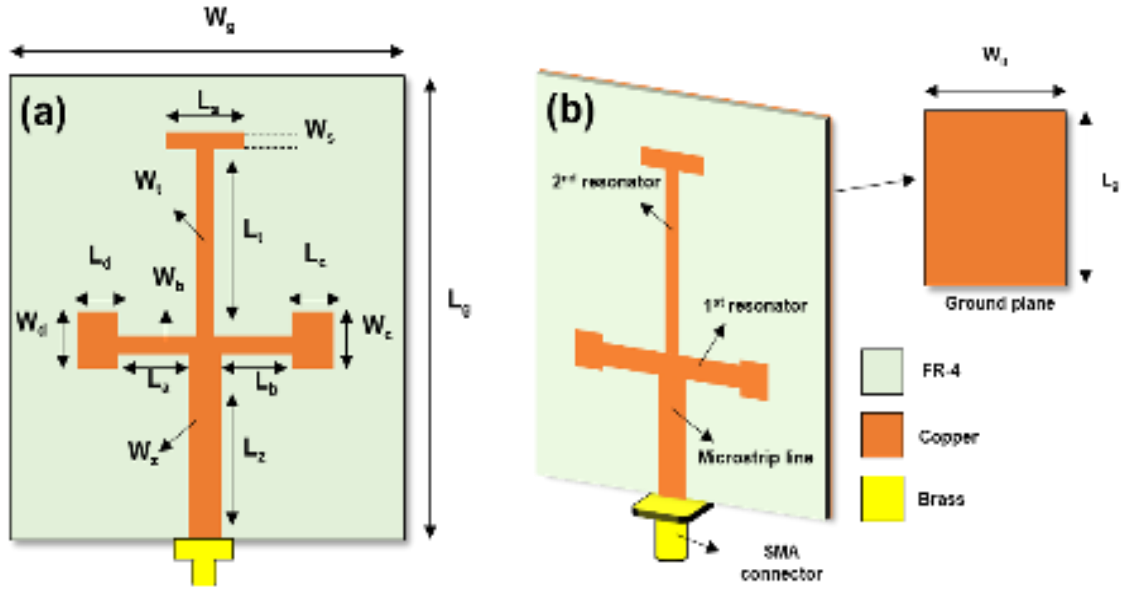


Figure 5.4 Structure of dual T-shaped resonator; (a) from front view (b) from perspective view

Table 5.2 Dimension of dual T-shaped resonator

Parameter	Value (mm)	Parameter	Value (mm)
W_z	3 mm	W_d	2 mm
L_z	15 mm	W_t	1 mm
L_a	16 mm	L_t	18 mm
L_b	3 mm	W_s	1 mm
L_c	15 mm	L_s	8 mm
L_d	16 mm	W_g	50 mm
W_c	13 mm	L_g	50 mm
W_z	3 mm	W_b	1 mm

Furthermore, the equivalent circuit model using $\mathbf{R}$ as resistors, $\mathbf{L}$ as inductors, and $\mathbf{C}$ as capacitors of the proposed resonator is shown in Figure 5.5 (a) where the 1st resonator is represented by $\mathbf{R}_1 = 1.341 \text{ k}\Omega$, $\mathbf{L}_1 = 3.010 \text{ nH}$ and $\mathbf{C}_1 = 1.260 \text{ pF}$ while the 2nd resonator is represented by $\mathbf{R}_2 = 0.379 \text{ k}\Omega$, $\mathbf{L}_2 = 5.530 \text{ nH}$ and $\mathbf{C}_2 = 1.530 \text{ pF}$. To control the impedance of the resonator, the transmission line was modeled and represented by $\mathbf{Z}_1 = 50 \text{ }\Omega$ with $\theta = 55.40$ which was connected to port 1 with an impedance of $50 \text{ }\Omega$. Furthermore, $\mathbf{C}_{g1} = 0.299$

pF and $C_{g2} = 0.378$ pF were proposed to prevent a short circuit between the resonator and the connector.

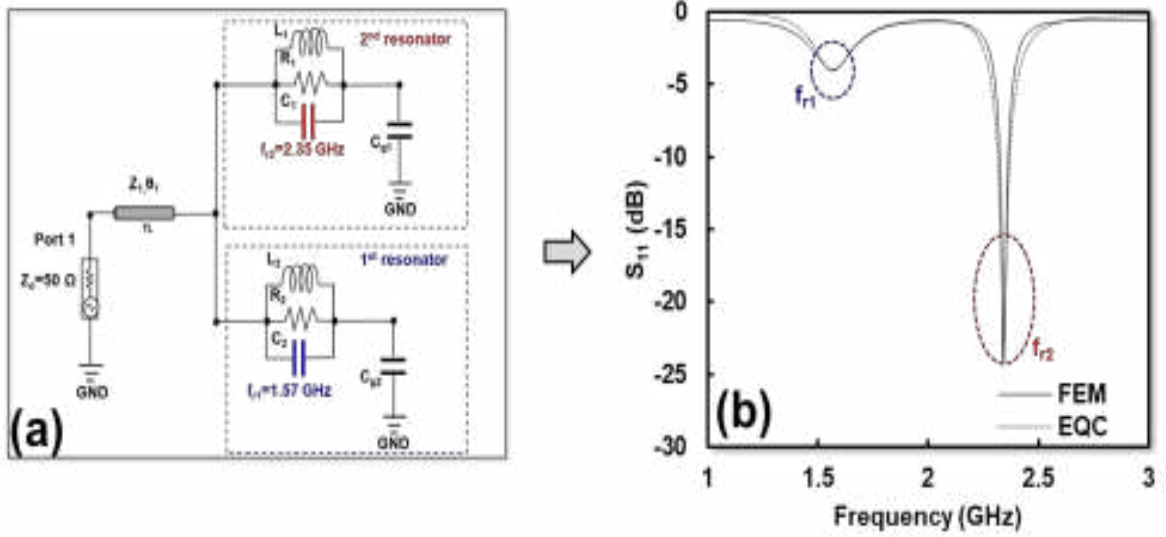


Figure 5.5 (a) Equivalent circuit model of dual T-shaped resonator, (b) simulation result of EQC vs FEM

Based on calculations using the equivalent circuit model, the resonators operate at $f_{r1} = 2.32$ GHz and $f_{r2} = 1.55$ GHz, respectively. These findings indicate that the equivalent circuit can be used to represent the proposed resonator and operates at frequencies around 1.2 and 2 GHz. Furthermore, Figure 5.5 (b) shows a comparison of the simulation based on Finite Element Model (FEM) and equivalent circuit model. It can be clearly seen that the experimental results obtained show good agreement for element model and full-wave EM simulation.

5.3 Potential location sensing area of dual T-shaped resonator

The sensing area was determined based on the concentration of the E and H-fields at each resonator. In this experiment, the concentration of the E and H-fields of proposed

resonator was simulated using HFSS 15.0 based on resonant frequency of $f_{r1} = 2.35$ GHz and $f_{r2} = 1.57$ GHz

Both resonators had different E- and H-field concentrations based on their respective resonant frequencies. The permittivity of the MUT was detected by placing the material at the sensing area location. The simulation results of the E- and H-field concentrations are shown in Figure 5.6(a), Figure 5.6(b), Figure 5.6 (c) and Figure 5.6(d).

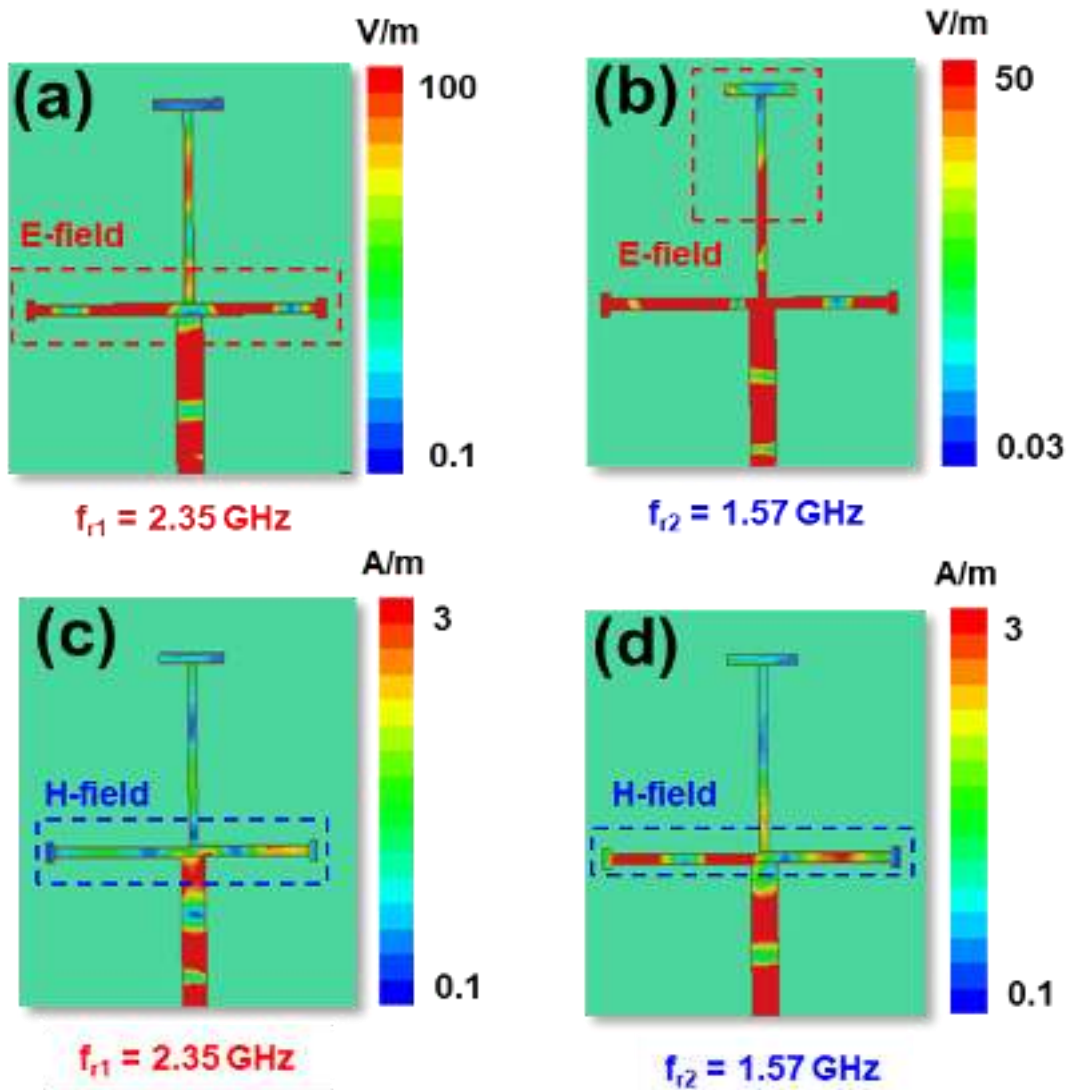


Figure 5.6 Simulation result of E and H-field concentration; (a) E-field at $f_{r1} = 1.57$ GHz, (b) E-field at $f_{r2} = 2.35$ GHz, (c) H-field at $f_{r1} = 1.57$ GHz, (d) H-field at $f_{r2} = 2.35$ GHz

Based on the simulation results using HFSS 15.0, the concentration of the E-field of the 1st resonator with $f_{r1} = 1.57$ GHz is in the open-ended stub while the H-field is in the middle arm of the 2nd resonator as shown in Figure 5.6 (a) and Figure 5.6 (c). These results show that the 1st resonator can function as MS with the sensing area located on the open-ended stub. On the other hand, the E-field and H-field concentrations of the 2nd resonator with $f_{r2} = 2.35$ GHz were in the same location on the middle arm of the proposed sensor as shown in Figure 5.6(b) and Figure 5.6 (d). These results indicate that the 2nd resonator can function as AS with a sensing area location on the middle arm of the proposed device where the maximum concentration of E-field and H-field was obtained. To validate the function of both resonators, a simulation of the radiation pattern was proposed to observe the performance of each as shown in Figure 5.7 (a) and Figure 5.7 (b).

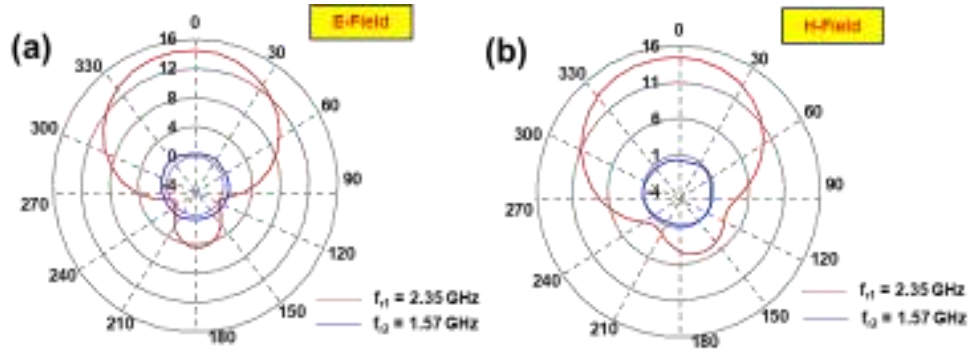


Figure 5.7 Radiation pattern of proposed resonator; (a) E-plane at $f_{r1} = 2.35$ GHz and $f_{r2} = 1.57$ GHz, (b) H-plane at $f_{r1} = 2.35$ GHz and $f_{r2} = 1.57$ GHz

Based on the simulation results, the radiation from the E and H fields of $f_{r1} = 2.45$ GHz is above 10 dB while for $f_{r2} = 1.57$ GHz it is around 1 dB. The simulation of the radiation pattern of the E-field and H-field of both resonators showed that the 1st resonator had maximum radiation compared to the 2nd resonator. These results confirmed that the 1st resonator can be proposed as AS while the 2nd resonator can be proposed as MS.

5.4 Scenario of contact and non-contact detection

The proposed sensor consists of dual T-shaped resonators with a single port operating at two resonance frequencies, the 1st resonator is $f_{r1} = 2.35$ GHz, and the 2nd resonator is $f_{r2} = 1.57$ GHz as mentioned in Figure 5.6. The low frequency is proposed as MS while the high frequency is proposed as AS based on reflection coefficient and radiation pattern based on resonant frequency as mentioned in chapter 3, sub-chapter 3.4.2. The proposed resonator has independent characteristics with different sensing areas so that contact and non-contact detection can be carried out separately. The placement of the MUT based on the location of the sensing area of the proposed resonator is shown in Figure 5.8.

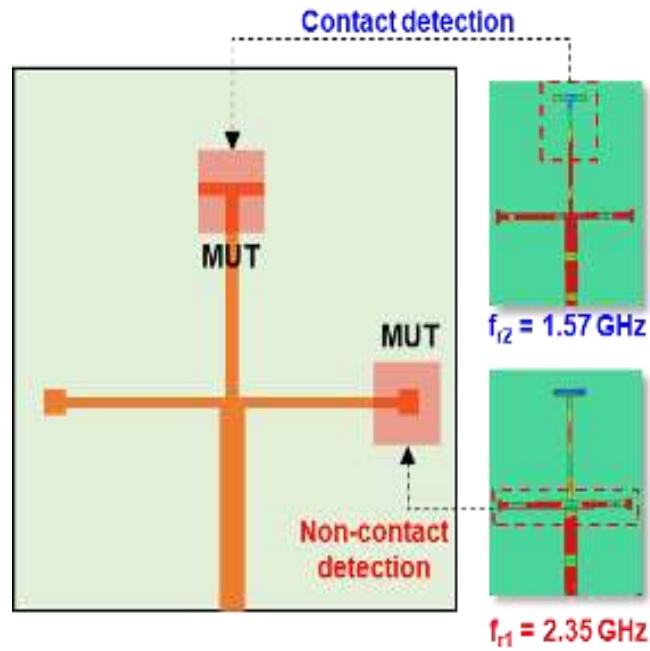


Figure 5.8 The placement of MUT based on sensing area of proposed sensor

MS and AS were determined based on the frequency shift on the reflection coefficient (S_{11}) where S_{11} of ≥ -10 dB was proposed as MS and S_{11} of ≤ -10 dB was proposed as AS. The sensing area was determined based on the concentration of the electric **E** and **H**-

fields of each resonator. Furthermore, the surface of the resonator which had a high concentration of E-field and a low concentration of H-field was used as an MS for contact detection. On the other hand, non-contact detection utilized the surface of the resonator which has the same location as E and H-fields. Generally, resonators that function as antennas transmit electromagnetic waves where the E and H-fields were in the same location, therefore it can be utilized for AS. The permittivity of the MUT was observed based on the shift in the resonant frequency of the resonator when the MUT was placed at each sensing area. The scenario of a contact and non-contact detection using MS and AS has been described in Figure 5.8 where the permittivity of the MUT was ϵ_r MUT, the air gap was ϵ_r gap (vacuum = 1), and the substrate of the resonator was ϵ_r sub.

5.5 Simulation of dual T-shaped resonator for contact and non-contact detection

The simulation process was carried out using HFSS 15.0 where two scenarios were proposed to detect the permittivity of MUTs both in contact and non-contact:

- (1) MUT was placed on the sensing area of the 1st resonator directly without an air gap as MS for contact detection while the 2nd resonator is in a vacuum.
- (2) MUT is placed on the sensing area of the 2nd resonator with an air gap as AS for non-contact detection while the 1st resonator is in a vacuum.

The placement of MUT and simulation results of contact detection using MS during scenario (1) with a permittivity range of 1 - 10 are shown in Figure 5.9 (a) and Figure 5.9 (b), respectively.

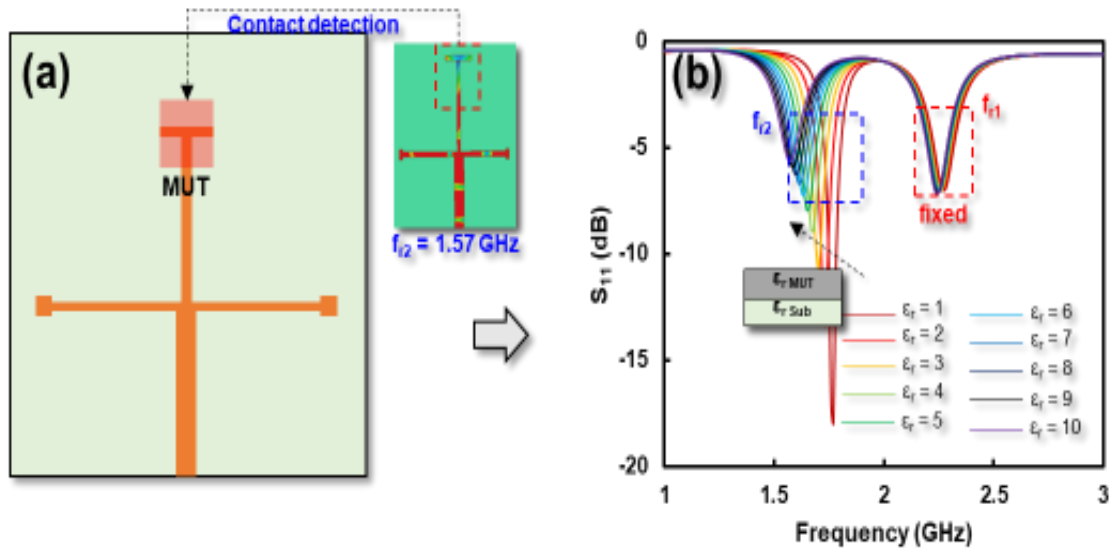


Figure 5.9 (a) Placement of MUT based on scenario (1), (b) simulation result during scenario (1)

The resonant frequency of MS (f_{r2}) for contact detection moves to the lower frequency from 1.770 GHz to 1.570 GHz while f_{r1} does not shift significantly while scenario (1) is applied as shown in Figure 5.9 (b). The placement of MUT and simulation result during scenario (2) are shown in Figure 5.10 (a) and Figure 5.10 (b).

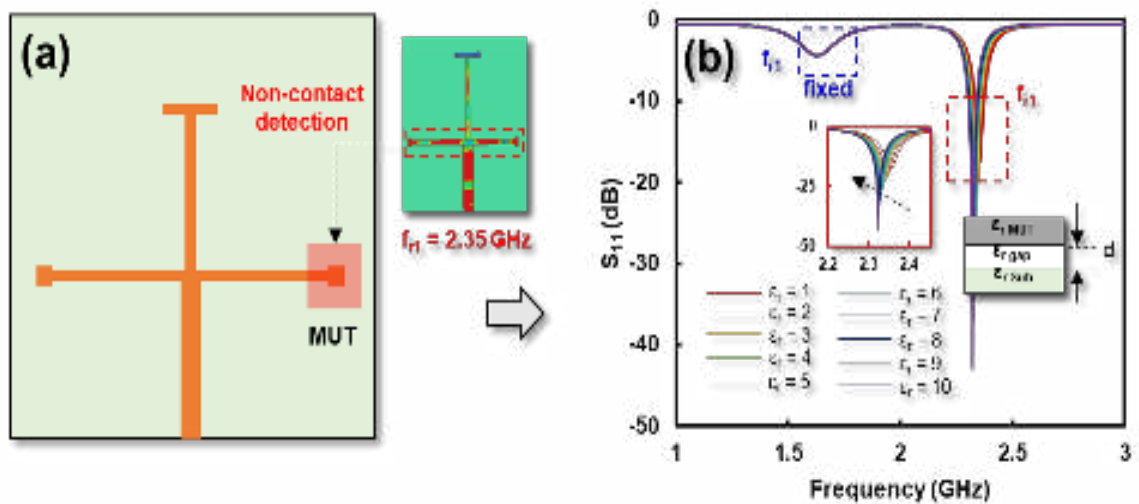


Figure 5.10 (a) Placement of MUT based on scenario (2), (b) simulation result during scenario (2)

Furthermore, the same characteristics were also obtained for non-contact detection using AS where f_{r2} moved to the lower frequency from 2.355 GHz to 2.322 GHz when the MUT was placed in non-contact with an air gap (d) of 0.5 mm as shown in Figure 5.10 (b) while f_{r1} does not shift significantly while scenario (2) is applied. This indicated that the two resonators had independent characteristics both in contact and non-contact detection. The correlation between f_{r1} and f_{r2} when the MUT loaded during scenario (1) and scenario (2) is shown in Figure 5.11 (a) and Figure 5.11 (b).

