Tunable Defected Ground Structure and Its Applications to Simultaneous Reconfigurable Communication and Partial Discharge Detection

by

Leong Chon Chio

Master of Science in Electrical and Electronics Engineering

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Faculty of Science and Technology University of Macau
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Co-Supervisor

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ABSTRACT

This thesis work is about the research of reconfigurable microstrip antenna for the partial discharge (PD) detection application. Detection of PD via UHF emission characterization at 500 MHz is an effective diagnostic tool for monitoring high voltage systems so as to predict the failure. Therefore, dual-band antenna with tunable 2nd operating band for simultaneous PD’s UHF emission detection and multi-band communications is needed. To accomplish this objective, frequency tuning techniques of antennas are reviewed with the emphasis on the use of defected grounded structure (DGS) and active elements on microstrip monopole antenna. An active dual-band microstrip monopole antenna is presented using varactor loaded DGS with Islands (DGSI) on feedline. This antenna’s 2nd operating frequency is tuned from 2.156 GHz to 2.78 GHz when varactor is biased with 5 V to 29 V DC respectively. Moreover, 10-dB matching bandwidth of 62 MHz (28.6%) and 64 MHz (23.0%) are recorded under the above bias conditions. For the sake of completeness of tuning element study, this thesis investigates on DGS stub-loaded structure which offers flexibility in frequency tuning while overall good impedance matching performance is kept. To this end, a novel microstrip monopole antenna using above DGS stub-loaded structure is implemented. Frequency tuning is achieved through the uses of MEMS switches so as to control the slot length of DGS. The prototype MEMS switched antenna reports 27% frequency tuning range from 0.91 GHz to 1.19 GHz for the 2nd operating frequency while the 1st operating frequency is kept unchanged at 0.49 GHz and -18.5 dB matching. Moreover, the bandwidth of 2nd operating band is changed from 382 MHz to 120 MHz for the same MEMS switches control status. The antenna’s measurement results agree with the simulations and fulfill the needs of the prototype PD detection.
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# Abstract ................................................................................................................................. ii
# Acknowledgement ................................................................................................................. iii
# Table of Contents ................................................................................................................... iv
# List of Figures........................................................................................................................... vii
# List of Tables............................................................................................................................ xii
# List of Abbreviations................................................................................................................. xiii

## Chapter 1 Introduction ............................................................................................................ 1

1.1 UHF Emission in Partial Discharge ..................................................................................... 3
1.2 Reconfigurable Antenna for Partial Discharge Detection ................................................. 5
1.3 Thesis Organization ............................................................................................................. 6
1.4 Thesis Significance and Originalities .................................................................................. 6
1.5 References ........................................................................................................................... 8

## Chapter 2 Microstrip Antenna Using Defected Ground Structure ...................................... 9

2.1 Basics of the Microstrip Antenna ....................................................................................... 9
2.1.1 Feeding Techniques for Microstrip Antenna .................................................................. 10
2.1.2 The Microstrip Patch Antenna ..................................................................................... 11
2.1.3 The Microstrip Monopole Antenna ............................................................................. 13
2.1.4 1 GHz Microstrip-Fed Square Monopole Design ......................................................... 14
2.2 Basics of Defected Ground Structure ............................................................................. 16
2.2.1 Slotline ......................................................................................................................... 16
2.2.2 Electromagnetic Band Gap (EBG) Structure ................................................................. 18
2.2.3 Defected Ground Structure ......................................................................................... 19
2.2.4 Equivalent Circuit Model of Defected Ground Structure Unit .................................... 20
2.3 Applications of the Defected Ground Structure ............................................................... 21
2.3.1 Spurious Suppressed Microstrip Dual-Mode Bandpass Filter Using Double U-Shaped DGS ................................................................................................................. 22
2.3.2 Application of Defected Ground Structure to Bandwidth Enhancement of Monopole Antenna ......................................................................................................................... 24
2.4 Summary ............................................................................................................................ 28
2.5 References ........................................................................................................................... 29
Chapter 3 Tunable Microstrip Monopole Antenna .................................................. 31
  3.1 Methodology of Frequency Tuning Technique 33
  3.1.1 Mechanically Tuning 33
  3.1.2 Electronically Tuning 36
    a) Varactor tuned antennas
    b) PIN diode tuned antennas
    c) MEMS Switch tuned antennas
  3.2 Parametric Analysis of Microstrip Monopole Antenna 45
    3.2.1 Analysis of the Radiation Patch 46
    3.2.2 Analysis of the Finite Ground Plane 48
    3.2.3 Analysis of the Antenna Feedline 50
  3.3 Reconfigurable Element Using Defected Ground Structure 51
    3.3.1 Length Variation of U-Shaped DGSI 54
    3.3.2 Width Variation of U-Shaped DGSI 54
    3.3.3 Length Variation of Islands 55
    3.3.4 Width Variation of Islands 56
    3.3.5 Tunable Bandstop Element Using Capacitive Loaded Double U-Shaped DGSI 56
  3.4 A Tunable Monopole Antenna Using Double U-Shaped Defected Ground Structure with Island 58
  3.5 Summary 61
  3.6 References 62

Chapter 4 Tunable Dual-Band Antenna Using DGS Stub Loaded Structure ............... 65
  4.1 Characteristic of Stub Loaded Structure 65
  4.2 DGS Stub Loaded Structure 70
  4.3 DGS Tunable Dual Band Monopole Antenna 73
    4.3.1 Simulations of Tunable DGS Stub-Loaded Antenna with Different DGS Slot Lengths 74
    4.3.2 Measurement Results of Tunable DGS Stub-Loaded Antenna 78
  4.4 Summary 82
  4.5 References 84

Chapter 5 Conclusions and Future Works ......................................................... 85
  5.1 Summary of the Thesis 85
  5.2 Future Works 86
  5.3 References 88
Appendix A Rogers RO4003 ................................................................. 89

Appendix B Infineon BB833 Varactor .............................................. 94

Appendix C Radant Rmsw101 SPST RF-Mems Switch .................. 97
## LIST OF FIGURES

| Figure 1.1 | RF emission measurement of PD for power transformer | 3 |
| Figure 1.2 | (a) Spark generator; (b) Measurement Setup | 4 |
| Figure 1.3 | PD emissions in UHF band. | 4 |
| Figure 1.4 | Measured $|S_{11}|$ of microstrip dual band monopole antenna. | 6 |
| Figure 2.1 | The configuration of microstrip patch antenna. | 10 |
| Figure 2.2 | Rectangular microstrip antenna fed by microstrip line. | 11 |
| Figure 2.3 | Basic shape of the microstrip patch antenna. | 12 |
| Figure 2.4 | The basic configuration of the monopole antenna. | 13 |
| Figure 2.5 | (a) A cylindrical shape used for approximation, (b) A basic microstrip monopole antenna. | 13 |
| Figure 2.6 | The microstrip square monopole antenna. | 14 |
| Figure 2.7 | The simulation result of proposed monopole antenna. | 15 |
| Figure 2.8 | The radiation pattern of monopole antenna. | 16 |
| Figure 2.9 | Slotline configuration. | 17 |
| Figure 2.10 | Coupled microstrip slotline configuration. | 17 |
| Figure 2.11 | The configuration of the Dumbbell Defected Ground Structure. | 19 |
| Figure 2.12 | The simulation results of an elementary DGS ($-|S_{11}| \cdots |S_{21}|$). | 20 |
| Figure 2.13 | The model of the dumbbell DGS unit. | 20 |
| Figure 2.14 | Proposed double U-shaped DGS unit. | 21 |
| Figure 2.15 | Dual bandgap characteristics of the double U-shaped DGS input feedline. | 22 |
| Figure 2.16 | Photograph of the proposed filter prototype. | 23 |
| Figure 2.17 | The simulation result of S-parameters of the proposed bandpass filter with/without DGS ($\cdots$ with DGS $\cdots$ conventional). | 23 |
Figure 2.18  The measurement of S-parameters of the proposed bandpass filter (\( |S_{11}| - |S_{21}| \)).  

Figure 2.19  Geometry of microstrip monopole antenna with U-shaped DGS.  

Figure 2.20  A single U-shaped DGS.  

Figure 2.21  The return loss of microstrip monopole antenna with a single U-shaped DGS.  

Figure 2.22  Effect of the parameter \( W_{\text{DGS2}} \) in impedance bandwidth 
\((-W_{\text{DGS2}} = 30 \text{ mm, } -W_{\text{DGS2}} = 28 \text{ mm, } -W_{\text{DGS2}} = 25 \text{ mm}).\)  

Figure 2.23  Effect of the parameter \( W_{\text{Gap2}} \) in impedance bandwidth (-\( W_{\text{Gap2}} = 3 \text{ mm, } -- W_{\text{Gap2}} = 6 \text{ mm, } ---- W_{\text{Gap2}} = 7 \text{ mm}.\)  

Figure 3.1  A varactor-loaded microstrip antenna.  

Figure 3.2  A PIN-loaded CPW dipole.  

Figure 3.3  A tunable microstrip antenna using piezoelectric actuator tuning system.  

Figure 3.4  Measured return loss of antenna versus frequency over a range of values of parasitic element spacing (-\( \bigtriangleup -2.28 \text{ mm, } ---5.10 \text{ mm, } -- 6.99 \text{ mm, } -- 9.59 \text{ mm, } -- 10.22 \text{ mm}.\)  

Figure 3.5  Measured radiation patterns over a range of parasitic element (a) E-plane and (b) H-plane (-\( \bigtriangleup -2.28 \text{ mm, } ---5.10 \text{ mm, } -- 6.99 \text{ mm, } -- 9.59 \text{ mm, } -- 10.22 \text{ mm}.\)  

Figure 3.6  Steerable leaky-wave antenna, (a) horizontally polarized antenna couples energy into leaky modes on the tunable impedance surface and (b) the mechanically tunable impedance surface consists of two printed circuit boards: a high-impedance ground plane, and a separate tuning layer.  

Figure 3.7  A mechanically tuned \( \lambda/2 \) quadrifilar helix antenna (QHA).  

Figure 3.8  Practical model of varactor.  

Figure 3.9  A varactor loaded CPW antenna.  

Figure 3.10  Tuning performance of the varactor loaded CPW antenna.  

Figure 3.11  Tuning Helical antenna using varactor.  

Figure 3.12  Tuning performance of Helical antenna using varactor with 50 mm diameter ground plane.  

Figure 3.13  Quasi-Yagi dipole antenna with length adjustable director.  

Figure 3.14  A steerable antenna (a) varactor loaded artificial magnetic conductor (AMC) surface and (b) electrically tunable antenna using varactor loaded AMC surface.  

Figure 3.15  Reconfigurable planar inverted-F antenna (a) 3D view and (b) zoom of radiator.
Figure 3.16  Tuning performances of planar inverted-F antenna (a) PIN diode off and (b) PIN diode on. 41
Figure 3.17  Patch antenna with PIN-diode tuned shorting posts. 42
Figure 3.18  Tuning performance of PIN-diode tuned slotted rectangular patch antenna. 42
Figure 3.19  Photo of MEMS switch. 43
Figure 3.20  Construction of RF MEMS switch (a) capacitive fixed-fixed beam type and (b) ohmic cantilever type. 43
Figure 3.21  Switch operation modes of MEMS switch. 44
Figure 3.22  A MEMS switches reconfigurable scan-beam single-arm spiral antenna. 44
Figure 3.23  Measured maximum beam direction and half-power beam width for different equivalent spiral arm. 45
Figure 3.24  Microstrip monopole antenna (a) Top side (b) Bottom side. 46
Figure 3.25  Simulation results of microstrip monopole antenna with different $L$ (- 30 mm, ·· 35 mm, ··· 40 mm and ---- 45 mm). 47
Figure 3.26  Operating frequencies vs. length of the radiation patch (- fundamental frequency and ··· 2nd harmonic frequency). 48
Figure 3.27  Simulation results of frequency with different width $W_g$ of the ground plane (- 38 mm, · 43 mm and ··· 48 mm). 49
Figure 3.28  Simulation results of frequency with different length $L_g$ of the ground plane (- 31 mm, ·· 36 mm and ··· 41 mm). 49
Figure 3.29  The simulation results of the operating frequency against ground plane dimension (- variation of $W_g$ and ··· variation of $L_g$). 50
Figure 3.30  The simulation results of the operating frequency against ground plane dimension (- variation of $W_g$ and ··· variation of $L_g$). 50
Figure 3.31  Simulation results of frequency with different length $L_f$ of the feedline (- 58 mm, ··· 56 mm, ··· 54 mm). 51
Figure 3.32  Operating frequencies vs. length of the feedline (- fundamental frequency and ··· 2nd harmonic frequency). 51
Figure 3.33  DGSI with double equilateral U-shaped defected ground pattern (a) Top side and (b) Bottom side. 52
Figure 3.34 Simulated $|S_{21}|$ of proposed DGSI with double equilateral U-shaped defected ground pattern.  

Figure 3.35 Simulated $|S_{21}|$ parameter of double U-shaped DGS unit with $W_1 = W_3 = 2$ mm, $W_2 = 2$ mm and different lengths ($- L_1 = 12$ mm & $L_2 = 8$ mm, $-- L_1 = 11$ mm & $L_2 = 8$ mm, $-- L_1 = 12$ mm & $L_2 = 7$ mm).  

Figure 3.36 Simulated $|S_{21}|$ parameter of double U-shaped DGS unit with $L_1 = 12$ mm, $L_2 = 8$ mm and different widths ($- W_1 = W_2 = W_3 = 2$ mm, $-- W_1 = W_3 = 3$ mm & $W_2 = 2$ mm, $-- W_1 = W_3 = 2$ mm & $W_2 = 3$ mm).  

Figure 3.37 Simulated $|S_{21}|$ parameter of island microstrip with $W = 3.5$ mm and different lengths ($- 18$ mm, $-- 20$ mm and $-- 22$ mm).  

Figure 3.38 Simulated $|S_{21}|$ parameter of island microstrips with $L = 20$ mm and different widths ($- 3$ mm, $-- 3.5$ mm and $-- 4$ mm).  

Figure 3.39 Proposed tunable DGSI bandstop element.  

Figure 3.40 Measured $|S_{21}|$ of the proposed tunable bandstop element using capacitive loaded DGSI with different bias voltage ($- 10$ V, $-- 5$ V, $--- 2$ V, $----- 0$ V).  

Figure 3.41 Circuit of the proposed tunable DGSI bandstop element (a) Top view and (b) Bottom view.  

Figure 3.42 Proposed monopole antenna with double U-shaped DGSI (a) Top and (b) Bottom.  

Figure 3.43 Photo of the prototype tunable monopole antenna.  

Figure 3.44 Measured $|S_{11}|$ of the proposed tunable DGSI monopole antenna with different bias voltage ($- 5$ V, $-- 10$ V, $--- 16$ V and $----- 29$ V).  

Figure 4.1 A stub-load resonator.  

Figure 4.2 The equivalent circuit in odd-mode of the SLR.  

Figure 4.3 Even-mode equivalent circuit.  

Figure 4.4 Simulated $|S_{21}|$ of SLS by varying the length $L_2$ ($- 1$ mm, $-- 5$ mm, $--- 10$ mm and $----- 15$ mm).  

