

Investigations into Impedance Tuners for the Next Generation of Communication Systems

by

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Author's Declaration

I hereby declare that I am the sole author of this thesis. This is a true copy of the thesis, including any required final revisions, as accepted by my examiners.

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Abstract

Novel communication systems, including 5G and the upcoming 6G cellular networks and software defined/cognitive radio technologies, call for compact, multi-band, and reconfigurable circuit components. Specifically, stringent requirements of 5G networks in terms of high data rates and spectrum efficiency have paved the way for the widespread incorporation of full-duplex radios, multi-band Power Amplifiers (PAs), and multiband antennas in communication systems. Simultaneously, the transceivers need to be tunable so that they can cover the numerous bands of the wireless networks. The implication in research is to push the existing fabrication technologies to their performance extremes, high precision tuning techniques, and to develop and devise reconfigurable subcomponents.

Impedance tuners are integral parts of any communication systems. Their ability to regulate the power transfer between different components has led to novel applications for these components in full-duplex front-ends. This includes suppressing the self-interference problem by improving the isolation between the transmitter (Tx) and receiver (Rx). Impedance tuners are also required to address the antenna impedance variation over the frequency and due to the variable environmental conditions. Moreover, an increasing number of multiband components such as antennas, PAs, power dividers, and baluns, are gaining momentum in communication systems. This highlights the need for multiband matching networks, which should simultaneously address the variable frequency bands (shift in frequency) and variable impedance of the different components. The requirements for such multiband impedance tuners include ability to handle variable frequency bands, wide impedance coverage at each frequency band, independent operation at each band, low loss, compact, and high power performance. The presented thesis is focused on novel applications for the impedance tuners. It presents a detailed analysis of doubly terminated impedance tuners and a novel approach for designing fixed and tunable dual-band matching networks using RF filters and phase-shifters. The intended frequency of operation is Sub-6 GHz frequency region, yet the discussions are kept applicable to other frequency ranges.

The thesis first presents a detailed analysis of doubly terminated impedance tuners in the form of tunable loaded-lines and π - networks. The intention here is to address the stringent requirements for the full-duplex transceiver. This includes improving the matching between the impedance-varying antenna and the circulator, as well as the isolation between circulator ports. Accordingly, Radio-Frequency Micro-Electro-Mechanical System (RF MEMS) packaged switches are used in a distributed π -network

matching network. Accurate modeling of the RF switch parasitic elements, as well as inclusion of the electrical length of the switches in the circuit design stage allows for low-loss and high precision switched matching networks. A doubly terminated prototype for achieving a 30 dB of matching at both ends is designed and tested at 2 GHz.

Moreover, Barium Strontium Titanate (BST) variable capacitors are employed in a CPW loaded line configuration to achieve at least 40 dB of matching between a variable antenna and a circulator with return losses of 15 dB and 18 dB, respectively. A prototype is designed and tested for operation between 3.4 GHz and 3.7 GHz frequencies. Accordingly, a detailed model for connecting Transmission Line (TL) segments is developed and optimized. The measurement results demonstrate that the designed matching network can achieve 40 dB matching over at least 20 MHz bandwidths over the band of 3.4-3.7 GHz with better than 0.6 dB of insertion loss.

In addition, a Reconfigurable Impedance Tuner (RIT) for full-duplex transceiver front-ends is presented. The RIT employs multi-port Phase Change Material (PCM) switches to improve the isolation between the circulator ports. These switches are extremely and have extremely low insertion loss over Sub-6 GHz frequencies. This allows for achieving compact, fully integrated, passive, and low loss RITs, where the main line is loaded by switched high quality factor capacitors. Employing 6 PCM switches allows for more than 2.6×10^5 switching states and a high precision uniform coverage. This allows for achieving circulator isolation of more than 40 dB over the 3.4-3.7 GHz frequency band, while negligibly affecting the matching.

In the second part of the thesis, a novel approach for designing multi-band matching networks is investigated. The novel concept is based on cascade configurations of filters and phase shifters to develop fixed and tunable dual-band matching networks. Two types of the fixed Dual-band Cascade Matching Network (DBCMN) are designed and developed. The first type of DBCMN uses individual filters and matching networks for each frequency band. This offers a simple and straightforward design procedure. The first type is demonstrated through a design example, which employs a Low Pass Filter (LPF), a High Pass Filter (HPF), and two microstrip-line phase-shifters. The second type combines the two filters and uses a dual-band phase-shifter. For this purpose, a Band Pass Filter (BPF) and dual-band Reflection Type Phase Shifter (RTPS) are used. The two types are compared and a few guidelines for selecting the appropriate type of the DBCMN according to the application are presented. The

advantages of the proposed DBCMN includes miniaturization, low loss performance, and independent matching of the two bands.

The most significant contribution of this thesis is the development of a design and implementation methodology for Dual-band Reconfigurable Impedance Matching Networks (DB-RIMNs). To our knowledge, this is the first time a tunable solution is offered for the cases where the impedance of the dual-band load and/or the intended matching frequency of the two bands are varied. First, the requirements for the proposed method are presented. This is followed by developing single band and dual-band tunable phase shifters. Then, circuit typologies for the proposed DB-RIMN are presented. Finally, the proposed DB-RIMN is implemented using tunable band pass and band stop filters, based on the frequency ratio of the two bands.

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Dedication

To Sara and Dario ...

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Chapter 1

Introduction, Motivation and Outline

1.1 Motivation

The fifth generation (5G) and the upcoming sixth generation (6G) of wireless networks employ a handful of novel subcomponents as the de-facto blocks for the transceivers and front-end modules. This includes beam steerable phased arrays, multiband Power Amplifiers (PAs), Self-Interference Cancelers, which have captured the interest of researchers due to their individual set of unique challenges. With an ever-increasing deployment of 5G networks across the globe, the need for addressing the stringent requirements of these components in terms of spectrum efficiency and coverage, reconfigurable performance, and multi-band operation is felt more than ever. On the other hand, novel telecommunication systems face a vast variety of challenges, including but not limited to, the dynamic environmental conditions, large operation bandwidths, integration possibilities, number of operational channels, and maintaining performance over different frequency bands. Therefore, there is an increasing demand for improving the performance of the existing components as well as introducing new methods to overcome these challenges.

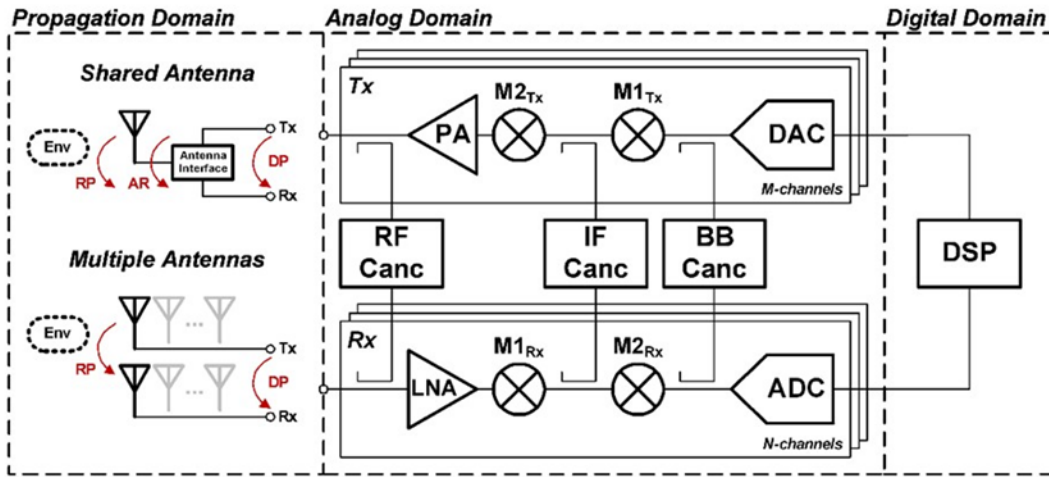


Figure 1-1 Full-duplex transceiver architecture highlighting three main domains for suppressing the interference signals [1].

As fundamental RF components, impedance tuners are two-port tunable devices that transfer the impedance of a termination into a desired impedance seen from the input. These components have a wide range of applications, including PA design, transistor characterization (i.e. source-pull and load-pull applications), and antenna impedance matching [2]. Over the recent years, the problem of matching has become increasingly more sophisticated than providing matching between two constant impedances. This

is associated with wide band operation of wireless systems and the fact that the input impedances of sources and loads do not remain constant over wide frequency bandwidths. Moreover, the stringent interference cancellation requirements of full-duplex systems that use circulators in the front-ends has been addressed by using impedance matching networks. Figure 1-1 depicts a generic full-duplex transceiver. It has been recently demonstrated that adding impedance tuners to the antenna interface can help in enhancing the cancellation between the transmit and receive channels in single antenna full duplex systems [3].

The 5G Sub-6 GHz band has fueled the recent surge in multi-band PA designs, which leverage small size technologies (e.g. 14nm FinFET CMOS [4]), Doherty amplifiers [5], and pre-distortion techniques [6], to name a few. Since PAs are the main driver for choosing a fabrication technology, the research for other RF components has followed case and numerous proposals have been put forward for multiband components, which range from multiband uplink transceivers [7] to quadruple band patch antennas [8].

Introduction of novel tunable devices such as Phase-Change Material (PCM) switches [9], as well as improving the frequency range of operation for technologies such as Barium Strontium Titanate (BST) varactors [10], offer superior performance in comparison with semiconductor tunable devices, which suffer from high loss and poor linearity performance. This has been timely to develop new impedance tuners for doubly terminated and dual-band applications for the next generation of wireless networks.

1.2 Objectives

The objectives of this research is to investigate the feasibility and limitations of approaching impedance tuners from a novel perspective of using filters to realize tunable matching networks with wide coverage of impedances for multi-band applications and for use in interference cancelation applications. The objectives also includes investigating different tunable technologies and their implications for realizing the proposed networks at sub-6 GHz band. Namely, the major objectives are broken down into:

- Development of low-loss miniaturized impedance tuners for doubly terminated applications

This task includes employing circuit architectures for impedance tuners and modifying them in compatibility with tuning elements suitable for sub-6 GHz operation. This task addresses the need for impedance tuners for self-interference cancellation applications. A variety of tuning elements such as varactors diodes, semiconductor switches, RF MEMS switches, BST Varactors, and PCM switches are investigated and employed.

- Development of a theoretical background and prototyping for filter-based dual-band matching networks

This task includes investigating the combination of filters and phase shifters and their performance as impedance tuners. It also involves investigating different block diagram combinations to realise dual-band matching networks. Moreover, this task includes circuit level investigation of trade-offs and optimization of physical and electrical properties of circuit elements. In addition, it is intended to derive the criteria for the performance of different alternative architectures and to demonstrate the methodology through design, fabrication, and measurement of different circuit prototypes.

- Development and demonstration of tunable dual-band matching networks

This task is the successor of the previous task and serves as the most innovative aspect of this thesis. The intention is to develop a set of comprehensive requirements for making the filter-based dual-band matching networks tunable. This objective also involves the development of suitable tunable sub-components such as tunable phase shifters, tunable band pass filters, and tunable band stop filters.

1.3 Thesis Outline

Chapter 1 of this thesis describes the motivation and objectives. Chapter 2 provides a brief survey of current research about impedance tuners and matching networks, identifies the frontline of research, and existing gaps in the research. Chapter 3 includes doubly terminated impedance tuners, which includes employment of MEMS switches for realizing switched matching networks, as well as developing distributed impedance tuner circuits for BST and PCM tuning devices for improving the effective isolation of circulators. Chapter 4 presents the novel methodology for designing dual-band reconfigurable impedance matching networks (DB-RIMNs) based on filters and phase shifters. This chapter also includes multiple realizations of the method for dual-band loads. Chapter 5 focuses on tunable DB-RIMNs and includes discussions about the choices of the filters and their implications on the frequency ratio between the two bands. This chapter also includes design and implementation of single-band and dual-band phase shifters. Finally, Chapter 6 summarizes the contributions achieved by this thesis.

Chapter 2

Literature Survey

This chapter is dedicated to review the current demand for impedance tuners and matching networks. The intention is to present an overarching overview of the current applications and identify existing gaps in the methods and research implementations. This requires covering a wide range of topics including novel applications of impedance tuners on a system level, circuit level solutions for realizing multi-band matching networks, as well as component level study of tuning elements and phase shifters.

2.1 Doubly Terminated Impedance Matching Networks

Pozar [11] counts three important roles for the impedance tuners and matching networks. First, they are essential for the delivery of maximum power between a load and a matched source. Secondly, they improve signal-to-noise ratio by matching receiver components such as antennas and low-noise amplifiers. Finally, they reduce phase and magnitude errors in distribution networks such as antenna arrays. Importantly, maximum power transfer is based on the assumption that the source is matched to a transmission line with characteristic impedance of Z_0 . This reference impedance simplifies the power transfer problem because a lossless matching network then needs to provide complex conjugate matching only on the load side. This singly-terminated problem [12] is the subject of academic textbooks and can be achieved using lumped elements networks, single stub, or double stub distributed networks [11], [13]. However, in practice the perfectly matched condition for the source is rarely satisfied because of the limited quality of circuit components as well as variation of termination impedances over frequency. Accordingly, the case where both source and load impedances represent some level of mismatch is known as doubly terminated networks. At low frequencies, and specially for real impedances, this problem is resolved using LC ladders (e.g. see [14]). The doubly terminated networks with real terminations are also extensively studied under microwave filter field [13].

In RF applications, both source and load impedances are often complex. A common practice for designing matching networks then involves transforming these impedances into real impedances so that they can be matched using existing doubly terminated networks. This can be done using two shunt reactive impedances to cancel the imaginary part of the source and load impedances. In [15], a π -network composed of two tunable shunt capacitors and a series fixed inductor is used as a matching network. Note that the combination of series inductors and shunt capacitors resembles the Low-Pass Filter (LPF) topology. Therefore, the π -network is interchangeably referred to as the low-pass matching network. The coverage achieved by this method is confined by the use of fixed inductor, which leads to a forbidden region on the

Smith chart. The impedances over this region cannot be matched using π - networks. An example of this forbidden region is shown in Figure 2-2.

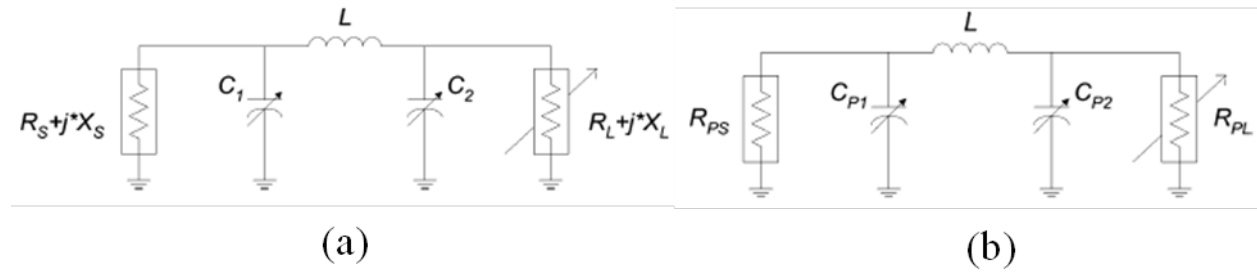


Figure 2-1 Transferring the complex source and load impedances into real impedances, (a) a π -network between the source and load, (b) equivalent circuit where the imaginary parts are compensate by the capacitors [15].

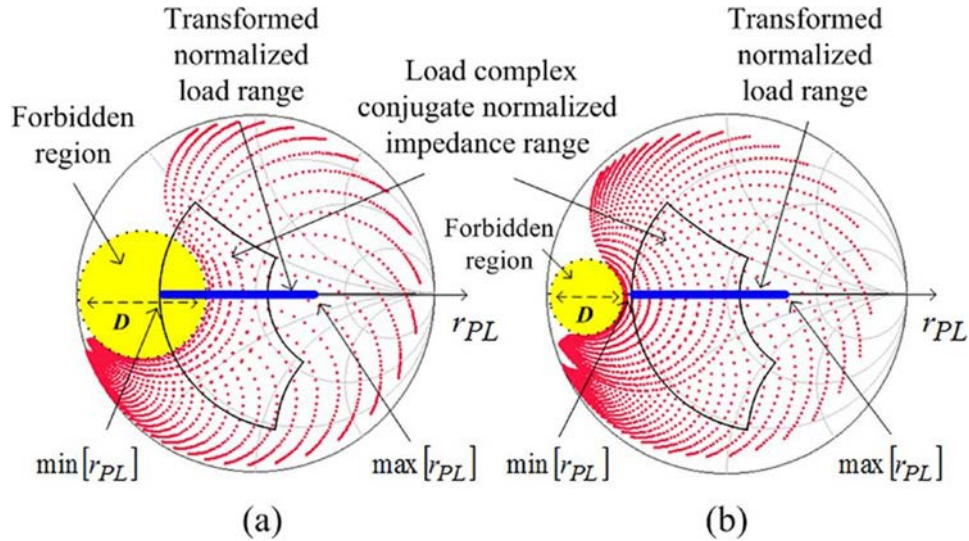


Figure 2-2 The forbidden region for a π - network (a) using a 2.5 nH inductor and (b) using a 1.7 nH inductor [15].

The distributed circuits for L-, π -, and T-network are realised using a series and shunt TL segment, two shunt TL segments and a series segment in between, and a shunt TL segment between two series segments, respectively. The electrical length and characteristic impedance of each TL segment can be modified as an independent degree of freedom. Note that four degrees of freedom are required for the case where both source and load impedances need complex conjugate matching (i.e. two conditions for the real part of impedances and two conditions for the imaginary parts). As a result, a unique L-section matching network is available for any two complex source and load pairs, while alternative solutions are possible for π - and T- networks [16]–[19].

These extra degrees of freedom has inspired some researchers to envision complimentary applications for matching networks. Notably, π - and T- networks have been employed to achieve simultaneous impedance matching and phase shift. This is studied in [16], where the ability of fixed (i.e. not tunable/single state) distributed matching networks to produce arbitrary transmission phase shifts is studied in detail. This includes simple structures such as a straight transmission line, as well as complex networks like asymmetric π - networks. Importantly, it identifies the fabrication limitation and assumes the characteristic impedance range of 20Ω to 120Ω as realizable. Correspondingly, the forbidden region for a given distributed circuit would expand to the cases that cannot be covered using realizable characteristic impedances. The range of forbidden impedances can be significantly improved by employing an additional stub in parallel with the load impedance or a series TL segment. Such elements can transfer a forbidden impedance into the allowed region. Thus, they are called load healers [20]. A comprehensive investigations of the allowed/forbidden regions and matching network coverage in general can be found in [21]–[23]. In the following section, recent techniques for realizing tunable impedance matching networks are briefly reviewed and the current shortcomings are identified.

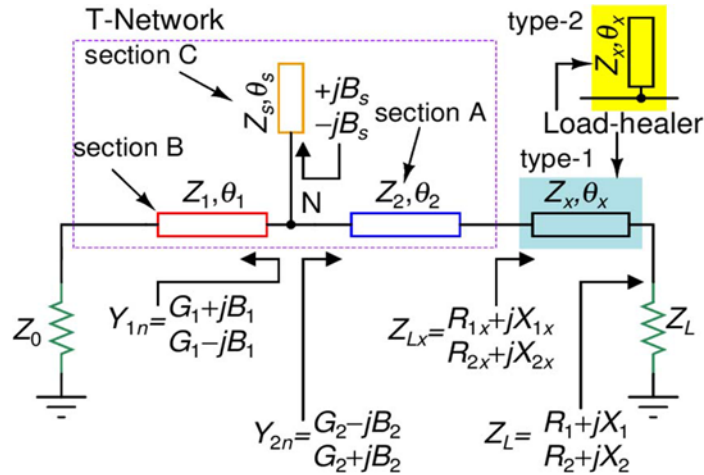


Figure 2-3 Incorporating load-healers in a conventional T-network matching design [20].

2.2 Reconfigurable Matching Networks

There are two major methods of realizing reconfigurable circuits. This includes employing continuously variable capacitors such as varactors and BST capacitors, and using RF switches such as PIN diodes and MEMS switches. In [24], reconfigurable matching networks are realized by employing varactors, which are continuously tunable, and PIN diodes, which are switchable. Note that it is possible to replace the series TL segment in π -networks with a T-network to mimic tunable inductor performance [25], [26]. These circuits are shown in Figure 2-4 and Figure 2-5. However, there are a handful of gaps in the presented

researches. Firstly, the impedance coverage achieved by this method is not studied. Therefore, given that the first step in designing these matching networks involves calculating the characteristic impedance for the fixed portion of the device, it is not understood how much the fabrication constraints will affect the coverage. Secondly, the relationship between the varactor tuning range and coverage is not explored. Thirdly, the examples are presented at the centre frequency of 1 GHz, where varactors have high-quality factor, PIN diodes have small parasitic components, and the components of the bias networks are non-resonant. Lastly, at this frequency the electrical length of the connections between the tuning elements and the fixed circuit is negligible, but as the frequency increases the electrical length of these parts become significant. The significance of these sections is not addressed in the literature. As a result, it remains to examine the designs at higher frequencies, where the quality factor of tuning elements degrade, junctions become electrically large, and designing bias networks become challenging.