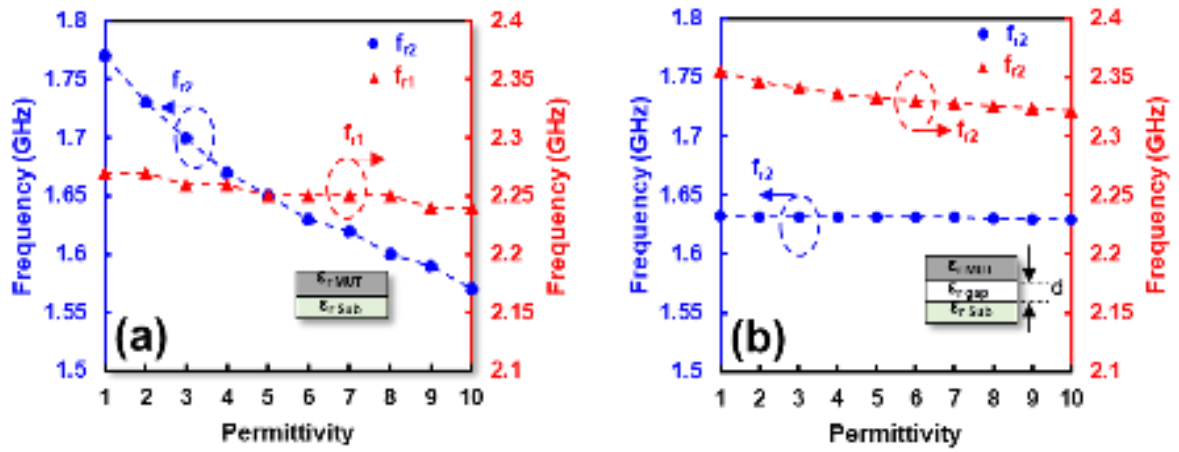


Figure 5.11 Correlation of f_{r1} and f_{r2} ; (a) during scenario (1), (b) during scenario (2)

The Δf and sensitivity of the resonator could be determined by comparing the difference between the resonant frequency of the lowest and highest permittivity of the MUT based on equation 3.48. The Δf and sensitivity of proposed sensor shown in Figure 5.12 (a) and Figure 5.12 (b) where the MS has a maximum sensitivity of 40 MHz / $\Delta \epsilon_r$ and Δf of 0.2 GHz / $\Delta \epsilon_r$ while AS is 9 MHz / $\Delta \epsilon_r$ and Δf of 0.033 GHz / $\Delta \epsilon_r$. These results indicate that MS were more sensitive than AS.

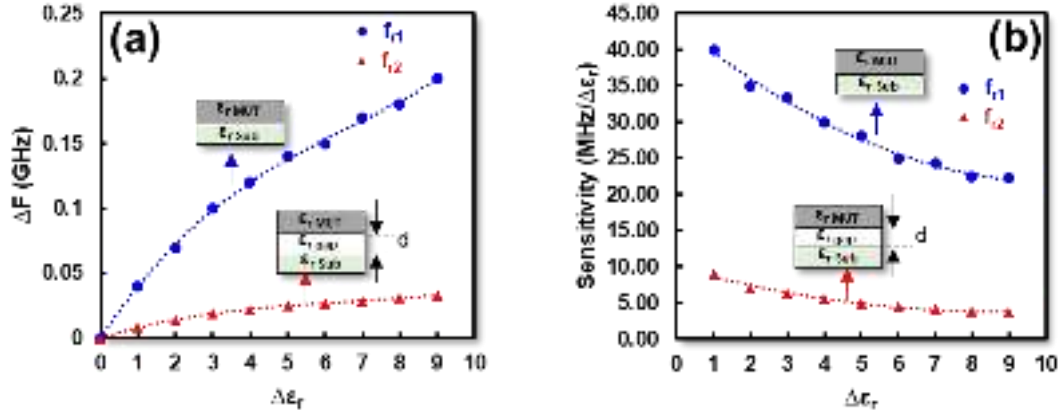


Figure 5.12 The Δf and sensitivity of proposed sensor based on simulation result; (a) Δf , (b) sensitivity

In this experiment, the effect of changes in $\tan \delta$ on both resonators was simulated and observed both for contact and non-contact detection as shown in Figure 5.13 (a) and Figure 5.13 (b) with a range of 0 – 0.1.

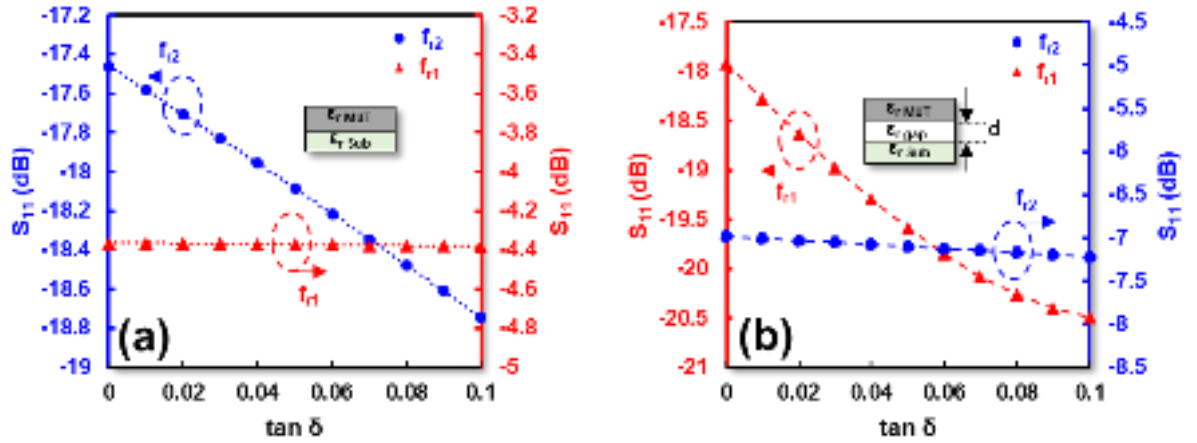


Figure 5.13 (a) Effect of $\tan \delta$ for MS, (b) Effect of $\tan \delta$ for AS with an air gap

Based on the simulation, MS had a better sensitivity for detecting changes in $\tan \delta$ compared to AS as shown in Figure 5.13 (a) and Figure 5.13 (b) with a shift in the S_{11} from -17.91 dB to -20.49 dB for contact detection while non-contact detection is -17.46 dB to -

18.74 dB. This finding shows that the proposed sensor had independent characteristics both with MS and AS.

In order to observe the effect of the air gap on non-contact detection, several experiments were carried out by changing the distance between the MUT and the sensor which is separated by an air gap (d). In this experiment, the distance from the air gap (d) observed was in the range of 0.5 - 3 mm while the observed parameters were Δ_f and sensitivity. Figure 5.14 (a) and Figure 5.14 (b) show the simulation results of Δ_f and sensitivity for an air gap of 0.5 - 3 mm.

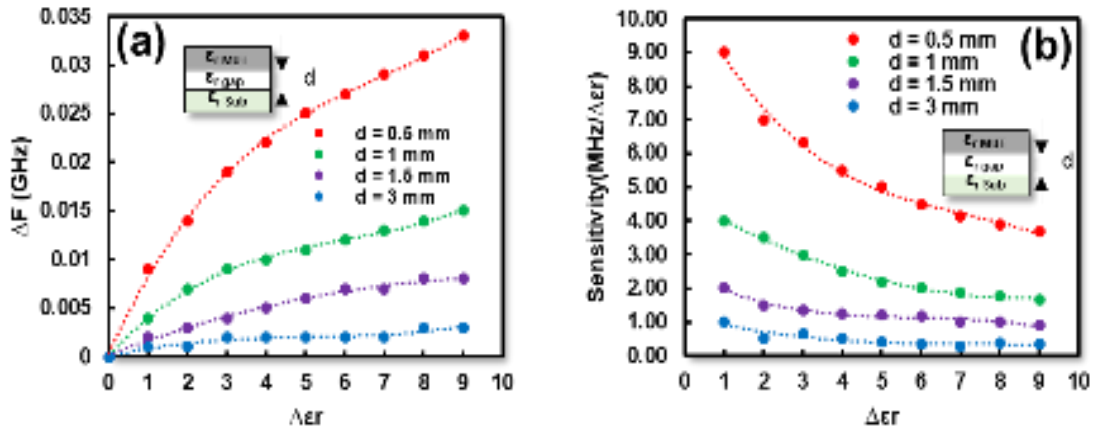


Figure 5.14 Effect of air gap for non-contact detection; (a) effect air gap with Δ_f , (b) effect air gap with sensitivity

Based on the simulation results with an air gap of 0.5 mm - 3 mm, the sensitivity range was 3.67 – 9 MHz / $\Delta\epsilon_r$ while for Δ_f was 0 - 0.033 GHz / $\Delta\epsilon_r$ as shown in Figure 5.14 (a) and Figure 5.14 (b). These findings showed that Δ_f and the sensitivity of AS for non-contact detection were greatly affected by the distance from the air gap (d). In other words, AS was more sensitive for non-contact detection with small air gaps. Furthermore, the effect of the air gap on the detection of tan delta is shown in Figures 5.15 (a) and Figure 5.15 (b).

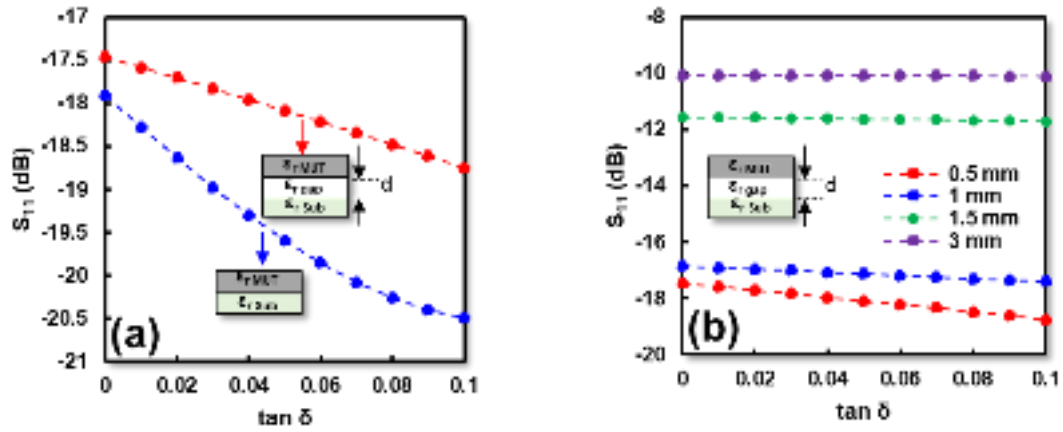


Figure 5.15 (a) Effect of $\tan \delta$ for AS and MS, (b) Effect of $\tan \delta$ for AS with an air gap of 0.5 – 3 mm

Another finding shows that the gap between the sensor and the MUT of AS greatly affected the value of S_{11} as shown in Figure 5.15 (a) and Figure 5.15 (b) where the S_{11} increased for small gap distances. Furthermore, S_{11} of AS shifted from -17.46 dB to -10.06 dB in line with changes in the air gap distance between the sensor and the MUT in the range of 0.5 mm - 3 mm. The best performance of AS was obtained with an air gap of 0.5 mm.

5.6 Fabrication and measurement of dual T-shaped resonator

Verification of the resonator shape was carried out in the laboratory using a Vector Network Analyzer with an ambient temperature of 25°. The stages and measurement process are explained in the next sub-chapter.

5.6.1 Measurement setup of MS and AS for contact and non-contact detection

The fabrication of the proposed device and MUT is shown in Figure 5.16 (a) where the MUTs used were RO5880 ($\epsilon_r = 2.2$, $\tan \delta = 0.009$), RO4003C ($\epsilon_r = 3.65$, $\tan \delta = 0.0027$), FR-4 ($\epsilon_r = 4.3$, $\tan \delta = 0.0265$), and RO3006 ($\epsilon_r = 6.15$, $\tan \delta = 0.0025$) with dimensions of

10 mm x 10 mm x 1.6 mm. A comparison of the simulation and measurement of the proposed resonator is shown in Figure 5.16 (b) where the two resonators had identical characteristics as dual-band resonators. However, there were still errors and frequency shifts where f_{r1} shifts from 2.35 GHz to 2.34 GHz while f_{r2} shifts from 1.57 GHz to 1.81 GHz. This was due to the variation of permittivity in the substrate, namely the range of ϵ_r of 4.25 – 4.35, and the error during the fabrication process.

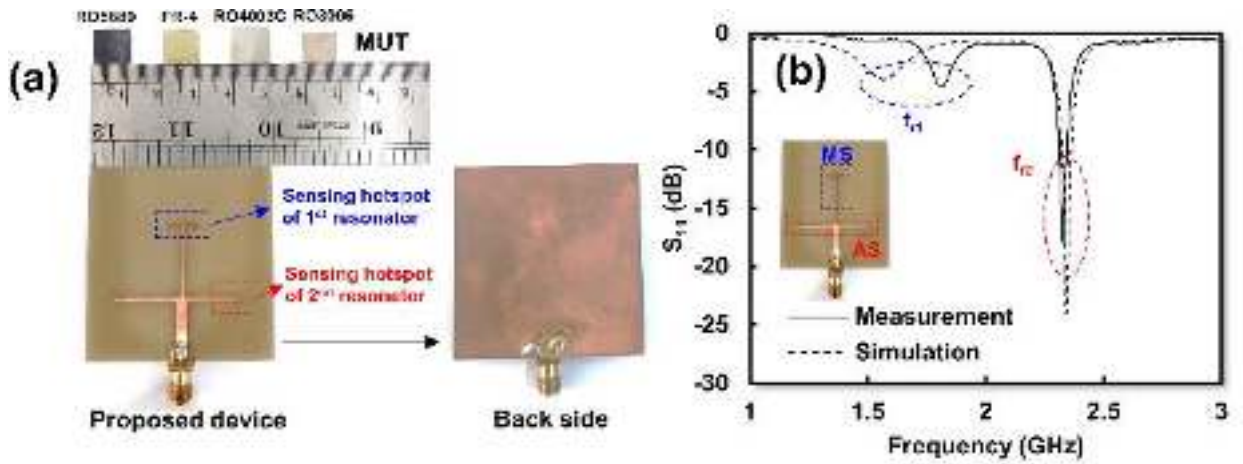


Figure 5.16 (a) Fabricated of proposed resonator, (b) simulation vs measurement

The measurement process was carried out using VNA which was connected directly to the proposed sensor using an SMA connector via port 1 with an impedance of 50 Ω as shown in Figure 5.17 (a). The frequency range used was 1.00 - 3.00 GHz with a sweep frequency of 0.001 GHz and an ambient temperature of 25° C. The measurement data was stored using a USB device connected to a VNA and was processed using a personal computer (PC).

To validate and measured the performance of MS and AS for contact and non-contact detection, two scenarios as shown in Figure 5.17 (a) were proposed:

- (1) MUT was placed on the sensing area of the 1st resonator without an air gap as MS

for contact detection while the 2nd resonator was in a vacuum.

- (2) MUT was placed on the sensing area of the 2nd resonator with an air gap as AS for non-contact detection while the 1st resonator was in a vacuum.

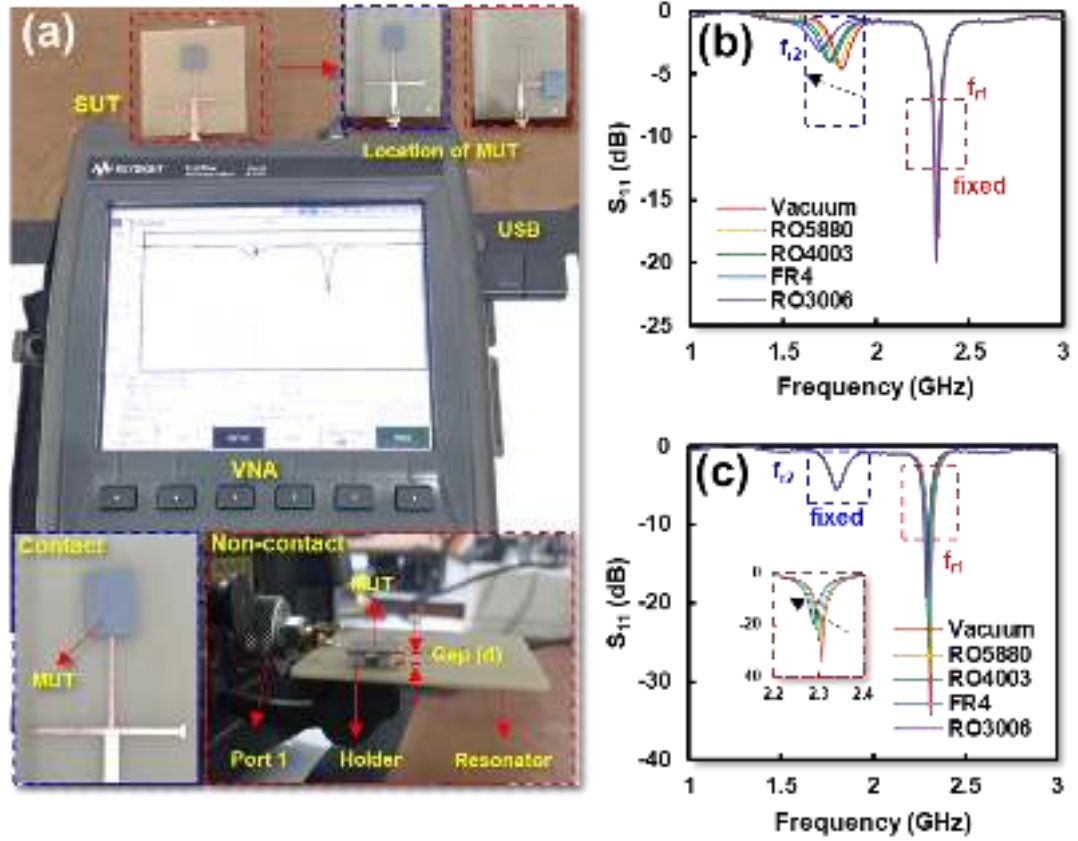


Figure 5.17 (a) Placement scenario of MUT, (b) measurement result of contact detection using MS, (c) measurement result of non-contact detection using AS with an air gap of 0.5 mm

Based on the measurements, for scenario (1) f_{r2} shifted to the lower frequency when the permittivity of the MUT increased, while f_{r1} was fixed. The same characteristics were obtained for scenario (2) where f_{r1} shifted to the lower frequency while f_{r2} was fixed for the permittivity range 1 - 6.15 as shown in Figure 5.17 (b) and Figure 5.17 (c). The independent performance of the resonators is shown in Figure 5.18 (a) and Figure 5.18 (b) where the

respective resonant frequencies of the two resonators did not experience a significant shift when they were used as MS and AS for contact and non-contact detection with the permittivity range of 1 - 6.15.

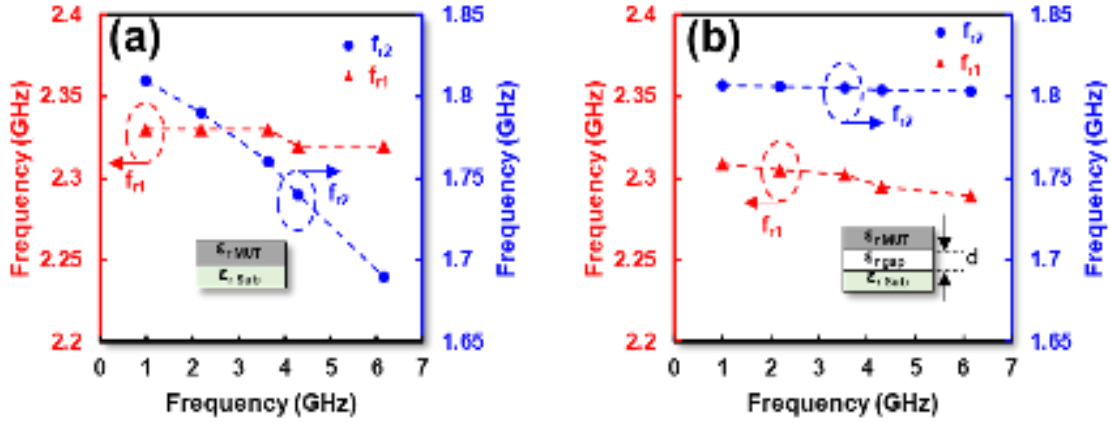


Figure 5.18 (a) Independent performance of MS for contact detection, (b) independent performance of AS for non-contact detection with an air gap of 0.5 mm

For contact detection using MS, f_2 shifted from 1.810 GHz to 1.690 GHz while f_1 shifted slightly from 2.340 GHz to 2.320 GHz. Furthermore, for AS, f_1 shifted from 2.309 GHz to 2.289 GHz while f_2 shifted slightly from 1.807 GHz - 1.804 GHz. These results also proved that MS are more sensitive than AS. Another interesting finding, the two resonators had independent characteristics and these results are in line with the simulation process.

5.6.2 Sensitivity and accuracy of MS and AS for contact and non-contact detection

The effect of the air gap with Δf and the sensitivity is also discussed in this sub chapter. The maximum sensitivity of MS was 23.30 MHz/ $\Delta\epsilon_r$ with Δf of 0.12 GHz/ $\Delta\epsilon_r$ while that of AS was 4.24 MHz/ $\Delta\epsilon_r$ and 0.02 GHz/ $\Delta\epsilon_r$ with an air gap of 0.5 mm as shown in Figure 5.19 (a) and Figure 5.19 (b). The air gap distance (d) greatly affected the sensitivity and Δf of the AS as shown in Figure 5.19 (c) and Figure 5.19 (d). The maximum sensitivity and Δf

of the AS with a gap of 0.5 mm - 1.5 mm were in the range of $0.001 \text{ GHz GHz}/\Delta\epsilon_r - 0.02 \text{ GHz GHz}/\Delta\epsilon_r$ and $0.39 \text{ MHz}/\Delta\epsilon_r - 4.24 \text{ MHz}/\Delta\epsilon_r$, respectively. This finding is in line with the simulation results where the AS is more sensitive with smaller air gaps.