Figure 4.5 Even-mode equivalent circuit of the SSLS.  

Figure 4.6 The geometric figure of DGS stub loaded structure.  

Figure 4.7 A DGS stub loaded structure.  

Figure 4.8 Simulated $|S_{21}|$ of DGS stub-loaded structure by varying DGS slot length $L$ ($- 10$ mm, $--- 20$ mm, $--- 30$ mm and $----- 45$ mm).
Figure 4.9  Proposed tunable dual band monopole antenna with U-shaped DGS stud load structure (a) Top view and (b) Bottom view.

Figure 4.10  The length of the DGS varied by short thin line.

Figure 4.11  Simulated matching frequency of reconfigurable antenna using thin line for different $L_{\text{DGS}}$ (--- 10 mm, --- 30 mm and – 45 mm).

Figure 4.12  The radiation pattern for $L_{\text{DGS}}$ = 10 mm (a) fundamental frequency, 500 MHz and (b) even-mode frequency, 1.19 GHz.

Figure 4.13  The radiation pattern for $L_{\text{DGS}}$ = 45 mm (a) fundamental frequency, 500 MHz and (b) even-mode frequency, 0.91 GHz.

Figure 4.14  Prototype tunable dual band DGS stub-loaded monopole antenna using MEMS switches. (a) Top view and (b) Bottom view.

Figure 4.15  The photo of the bond wired MENS switch.

Figure 4.16  Comparisons of simulated and measured $|S_{11}|$ of tunable DGS slot-loaded monopole antenna by varying DGS slot length $L_{\text{DGS}}$. (a) $L_{\text{DGS}}$ = 10 mm, (b) $L_{\text{DGS}}$ = 30 mm and (c) $L_{\text{DGS}}$ = 45 mm. (--- simulation and – measurement)

Figure 4.17  Comparisons of measured $|S_{11}|$ of tunable DGS slot-loaded monopole antenna by varying DGS slot length $L_{\text{DGS}}$. (--- 10 mm, --- 30 mm and – 45 mm).
LIST OF TABLES

Table 2.1 The comparisons of different patch antennas 12
Table 3.1 PIN diodes on-off combination for antenna in Figure 3.17 42
Table 3.2 Dimensions of conventional microstrip monopole square antenna (mm) 46
Table 3.3 Dimensions of the proposed tunable monopole antenna (mm) 59
Table 4.1 The dimensions of the DGS stub loaded structure (mm) 72
Table 4.2 Dimensions of the proposed tunable dual band monopole antenna (mm) 75
Table 4.3 MEMS switches status and its corresponding DGS slot length 79
Table 4.4 Comparisons of the simulation and measurement results for fundamental frequency of antenna with different $L_{DGS}$ 81
Table 4.5 Comparisons of the simulation and measurement results for even-mode frequency of antenna with different $L_{DGS}$ 81
<table>
<thead>
<tr>
<th>Acronym</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>AMC</td>
<td>Artificial magnetic conductor</td>
</tr>
<tr>
<td>BW</td>
<td>Bandwidth</td>
</tr>
<tr>
<td>CPW</td>
<td>Coplanar waveguide</td>
</tr>
<tr>
<td>DGS</td>
<td>Defected ground structure</td>
</tr>
<tr>
<td>DGSI</td>
<td>Defected ground structure with islands</td>
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<tr>
<td>EBG</td>
<td>Electromagnetic bandgap structure</td>
</tr>
<tr>
<td>OSLC</td>
<td>Open Stub loaded structure</td>
</tr>
<tr>
<td>PD</td>
<td>Partial Discharge</td>
</tr>
<tr>
<td>PBG</td>
<td>Phononic bandgap structure</td>
</tr>
<tr>
<td>PIFAs</td>
<td>Planar inverted F-type antenna</td>
</tr>
<tr>
<td>QHA</td>
<td>Quadrifilar helix antenna</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency</td>
</tr>
<tr>
<td>RMSA</td>
<td>Rectangular microstrip antenna</td>
</tr>
<tr>
<td>SLR</td>
<td>Stub loaded resonator</td>
</tr>
<tr>
<td>SLC</td>
<td>Stub loaded structure</td>
</tr>
<tr>
<td>SSLC</td>
<td>Single stub loaded structure</td>
</tr>
</tbody>
</table>
Electricity is the most important energy source in the modern society. It penetrates deeply into the human’s daily life and most of the activities cannot be carried out without the supply of the electricity power. Therefore, the power failure in different levels causes serious problems. For example, a failure of electrical power substation, which knocked out power to 630,000 customers around Nagasaki Prefecture and some of Saga Prefecture, include Nagasaki, Sasebo, Isahaya and Karatsu in Japan in April 1998. In 2005, due to a fault of the main transmission cable, Malaysia electricity blackout crisis happened where many states of Malaysia’s northern peninsular, including Perak, Penang, Kedah, and Perlis had no electricity and great economic lose are caused. Therefore, the reliability of the power facilities is always a major concern.

Traditionally, to maintain the system operates reliable, facilities are required to regular check-up within their mean time between failures (MTBF), however the complexity of the environment affects the effectiveness of maintenance while replacing facilities prior the MTBF results increasing of cost. To this end, efforts are made to monitor the healthy condition of the power facilities and predict the failure before power outage is occured. In
the high power facilities, the devastated failure phenomenon is concealed in facilities indeed, which is called partial discharge (PD). Due to the operation of high voltage, PD will be caused and it is an electric discharge that only partially bridges the insulation between conductors and it is known as one of the major factors to accelerate degradation of electrical insulation. The process of deterioration can propagate and develop, until the insulation is unable to withstand the electrical stress, leading to flashover. The ultimate failure of high voltage (HV) facilities is often sudden and catastrophic, producing major damage and network outages. Therefore, PD activity provides clear evidence that an asset is deteriorating in a way that is likely to be failure. Depending on causes and effects, PD detection is proposed to monitor PD activities in HV facilities such as transformers, on-load tap changers, bushings, switchgear, cables and so forth. It is reported that the sustainable occurrence of the PD will lead the facility in high possibility of failure [1.1]-[1.2]. Therefore, PD detection has been considered as one of the most effective diagnostic tools for insulation monitoring of high voltage systems so as to achieve the failure prediction.

During the occurrence of the PD phenomenon, various physical quantities, e.g. electromagnetic wave (in UHF band) and ultraviolet (UV) and so forth are emitted. Relied on these physical quantities, various methods were proposed to examine/detect the partial discharge phenomenon. Some detections methods measure the voltage drop, current impulse while others sense the emission of UHF, ultrasonic and optics [1.3]-[1.6]. Among these methods, the UHF method offers the non-invasive and real-time detection. This UHF method and its physical characteristic will be discussed in following section [1.1]. Additionally, the phenomenon of PD occurs sporadically, randomly, and unexpectedly. A timely detection is required in order to achieve high reliability monitor and the detected signals are required to be centralized through various communications mean to provide the time-based monitoring.
1.1 **UHF Emission in Partial Discharge**

Electromagnetic wave emission is associated with PD. Therefore, RF emission method is an important approach to detect PD and should be applied not only in GAS-INSULATED substations (GISs) but also in power transformer as shown in Figure 1.1, cable and so forth. To this end, the RF emission technique becomes a welcome tool of PD monitoring, which has been widely applied in worldwide with excellent result both during high voltage on-site commissioning tests and condition monitoring during service [1.7]-[1.8].

![Figure 1.1 RF emission measurement of PD for power transformer.](image)

For the simulation of the PD phenomenon and detection measurement, a HV spark generator as shown in Figure 1.2(a) is used for experiment. It can generate spark and simulate the occurrence of PD under four states: 5 kV, 10 kV, 20 kV and 50 kV. In this experiment, 50 kV state is used and electrodes are adjusted with 1 cm separation. In order to detect the wideband RF emission of the PD, Anritsu MT8222A spectrum analyzer is used with a wideband antenna with frequency range from 0.5 GHz to 7.5 GHz to capture the radio spectrum. The measurement equipments were located with 1.5 m apart the PD generator as shown in Figure 1.2(b). Figure 1.3 shows the captured spectrum from range of 500 MHz to 3 GHz.
In Figure 1.3, it is obvious that high counts of emissions are detected within the frequency band lower than 1 GHz. For example, it is observed that -52 dBm is measured at 500 MHz. However, 800 MHz to 2 GHz frequency band is allocated for various mobile communication services in general and it is difficult to distinguish. Moreover, the emissions between 2 GHz to 3 GHz have lower emission count. Against this background, the lower UHF band is selected for the RF emission PD detection in the prototype PD detection system whilst a tunable upper UHF band is kept for communication.
1.2 RECONFIGURABLE ANTENNA FOR PARTIAL DISCHARGE DETECTION

As introduced, a prototype PD detection system is developed in this work to sense the PD RF emissions at UHF band. With the help of the cognitive wireless network which enable to carry out the communication through different networks dynamically, the flexibility and reliability can be improved [1.9]. To realize such adaptive communication node, tunable or namely reconfigurable RF components are the key components which enable the multi-band/multi-mode transceiver to operate in different frequency bands. For example, as shown in Figure 1.4 by applying a dual-band antenna with tunable frequency band to the PD detection system, PD sensing at 500 MHz can be kept while the tunable higher frequency enables the physical communication for the multi-standards transceiver. Hence the tunable antenna is a key component. To achieve the frequency agility of the antenna, different methodologies were proposed through alteration of antennas’ physical dimension, substrate parameter variation, electronic switches and tunable devices and so forth. Moreover, the bandgap element namely Defected Ground Structure (DGS) had been proposed for the RF/MW components/circuits to elevate the overall system performance recently. These DGS elements are implemented on the ground plan of the circuit board and this benefits the compact circuit implementation indeed. However, the studies of frequency agility of the DGS element are still rare. Therefore, the objectives of this research work are to investigate the frequency agility of the DGS elements and its application on tunable antenna for the PD detection.
1.3 **THESIS ORGANIZATION**

In addition to this introductory chapter, there are four chapters in this thesis. In Chapter 2, the basic microstrip antenna will be reviewed. Then the defected ground structures (DGS) of electromagnetic bandgap structure will be discussed together with the studies of some notable DGS examples. A novel U-shaped DGS is proposed in this chapter and its applications to guided-wave and radiated-wave components will be presented. In Chapter 3, some frequency tuning techniques for antennas will be reviewed and then frequency tuning of the microstrip monopole antenna is studied particularly. A varactor tuned microstrip monopole antenna is presented based on a novel defected ground structures with island (DGSI) and experimentally characterized. In Chapter 4, a new DGS stub loaded structure will be proposed and analyzed. Using the novel DGS structure; a dual-band microstrip monopole antenna with tunable upper band is implemented and measured. Finally, the conclusions will be drawn together with the future perspectives in Chapter 5.

1.4 **THESIS SIGNIFICANCE AND ORIGINALITIES**

In this thesis work, the defected grounded structure (DGS) is studied and summarized to explore its applications to various guided wave and radiated applications. The tuning
techniques for reconfigurable/tunable antennas were studied and summarized. Besides the above basic studies, there are two originalities in this thesis. The first originality presents the uses of varactor loaded defected ground structures with island (DGSI) on feedline of microstrip monopole antenna and achieves the frequency tuning in L-band. The second originality is about the novel DGS stub-loaded structure which offers the frequency agility whilst maintaining impedance matching performance. The analytical formulation of DGS stub-loaded structure is studied in this work. Integrating the proposed DGS structure with the RF MEMS switches, dual-band microstrip antenna with electrically tuned upper band is implemented for the application of partial discharge detection application. The above originalities result the following publications.

1.5 REFERENCES


The first microstrip antenna had been developed by Munson and Howell [2.1]-[2.3] in early of 70’s. Extensive researches have been carried out for exploiting its advantages such as low weight, small volume and ease of fabrication [2.4]-[2.8]. There are many types of the microstrip antennas such as microstrip patch antennas [2.9], microstrip planar inverted-F-type antennas (PIFAs) [2.10], microstrip loop antennas [2.11], microstrip wire antennas [2.12] and microstrip monopole antennas [2.13] and so forth.

2.1 Basics of The Microstrip Antenna

The simplest form of the microstrip antenna consists of a radiating patch on one side of a dielectric substrate and a ground plane on the opposite. The Figure 2.1 shows the top and the side views of a rectangular microstrip antenna (RMSA). This kind of the microstrip antenna is the patch antenna. The radiation patch of the patch antenna is calculated by \( L \times W \) suspended parallel with a ground plane. The substrate with thickness \( h \) is inserted.
between the radiation patch and the ground plane. A cylindrical 50 Ω probe excites the radiation patch through the ground plane. The radiation from the microstrip patch antenna can occur from the fringing fields between the periphery of the patch and the ground plane. The length \( L \) of the rectangular patch for the fundamental TM\(_{10}\) mode excitation is slightly shorter than \( \lambda/2 \), where \( \lambda \) is the wavelength for the substrate \( \varepsilon_r \) used. The fringing fields from the patch to the ground plane are not only restricted in the dielectric, but also exist in the air.

2.1.1 FEEDING TECHNIQUES FOR MICROSTRIP ANTENNA

Feeding technique is an important design parameter of the microstrip antenna which can influence the input impedance and characteristics of the antenna. There are two types of feeding technique used. One is either the common coaxial or probe feed and the other is about the microstrip line feed. The coaxial or probe feed arrangement is shown in Figure 2.1. The conductor wire of the coaxial connector is connected from the ground directly passed to the patch. The advantage of this feeding technique is that it can easily match with its

![Figure 2.1 The configuration of microstrip patch antenna.](image-url)
input impedance by placing the probe at different location on the patch. But, there is a hole drilled in the substrate. The microstrip feed line feed which is shown in Figure 2.2. In this feeding technique, the patch is placed on the top of the substrate and excited by the microstrip line feed. This feeding technique has the advantage that it can be etched on the same substrate to form a completely planar structure. The drawback is the radiation from the feed line. The thickness of the substrate also affects the bandwidth of the microstrip antenna. In the case of the coaxial feed, longer probe length makes the input impedance more inductive, leading to the matching problem.

Figure 2.2 Rectangular microstrip antenna fed by microstrip line.

Except the probe feed and microstrip feed line, there are another feedings available in the microstrip antennas such as aperture coupling [2.16] and proximity coupling [2.17]. Actually the direct feed is the most common method used in modern antenna design. The two common microstrip antennas will be discussed in the following section.

2.1.2 The Microstrip Patch Antenna

Microstrip patch antennas have attracted considerable interest since the 1970s. Nowadays, microstrip patch antennas are used in many different applications because of their low profile, light weight, ease of fabrication, and relatively low cost. It consists of a radiating patch on the top and a ground plane on the back as in Figure 2.1. The operating frequency of the antenna can be controlled by the size of the radiation patch and the shape of the radiation patch. The shape of the radiation can be a square, rectangular, circular,
triangular and annular ring as shown in Figure 2.3.

![Square](image1.png) ![Rectangular](image2.png) ![Circular](image3.png)  
Square  Rectangular  Circular

![Triangular](image4.png)  ![Annular ring](image5.png)  ![Square ring](image6.png)

Triangular  Annular ring  Square ring

Figure 2.3 Basic shape of the microstrip patch antenna.