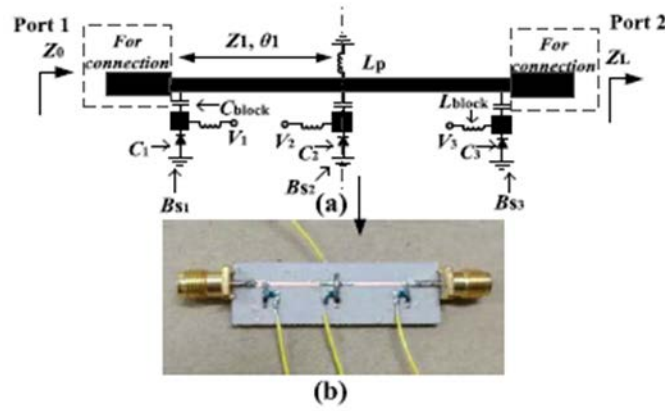


Figure 2-4 The tunable matching network using varactors, (a) circuit topology and (b) fabricated prototype [24].

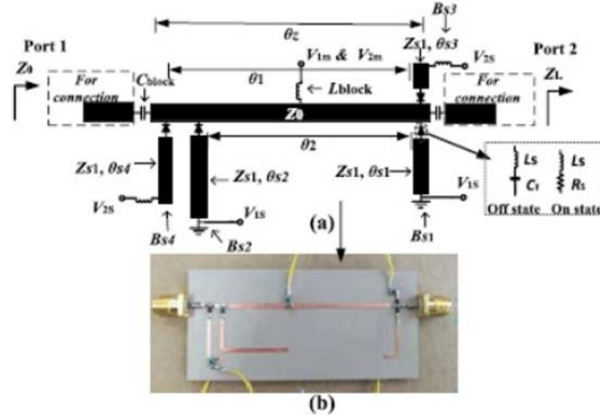


Figure 2-5 The switched matching network using PIN diodes, (a) circuit topology and (b) fabricated prototype [24].

The switched matching network presented in [24] is referred to as a loaded line. This is one of the four typical circuit topologies of tunable matching networks, including tunable π -network, loaded lines, coupler based tunable matching network, and circulator based matching networks. Choice of the proper topology depends on the application, frequency range, choice of tuning elements, bandwidth of operation, and acceptable insertion loss. For example, in [27], the switched implementation of these topologies, as illustrated in Figure 2-6, are compared in terms of the number of states required for achieving desired coverage and it was concluded that the circulator based matching networks achieve a more uniform coverage using the same number of switches. However, coupler based and circulator based matching networks are often much bulkier than the other counterparts. In addition, couplers are often narrowband and circulators may impose higher losses. Therefore, reconfigurable π -network and loaded lines are preferred. Note that, a practical method for realizing π -networks is replacing the series TL segment (i.e. the inductor in lumped circuit) with a T- section with 45° arms because realization of tunable inductors is challenging. An example of tunable inductors using MEMS switches is presented in [28]. According, as seen in Figure 2-4 [24], this implementation of π -network become a special case for the loaded line, where the impedances are applied at 45° intervals.

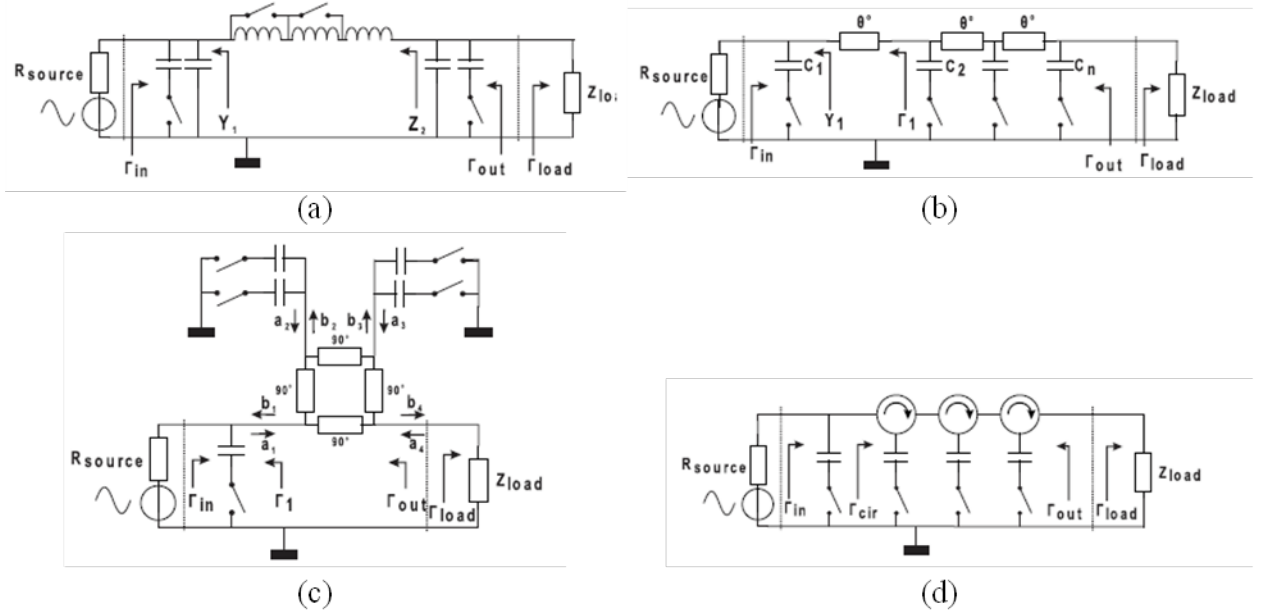


Figure 2-6 Switched matching networks (a) switched π -network, (b) Switched loaded line, (c) switched coupler based matching network (using branch-line coupler), (d) switched circulator based matching network [27].

2.3 Quest for Sub-6 GHz Reconfigurable Matching Networks

A comparison between early generations of RF-MEMS switches, semi-conductor switches, and PIN diodes has shown that the MEMS switches have superior performance in terms of capacitance at the off-state, resistance at on-state, significantly better switching ratio (i.e. the off/on ratio of insertion losses), and power consumption [29]. In [30], a comparison between RF MEMS switches and PIN diode based, FET based, Ferrite based, and MMIC based RF switches has shown that not only the results presented in [29] have been improved, but the performance of the MEMS switches in terms of metrics such as weight, size, insertion loss, isolation, and reliability of fabrication all have been improved. Specially, the demonstrated frequency range of operation from L-band to mm-wave bands makes these switches a viable option for realizing mm-wave impedance matching networks. The distributed loaded lines that use MEMS devices as tuning elements are referred as Distributed MEMS Transmission Line (DMTL) structure, which have applications in designing matching networks and phase shifters [31].

In [32], a unique approach to DMTLs is presented, where a straight CPW line is periodically loaded by defects in the ground plane. This is shown in Figure 2-7. The ground defects are equipped with MEMS switches. Therefore, when the MEMS switches are in on state, the ground defect is eliminated. In off state, each ground defect is equivalent to a parallel RLC component with a resistance of roughly 1k, inductance of roughly 47 pH, and capacitance of 22 pF. These small values help the circuit to achieve a considerable impedance coverage beyond 24 GHz. This DMTL matching network has an exceptionally low average insertion loss of about 0.2 dB over the frequency range between 20 GHz and 40 GHz. However, modifying this design for Sub-6 GHz applications is challenging because of the small equivalent circuit components associated with the ground defects.

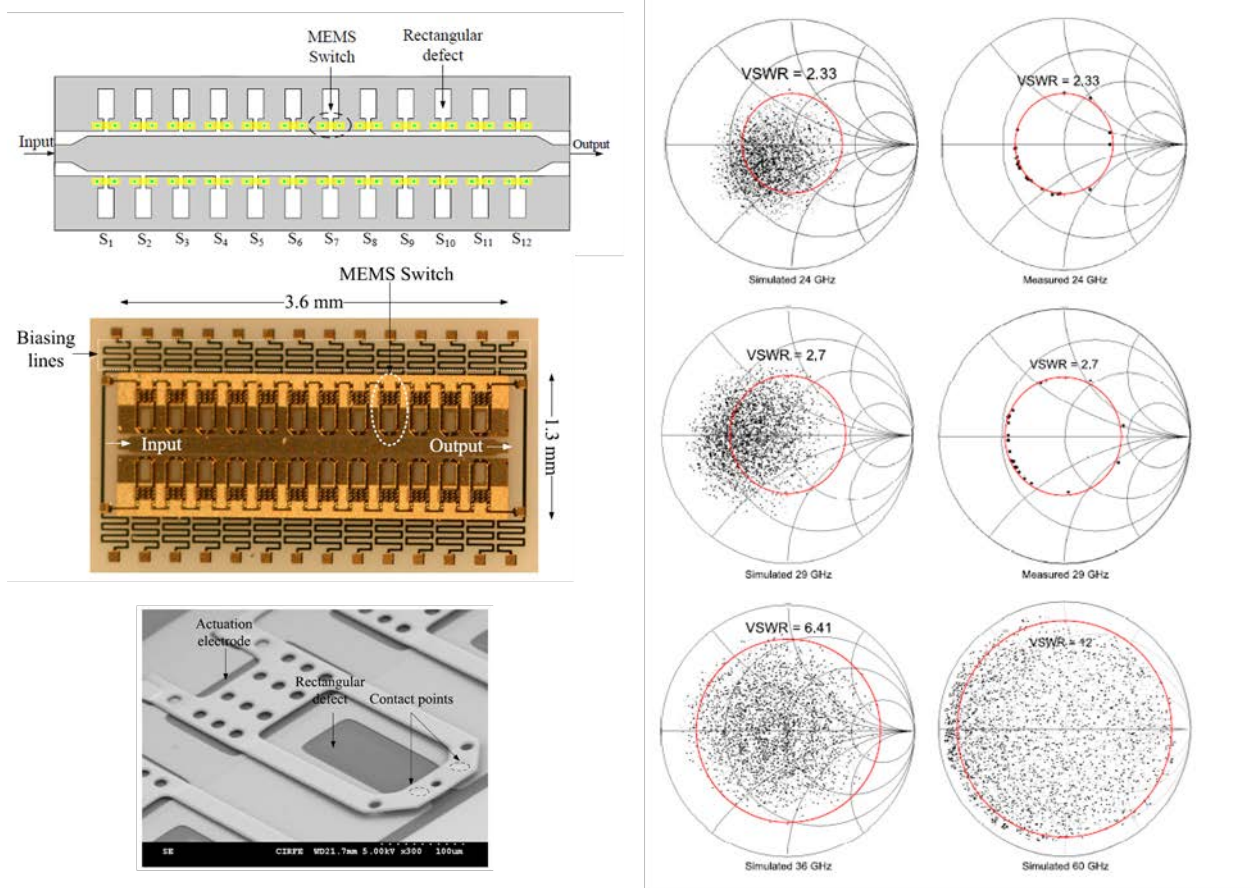


Figure 2-7 DMTL matching network with defected ground slots [32].

In [33], a miniaturized DMTL matching network is presented. In this work, a slow-wave CPW line is used as main TL for miniaturization. Moreover, tri-state RF MEMS switches are used to load the circuit. The fabricated circuit and the measured impedance coverage are shown in Figure 2-8. As can be seen, the use of slow-wave structures and higher loading by the MEMS switched capacitors has resulted in a coverage at lower frequencies compared to the previous DMTL, while maintaining a relatively similar size. However, the use of slow-wave structures as well as direct influence of the switches on the main line has resulted in significantly higher losses compared to the previous design. Moreover, the coverage at 5 GHz is significantly smaller than the higher frequencies. The DMTL matching network is further optimised in [34] to shrink its size by 25% and adapt it for low frequency applications. However, the impedance coverage is still small over the sub- 6GHz bands and the circuit is lossy because of the silicon substrate.

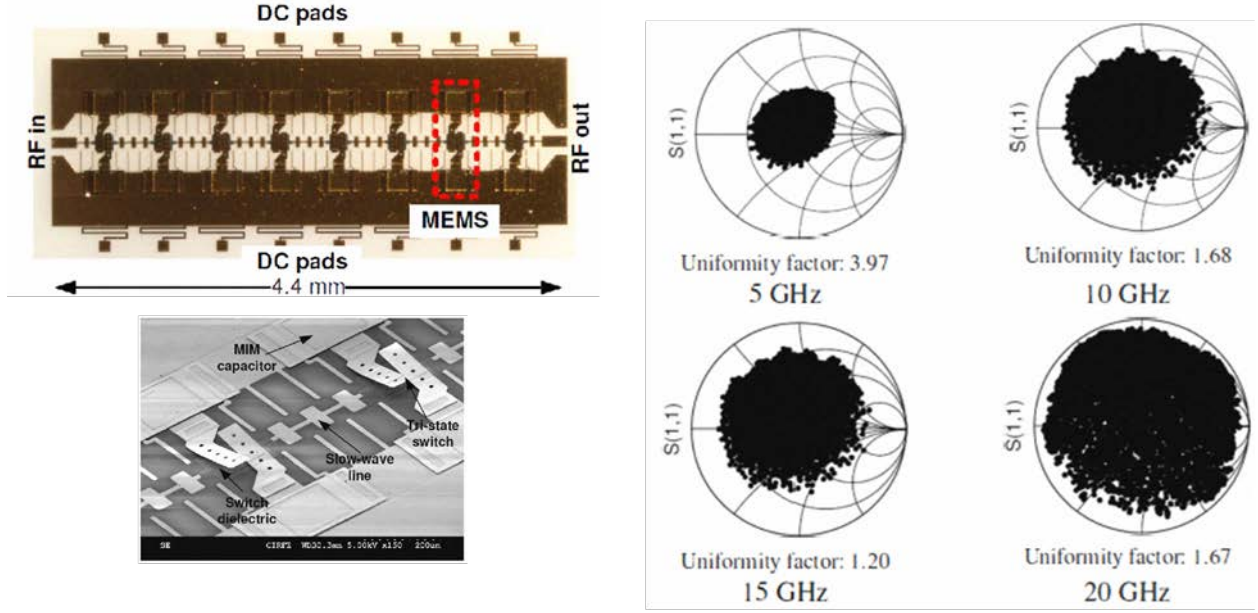


Figure 2-8 DMTL matching network using Dual-beam MEMS and slow-wave structures [33].

In [35], the lossy performance of the DMTL matching network is improved by removing the lossy substrate under the slow-wave CPW line. This is done by Ion Etching process and creating trenches below the devices, as shown in Figure 2-9. The resulting slow-wave structure has an effective dielectric constant between the effective dielectric constant of ordinary CPW line and suspended CPW line. Therefore, improving the loss performance has decreased the electrical length of the TL segments. Correspondingly, the coverage of this matching network has shifted to mm-wave region.

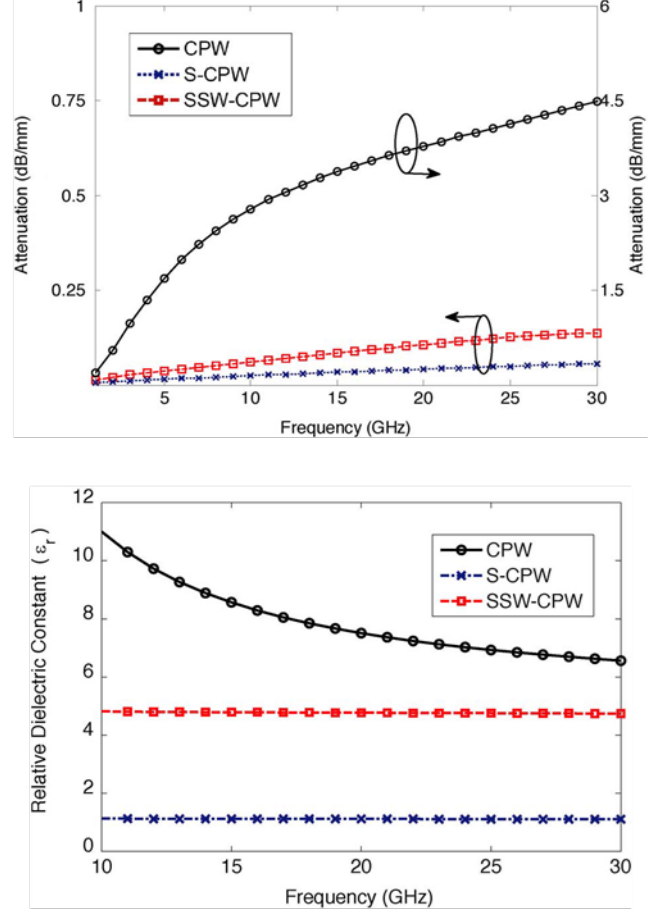
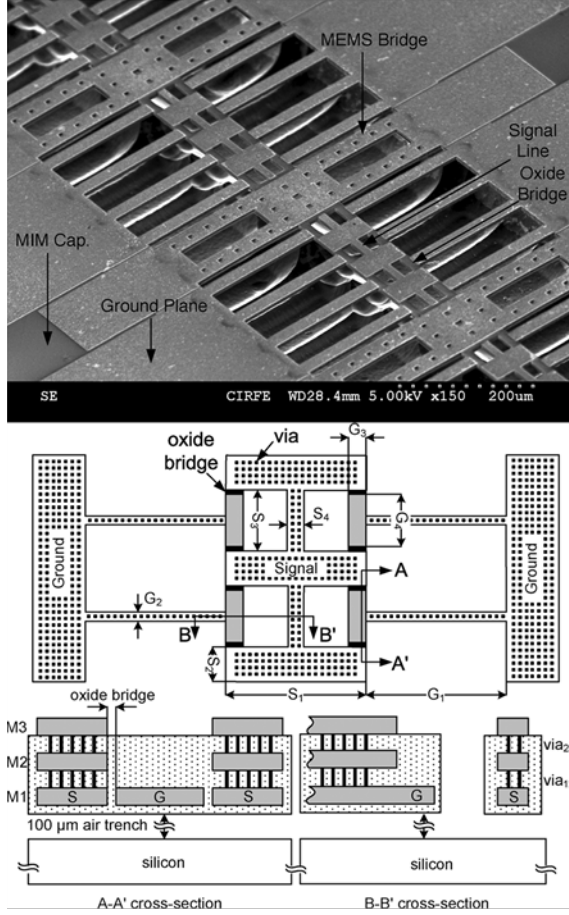


Figure 2-9 The suspended slow-wave DMTL matching network in a standard CMOS- MEMS [35].

In [36], a monolithic 4 tuning element DMTL matching network is presented which employs 70° TL segments. This work uses 4 MEMS varactors to load the main line. Each MEMS varactors is composed of a parallel plate capacitor and the gap between the plates is precisely controlled using a Chevron actuator. The individual varactors have a tuning range of 190fF- 0.8 pF. In addition, high resistivity silicon substrate is used to reduce the losses. This design has a wide impedance coverage at 6 GHz and an insertion loss of less than 1 dB over the Sub-6GHz region. This device and its performance are shown in Figure 2-10.

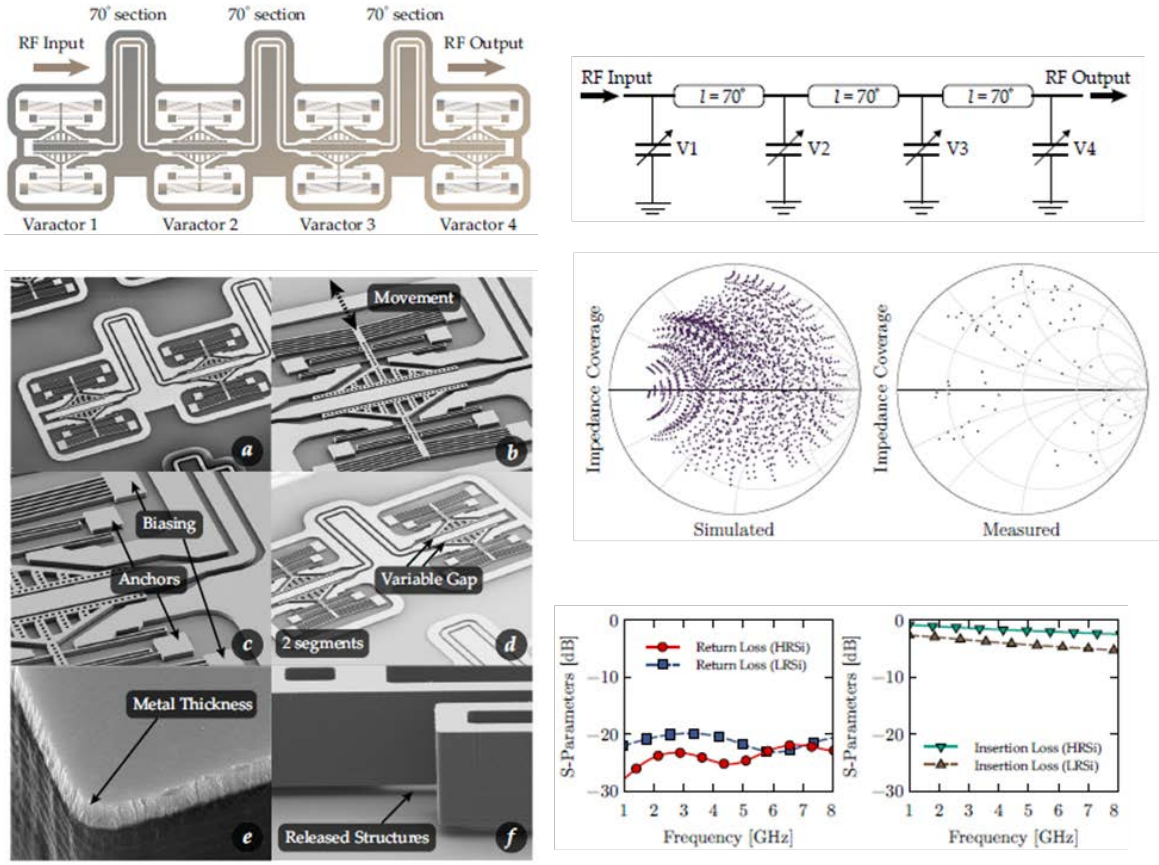


Figure 2-10 Fully integrated RF MEMS varactors matching network [36].