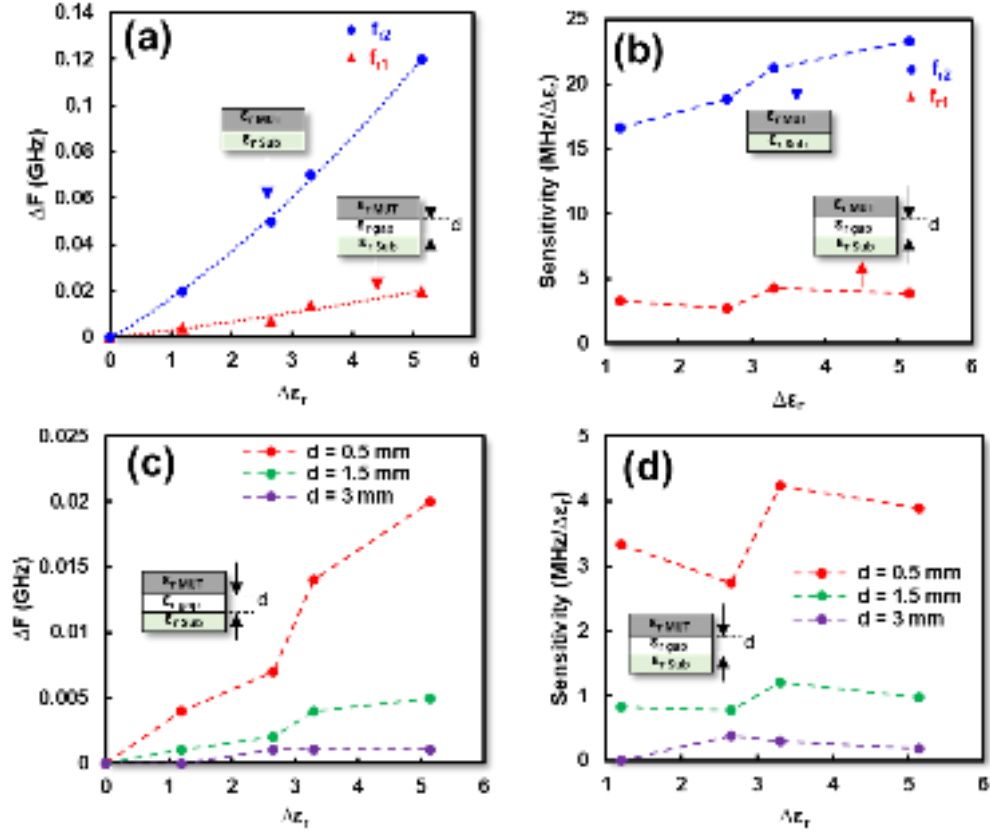


Figure 5.19 (a) Comparison of sensitivity for MS and AS, (b) correlation between air gap and sensitivity for AS, (c) comparison of ΔF for MS and AS, (d) correlation between air gap and ΔF for AS

Furthermore, the difference between the permittivity of the MUT and the reference is expressed as $\Delta\epsilon_r$. Generally, the reference permittivity used is a vacuum with $\epsilon_r = 1$. The normalized sensitivity (NS) was calculated based on equation 3.50. The normalized sensitivity (NS) of the proposed MS and AS sensor was 1.15% and 0.16% with a permittivity range of 1 - 6.15. Furthermore, the accuracy of MS and AS for contact and non-contact detection was obtained by using polynomial equations. The permittivity from the datasheet

was used as a reference and was compared to the permittivity calculated based on the polynomial equation. The comparison of the permittivity values from the datasheet and the calculation from MS and AS are shown in Figure 5.20 (a) and Figure 5.20 (b).

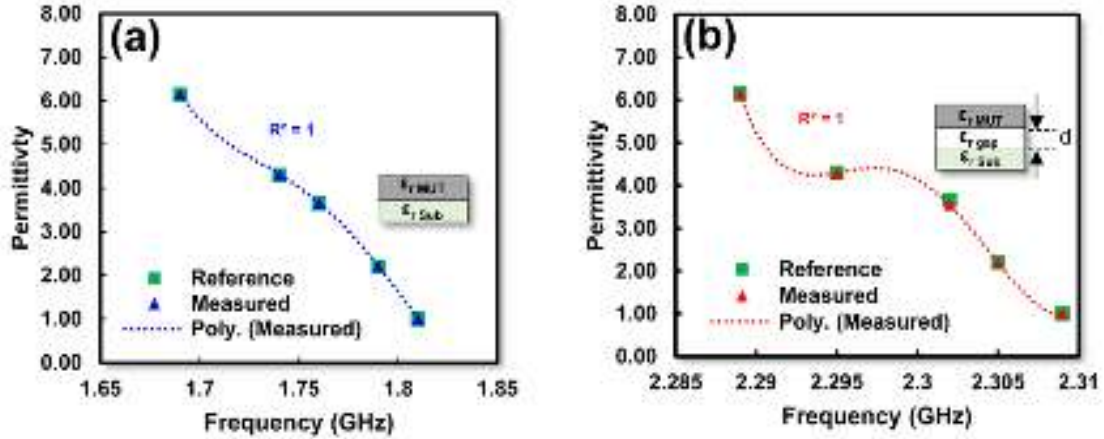


Figure 5.20 (a) Accuracy of MS compared with reference permittivity, (b) accuracy of AS compared with reference permittivity

Based on measurements and calculations, the permittivity of the MUT was determined for MS (ϵ_{r1}) and AS (ϵ_{r2}) with the following equation:

$$\epsilon_{r1} = a_1 f_{ms}^4 - a_2 f_{ms}^3 + a_3 f_{ms}^2 - a_4 f_{ms} + a_5 \quad (5.1)$$

$$\epsilon_{r2} = b_1 f_{as}^4 - b_2 f_{as}^3 + b_3 f_{as}^2 - b_4 f_{as} + b_5 \quad (5.2)$$

where f_{ms} is the resonant frequency of MS with $a_1 = 41667$, $a_2 = 294643$, $a_3 = 780982$, $a_4 = 919666$, and $a_5 = 405977$ as shown in Eq. (5.1). On the other hand, f_{as} represented the resonant frequency of AS with $b_1 = 4 \times 10^8$, $b_2 = 3 \times 10^9$, $b_3 = 1 \times 10^{10}$, $b_4 = 2 \times 10^{10}$ and $b_5 = 1 \times 10^{10}$ as shown in Eq. (5.2). Table 5.3 shows the accuracy of permittivity detection for MS and AS for permittivity from 1 - 6.15. From the measurement results, the average accuracy of the MS and AS with an air gap of 0.5 mm is 99.30% and 99.31%, respectively.

These results indicate that both sensors had high accuracy in detecting the permittivity of the MUT. The overall performance of MS and AS including sensitivity and ΔF and accuracy is shown in Table 5.3.

Table 5.3 Performance of proposed MS and AS with an air gap of 0.5 mm

MUT	MS			AS		
	ΔF (GHz/ $\Delta\epsilon_r$)	S (MHz/ $\Delta\epsilon_r$)	Acc (%)	ΔF (GHz/ $\Delta\epsilon_r$)	S (MHz/ $\Delta\epsilon_r$)	Acc (%)
Vacuum	-	-	98.001	-	-	99.641
RO5880	0.020	16.161	99.091	0.004	3.330	99.842
RO4003C	0.050	18.862	99.732	0.007	2.741	97.170
FR-4	0.070	21.213	99.772	0.014	4.240	99.932
RO-3006	0.120	23.301	99.843	0.020	3.882	99.953

Table 5.3 shows that the performance of MS was more sensitive than AS with an air gap of 0.5 mm. Furthermore, both sensors had high accuracy and independent characteristics for both MS and AS for contact and non-contact detection.

5.7 Experimental validation with existing works

In this paper, integrated MS and AS based on dual T-shaped resonators is presented. MS and AS were proposed for permittivity detection of MUT for contact and non-contact detection, respectively. In addition, this paper introduces a dual T-shaped resonator structure that has independent characteristics using a single port. The proposed sensor also has two different sensing areas so that it can be proposed for contact and non-contact measurements independently. Table 5.4 shows a comparison between the proposed work and the existing sensors.

Table 5.4 presents that the proposed device has novelty by proposing integration of MS and AS. Another novelty, the dual T-shaped resonator has independent characteristics so that it can be used for contact and non-contact detection independently. Moreover, this study also presents the correlation between the distance from the air gap to the accuracy and sensitivity of AS for non-contact detection. Based on the measurements, an air gap (d) of 0.5 – 1.5 mm is proposed to separate the sensor and MUT for non-contact detection while the best performance for AS is obtained with an air gap of 0.5 mm.

Table 5.4 Performance of proposed MS and AS with an air gap of 0.5 mm

Ref	Model	Range of permittivity (ϵ_r)	Material Under Test (MUT)	Resonance frequency (GHz)			Independent performance	Num. of port	Range of gap (mm)	Contact	Non-contact	Integrated Function MS & AS	Design complexity
				f_{r1}	f_{r2}	f_{r3}							
(Piekarz et al., 2017)	Couple line section	1 – 10.2	Solid	1.50	-	-	No	2	-	Yes	No	No	High
(Ansari, Jha and Akhtar, 2015)	SRR	1 – 3	Solid	2.65	-	-	No	2	-	Yes	No	No	Moderate
(Ebrahimi, Scott and Ghorbani, 2018b)	LC resonators	1 – 10.2	Solid	2.00	-	-	No	2	-	Yes	No	No	High
(Lim, Kim and Hong, 2018)	Microstrip line ring resonator	1 – 3.48	Solid	4.89	9.81	-	No	2	-	Yes	No	No	High
(Behdani et al., 2020)	Aperture Coupling	1 – 12.85	Solid	9.50	12.30	-	No	1	-	Yes	No	No	High
(Haq et al., 2019)	CSRR resonator	1 – 11.90	Solid	5.35	7.99	-	No	2	-	Yes	No	No	Moderate
(Yeo and Lee, 2019b)	Slot loaded	1 – 10.2	Solid	2.50	-	-	No	1	-	Yes	No	No	Moderate
(Yang et al., 2016)	SC-TCSRR	1 – 4.4	Solid	1.42	2.65	4.40	No	2	0.10 – 0.50	No	Yes	No	Moderate
(Ali et al., 2021)	IDC	3 – 6	Solid	1.38	-	-	No	2	0.02 – 0.30	No	Yes	No	High
This work	Dual T-Shaped	1 – 6.15	Solid	1.52	2.35	-	Yes	1	0.50 – 1.50	Yes	Yes	Yes	Low

5.8 Concluding remarks

In this chapter, we have successfully integrated MS and AS for contact and non-contact characterization of solid materials with a permittivity range of 1 – 6.15. The proposed sensor is based on a dual T-Shaped resonator with a single port which operates at $f_{r1} = 2.34$ GHz and $f_{r2} = 1.81$ GHz. In addition, the proposed device has independent characteristics so that it can be used for contact and non-contact measurements separately. From the measurements, the average accuracy was 99.30 % and 99.31 % for MS and AS with an air gap (d) of 0.5 mm, respectively. The normalized sensitivity of the two sensors was 1.16 % and 0.16 % while the range of the air gap (d) for AS was 0.5 - 1 mm. This integration of MS and AS can be recommended as an excellent solution for contact and non-contact material characterization in the biomedical, quality control, and materials science industries.

CHAPTER 6

MULTIFUNCTIONAL MICROWAVE SENSOR AND ANTENNA BASED ON MULTICASCODE T-SHAPED RESONATOR

This chapter presents the development of our proposed microwave sensor based on multicascode T-shaped using single port planar resonator. The discussion in this chapter begins with an introduction to explain the main contribution of the proposed work, design and simulation of multicascode T-shaped resonator, determining the location of sensing areas for microwave sensor, scenarios for placing MUTs for microwave sensor and antenna for data transfer capabilities. The discussion continued with observing the performance of the microwave sensor including the accuracy, sensitivity and while characterize the MUT. In the other hand, the performance of the antenna is indicated by the radiation pattern and data transfer capability using the access point. Furthermore, the final discussion of this chapter is validating the performance of the proposed sensor and antenna through a measurement process and comparing it with existing work to show the novelty of the proposed work.

6.1 Introduction

In this chapter, development of multifunctional device as an antenna and permittivity sensor in a single port resonator has been presented. The sensor consists of three T-shaped resonators operating at three different resonant frequencies at 2.45 GHz, 2.00 GHz, and 1.50 GHz, respectively. The proposed sensor has independent characteristics so that changes that occur in one resonator will not affect the other.

This chapter describes in detail starting from the working principle of the T resonator, sensor model development, equivalent circuit, simulation, and measurement of the proposed sensor. Furthermore, a mathematical model using 2nd order polynomial is proposed to determine the permittivity of the MUT. The average accuracy and normalized sensitivity are described in this paper comprehensively to evaluate the performance of the proposed sensor compared to the previously proposed sensor. The radiation pattern of the antenna at a resonant frequency of 2.45 GHz is also described to see the performance of the proposed resonator.

6.2 Design and simulation of multicascode T-shaped resonator

The design and simulation of the multicascode T-shaped resonator was carried out using HFSS 15.0 while the equivalent circuit used AWR 2009. The stages and simulation process are explained in the next sub-chapter.

6.2.1 Multicascode T-shaped resonator with single port

The proposed resonator consists of three T-shaped resonators connected directly to a microstrip line with an impedance of 50 Ω . The multicascode T-shaped resonator is designed to operate at $f_{r1} = 2.46$ GHz and $f_{r2} = 2.01$ GHz and $f_{r3} = 1.47$ GHz. The development stages of the multicascode T-shaped resonator have been described in Figure 3.21 in chapter 3. The development consists of three stages which indicate the resonant frequency of each resonator. It should be noted, the resonant frequency of each resonator is determined based on the length of the arm of the T-shape resonator. Furthermore, the resonant frequency of each resonator can be determined based on the transmission line model.

Based on the similar approach as the dual T-shaped resonator described in chapter 5, the resonant frequency can be determined based on the length of the arm of the resonator which has a high concentration of electric field based on each resonant frequency. The structure and E-field simulation results of the multicascode T-shaped resonator at $f_{r1} = 2.46$ GHz and $f_{r2} = 2.01$ GHz and $f_{r3} = 1.47$ GHz are shown in Figure 6.1 (a), Figure 6.1 (b), Figure 6.1 (c) and Figure 6.1 (d).

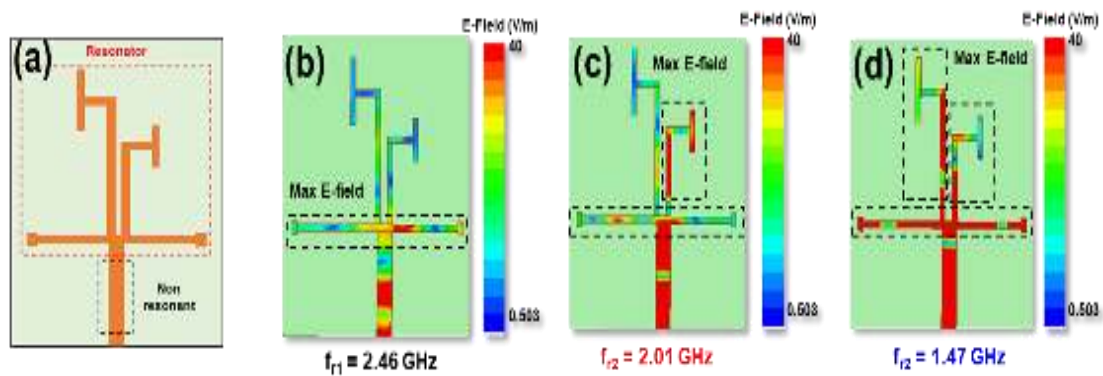


Figure 6.1 Structure and simulation of E-field of dual T-shaped resonator; (a) structure of multicascode T-shaped resonator, (b) E-field at $f_{r1} = 2.46$ GHz, (c) E-field at $f_{r2} = 2.01$ GHz, (d) $f_{r3} = 1.47$ GHz

Figure 6.1 (a) shows the structure of a multicascode T-shaped resonator consisting of a 1st, 2nd and 3rd resonator connected by a microstrip line with an impedance of 50 Ω . The highest concentration of the electric field of the resonator at $f_{r1} = 2.46$ GHz is around the arms of the 1st resonator as shown in Figure 6.1 (b) while for $f_{r2} = 2.01$ GHz it is around both arms of the 1st and 2nd resonator as shown in Figure 6.1 (c) and for $f_{r3} = 1.54$ GHz it is around the arms of 1st, 2nd and 3rd resonator as shown in Figure 6.1 (d).

These findings indicate that the 1st and 2nd resonators work together to generate $f_{r2} = 2.01$ GHz and for $f_{r3} = 1.47$ GHz, all the resonators work together. Based on equation 3.14

and equation 3.15, the width (W_z) of feedline of the proposed resonator can be determined as follows:

$$B = \frac{377\pi}{2Z_0\sqrt{\epsilon_r}} = \frac{(377)(3.14)}{2(50)\sqrt{4.3}} = 5.72$$

$$W_z = \frac{2}{\pi} \left[B - 1 - \ln(2B - 1) + \frac{\epsilon_r - 1}{2\epsilon_r} \left\{ \ln(B - 1) + 0.39 - \frac{0.61}{\epsilon_r} \right\} \right]$$

$$W_z = \frac{2}{3.14} \left[(5.72) - 1 - \ln(10.44 - 1) + \frac{4.3 - 1}{2(4.3)} \left\{ \ln(5.72 - 1) + 0.39 - \frac{0.61}{4.3} \right\} \right]$$

$$= 3.1 \text{ mm}$$

Based on the calculation, the width of the microstrip line (W_z) is 3.1 mm with an impedance of 50 Ω . Furthermore, the length and width of the arm of the resonator is obtained from the optimization process using EM simulation software which represents the resonant frequency of the resonator as shown in Table 6.1.

It should be noted, the three resonators have different dimensions and work together to generate their respective resonant frequencies. In the first stage, the first resonator operates at $f_{r1} = 2.46$ GHz while in the second stage the addition of the second resonator generates $f_{r2} = 2.02$ GHz and in the final stage the addition of the third resonator generates $f_{r3} = 1.47$ GHz. Furthermore, the resonant frequency of the resonator can be determined using a transmission line model based on the arm of the resonator which has a high electric field concentration at each resonant frequency as shown in Figure 6.2.

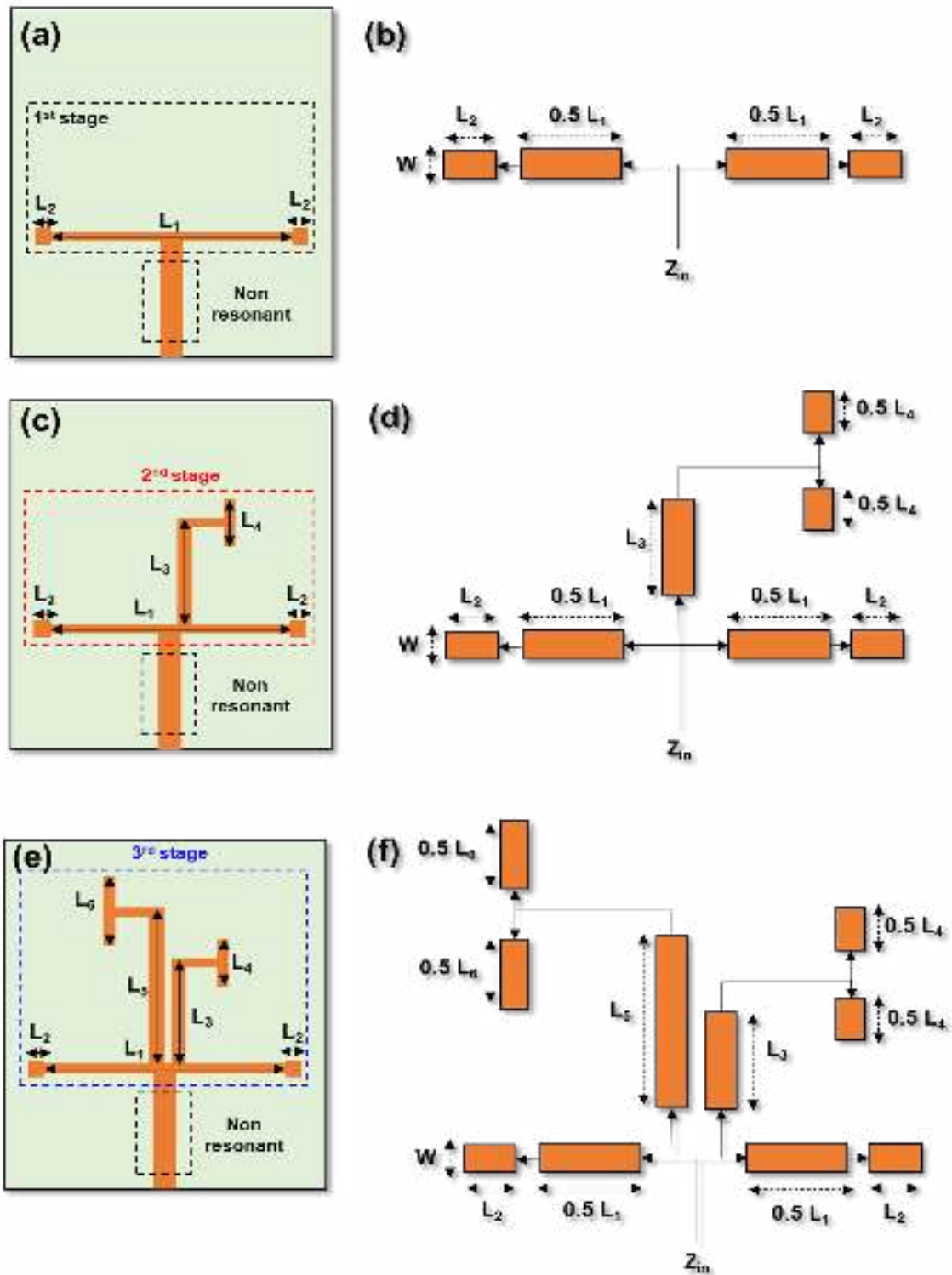


Figure 6.2 Structure and transmission line model multicascode T-shaped resonator; (a) structure of 1st stages, (b) transmission line of 1st stage (c) structure of 2nd stage, (d) transmission line model of 2nd stage, (e) structure of 3rd stages, (f) transmission line model of 3rd stage

Figure 6.2 (a), Figure 6.2 (c) and Figure 6.2 (e) show the structure of the resonator in the 1st, 2nd and 3rd stage which is represented by L_1 , L_2 , L_3 , L_5 , L_6 , L_7 and L_8 which are connected to Z_{in} . The transmission line model of the 1st, 2nd and 3rd stages is shown in Figure 6.1 (b), Figure 6.1 (d) and Figure 6.1 (f). The length (L) and width (w) of the arm of the resonator is obtained from the optimization process using EM simulation software which represents the resonant frequency of the resonator as shown in Table 6.1.

Table 6.1 Optimized dimensions of the arm of the proposed resonator

Parameter	Dimension (mm)
L_1	32
L_2	1
L_3	12.5
L_4	1
L_5	6
L_6	19
L_7	1
L_8	9
w	1

Furthermore, the resonant frequency of each resonator can be determined based on the equation 3.4 and 3.45 as follows:

$$\varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left[1 + 12 \frac{h}{w} \right]^{-\frac{1}{2}}$$

$$\varepsilon_{eff} = \frac{4.3 + 1}{2} + \frac{4.3 - 1}{2} \left[1 + 12 \frac{1.6}{1} \right]^{-\frac{1}{2}} = 3.017$$

Therefore, the resonant frequency of proposed resonator can be determined as follows:

$$f_{r1} = \frac{nc}{2(L_1 + 2L_2)\sqrt{\varepsilon_{eff}}}$$

$$f_{r1} = \frac{(1)(3 \times 10^8)}{2(32 + 2)\sqrt{3.017}} = 2.54 \text{ GHz}$$

$$f_{r2} = \frac{nc}{2(L_1 + 2L_2 + L_3 + L_4 + L_5)\sqrt{\varepsilon_{eff}}}$$

$$f_{r3} = \frac{(1)(3 \times 10^8)}{2(32+2+12.5+1+6)\sqrt{3.017}} = 1.62 \text{ GHz}$$

$$f_{r3} = \frac{nc}{2(L_1+2L_2+L_6+L_7+L_8)\sqrt{\epsilon_{eff}}}$$

$$f_{r2} = \frac{nc}{2(32+2+19+1+9)\sqrt{\epsilon_{eff}}} = 1.37 \text{ GHz}$$

To validate the calculations, a simulation with EM simulation software is proposed to observe the resonant frequency and performance of the resonator. The resonant frequency of each resonator can be modelled and simulated using EM simulation based on the arm of the resonator with the highest electric field concentration as shown in Figure 6.4 (a), Figure 6.4 (b) and Figure 6.4 (c).