In the past few years, the above patterns of the radiation patch have been intensively investigated. Table 2.1 shows the comparisons of different shape of the microstrip patch antenna. The rectangular and square microstrip antennas are the simplest and widely used microstrip antennas configuration. It offers the wide impedance bandwidth, but the size is concerned. The size of the circular patch with inherent symmetry is slightly smaller than that of the rectangular and square ones. The disadvantages of this shape are low gain and narrow bandwidth. For triangular patches, the size is smaller compared to the rectangular and circular shapes. The disadvantage is about low gain with higher cross-polarization. Finally, annular ring is the smallest shape.

<table>
<thead>
<tr>
<th>Table 2.1 The comparisons of different patch antennas</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Rectangular &amp; Square</strong></td>
</tr>
<tr>
<td>Advantage</td>
</tr>
<tr>
<td>Disadvantage</td>
</tr>
</tbody>
</table>
2.1.3 The Microstrip Monopole Antenna

The monopole is a well known antenna structure used in mobile communication systems. The monopole is usually fed by a transmission line above a ground plane as a vertical element. The configuration is shown in Figure 2.4. The radiation patch is placed vertically to the finite ground plane. A probe feed is used to feed the radiation patch to the ground plane.

![Diagram of a microstrip monopole antenna](image)

Different sizes of the radiation patch yield different operation frequency. Increasing the size of the ground plane of microstrip antenna, it would affect the input impedance and/or bandwidth.

For a planar monopole antenna, the frequency can be approximately calculated by cylindrical calculation method.

![Diagram of a cylindrical shape](image)
As an equivalent cylindrical monopole antenna of same height $L$ and equivalent radius $r$:

$$2\pi L = WL$$  \hspace{1cm} (2.1)

The length of the monopole can be determined by

$$L = 0.24\lambda F$$  \hspace{1cm} (2.2)

Where

$$F = \frac{L}{r} = \frac{L}{1 + \frac{L}{r}}$$  \hspace{1cm} (2.3)

So, we can obtain the frequency

$$f_L = \frac{c}{\lambda} = \frac{7.2}{L + r}$$  \hspace{1cm} (2.4)

An example of microstrip monopole antenna can be designed and optimized by full-wave electromagnetic solver as in section 2.1.4.

2.1.4 1 GHZ MICROSTRIP-FED SQUARE MONOPOLE DESIGN

The microstrip monopole antenna consists of a square radiating patch ($W_p = L_p$) connected to
the microstrip transmission line with the dimension $L_m$ and $W_m$. On the bottom of the antenna, it has a finite ground plane with width $W_a$ and length $L_g$. As an example, a microstrip monopole antenna with square path operated at 1 GHz using probe fed technique is considered. We choose the material as RO4003 with the dielectric constant $\varepsilon_r = 3.38$. The operating frequency of the antenna can be controlled by the size of the radiation patch, after the calculations according to (2.1) to (2.4), the dimensions of radiation patch are $W_p = 30$ mm and $L_p = 30$ mm. The length $L_m$ and width of the transmission lines $W_m$ are 57.8 mm and 4 mm respectively. The size of the finite ground plane is as compact as 31 mm × 38 mm. Overall the circuit size of this microstrip antenna is 38 mm × 87.8 mm. By Full-wave electromagnetic simulator IE3D, this monopole antenna is simulated and the simulation result is shown in the Figure 2.7 and Figure 2.8. The operating frequency of the monopole antenna is at 1 GHz with 138 MHz as impedance bandwidth.

![Figure 2.7 The simulation result of proposed monopole antenna.](image-url)
2.2 Basics of Defected Ground Structure

Defected Ground Structure (DGS) is a simple microwave structure with a transmission zero or bandgap at finite frequency. Defected Ground Structure is very similar to slotline and photonic band gap structure. The bandgap and slotline structures are very useful in rejecting the unwanted response of guided wave and radiated wave components by using the characteristic of the transmission zero they produce.

2.2.1 Slotline

A Slotline is a planar transmission structure proposed by Cohn in 1968. The basic configuration is a slot etched in the metallization on one side of the substrate; the other side consists of the ground plan as in Figure 2.9. In analysis of the slotline, it is known that the wave propagates along the slot with the major electric field component oriented across the slot in the plane of metallization on the dielectric substrate. The Figure 2.10 shows the configuration of the coupled microstrip slotline. There are two different types of modes propagate in transmission line. Quasi-microstrip (even-mode) and Quasi-slotline (odd-mode) are discussed. In even-mode, the field distribution is similar to that of a microstrip line and

Figure 2.8 The radiation pattern of monopole antenna.
if the field distribution similar to that in a slotline is the odd mode.

![Slotline configuration](image1)

**Figure 2.9** Slotline configuration.

![Coupled microstrip slotline configuration](image2)

**Figure 2.10** Coupled microstrip slotline configuration.

The effective dielectric constants and the characteristic impedances are thus discussed. In even-mode, the effective dielectric constants decrease and the characteristic impedance increases with the wider slot width. Therefore it can be used to realize high impedance of the antenna. The odd mode is less sensitive to the presence of the strip; most of the energy in this mode is confined near the slot.

The couple microstrip slotlines are the same as the bandgap structure, Electromagnetic Band Gap Structure, is discussed in the following section.

### 2.2.2 Electromagnetic Band Gap (EBG) Structure

Slotlines can be included in microstrip circuits by etching them in the ground plane of...
microstrip circuits. A similar structure that etches some periodic patterns in the ground plane was proposed to offer a controllable bandgap at some finite frequency through disturbing the ground plane shield current distribution. This structure is initially developed for a bandgap in the optical frequency band, and thus named phontonic bandgap structure (PBG). In 1998, the PBG structure was first proposed by Itoh et al. that offered a rejection band for electromagnetic transmission, and that is referred as electromagnetic bandgap structure (EBG). This structure uses a microstrip line on the top and some uniform circles periodically etched in the ground plane.

Since the substrate and the photonic bandgap materials are periodic dielectrics, the PBG technology could apply into the microstrip circuit design. The bandgap in photonic bandgap materials represents the forbidden energy range where wave behaving photons cannot be transmitted through the materials. So it can be used to affect and control the movement of electromagnetic waves. The electromagnetic bandgap structures are artificial periodic (or sometimes non-periodic) objects that prevent/assist the propagation of electromagnetic waves in a specified band of frequency for all incident angles and all polarization states. If EBG structure is used, advantage are low profile, light weight and low fabrication cost. The EBG structure consists of a microstrip line on the top of the dielectric and some uniform circles etched in the ground. An etched pattern looks like a cross-shaped slot, to achieve an extension of the bandgap and thus being useful to suppress the harmonic passbands of the conventional parallel coupled-line bandpass filters. So the EBG structures using the slotted ground technique are difficult to use in the design of microwave components due to the lack of accurate modeling. To simplify the slotted pattern of the ground, a new structure was proposed in 2000 by Kim et al., named defected ground structure (DGS).

2.2.3 DEFECTED GROUND STRUCTURE
The Defected Ground Structure is one of the electromagnetic bandgap structures. The basic structure of the DGS has a microstrip line on the top and a series a dumbbell-shaped pattern periodically etched in the ground plane. As shown in Figure 2.11, the elementary dumbbell-shaped DGS unit has two square slots connected with a thinner slotline in the middle. The frequency response of such elementary unit exhibits a bandgap characteristic imposed by a transmission zero that can be analyzed and modeled by a parallel LC equivalent circuit, where the inductance \( L \) and Capacitance \( C \) are controlled by the dimension of the square slots and the middle slotline of the dumbbell pattern respectively. An elementary dumbbell-shaped DGS example with dimensions as \( L_{D1} = L_{D2} = W_{D2} = 28 \) mm, \( W_{D1} = 2 \) mm and \( W_m = 3.452 \) mm is simulated, and the simulation result is shown in Figure 2.12. It is observed that this unit provides a bandgap at 900 MHz with 420 MHz 10-dB rejection bandwidth.
2.2.4 Equivalent Circuit Model of Defected Ground Structure Unit

The model of the dumbbell DGS unit is based on a parallel inductive and capacitive (LC) circuit, as shown in Figure 2.13.

At the resonance frequency,

\[ \omega_z = \frac{1}{\sqrt{LC}} \quad (2.5) \]

As observed in Figure 2.12 a transmission zero is yielded. So the frequency of the transmission zero \( f_z \) and its 3-dB rejection bandwidth \( \Delta f_z \) are expressed, related to the inductance and capacitance values

\[ C = \frac{1}{4\pi Z_0 \Delta f_z} \quad (2.6) \]
\[ L = \frac{1}{(2\pi f_z)^2 C} \]  

(2.7)

where \( Z_0 \) is the characteristic impedance.

In this model, the rejection bandwidth is dependent on the capacitance \( C \) only, whilst the transmission zero is related to both the \( L \) and \( C \) values. Applications of these DGS elements are then discussed in the following section.

2.3 Applications of The Defected Ground Structure

Different from the traditional dumbbell shaped DGS with single transmission zero, we proposed Double U-Shaped Defected Ground Structure. In this structure, a double equilateral U-shaped DGS unit has been studied to offer a dual-zero characteristic (namely lower and upper transmission zeros). This DGS contains a microstrip line on the top and two equilateral U-shaped slotlines etched in the ground plane with the smaller ones embedded inside the larger ones like the Figure 2.14. By adjusting the DGS geometrical parameters, the two transmission zeros can be easily controlled. It was found that the lower transmission zero is resulted from the longer U-shaped slot whereas the upper zero is due to the shorter one. This implies that the independence on the control of these zeros. In addition, both zeros will be distanced to lower frequencies when the end sections of the U-shaped slots have larger widths. Or, they will be relocated to higher frequencies together if the widths of the slotline of middle sections are increased.

![Figure 2.14 Proposed double U-shaped DGS unit.](image)
The above DGS unit is generalized, as shown in Figure 2.14 and Figure 2.15 by ignoring the constraints of equilateral in each U-shaped slot and the open-end alignment of the two U-shaped slots. Each U-shaped slot is symmetrically etched under the microstrip line. Based on the similar tuning properties mentioned above, the transmission zeros of the generalized double U-shaped DGS unit can be tuned for different frequencies. The lower zero frequency is inversely proportional to the length of the larger U-shaped slot ($L_A = 2L_{A1} + L_{A2}$) whilst the upper zero frequency can be controlled by the length ($L_B = 2L_{B1} + L_{B2}$). The width parameters $W_{A1}$, $W_{A2}$, $W_{B1}$ and $W_{B2}$ can relocate both the zeros together to lower and upper frequencies regime.

2.3.1 SPURIOUS SUPPRESSED MICROSTRIP DUAL-MODE BANDPASS FILTER USING DOUBLE U-SHAPED DGS

The above double U-shaped DGS can be used to achieve quasi-elliptic bandpass response with wide stopband. A microstrip dual-mode bandpass filter with multiple-spurious-suppression is presented [2.18]. The input and output feedlines of this
bandpass filter is etched with a compact double U-Shaped defected ground structure (DGS) which offers dual transmission zeros. By locating these transmission zeros properly, the intrinsic harmonic components can be significantly suppressed. An example dual-mode bandpass filter as in Figure 2.16 centered at 2 GHz and 4.4% fractional bandwidth on Rogers RO4003 substrate with the above input and output DGS feedlines is studied. The simulation result is shown in Figure 2.17. The spurious responses of the dual mode bandpass filter are suppressed with overall 20-dB attenuation, whilst the fundamental passband is kept. The bandpass response is located at 2.04 GHz. The two transmission zeros near the band edges of the resultant filter zeros are placed at 1.84 GHz and 2.29 GHz. There are a little different from their original frequencies at 1.86 GHz and 2.26 GHz.

The measurement result has been recorded in Figure 2.18. Obviously, there is very good
agreement between the simulation (Figure 2.17) and the measurement (Figure 2.18) results.
This filter prototype shows a passband at 2.04 GHz with two transmission zeros at 1.79 GHz
and 2.31 GHz. The passband has an insertion loss of about 1.7 dB and 30 dB matching.
Spurious responses are suppressed with more than 20-dB, resulting in a wide stopband from
2.5 GHz up to 9 GHz.

2.3.2 APPLICATION OF DEFECTED GROUND STRUCTURE TO BANDWIDTH
ENHANCEMENT OF MONOPOLE ANTENNA

The usage of the simple microstrip monopole antenna in modern mobile communications is
preferable because of its compactness. But, it suffers from narrow impedance bandwidth.
The impedance bandwidth improvement is realized by not only different patch shape but
also the matching network incorporated with the microstrip feedline. According to the open
literatures, the patterns of the monopole were in the form of triangular- and disc-shaped so
as to improve impedance bandwidth. A simple microstrip stub served as the impedance
matching element and provided around 13% bandwidth enhancement when compared with
the traditional design. For much wider bandwidth implementation, additional stub could be
added but the size is a lingering doubt.

Recently, a new impedance bandwidth improvement is proposed by DGS. A single stub feedline etched with double U-shaped DGS was employed to enhance the impedance bandwidth of a microstrip-fed monopole antenna. Because of the higher mode suppression by the bandgap characteristics of DGS, etching this DGS underneath the simple microstrip feedline, impedance bandwidth broadening can be obtained. A prototype microstrip monopole antenna operated at $L$-band was demonstrated with good 10-dB impedance bandwidth improvement of 112.4% when compared with that of traditional monopole antenna.

A simple DGS monopole as shown in Figure 2.19 is proposed by a single U-shaped DGS [2.19]. Different from the stub feedline, a simple microstrip feedline can be applied. The microstrip feedline monopole antenna printed on a grounded substrate and it has a simple square patch. Followed the basic monopole antenna design formulation, a basic monopole antenna is with length $L_a = 88$ mm and with width $W_a = 38$ mm is designed. Under this microstrip feedline, a U-shaped DGS is etched on a finite ground plane. The U-shaped slotted pattern is shown in Figure 2.20 and it has the dimensions as the width $W_{gap} = 2$ mm,
\( W_{\text{DGS2}} = 30 \text{ mm} \) and also with the height of U-slot \( L_{\text{DGS}} = 20 \text{ mm} \) and \( W_{\text{DGS1}} = 34 \text{ mm} \). In the present study, the above monopole has been etched on a Roger substrate (RO4003) with a dielectric constant of 3.38, and a thickness of 1.524 mm.

By full wave electromagnetic solver, the above DGS monopole antenna has been simulated and its 10-dB impedance bandwidth is studied in Figure 2.21, a wide bandwidth from 2.6 GHz to 3.9 GHz is exhibited.

![Figure 2.20 A single U-shaped DGS.](image)

The impedance bandwidth of the designed antenna can be adjusted by tuning the length of the U-Shaped DGS. As shown in Figure 2.20, the bandwidth characteristic is investigated for
a change of the DGS width parameter $W_{DGS2}$ from 30 mm to 25 mm. When $W_{DGS2}$ is decreased, the impedance bandwidth is deteriorated from 1.3 GHz to 0.4 GHz. It is reported that the bandwidth is improved from 1.3 GHz to 0.9 GHz for 1 mm and 2 mm change. In Figure 2.21, the incremental change in length $W_{gap2}$ is tuned from 3 mm to 7 mm, the bandwidth is varied inversely. It decreases from 1.35 GHz to 1.2 GHz.