In addition to the integrated MEMS based matching networks that struggle to achieve coverage over Sub-6 GHz region, efforts are under way to expand the frequency range of matching network that rely on packaged tuning elements to cover the Sub-6 GHz region. The restraints for these devices is mainly due to the degradation of the performance of the tuning elements, lumped circuit elements, and distributed components with respect to frequency. The current state of the art commercially available packaged MEMS tunable devices are available for frequency ranges below 3 GHz (e.g. see products from wiSpry and Qorvo [37], [38]). A similar trend is observed for the other counterpart tunable devices such as BST tunable capacitors and semiconductor tunable capacitors (e.g. see products from OnSemi under the name Passive Tunable Integrated Circuits [39] and pSemi under the name Digital Tunable Capacitors [40]). This is limitation is because of the self resonance frequency of these devices, which occur over the higher end of the Sub-6 GHz band (e.g. refer to the S-parameters of the SMV1405 and SMV1232 varactors [41]).

In [42], a π -network tunable matching network is presented, which uses BST variable varactors. Moreover, off-chip discrete lumped inductors using Integrated Passive Device (IPD) are used to increase

the thickness of metal layers. The thick metal inductors have a higher quality factor compared to inductors realized using regular CMOS process and are used to improve the insertion loss performance of the matching network. The circuit, characterization of its inductor and measurement results for a few matched states is presented in Figure 2-11. The frequency range for this matching network is between 2 GHz and 3 GHz. The insertion loss for the presented cases is between 1dB and 1.3dB. As can be seen the quality factor of the inductor decreases as the frequency increases. Therefore, this design is not suitable for the higher end of the Sub-6 GHz band.

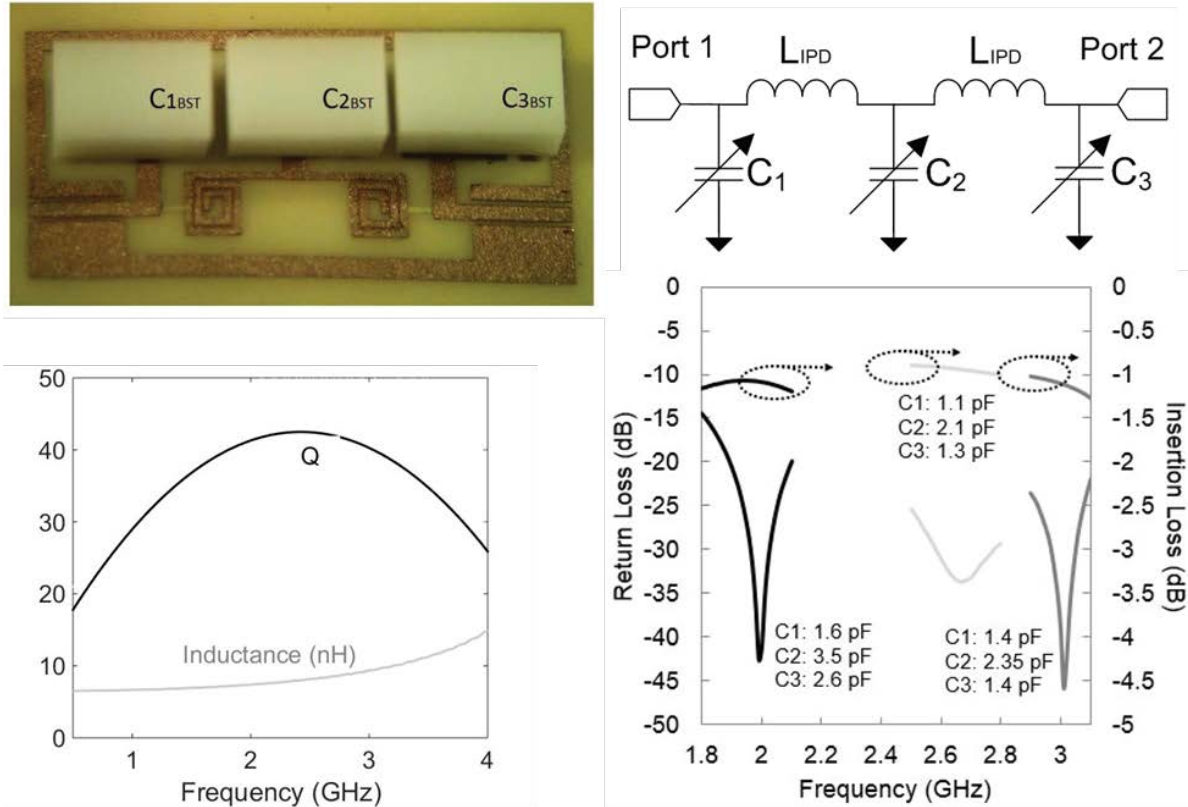


Figure 2-11 Matching network Using BST Varactors and thick metal discrete inductors.

The reviewed literature in this section suggest that fully integrated MEMS matching networks are suitable for frequencies beyond the Sub-6 GHz region because of the size constraints, fabrication limitation, and limited tuning range. On the other hand, lumped element matching networks, which employ packaged tuning devices, are not suitable for the higher frequencies of the band. A gap over the higher end of the Sub-6 GHz spectrum is identified. To the knowledge of the author, no tunable matching network has been reported for this band. The issues associated with this band was identified in this section. However, with surge of novel technologies such as Phase Change Material (PCM) switches [43], advances in RF MEMS switches (e.g. products from MenloMicro [44]) and improved performance of PCM variable capacitors

promise for hybrid solutions for filling this gap. Accordingly, a handful of solutions will be presented in the next chapter.

2.4 Self-interference Cancellation Using Impedance Tuners

Self-interference cancellation modules are an integral part of full duplex transceivers. Full-duplex wireless communication requires extreme levels of isolation between transmission paths (Tx) and receive paths (Rx) to ensure the operation of the receiver is not interfered by the leaked signal from Tx. One prominent source of interference is circulator leakage, which is a consequence of limited isolation between different ports of the circulator [45]. Therefore, a number of techniques have been presented to enhance the Tx/Rx isolation in single antenna front-ends [46], [47]. In [48], the cancellation is improved by 25 dB by adding a phase shifter between antenna and circulator. In [49], isolation was increased using antenna polarization was used to increase the isolation. Similarly, adding a matching network can improve the performance by up to 20 dB. As a common practice, antenna and circulators are described by the magnitude of S-parameters (i.e. insertion loss and return loss). Since these components are designed separately, matching networks are needed to ensure proper impedance matching between loads located on arbitrary locations on constant voltage standing wave ratio VSWR circles [48].

Recently, a method of increasing the isolation between Tx and Rx using an impedance tuner is presented [50]. As shown in Figure 2-12, the tuner is inserted between the antenna port and the circulator. The tuner in this case is not intended for maximum power transfer. Instead, it is used in conjunction with the S-parameters of the circulator to eliminate leaked signal from Tx(1) to Rx(2). This is done by tuning the parameters of the impedance tuner such that the input reflection coefficient of the impedance tuner, γ , is equal to $\frac{\beta^2 \alpha}{\alpha^2 - \Gamma \beta}$, where β , α , and Γ are isolation, attenuation, and reflection coefficients associated to the circulator, respectively. Applying this condition negligibly influences the power transfer from Tx to antenna and for antenna to Rx, but leads to significant improvement in the isolation [51]. This application will be further investigated in section 3.5 of this dissertation.

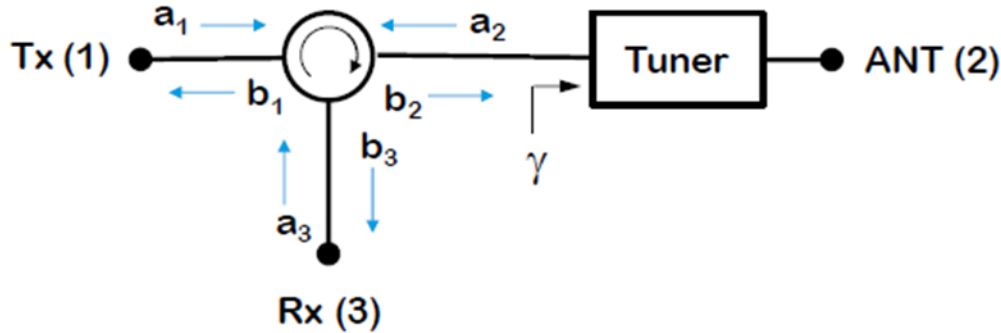


Figure 2-12 Improving the Tx/Rx isolation using an impedance tuner [3], [50].

In [3], a close loop control circuitry, without the need for additional Digital Signal Processing (DSP) for tuning the device, or couplers for isolating the paths, for controlling the impedance tuner is presented. The impedance tuner used is composed of a lumped component fixed π - network, a series and a shunt tunable BST varactors L-network, and an inductor load healer on the antenna side. This configuration achieves a 33 dB self-interference cancellation improvement at a 30 MHz bandwidth around 930 MHz frequency. The penalty is a 0.7 dB degradation in insertion loss performance. This design works at a single frequency and since it is based on lumped components, it is not suitable for Sub-6 GHz band. The circuit and its measurement results are shown in Figure 2-13.

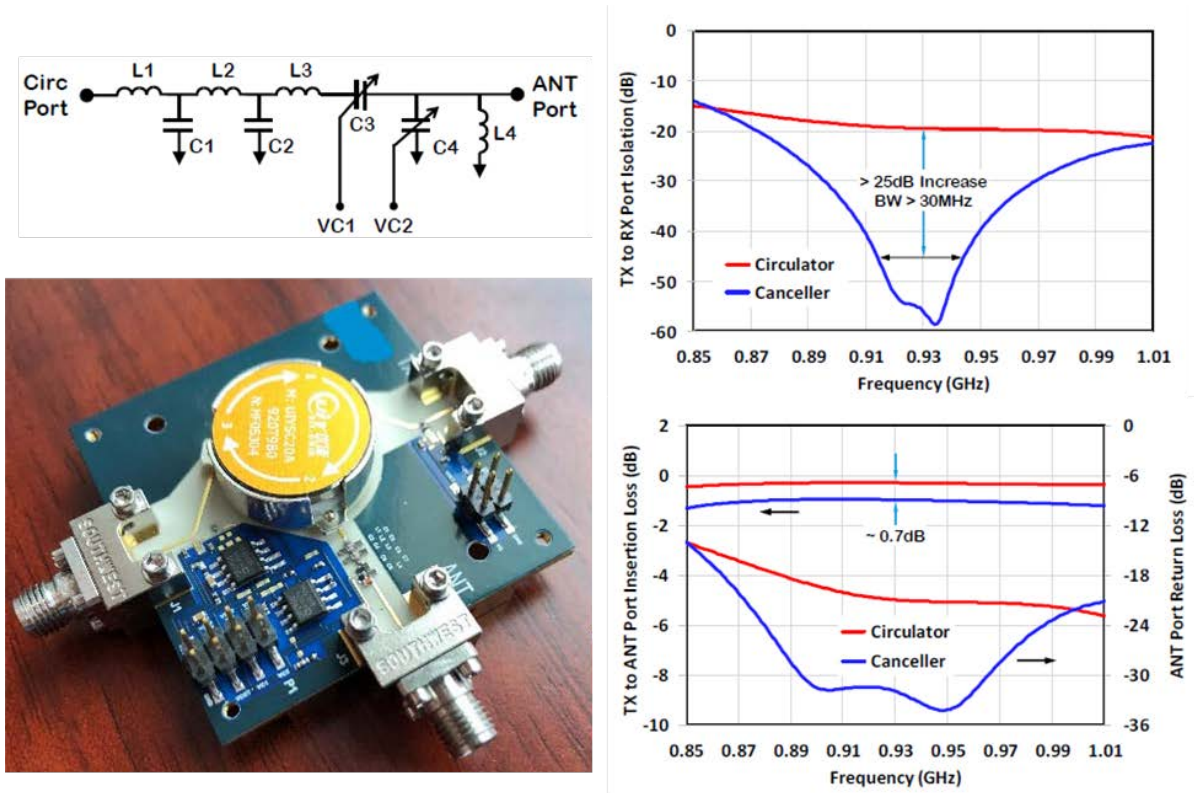


Figure 2-13 The BST varactor based circulator self-interference canceller [3].

In [51], an impedance tuner composed of two T-networks in cascade configuration is presented. This work is intended to improve the isolation for a circulator intended to operate at 2.5 GHz. A commercial antenna is used to demonstrate the performance of this design. The circuit topology, a device under test and its performance between 2.2 GHz and 2.8 GHz is shown in Figure 2-14. This design is intended to improve isolation for all antenna impedances with better than 10 dB return loss. However, such coverage is not demonstrated and the design methodology is ambiguous in a sense that it is not clear if a constant reflection-

coefficient is assumed for the circulator or if the design can be used for other antenna/circulator combinations. A demonstration method for the performance of impedance tuners to improve isolation is needed which shows the range of antenna impedances and circulator reflection coefficients covered by the tuner and the degree of improvement for a variety of cases. Moreover, this design is not suitable for the Sub-6 GHz region. Scaling this design for higher frequencies requires addressing the parasitic, circuit junctions, and the added lengths of the connections. Such gaps are addressed in section 3.5 of this work, where alternative designs suitable are presented and their performance is demonstrated for Sub-6 GHz applications.

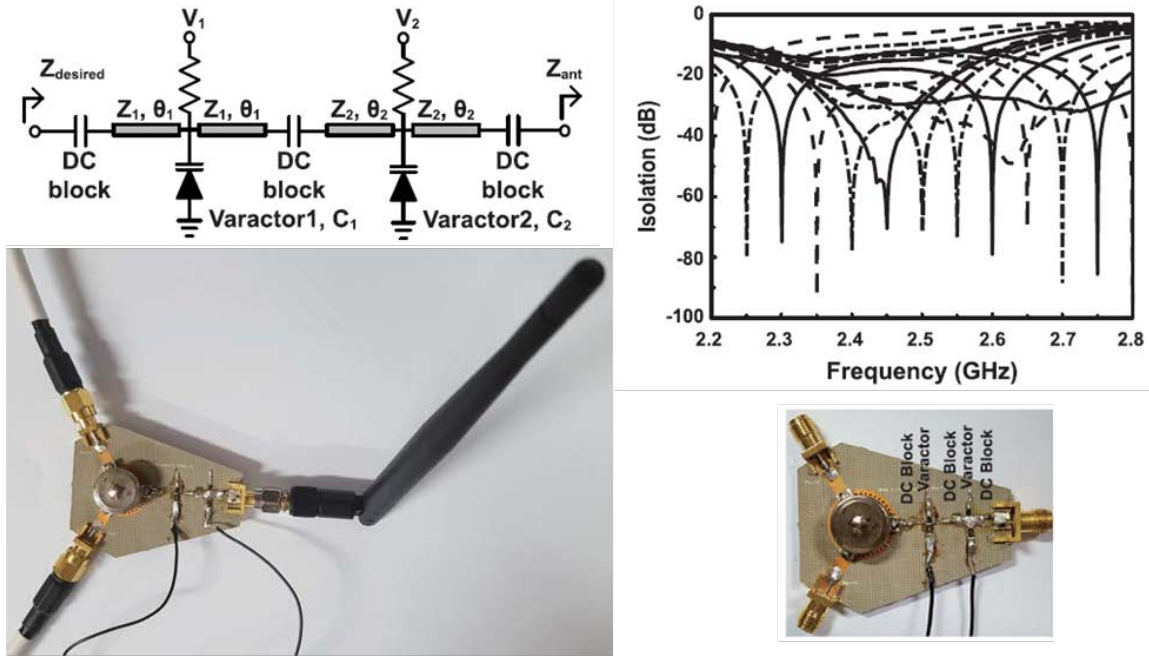


Figure 2-14 A reflection coefficient controller to enhance circulator isolator by 70 dB [51].

2.5 Dual-band Matching Networks

Similar to the doubly terminated designs, π -networks have been used in designing dual-band matching networks due to their degrees of freedom. In essence, a dual-band matching network problem requires 4 variables to address real and imaginary part of a load at two frequencies. In [52], a π -network is used in 2 steps to provide dual-band matching. This is shown in Figure 2-15. In first step, a shunt stub is used to cancel the imaginary part of load admittance at the two frequencies. This is done by modifying the impedance and electrical length of the stub. The input admittance seen after this stub is purely real and can be matched using an L-network. The difference between this design and the conventional double stub matching is that the series TL segment in the conventional double stub-matching network is a 50 Ω TL,

while here the characteristic impedance of this segment is also taken into account. An inherent issue with this solution is that it may require unrealizable characteristic impedances. For example, characteristic impedances as high as 200Ω are calculated in the design examples [52].

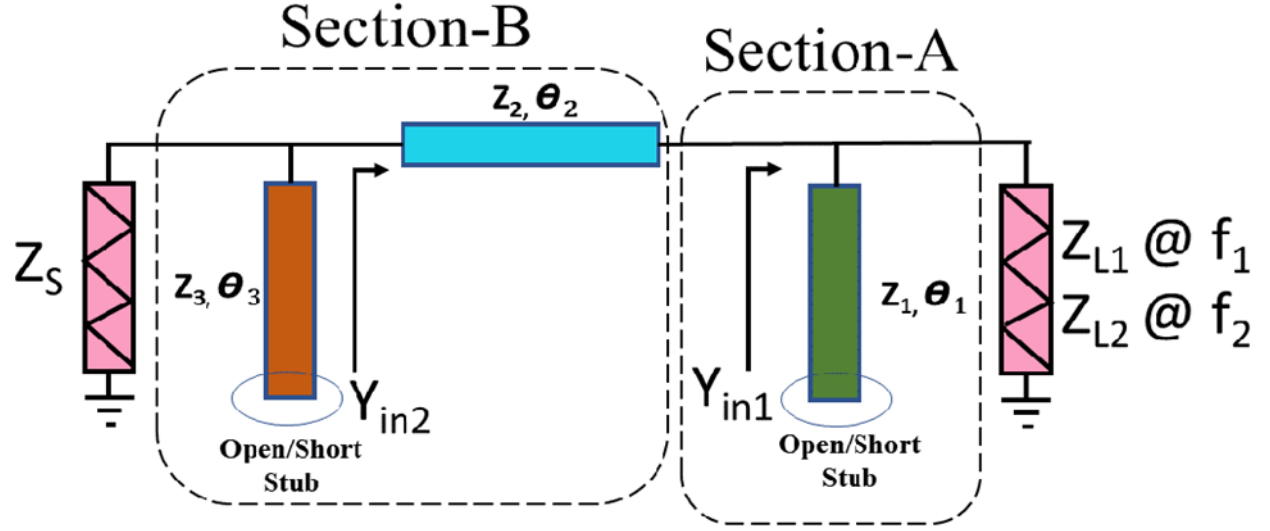


Figure 2-15 A two section π -network for dual-band impedance matching applications [52].

In [53], three coupled line wave slowing structures are used. This is depicted in Figure 2-16. Such structures are constructed by connecting one end of a conventional coupled line coupler together; use the other ends as input, and output ports. These components are commonly referred to as C-sections and have a range of applications including filter design and commensurate delay lines [54]. However, the use of this solution for dual band impedance matching is limited to only real impedances. Moreover, this method may require high even and odd impedances for the coupled line segments. Finally, although matched bandwidths as wide as 50% is reported for this design. However, as the bandwidth of the signals increase, the load impedances can no longer be assumed constant and vary over the band, which renders the wideband performance of this design.

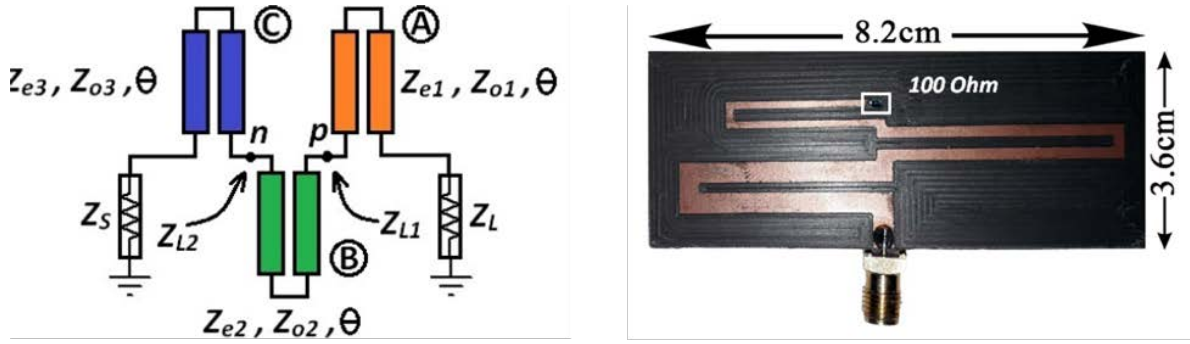


Figure 2-16 Enhancing the bandwidth of the dual-band matching network using coupled-line wave slowing [53].