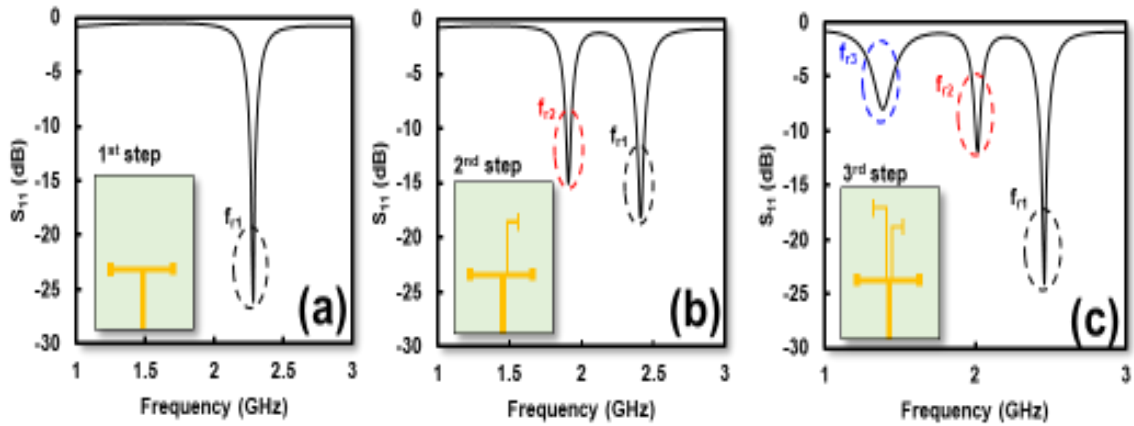


Figure 6.3 (a) 1st step design of proposed resonator with single band (b) 2nd step design of proposed resonator with dual-band, (c) 3rd step design of proposed resonator with triple-band

The 1st stage model shows a T-shaped resonator design which was connected directly to the microstrip line and connector with an impedance of 50 ohms that operates at only one resonant frequency of 2.46 GHz. In the 2nd stages, a resonator which has an identical shape to the 1st resonator was added to produce a new resonant frequency of 2.01 GHz. Finally, in

the 3rd stage, the last resonator was added which functions to produce a new resonant frequency of 1.47 GHz.

6.2.2 Structure of multicascode T-shaped resonator with single port

The proposed sensor consists of three T-shaped resonators with a multicascode configuration. Each resonator operates at a different resonant frequency and is connected to port 1 using an SMA connector with an impedance of 50 Ω . The substrate used is FR-4 Epoxy with a dielectric constant (ϵ_r) of 4.3, a thickness (h) of 1.6 mm, and a $\tan \delta$ of 0.0265. The structure of multicascode T-shaped resonator is shown in Figure 6.4. while the overall dimension shows in Table 6.2.

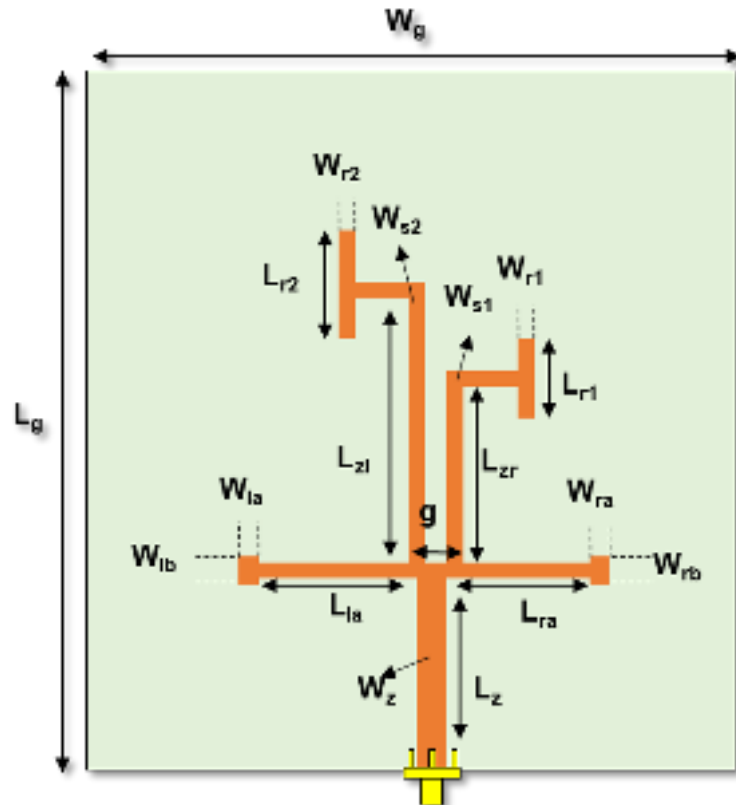


Figure 6.4 Structure of multicascode T-shaped resonator

Table 6.2 Dimension of proposed multicascode T-shaped resonator

Parameters	Dimension (mm)	Parameters	Dimension (mm)
L_z	15	W_{lb}	2
W_{r1}	1	W_{rb}	2
L_{r1}	6	L_{zr}	11.5
W_{s1}	1	L_{zl}	18
W_{r2}	1	L_{ra}	13
L_{la}	15.5	g	0.5
L_{r2}	9.5	W_g	50
W_{s2}	1	L_g	50
W_{la}	1	W_{ra}	1

The development equivalent circuit model of the proposed resonator has been mentioned in chapter 3, Figure 3.18 consists of a combination of resistor, capacitor, and inductor. The 1st resonator represented operates at a resonant frequency of 2.46 GHz by $R_1 = 1.47 \text{ k}\Omega$, $L_1 = 2.65 \text{ nH}$ and $C_1 = 3.16 \text{ pF}$ while $C_{l1} = 1.06 \text{ pF}$ and $L_{l1} = 1.705 \text{ nH}$ are functioned as impedance matching circuits for high frequencies.

Furthermore, the 2nd and 3rd resonators operated at resonant frequencies of 2.01 GHz and 1.47 GHz are represented by $R_3 = 0.64 \text{ k}\Omega$, $C_3 = 2.59 \text{ pF}$, $L_3 = 1.88 \text{ nH}$ and $R_2 = 0.21 \text{ k}\Omega$, $C_2 = 2.68 \text{ pF}$, $L_2 = 4.07 \text{ nH}$ while $C_{l3} = 0.68 \text{ pF}$, $L_{l3} = 2.62 \text{ nH}$, $C_{l2} = 2.42 \text{ pF}$ and $L_{l2} = 4.4 \text{ nH}$ functioned as impedance matching circuits for low frequencies, respectively. Furthermore, $C_{g1} = 1.11 \text{ pF}$ and $C_{g2} = 0.38 \text{ pF}$ are added to show that the proposed resonator is an open circuit where C_{g1} and C_{g2} show a gap between the resonator and ground plane.

Figure 6.5 shows that the simulation results of the FEM simulation are in line with the proposed equivalent circuit model. The proposed resonator between the FEM simulation and the equivalent circuit model has identical characteristics with three different resonant frequencies.

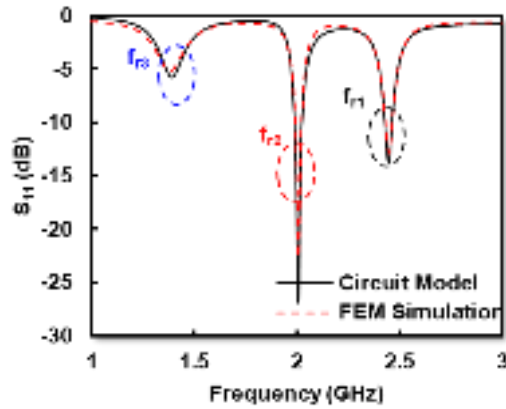


Figure 6.5 Simulation from equivalent circuit model and FEM

6.3 Potential location sensing area of multicascode T-shaped resonator

Surface current distribution refers to the flow of electrical current that occurs on the surface of a conductive material. In the context of microwave sensors, the surface current distribution at resonant bands refers to the distribution of current that occurs on the surface of the sensor at the frequencies at which the sensor resonates. At the resonant frequency, the electromagnetic field produced by the sensor becomes highly concentrated, which causes an accumulation of charge on the surface of the sensor. This accumulation of charge results in a flow of electrical current along the surface of the sensor, which can be visualized as a distribution of current density.

In this paper, the placement location of the MUT depends on the concentration of the E-field of the resonator. Generally, the sensing area is determined from the highest E-field of the resonator structure. Furthermore, the concentration of the H-field showed the characteristics of the resonator as an antenna. To determine the maximum E-field and H-field, a simulation process was carried out on the three resonant frequencies of the proposed

resonator using HFSS 15.0 as shown in Figure 6.6 (a), Figure 6.6 (b), Figure 6.6 (c) and Figure 6.6 (d).

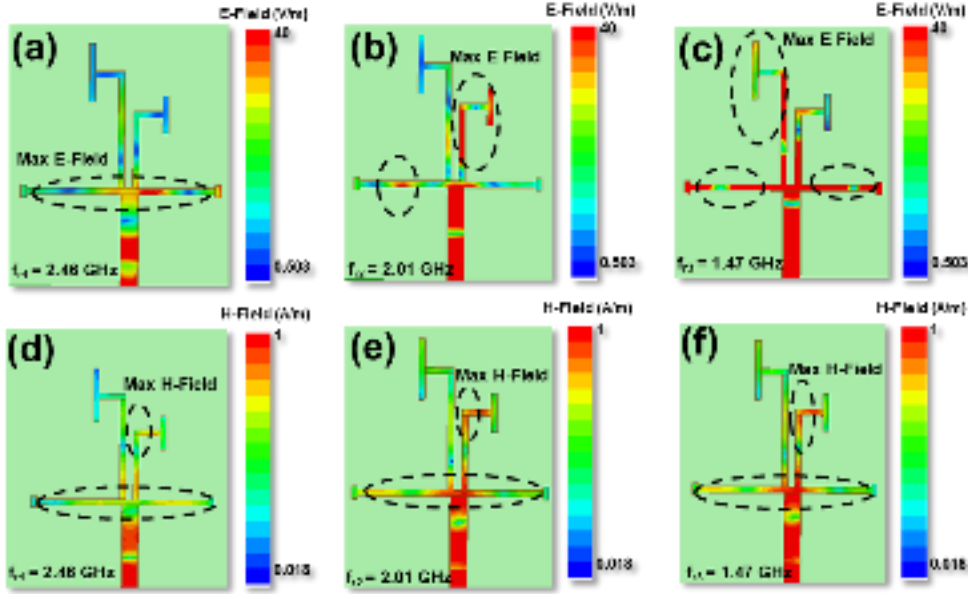


Figure 6.6 E-field and H-field concentration of proposed resonator (a) E-field at $f_{r1} = 2.46$ GHz (b) E-field at $f_{r2} = 2.01$ GHz, (c) (b) E-field at $f_{r3} = 1.39$ GHz, (d) H-field at $f_{r1} = 2.46$ GHz, (e) H-field at $f_{r2} = 2.01$ GHz, (f) H-field at $f_{r3} = 1.47$ GHz

Figure 6.6 (a) and Figure 6.6 (d) show the maximum E-field and H-field at $f_{r2} = 2.46$ GHz at the same location on the arm of the 1st resonator. This showed that the 1st resonator can be used as an antenna because it has the E-field and H-Field concentrated in the same location. Furthermore, Figure 6.6 (b) and Figure 6.6 (e) show that the location of the maximum E-field at $f_{r2} = 2.01$ GHz is in the open-ended stub of the 2nd resonator while the maximum H-field is in the arms of the 1st and 2nd resonators. On the other hand, the location of the maximum E-field at $f_{r3} = 1.47$ GHz is on the arm and the open-ended stub of the 3rd resonator while the maximum H-field is on the arm of the 1st resonator and 2nd resonator as shown in Figure 6.6 (c) and Figure 6.6 (f).

Furthermore, to observe the performance of the proposed resonator, radiation patterns of the three resonant frequencies were simulated using HFSS 15.0 as shown in Figure 6.7 (a), Figure 6.7 (b), and Figure 6.7 (c) where the E-plane was indicated in blue marker and H- plane with a red marker.

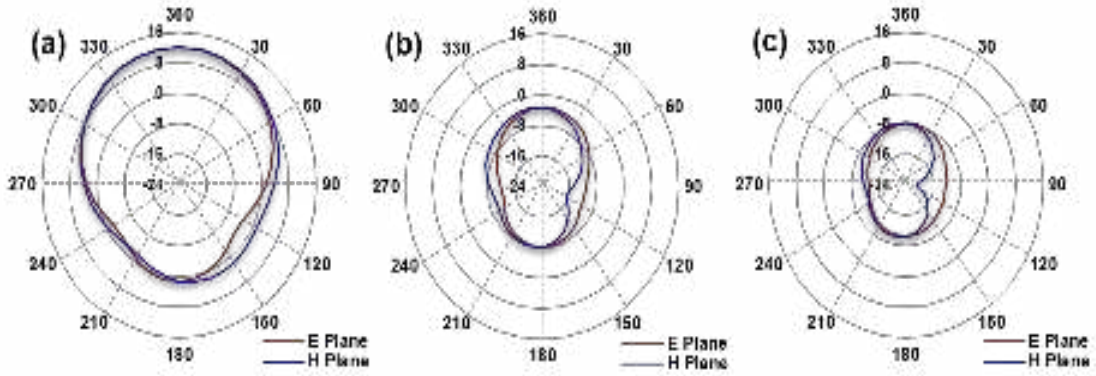


Figure 6.7 Radiation pattern of proposed resonator (a) E-plane and H-Plane at $f_{r1} = 2.46$ GHz (b) E-plane and H-Plane at $f_{r2} = 2.01$ GHz, (c) E-plane and H-Plane at $f_{r3} = 1.47$ GHz

The simulation results showed that the resonator radiated electromagnetic waves at all three resonant frequencies as shown in Figure 6.7. However, maximum electromagnetic wave radiation was only generated from the 1st resonator which operated at a resonant frequency of $f_{r1} = 2.46$ GHz with maximum radiation is 12.09 dB while that for $f_{r2} = 2.02$ GHz and $f_{r3} = 1.47$ GHz is -3.81 dB and -7.83 dB, respectively. This shows that the 1st resonator could be used as an antenna according to the performance of the radiation pattern obtained from the simulation results while the 2nd resonator and 3rd resonators could be used as permittivity sensors.

Based on the simulation process and observations of the E-field and H-field concentrations of the proposed resonator, the location of the MUT is shown in Figure 6.8 (a).

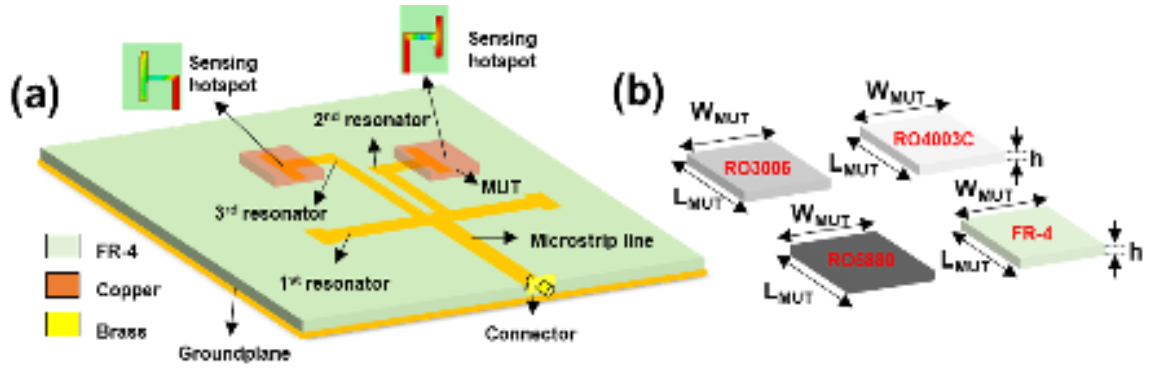


Figure 6.8 (a) Placement MUT of proposed resonator in 2nd and 3rd resonator, (b) dimensions of MUT

Figure 6.8 (a) shows that the placement location of the MUT was in the open-ended stub of the 2nd and 3rd resonators while the 1st resonator was used as an antenna that functions to radiate electromagnetic waves. The MUT used is represented by W_{MUT} , L_{MUT} and h with the overall dimensions were $(10 \times 10 \times 1.6) \text{ mm}^3$ and were placed on the surface of the open-ended stubs of the 2nd and 3rd resonators as shown in Figure 6.8 (b).

6.4 Working principles of a multifunctional microwave sensor with an antenna

The proposed microwave sensor operates at three different resonant frequencies. High frequency represented as $f_{r1} = 2.45 \text{ GHz}$ is used as an antenna while low frequency is represented by $f_{r2} = 2.00 \text{ GHz}$ and $f_{r3} = 1.50 \text{ GHz}$ is used as a permittivity sensor as shown in Figure 6.9 (a) and Figure 6.9 (b). The three resonant frequencies of the proposed device are independent so that they do not affect each other when disturbed. The permittivity sensor is used for the characterization of solid materials simultaneously at two different sensing areas while the antenna is used to transfer measurement data using a wireless network as shown in Figure 6.9 (c) and Figure 6.9 (d).

The performance of both functions of the proposed device was observed by performing several measurements in the laboratory using supporting equipment such as vector network analyzer (VNA), access point, USB wireless and computer. The main contribution of this paper is to produce a multifunctional microwave sensor as a permittivity sensor and antenna that works for the characterization of solid materials and data transfer capabilities.

In addition, the permittivity sensor has two different sensing areas so that it can be used to determine the permittivity of two different types of MUT simultaneously. The working principle of the proposed device is shown in Figure 6.9.

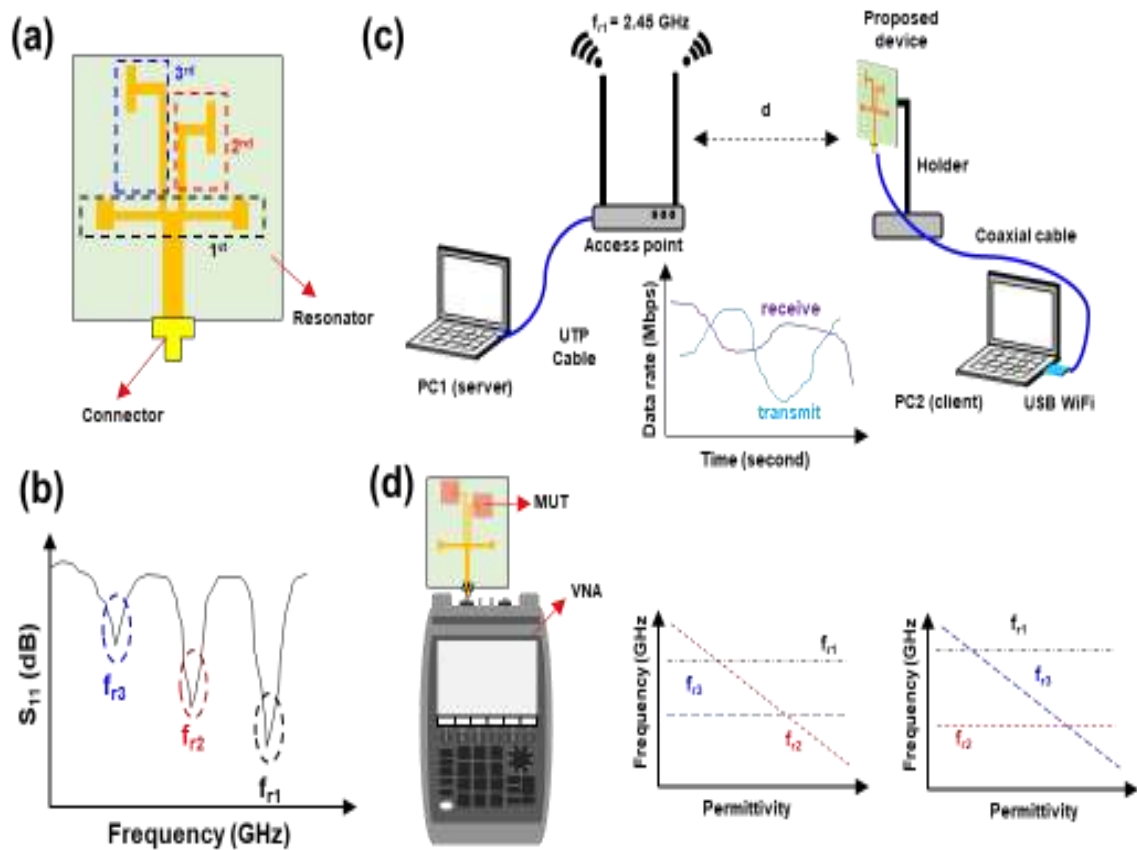


Figure 6.9 (a) Structure of proposed device, (b) resonant frequency of proposed device, (c) proposed device as an antenna (d) proposed device as permittivity sensors

The microwave sensor utilizes the electric field to determine the sensing area of the sensor from the proposed device and its performance is observed from correlation of resonant frequency with permittivity of MUT obtained during the measurement process. Furthermore, the proposed device functions as an antenna by utilizing the concentration of the magnetic field from the resonator and its performance is observed from the data rates in sending and receiving data through a wireless network to send measurement data.

6.5 Simulation and measurement of multifunctional microwave sensor and antenna

To determine the permittivity of the MUT, a simulation process was carried out by placing the MUT on the surface of the 2nd and 3rd resonators. The simulation used two following scenarios:

- (1) The MUT was placed on the surface of the 2nd resonator while the 3rd resonator was empty (vacuum).
- (2) The MUT was placed on the surface of the 3rd resonator while the 2nd resonator was empty (vacuum).

The simulation results are shown in Figure 6.10 (a) and Figure 6.10 (b). The permittivity range used in the simulation process was $\epsilon_r = 1 - 10$ with $\tan \delta = 0$. Figure 6.10 (a) shows that in scenario (1) the resonant frequency of the 1st and 3rd resonators were fixed while the resonant frequency of the 2nd resonator shifted to a low frequency in line with the change in permittivity of the MUT. Furthermore, Figure 6.10 (b) shows that in scenario (2), the frequency of the 1st and 2nd resonators was fixed while the frequency of the 3rd resonator shifted to a lower frequency in line with the change in permittivity of the MUT.