In this work, a simple monopole antenna with a single U-shaped DGS is proposed. By this design, the prototype antenna obtains the 10-dB return loss from 2.6 GHz to 3.9 GHz that is
much wider than that of traditional design.

2.4 SUMMARY

In this chapter, the basic microstrip antenna is reviewed. To improve the performance of these microstrip antennas, the electromagnetic bandgap structure is discussed with emphasis on the defected ground structure. Some notable DGS examples were reviewed and a U-shaped DGS and its applications to guided wave and radiated wave components were presented.
2.5 REFERENCES


Modern communication systems operated at different standards are greatly benefited from the use of the tunable antenna because of its re-configurability of operating frequency from 10% to 60% in general. In example [3.1], a notable microstrip antenna with frequency agility was designed at 2 GHz. As demonstrated, the operating frequency was varied from 850 MHz to 2.4 GHz. Such tunable antenna finds applications in compact transceiver design of GSM850/GSM900/GPS/DCS/PCS/UMTS/WLAN. To determine the performance of a frequency agile antenna, the frequency tuning range, the type of frequency tuning (i.e. discrete or continuous) and radiation pattern are concerned.

In past ten years, there were some frequency agile techniques and these antennas were tuned by the alteration of physical dimension, substrate parameter variation, electronic switches and tunable devices and so forth. Changing the physical dimensions of the antenna by mechanical actuators, piezoelectric actuators and voltage-controlled membranes, continuous frequency tuning and low losses were realized [3.2]. This technique suffers from not only the slow response time and high actuation voltage (>100 V) but also narrow tuning range (<10%). Varying substrate parameters by bulky ferroelectrics, ferrites and liquid crystals is one of a continuous frequency tunings which offers the wide tuning range (>100%)
and excess losses under the high bias voltage. Another continuous frequency tuning method is provided by using the active element, such as varactor. These active circuits have reached the tuning range up to 60% by applying the low-to-medium bias voltage (<30 V). The discrete frequency tuning is performed by adding the active elements PIN diodes and MEMS switches in the microwave circuit which provided a wide tuning range and low losses. Therefore, the continuous frequency tuning and discrete frequency tuning are commonly used techniques in frequency agility microstrip antenna. For example, a latest design consisting of the stacked patch configurations is shown in Figure 3.1, where four varactors are added to implement along the non-radiating edges of the patch [3.3]. This type of loading changes the propagation characteristics and slows down the phase velocity along the non-radiating edges. Therefore, the mechanism is similar to the loaded line phase shifter. From the results, a 63% tuning range was achieved when the bias voltage was varied between 0 - 20 V. The radiation efficiency varies between about 80% - 90% over the tuning range.

![Figure 3.1 A varactor-loaded microstrip antenna [3.3].](image)

Besides the continuous frequency tuning with varactors, PIN diodes were used to adjust the size of the radiators for discrete frequency tuning. As shown in Figure 3.2, 8 PIN diodes
which were placed across the arms of a coplanar waveguide (CPW) dipole, thus the effective length of the dipole can be controlled by the different combinations of diodes switched on or off state. The antenna operating frequency was varied from 2.96 GHz to 7.96 GHz with an average bandwidth 15.2% [3.4].

![PIN-loaded CPW dipole](image)

**Figure 3.2 A PIN-loaded CPW dipole [3.4].**

### 3.1 Methodology of Frequency Tuning Technique

To control the operation frequency of the antenna, there are various tuning methods and these are relied on the control of the antennas’ structure likes radiator, ground plane, directive/reflective elements structure and dielectric materials. The techniques based on mechanical actuator and electrical tuning elements are widely used. These typical tuning techniques will be reviewed in following section.

#### 3.1.1 Mechanically Tuning

As reported in [3.5], a piezoelectric actuator was applied to a microstrip antenna and the mechanical displacement of the parasitic director was controlled. With this adjustable parasitic director, the operating frequency, bandwidth, and gain of antenna can be changed according to the spacing between the driven and parasitic elements. The photo of this example antenna is depicted in Figure 3.3.
By adjusting the space between the driven and parasitic element from 2.28 mm to 10.22 mm, the above antenna has a tuning operating frequency from 2.79 GHz to 2.92 GHz. Moreover, the beamwidth in the E-plane is decreased simultaneously while characteristic of the H-plane is kept. The measured results are shown in Figure 3.4 - Figure 3.5 respectively.
Different from the above example tuning the distance between the antenna driven and director, another antenna example of steerable leaky-wave antennas shown in Figure 3.6 mechanically moves a tuning layer across the stationary high-impedance surface and thus the capacitance between the overlapping plates is varied [3.6]. As a result, the resonance frequency of the surface is tuned. Eventually, the band structure is shifted in frequency, which changes the tangential wave vector of the leaky waves for a fixed frequency.

Finally, the mechanical tuning method is suitable to apply onto the antenna with 3-dimensional structure. For instance, by introducing the tunable gap which varies the axial length and radial gap between the overlapping volutes at the center of the helical sections of a λ/2 quadrifilar helix antenna (QHA) as depicted in Figure 3.7, the antenna yields a 28% impedance bandwidth which improves the bandwidth of a conventional QHA by 9 times. Moreover, a 16% bandwidth with a front to back ratio of 14 dB is achieved while the size of the antenna is reduced by 5-14% [3.7].
3.1.2 **Electronically Tuning**

Electrical tuning is a widely used method besides the above mechanical one. An electronically tuned antenna can vary its operating frequency, bandwidth and/or radiation patterns dynamically through the electrical biased signal. To achieve the above tuning capabilities, active devices such as varactor, PIN diode and MEMS switch were commonly applied to different components of antennas, i.e. radiator, ground structure and feeding network and so forth.

a) **Varactor tuned antennas**

The varactor or so called varactor diode is a type of semiconductor diode which presents the voltage-dependent capacitor properties. It is built upon the voltage dependency of the depletion layer capacitance of the diode pn-junction. The width of the diode depletion layer can be controlled by reverse biased voltage at levels without current pass through. Therefore, the depletion layer acts as a capacitor and the diode behaves as a variable capacitor against the bias voltage. A practical varactor equivalent circuit model is shown in Figure 3.8, besides the variable capacitance $C_J$, the diode series resistance $R_S$ and some unavoidable parasitic capacitance and inductance $C_P$ and $L_P$ are existed.

![Figure 3.7 A mechanically tuned λ/2 quadrifilar helix antenna (QHA) [3.7].](image-url)
Based on this variable capacitance property, the varactor is widely used in the microwave guided wave and radiated wave circuits for frequency tuning [3.8]-[3.9]. As shown in Figure 3.9, a varactor is capacitively loaded to the radiating edge of a CPW patch antenna, the effective length of the antenna increases and thus decreases the resonant frequency of the antenna [3.10]. The tuning performance of this antenna is shown in Figure 3.10 in which the antenna reports a tuning range from 4.92 GHz to 5.40 GHz while the bias voltage is changed from 0 V to 19.5 V respectively.
In addition, varactor can be used on the feeding element of antenna for frequency tuning. As reported in [3.11], a varactor is series connected to the feed of a Helical antenna for DVB-H application. The antenna structure and its tunable performance are shown in Figure 3.11 and 3.12 respectively. In Figure 3.12, the antenna shows a tuning range from 490 MHz to 600 MHz whilst the varactor is reversely biased from 1 V to 20 V.
Except the radiating and feeding elements, varactor can be applied to the reflecting and
directing elements for the beam steering antennas [3.12]-[3.14]. In Figure 3.13, a varactor with capacitance $C_2$ is loaded on the director of a Quasi-Yagi dipole antenna and a 35% continuously frequency-tuning bandwidth is measured within 1.80 GHz to 2.45 GHz [3.15]. Moreover, with the help of this length-adjustable director, it allows the endfire pattern to maintain a relative high gain over the tuning bandwidth.

![Quasi-Yagi dipole antenna with length adjustable director](image)

Similarly, a varactor loaded high-impedance surface is applied to form a steerable antenna as shown in Figure 3.14. In this antenna, a conventional wideband bow-tie antenna is mounted on top of an active artificial magnetic conductor (AMC) [3.16]. The varactors are placed between the metallic elements and the ground plane. With different bias voltages, the antenna can be tuned over the S-band. Moreover, when applying the suitable biasing voltage to the active elements of the AMC surface based on the leaky radiation principles, beam scanning operation over each working frequency can be yielded. However, the non-linear effect due to the presence of active element will be a concern though varactors are widely used on various tunable antennas.
b) PIN diode tuned antennas

Besides the varactor diode, PIN diode is another active element widely used for the high frequency tuning circuits. Unlike the varactor which obtains the tuning relied on the variable capacitive loading on the circuit, PIN diode presents a high-frequency resistance which is inversely proportional to the DC bias current through the diode. Therefore, PIN diode can acts as a variable resistor with proper bias. For instance, typical PIN diode exhibits capacitance under zero or reverse bias and thus RF signal is blocked. On the other hand, when it is forward biased, a low high-frequency resistance can be yielded and good RF conductor is obtained.

Using this RF switch, tuning/switchable antennas were proposed [3.17]-[3.19]. For example, a reconfigurable planar inverted-F antenna was proposed as shown in Figure 3.15 in which the PIN diode is used to connect between the radiator and an additional radiator [3.20]. When PIN diode is off, the antenna operates in 1.85 GHz - 2.18 GHz and 5.15 GHz - 5.825 GHz for USPCS/WCDMA and WLAN bands respectively. When PIN diode is on, the upper band of the antenna will lower down to 3.4 GHz - 3.6 GHz for WiMAX while the lower band is kept. The tuning performances of the antenna on two modes are shown in Figure 3.16.
Moreover, PIN diode can be used to control the short-circuit post of the patch antenna and thus yield tuning capability. As shown in Figure 3.17, the slotted rectangular patch of patch antenna is loaded by number of posts which are short to the ground plane at the patch edge [3.21]. These shorting posts connectivity are controlled by the PIN diodes and antenna is tuned in sub-bands from 620 MHz to 1150 MHz as illustrated in Figure 3.18. The antenna’s operating frequency and its tuning states based on the PIN diode switches on-off combination are summarized in Table 3.1.
Figure 3.17 Patch antenna with PIN-diode tuned shorting posts [3.21].

Figure 3.18 Tuning performance of PIN-diode tuned slotted rectangular patch antenna [3.21].

Table 3.1 PIN diodes on-off combination for antenna in Figure 3.17 [3.21]

<table>
<thead>
<tr>
<th>State</th>
<th>SW1</th>
<th>SW2</th>
<th>SW3</th>
<th>Operating Frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>On</td>
<td>Off</td>
<td>Off</td>
<td>620 MHz</td>
</tr>
<tr>
<td>2</td>
<td>Off</td>
<td>On</td>
<td>Off</td>
<td>810 MHz</td>
</tr>
<tr>
<td>3</td>
<td>On</td>
<td>On</td>
<td>Off</td>
<td>910 MHz</td>
</tr>
<tr>
<td>4</td>
<td>Off</td>
<td>On</td>
<td>On</td>
<td>1150 MHz</td>
</tr>
</tbody>
</table>

c) **MEMS switch tuned antennas**

Among the tuning elements for electrically tuned antenna, MEMS switch is an alternative. MEMS switch is a switch component which is built based on the Microelectromechanical systems (MEMS) technology. A photo of the MEMS switch is depicted in Figure 3.19.
Different from the operation of varactor and PIN diode, MEMS switch is indeed a technology of very small mechanical devices instead of electronic devices even both are driven by electricity though. The MEMS technology was proposed in the late of 1950s and MEMS devices have size from 20 µm to 1 mm in general. Due to some physical effects such as electrostatics, wetting dominate volume effects and large surface area to volume ratio, etc., the standard constructions in such small size are not useful usually. Till the device could be fabricated using the modified semiconductor fabrication technologies, this made MEMS technology practical and various small mechanical devices were implemented. e.g. electromechanical monolithic resonator was one of the early examples of MEMS devices.

There are two major constructions of RF MEMS switch; i.e. capacitive fixed-fixed beam switch and ohmic cantilever switch as shown in Figure 3.20. The former is relied on a micro-machined capacitor with a moving top electrode while later is capacitive in the up-state, but makes an ohmic contact in the down-state.
Figure 3.21(a), the switch is fabricated in Normal On status which allows the RF signal pass through. When bias voltage is applied, the electrostatic force from the electrode will make the beam shunt to the ground of the substrate and RF signal is off. On the other hand, the ohmic cantilever switch as shown in Figure 3.21(b) behaves as open circuit without bias voltage. When bias voltage is applied, the beam will be forced to contact the other end of the signal line by the electrode and thus a series switch is implemented.

For example, reconfigurable scan-beam antenna using RF-MEMS switches integrated into the single-arm spiral antenna as illustrated in Figure 3.22 was reported in [3.22]. By proper control these MEMS switches, various equivalent spiral antenna arm is yielded and resulted in the control of maximum beam direction as shown in Figure 3.23.
Except applying the MEMS switch on the antenna arm, the MEMS switch was proposed to be used on the antenna ground plane to switch the antenna frequency bands and polarization. As reported in [3.23], the MEMS switches were applied on the slotted ground plan of microstrip antenna. Therefore, frequency bands operation between 1.575 GHz and 1.615 GHz are obtained while RHCP and LHCP radiation polarization are maintained respectively. As reported in above sections, different tuning methods and devices such as varactors, PIN diodes and MEMS switches were reviewed for design of tunable antennas with different pros and cons. In the following sections, a tunable antenna based on the microstrip monopole will be designed and discussed.

### 3.2 Parametric Analysis of Microstrip Monopole Antenna

There are numbers of antenna structure that can be used to implement the tunable antenna. Among these structures, microstrip monopole antenna is easy to design while it inherits the advantages of the planar circuit. Therefore, to design a tunable antenna, a microstrip monopole antenna which has been discussed in Chapter 2 is selected as the basic antenna structure. It is known that the geometry/dimension of the microstrip monopole antenna’s...
radiation patch and the finite ground plane can be used to control its operation frequency. In this section, the parametric analyses of radiation driven, ground plane and feeding element are carried out respectively to investigate its performance on frequency tuning. An example 1 GHz conventional microstrip square monopole antenna is designed on the Rogers RO4003 high frequency laminate as shown in Figure 3.24. Its radiation patch size is $L \times L$ and the finite ground plane has the size of $W_g \times L_g$ respectively while the total antenna size is $L_c \times W_c$. In addition, a 50 $\Omega$ microstrip feedline is used to feed this square patch. Based on the substrate parameters, i.e. relative dielectric constant $\varepsilon_r = 3.38$ and substrate height $h = 1.524$ mm, etc. The antenna’s dimensions are summarized in Table 3.2.