Similarly, design equations for a three-section TL matching network were presented in [55]–[58]. The schematic of this dual-band impedance matching network is shown in Figure 2-17. The presented design equations are tedious and have remained subject of debate for 10 years. Briefly, the input impedances seen after the third section are complex conjugate of each other at the two frequencies. Then the next two sections transfer the conjugate impedance pair into the source impedance [59]. The matching for conjugate pairs using a two section transmission line is presented in [60]. Similar to the previous dual-band methods, this method may need high impedance TL segments.

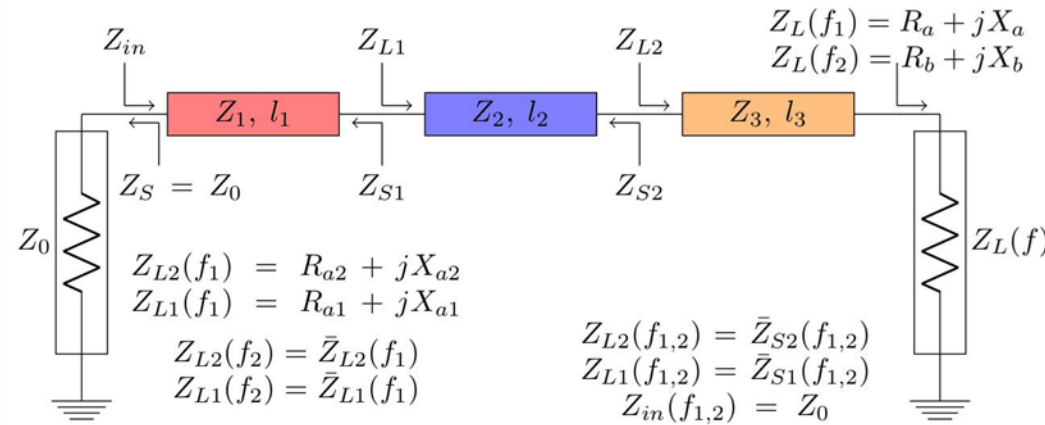


Figure 2-17 The three-section dual-band matching network [58].

More recent methods of designing dual-band matching networks employ a mix of lumped and TL elements in addition to increasing the number of segments. In [61], a double segment TL segment, a double stub resonator, and a L-network are used to achieve a dual-band impedance tuner. This is shown in Figure 2-18. The L-network is composed of a mixed of TL segments and high quality factor lumped components. Incorporation of the resonator has added two transmission zeros; one between the two frequencies and one above the higher frequencies. This design is capable of achieving dual band matching without the need of

high or low impedance TL segments. However, adding additional segments has significantly enlarged the size of the circuit.

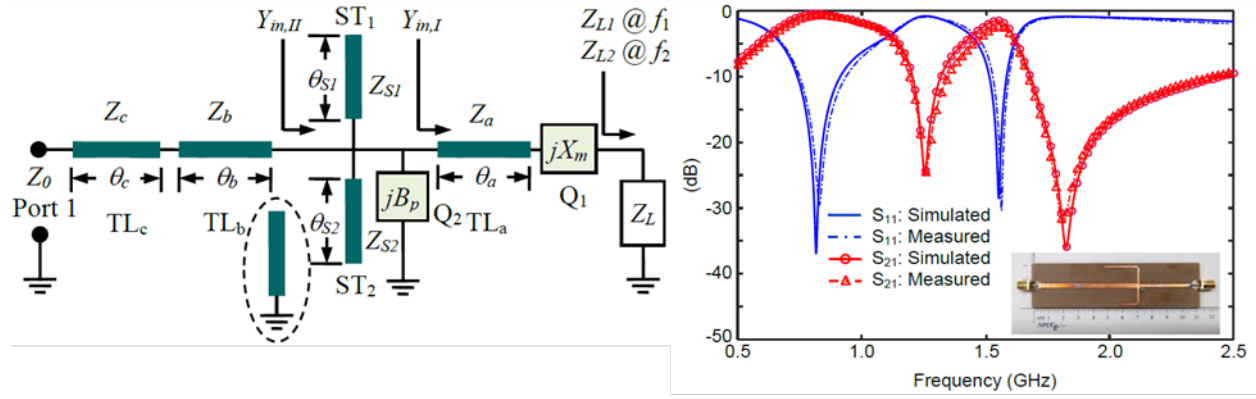


Figure 2-18 The mixed Impedance tuner with embedded transfer zeros [61].

Collectively, the above-mentioned methods are developed for simultaneous matching for specific frequencies and specific load impedances. Therefore, the entire design must be updated by each change in frequency or load impedances. Due to the limited degrees of freedom, these methods have limited use in designing multiband matching networks. More fundamentally, existing dual-band circuits are not suitable for tunability because of their inherent dependency on the characteristic impedance of the TL segments. Moreover, the authors are not aware of any prior works that are capable of match variable impedance dual-band loads. The same is true for the case where the two frequencies need to be tuned. To our knowledge, current possible solution may involve using dual-band reconfigurable filters and add shunt load heaters. Therefore, a demand for a novel methodology for designing dual-band impedance matching networks is identified. In chapter 4, a new methodology based on filters and phase shifters is presented. In chapter 5, arguments for making the novel method tunable in presented.

Chapter 3

Doubly Terminated Impedance Tuners

This chapter deals with novel applications for single-band reconfigurable impedance tuners. Such applications have the potential to be employed in full-duplex transceivers as a solution for the self-interference cancellation. First, the use of these devices to achieve matching between two independently variable impedances is studied. This problem is known as matching of doubly terminated networks. Second, an application for improving the isolation between transmit and receive paths is studied. For clarity, in the remainder of this work, the circuits that are intended for complex conjugate matching are referred to as Impedance Matching Networks (IMNs), while other circuits that modify input impedance/reflection coefficient are called Impedance Tuners (ITs). Note that IMNs are a category of ITs. Accordingly, reconfigurable IMN and ITs are referred to as RIMNs and RITs, respectively.

Full-duplex transceivers are capable of transmitting and receiving signals simultaneously and over the same frequency band, which enhance the spectrum efficiency of communication systems from the fifth generation onwards. However, since transmit signal is dramatically stronger than the received signal, any leakage of it into the receive path masks the received signal. This is known as self-interference challenge. In theory, self-interference can be suppressed using signal processing since, the masking interference signal is essentially known. However, to avoid ADC saturation, approximately 50-60 dB of self-interference cancellation is needed using the stages before the ADCs. This is achieved at various stages by a variety of techniques collectively known as analog and passive self-interference cancellation (in contrast with digital cancellation). In this chapter, two potential applications of RITs for analog interference cancellation are studied and implemented. This includes improving the matching between antenna and circulators, and using reconfigurable loaded lines to improve the Tx/Rx isolation for circulators, while providing matching between antenna and circulators. A variety of implementation techniques such as using switched length stub loaded line, and tuning elements including RF MEMS switches [26], BST varactors, and PCM switches are discussed and used.

3.1 Suppressing Self-interference Using RITs

In this section, we investigate the causes of self-interference associated with the use of the same band for transmit and receive circuits. Two general topologies of antennas are considered for the front-end of full-duplex RF transceivers. In the first topology, transmit and receive paths are separated and each path employs a separate antenna or antenna array. The self-interference cancellation in this type of transceivers has been studied extensively and can be implemented effectively by modifying the distance between the

two antennas, placing the antennas on the nulls of the radiation pattern of each other, as well as using opposing polarization. However, these techniques require at least two antennas that are physically separated from each other and large areas. A more space efficient topology employs a single antenna and uses a circulator for separating transmit from receive paths. For convenience, the studied scenario is limited to the latter topology.

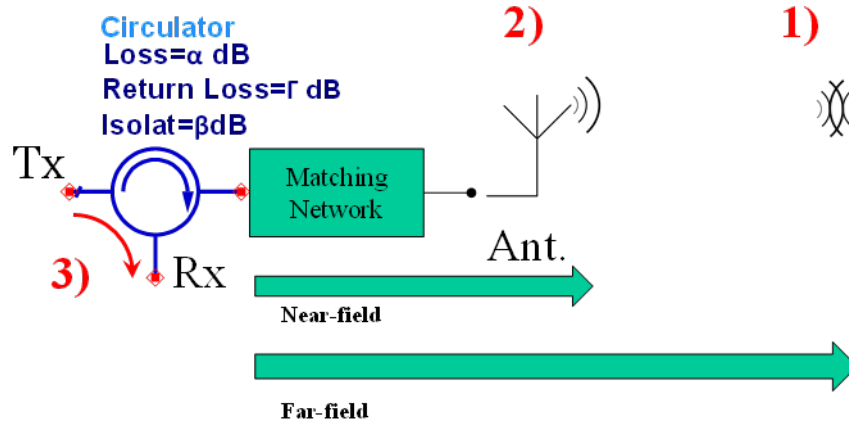


Figure 3-1 Sources of self-interference in a conventional single antenna front-end.

As presented in Figure 3-1, there are three possibilities where transmit signal can leak into the receive path. These include far-field of the antenna, near-field of the antenna, and at the circulator, which are highlighted in Figure 3-1 by numbers 1 to 3, respectively. The far-field reflection from the surrounding environment is often subject to the path loss and scattering from lossy obstacles, which results in weak self-interference. Therefore, this source of self-interference can be suppressed using digital cancellation. The near-field reflections is due to the imperfect antenna performance as well as placement of obstacles in the near-field region. External objects in this region alter the reactive power in the near field, which significantly influence the imaginary part of the antenna input impedance [62]. Finally the non-ideal performance of the circulator is associated with limited isolation between transmit and receive ports. The limited isolation is highlighted as β in Figure 3-1. Commercial circuit board compatible circulators often have an isolation around 20 dB over their suggested bandwidth. The S-parameters for two commercially available circulators are presented in Figure 3-2.



Figure 3-2 Two examples of performance characterization for commercial Sub-6 GHz circulators obtained directly from the manufacturers, (a) JQL Circulators (www.jqlelectronics.com) and (b) UIY Circulators (www.uiy.com).

Addressing these sources of self-interference using fixed circuitry is challenging and therefore, reconfigurable components are suitable solutions for the self-interference problem. In this chapter, two solutions for the self-interference in the near-field and circulator are presented. First, RIMNs are used to improve the matching between an antenna and a circulator. Second, RITs are studied for improving the isolation between receive and transmit paths of a circulator. These networks are implemented using various solutions, including switched or continuously tunable loaded lines, and tuning elements including RF-MEMS switches, Semiconductor Switches, Semiconductor Varactors, BST Varactors, and PCM switches.

3.2 Realizing Reconfigurable Stub Impedance Matching Networks Using RF-MEMS Switches

The impedance matching between an antenna and a circulator is a specific case for the doubly terminated matching network problem, which is shown in Figure 3-3. Here, both Z_s and Z_l are considered as variable impedances. For clarity, we refer to return loss as Γ_{dB} in decibels, and to the reflection coefficient as Γ in linear scale. In practice, these two terms are used as analogous to each other as the return loss is defined as:

$$\Gamma_{dB} = RL = -20 \log|\Gamma| \text{ dB}, \quad 3-1$$

which its value ranges from 0 dB to ∞ dB for a passive load [11], [63].

An impedance that is identified only by its return loss can be illustrated on the Smith chart using a constant $|\Gamma|$ circle. To draw such circle, we define $Z_l = a + jb = re^{j\theta}$, where a and b are real and

imaginary part, and r and θ are magnitude and phase of the load impedance, respectively. Then, for a load Z_l attached to a TL with characteristic impedance of Z_0 , we have:

$$\text{Constant } \Gamma \text{ Circle: } |\Gamma| = cte = \left| \frac{Z_l - Z_0}{Z_l + Z_0} \right| = \frac{\sqrt{(a - Z_0)^2 + b^2}}{\sqrt{(a + Z_0)^2 + b^2}} \quad 3-2$$

On the other hand, VSWR is related to Γ through:

$$VSWR = \frac{1 + |\Gamma|}{1 - |\Gamma|} \quad 3-3$$

Correspondingly, for a constant VSWR or Γ , one can specify the range of a as:

$$a \in [1/vswr, vswr], \quad 3-4$$

and for each value of a , b would have two values of

$$b = \pm \sqrt{\frac{\Gamma(a^2 + Z_0^2 + 2az_0) - (a^2 + Z_0^2 - 2az_0)}{1 - \Gamma^2}}, \quad 3-5$$

as shown in Figure 3-4.

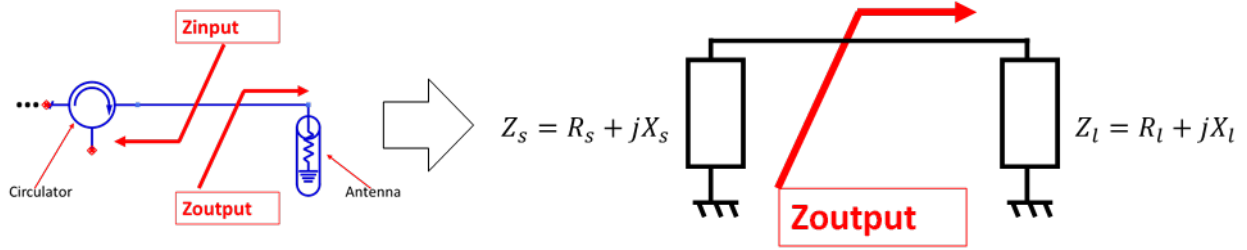


Figure 3-3 Impedance matching between antenna and circulator

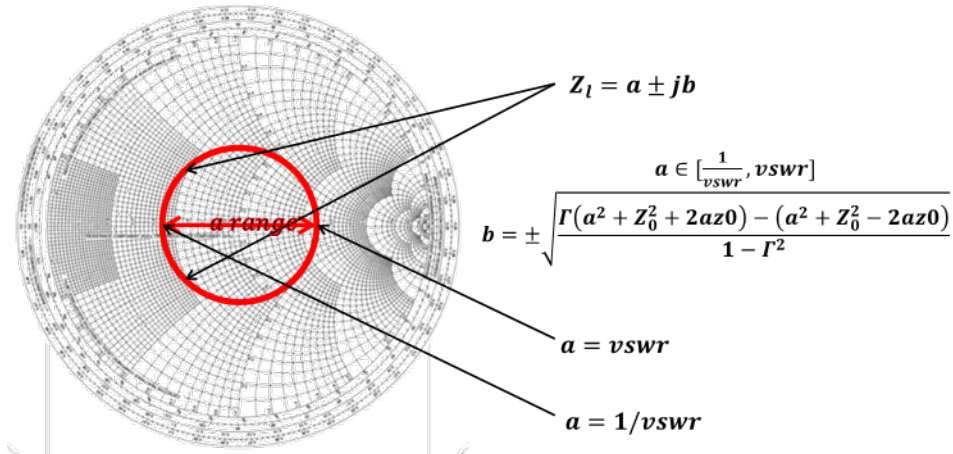


Figure 3-4 representation of constant return loss circles based on real and imaginary parts of the impedance

This representation is especially useful for the case of doubly terminated networks where both impedances are specified based on their VSWR (or return loss) to determine the required coverage. If an impedance is only described by its return loss or reflection coefficient, Γ , then it is represented by a circle on the Smith chart. If a load is expected to have a return loss better or worse than Γ_{dB} , then its potential location will be outside this constant $|\Gamma|$ circle, respectively. A few impedances with the associated with constant Γ_{dB} circles and different phases are listed in Table 3-1. For consistency, Z_0 is set to 50Ω whenever its numerical value is used throughout this thesis. Constant Γ_{dB} circles for values between 18dB and 35 dB are drawn in Figure 3-5 as well.

Table 3-1 Calculated impedances over constant return loss circles

ϕ RL, VSWR	0	$\pi/7$	$2\pi/7$	$3\pi/7$	$4\pi/7$	$5\pi/7$	$6\pi/7$	π
18 dB, 1.288	38.8	42.5 + 8.9i	46.1 + 11.6i	49.8 + 12.6i	53.4 + 12.6i	57.1 + 11.6i	60.7 + 8.9i	64.4
20 dB, 1.220	40.9	43.8 + 7.0i	46.6 + 9.1i	49.5 + 10.0i	52.4 + 9.9i	55.3 + 9.1i	58.2 + 7.0i	61.1
25 dB, 1.119	44.7	46.3 + 3.9i	47.9 + 5.0i	49.5 + 5.5i	51.1 + 5.5i	52.7 + 5.0i	54.3 + 3.9i	55.9
30 dB, 1.065	46.8	47.8 + 2.2i	48.7 + 2.8i	49.6 + 3.1i	50.5 + 3.1i	51.4 + 2.8i	52.4 + 2.2i	53.3
35 dB, 1.036	48.2	48.7 + 1.2i	49.2 + 1.6i	49.7 + 1.7i	50.2 + 1.7i	50.8 + 1.6i	51.3 + 1.2i	51.8

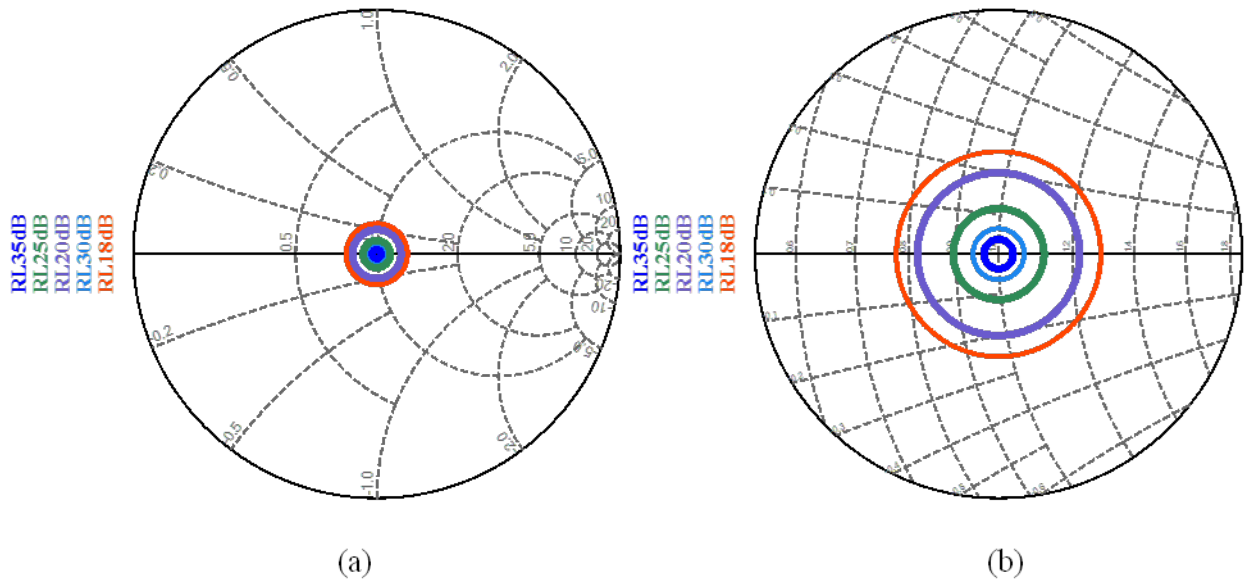


Figure 3-5 constant return loss circles on the Smith chart, (a) original view and (b) enlarged view highlighting the centre portion of the Smith chart between the values of 0.5 and 2 on the real axis.

The generalized problem of doubly terminated matching network is illustrated in Figure 3-6, where two independent impedances are connected to each other using a transmission line segment. Accordingly, the values of Γ_l and Γ_s are calculated with respect to the characteristic impedance of the connecting line, Z_0 . If the electrical length of the line is not significant, then the input reflection coefficient can be expressed as the following:

$$\Gamma_{in} = \frac{Z_l - Z_s}{Z_l + Z_s} = \frac{\frac{1+\Gamma_l}{1-\Gamma_l} - \frac{1+\Gamma_s}{1-\Gamma_s}}{\frac{1+\Gamma_l}{1-\Gamma_l} + \frac{1+\Gamma_s}{1-\Gamma_s}} \quad 3-6$$

which with basic algebra simplifies to:

$$\Gamma_{in} = \frac{\Gamma_l - \Gamma_s}{1 - \Gamma_l \Gamma_s} \quad 3-7$$

Similarly, $\Gamma_{in} = -\Gamma_{out}$. The equation 3-7 represents an annular ring around the $|\Gamma_l|$ circle on the Smith chart, which is illustrated in Figure 3-7.