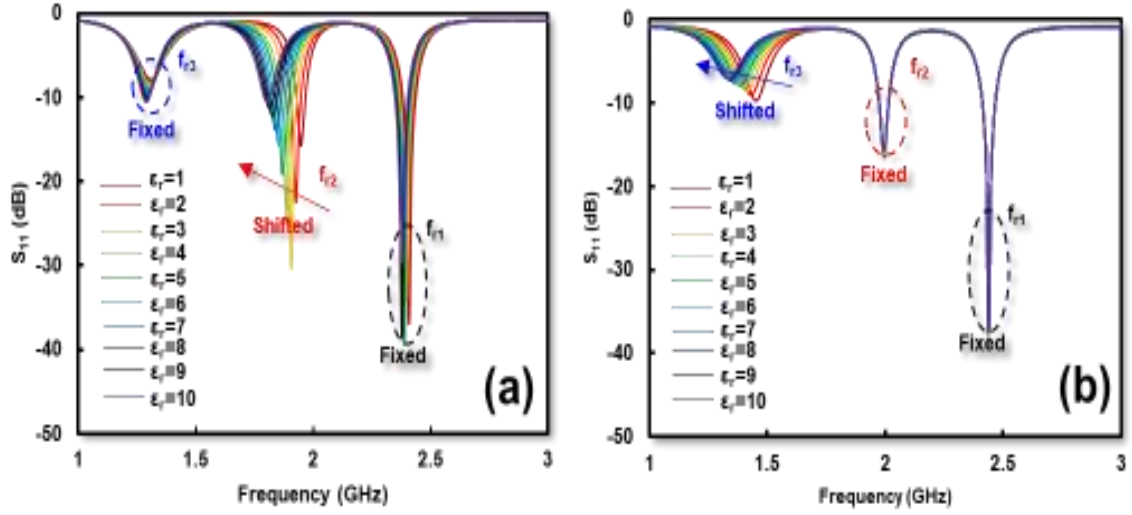


Figure 6.10 Simulation of proposed resonator while MUT loaded in 2nd and 3rd resonator; (a) during scenario 1, (b) during scenario 2

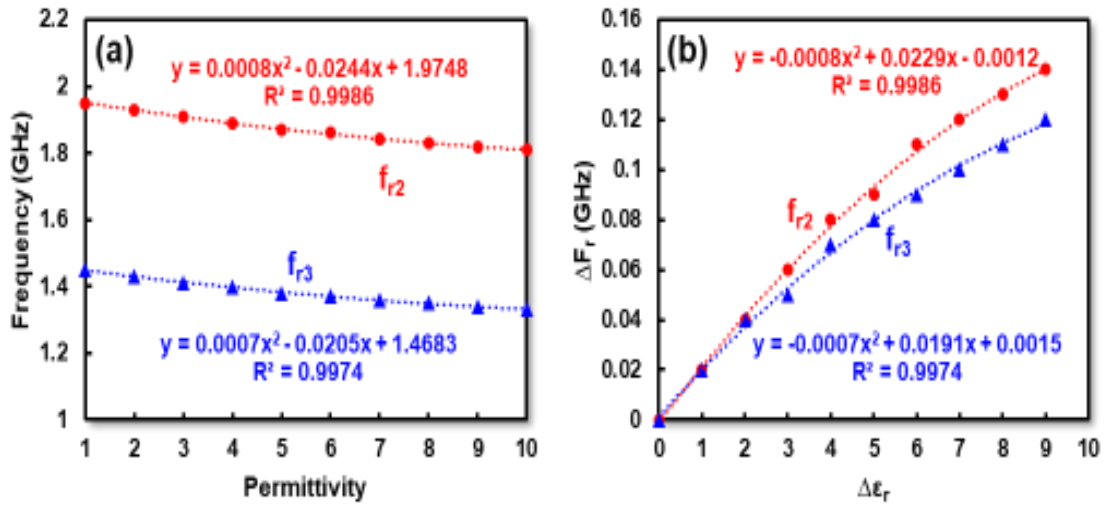


Figure 6.11 (a) The effect of the change in permittivity on the resonant frequency of the resonators; (b) the maximum frequency shift (Δf_r) of the proposed sensors

The effect of the change in permittivity of the MUT on the resonant frequencies of the 2nd and 3rd resonators during scenarios 1 and 2 is shown in Figure 6.11 (a) while the frequency shift is shown in Figure 6.11 (b). Furthermore, Figure 6.11 (a) shows that the

resonant frequency of the 2nd resonator shifted from 1.95 GHz to 1.81 GHz while the frequency of the 3rd resonator shifts from 1.45 GHz to 1.33 GHz in line with the changes in the permittivity of the MUT from the range 1-10. Furthermore, the maximum frequency shift (Δf_r) of the 2nd resonator and 3rd resonator was 0.12 GHz and 0.14 GHz as shown in Figure 6.11 (b).

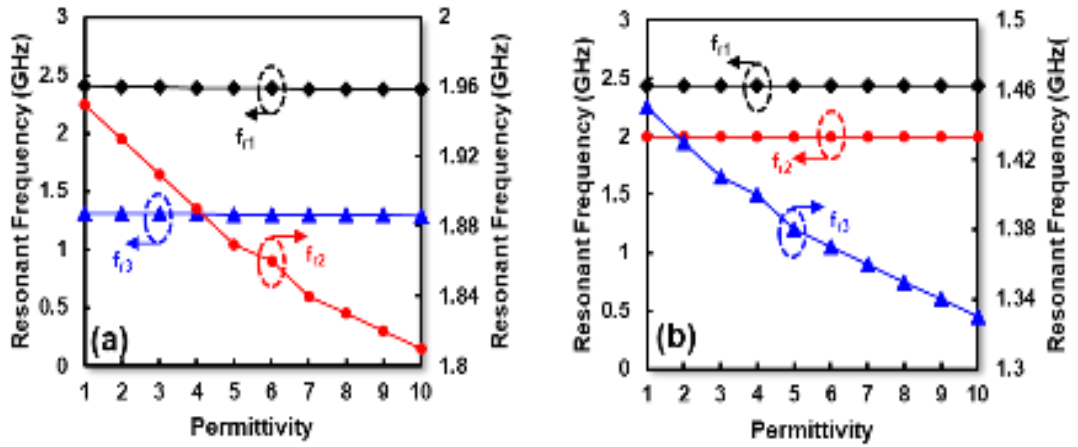


Figure 6.12 Independent performance of the resonator when the permittivity changes; (a) during scenario 1, (b) during scenario 2

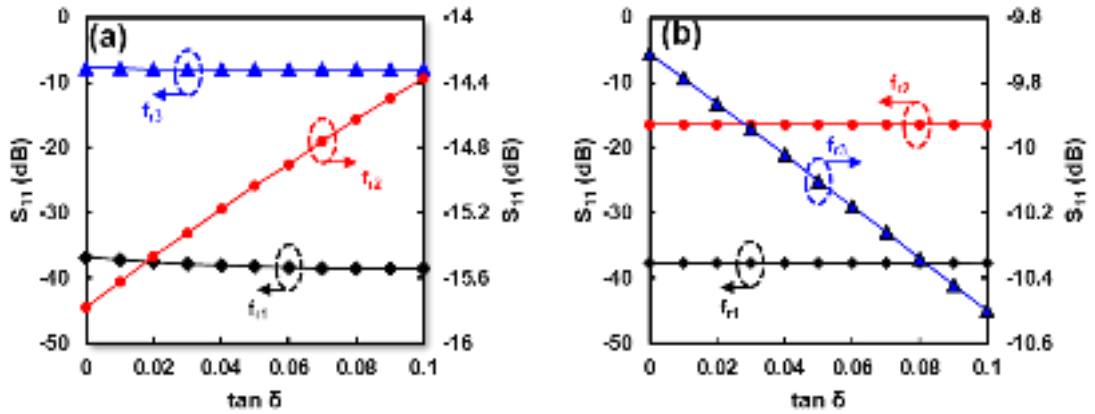


Figure 6.13 The effect of the change in $\tan \delta$ on the resonant frequency of resonators; (a) during scenario 1, (b) during scenario 2

The simulation results from scenarios (1) and (2) showed that the proposed 2nd and 3rd resonators had independent characteristics where the permittivity change only occurred when the resonator was loaded MUT at the location of the sensing area as shown in Figure 6.12 (a) and Figure 6.12 (b). In addition, the proposed sensor generated different sensing area locations for the 2nd and 3rd resonators and independent characteristics of reflection coefficient when $\tan \delta$ changed from 0–0.01 as shown in Figure 6.13 (a) and Figure 6.13 (b).

Figure 6.13 (a) shows that the reflection coefficient of f_{r2} shifts was in line with the change in $\tan \delta$ from -15.77 dB to -14.56 dB while f_{r1} and f_{r3} were fixed at -36.90 dB and -7.98 dB. Similarly, the reflection coefficient of f_{r3} shifted from -9.71 dB to -10.50 dB while f_{r1} and f_{r2} were fixed at -37.70 dB and 16.43 dB as shown in Figure 6.13 (b). This proves that the proposed sensor had independent characteristics based on scenarios (1) and (2). Therefore, the sensor can be used to detect the permittivity of two types of MUTs simultaneously.

6.6 Fabrication and measurement of multicascode T-shaped resonator

The measurement process was carried out using a Vector Network Analyzer (VNA) which is connected directly to the proposed sensor using an SMA connector via port 1 with an impedance of 50 Ω as shown in Figure 6.14(a). The frequency range used is 1.00 - 3.00 GHz with a sweep frequency of 0.01 GHz and an ambient temperature of 25° C. The measurement data was stored using a USB device connected to a VNA and were processed using a personal computer (PC).

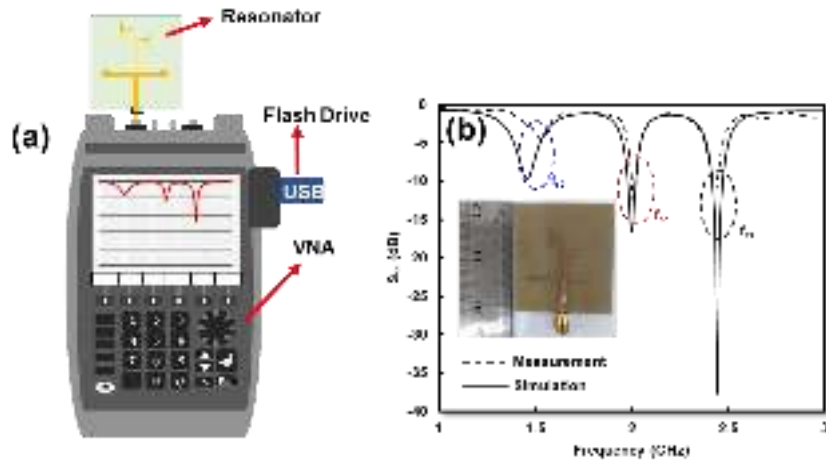


Figure 6.14 (a) Measurement setup of proposed resonator, (b) simulation and measurement result of proposed resonator

The comparison of the resonant frequency and reflection coefficients of the proposed resonator from the simulation and measurement process is shown in Figure 6.14 (b). However, there was still a slight difference between simulation and measurement where f_{r1} shifted from 2.46 GHz to 2.45 GHz, f_{r2} from 2.01 GHz to 2 GHz, and f_{r3} from 1.47 GHz to 1.5 GHz. This was due to the variation of permittivity in the substrate, i.e., range of ϵ_r was 4.25 – 4.35 and also the error during the fabrication process. Overall, the errors in the measurement and simulation processes were 0.41%, 0.50% and 2.04%, respectively.

The measurement process scenario used to observe the performance of the proposed sensor is as follows:

- (1) MUT A placed on the 2nd resonator was fixed (RO5880) while MUT B was changed randomly.
- (2) MUT B placed on the 3rd resonator was fixed (RO5880) while MUT A was changed randomly.

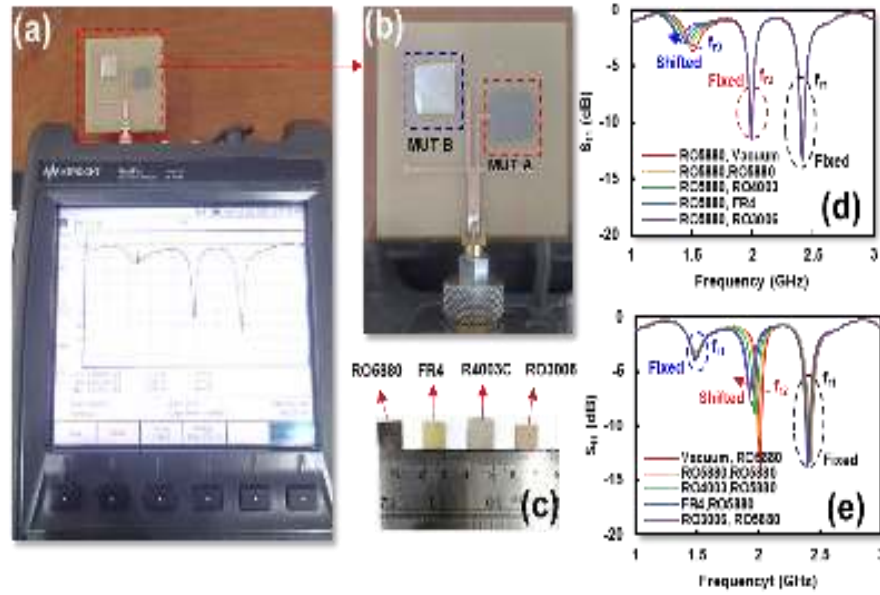


Figure 6.15 (a) Measurement setup of simultaneous MUT loaded in 2nd resonator and 3rd resonator, (b) measurement result of proposed resonator during scenario 1, (c) measurement result of proposed resonator during scenario 2

Furthermore, to observe the performance of the proposed sensor, the measurement process was carried out by placing the MUT simultaneously in each sensing area. The MUT placed on the 2nd resonator was represented as MUT A while MUT B is placed on the 3rd resonator as shown in Figure 6.15 (a) and Figure 6.15 (b). The dimensions of the MUT used are shown in Figure 6.15 (c) which are 10 x 10 x 1.6 mm³. Furthermore, Figure 6.15 (d) shows that in scenario (1), f_{r3} shifted to a low frequency in line with the change in permittivity of the MUT while f_{r1} and f_{r2} were fixed. Similarly, the reflection coefficients of f_{r1} and f_{r2} were also fixed because the MUT placed on the 3rd resonator is fixed (RO5880). Meanwhile, in (2), f_{r2} shifted to a low frequency in line with the change in permittivity of the MUT, while f_{r1} and f_{r3} were fixed in line with the reflection coefficient as shown in Figure 6.15 (e). The correlation between each resonant frequency of the resonator to the change in the permittivity of the MUT is shown in Figure 6.16 (a) and Figure 6.16 (b).

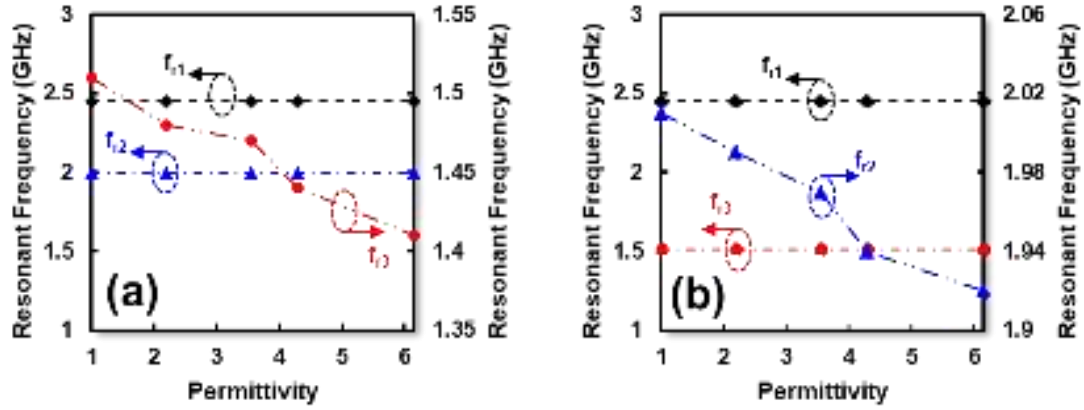


Figure 6.16 (a) Correlation of f_{r1} , f_{r2} and f_{r3} at scenario 1, (b) correlation of f_{r1} , f_{r2} and f_{r3} at scenario 2

Figure 6.16 (a) shows that f_{r3} shifted to a lower frequency from 1.5 GHz to 1.41 GHz while f_{r2} and f_{r1} were fixed at 2 GHz and 2.45 GHz in scenario (1). Furthermore, Figure 6.16 (b) shows that f_{r2} shifted to a lower frequency from 2 GHz to 1.92 GHz while f_{r1} and f_{r3} were fixed at 2.45 GHz and 1.5 GHz in scenario (2). This proves that the proposed sensor had independent characteristics and was not affected when the MUT was loaded simultaneously.

Furthermore, to observe the performance of the 1st resonator which functions as an antenna, the radiation pattern when the MUT was placed simultaneously in both sensing areas according to scenarios (1) and (2) are shown in Figure 6.17 (a), Figure 6.17 (b), Figure 6.17 (c), and Figure 6.17 (d). The radiation pattern from the antenna during scenario 1 was fixed as shown in Figure 6.17 (a) and Figure 6.17 (b), while Figure 6.17 (c) and Figure 6.17 (d) show that the radiation pattern from the antenna changed slightly in scenario (2). However, the characteristics of the radiation pattern obtained from scenario 1 and scenario (2) were still identical to each other and there was no significant change when the MUT was placed simultaneously in the sensing area of the proposed sensor.

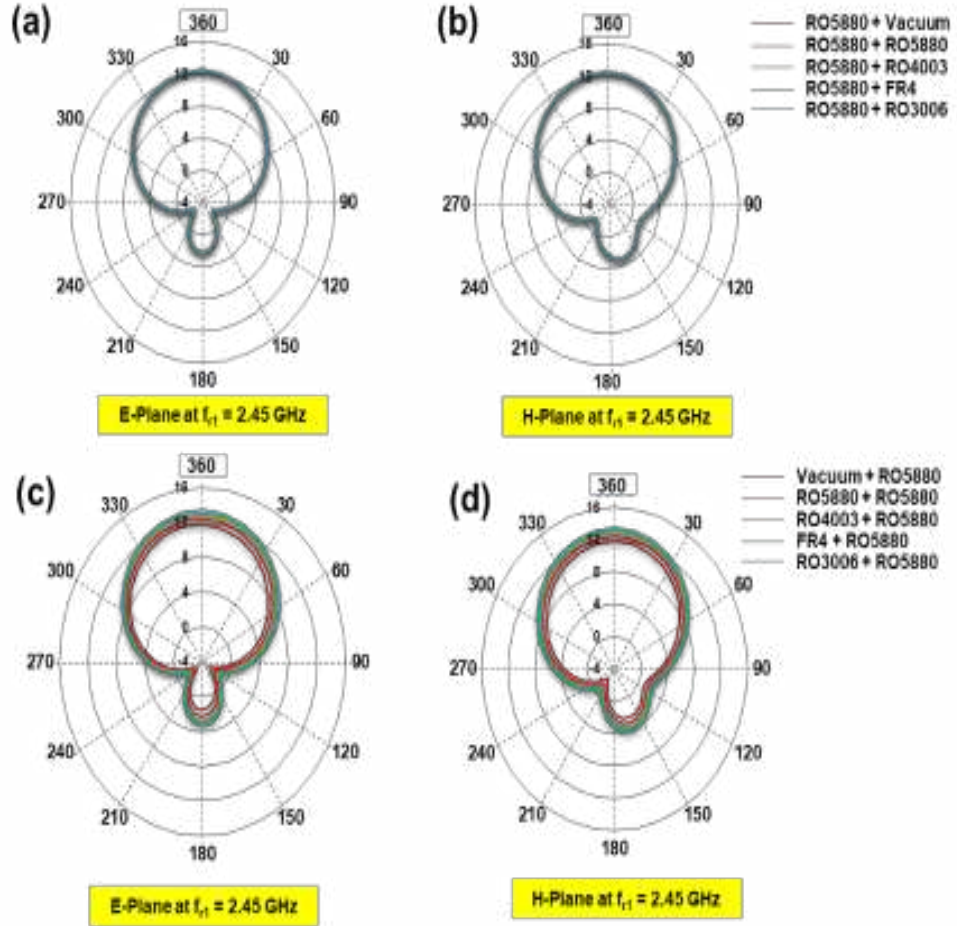


Figure 6.17 (a) Radiation pattern of proposed resonator; (a) E-plane at $f_{r1}=2.45$ during scenario 1 (b) H-plane of proposed resonator at $f_{r1}=2.45\text{GHz}$ during scenario 1, (c) E-plane of proposed resonator at $f_{r1}=2.45$ GHz during scenario 2, (d) H-plane of proposed resonator at $f_{r1}=2.45$ GHz during scenario 2

The comparison of the radiation pattern from the simulation results and the measurement process at $f_{r1} = 2.45$ GHz is shown in Figure 6.18 (a) while Figure 6.18 (b) shows the measurement setup of the radiation pattern in the anechoic chamber. Furthermore, Figure 6.18 (a) shows that there is good agreement between the radiation patterns from the measurement and simulation results. The measurement process was carried out in the anechoic chamber where the Device Under Test (DUT) was positioned as a receiver (Rx) while the reference antenna was positioned as a transmitter (Tx) which was separated by a

distance of $d = 60$ cm as shown in Figure 6.18 (b). The proposed resonator produced a directional radiation pattern with a maximum radiation of 13 dB at a resonant frequency of 2.45 GHz.

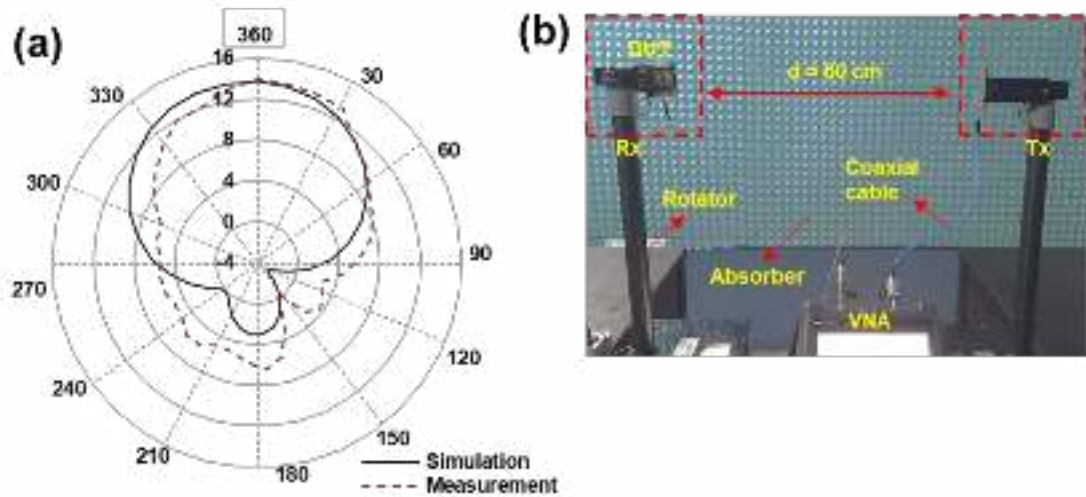


Figure 6.18 (a) The radiation pattern of the antenna at $f_1 = 2.45$ GHz, (b) the measurement setup of the radiation pattern in the anechoic chamber

6.7 Validation and performance of multicascode T-shaped resonator

Verification of the resonator shape was carried out in the laboratory using a Vector Network Analyzer with an ambient temperature of 25°. The stages and measurement process are explained in the next sub-chapter.