<table>
<thead>
<tr>
<th>$W_g$</th>
<th>$L_g$</th>
<th>$L_c$</th>
<th>$W_c$</th>
<th>$L$</th>
<th>$L_f$</th>
</tr>
</thead>
<tbody>
<tr>
<td>38</td>
<td>31</td>
<td>88</td>
<td>38</td>
<td>30</td>
<td>58</td>
</tr>
</tbody>
</table>

![Figure 3.24 Microstrip monopole antenna (a) Top side (b) Bottom side.](image)

**3.2.1 Analysis of the Radiation Patch**

Based on the dimensions listed in Table 3.2, the antenna is simulated by EM simulator. Its $|S_{11}|$ are shown in Figure 3.25 and its fundamental operating frequency is located in 1 GHz.
By adjusting the length of the patch $L$ from 30 mm to 45 mm, its operation frequency is relocated from 1 GHz to 0.83 GHz. It is obvious that the antenna’s fundamental operating frequency is inversely proportional to the patch size. Besides the fundamental operating frequency, it can be found that the 2\textsuperscript{nd} operating frequency band of the antenna will also be reallocated by varying the patch size. To summarize these variations, the patch length $L$ against the operating frequency is plotted in Figure 3.26. In this figure, it is obvious that when $L$ is increased from 30 mm to 45 mm by 5 mm, the fundamental operating frequency of the antennas is monotonically decreasing from 0.99 GHz to 0.93 GHz. Moreover, the 2\textsuperscript{nd} operating frequency is decreased from 4.64 GHz to 4.34 GHz simultaneously.

Figure 3.25 Simulation results of microstrip monopole antenna with different $L$ (- 30 mm, --- 35 mm, --- 40 mm and ---- 45 mm).
3.2.2 ANALYSIS OF THE FINITE GROUND PLANE

The shape and dimensions of the antenna ground plane will affect the performance of the antenna also. For the microstrip monopole antenna simulated, its size is $88 \times 38$ mm$^2$ ($L_x \times W_x$) while the finite ground plane has size of $L_y = 38$ mm and $W_y = 31$ mm under the feedline. Keeping the radiation patch unchanged with the size of $30 \times 30$ mm$^2$, the simulated results of the operating frequency against the length $L_y$ and the width $W_y$ of the finite ground plane are shown in Figure 3.27 and Figures 3.28 respectively. In Figure 3.27, the operating frequency of the antenna is inversely changed from 0.99 GHz to 0.97 GHz when $W_y$ is increased from 38 mm to 48 mm. In Figure 3.28, the operating frequency of the antenna is elevated from 0.99 GHz to 1.08 GHz when $L_y$ is increased from 31 mm to 41 mm. It is obvious that the operating frequency is much sensitive to the length $L_y$. Similar to the study of the radiation patch; it is found that the change of the ground plane size, the 2nd harmonic resonant frequency will be affected also and its parametric analyses results against the changes of $L_y$ and $W_y$ are summarized in Figure 3.29 and Figure 3.30 respectively.
Figure 3.27 Simulation results of frequency with different width $W_g$ of the ground plane ($\sim 38$ mm, $\sim 43$ mm and $\sim 48$ mm).

Figure 3.28 Simulation results of frequency with different length $L_g$ of the ground plane ($\sim 31$ mm, $\sim 36$ mm and $\sim 41$ mm).
3.2.3 ANALYSIS OF THE ANTENNA FEEDLINE

Typically, the microstrip monopole antenna is fed by a 50 Ω microstrip line as shown in Figure 3.24, the length of this feedline affects the antenna performance indeed and thus it is studied in this section. By changing the length of this feedline, the simulated operating frequency of the antenna is shown in Figure 3.31 below.
3.3 RECONFIGURABLE ELEMENT USING DEFECTED GROUND STRUCTURE

Based on the above parametric analyses, frequency agility of the microstrip monopole antenna can be yielded by proper control of its resonant structure. But the study of
frequency tuning through the non-resonant structure likes the feedline is rare. As discussed in Chapter 2, the frequency properties of feedline can be controlled by the DGS. However, due to the intrinsic limitation of the DGS which is located on the ground plane of the microstrip structure, the on-line frequency tuning controllability is difficult. To realize such tuning, DGSI (DGS with Islands) is one of DGS structures with some islands around the microstrip signal line was proposed [3.24]. Because of the coupling between islands and microstrip signal line, reactive loading is easily applied onto the signal line affecting the frequency characteristics. A single zero frequency tuning was demonstrated in the DGSI with capacitive loading and; its equivalent circuit was studied [3.25]. Based on this study of DGSI, an electrically tunable bandstop element using double U-shaped DGS with island was presented [3.26].

The proposed double U-shaped DGS with island (DGSI) is shown in Fig. 3.33. It has a 50Ω microstrip line on the top and two U-shaped patterns that are symmetrically etched in the ground plane. Each U-shaped pattern consists of three etched lines with different widths ($W_1$, $W_2$ and $W_3$). Using different lengths of these U-shaped patterns ($L_1$ and $L_2$, where $L_1 > L_2$), these two DGS patterns can be embedded with the open-end alignment. In addition, two islands are placed above and below the microstrip signal line. Reactive loading can be applied to the islands in order to tune the frequency response of the DGS structure.

![Figure 3.33 DGSI with double equilateral U-shaped defected ground pattern (a) Top side and (b) Bottom side.](image)
As reported in [3.26], the proposed structure also offers two transmission zeros. An example DGSI unit has been simulated with IE3D on the Rogers RO4003 substrate with relative dielectric constant $\varepsilon_r$ of 3.38 and a thickness $h$ of 1.524 mm. The simulation results are shown in Fig. 3.34, in which it clearly records two finite zeros at distinct frequencies $f_1$ and $f_2$. When $W_1 = W_2 = W_3 = 2$ mm, $L_1 = 12$ mm and $L_2 = 8$ mm, the island lines dimensions are fixed at $W = 3.5$ mm, $L = 20$ mm and gap $s = 0.2$ mm. The two zeros are located at 2.07 GHz and 2.7 GHz and they yield two bandgaps having about 20.1% and 5.8% 10 dB-bandwidth. Like the reported DGSI, these two attenuation poles will be distanced to lower frequencies when longer lengths $L_1$ and $L_2$ are used. The variation of these bandgaps can also be observed by the reactive loading to the island lines. The phenomenon is mainly due to coupling between the signal line and the island microstrips. In the proposed DGSI unit above, the bandgaps can be tuned by the DGSI geometry; like the length of the U-shaped pattern ($L_1$ and $L_2$) and its widths ($W_1$, $W_2$ and $W_3$) and; length and width of the island ($L$ and $W$).

![Figure 3.34 Simulated $|S_{21}|$ of proposed DGSI with double equilateral U-shaped defected ground pattern.](image)
3.3.1 LENGTH VARIATION OF U-SHAPED DGSI

When the widths are fixed and the lengths ($L_1$ and $L_2$) vary from 12 mm to 11 mm and 8 mm to 7 mm, the dual attenuation pole characteristic is studied. From Fig. 3.35, it is clear that the two attenuation poles can be relocated by adjusting lengths of the DGSI. The individual length adjustment can control the pole independently, and thus will be more beneficial for the bandgap design and tuning. As such, distinct transmission zeros can be simply designed with longer and shorter lengths ($L_1$ and $L_2$).

![Figure 3.35 Simulated $|S_{21}|$ parameter of double U-shaped DGS unit with $W_1 = W_3 = 2$ mm, $W_2 = 2$ mm and different lengths (--- $L_1 = 12$ mm & $L_2 = 8$ mm, -- $L_1 = 11$ mm & $L_2 = 8$ mm, --- $L_1 = 12$ mm & $L_2 = 7$ mm).](image)

3.3.2 WIDTH VARIATION OF U-SHAPED DGSI

For the width variation of U-Shaped DGS from 2 mm to 3 mm, like the effect reported in [3.27], these two attenuation poles will be distanced to lower frequencies when larger widths $W_1$ and $W_3$ are used. Both the attenuation poles will be relocated to higher frequencies if only the width $W_2$ is increased. As illustrated in Figure 3.36, the change of these width is plotted against the transmission zero variation.
3.3.3 **Length Variation of Islands**

In addition to the above DGSI geometrical variation, the length variation of island microstrips is studied. When it is changed from 18 mm to 20 mm, the transmission zero at 2.06 GHz is varied from 2.16 GHz to 1.96 GHz as in Figure 3.37. In contrary to such variation, the transmission zero would be relocated to higher frequency.

---

**Figure 3.36** Simulated $|S_{21}|$ parameter of double U-shaped DGS unit with $L_1 = 12$ mm, $L_2 = 8$ mm and different widths (− $W_1 = W_2 = W_3 = 2$ mm, $W_1 = W_2 = 3$ mm & $W_3 = 2$ mm, $W_1 = W_2 = 2$ mm & $W_3 = 3$ mm).

**Figure 3.37** Simulated $|S_{21}|$ parameter of island microstrip with $W = 3.5$ mm and different lengths (− 18 mm, −− 20 mm and −−− 22 mm).
3.3.4 Width Variation of Islands

Similar to the study of the length of the islands microstrip, transmission zero at 2.06 GHz is elevated from 2 GHz to 2.12 GHz as shown in Figure 3.38 whilst the width of the islands is changed from 3 mm to 4 mm and length are fixed at 20 mm. Therefore, the transmission zeros can be placed by controlling the geometry of the islands microstrip.

![Figure 3.38 Simulated \( |S_{21}| \) parameter of island microstrips with \( L = 20 \text{ mm} \) and different widths \((-3 \text{ mm}, \ldots , 3.5 \text{ mm and } -4 \text{ mm})\).](image)

3.3.5 Tunable Bandstop Element Using Capacitive Loaded Double U-shaped DGSI

Base on the above parametric analyses, it is obvious that the transmission zeros can be relocated significantly by tuning the length of the island microstrips of the proposed DGSI structure. A tunable bandstop element can be implemented by incorporating the DGSI structure with varactor which is terminated at one end of the island microstrip intuitively and this proposed structure is shown in Figure 3.39. To verify the tunable ability of the proposed DGSI element, an experimental tunable DGSI bandgap element with two transmission zeros at 2 GHz and 2.6 GHz is implemented using the Rogers RO4003 substrate.
with relative dielectric constant $\varepsilon_r = 3.38$ and substrate height $h = 1.524$ mm. Its geometry is shown Figure 3.39 and Infineon BB833 varactor is used.

The measurement results of the prototype circuit have been recorded in Figure 3.40. In this figure, the black curve shows the transfer characteristic of the elements with 10 V bias voltages. The two designed transmission zeros are located at 2.07 GHz and 2.7 GHz respectively. Additional zero is introduced. The lower transmission zero contributes a bandgap with 10-dB bandwidth of 416 MHz at 2.07 GHz whilst the upper zero forms the bandgap at 2.7 GHz with 157 MHz bandwidth.
By decreasing the bias voltage from 10 V to 5 V, 2 V and 0 V, the reverse bias capacitance of the varactor is increasing and these two transmission zeros can be relocated to lower frequencies accordingly. Meanwhile, the bandwidths of both bandgaps are changed from 60 MHz to 340 MHz and 170 MHz to 184 MHz respectively. Besides the two transmission zeros which obtained from the double U-shape DGS pattern, an extra bandgap is obtained from this structure due to the use of the varactor and it is located at 1.3 GHz with 58 MHz 10-dB attenuation bandwidth. By changing the bias voltage, this bandgap can be tuned to 1.8 GHz and its bandwidth is changed from 58 MHz to 342 MHz. The top view and bottom view photos of the prototyped tunable DGSI are shown in Figure 2.41(a) and (b) respectively.

![Top view and Bottom view photos](image)

Figure 3.41 Circuit of the proposed tunable DGSI bandstop element (a) Top view and (b) Bottom view.

### 3.4 A Tunable Monopole Antenna Using Double U-Shaped Defected Ground Structure with Islands

As reported in previous section, the varactor loaded DGSI element can electrically tune the transmission zeros of the bandstop element which is developed based on the microstrip line. It is mainly relied on the tuning of the matching impedance of the transmission line structure. Therefore, the above proposed structure is further applied to the study of reconfigurable antenna in this section. With the help of the varactor loaded DGSI structure and parametric analyses in Section 3.2, a tunable antenna achieves frequency tuning of the 2nd frequency band while the fundamental one is kept unchanged is proposed and reported
For *in-situ* frequency tuning, the varactor is proposed to terminate at the end of the island microstrip as shown in Figure 3.42. With the terminated varactor, the electrical length of island microstrip can be varied and thus; the antenna’s operating frequency and bandwidth are tuned indirectly.

To verify such tuning ability of the proposed active antenna, Infineon BB833 varactor is used and the prototype antenna with dimensions listed in Table 3.3 is fabricated on the RO4003 substrate with dielectric height $h = 1.524$ mm. The photo of the prototyped antenna is shown in Figure 3.43.

<table>
<thead>
<tr>
<th>$L_1$</th>
<th>$L_2$</th>
<th>$L_3$</th>
<th>$L_4$</th>
<th>$W_1$</th>
<th>$W_2$</th>
<th>$W_3$</th>
<th>$S$</th>
</tr>
</thead>
<tbody>
<tr>
<td>88</td>
<td>30</td>
<td>30</td>
<td>22</td>
<td>4</td>
<td>4</td>
<td>4</td>
<td>0.5</td>
</tr>
<tr>
<td>$L_5$</td>
<td>$L_6$</td>
<td>$W_4$</td>
<td>$W_5$</td>
<td>$W_6$</td>
<td>$W_7$</td>
<td>$W_8$</td>
<td></td>
</tr>
<tr>
<td>20</td>
<td>14</td>
<td>14</td>
<td>24</td>
<td>2</td>
<td>2</td>
<td>2</td>
<td></td>
</tr>
</tbody>
</table>

Figure 3.42 Proposed monopole antenna with double U-shaped DGSIS (a) Top and (b) Bottom.
The experimental return loss measurements of the prototype antenna with bias voltage in the range from 29 V to 5 V are recorded and plotted in Figure 3.44. According to the previous study, it is expected that with the varactor termination, the 2\textsuperscript{nd} matching frequency of the antenna is lowered to 2.78 GHz. It is due to the capacitive loading which is contributed by the varactor. The 2\textsuperscript{nd} operating frequency is located at 2.78 GHz when 29 V bias voltage is applied. Moreover, the 10-dB bandwidth of 64 MHz is obtained whilst 16 dB matching level is recorded. When bias voltage is decreased to 16 V and this increases the capacitance of the varactor indeed, the matching frequency is relocated to 2.628 GHz, matching level is now better than 21.2 dB but there is 62 MHz bandwidth shrink compared with that of condition using 29 V. When bias voltage is further decreased to 10 V and 5 V respectively, antenna’s operating frequency is lowered to 2.46 GHz and 2.156 GHz. Also, 44 MHz and 15 MHz 10-dB bandwidth are recorded for the above bias voltages variation. It shows the frequency of the 2\textsuperscript{nd} frequency band of the antenna.
Moreover, the antenna’s fundamental operating frequency around 1 GHz is expected to be relocated while different bias voltage is applied. However, it is the 1st operating frequency around 1 GHz is significantly degraded due to the lacking optimization of feedline with varactor model.