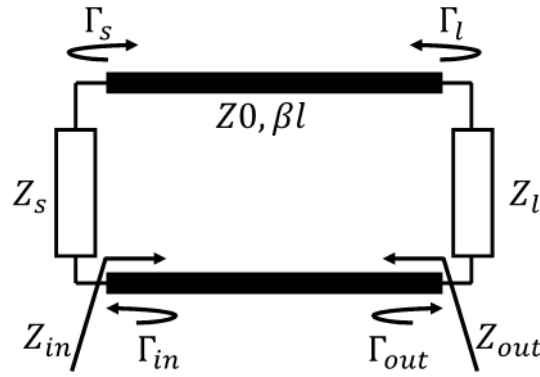


Figure 3-6 Doubly Terminated Matching Network

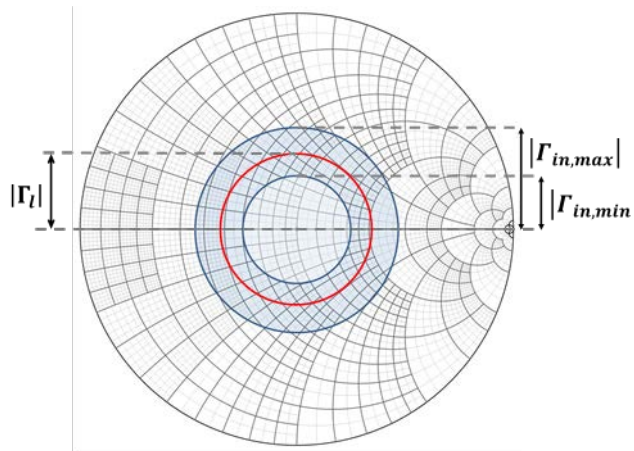


Figure 3-7 The required coverage area for doubly terminated matching networks

The complex conjugate matching at the input requires $\Gamma_s = \Gamma_{in}^*$. Identical areas are covered by Γ_{in} and Γ_{in}^* due to the symmetry of the annular ring. Therefore, the shaded area shown in Figure 3-7 is the required coverage area for matching a source and load described by their return loss of $\Gamma_{s,dB}$ and $\Gamma_{l,dB}$.

To determine the upper and lower ends of the ring we can write 3-7 in polar format as the following:

$$\Gamma_{in} = \frac{|\Gamma_l|e^{i\theta_l} - |\Gamma_s|e^{i\theta_s}}{1 - |\Gamma_l||\Gamma_s|e^{i(\theta_l + \theta_s)}} \quad 3-8$$

Based on the triangular inequality, absolute value of sum of two complex numbers is always smaller or equal to the sum of their individual absolute values and is greater or equal to their subtraction. Therefore, the numerator of 3-8 is maximized when θ_l and θ_s are π apart, and minimizes when they are in phase. Moreover, the product of $|\Gamma_l||\Gamma_s|$ is negligibly small with respect to one in the denominator for small values of $|\Gamma_l|$ and $|\Gamma_s|$. Accordingly, the annular ring of Γ_{in} is bound by two circles with the radii of $|\Gamma_l| + |\Gamma_s|$ and $|\Gamma_l| - |\Gamma_s|$. Note that when $|\Gamma_l| = |\Gamma_s|$, then $0 < |\Gamma_{in}| < 2|\Gamma_l|$. Moreover, when either Γ_l or Γ_s are determined by a return loss better than $\Gamma_{l,dB}$ or $\Gamma_{s,dB}$, then the internal radius reaches zero and conversely, when they are determined by a return loss worse than $\Gamma_{l,dB}$ or $\Gamma_{s,dB}$, then the external radius reaches one. As an example, Figure 3-8 depicts the coverage needed for matching a source and a load with 24 dB and 18 dB return losses.

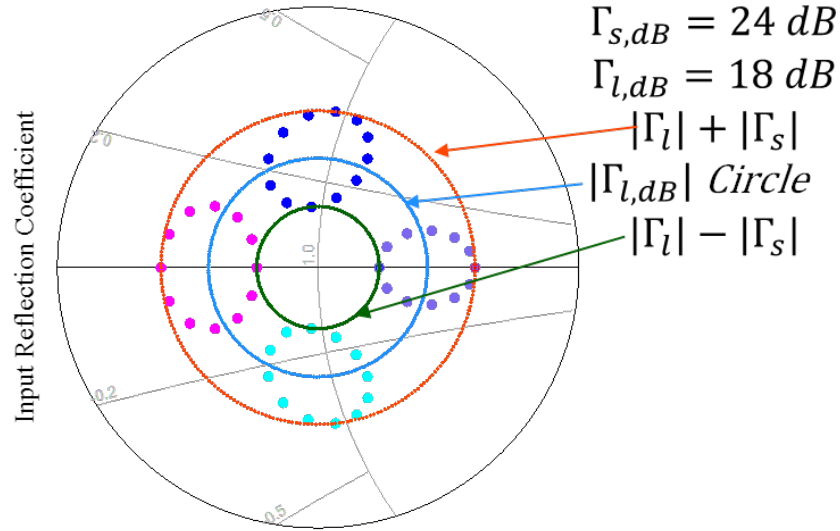


Figure 3-8 impedance distribution on the Smith chart when $\Gamma_{s,dB} = 24 \text{ dB}$ and $\Gamma_{l,dB} = 18 \text{ dB}$. For clear reading, θ_l is set to $0, \pi/2, \pi$, and $3\pi/2$. For each θ_l , θ_s is swept between 0 and 2π , in $\pi/6$ steps.

Introduction of a passive, reciprocal, symmetrical, and lossless matching network between the source and load results in alteration in distribution of the Γ_{in} on the Smith chart. These alterations include iterative

movement along constant susceptance or capacitance lines, by using shunt inductors or capacitors, respectively, and rotation around the centre using transmission line segments in series. Other operations including use of resistors, series capacitors, series inductors, and transformers are exempt due to introduction of losses, difficulties in realization, redundancy of design, and loss of symmetry (for bidirectional operation), respectively. In other words, such a matching network cannot perform tasks such as shrinking, expanding, or modifying the uniformity for a given impedance distribution. As a result, it is not possible to bring the entire impedance distribution within a prescribed vicinity around the centre of the Smith chart, unless the distribution already contained within the desired vicinity, in which case a matching network is no longer needed. Instead, one can employ a multistate matching network that transfers a portion of the original impedance distribution near the centre of the Smith chart at each state and the desired coverage is achieved by the combination of different states.

3.3 Switched Stubs Matching Network Based on MEMS SPSTs

The method described in the previous section is realized using a combination of π - and T- networks and replace their fixed elements with tuning elements. These networks provide coverage for the entire Smith chart. They can be realized at low frequencies by using lumped components such as varactors and tunable inductors. An example of a π -network doubly terminated matching network for application below 1 GHz is presented in Figure 3-9 [17]. Here, all the tuning elements are realized using tunable capacitors. The tunable inductor is realized using a resonator consisting of an inductor in parallel with a smaller tunable capacitor. Then, increasing the capacitance results in reduced effective inductance.

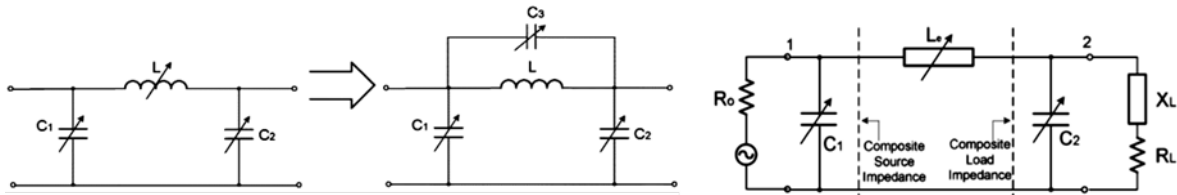


Figure 3-9 A tunable lumped matching network and the realization of series inductor using a tunable capacitor.

Due to the degradation of quality factor of lumped components, matching networks are realized using transmission line segments at higher frequencies. A conventional realization for the π -network is the well-studied double-stub matching networks. Compared to lumped element realization, an advantage of this method of realization is that the length and width of transmission line segments can be modified for open circuit, short circuit, inductive, and capacitive behaviour [64]. In the following, a switched transmission line matching network that is published in [65], is presented.

The proposed matching network is shown in Figure 3-10. Implementing series switched length transmission lines requires an even number of multistate switches placed at its both ends to select different lengths for transmission lines, which degrades the insertion loss performance of the matching network. Instead, an alternative solution is employed which required only one switch on a shunt transmission line segment to mimic a series segment. This is achieved by introducing $\lambda/4$ TL segments on each side of the switched-length shunt transmission line. Therefore, the segment shown as Z_2 in Figure 3-10 is realized using a T- network with two 90° arms and a switched length short-circuited stub. This method has been used in RF circuits, for example to convert shunt capacitors into series inductors [66].

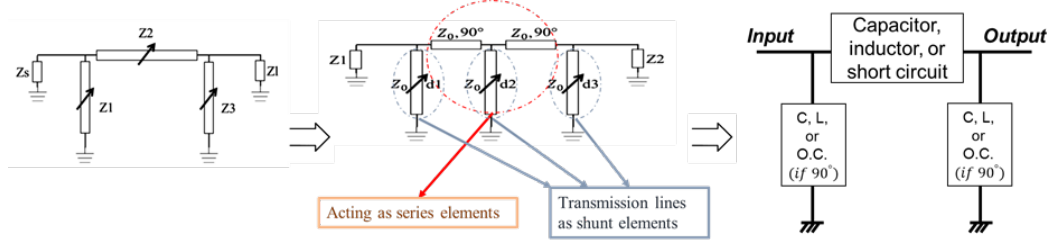


Figure 3-10 Realizing a reconfigurable π - network using switched length transmission lines

The three switched length stubs are terminated in short circuit to eliminate loading between different states as well as avoiding fringing effects. Subsequently, by employing n switches, each stub has $n+1$ states. Moreover, during the tuning and optimization steps, only the electrical length of these stubs was used to relax the frequency dependency. Furthermore, by including a 90° state for all the stubs, an effective bypass state is included in states where the matching network acts as a 50Ω transmission line segment, which connects the source and load without any additional loading. This is achieved because at 90° , the short-circuited stubs behave as open circuit terminations seen from the main branch side. The range of electrical lengths and switching steps are presented in Table 3-2. These values are derived based on the considerations for the number of states, available switches, control circuitry, and matching coverage.

Table 3-2 Lengths and switching steps of the stubs

Stubs	Range	Steps
d1 and d3	60° — 120°	$6 \times 10^\circ$
d2	65° — 115°	$6 \times 8.33^\circ$

A prototype RF-MEMS switches, MM-3100 by Menlomicro [44], is used in this design because each chip contains 6 independently controlled high power single pole single through (SPST) switches. Each stub has 7 switch states. The proposed design is capable of providing better than 30 dB of matching for a set of source and loads that are defined by 18 dB return loss. An accurate S-parameter model provided by the manufacturer is used to represent the MEMS switches. Moreover, the impedance distributions are generated using Batch Simulation Controller in Keysight ADS schematic environment. To demonstrate the achievable

coverage, the impedance distribution for 18 dB return losses is generated, as presented in Figure 3-11. Note that the target 30 dB matching is highlighted in this figure. Next step is to demonstrate that by introducing the proposed matching network, the entire impedance distribution can be transferred to the target matched circle using different matching states. This is shown Figure 3-12 for a few states.

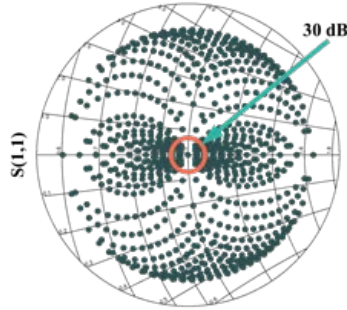


Figure 3-11 Impedance distribution for 18 dB return loss source and loads.

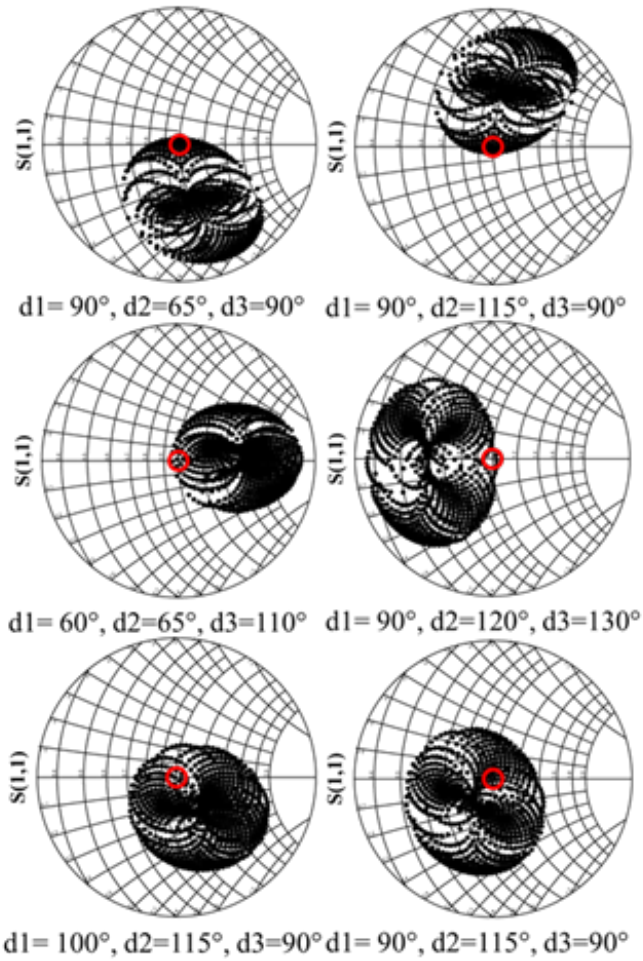


Figure 3-12 Achievable matching coverage and the associated switching states.

3.3.1 Operational Bandwidth

The dimensions presented in Table 3-2 are optimized for wideband operation. However, these properties are not unique; meaning there are other combinations of transmission line lengths that may provide the desired coverage. For example, the properties expressed in Figure 3-13 provide the required coverage. However, as depicted in Figure 3-13, it provides the coverage over a narrow band vicinity of the f_0 . On the other hand, termination of the stubs when they are separated by short line segments is problematic and difficult to fabricate. The coverage of the proposed design at different frequencies is presented in Figure 3-17. The accurate model for switches is used to generate these charts. As can be seen, around the centre frequency the complete coverage is achievable. However, at lower frequencies the 18 dB circle is not entirely covered. Moreover, at higher frequencies the coverage becomes excessively porous.

Table 3-3 Alternative dimensions for the proposed matching network

Stubs	Range	Number of Steps
d1	55°—75°	7 steps
d2	75°—105°	7 steps
d3	105°—125°	7 steps

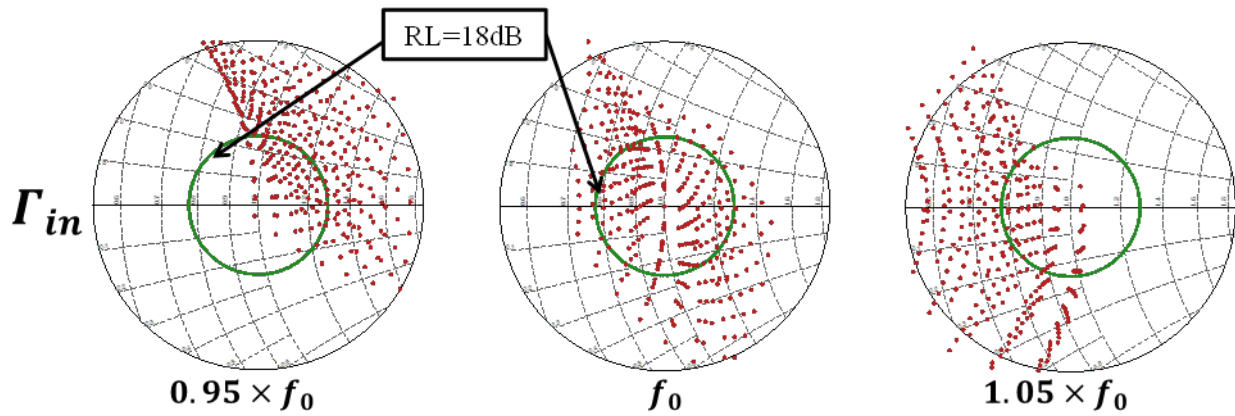


Figure 3-13 Coverage over different frequencies for the alternative matching network dimensions.

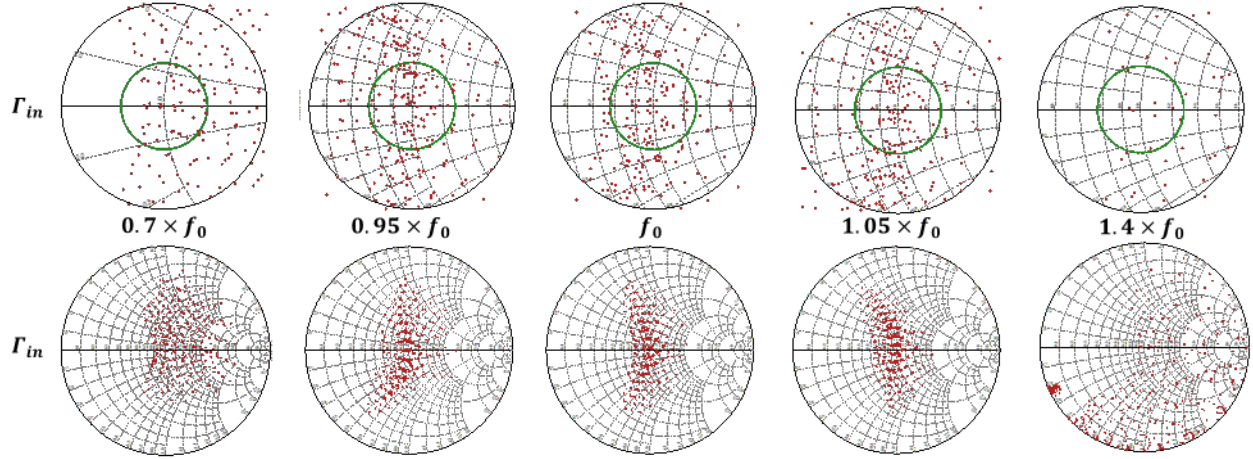


Figure 3-14 Achievable coverage by the proposed matching network over frequency.

3.3.2 Circuit Design and Optimization

A distributed circuit realisation is designed using 50 mil thick Rogers 3210 substrates with dielectric constant of 9.8, and loss tangent of 0.0027. The copper cladding is 1oz, which is plated with gold using ENIG process. The prototype circuit is designed at 2 GHz due to the low loss performance of the RF switches. The transmission line segments are GCPW with 300 μm wide signal lines and 200 μm gaps.

Figure 3-15 depicts the accurate circuit model used for the switches. It consists of parasitic elements as well as transmission line segments at the input and output to account for packaging and physical size of the switches. To account for these extra lengths a switching mechanism is used. The intention is to design a sub-circuit that behaves as open circuit termination when the switch is OFF. In contrast, it should behave as a short circuit termination when the switch is in ON state. The switching mechanism is illustrated in Figure 3-16. The switch is extended by an additional open circuited transmission line segment, which in conjunction with the internal length of the switch act as a short circuit termination for the switch. Moreover, an additional length is added before the switch to form a half wavelength transformer before the switch. Therefore, a short circuit and an open circuit termination are delivered at the input when the switch is in ON and OFF state, respectively.

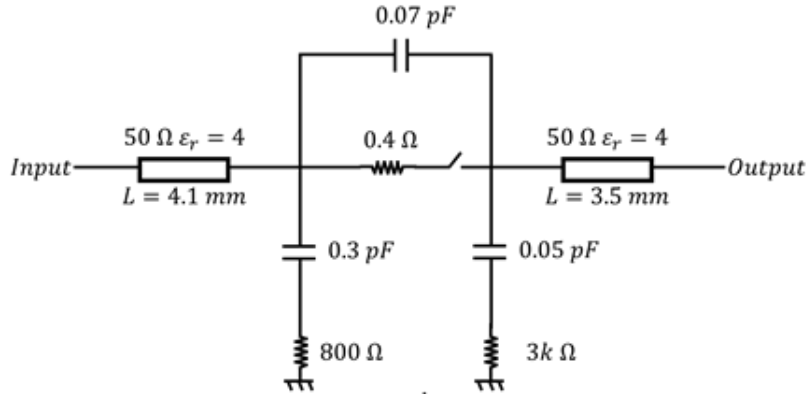


Figure 3-15 The accurate circuit model for the individual RF MEMS switches

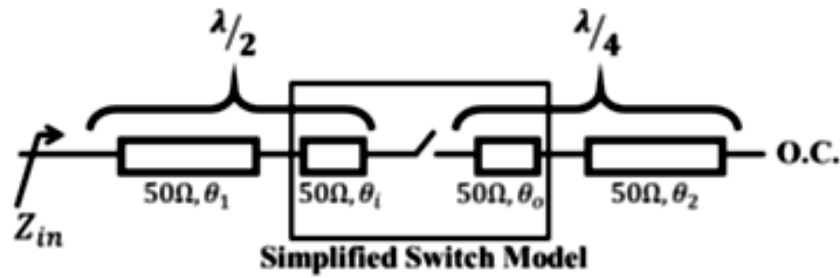


Figure 3-16 Switching mechanism for compensating for the undesired lengths of the switches

The switch chips, each contain six SPST switches and are attached to the stubs, as shown in Figure 3-17. The transmission line segments are bent to have the same lengths. The PCB circuit is designed and optimized in ANSYS HFSS environment and a 20 port touchstone file is used to transfer the simulation results into Keysight ADS for further analysis. The ports included 2 external ports for the input and output and 18 internal ports for the input and output of switches. The ADS environment is used so that the accurately model the switches using circuit elements. Moreover, the input and output ports are conveniently terminated in variable impedances. The optimized circuit parameters are listed in Table 3-4. Furthermore, the simulation results for a few states and their state number are shown in Figure 3-18. Each state is identified by a three digit number. Each number represent the switched state of one of the stubs. The digits show the closest switch to the main branch that is in ON state. Correspondingly, the number 1 means the first switch is ON and number 7 means non of the switches are ON. The states shown in Figure 3-18 show the edges of the impedance distribution can be matched.