6.7.1 Accuracy and normalized sensitivity

The permittivity of the MUT was determined using a 2nd-order polynomial equation that expressed the relationship between the permittivity and the resonant frequency of the proposed sensor when the MUT was placed on the sensing area as shown in Figure 6.19 (a) and Figure 6.19 (b). Furthermore, the accuracy of the sensor was obtained by comparing the

calculated permittivity using a 2nd order polynomial with the reference permittivity from the datasheet.

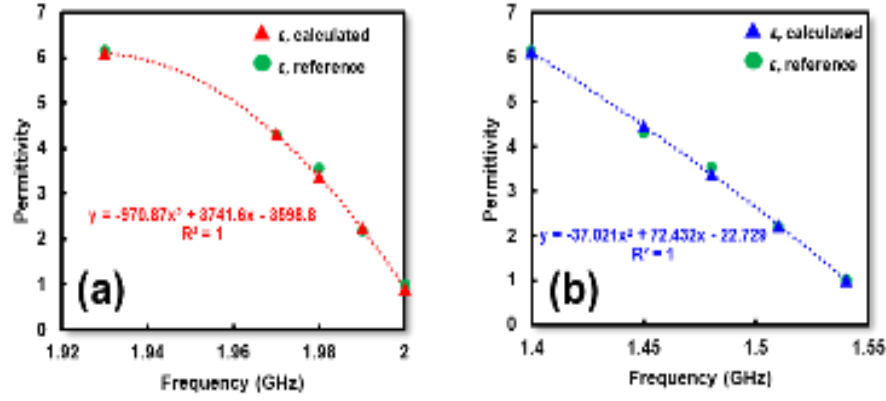


Figure 6.19 (a) accuracy of 2nd resonator compared with reference permittivity, (b) accuracy of 3rd resonator compared with reference permittivity

Based on Figure 6.19 (a) and Figure 6.19 (b), the permittivity of the MUT was determined using the 2nd order polynomial which is expressed as follows:

$$\epsilon_{r1} = -970.87 f_{r2}^2 + 3741.6 f_{r2} - 3598.8 \quad (6.1)$$

$$\epsilon_{r2} = -37.021 f_{r3}^2 + 72.432 f_{r3} - 22.729 \quad (6.2)$$

The overall comparison of the permittivity between the measurement process and the datasheet is shown in Table 6.3.

Table 6.3 Comparison of the value of the permittivity between the measurement process and the datasheet

MUT	Parameter						
	Ref. Permittivity	Meas. Permittivity		Error (%)		Accuracy (%)	
		ϵ_{r1}	ϵ_{r2}	ϵ_{r1}	ϵ_{r2}	ϵ_{r1}	ϵ_{r2}
Vacuum	1.00	0.92	1.02	8.00	1.73	92.00	98.27
RO5880	2.20	2.24	2.23	1.89	1.44	98.11	98.56
RO4003	3.65	3.37	3.38	5.09	4.80	94.91	95.20
FR4	4.30	4.30	4.46	0.06	3.73	99.94	96.27
RO3006	6.15	6.09	6.11	0.90	0.58	99.10	99.42

Table 6.3 shows that the proposed sensor has a high accuracy in determining the permittivity of the MUT compared to the permittivity of the datasheet. The average accuracy of the proposed sensor is 96.81% and 97.54% with a permittivity range of 1 - 6.15.

The MUT placed in the sensing area area perturbed the E-field. The interaction of the MUT to the resonator can be assumed as a capacitive load that causes the resonant frequency to shift. Furthermore, the sensitivities of the 2nd and 3rd resonators are shown in Figure 6.20 (a) and Figure 6.20 (b). The maximum sensitivities of the 2nd resonator and 3rd resonator were 13.59 MHz/ $\Delta\epsilon_r$ and 27.27 MHz/ $\Delta\epsilon_r$, and the average sensitivities were 9.71 MHz/ $\Delta\epsilon_r$ and 25.75 MHz/ $\Delta\epsilon_r$.

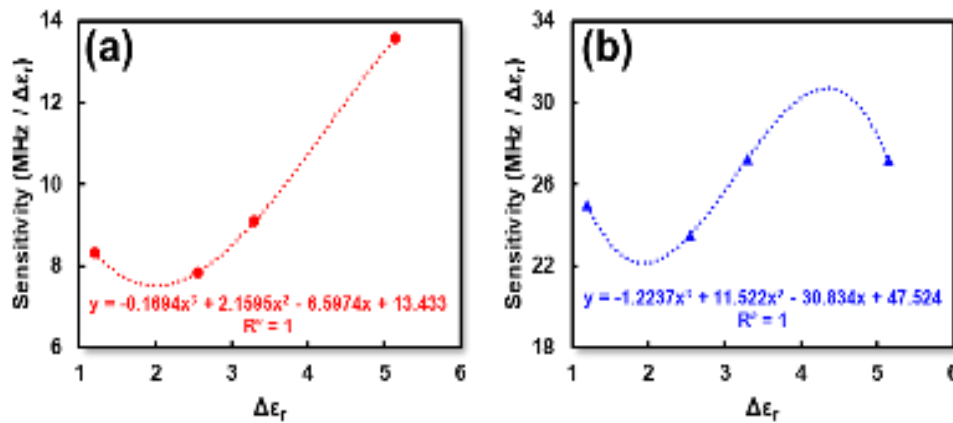


Figure 6.20 (a) Sensitivity of the 2nd resonator at $f_2 = 2.00$ GHz, (b) sensitivity of the 3rd resonator at $f_3 = 1.50$ GHz

Furthermore, the difference between the permittivity of the MUT and the reference is expressed as $\Delta\epsilon_r$. Generally, the reference permittivity used is a vacuum with $\epsilon_r = 1$. Based on Equation 3.50, the normalized sensitivity (NS) of the proposed sensor was 0.68% and 1.77% with a permittivity range of 1 - 6.15.

6.7.2 Data transfer capabilities of multicascode T-shaped resonator

Furthermore, to demonstrate the performance of the antenna when transmitting data, measurements of S_{11} and S_{21} are proposed as shown in **Figure 6.21**.

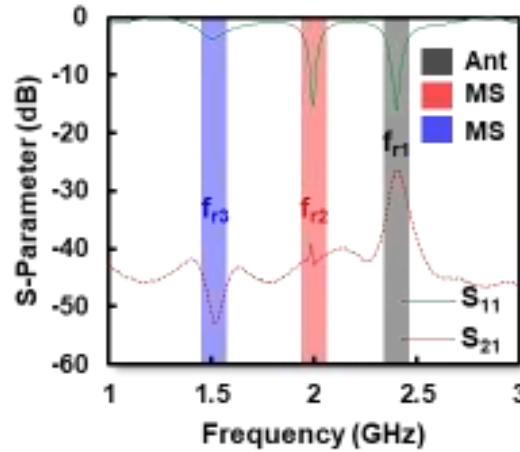


Figure 6.21 Measurement of S_{11} and S_{21} from proposed resonator

Figure 6.21 shows S_{11} and S_{21} of the proposed resonator when functioning as an antenna (Ant) having the same resonant frequency at $f_{r1} = 2.45$ GHz. Based on the measurement results, the magnitude of the mutual coupling (S_{21}) of the resonator increases when the resonator functions as an antenna (Ant) at $f_{r1} = 2.45$ GHz while at $f_{r2} = 2.00$ GHz and $f_{r3} = 1.50$ GHz as microwave sensor (MS) it has a lower magnitude. This shows that the resonator has functioned as an antenna at $f_{r1} = 2.45$ GHz while $f_{r2} = 2$ GHz and $f_{r3} = 1.50$ GHz can be used as permittivity sensors.

Table 6.4 The value of S_{11} and S_{21} of proposed resonator

Resonator	Resonance Frequency	S_{11} (dB)	S_{21} (dB)
1 st	2.45 GHz	-14.68	-26.73
2 nd	2.00 GHz	-15.31	-40.65
3 rd	1.50 GHz	-3.69	-52.87

Table 6.4 shows the value of S_{11} and S_{21} of proposed resonator for f_{r1} , f_{r2} and f_{r3} , respectively. Furthermore, to observe the performance of the proposed microwave sensor as an antenna, the measurement process is performed as shown in Figure 6.22 (a) and Figure 6.22 (b).

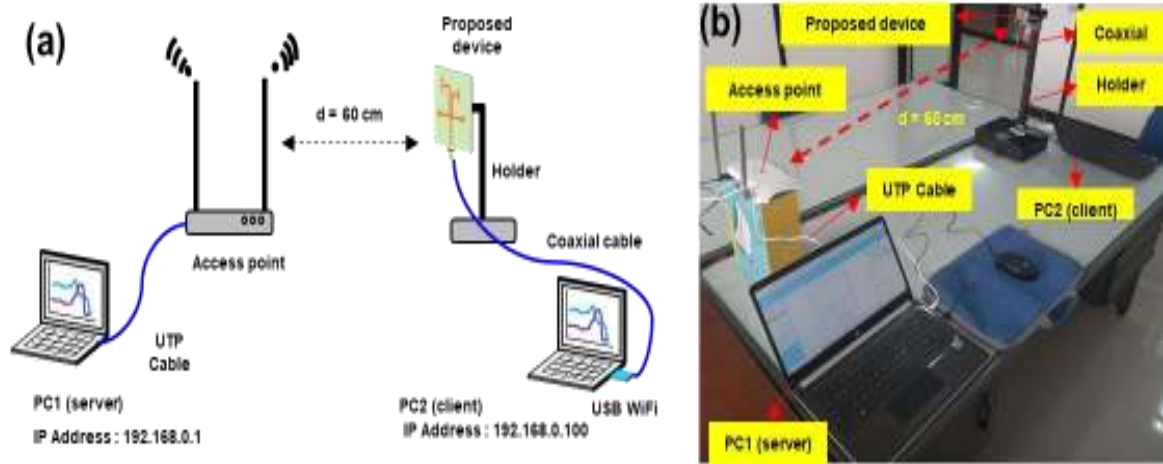


Figure 6.22 Measurement setup from the proposed device for data transfer; (a) illustration of the configuration and measurement process, (b) real measurement configuration

Figure 6.22 (a) shows the configuration of the measurement process where the proposed device is connected to PC2 (client) using USB WLAN while the server located on PC1 is directly connected to an access point operating at a resonance frequency of 2.45 GHz, respectively. Furthermore, PC1 is configured as a gateway with IP address 192.168.0.1 while PC2 as a client with IP address 192.168.0.100. The distance from the server and client is represented by $d = 60 \text{ cm}$ where the client's antenna is placed in the holder and connected directly to the USB WLAN using a coaxial cable with an impedance of 50 Ohm. In addition, PC1 as a server is connected to the access point using a UTP cable as shown in Figure 6.22 (b). The performance of the proposed device is observed with the throughput monitor available on the TP Link - WR840N access point. In this experiment, the server will send

measurement data to the client and vice versa via a wireless network with a file size of 150 MB. In other words, the proposed device functions as a sender and receiver of data transfer. The measurement results from the proposed device are shown in Figure 6.23 (a) and Figure 6.23 (b).

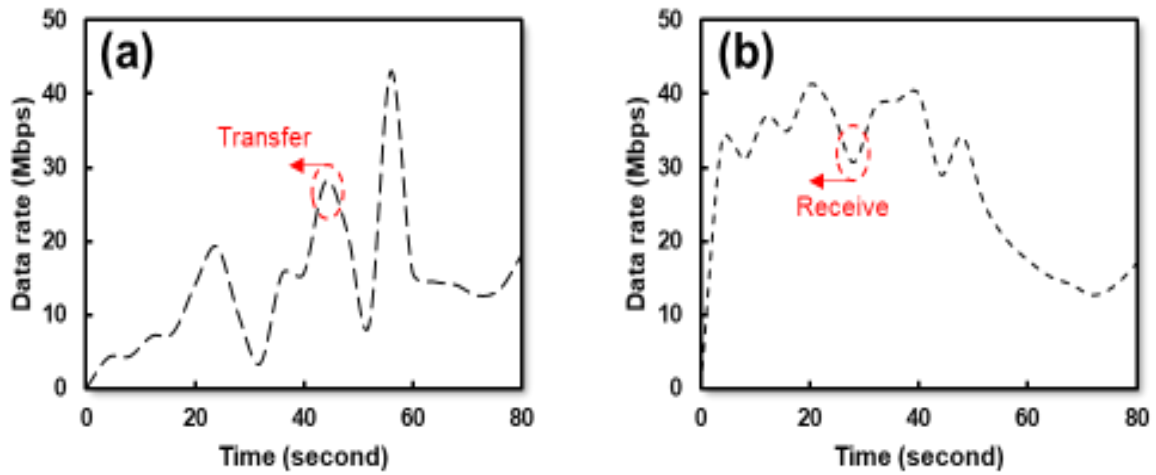


Figure 6.23 Data transfer performance from the proposed device; (a) transfer data rates, (b) receive data rates

Figure 6.23 (a) shows the maximum transfer data rates from the proposed device was 43.01 Mbps while maximum receive data rates was 41.32 Mbps as shown in Figure 6.23 (b), respectively. Furthermore, the average for transfer and reception data rates was 13.32 Mbps and 26.22 Mbps. These results indicate that the proposed device has functioned as an antenna to send and receive data using a wireless network with good performance. The speed of transfer and reception data rates is greatly influenced by the quality of the signal and the size of the data packet being sent. This finding proves that the proposed microwave sensor can function as an antenna for data transfer capabilities.

6.8 Validation with existing works

To demonstrate the novelty and contribution of the proposed sensor, a comparison with the previously described sensors is comprehensively shown in Table 6.5. Based on the comparison with previous studies, the novelty and main contribution of this research was shown by its multifunctionality as a permittivity sensor and an antenna for data transfer capabilities. In addition, the sensor had independent characteristics with two different sensing areas. Therefore, it can be used to detect the permittivity of two different types of MUT simultaneously. The proposed sensor had high accuracy with a compact and simple structure using a single layer thus allowing a simple measurement and easy integration with other devices.

Table 6.5 Comparison with sensors from previous studies

Ref.	Model	Parameter															
		Range of ϵ_r	MUT	Frequency (GHz)			Average of Accuracy (%)			NS (%)			Integrated with antenna	Layer	Sensing area	Independent performance	Simultaneous measurement
				f_{r1}	f_{r2}	f_{r3}	f_{r1}	f_{r2}	f_{r3}	f_{r1}	f_{r2}	f_{r3}					
(Armghan et al., 2021)	CSRR resonator	1-11.9	Solid	5.35	7.99	-	94.24	97.76	-	3.40	3.53	-	-	1	1	-	-
(Wang et al., 2021)	Interdigital capacitors	1-10.5	Solid	2.51	-	-	99.40	-	-	3.98	-	-	-	1	1	-	-
(Lim, Kim and Hong, 2018)	Microstrip line ring resonator	1-3.48	Solid	4.89	9.81	-	92.50	-	-	13.00	26.00	-	-	1	1	-	-
(Yeo and Lee, 2019a)	Interdigital capacitors	1-10.2	Solid	1.50	-	-	95.30	-	-	4.40	-	-	-	1	1	-	-
(Lee and Yang, 2014)	CSRR	1-10.2	Solid	2.4	-	-	97.79	-	-	NA	-	-	-	1	1	-	-
(Ebrahimi, Scott and Ghorbani, 2018b)	LC resonators	1-10.2	Solid	2.00	-	-	99.90	-	-	3.28	-	-	-	1	2	-	-
(Ansari, Jha and Akhtar, 2015)	SRR	1-3	Solid	2.65	-	-	97.71	-	-	5.38	-	-	-	1	1	-	-
(Piekarz et al., 2017)	Couple line section	1-10.2	Solid	1.50	-	-	97.50	-	-	NA	-	-	-	1	1	-	-
(Yeo and Lee, 2019b)	Slot loaded	1-10.2	Solid	2.50	-	-	98.62	-	-	5.65	-	-	-	1	1	-	-
(Behdani et al., 2020)	Apperture coupling antenna	1-12.85	Solid	9.54	12.30	-	-	92.30	-	-	0.64	-	yes	3	1	-	-
This work	Multicascode T-shaped resonator	1-6.15	Solid	2.45	2.00	1.50	-	96.81	97.54	-	0.68	1.77	yes	1	2	yes	yes

6.9 Concluding remarks

In this research, multifunctional dual-band microwave sensors and antenna using multicascode T-shaped resonators for simultaneous measurement of solid materials and data transfer capabilities successfully designed and implemented. The 1st resonator operating at $f_{r1} = 2.45$ GHz was used as an antenna while the 2nd and 3rd resonators operating at $f_{r2} = 2$ GHz and $f_{r3} = 1.5$ GHz were used as a permittivity sensor. The proposed sensor had independent characteristics with two different sensing areas so that it can be used to determine the permittivity of two different types of MUTs simultaneously. From the measurement results, the average accuracy of the proposed sensor was 96.81% and 97.54% while the normalized sensitivity (NS) was 0.68 % and 1.77% with a permittivity range of 1-6.15. Furthermore, the performance of the antenna did not change significantly when the MUT was placed simultaneously on both sensing areas. In addition, the proposed microwave sensor can also function as an antenna for data transfer capabilities with a average transfer and reception data rates of 13.32 Mbps and 26.22 Mbps with file size of 150 MB. The proposed sensor can be recommended as a multifunctional sensor that can be integrated with other devices via wireless for the food industry, biomedical industry, and pharmaceutical industry.

CHAPTER 7

CONCLUSION AND RECOMMENDATIONS FOR FUTURE RESEARCH

This chapter provides conclusions and suggestions for future work based on the three topics discussed in chapter 4, chapter 5 and chapter 6.

7.1 Conclusion

This thesis contributes to observing the interaction between the characteristics and performance of planar resonators and the permittivity of solid materials. The planar resonator is proposed as a microwave sensor to detect the permittivity of solid materials using the perturbation theory by placing the sample in the sensing area of the sensor which is determined by the highest electric field based on the resonant frequency of the resonator. The sensors are designed to operate at different resonant frequencies and have different sensing area locations. Furthermore, the permittivity of the sample is extracted using a polynomial equation and verified with the permittivity of the datasheet. The original contribution of this thesis is presented in three parts as follows:

In the first part, a microwave sensor is developed utilizing a planar resonator with a single port based on the dual U-shaped. The dual U-shaped resonator has been introduced to produce two different sensing areas at the resonant frequency, namely $f_{r1} = 1.20$ GHz and $f_{r2} = 2.10$ GHz. The two sensing areas are located at different locations and have independent characteristics so that they can be used to characterize materials independently and simultaneously. The proposed sensor had a very high accuracy of 99.02 % and 96.44 % compared with permittivity from datasheet with range of 1 – 4.3, respectively. Furthermore, the structure of the proposed sensor using a single port makes it possible to utilize it as an antenna. However, the radiation of proposed resonator is still low, so it is

necessary to optimize it so that it can transmit electromagnetic waves. In addition, the sample being characterized is still placed directly on the surface of the sensor so that it can be corrosive and damage the surface of the resonator.

In the second part, dual T-shaped resonators are introduced to integrate the microwave sensor (MS) and antenna sensor (AS). The sensor is designed to operate at dual frequencies, namely $f_{r1} = 2.35$ GHz and $f_{r2} = 1.81$ GHz. The functionality of the microwave sensor and antenna sensor is determined based on the reflection coefficient and the maximum radiation from the resonator at each resonance frequency where the microwave sensor is used for contact detection and the antenna sensor is used for non-contact detection. MS and AS were determined based on the frequency shift on the reflection coefficient (S_{11}) where S_{11} of ≥ -10 dB was proposed as MS and S_{11} of ≤ -10 dB was proposed as AS. The sensing area was determined based on the concentration of the electric E and H-fields of each resonator. Furthermore, the surface of the resonator which had a high concentration of E-field and a low concentration of H-field was used as an MS for contact detection. On the other hand, non-contact detection utilized the surface of the resonator which has the same location as E and H-fields. Generally, resonators that function as antennas transmit electromagnetic waves where the E and H-fields were in the same location, therefore it can be utilized for AS. The proposed sensor has independent characteristics with different sensing area locations depending on the resonant frequency. Based on the measurement results, the microwave sensor and antenna sensor have good performance with an accuracy of 99.30 % and 99.31 % for contact and non-contact detection with permittivity range of 1 – 6.15, respectively. Interestingly, non-contact sensors can also detect permittivity of samples with air gaps in the range of 0.5 - 1.50 mm. It should be noted, the air gap greatly affects the sensitivity of the sensor during non-contact detection. These findings indicate that the proposed work has successfully integrated microwave sensors and antenna sensors for contact and non-contact detection

and is very useful for the food industry, biomedical industry and quality control of materials.

In the final part, a multicascode T-shaped resonator is proposed as a microwave sensor and antenna for solid material characterization and data transfer capability. The proposed sensor is designed to operate at three different frequencies namely $f_{r1} = 2.46$ GHz, $f_{r2} = 2.01$ GHz and $f_{r3} = 1.57$ GHz with different sensing area locations. The functionality of the microwave sensor and antenna is determined based on the maximum radiation of the resonator at each resonant frequency. The high frequency (f_{r1}) is used as an antenna for data transfer purposes while the low frequency (f_{r2} and f_{r3}) is used to characterize two types of samples with different permittivities independently and simultaneously. The proposed sensor had independent characteristics with two different sensing areas so that it can be used to determine the permittivity of two different types of MUTs simultaneously. From the measurement results, the average accuracy of the proposed sensor was 96.81% and 97.54% while the normalized sensitivity (NS) was 0.68 % and 1.77% with a permittivity range of 1-6.15. Interestingly, the performance of the antenna did not change significantly when the MUT was placed simultaneously on both sensing areas. In addition, the proposed microwave sensor can also function as an antenna for data transfer capabilities with an average transfer and reception data rates of 13.32 Mbps and 26.22 Mbps with file size of 150 MB. The proposed sensor can be recommended as a multifunctional sensor that can be integrated with other devices via wireless for the food industry, biomedical industry, and pharmaceutical industry.