### 3.5 SUMMARY

In this chapter, the methods of the frequency tuning technique of antenna have already been studied with emphasis on the microstrip monopole antenna. It is known that the dimensions of the radiation patch and the feedline are more sensitive to the operation frequency. Based on this study, a varactor tuned microstrip monopole antenna is proposed by applying an active DGS1 structure on its feedline.
3.6 REFERENCES


In this chapter, a stub loaded structure based on DGS is introduced and used to design the tunable dual-band antenna. The property of the stub-loaded structure is analyzed theoretically and confirmed by full-wave electromagnetic simulation [4.1]. In the transformer design [4.2], the stub-loaded structure easily made the transformer to control the resonant frequencies and its resonant frequencies can be separated into two mode, odd-mode and even-mode. In these cases, the resonant frequencies of even-mode can be conveniently tuned by the structure, while those of the odd-mode remain almost the same. The characteristic of the stub-loaded structure is discussed in following sections.

4.1 CHARACTERISTIC OF STUB LOADED STRUCTURE

The stub-loaded structure can be classified into two models, Open Stub-Loaded Structure (OSLS) and Shot Stub-Loaded Structure (SSLS), either of which comprises a half-wavelength resonator and an open stub or a short stub shunted at the midpoint.

For the OSLS analysis, it is known that the Stub-Loaded Structure (SLS) consists of a
conventional microstrip half wavelength resonator and an open stub as Figure 4.1, where $Y_1$, $L_1$, $Y_2$ and $L_2$ denote the characteristic admittances and lengths of the microstrip line and open stub, respectively. The open stub is shunted at the midpoint of the microstrip line [4.1]. This resonator has easily controlled resonant frequencies. Its resonant frequencies of even-mode can be tuned, while those of the odd-mode remain almost the same. The analysis of the stub-loaded resonator is studied as follows:

![Figure 4.1 A stub-load resonator [4.1].](image)

In odd-mode state, the middle of the SLS is shorted to the ground as shown in Figure 4.2.

![Figure 4.2 The equivalent circuit in odd-mode of the SLR [4.1].](image)

The resulting input admittance $Y_{in,odd}$ can be written as

$$Y_{in,odd} = Y_1 \frac{j \tan \left( \frac{\beta L_1}{2} \right)}{\beta L_1}$$

(4.1)

where $\beta = \theta_1$.

From resonance condition of $Y_{in,odd} = 0$,

According to the theory of the microstrip transmission line, it is known that

$$\lambda = \frac{c}{f \sqrt{\varepsilon_{eff}}}$$

(4.2)

or
When \( \lambda = 4\ell \)

\[
f = \frac{c}{4\ell \sqrt{\varepsilon_{eff}}} \tag{4.4}
\]

In Figure 4.2, we got \( \ell = \frac{L_2}{2} \) and rewrite the equation as

\[
f = \frac{c}{2L_1 \sqrt{\varepsilon_{eff}}} \tag{4.5}
\]

so the odd-mode resonant frequencies can be deduced as

\[
f_{\text{odd}} = \frac{(2n-1)c}{2L_1 \sqrt{\varepsilon_{eff}}} \tag{4.6}
\]

where \( n = 1, 2, 3, \ldots \), \( c \) is the speed of light in free space and \( \varepsilon_{eff} \) denotes the effective dielectric constant of the substrate. It is seen that the odd-mode resonant frequencies are not affected by the open stub.

Figure 4.3 Even-mode equivalent circuit.

The SLR is bisected in the middle symmetrically to form the even-mode and open to the air.

The Figure is shown in Figure 4.3. The input admittance \( Y_{in,\text{even}} \) for even-mode can be approximately obtained as
\[
Y_{in,\text{even}} = jY_1 \frac{2Y_1 \tan\left(\frac{\theta_1}{2}\right) + Y_2 \tan \theta_2}{2Y_1 - Y_2 \tan\left(\frac{\theta_1}{2}\right) \tan \theta_2}
\]

where \( \theta_2 = \beta L_2 \),

Due to the resonance condition of \( Y_{in,\text{even}} = 0 \), and \( \ell = \frac{L_1}{2} + L_2 \), therefore we can rewrite the equation (4.4) as

\[
f = \frac{c}{(2L_1 + 4L_2) \sqrt{\varepsilon_{\text{eff}}}}
\]

So the even-mode resonance frequencies can thus be derived as

\[
f_{\text{even}} = \frac{2nc}{2(L_1 + 2L_2) \sqrt{\varepsilon_{\text{eff}}}} = \frac{nc}{(L_1 + 2L_2) \sqrt{\varepsilon_{\text{eff}}}}
\]

To verify the above results, the full wave electromagnetic simulation is carried out by using IE3D. The length \( L_1 \) is fixed at 40 mm and varying the stub length \( L_2 \) from 1 mm to 20 mm.

From the simulation results in Figure 4.4, it is shown that the fundamental even-mode resonant frequency can be shifted within a wide range and the fundamental odd-mode is kept in the fixed frequency.
As for the Short Stub-Loaded Structure (SSLS), the first resonant frequency is at even-mode.

The equivalent circuit is thus shown in Figure 4.5 and the resonant condition is given by

\[
\tan \theta_3 \tan \left( \frac{\theta_4}{2} \right) = \frac{Y_3}{2Y_4} \quad (4.10)
\]

where \( \theta_3 \) and \( \theta_4 \) are the electric length of the half wavelength resonator in the SSLS and the open stub, respectively. Also, for the special case of \( Y_3 = 2Y_4 \), the resonant frequency here can be derived as

\[
f_{SSLR_{\text{even}}} = \frac{c}{2(L_3 + 2L_4)\sqrt{\varepsilon_{\text{eff}}}} \quad (4.11)
\]

In the case where the first odd-mode is excited in the SSLS, the resonant characteristic is almost the same as that of the OSLS, which is mainly determined by the half-wavelength resonator. Therefore, the second resonant frequency of the SSLS can be given by

\[
f_{SSLR_{\text{odd}}} = \frac{c}{2L_3\sqrt{\varepsilon_{\text{eff}}}} \quad (4.12)
\]
4.2 DGS Stub-Loaded Structure

A stub-loaded structure or resonator is formed by etching the U-Shaped DGS pattern on its ground as shown in Figure 4.6. According to the chapter 2 and chapter 3, it is known that the DGS structure placed under the microstrip stub line would affect the characteristic impedance and the electrical length of the stub line and those circuits can easily be controlled. Therefore, the DGS is embedded under the open with short wavelength stub to controlled the even-mode resonator frequency of the SLS.

In the even-mode analysis, the input admittance \( Y_{in, even} \) for even-mode can be approximately obtained as (4.7)

\[
Y_{in, even} = jY_1 \frac{2Y_1 \tan \left( \frac{\theta_1}{2} \right) + Y_2 \tan \theta_2}{2Y_1 - Y_2 \tan \left( \frac{\theta_1}{2} \right) \tan \theta_2}
\]

(4.12)

\[
2Y_1 \tan \left( \frac{\theta_1}{2} \right) + Y_2 \tan \theta_2 = 0
\]

(4.13)

\[
\cot \left( \frac{\theta_1}{2} \right) \tan \theta_2 = -\frac{2Y_1}{Y_2}
\]

(4.14)

For tuning DGS, the characteristic impedance ratio is assumed as \( m \) and thus.

\[
\cot \left( \frac{\theta_1}{2} \right) \tan \theta_2 = -\frac{2Y_1}{Y_2} = -m
\]

(4.15)

\[
\tan \theta_2 = -m \tan \left( \frac{\theta_1}{2} \right)
\]

(4.16)
\[
\theta_2 + \frac{\theta_1^3}{3} = -m \left( \pi - \frac{\theta_1}{2} \right) + \frac{\left( \pi - \frac{\theta_1}{2} \right)^3}{3}
\] (4.17)

\[
\theta_2 + m_1 \theta_1 \approx m_0
\] (4.18)

Assume the condition of the even-mode is \( \theta_1 = 2\theta_2 \), and

\[
\theta_2 + 2m_1 \theta_2 \approx m_0
\] (4.19)

\[
\theta_2(1 + 2m_1) \approx m_0
\] (4.19a)

\[
\beta L_2 (1 + 2m_1) \approx m_0
\] (4.19b)

where \( \beta = \frac{2\pi}{\lambda} \) and \( \lambda = \frac{c}{\sqrt{\epsilon_{\text{eff}}}} \),

\[
\frac{2\pi \sqrt{\epsilon_{\text{eff}}}}{c} L_2 (1 + 2m_1) \approx m_0
\] (4.19c)

\[
\frac{2\pi \sqrt{\epsilon_{\text{eff}}}}{c} \approx \frac{m_0}{1 + 2m_1}
\] (4.19d)

Finally, we have

\[
f_{\text{DGS, even}} = \frac{c m_0}{2\pi L_2 \sqrt{\epsilon_{\text{eff}} (1 + 2m_1)}}
\] (4.20)

The frequency of the even-mode can be calculated by the equation (4.20). \( L_2 \) denote the length of the microstrip stub on the resonator.

Comparing (4.2) to (4.9), it is observed that additional degree of freedom in controlling even-mode frequency is obtained by DGS stub. In fact, it is anticipated the control of characteristic impedance may benefit the antenna impedance matching that is not easily realized by stub length change alone.

The dimensions of the DGS stub-loaded structure as shown Figure 4.7 is listed Table 4.1, this DGS stub loaded structure is now simulated. The simulation results of \(|S_{21}|\) are shown in Figure 4.8. The stub-loaded structure has two resonant frequencies at 0.97 GHz and 1.96 GHz.
GHz.

![Figure 4.7 A DGS stub loaded structure.](image)

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<table>
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</thead>
<tbody>
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<td>$L_2$</td>
<td>$W_1$</td>
<td>$W_2$</td>
<td>$L$</td>
<td>$W_{\text{gap}}$</td>
</tr>
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<td>30</td>
<td>4</td>
<td>3</td>
<td>10</td>
<td>1.5</td>
</tr>
</tbody>
</table>

The traditional method of tuning even-mode frequency is studied by varying the length of the microstrip stub directly. In the previous study, it is known that the DGS can be used to affect the characteristic impedance of the microstrip. It means that the DGS can vary the length and the width of the microstrip anew. Therefore, the DGS pattern is thus etched under the microstrip stub.

The dimension of the DGS structure will affect the performance of the resonators. For the microstrip resonators simulated, the DGS has the dimension of $L = 10$ mm, $W_{\text{gap}} = 1.5$ mm and $L_{\text{DGS}} = 36$ mm. The simulation results of the operating frequency against the length $L$ of the DGS structure is shown in Figure 4.8.
From the simulation result in Figure 4.8, the fundamental frequency of the resonator is kept, and the second operation frequency of the resonator is just inversely changed from 1.97 GHz to 1.96 GHz when the length of the DGS length $L$ is increased from 10 mm to 45 mm. It is obvious that the DGS performs similar even-mode frequency variation [4.1], the odd-mode frequency is kept and the even-mode frequency can be controlled by the dimension of DGS.

**4.3 Tunable DGS Stub-Loaded Dual Band Monopole Antenna**

As reported in previous section, tunable antenna can be performed by the electrical tuning with the varactor loaded on the top of the substrate. In the past decade years, many tunable or reconfigurable circuit design either in guided wave circuit or the radiated wave circuit always used some active element, such as varactors, PIN diodes and MEMS switches loaded on the microstrip line [4.3] - [4.10]. The example showed that the varactor loaded the DGSI element for varying the frequency in the previous chapter. In this chapter we studied an alternative active element - MEMS switch. It is known that MEMS switch can implement the frequency tuning in different microwave circuits. The principle of the
operation is very simply as a switch - on and off. Therefore, the MEMS switch can be used to connect the microstrip line directly. It can use to change the any dimension in different places of the microwave structure like the radiation patch [4.11], the length of the microstrip feedline, the length of the couple island [4.12] and so forth. For in-situ frequency tuning, the MEMS switch is proposed to vary the physical length of the DGS dimension. The analysis of the DGS stub-loaded structure is studied. It is known that the length of the DGS can vary the operating frequency.

4.3.1 SIMULATIONS OF TUNABLE DGS STUB-LOADED ANTENNA WITH DIFFERENT DGS SLOT LENGTHS

By applying proposed DGS stub-loaded structure for a tunable antenna with fundamental operating frequency at 500 MHz and even-mode operating frequency at 1 GHz; the antenna structure will has the structure as shown in Figure 4.9 and its dimensions are listed in Table 4.2 where the RO4003 substrate with height $h = 1.524$ mm) is used. In this antenna structure, a U-shape slotline is etched on the ground plane under a stub which is loaded on the antenna’s feedline.

![Figure 4.9 Proposed tunable dual band monopole antenna with U-shaped DGS stud load structure](image)

(a) Top view and (b) Bottom view.
Table 4.2 Dimensions of the proposed tunable dual band monopole antenna (mm)

<table>
<thead>
<tr>
<th>$L_2$</th>
<th>$L_r$</th>
<th>$L_f$</th>
<th>$L_{stub}$</th>
<th>$W_s$</th>
<th>$W_f$</th>
<th>$W_{stub}$</th>
<th>$W_r$</th>
</tr>
</thead>
<tbody>
<tr>
<td>210.8</td>
<td>109.8</td>
<td>101</td>
<td>30</td>
<td>114</td>
<td>4</td>
<td>3</td>
<td>84</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>$L_2$</th>
<th>$L_{DGS}$</th>
<th>$L_{DGS2}$</th>
<th>$W_{DGS}$</th>
<th>$W_s$</th>
<th>$g$</th>
</tr>
</thead>
<tbody>
<tr>
<td>97</td>
<td>10</td>
<td>36</td>
<td>1.5</td>
<td>114</td>
<td>2.5</td>
</tr>
</tbody>
</table>

To achieve the frequency agility, the DGS slot length $L_{DGS}$ is required to tune with different lengths. Moreover, MEMS switches are used to imitate the metal strips which terminate the slot and shorten its length electrically. In order to predict the antenna tuning performance, antenna with different slot lengths are simulated by the full-wave electromagnetic simulation and two thin metal strips are used to simulate the function of the MEMS switches as depicted in Figure 4.10. By adjusting the $L_{DGS}$ from 45 mm to 10 mm, the even-mode frequency of the antenna is varied from 1 GHz with 327 MHz impedance bandwidth to 1.22 GHz with 131 MHz impedance bandwidth while the fundamental operating frequency is kept at 0.5 GHz as shown in Figure 4.11. These simulation results show that the fundamental frequency of the dual band monopole antenna is independent of the length of the DGS. And the even-mode frequency is dependent on the length of the DGS as expected. It has 210 MHz tuning range of the antenna.