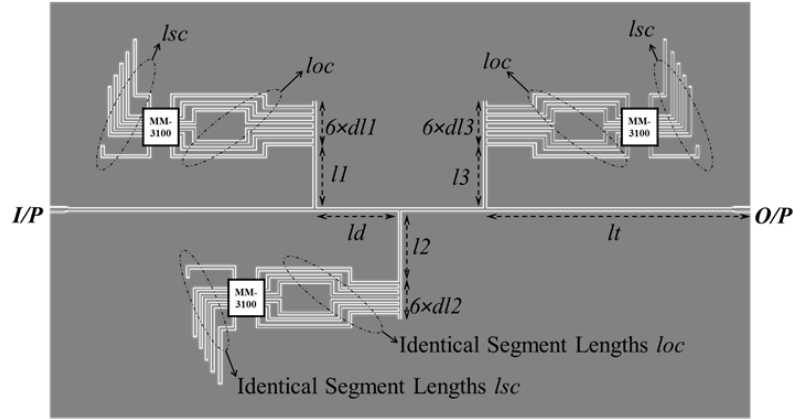


Figure 3-17 A detailed schematic of the fabricated PCB

Table 3-4 Design properties of the PCB

$l1$ (mm)	$l2$ (mm)	$dl1$ (mm)	$dl2$ (mm)	loc (mm)	lsc (mm)
11.19	12.7	1.865	1.55	30.7	13.3
ld (mm)	lt (mm)	MM-3100 chip size (mm ²)			
16.78	45	6×6			

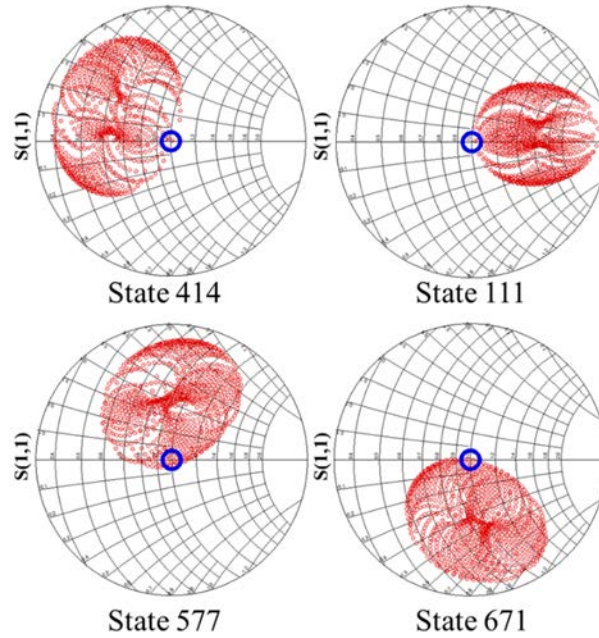


Figure 3-18 Simulation results for achieved coverage in marginal states. These states show that the edges of the impedance distribution can be matched.

3.3.3 Fabrication and Measurements

The proposed design is fabricated and assembled by local companies and was tested in Centre for Integrated RF Engineering (CIRFE) labs. The PCB finish was immersion gold (ENIG), which was required for the

assembly of MEMS chips. The chips are connected to the control circuit using compact 1.27 mm pin headers. The assembled device is shown in Figure 3-19. The control circuit uses SPI protocol to control the state of switches. The proposed device cannot be tested directly in the lab because of the variable source and load requirements. Accordingly, the circuit is terminated in 50Ω source and load port impedances for measurements and the exported data are used in ADS for modifying the termination impedances. The device is tested using N5242A PNA, which is calibrated to 50Ω load impedances. For post processing, a few states are randomly selected and their corresponding matched loads are selected from impedances within the 30 dB circles. Then the corresponding return loss and insertion loss are compared.

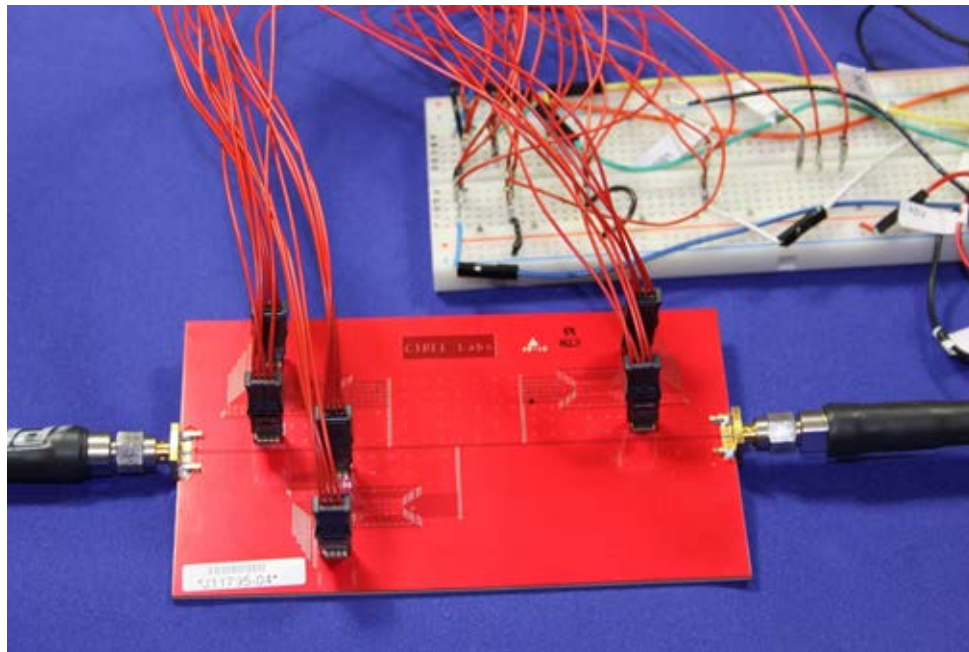


Figure 3-19 Fabricated doubly terminated matching network under test.

Wide band and narrow band measurement results are depicted in Figure 3-20. As can be seen the 30 dB matching target is achieved for all the states. It is noticeable that apart from state 777, which shows the exact anticipated performance at 2 GHz, other states are slightly shifted toward higher frequencies. This miniscule discrepancy between state 777 (all switches OFF) and the rest of states may be caused by switch malfunction or small discrepancies in the parasitics modeled for the switches. The insertion loss is consistent for all the states and differs by less than 0.5 dB. Moreover, the matched bandwidth for different states ranges from 20 MHz to 170 MHz, which is wider than the requirements for LTE applications and is significantly wider than competing designs. Finally, the limited bandwidth of this design is caused by use of multiple half and quarter wavelength transformers in the circuit. Termination of the stubs with bare die MEMS switches or designing a fully integrated circuit eliminates the need for these sections and improves the operational bandwidth performance.

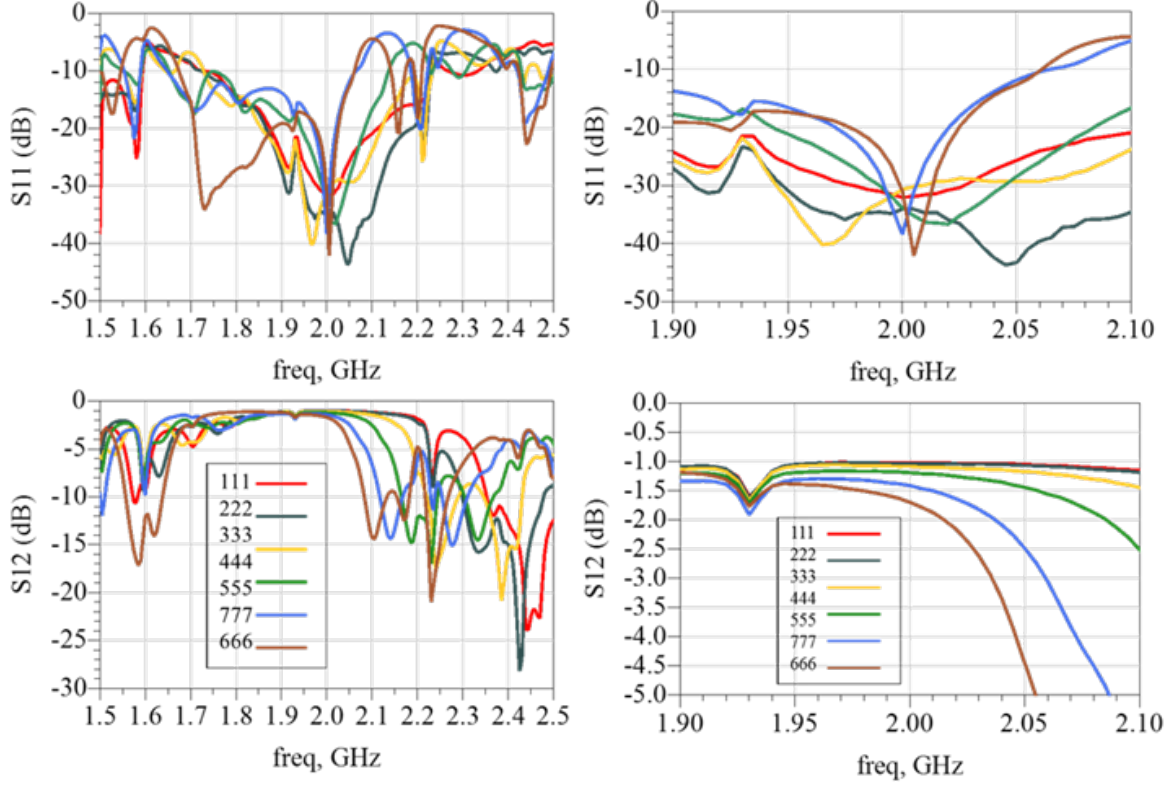


Figure 3-20 Measurement S11 and S21 for wideband (left) and narrowband (right) operation.

3.4 Demonstrating the Performance of Doubly Terminated Matching Networks Using Singly Terminated Circuits

The method presented in section 3.3 does not show the coverage between the states. Moreover, it needs simultaneous matching between the antenna and circulator and its input and output cannot be tuned independently. In this section, we investigate another method, which uses smaller singly terminated circuits to show the performance of the complete matching network. The main intention, apart from the proof of the concept, is to develop a method for demonstrating the coverage and the insertion/return loss performance of the proposed device. This is done by proposing a matching network for a circulator and an antenna with return losses of better than 18 dB and 15 dB respectively. The target is 40 dB return loss over matching bandwidths of 20 MHz. The frequency of operation is between 3.4 GHz and 3.7 GHz, which is a globally allocated 5G band, known as mid-band or Sub-6 GHz band [67].

Figure 3-21 shows a doubly terminated matching network on the left and a break down version on the right, which is composed of two smaller matching networks with reflection coefficients Γ_1 and Γ_2 in the middle; looking toward the antenna and circulator, respectively. In this figure, Γ_A and Γ_C are calculated based on a 50 Ω line. The middle coefficients can be expressed as the following:

$$\Gamma_1 = S_{11}^A + \frac{S_{21}^A S_{12}^A \Gamma_A}{1 - S_{22}^A \Gamma_A} \quad 3-9$$

$$\Gamma_2 = S_{22}^C + \frac{S_{21}^C S_{12}^C \Gamma_C}{1 - S_{11}^C \Gamma_C} \quad 3-10$$

Moreover, the input and output reflection coefficients are written as:

$$\Gamma_{inC} = S_{11}^C + \frac{S_{21}^C S_{12}^C \Gamma_1}{1 - S_{22}^C \Gamma_1} \quad 3-11$$

$$\Gamma_{inA} = S_{22}^A + \frac{S_{21}^A S_{12}^A \Gamma_2}{1 - S_{11}^A \Gamma_2} \quad 3-12$$

The complex conjugate matching condition at the centre calls for $\Gamma_1 = \Gamma_2^* \Rightarrow |\Gamma_1| = |\Gamma_2|$ & $\angle \Gamma_1 = -\angle \Gamma_2$. A special case is when $\Gamma_1 = \Gamma_2 = 0$. This leads to simplification of 3-11 and 3-12 to:

$$\Gamma_{inC} = S_{11}^C \text{ and } \Gamma_{inA} = S_{11}^A \quad 3-13$$

Accordingly, to address the complex conjugate condition at the circulator's side, one needs to modify the S_{11}^A so that:

$$S_{11}^A = \Gamma_C^* \quad 3-14$$

Note that 3-14 is valid when $\Gamma_1 = \Gamma_2 = 0$. This can be achieved by setting $S_{22}^A = \Gamma_A^*$. Then 3-9 simplifies to:

$$\Gamma_1 = S_{11}^A + \frac{S_{21}^A S_{12}^A S_{22}^{A*}}{1 - S_{22}^A S_{22}^{A*}} \quad 3-15$$

Moreover, imposing the unitary condition for the lossless and reciprocal networks, we have:

$$\begin{cases} S_{12}^A S_{22}^{A*} = -S_{21}^{A*} S_{11}^A, \\ 1 - |S_{22}^A|^2 = |S_{21}^A|^2 \end{cases} \quad 3-16$$

Plugging 3-16 in 3-15 and then eliminating the equal terms leads to

$$\Gamma_1 = S_{11}^A - \frac{S_{21}^A S_{21}^{A*} S_{11}^A}{1 - S_{22}^A S_{22}^{A*}} = S_{11}^A - \frac{|S_{21}^A|^2 S_{11}^A}{1 - |S_{22}^A|^2} = 0 \quad 3-17$$

Similarly, for $S_{11}^C = \Gamma_C^*$, $\Gamma_2 = 0$.

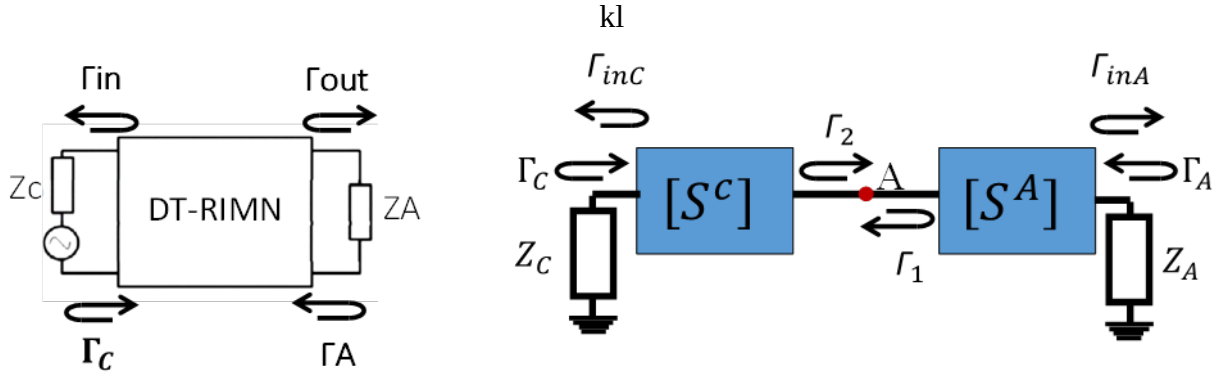


Figure 3-21 Doubly terminated matching networks and reflection coefficients.

Now assume that the doubly terminated matching network in Figure 3-21 is divided into two separate networks of S^C and S^A , and these networks are detached from each other at the point A. If these two networks are separately tuned to achieve $\Gamma_1 = 0$ and $\Gamma_2 = 0$ and then re-attached to each other, then complex conjugate matching is achieved at both ends. Although, this representation seems trivial, it has a very useful application in demonstrating the impedance coverage of reconfigurable doubly terminated matching networks. For this purpose, one needs to singly terminate the tunable S^C and S^A networks in Z_C and Z_A , respectively. Then, it should be demonstrated that for different values of Z_C and Z_A , there is at least one state for S^C and S^A networks that leads to $\Gamma_2 = 0$ and $\Gamma_1 = 0$, respectively.

3.4.1 Continuously Tunable Doubly Terminated RIMNs

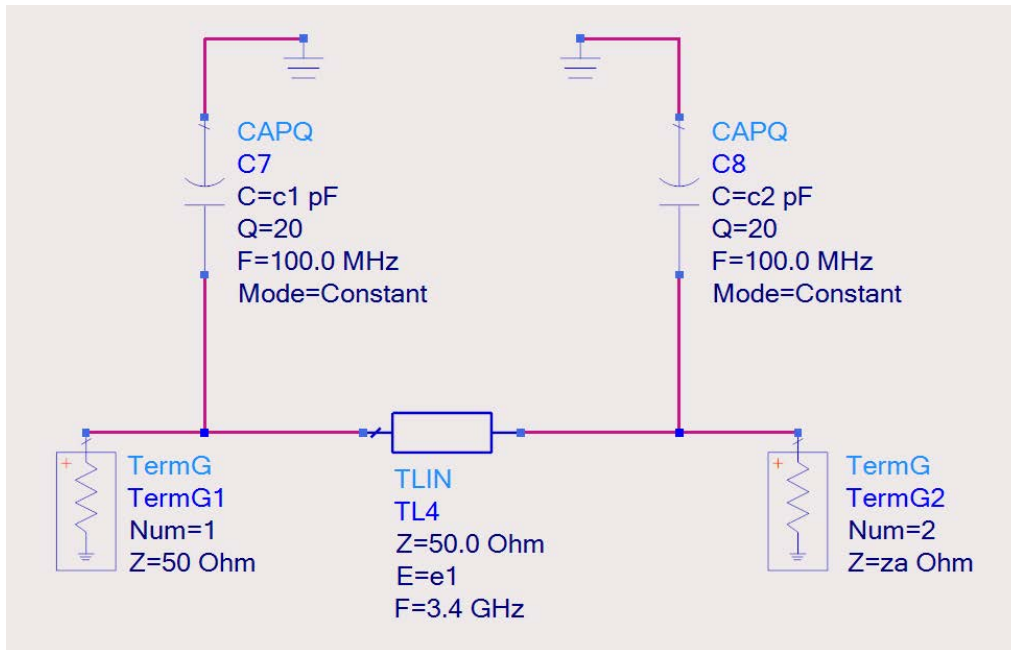


Figure 3-22 Singly Terminated RIMN Schematic in ADS Environment.

In this section, we investigate the circuit level performance of a capacitor loaded circuit prototype. Figure 3-22 shows a simple loaded line, which employs two tunable capacitors with non-ideal quality factor. The tuning range for the capacitors is 0.4 pF to 2.5 pF, which corresponds to the characterized performance of SMV2019-000 varactors presented in chapter 3. The circuit is simulated at 3.4 GHz and two capacitors are separated by a 45° transmission line segment. Figure 3-23 shows the achievable coverage when antenna impedance has a return loss of 15 dB but with different phases. For convenience and simplicity, the phases are 90° apart. Figure 3-24 shows the coverage for when the circulator return loss is 18 dB. As can be seen both these figures demonstrate that the centre of the Smith chart is covered for the studied antenna and circulator impedances.