7.2 Suggestion for future works

Multifunctional and independent microwave sensors have several advantages, for example, they can be used for simultaneous characterization and are also easy to integrate

with other devices such as internet-of-things (IoT), Wi-Fi and Zigbee. Therefore, the recommendation for future work from this study is to implement sensors and antennas simultaneously for real time measurements as shown in Figure 7.1.

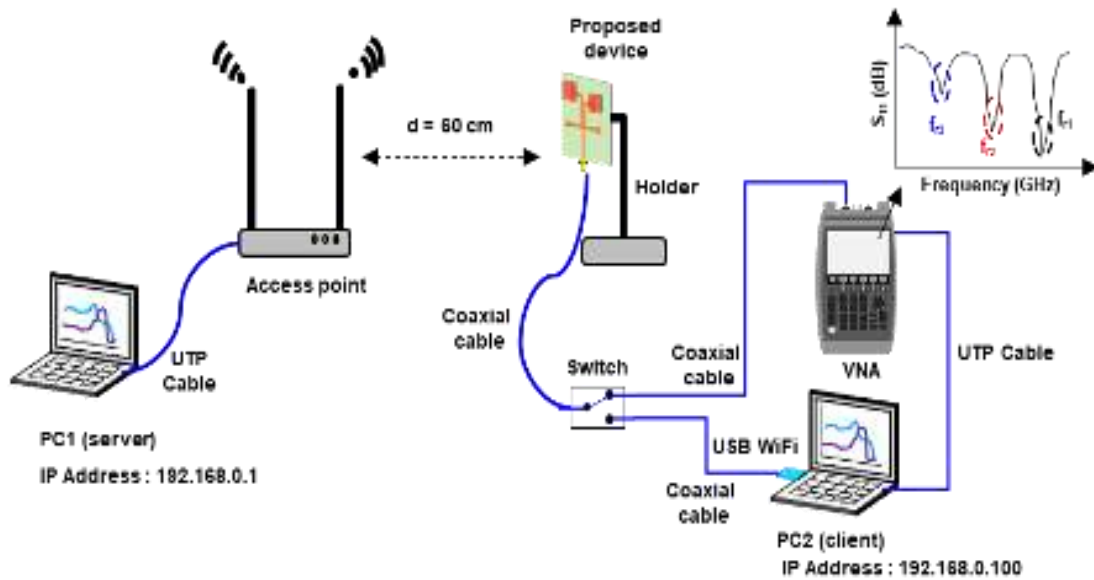


Figure 7.1 Possibility scenario of permittivity detection and data transfer using proposed device

Figure 7.1 shows the possibility scenario of the permittivity detection process from MUT and data transfer using the proposed device. The permittivity of the MUT is detected using proposed device as microwave sensor connected to the VNA, the measurement data is stored on a PC2 which is also connected to the VNA using a UTP cable. Furthermore, to change the function of the proposed device as an antenna, a switch is proposed as a selector. Data from the measurement process is sent using the proposed device as an antenna connected to USB Wi-Fi operating at a resonant frequency of 2.45 GHz to a PC server connected to the access point. Furthermore, data transfer capability is monitored on PC1 as a server using an access point. Therefore, real-time permittivity detection using the proposed device simultaneously as sensor and antenna can be recommended as a future work.

Another recommendation for future work is to utilize the independent characteristics of the sensor to simultaneously characterize low and high permittivity samples for example solid materials and liquid materials. Generally, microwave sensors are only used for the detection of solid or liquid materials. For this reason, by using independent characteristics, sensors can be designed to simultaneously detect solid and liquid materials at different sensing area locations simultaneously as shown in Figure 7.2.

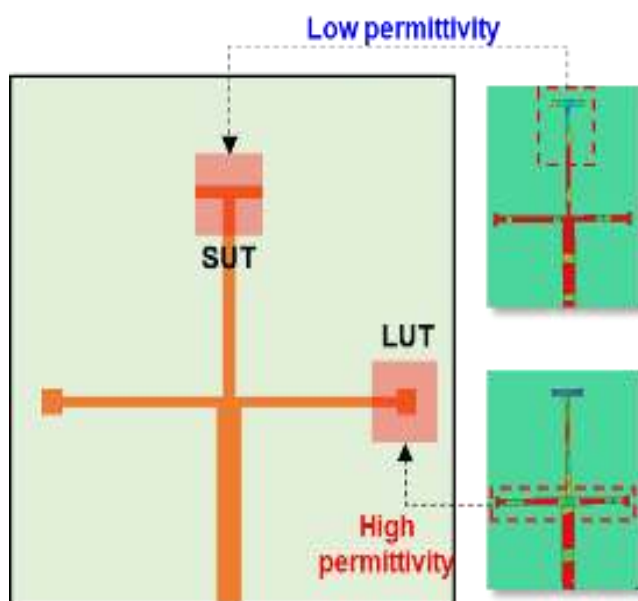


Figure 7.2 Possibility scenario of permittivity detection and data transfer using proposed device

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APPENDICES

Appendix A

Datasheet permittivity of material under test

Datasheet of FR-4

Welcome to Nam Hing Industrial Laminate Ltd.

http://www.nh-laminate.com.hk/our_products/fr4-4806.htm



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Copper-Clad Glass-Epoxy Laminates :

Type	NHL-4806	Conditions
ANSI (NEMA) Grade	FR4	
UL File No.	E74035	
Colour	Natural	
Nominal Thickness and Tolerance	0.4mm (0.016 in) +/- 0.063mm (0.0025 in) 0.6mm (0.023 in) +/- 0.063mm (0.0025 in) 0.8mm (0.031 in) +/- 0.089mm (0.0035 in) 1.2mm (0.047 in) +/- 0.130mm (0.0051 in) 1.6mm (0.062 in) +/- 0.130mm (0.0051 in) (for nominal thickness not shown, the tolerance for the next greater thickness shall apply)	
Standard Sheet Size	1,020mm x 1,220mm +/- 5mm	
Copper Cladding	135 g/m ² (1/2 oz/ft ²), 18 mm ; 305 g/m ² (1 oz/ft ²), 35 mm ; 610 g/m ² (2 oz/ft ²), 70 mm ; one- or two-sided electrolytic copper cladding	
Max. Operating Temp (Tg)	140 °C (Electrical) +/- 5 °C 110 °C (Mechanical) +/- 5 °C	
Flame Class	94-V0	UL-94
Specific Gravity	1.185	
Water Absorption	0.10% (For 1.6mm)	
Solder Float (260 °C)	Over 12 sec	

Peel Strength	1.8 N/mm (Normal state) 1.7 N/mm (Thermal shock, 550 °F 10s) 1.6 N/mm (Post exposure to process solution)	IPC-TM-650 (For 1/2 oz Cu)
Flexural Strength	618.3 N/mm ² For 0.8mm 735.1 N/mm ² For 1.6mm	ASTM D 790
Insulation Resistivity	8.47×10^{11} W (Normal state)	IPC-TM-650
Volume Resistivity	7.5×10^{10} W m (Normal state, 500Vdc) 5.2×10^{10} W m (Damp resistance test)	IPC-TM-650
Dielectric Constant	4.3 (Damp heat, steady state 96h, 1Mhz)	IPC-TM-650
Dielectric Strength	29.4kv/mm (Normal state) 24.0kv/mm (Moist state)	ASTM D 149
Dispation Factors	0.0265 (Damp heat, steady state 96h, 1Mhz)	IPC-TM-650
Bow & Twist	0.31% Bowed and No Twist	IPC-TM-650
High Current Arc Resistance Test	17 arc for 0.8mm 18 arc for 1.6mm	UL 746
Hot Wire Ignition	120+ sec to ignition for 0.8mm 120+ sec to ignition for 1.6mm	UL 746
Arc Resistance	93 sec	ASTM D 495
High Voltage Arc Resistance	300+ sec	UL 746
Arc Tracking Rate	368mm/min for 0.8mm 267mm/min for 1.6mm	UL 746
Comparative Tracking Index	175	IEC 112

The properties described herein are based upon test data generated at independent laboratories and is believed to be accurate. However, the final determination regarding the suitability of the material described herein for any contemplated use is the sole responsibility of the end user. There is no warranty, expressed or implied.

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Datasheet of RO-5880



RT/duroid® 5870 /5880 High Frequency Laminates



RT/duroid® 5870 and 5880 glass microfiber reinforced PTFE composites are designed for exacting stripline and microstrip circuit applications.

The randomly oriented microfibers result in exceptional dielectric constant uniformity.

The dielectric constant of RT/duroid 5870 and 5880 laminates is uniform from panel to panel and is constant over a wide frequency range.

Its low dissipation factor extends the usefulness of RT/duroid 5870 and 5880 laminates to Ku-band and above.

RT/duroid 5870 and 5880 laminates are easily cut, sheared and machined to shape. They are resistant to all solvents and reagents, hot or cold, normally used in etching printed circuits or in plating edges and holes.

Normally supplied as a laminate with electrodeposited copper of $\frac{1}{16}$ to 2 ounces/ft.² (6 to 70µm) or reverse treated EDC on both sides, RT/duroid 5870 and 5880 composites can also be clad with rolled copper foil for more critical electrical applications. Cladding with aluminum, copper or brass plate may also be specified.

When ordering RT/duroid 5870 and 5880 laminates, it is important to specify dielectric thickness, tolerance, rolled, electrodeposited or reverse treated copper foil, and weight of copper foil required.

Data Sheet



Features:

- Lowest electrical loss for reinforced PTFE material
- Low moisture absorption
- Isotropic
- Uniform electrical properties over frequency
- Excellent chemical resistance

Some Typical Applications:

- Commercial Airline Broadband Antennas
- Microwave and Stripline Circuits
- Millimeter Wave Applications
- Military Radar Systems
- Missile Guidance Systems
- Point to Point Digital Radio Antennas

Datasheet of RO-4003C



RO4000® Series High Frequency Circuit Materials

RO4000® hydrocarbon ceramic laminates are designed to offer superior high frequency performance and low cost circuit fabrication. The result is a low loss material which can be fabricated using standard epoxy/glass (FR-4) processes offered at competitive prices.

The selection of laminates typically available to designers is significantly reduced once operational frequencies increase to 500 MHz and above. RO4000 material possesses the properties needed by designers of RF microwave circuits and matching networks and controlled impedance transmission lines. Low dielectric loss allows RO4000 series material to be used in many applications where higher operating frequencies limit the use of conventional circuit board laminates. The temperature coefficient of dielectric constant is among the lowest of any circuit board material (Chart 1), and the dielectric constant is stable over a broad frequency range (Chart 2). For reduced insertion loss, LoPro® foil is available (Chart 3). This makes it an ideal substrate for broadband applications.

RO4000 material's thermal coefficient of expansion (CTE) provides several key benefits to the circuit designer. The expansion coefficient of RO4000 material is similar to that of copper which allows the material to exhibit excellent dimensional stability, a property needed for mixed dielectric multi-layer boards constructions. The low Z-axis CTE of RO4000 laminates provides reliable plated through-hole quality, even in severe thermal shock applications. RO4000 series material has a Tg of >280°C (536°F) so its expansion characteristics remain stable over the entire range of circuit processing temperatures.

RO4000 series laminates can easily be fabricated into printed circuit boards using standard FR-4 circuit board processing techniques. Unlike PTFE based high performance materials, RO4000 series laminates do not require specialized via preparation processes such as sodium etch. This material is a rigid, thermoset laminate that is capable of being processed by automated handling systems and scrubbing equipment used for copper surface preparation.

RO4003C™ laminates are currently offered in various configurations utilizing both 1080 and 1674 glass fabric styles, with all configurations meeting the same laminate electrical performance specification. Specifically designed as a drop-in replacement for the RO4003C™ material, RO4350B™ laminates utilize RoHS compliant flame-retardant technology for applications requiring UL 94V-0 certification. These materials conform to the requirements of IPC-4103, slash sheet /10 for RO4003C, see note #1 for RO4350B slash sheet determination.



Data Sheet



FEATURES AND BENEFITS:

RO4000 materials are reinforced hydrocarbon/ceramic laminates - not PTFE.

- Designed for performance sensitive, high volume applications

Low dielectric tolerance and low loss

- Excellent electrical performance
- Allows applications with higher operating frequencies
- Ideal for broadband applications

Stable electrical properties vs. frequency

- Controlled impedance transmission lines
- Repeatable design of filters

Low thermal coefficient of dielectric constant

- Excellent dimensional stability

Low Z-axis expansion

- Reliable plated through holes

Low in-plane expansion coefficient

- Remains stable over an entire range of circuit processing temperatures

Volume manufacturing process

- RO4000 laminates can be fabricated using standard glass epoxy processes
- Competitively priced

CAF resistant

SOME TYPICAL APPLICATIONS:

- Cellular Base Station Antennas and Power Amplifiers
- RF Identification Tags
- Automotive Radar and Sensors
- LNBs for Direct Broadcast Satellites



Property	Typical Value		Direction	Units	Condition	Test Method
	RO4003C	RO4320B				
Dielectric Constant, ϵ_r Process	3.38 ± 0.05	3.38 ± 0.05	Z	-	10 GHz/23°C	IPC TM 650 2.5.5.5 Clamped Sample
ϵ_r Dielectric Constant, ϵ_r Design	3.55	3.66	Z	-	8 to 40 GHz	Differential Phase Length Method
Dissipation Factor tan, d	0.0027 0.0027	0.0037 0.0037	Z	-	10 GHz/23°C 2.5 GHz/23°C	IPC TM 650 2.5.5.5
Thermal Coefficient of ϵ_r	+40	+59	Z	ppm/°C	30°C to 150°C	IPC TM 650 2.5.5.5
Volume Resistivity	1.7 E 10 ¹²	1.2 E 10 ¹²	-	MΩ-cm	COND A	IPC TM 650 2.5.17.1
Surface Resistivity	4.2 E 10 ¹²	5.7 E 10 ¹²	-	MΩ	COND A	IPC TM 650 2.5.17.1
Electrical Strength	21.2 (1968)	21.2 (1968)	Z	kV/mm (V/mil)	0.51mm (0.020")	IPC TM 650 2.5.6.3
Tensile Modulus	19,690 (2,830) 19,450 (2,821)	10,707 (2,432) 14,183 (2,093)	X Y	MPa (ksi)	RT	ASTM D638
Tensile Strength	129 (20.2) 100 (14.5)	202 (29.5) 130 (18.8)	X Y	MPa (ksi)	RT	ASTM D638
Flexural Strength	270 (40)	255 (37)	-	MPa (ksi)	-	IPC TM 650 2.4.4
Dimensional Stability	<0.3	<0.5	X,Y	mm/m (in/inch)	after etch +125/130°C	IPC TM 650 2.4.39A
Coefficient of Thermal Expansion	11 14 46	10 12 42	X Y Z	ppm/°C	-55 to 200°C	IPC TM 650 2.4.41
T _g	>280	>280	-	°C (Mk)	A	IPC TM 650 2.4.24.2
T _d	425	390	-	°C TGA	-	ASTM D3490
Thermal Conductivity	0.71	0.69	-	W/m·K	80°C	ASTM C118
Moisture Absorption	0.06	0.06	-	%	48 hrs Immersion 0.000" sample Temperature 30°C	ASTM D570
Density	1.75	1.86	-	g/cm ³	23°C	ASTM D792
Copper Peel Strength	1.05 (0.0)	0.80 (0.0)	-	N/mm (g/ft)	after solder float 1 sec. HIC Peel	IPC TM 650 3.4.8
Flammability	UL94	UL94	-	-	-	UL 94
Lead-Free Process Compatible	Yes	Yes	-	-	-	-

NOTES:

- RO4003C and laminates have a process DR of 0.03 ± 0.05 and are in conformance with IPC-4101A/240. All other RO4003C laminate thicknesses are 7.1 and 340 compliant.
- The design DR is an average number from several different tested lots of material and on the most common thickness. If more detailed information is required, please contact Rogers Corporation or refer to Rogers' technical papers in the Rogers Technology Support Hub available at <http://www.rogerscorp.com>.
- RO4320B copper laminate does not share the same UL designation as standard RO4003C laminates. A separate UL qualification may be necessary.

Typical values are a representation of an average value for the population of the property. For specification values contact Rogers Corporation.

RO4003C LCP laminate uses a modified version of the RO4003C resin system to bond reverse seeded fill. Values shown above are RO4003C laminates without the addition of the LCP resin. Copper foil results in an overall thickness increase of approximately 0.0007" (18µm) per core.

The copper resin DR is approximately 2.4. However, when used in combination with the base laminate system, the average design DR noted in the data sheet table should be used. (The design DR value decreases by about 0.1 as the core thickness decreases from 0.001" to 0.004").

Prolonged exposure in an oxidative environment may cause changes to the dielectric properties of hydrocarbon based materials. The rate of change increases at higher temperatures and is highly dependent on the circuit design. Although Rogers' high frequency materials have been used successfully in innumerable applications and reports of oxidation resulting in performance problems are extremely rare, Rogers recommends that the customer evaluate each material and design combination to determine fitness for use over the entire life of the end product.

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Standard Thicknesses	Standard Panel Sizes	Standard Claddings
RO4001C: 0.008" (0.203mm) +/- 0.0010" 0.012" (0.305mm) +/- 0.0010" 0.016" (0.406mm) +/- 0.0010" 0.020" (0.508mm) +/- 0.0010" 0.032" (0.813mm) +/- 0.0020" 0.060" (1.524mm) +/- 0.0040"	24" X 12" (610 X 437 mm) 24" X 24" (610 X 562 mm) 24" X 36" (610 X 913 mm) 48" X 36" (1219 X 913 mm) *Additional panel sizes available	<u>Electrodeposited Copper Foil</u> 1 to 2 (18µm) Hx/WH 1 to 2 (33µm) Hx/WH *Additional cladding weights are available
RO4350B: 0.004" (0.10mm) +/- 0.0007" 0.006" (0.17mm) +/- 0.0007" 0.010" (0.25mm) +/- 0.0010" 0.020" (0.51mm) +/- 0.0015" 0.030" (0.76mm) +/- 0.0020" 0.060" (1.52mm) +/- 0.0040"		
*Additional non-standard thicknesses available from 0.0002" - 0.080" in varying increments		

*Contact Customer Service or Sales Engineering to inquire about additional available product configurations

The information in this data sheet is intended to assist you in designing with Rogers' circuit materials. It is not intended to and does not create any warranties express or implied, including any warranty of merchantability or fitness for a particular purpose or that the results shown on this data sheet will be achieved by a user for a particular purpose. The user should determine the suitability of Rogers' circuit materials for each application.

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Helping power, protect, connect our world®



RO3000® Series Circuit Materials

RO3003™, RO3006™, RO3010™ and RO3035™
High Frequency Laminates

RO3000® high frequency circuit materials are ceramic-filled PTFE composites intended for use in commercial microwave and RF applications. This family of products was designed to offer exceptional electrical and mechanical stability at competitive prices.

RO3000 series laminates are ceramic-filled PTFE based circuit materials with mechanical properties that are consistent regardless of the dielectric constant selected. This allows the designer to develop multi-layer board designs that use different dielectric constant materials for individual layers, without encountering warpage or reliability problems.

RO3000 materials exhibit a coefficient of thermal expansion (CTE) in the X and Y axis of 17 ppm/°C. This expansion coefficient is matched to that of copper, which allows the material to exhibit excellent dimensional stability, with typical etch shrinkage (after etch and bake) of less than 0.5 mils per inch. The Z-axis CTE is 24 ppm/°C, which provides exceptional plated through-hole reliability, even in severe thermal environments. The dielectric constant versus temperature for RO3003™ and RO3035™ materials is very stable (Chart 1).

RO3000 series laminates can be fabricated into printed circuit boards using standard PTFE circuit board processing techniques, with minor modifications as described in the application note "Fabrication Guidelines for RO3000 Series High Frequency Circuit Materials."



Data Sheet



Features and Benefits:

Low dielectric loss (RO3003™ laminates)

- Laminates can be used in applications up to 27 GHz

Excellent mechanical properties versus temperature

- Reliable stripline and multi-layer board constructions

Uniform mechanical properties for a range of dielectric constants

- Ideal for multi-layer board designs with a range of dielectric constants
- Suitable for use with epoxy glass multi-layer board hybrid designs

Stable dielectric constant versus temperature and frequency (RO3003™ laminates)

- Ideal for band pass filters, microstrip patch antennas, and voltage controlled oscillators

Low in-plane expansion coefficient (match to copper)

- Allows for more reliable surface mounted assemblies
- Ideal for applications sensitive to temperature change
- Excellent dimensional stability volume manufacturing process
- Economical laminate pricing

Some Typical Applications:

- Automotive radar applications
- Global positioning satellite antennas
- Cellular telecommunications systems / power amplifiers and antennas
- Patch antennas for wireless communications
- Direct broadcast satellites
- Database on cable systems
- Remote meter readers
- Power backplanes

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Property	Typical Value ⁽¹⁾				Direction	Unit	Condition	Test Method
	RO3003	RO3035	RO3006	RO3010				
Dielectric Constant, ϵ_r Process	3.00 ± 0.04	3.30 ± 0.03	3.13 ± 0.13	10.2 ± 0.82	Z	-	10 GHz 23°C	IPC-89-600 2.3.5.3 Clamped Stripline
Dielectric Constant, ϵ_r Design	3.00	3.30	3.30	11.20	Z	-	8 GHz - 40 GHz	Differential Phase Length Method
Dissipation Factor, $\tan \delta$	0.0010	0.0015	0.0020	0.0022	Z	-	10 GHz 23°C	IPC-89-600 2.3.5.3
Thermal Coefficient of ϵ_r	-1	-43	-282	-390	Z	ppm/°C	10 GHz -50 to 150°C	IPC-89-600 2.3.5.3
Dimensional Stability	-0.08 0.07	-0.11 0.11	-0.27 -0.15	-0.25 -0.21	X Y	mm/m	COND A	IPC-6013 2.2.4
Volume Resistivity	10 ¹²	10 ¹²	10 ¹²	10 ¹²		MΩ·cm	COND A	IPC 2.3.17.1
Surface Resistivity	10 ¹²	10 ¹²	10 ¹²	10 ¹²		MΩ	COND A	IPC 2.3.17.1
Tensile Modulus	958 623	1023 1066	1496 1202	1902 1324	X Y	GPa	23°C	ASTM D3039
Moisture Absorption	0.04	0.04	0.02	0.03	-	%	D48/30	IPC-89-600 2.6.2.1
Specific Heat	0.9		0.86	0.9		J/g·K		Calculated
Thermal Conductivity	0.58	0.50	0.79	0.95	-	W/m·K	23°C	ASTM D5470
Coefficient of Thermal Expansion (-35 to 260°C)	17 10 25	17 17 24	17 17 24	18 11 16	X Y Z	ppm/°C	23°C/30% RH	IPC-19-600 2.4.41
Td	500	500	500	500		°C TGA		ASTM D3850
Density	2.1	2.1	2.0	2.0		g/cm ³	23°C	ASTM D792
Copper Peel Strength	12.7	10.2	7.3	9.4		lb/in	1 on 30C after solder flux	IPC-TM-2.4.8
Flammability	V-0	V-0	V-0	V-0				UL 94
Lead-Free Process Compatible	YES	YES	YES	YES				

NOTES:

(1) Typical values are a representation of an average value for the population of the property. For specification values contact Rogers Corporation.