Figure 4.10 The length of the DGS varied by short thin line.
In addition, the radiation patterns of the antenna for $L_{\text{DGS}} = 10$ mm and $45$ mm are simulated and illustrated in Figure 4.12 and Figure 4.13 respectively. From these figures, the proposed antenna shows a monopole likes radiation pattern. In Figure 4.12(a) and Figure 4.13(a), the antenna has antenna gain about 2 dBi at the fundamental operating frequency of 500 MHz for both $L_{\text{DGS}} = 10$ mm and $L_{\text{DGS}} = 45$ mm. In Figure 4.12(b), it shows the radiation patterns of even-mode operating frequency at 1.19 GHz ($L_{\text{DGS}} = 10$ mm), the antenna has gain about 4 dBi. In Figure 4.13(b) the antenna is decreased to about 3 dBi at 0.91 GHz when $L_{\text{DGS}}$ is increased to 45 mm. Moreover, these simulation results show the cross-polarization of the antenna are about -8 dBi and -7 dBi at 1.19 GHz ($L_{\text{DGS}} = 10$ mm) and 0.91 ($L_{\text{DGS}} = 45$ mm) GHz respectively.
Figure 4.12 The radiation pattern for \( \text{L}_{\text{DGS}} = 10 \text{ mm} \)
(a) fundamental frequency, 500 MHz and (b) even-mode frequency, 1.19 GHz.
4.3.2 MEASUREMENT RESULTS OF TUNABLE DGS STUB-LOADED ANTENNA

To verify the above simulation results, the proposed tunable DGS stub-loaded antenna is implemented on the designed substrate with Radant SPST MEMS switches RMSW101. The prototype antenna’s photo is depicted in Figure 4.14. Two pairs of MEMS switches SW1 and
SW2 are bond wired to the surrounding pads on the top side of antenna as shown in Figure 4.15. These pads are located at 10 mm and 30 mm apart the bend of the DGS slot respectively and via connecting to the ground plane.

![Figure 4.14 Prototype tunable dual band DGS stub-loaded monopole antenna using MEMS switches. (a) Top view and (b) Bottom view.](image)

![Figure 4.15 The photo of the bond wired MEMS switch.](image)

Then the prototype antenna is measured with different DGS slot lengths by controlling the on-off status of MEMS switches pairs SW1 and SW2. The corresponding DGS slot length for different MEMS switches on-off status are summarized in Table 4.3. The measured $|S_{11}|$ of the antenna are shown in Figure 4.16.

<table>
<thead>
<tr>
<th>SW1</th>
<th>SW2</th>
<th>$L_{DGSS}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Off</td>
<td>Off</td>
<td>45 mm</td>
</tr>
<tr>
<td>Off</td>
<td>On</td>
<td>30 mm</td>
</tr>
<tr>
<td>On</td>
<td>Off</td>
<td>10 mm</td>
</tr>
</tbody>
</table>

In Figure 4.16(a), it shows the measured results for $L_{DGSS} = 10$ mm. The measured
fundamental frequency of antenna is located at 0.55 GHz whilst the even-mode frequency is located at 1.19 GHz with 120 MHz impedance bandwidth and 12.27 dB return loss. Similarly, the experimental measurements for $L_{\text{DGS}} = 30$ mm and 45 mm are also illustrated in Figure 4.16 (b) and (c) respectively. Table 4.4 summarizes the simulated and measured results of operating frequencies, bandwidth and return losses for fundamental frequency and the even-mode one are listed in Table 4.5.
Figure 4.16 Comparisons of simulated and measured $|S_{11}|$ of tunable DGS slot-loaded monopole antenna by varying DGS slot length $L_{DGS}$. (a) $L_{DGS} = 10$ mm, (b) $L_{DGS} = 30$ mm and (c) $L_{DGS} = 45$ mm. (--- simulation and - measurement)

Table 4.4 Comparisons of the simulation and measurement results for fundamental frequency of antenna with different $L_{DGS}$

<table>
<thead>
<tr>
<th>$L_{DGS}$ (mm)</th>
<th>Simulation</th>
<th>Measurement</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Frequency (GHz)</td>
<td>Bandwidth (MHz)</td>
</tr>
<tr>
<td>10</td>
<td>0.5</td>
<td>112</td>
</tr>
<tr>
<td>30</td>
<td>0.5</td>
<td>92</td>
</tr>
<tr>
<td>45</td>
<td>0.49</td>
<td>70</td>
</tr>
</tbody>
</table>

Table 4.5 Comparisons of the simulation and measurement results for even-mode frequency of antenna with different $L_{DGS}$

<table>
<thead>
<tr>
<th>$L_{DGS}$ (mm)</th>
<th>Simulation</th>
<th>Measurement</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Frequency (GHz)</td>
<td>Bandwidth (MHz)</td>
</tr>
<tr>
<td>10</td>
<td>1.24</td>
<td>130</td>
</tr>
<tr>
<td>30</td>
<td>1.10</td>
<td>230</td>
</tr>
<tr>
<td>45</td>
<td>1.00</td>
<td>320</td>
</tr>
</tbody>
</table>

From Figure 4.16, Table 4.4 and Table 4.5, the measured results show good agreement with the simulated one. The measured ripple is caused by the reflections from the measurement environment. The fundamental and even-mode frequencies record 50 MHz and 90 MHz.
Frequency shift in maximum for $L_{DGS}$ is 10 mm and 45 mm respectively. The measurement results for different $L_{DGS}$ are re-plotted in Figure 4.17 and it shows the fundamental frequency stays around at the designed frequency 0.5 GHz. The even-mode frequency achieves 280 MHz frequency tuning range from 1.91 GHz to 0.91 GHz and bandwidth are varied from 120 MHz to 382 MHz. All measurements maintain the overall good matching level. However, due to the lacking of measurement facilities, the radiation patterns are not measured.

![Figure 4.17 Comparisons of measured $|S_{11}|$ of tunable DGS slot-loaded monopole antenna by varying DGS slot length $L_{DGS}$.](image)

The prototype tunable dual band monopole antenna covers the frequency range from 1 GHz to 1.2 GHz. The comparison of the measurement and simulation is shown in Figure 4.16. The Figure 4.16a shows that the result of the antenna which the length of the DGS $L$ is set at 10 mm is simulated. The fundamental frequency is located at 500 MHz with 112 MHz impedance bandwidth and 23.38 dB return loss. The even-mode frequency is located at 1.24 GHz with 130 MHz impedance bandwidth and 34.85 dB return loss. Compare with the measurement results, the fundamental frequency was found at 550 MHz with 29 MHz impedance bandwidth and 27.48 dB return loss. And the even-mode is relocated at 1.19 GHz with 120 MHz impedance bandwidth and 12.27 dB return loss. The radiation patterns are shown in
Figure 4.12 to Figure 4.13.

4.4 SUMMARY

In this chapter, the DGS stub loaded structure is proposed and analyzed analytically. Based on the analysis results, it is observed that additional degree of freedom in controlling even-mode frequency is obtained by DGS stub and it is anticipated that the control of characteristic impedance may benefit to the antenna impedance matching. Using the proposed DGS stub-loaded structure, a tunable dual-band microstrip monopole is designed and implemented with the use of MEMS switches. The MEMS switches are controlled in pair to adjust the slot length of the DGS stub-loaded structure so as to achieve the frequency agility. The antennas are fabricated and experimentally characterized with good agreement to the simulation results. By tuning the MEMS switches, it demonstrates the uses of the MEMS switches for DGS slot length control and the length is switched from 10 mm to 45 mm, and thus the even-mode frequency of the antenna is tuned from 1.19 GHz to 0.91 GHz whilst overall good matching is maintained at -13 dB level.
4.5 REFERENCES


5.1 SUMMARY OF THE THESIS

Recently, the dual-band antenna has been studied extensively as a key component in modern communications. In particular, the added function of frequency agility of the above dual-band antenna is still under active research. To this end, this thesis reports the research and development of tunable antenna using defected grounded structure (DGS) and some active elements namely varactors and MEMS switches. The applications of the proposed dual-band antenna for high voltage facilities’ partial discharge (PD) monitoring are also highlighted in order to demonstrate its usefulness. Two DGS antennas were designed and experimentally characterized and, the first operating frequency were at around 0.5 GHz/1 GHz respectively whilst the second operating frequency was designed to meet modern multi-band communications, for example, GSM, WCDMA, RFID and WiFi.

Chapter 1 introduces the importance of PD and its UHF characteristics. The need of tunable dual-band antenna is demonstrated with some practical spectrum measurement of spark
generator. Chapter 2 goes to the discussion of the microstrip antenna and latest electromagnetic bandgap structure with special emphasis on DGS. Some DGS applications to both guided- and radiated-wave components were also reported. In particular, a new DGS using U-shape is proposed and its usefulness is demonstrated by examples of filter and antenna.

In chapter 3, different antennas with frequency agility have been reviewed with emphasis on the microstrip monopole antenna. Followed the study of DGS in last chapter, a varactor tuned microstrip monopole antenna is proposed by applying a novel defected grounded structure with islands (DGSI) structure on its feedline. Overall good tunable range is experimentally verified by a monopole antenna with operating frequency range between 2.156 GHz and 2.78 GHz. Chapter 4 studies the use of MEMS switched and DGS for tunable antenna design and a novel DGS stub-loaded structure is presented. This structure can be tuned by MEMS switched controlled DGS dimension and it is experimentally verified by a 500 MHz/1 GHz dual-band monopole antenna. This fundamental operating frequency of the antenna is kept at 0.49 GHz for the PD detection whilst the second operating frequency is switched from 0.91 GHz to 1.01 GHz and 1.19 GHz respectively by controlling the on/off status of two pairs of MEMS switches.

## 5.2 Future Works

Concerning the dual-band antenna proposed in this work, the first and second operating frequencies are controlled by radiation patch and DGS respectively. However, there are some directions for future development should be focused:

- Design of dual-band antenna - apply the DGS stub loaded structure to other compact microstrip dual-band monopole antennas feedline for frequency tuning and relate the antenna design with the design of DGS stub loaded structure [5.1] - [5.3].

- MEMS switch modeling - due to the presence of the parasitic elements from the
MEMS devices’ physical structure and inductance from bond wires, frequency shift is observed in experimental results when compared with the simulations one in general. To further elevate the accuracy circuit with MEMS switch, the circuitry and EM models of MEMS switch can be studied and accounted into the antennas’ full-wave simulations [5.4] - [5.8].

- Integration of dual-band antenna with other components in PD detection - to further reduce the size of the PD detection system, antenna can be integrated with other guided wave components, i.e. filter, diplexer and so forth [5.9] - [5.11].
5.3 REFERENCES


APPENDIX A

ROGERS RO4003
Advanced Circuit Materials

RO4000® Series High Frequency Circuit Materials

<table>
<thead>
<tr>
<th>Features:</th>
<th>Benefits:</th>
</tr>
</thead>
<tbody>
<tr>
<td>RO4000® materials are reinforced hydrocarbon/ceramic laminates</td>
<td>• Designed for performance sensitive, high volume applications</td>
</tr>
<tr>
<td>Low dielectric tolerance and low loss</td>
<td>• Excellent electrical performance</td>
</tr>
<tr>
<td>Stable electrical properties vs. frequency</td>
<td>• Always maintains similar characteristics over a range of frequencies</td>
</tr>
<tr>
<td>Lead-free process compatible</td>
<td>• No blistering or deformation</td>
</tr>
<tr>
<td>Low Z-axis expansion</td>
<td>• Reliable plated through holes</td>
</tr>
<tr>
<td>Low in-plane expansion coefficient</td>
<td>• Remains stable over an entire range of circuit processing temperatures</td>
</tr>
<tr>
<td>Volume manufacturing process</td>
<td>• RO4000 laminates can be fabricated using standard glass epoxy processes</td>
</tr>
<tr>
<td></td>
<td>• Competitively priced</td>
</tr>
</tbody>
</table>

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

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). 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 multilayer boards constructions. The low Z-axis CTE of RO4000 laminates provides reliable printed through-hole quality, even in extreme thermal shock environments. RO4000 series material has a Tg of 356°C (673°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 specialised 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.

RO4000™ 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 and /11 for RO4350B materials.

The world runs better with Rogers."
Chart 1: RO4000 Series Materials
Dielectric Constant vs. Temperature

Chart 2: RO4000 Series Materials
Dielectric Constant vs. Frequency

Chart 3: Microstrip Insertion Loss (0.030" Dielectric Thickness)
<table>
<thead>
<tr>
<th>Property</th>
<th>Typical Value</th>
<th>Direction</th>
<th>Units</th>
<th>Condition</th>
<th>Test Method</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dielectric Constant, εr (Process specification)</td>
<td>3.38 ± 0.05</td>
<td>f) 3.48 ± 0.05</td>
<td>Z</td>
<td>--</td>
<td>IPC-TM-650 2.5.5.3</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>10 GHz/23°C</td>
<td>Printed Circuit Board</td>
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<td></td>
<td></td>
<td></td>
<td></td>
<td>2.5 GHz/23°C</td>
<td>Full Sheet Resonance</td>
</tr>
<tr>
<td>III Dielectric Constant, εr (Recommended for use in circuit design)</td>
<td>3.55</td>
<td>3.66</td>
<td>Z</td>
<td>--</td>
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</tr>
<tr>
<td></td>
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<td></td>
<td></td>
<td>FSR/23°C</td>
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<td></td>
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<td>Dissipation Factor tan δ</td>
<td>0.0027</td>
<td>0.0037</td>
<td>Z</td>
<td>--</td>
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<tr>
<td></td>
<td>0.0021</td>
<td>0.0031</td>
<td></td>
<td>10 GHz/23°C</td>
<td>Printed Circuit Board</td>
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<td></td>
<td>2.5 GHz/23°C</td>
<td>Full Sheet Resonance</td>
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<tr>
<td>Thermal Coefficient of εr</td>
<td>+40</td>
<td>+50</td>
<td>Z</td>
<td>ppm/°C</td>
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<td></td>
<td></td>
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<td></td>
<td>-50°C to 150°C</td>
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<td>1.7 x 10^12</td>
<td>1.2 x 10^12</td>
<td>MΩ•cm</td>
<td>COND A</td>
<td>IPC-TM-650 2.5.17.1</td>
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<td></td>
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<td>Printed Circuit Board</td>
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<tr>
<td>Surface Resistivity</td>
<td>4.2 x 10^10</td>
<td>5.7 x 10^10</td>
<td>Ω</td>
<td>COND A</td>
<td>IPC-TM-650 2.5.17.1</td>
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<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>Printed Circuit Board</td>
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<tr>
<td>Electrical Strength</td>
<td>31.2 (780)</td>
<td>31.2 (780)</td>
<td>Z</td>
<td>KV/mm</td>
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<tr>
<td></td>
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<td></td>
<td></td>
<td>0.51 mm (0.020”)</td>
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<tr>
<td>Tensile Modulus</td>
<td>22.58 (3000)</td>
<td>11.47 (1644)</td>
<td>Y</td>
<td>MPa</td>
<td>ASTM D638</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
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<td>MPa [psi]</td>
<td>Printed Circuit Board</td>
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<tr>
<td>Tensile Strength</td>
<td>141 (20.4)</td>
<td>175 (25.4)</td>
<td>Y</td>
<td>MPa</td>
<td>ASTM D638</td>
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<td></td>
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<td></td>
<td>MPa [psi]</td>
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<td>Flexural Strength</td>
<td>276 (40)</td>
<td>255 (30)</td>
<td></td>
<td>MPa</td>
<td>ASTM D638</td>
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<td>MPa [psi]</td>
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<td>X,Y</td>
<td>mm/m</td>
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<td></td>
<td></td>
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<td></td>
<td>mm/inch</td>
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<td>14</td>
<td>X</td>
<td>ppm/°C</td>
<td>@ 68°C IPC-TM-650 2.4.4.4</td>
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<td>46</td>
<td>55</td>
<td>Y</td>
<td>ppm/°C</td>
<td>@ 25°C IPC-TM-650 2.4.4.4</td>
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<td>&gt;280</td>
<td></td>
<td>°C</td>
<td>DSC A IPC-TM-650 2.4.24</td>
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<td></td>
<td></td>
<td>Printed Circuit Board</td>
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<tr>
<td>1d</td>
<td>425</td>
<td>590</td>
<td></td>
<td>°C</td>
<td>ASTM D3550</td>
</tr>
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<td></td>
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<td>Printed Circuit Board</td>
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<tr>
<td>Thermal Conductivity</td>
<td>0.31</td>
<td>0.69</td>
<td></td>
<td>W/m·K</td>
<td>ASTM C518</td>
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<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>Printed Circuit Board</td>
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<td>Moisture Absorption</td>
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<td>0.06</td>
<td>%</td>
<td>48 hrs immersion</td>
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<td></td>
<td></td>
<td></td>
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<td>0.06° sample</td>
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<td>Density</td>
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<td>1.86</td>
<td>g/cm³</td>
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<td>ASTM D792</td>
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<tr>
<td>Copper Peel Strength</td>
<td>1.06 (6.0)</td>
<td>0.98 (5.0)</td>
<td>N/mm²</td>
<td>after solder final 1 oz. EDC Foil</td>
<td>IPC-TM-650 2.4.8</td>
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<td></td>
<td></td>
<td></td>
<td>28°C</td>
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<td>Flammability</td>
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<td>V-0</td>
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<td></td>
<td>UL 94</td>
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<td></td>
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<td></td>
<td>Printed Circuit Board</td>
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</tbody>
</table>

1. **Note on 94V-0** RO4350B Lofro™ laminates do not share the same UL designation as standard RO4350B 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.