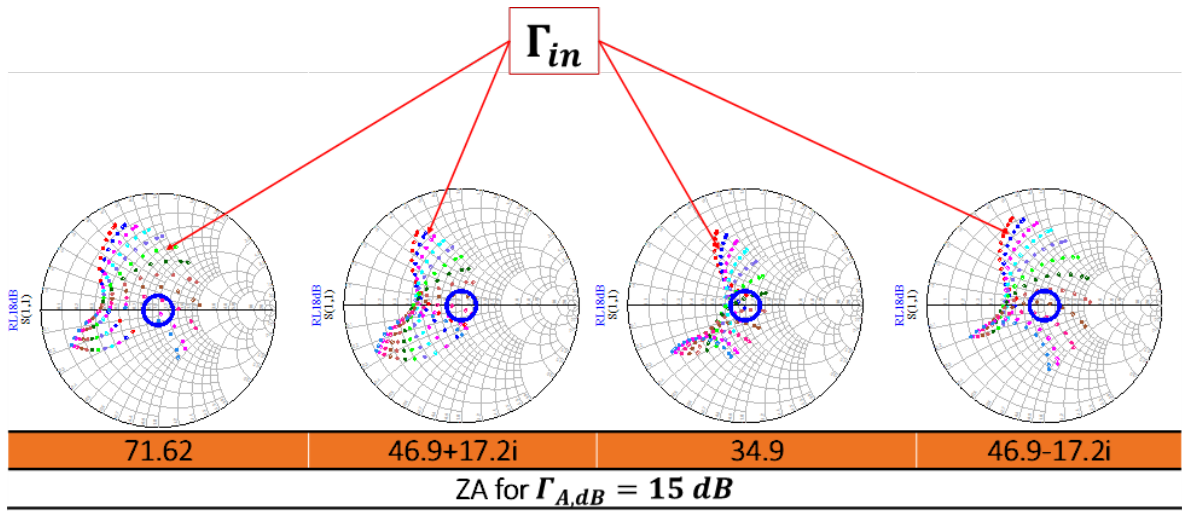


Figure 3-23 Input reflection coefficient for different RIMN states and antenna impedances

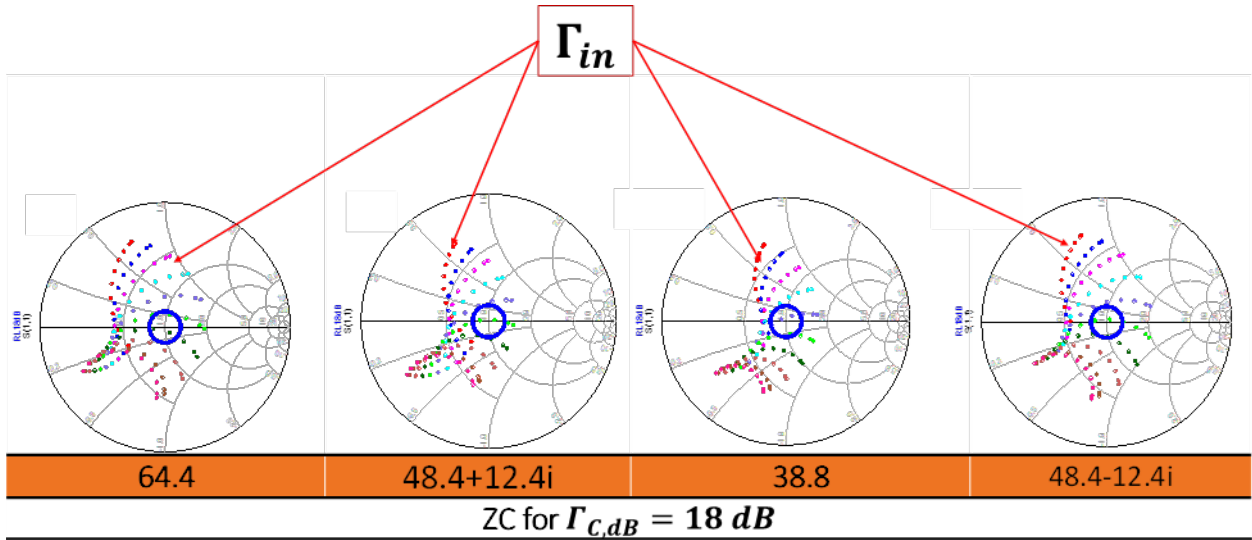


Figure 3-24 Input reflection coefficient for different RIMN states and circulator impedances

Next step in demonstrating the performance of the doubly terminated network is to attach the two circuit in cascade and terminate the input and output with the antenna and circulator impedances. The resulting network is composed of four tunable capacitors, and two 45° transmission lines. However, since the two capacitors in the middle are connected in shunt, they can be replaced by a single capacitor with twice the tuning range. Therefore, the final circuit is composed of two 45° line segments that separate three tunable capacitors. The centre capacitors also has twice the tuning range of the outer capacitors.

Figure 3-25 shows the schematic of the proposed matching network. The electrical length $e2$ will not affect the matching coverage because it only adds a phase shift to the input and output sections. This length is often determined based on the input and output connectors (probe pads or SMA connectors). The circuit model is designed on 25 mil thick Alumina substrate with $\epsilon_r = 9.9$ and $\tan \delta = 1 \times 10^{-4}$. The capacitance range of values mimic the performance of Skyworks SMV 2019-000 and SMV 2022-000 varactors. Accordingly, the microstrip transmission lines are simulated with width of $600 \mu\text{m}$ and the lengths of $500 \mu\text{m}$ for the outer segments and $6300 \mu\text{m}$ for the middle segments.

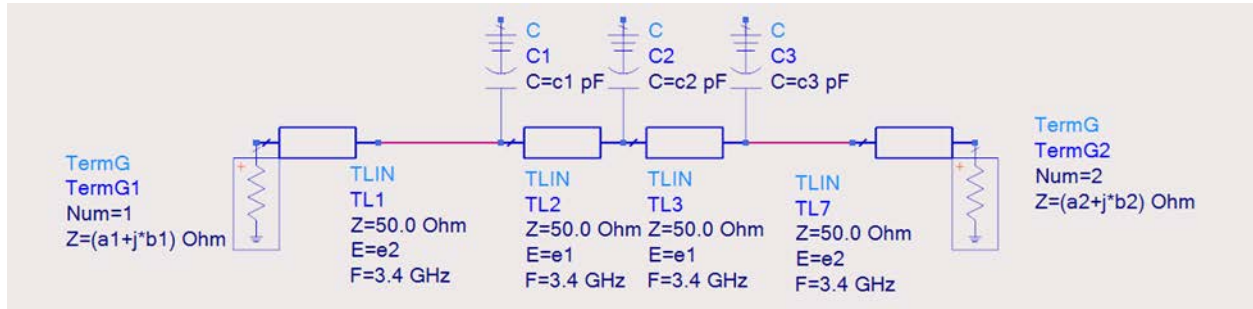


Figure 3-25 Schematic view of the proposed capacitive loaded line.

Figure 3-26, Figure 3-27, and Figure 3-28 show the performance of the proposed design at three frequencies. In these figures, each cell represents a combination of antenna and circulator impedances, where the rows represent different circulator impedances and columns represent different antenna impedances. As can be seen, 20 MHz bandwidth for 40 dB return loss is achieved in almost all states. The few exceptions, where matching bandwidth is not achieved is due to local minimums in optimization cost function response, which can be resolved by choosing a different set of initial values for the capacitors.

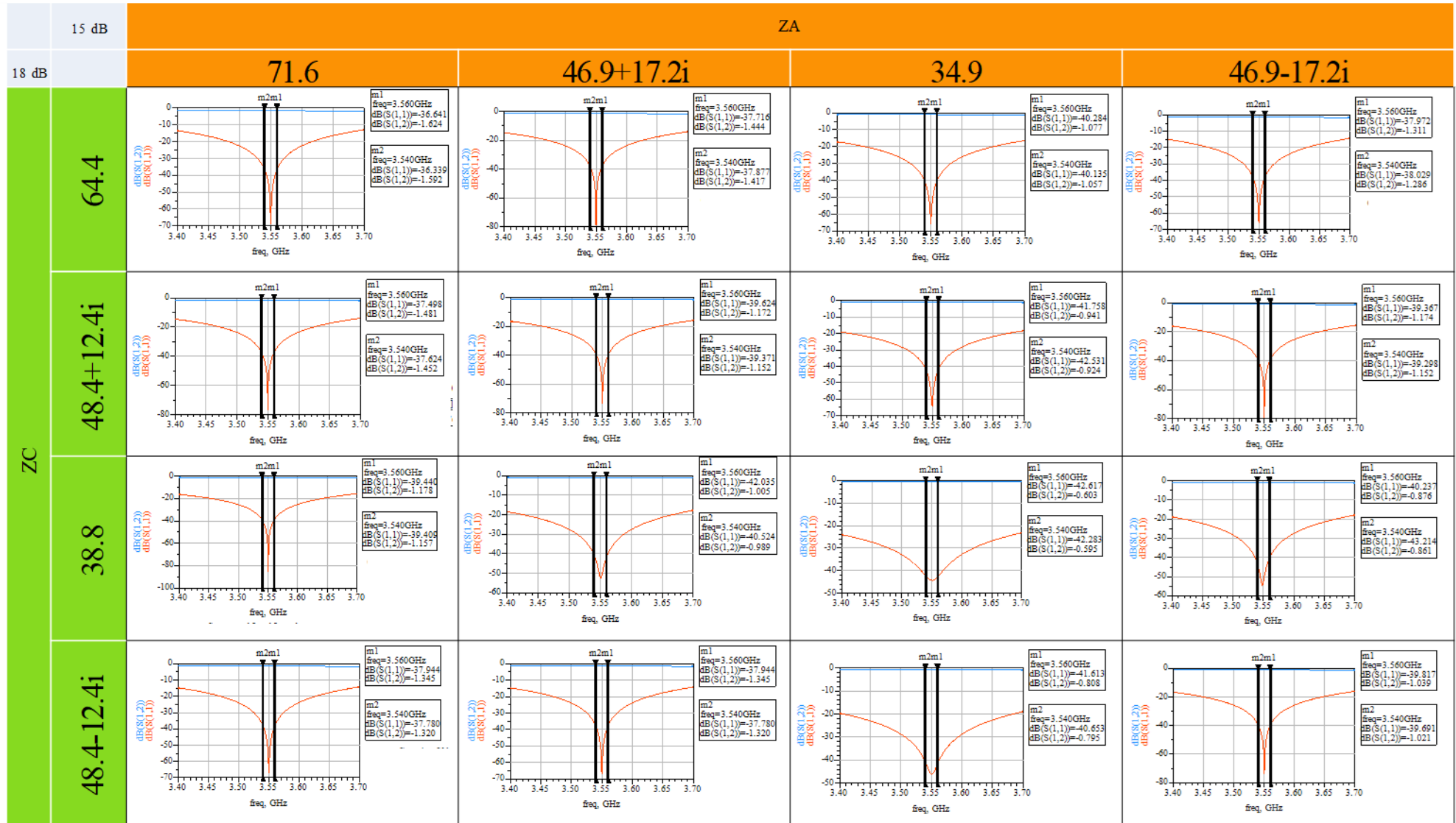


Figure 3-26 S-parameters for a combination set of antenna and circulator impedances. Matched around 3.55 GHz.

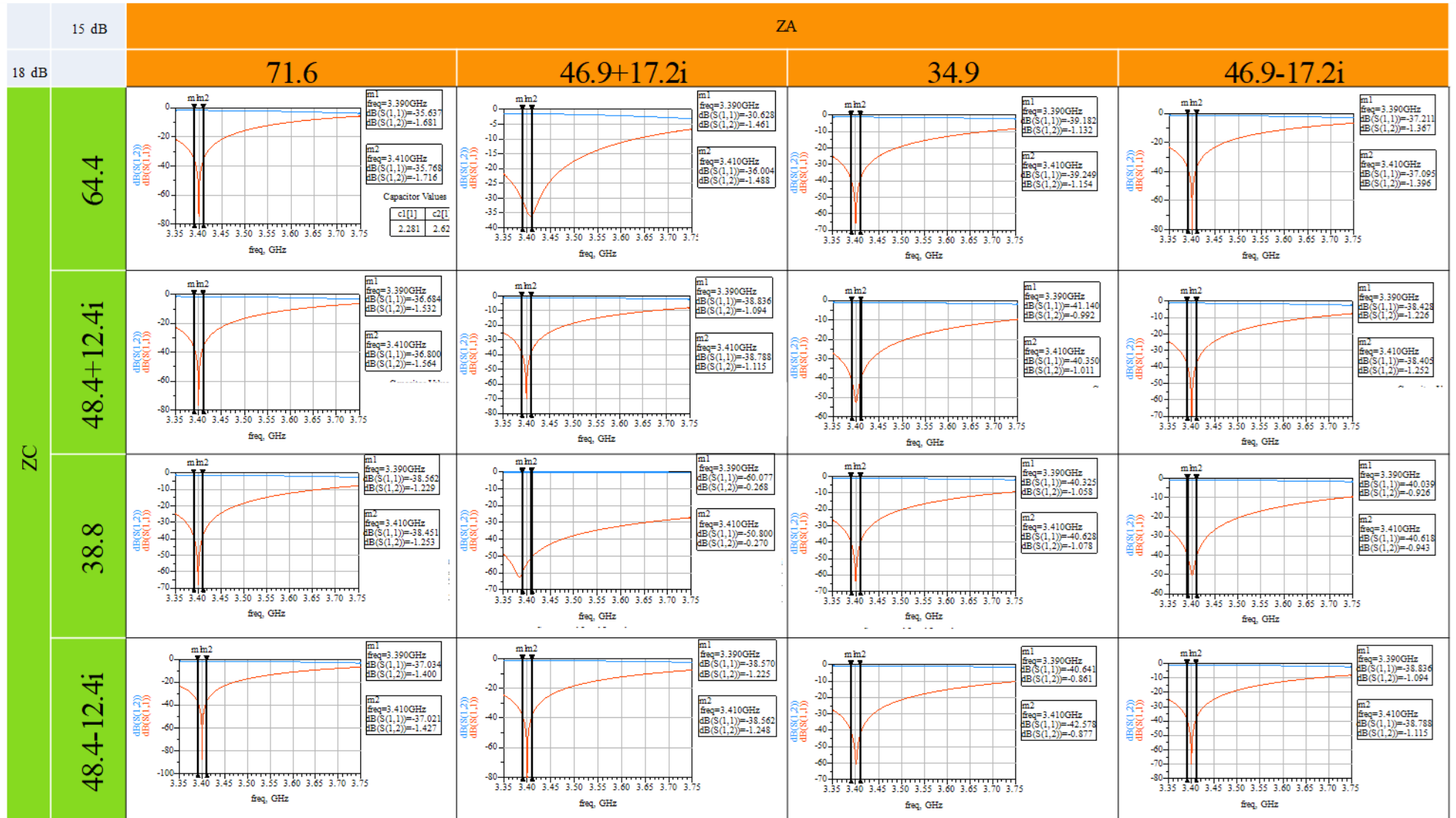


Figure 3-27 S-parameters for a combination set of antenna and circulator impedances. Matched around 3.4 GHz.

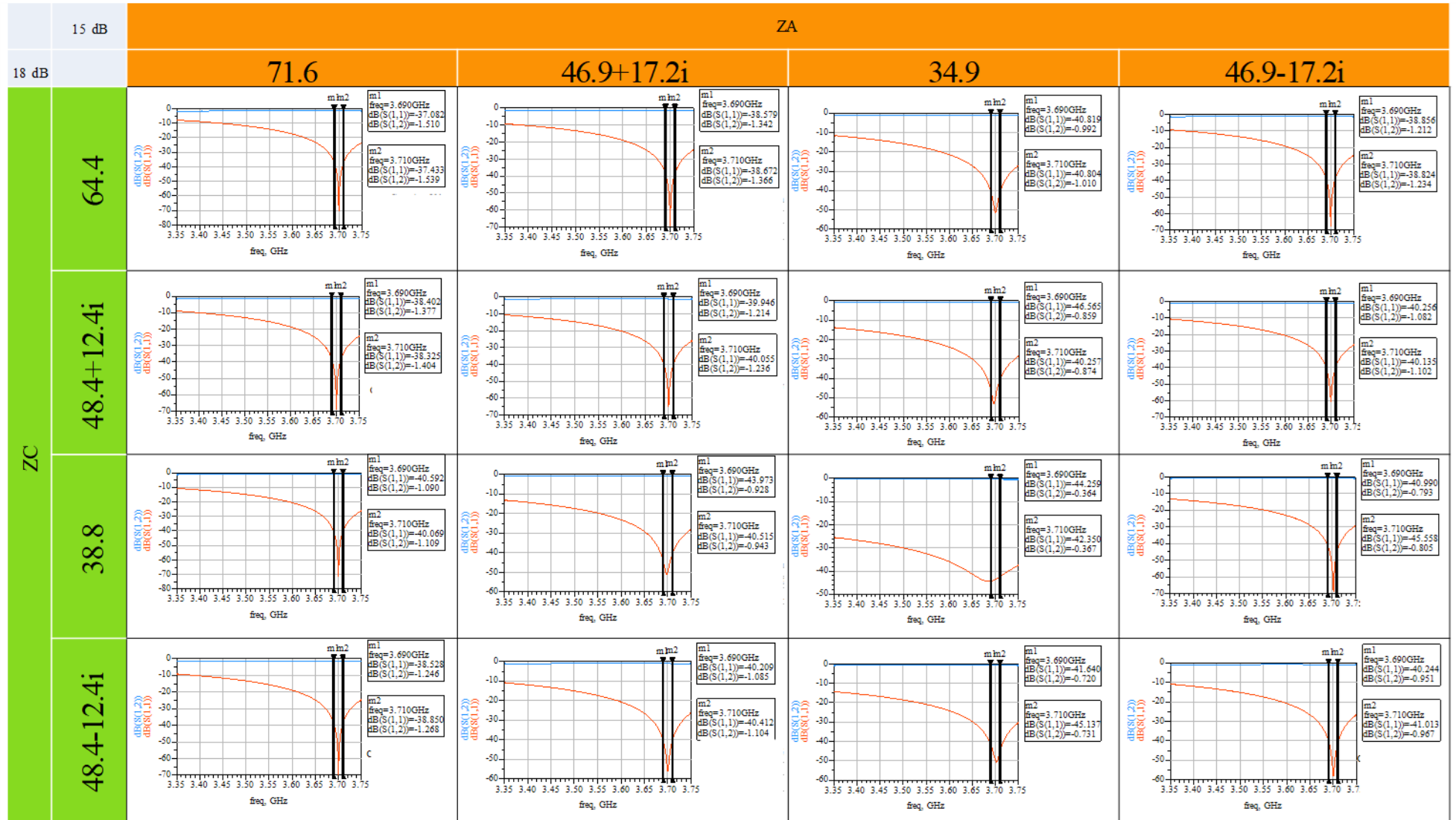


Figure 3-28 S-parameters for a combination set of antenna and circulator impedances. Matched around 3.7 GHz.

3.4.1.1 Complete Circuit Model for the Loaded Line Devices

The implementation of the circuit shown in Figure 3-25 poses a few challenges that need to be addressed. First, implementing ground termination in microstrip technology requires vias. Vias add a level of unwanted inductive loading, requiring additional processing for high dielectric constant substrates, and degrade the quality factor of the capacitors. Second, the T-junctions between the main branches and the capacitors are not modeled. Third, the connection between the T-junctions and the capacitors form a short transmission line segment, which its impedance and electrical length are dictated by the physical constraints of the circuit. Fourth, microstrip lines require metallization on the backside of the substrates, which is not suitable for technologies other than PCB (e.g. electroplating and metal sputtering, etc.). Accordingly, it is more convenient to implement the circuit using CPW transmission lines, because they are fabricated on a single layer, do not need via conductors for grounded terminations, and their characteristic impedance are freely modified by tuning the width and gap of the line.

Figure 3-29 illustrates a detailed equivalent circuit for a loaded line circuit (shown on the right). It is composed of multiple T-junctions that are separated by horizontal CPW line segments. Each T-junction is also connected to a tuning element (i.e. tunable capacitor) using different transmission line segments. In Figure 3-29, the T-junction is highlighted by number 1, the horizontal line by number 2, the connecting transmission line inside the gap of the horizontal segments by number 3, the connecting segment intruding into the ground plane by number 4, and the port for capacitors by number 5. This is a basic model and does not encapsulate the interactions between different segments (e.g. coupling between different segments and fringing effects). However, this model becomes useful when the tuning element is not moved into the ground plane and the gap of section 2 is wide. In other words, this model is useful when section 4 is removed and the section 3 has a shorter length than the gap of section 2. In this case, the conventional T-junction models (composed of a shunt capacitor at the centre of the junction and three inductors for the three ports [68]) are no longer valid because of the removed ground plane. Moreover, for the conventional models of T-junction to be accurate, the three ports must be far from the junction to prevent perturbation and influencing the shunt capacitor in the conventional model. However, in a case where the section 3 and 4 are short, the interaction between the tuning element and the junction makes the conventional method unreliable [68]. Accordingly, we use this method for two applications. First, we use it for a DT-RIMN, which has a wide signal trace, narrow gaps, and short transmission line segments between the main branch and the tunable capacitors. Second, we investigate a design with high-impedance main branch segments with wide gaps and short junction segments for realizing a continuously tunable 90° phase shifter, which will be presented in Chapter 5. These circuits employ BST tunable capacitors instead of semiconductor varactors

because BSTs have higher quality factor compared with the semiconductors at higher frequencies and are usable over the 3.4 GHz to 3.7 GHz frequency range, can be conveniently assembled, and do not require bias circuits, which dramatically simplifies the realization of the circuits compared to the semiconductor varactors based circuits.