(2) The design DR is an average number from several different tested lots of material and on the most common thicknesses. If more detailed information is required, please contact Rogers Corporation or refer to Rogers' technical papers in the Rogers Technology Support Hub available at <http://www.rogerscorp.com/techhub>.

Standard Thicknesses	Standard Panel Sizes	Standard Claddings
RO3004 0.005" (0.13mm) $\pm$ 0.0005" 0.007" (0.18mm) $\pm$ 0.0007" 0.008" (0.21mm) $\pm$ 0.0010" 0.010" (0.25mm) $\pm$ 0.0015" 0.012" (0.32mm) $\pm$ 0.0020" RO3006/RO3010 0.005" (0.13mm) $\pm$ 0.0005" 0.007" (0.18mm) $\pm$ 0.0007" 0.008" (0.21mm) $\pm$ 0.0010" 0.010" (0.25mm) $\pm$ 0.0015" 0.012" (0.32mm) $\pm$ 0.0020" RO3013 0.007" (0.18mm) $\pm$ 0.0007" 0.010" (0.25mm) $\pm$ 0.0010" 0.012" (0.32mm) $\pm$ 0.0015" *Additional non-standard thicknesses available from 0.005" to 0.149" in increments of 0.005"	RO3005/RO3006/RO3010/RO3015 12" X 18" (305 X 457mm) 24" X 18" (610 X 457mm) RO3003 0.005" and 0.010" 12" X 18" (305 X 457mm) 24" X 18" (610 X 457mm) 24" X 21" (610 X 533mm) *Additional panel sizes available	RO3004 <u>Electrodeposited Copper Foil</u> 15 oz. (18µm) (H4/H6) 1 oz. (33µm) (H1/H11) <u>Rolls Copper Foil</u> 15 oz. (18µm) (AP/H6) 1 oz. (33µm) (A1/A12) RO3005/RO3010/RO3015 <u>Electrodeposited Copper Foil</u> 15 oz. (18µm) (H4/H6) 1 oz. (33µm) (H1/H11) *Additional claddings, such as heavy metal, resistive foil and underclad are available

*Contact Customer Service or Sales Engineering to inquire about additional available product configurations

The information in this data sheet is intended to assist you in designing with Rogers' circuit materials. It is not intended to and does not create any warranties express or implied, including any warranty of merchantability or fitness for a particular purpose or that the results shown on this data sheet will be achieved by a user for a particular purpose. The user should determine the suitability of Rogers' circuit materials for each application.

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Revised 1509 000122 Publication P32-130

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Appendix B

Data of independent performance of dual U-shaped resonator

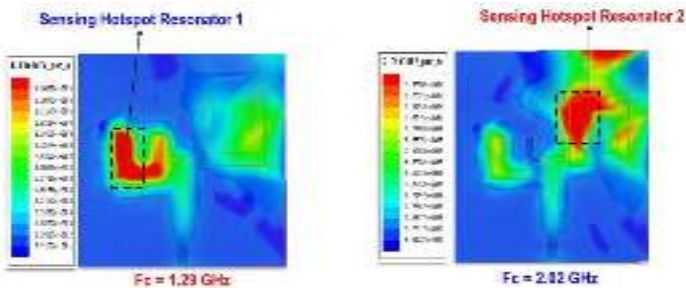
Independent performance of C₁

C
1

No	X1	X2	Y1	Y2
1	1	1	1.21	2.1
2	1	2.2	1.21	2.09
3	1	3.38	1.21	2.07
4	1	4.3	1.21	2.04
5	2.2	1	1.2	2.1
6	2.2	2.2	1.2	2.08
7	2.2	3.38	1.2	2.07
8	2.2	4.3	1.2	2.05
9	3.38	1	1.2	2.1
10	3.38	2.2	1.2	2.09
11	3.38	3.38	1.2	2.08
12	3.38	4.4	1.2	2.07
13	4.3	1	1.19	2.1
14	4.3	2.2	1.19	2.08
15	4.3	3.38	1.19	2.07
16	4.3	4.3	1.19	2.05



Correlation of C ₁		
	Y1	Y2
X1	-94%	7%
X2	0.0%	-92%



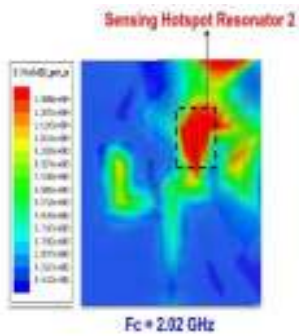
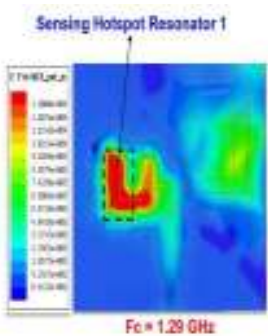
Indepedendent performance of C₂

C₂

No	X1	X2	Y1	Y2
1	1	1	1.21	2.1
2	1	2.2	1.21	2.09
3	1	3.38	1.21	2.08
4	1	4.3	1.21	2.06
5	2.2	1	1.2	2.1
6	2.2	2.2	1.2	2.08
7	2.2	3.38	1.2	2.07
8	2.2	4.3	1.2	2.06
9	3.38	1	1.2	2.1
10	3.38	2.2	1.2	2.08
11	3.38	3.38	1.2	2.07
12	3.38	4.4	1.2	1.06
13	4.3	1	1.19	2.1
14	4.3	2.2	1.19	2.08
15	4.3	3.38	1.19	2.07
16	4.3	4.3	1.19	2.06



Correlation of C ₂		
	Y1	Y2
X1	-94%	-14%
X2	0%	-40%



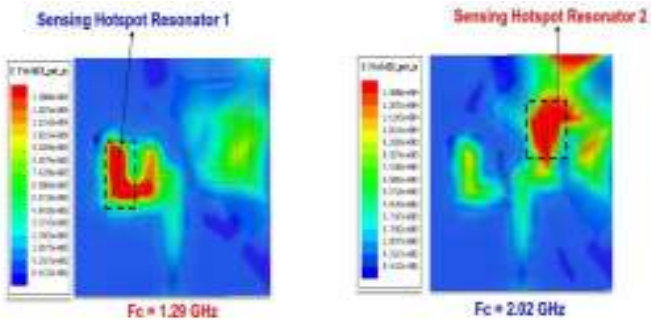
Independent performance of C₃

C₃

No	X1	X2	Y1	Y2
1	1	1	1.21	2.1
2	1	2.2	1.2	2.09
3	1	3.38	1.2	2.08
4	1	4.3	1.19	2.07
5	2.2	1	1.2	2.1
6	2.2	2.2	1.2	2.08
7	2.2	3.38	1.19	2.08
8	2.2	4.3	1.19	2.07
9	3.38	1	1.2	2.1
10	3.38	2.2	1.2	2.1
11	3.38	3.38	1.2	2.1
12	3.38	4.4	1.19	2.07
13	4.3	1	1.19	2.1
14	4.3	2.2	1.19	2.09
15	4.3	3.38	1.19	2.08
16	4.3	4.3	1.19	2.07



Correlation of C ₃		
	Y1	Y2
X1	-49%	10%
X2	-59%	-86%

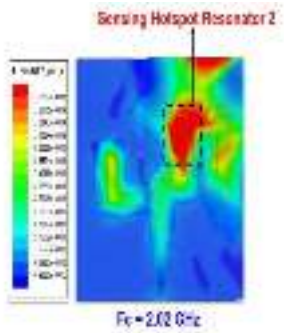
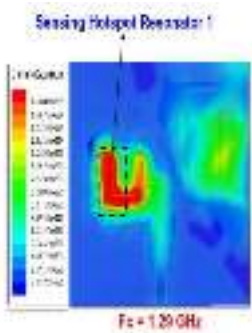


Independent performance of C₄

C ₄				
No	X1	X2	Y1	Y2
1	1	1	1.21	2.1
2	1	2.2	1.21	2.09
3	1	3.38	1.21	2.07
4	1	4.3	1.21	2.04
5	2.2	1	1.2	2.1
6	2.2	2.2	1.2	2.08
7	2.2	3.38	1.2	2.07
8	2.2	4.3	1.2	2.05
9	3.38	1	1.2	2.1
10	3.38	2.2	1.2	2.09
11	3.38	3.38	1.2	2.08
12	3.38	4.4	1.2	2.07
13	4.3	1	1.18	2.09
14	4.3	2.2	1.18	2.08
15	4.3	3.38	1.18	2.07
16	4.3	4.3	1.18	2.05



Correlation of C ₄		
	Y1	Y2
X1	-90.2%	3.0%
X2	0.1%	-90.7%



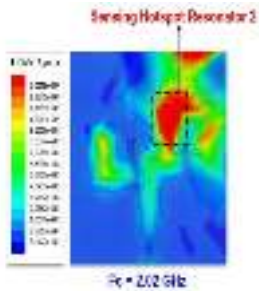
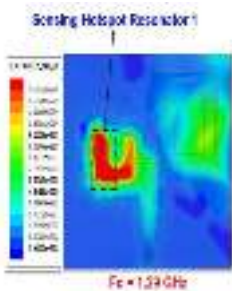
Independent performance of C₅

C₅

No	X1	X2	Y1	Y2
1	1	1	1.21	2.1
2	1	2.2	1.21	2.09
3	1	3.38	1.21	2.07
4	1	4.3	1.21	2.04
6	2.2	1	1.2	2.1
7	2.2	2.2	1.2	2.08
8	2.2	3.38	1.2	2.07
9	2.2	4.3	1.2	2.04
10	3.38	1	1.2	2.1
11	3.38	2.2	1.2	2.09
12	3.38	3.38	1.2	2.08
13	3.38	4.4	1.2	2.07
14	4.3	1	1.18	2.1
15	4.3	2.2	1.18	2.08
16	4.3	3.38	1.18	2.07
17	4.3	4.3	1.18	2.05



Correlation of C ₅		
	Y1	Y2
X1	-90%	8%
X2	0%	-91%



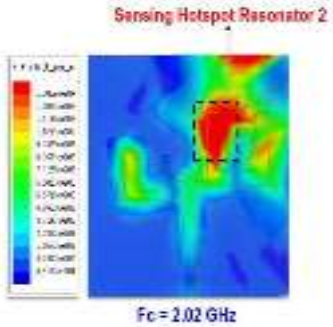
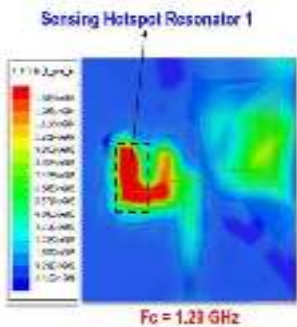
Independent performance of C₆

C₆

No	X1	X2	Y1	Y2
1	1	1	1.21	2.1
2	1	2.2	1.2	2.09
3	1	3.38	1.2	2.08
4	1	4.3	1.19	2.07
6	2.2	1	1.2	2.1
7	2.2	2.2	1.19	2.09
8	2.2	3.38	1.19	2.08
9	2.2	4.3	1.18	2.07
10	3.38	1	1.2	2.1
11	3.38	2.2	1.19	2.09
12	3.38	3.38	1.19	2.08
13	3.38	4.4	1.18	2.07
14	4.3	1	1.18	2.1
15	4.3	2.2	1.18	2.09
16	4.3	3.38	1.18	2.08
17	4.3	4.3	1.17	2.07



Correlation of C ₆		
	Y1	Y2
X1	-72%	0%
X2	-56%	-100%



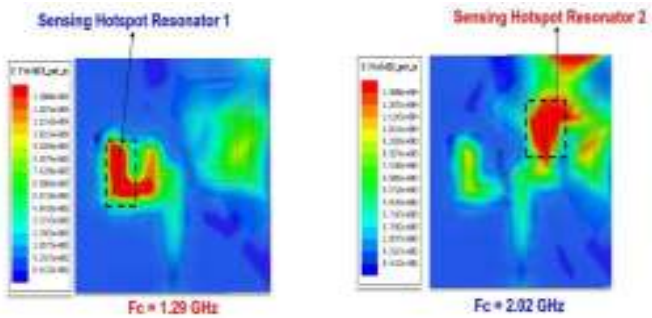
Independendent performance of C₇

C
7

No	X1	X2	Y1	Y2
1	1	1	1.21	2.1
2	1	2.2	1.21	2.09
3	1	3.38	1.21	2.07
4	1	4.3	1.21	2.04
5	2.2	1	1.2	2.1
6	2.2	2.2	1.2	2.08
7	2.2	3.38	1.2	2.07
8	2.2	4.3	1.2	2.04
9	3.38	1	1.19	2.1
10	3.38	2.2	1.19	2.08
11	3.38	3.38	1.19	2.08
12	3.38	4.4	1.19	2.07
13	4.3	1	1.18	2.09
14	4.3	2.2	1.18	2.08
15	4.3	3.38	1.18	2.07
16	4.3	4.3	1.17	2.06



Correlation of C ₇		
	Y1	Y2
X1	-98%	7%
X2	-7 %	-89%



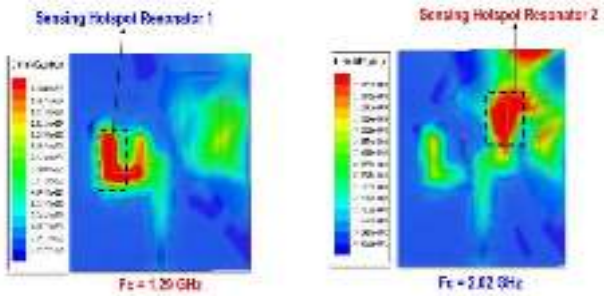
Independent performance of C₈

C₈

No	X1	X2	Y1	Y2
1	1	1	1.21	2.1
2	1	2.2	1.21	2.08
3	1	3.38	1.21	2.07
4	1	4.3	1.2	2.05
5	2.2	1	1.2	2.1
6	2.2	2.2	1.2	2.07
7	2.2	3.38	1.2	2.07
8	2.2	4.3	1.2	2.06
9	3.38	1	1.19	2.1
10	3.38	2.2	1.19	2.08
11	3.38	3.38	1.19	2.08
12	3.38	4.4	1.19	2.06
13	4.3	1	1.18	2.09
14	4.3	2.2	1.18	2.08
15	4.3	3.38	1.18	2.07
16	4.3	4.3	1.18	2.05



Correlation of C ₈		
	Y1	Y2
X1	-97%	-1%
X2	-8%	-93%



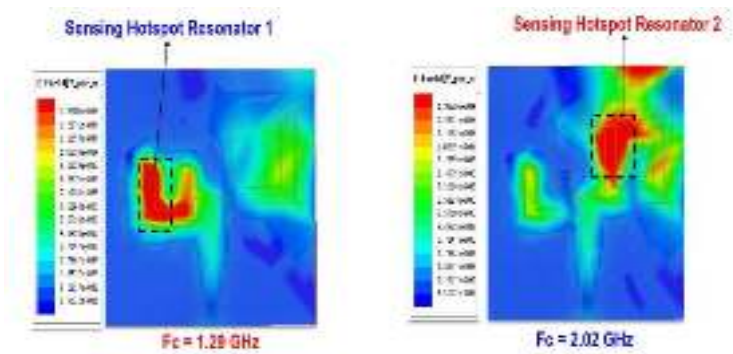
Independent performance of C₉

C₉

No	X1	X2	Y1	Y2
1	1	1	1.21	2.1
2	1	2.2	1.2	2.09
3	1	3.38	1.2	2.08
4	1	4.3	1.19	2.07
5	2.2	1	1.2	2.1
6	2.2	2.2	1.19	2.09
7	2.2	3.38	1.19	2.08
8	2.2	4.3	1.19	2.06
9	3.38	1	1.19	2.1
10	3.38	2.2	1.19	2.09
11	3.38	3.38	1.18	2.09
12	3.38	4.4	1.18	2.07
13	4.3	1	1.18	2.09
14	4.3	2.2	1.17	2.08
15	4.3	3.38	1.17	2.08
16	4.3	4.3	1.16	2.07



Correlation of C ₈		
	Y1	Y2
X1	-85%	-9%
X2	-42%	-90%



Appendix C

Figure of measurement process in laboratorium



Appendix D

Derivated of Polynomial Equation

Curve fitting using regression is a statistical technique used to find the best-fitting curve that describes the relationship between two variables. The goal is to find a mathematical equation that represents the data in the best possible way. Furthermore, a polynomial equation is a mathematical expression consisting of variables, coefficients, and exponents, combined using addition, subtraction, and multiplication operations. Polynomial equations find applications in various fields, including physics, engineering, computer science, and finance. They are used to model and solve problems involving relationships between variables, curve fitting, and optimization. For instance, quadratic equations are commonly used to describe the trajectory of projectiles, while cubic and higher-degree polynomials might model more complex phenomena. Polynomial equations are also used in numerical analysis, interpolation, and approximation techniques. The general form of a polynomial equation can be determined as follows:

$$P(x) = a_n x^n + a_{n-1} x^{n-1} + \dots + a_1 x + a_0$$

where :

- **$P(x)$** is the polynomial function
- **n** is the degree of the polynomial, which is the highest power of **(x)**
- **$a_n, a_{n-1}, \dots, a_1, a_0$** are coefficients, where a_n is the coefficient of the term with $x^{(n)}$, a_{n-1} is the coefficient of the term with $x^{(n-1)}$
- **x** is the variable

To determine the values $a_0, a_1, a_2, a_3, \dots$ we use the following matrix:

$$\begin{bmatrix} n & \sum x & \sum x^2 & \sum x^3 & \dots & \sum x^n \\ \sum x & \sum x^2 & \sum x^3 & \sum x^4 & \dots & \sum x^{n+1} \\ \sum x^2 & \sum x^3 & \sum x^4 & \sum x^5 & \dots & \sum x^{n+2} \\ \sum x^3 & \sum x^4 & \sum x^5 & \sum x^6 & \dots & \sum x^{n+3} \\ \dots & \dots & \dots & \dots & \dots & \dots \\ \sum x^n & \sum x^{n+1} & \sum x^{n+2} & \sum x^{n+3} & \dots & \sum x^{n+m} \end{bmatrix} \begin{bmatrix} a_0 \\ a_1 \\ a_2 \\ a_3 \\ \dots \\ a_n \end{bmatrix} = \begin{bmatrix} \sum x^0 y \\ \sum x^1 y \\ \sum x^2 y \\ \sum x^3 y \\ \dots \\ \sum x^n y \end{bmatrix}$$

Furthermore, the derivation of equations 6.1 and 6.2 uses a second order polynomial approach where the x-axis parameters are represented by frequency and the y-axis parameters are represented by permittivity. Table 1 and table 2 show the values of the x-axis and x-axis parameters for equations 6.1 and 6.2.

Table 1. Value of x and y axis for equation 6.1

X	Y
1	2.01
2.2	1.99
3.55	1.97
4.3	1.94
6.15	1.92

Table 2. Value of x and y axis for equation 6.2

X	Y
1	1.51
2.2	1.48
3.55	1.47
4.3	1.44
6.15	1.41

Furthermore, equations 6.1 and 6.2 can be modeled as follows:

For equation 6.1

$$\begin{bmatrix} n & \sum x & \sum x^2 \\ \sum x & \sum x^2 & \sum x^3 \\ \sum x^2 & \sum x^3 & \sum x^4 \end{bmatrix} \begin{bmatrix} a_0 \\ a_1 \\ a_2 \end{bmatrix} = \begin{bmatrix} \sum x^0 y \\ \sum x^1 y \\ \sum x^2 y \end{bmatrix}$$

Table 3 Value of matrix from equation 6.1

X	Y	X ²	X ³	X ⁴	X ⁰ *Y	X ¹ *Y	X ² *Y
2.01	1	4.04	8.12	16.32	1.00	2.01	4.04
1.99	2.2	3.96	7.88	15.68	2.20	4.38	8.71
1.97	3.55	3.88	7.65	15.06	3.55	6.99	13.78
1.94	4.3	3.76	7.30	14.16	4.30	8.34	16.18
1.92	6.15	3.69	7.08	13.59	6.15	11.81	22.67
9.83		19.33	38.03	74.82	17.20	33.53	65.38

Therefore based on table 3, the value of **a₀**, **a₁** and **a₂** for equation 6.1 can be determined using following equation:

$$\begin{bmatrix} 5 & 9.83 & 19.33 \\ 9.83 & 19.33 & 38.03 \\ 19.33 & 38.03 & 74.82 \end{bmatrix} \begin{bmatrix} a_0 \\ a_1 \\ a_2 \end{bmatrix} = \begin{bmatrix} 17.20 \\ 33.53 \\ 65.38 \end{bmatrix}$$

Based on following equation, **a₀**, **a₁** and **a₂** can be determined **a₀** = -970.87, **a₁** = 3741.6 and **a₃** = -3598.8

For equation 6.2

$$\begin{bmatrix} n & \sum x & \sum x^2 \\ \sum x & \sum x^2 & \sum x^3 \\ \sum x^2 & \sum x^3 & \sum x^4 \end{bmatrix} \begin{bmatrix} a_0 \\ a_1 \\ a_2 \end{bmatrix} = \begin{bmatrix} \sum x^0 y \\ \sum x^1 y \\ \sum x^2 y \end{bmatrix}$$

Table 4. Value of matrix from equation 6.2

X	Y	X²	X³	X⁴	X⁰*Y	X¹*Y	X²*Y
1.51	1	2.28	3.44	5.20	1.00	1.51	2.28
1.48	2.2	2.19	3.24	4.80	2.20	3.26	4.82
1.47	3.55	2.16	3.18	4.67	3.55	5.22	7.67
1.44	4.3	2.07	2.99	4.30	4.30	6.19	8.92
1.41	6.15	1.99	2.80	3.95	6.15	8.67	12.23
7.31		10.69	15.65	22.92	17.20	24.85	35.91

Therefore based on Table 4, the value of $\mathbf{a_0}$, $\mathbf{a_1}$ and $\mathbf{a_2}$ for equation 6.2 can be determined using following equation:

$$\begin{bmatrix} 5 & 7.31 & 10.69 \\ 7.31 & 10.69 & 15.65 \\ 10.69 & 15.65 & 22.92 \end{bmatrix} \begin{bmatrix} a_0 \\ a_1 \\ a_2 \end{bmatrix} = \begin{bmatrix} 17.20 \\ 24.85 \\ 35.91 \end{bmatrix}$$

Based on following equation, $\mathbf{a_0}$, $\mathbf{a_1}$ and $\mathbf{a_2}$ can be determined $\mathbf{a_0} = -37.021$, $\mathbf{a_1} = 72.342$ and $\mathbf{a_3} = -22.729$.