RO4000 Lofro laminate uses a modified version of RO4000 resin system to bond reverse treated foil. Values shown above are RO4000 laminates with out the addition of the Lofro resin. For double sided board, the Lofro foil results in a thickness increase of approximately 0.0005” (0.013mm) and the DK is apparent matery 2.4. Therefore, effective DK is highly dependent on core thickness.

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.
World Class Performance

Rogers Corporation (NYSE: ROG), headquartered in Rogers, Conn., is a global technology leader in the development and manufacture of high performance, specialty material-based products for a variety of applications in diverse markets including portable communications, infrastructure, computer and office equipment, consumer products, ground transportation, aerospace and defense. In an ever-changing world, where product design and manufacturing often take place on different sides of the planet, Rogers has the global reach to meet customer needs. Rogers operates facilities in the United States, Europe and Asia. The world runs better with Rogers.

CONTACT INFORMATION:

USA: Rogers Advanced Circuit Materials, G0-1002 certified Tel: 480-961-3822 Fax: 480-961-4533

Belgium: Rogers Suisse SA Tel: +32 92-243-611

Japan: Rogers Japan Inc. Tel: 81-3-5200-2700 Fax: 81-3-5200-0271

Taiwan: Rogers Taiwan Inc. Tel: 886-2-86609856 Fax: 886-2-86609357

Korea: Rogers Korea Inc. Tel: 82-31-716-6112 Fax: 82-31-716-6208


China: Rogers (Shanghai) International Trading Co., Ltd (Shanghai Office) Tel: 86-21-62175599 Fax: 86-21-62677913

China: Rogers (Shanghai) International Trading Co., Ltd (Beijing Office) Tel: 86-10-5820-7667 Fax: 86-10-5820-7997

China: Rogers International Trading Co., Ltd (Shenzhen Office) Tel: 86-755-8236-6060 Fax: 86-755-8236-6123

The information in this data sheet is intended to assist you in designing with Rogers’ circuit material laminates. 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 material laminates for each application.

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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Revised 05/2010 0891-0110-0.5CC

Publication #92-004

<table>
<thead>
<tr>
<th>Standard Thickness</th>
<th>Standard Panel Size</th>
<th>Standard Copper Cladding</th>
</tr>
</thead>
<tbody>
<tr>
<td>RO4003C: 0.006&quot; (0.203mm), 0.012&quot; (0.305mm), 0.016&quot; (0.406mm), 0.020&quot; (0.508mm), 0.022&quot; (0.568mm), 0.025&quot; (0.635mm), 0.027&quot; (0.686mm), 0.030&quot; (0.762mm), 0.060&quot; (1.524mm), 0.120&quot; (3.048mm)</td>
<td>12&quot; X 18&quot; (305 X 457 mm)</td>
<td>1/8 oz. (17μm), 1 oz. (35μm) and 2 oz. (70μm) electrodipped copper foil. LaPro Reverse Treated EDC for PIM Sensitive Applications: 1/4 oz. (17μm), 1 oz. (35μm) Note: LaPro EDC foil adds .00035&quot; to the panel thickness per side.</td>
</tr>
<tr>
<td>RO4350B: 0.004&quot; (0.101mm), 0.0064&quot; (0.168mm), 0.010&quot; (0.254mm), 0.0139&quot; (0.353mm), 0.0164&quot; (0.412mm), 0.020&quot; (0.508mm), 0.030&quot; (0.762mm), 0.060&quot; (1.524mm)</td>
<td>12&quot; X 18&quot; (305 X 457 mm)</td>
<td></td>
</tr>
<tr>
<td>Material clad with LaPro foil adds .0007&quot; (0.018mm) to dielectric thickness</td>
<td>24&quot; X 18&quot; (610 X 457 mm)</td>
<td></td>
</tr>
<tr>
<td>24&quot; X 36&quot; (610 X 915 mm)</td>
<td>48&quot; X 36&quot; (1.224 m X 915 mm)</td>
<td></td>
</tr>
<tr>
<td>0.04&quot; material in a panel size larger than 24&quot; X 18&quot; (610 X 457 mm)</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
SIEMENS

Silicon Tuning Diode BB 833

Features

- Extended frequency range up to 2.5 GHz; special design for use in TV-SAT indoor units
- High capacitance ratio

<table>
<thead>
<tr>
<th>Type</th>
<th>Ordering Code (tape and reel)</th>
<th>Pin Configuration 1</th>
<th>Pin Configuration 2</th>
<th>Marking</th>
<th>Package</th>
</tr>
</thead>
<tbody>
<tr>
<td>BB 833</td>
<td>Q62702-B628</td>
<td>C</td>
<td>A</td>
<td>white X</td>
<td>SOD-323</td>
</tr>
</tbody>
</table>

Maximum Ratings

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Values</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Reverse voltage</td>
<td>$V_r$</td>
<td>30</td>
<td>V</td>
</tr>
<tr>
<td>Reverse voltage ($R \geq 5 , k\Omega$)</td>
<td>$V_{IR}$</td>
<td>35</td>
<td>mA</td>
</tr>
<tr>
<td>Forward current</td>
<td>$I_F$</td>
<td>20</td>
<td>mA</td>
</tr>
<tr>
<td>Operating temperature range</td>
<td>$T_{op}$</td>
<td>$-55 \ldots +150$</td>
<td>°C</td>
</tr>
<tr>
<td>Storage temperature range</td>
<td>$T_{st}$</td>
<td>$-55 \ldots +150$</td>
<td>°C</td>
</tr>
</tbody>
</table>

Thermal Resistance

- Junction - ambient $R_{J, \infty} \leq 450 \, \Omega$ K/W

Semiconductor Group 1  04.96
**Electrical Characteristics**

at $T_A = 25 \, ^\circ C$, unless otherwise specified.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Values</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Reverse current</td>
<td>$I_R$</td>
<td>–</td>
<td>20 nA</td>
</tr>
<tr>
<td>$V_R = 30 , V$</td>
<td></td>
<td>–</td>
<td></td>
</tr>
<tr>
<td>$V_R = 30 , V$, $T_A = 85 , ^\circ C$</td>
<td></td>
<td>–</td>
<td></td>
</tr>
<tr>
<td>Diode capacitance</td>
<td>$C_T$</td>
<td>8.5 pF</td>
<td>9.3 pF</td>
</tr>
<tr>
<td>$f = 1 , MHz$, $V_R = 1 , V$</td>
<td></td>
<td>0.6 pF</td>
<td>0.75 pF</td>
</tr>
<tr>
<td>$V_R = 28 , V$</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Capacitance ratio</td>
<td>$C_{T1}$</td>
<td>11 pF</td>
<td>12.4 pF</td>
</tr>
<tr>
<td>$f = 1 , MHz$, $V_R = 1 , V$, 28 V</td>
<td>$C_{T20}$</td>
<td>–</td>
<td>–</td>
</tr>
<tr>
<td>Capacitance matching</td>
<td>$\Delta C_T$</td>
<td>–</td>
<td>–</td>
</tr>
<tr>
<td>$f = 1 , MHz$, $V_R = 1 , V$ ... 28 V</td>
<td>$C_T$</td>
<td>–</td>
<td>–</td>
</tr>
<tr>
<td>Series resistance</td>
<td>$R_s$</td>
<td>–</td>
<td>1.8 Ω</td>
</tr>
<tr>
<td>$C_T = 9 , \mu F$, $f = 470 , MHz$</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Series inductance</td>
<td>$L_s$</td>
<td>–</td>
<td>–</td>
</tr>
</tbody>
</table>

**Diode capacitance** $C_T = f (V_R)$

\[ f = 1 \, MHz \]
APPENDIX C

RADANT RMSW101 SPST RF-MEMS SWITCH
The RMSW101™ is a Single Pole Single Throw (SPST) Reflective RF Switch utilizing Radant’s break-through MEMS technology that delivers high linearity, high isolation and low insertion loss in a chip-scale package configuration.

This device is ideally suited for use in many applications such as RF and microwave multi-throw switching, radar beam steering antennas, phase shifters, RF test instrumentation, ATE, cellular, and broadband wireless access.

### Features
- Low Insertion Loss (0.24 dB typical @ 2.4 GHz)
- High Isolation (27 dB typical @ 2.4 GHz)
- Near Zero Harmonic Distortion
- No Quiescent Power Dissipation
- Long Life (typical lifetime >100 billion cycles @ 30 dBm, >1 billion cycles @ 33 dBm)
- Hermetically sealed die designed for die-attach and wire-bond to board. Please contact us for other packaging options.

### Description

### Typical Device Specifications

<table>
<thead>
<tr>
<th>Insertion Loss</th>
<th>2 GHz</th>
<th>4 GHz</th>
<th>10 GHz</th>
<th>Lifecycle</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC</td>
<td>&lt; 0.23 dB</td>
<td>&lt; 0.26 dB</td>
<td>&lt; 0.32 dB</td>
<td>Cold-switched, 30 dBm</td>
</tr>
<tr>
<td>2 GHz</td>
<td>&lt; 26 dB</td>
<td>&gt; 21 dB</td>
<td>&gt; 12 dB</td>
<td>Cold-switched, 35 dBm</td>
</tr>
<tr>
<td>4 GHz</td>
<td>&gt; 26 dB</td>
<td>&gt; 21 dB</td>
<td>&gt; 12 dB</td>
<td>Cold-switched, 36 dBm</td>
</tr>
<tr>
<td>10 GHz</td>
<td>&gt; 26 dB</td>
<td>&gt; 21 dB</td>
<td>&gt; 12 dB</td>
<td>Hot-switched, 10 dBm</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>Hot-switched, 20 dBm</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>&gt; 10^11 cycles</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>&gt; 10^10 cycles</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>&gt; 10^9 cycles</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>&gt; 10^8 cycles</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Isolation</th>
<th>DC</th>
<th>2 GHz</th>
<th>4 GHz</th>
<th>10 GHz</th>
<th>Control</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC</td>
<td>&gt; 1 GΩ</td>
<td>&gt; 26 dB</td>
<td>&gt; 21 dB</td>
<td>&gt; 12 dB</td>
<td>Gate-Source Voltage (on)</td>
</tr>
<tr>
<td>2 GHz</td>
<td>&gt; 26 dB</td>
<td></td>
<td></td>
<td></td>
<td>Gate-Source Voltage (off)</td>
</tr>
<tr>
<td>4 GHz</td>
<td>&gt; 21 dB</td>
<td></td>
<td></td>
<td></td>
<td>Control Power, steady-state</td>
</tr>
<tr>
<td>10 GHz</td>
<td>&gt; 12 dB</td>
<td></td>
<td></td>
<td></td>
<td>Control Power, 1 kHz cycle rate</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Return Loss</th>
<th>DC</th>
<th>2 GHz</th>
<th>4 GHz</th>
<th>10 GHz</th>
<th>Switching speed</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC</td>
<td></td>
<td>&gt; -25 dB</td>
<td>&lt; -22 dB</td>
<td>&lt; -20 dB</td>
<td>On</td>
</tr>
<tr>
<td>2 GHz</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>Off</td>
</tr>
<tr>
<td>4 GHz</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10 GHz</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Input IP3</th>
<th>DC</th>
<th>900 MHz and 900 MHz up to +5 dBm</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>&gt; 65 dB</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Operating temperature</th>
<th>Maximum</th>
<th>Minimum</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>85 °C</td>
<td>-40 °C</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Storage temperature</th>
<th>Maximum</th>
<th>Minimum</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>150 °C</td>
<td>-55 °C</td>
</tr>
</tbody>
</table>

Notes:
1. All RF measurements were made in a 50 Ω system.
2. Measurements include bond-wires from die to test-board.
SPST RF-MEMS Switch, DC to 12 GHz  
RMSW101™

**Typical RF Performance**

![Graph showing RF performance with frequency and isolation/return loss on the y-axis and frequency on the x-axis.]

*Measurement results include bond wires*

**Absolute Maximum Ratings**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum Temperature</td>
<td>290°C</td>
</tr>
<tr>
<td>(10 seconds)</td>
<td>250°C</td>
</tr>
<tr>
<td>(120 seconds)</td>
<td></td>
</tr>
<tr>
<td>Maximum Voltage, Gate-Source</td>
<td>±/+ 10 V</td>
</tr>
<tr>
<td>Maximum Voltage, Drain-Source</td>
<td>±/+ 100 V</td>
</tr>
</tbody>
</table>

**Recommended Application**

1. Resistors RS and RD (40 kΩ-100 kΩ) or inductors LS and LD should be used to provide a path to DC Ground from Source and Drain.
2. VG may be of either polarity.
3. VG rise-time should be at least 10 μs for optimal lifetime.
4. Please refer to "Application Note for Test and Handling of SPST RF-MEMS Switches" for more information. Contact us for driver solutions.

- Phone: 978-562-3866
- Fax: 978-562-6277
- Email: sales@radantmems.com
- Visit www.radantmems.com
**Nominal Device Dimensions**

Dimensions are in micrometers. Please contact us for a footprint in .gds or .dxf format.

**Static Sensitivity**

This device has an ESD (HBM) sensitivity of 100 V. Use proper ESD precautions when handling. Please refer to "Application Note for Test and Handling of SPST RF-MEMS Switches" for more information.

**Die Assembly**

The gold backside-metallization on the die is designed to be mounted with electrically conductive silver epoxy, or with a lower temperature solder which does not consume gold. Bond pads on the die are made of gold. Ball-bonds should be utilized to attach gold or Aluminum 1 mil wires. Please refer to "Application Note for Test and Handling of SPST RF-MEMS Switches" for more information.

---

- Phone: 978-562-3866
- Fax: 978-562-6277
- Email: sales@radantmems.com
- Visit: www.radantmems.com