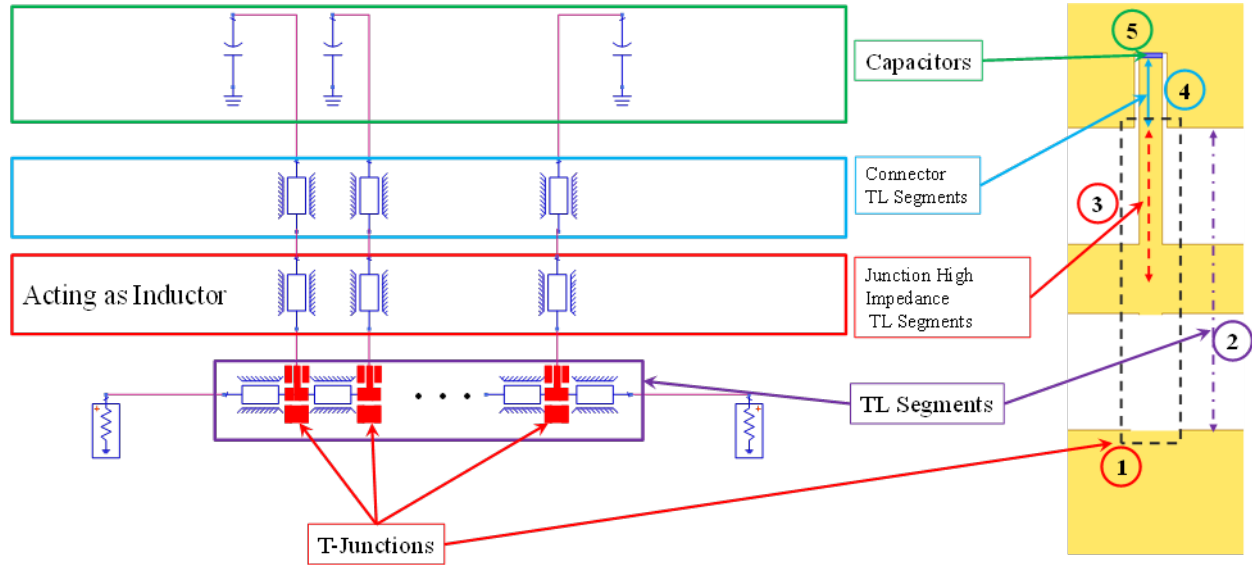


Figure 3-29 Analogy between circuit and EM model for the CPW loaded line.

3.4.1.2 BST Characterization

The BST capacitors used for the proposed circuits are prototypes from ON-Semi, under the commercial name of Paratek HF Passive Tunable Integrated Circuits (PTICs). They have a nominal ratio of maximum capacitance to minimum capacitance of 4.65. The prototypes have a higher quality factor than the commercially available devices. Therefore, test circuits are developed in CIRFE labs to characterize samples with maximum capacitance values of 1.0pF, 1.2pF, 1.5pF, 1.8pF, 2.2pF, 2.7pF, 4.7pF, and 8.2pF. The singly terminated characterization circuit, the circuit model for the tuning element, and the internal block diagram of the BST capacitors are presented in Figure 3-30. The BST variable capacitors do not need external biasing network and the devices operate when attached to a bias voltage.

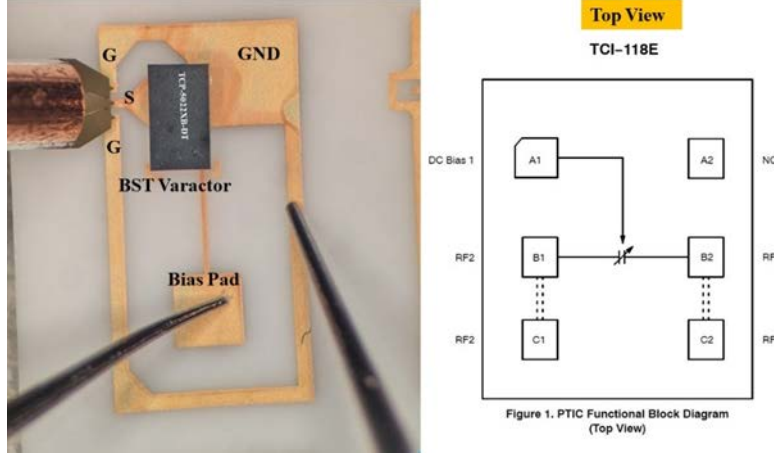


Figure 3-30 Characterization of BST tunable capacitors.

Measured capacitance for the 1.5 pF and 2.2 pF BST varactors are presented in Figure 3-31. As can be seen, a negative capacitance value is recorded for each state after the self resonance frequency (SRF). Moreover, the resonances for the 2.2 pF device occur at lower frequencies compared to the 1.5 pF BST. The capacitance vs voltage curves at 3.5 GHz are presented in Figure 3-32. The recorded lowest capacitance are 0.71 pF and 0.51 pF, recorded for 17V bias voltage, for the 2.2 pF and 1.5 pF, respectively. The recorded highest capacitance is significantly larger than the nominal values. This is because the capacitance value near the SRF tends to increase. However, for these states the quality factor becomes significantly small. This is shown in Figure 3-33. As can be seen, the bias voltages greater than 8V and 4V result in quality factors of more than 10 for the 2.2 pF and 1.5 pF, respectively. A similar trend is recorded for the other available BSTs, which are omitted here to reduce the volume of this thesis. Wherever needed in the reminder, the BST varactors are modelled by a variable capacitor with a constant limited quality factor of 20~30 over the frequency range.

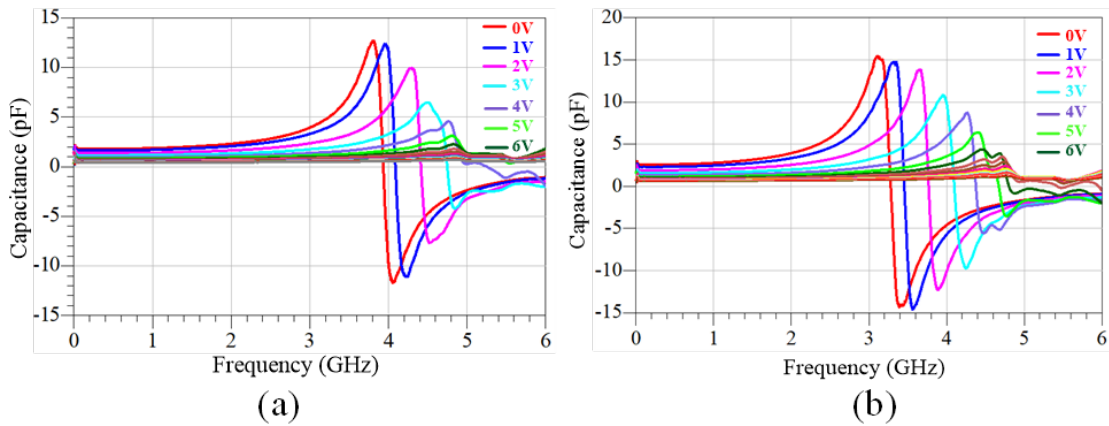


Figure 3-31 Measured capacitance of the BST capacitors, (a) 1.5 pF and (b) 2.2 pF.

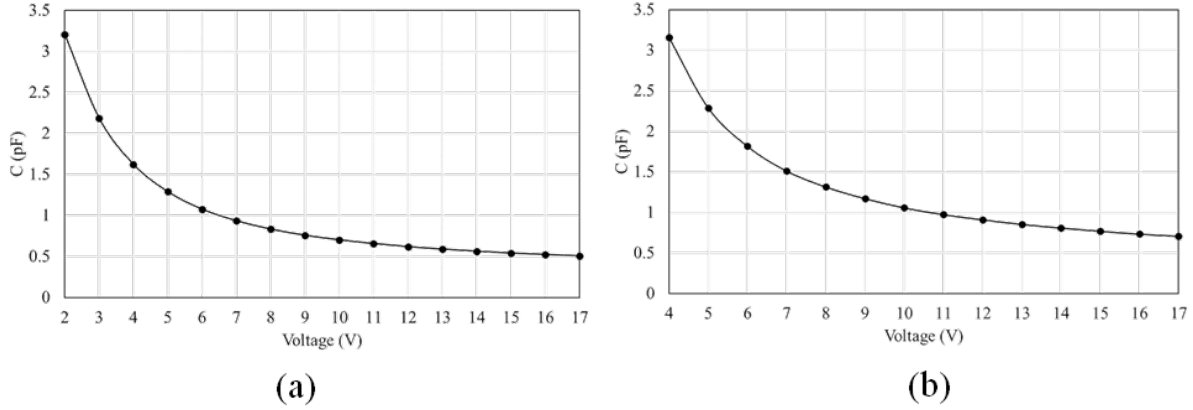


Figure 3-32 Measured capacitance vs bias voltage at 3.5 GHz, (a) 1.5 pF and (b) 2.2 pF.

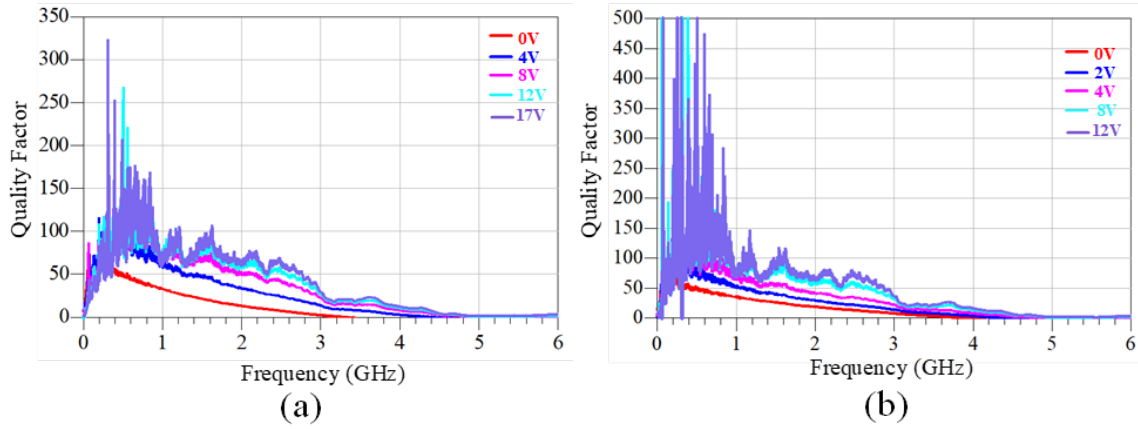


Figure 3-33 Measured quality factor of the BST capacitors, (a) 1.5 pF and (b) 2.2 pF.

3.4.2 Realising DT-RIMNs Using BST Loaded Lines

A circuit prototype DT-RIMN is designed, fabricated, assembled, and tested in CIRFE labs, which employs BSTs as its tuning elements. While the range requirement for the capacitance values that was discussed in the previous section cannot be exactly met using the characterized BSTs. However, the coverage can be achieved by optimizing the characteristic impedance and the length of the TL segments. The circuit is optimized for using 1.8 pF and 2.2 pF BSTs, as well as miniaturization. Accordingly, it is realized using 37Ω lines and 37° segments on 25mil alumina wafers. The initial circuit design is based on CPW lines using $3\mu\text{m}$ of ideal gold on top of the substrate. The layout of the circuit is illustrated in Figure 3-34. The entire circuit and its DC bias pads occupy an area of 10.4 mm by 4.8 mm. The area of the circuit, excluding the bias pads, is 10 mm by 1.4mm.

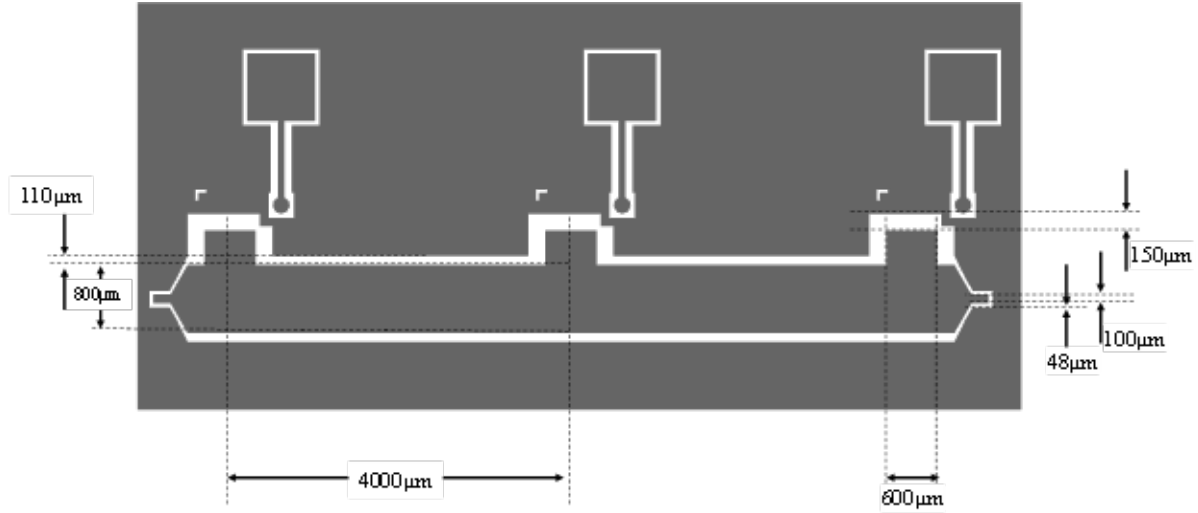


Figure 3-34 Schematic view of the DT-RIMN with its dimensions.

The full-wave simulated circuit is terminated by variable capacitors with constant quality factor of 30. The simulated distribution of the input and output reflection coefficients are plotted at 3.4 GHz, 3.55 GHz, and 3.7 GHz in Figure 3-35. For clarity 18 dB constant return loss at the input, 15 dB constant return loss at the output, as well as 10 dB return loss at both ends are highlighted. Note that $18\text{ dB} = 0.1259$ and $15\text{ dB} = 0.1778$. Therefore, the $|T_i| + |T_s| = 0.3037 = 10.35\text{ dB} \cong 10\text{ dB}$. Since the coverage over the frequency range covers well beyond the 10 dB circle, the coverage is demonstrated using the calculations in section 3.2 as well.

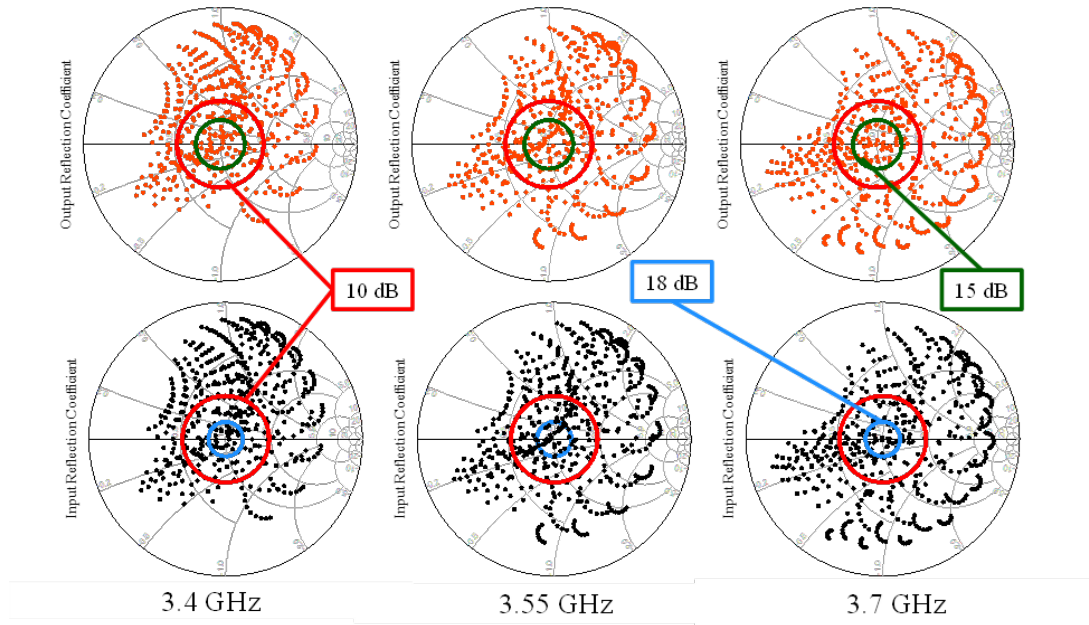


Figure 3-35 The simulated overage for the continuously tunable RIMN using BSTs.

Two copies of the proposed DT-RIMN are fabricated on 25 mil alumina substrate using e-beam evaporation and electroplating processes to develop 150nm and 3 μ m of gold, respectively. The bulk conductivity of the gold for the thinner circuit is measured as 2.18×10^7 U/m. Since the Main TL segments are low impedance (i.e. 37 Ω) the conductor loss for the thin gold circuit is significantly higher than its electroplated counterpart. This dramatically shrinks the coverage area. The coverage area for the circuit with the thin 150nm gold is presented in Figure 3-36. As can be seen, the 10 dB return loss circle is not covered entirely on the Smith chart at 3.4 GHz and 3.55 GHz. Therefore, this circuit is omitted from the measurement results.

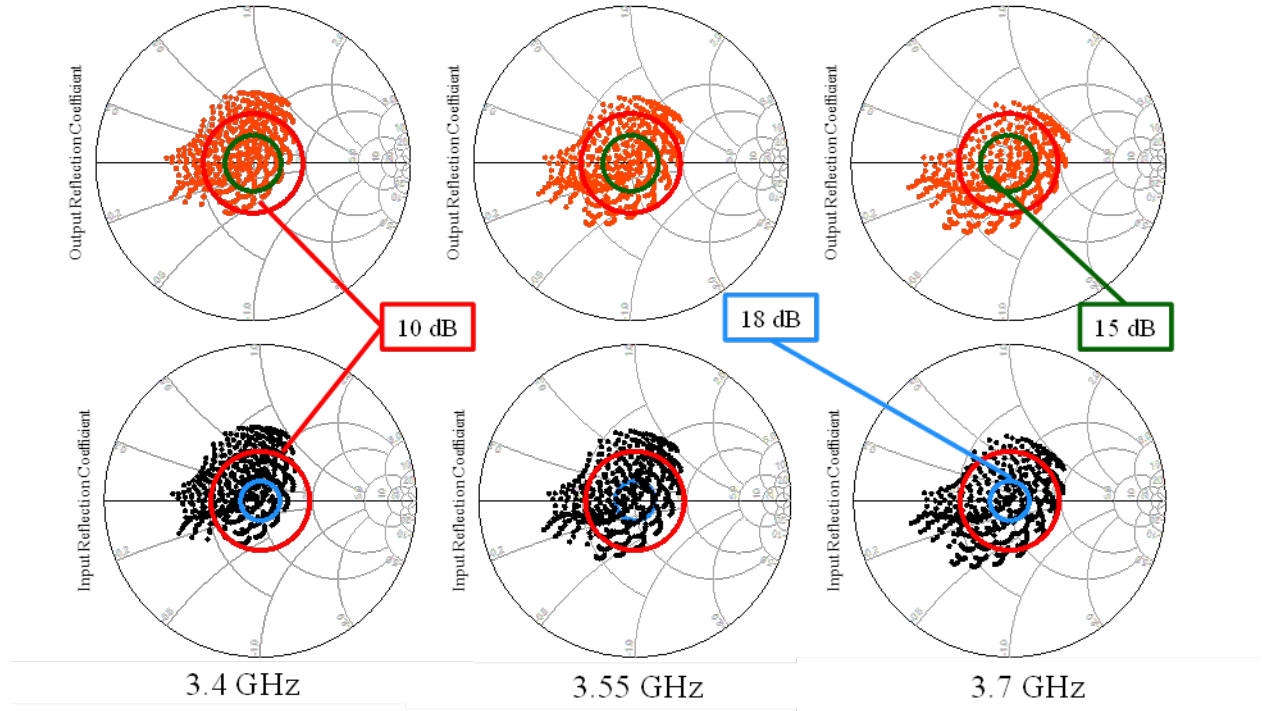
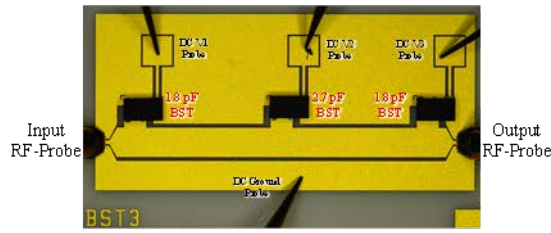


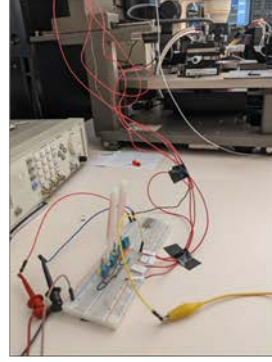
Figure 3-36 The degraded overage for the continuously tunable RIMN using BSTs with thin gold circuit.

3.4.2.1 Fabrication and Measurement Results

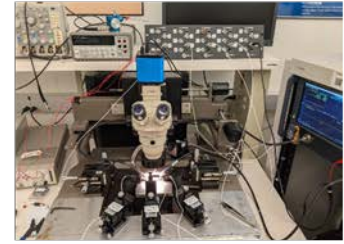
The circuit is fabricated by electroplating 3 μ m of 99.99% purity gold over a 150nm thin layer of gold, which is deposited on the substrate using e-beam evaporation. The BST capacitors are assembled using silver epoxy. The BSTs are independently controlled using DC probes. A single voltage source is set to 19V and is employed for the biasing. Then, three 10k Ω variable potentiometers are used to divide the source voltage into required voltages for each BST capacitor. Photos of the test setup and the DT-RIMN under test conditions are presented in Figure 3-37.



(a)



(b)



(c)

Figure 3-37 Fabricated and assembled DT-RIMN under test, (a) Device under test, (b) bias voltage control, and (c) DC and RF probes on the probe station.

A total number of 62 states are measured. For simplicity, the measurements are done only on integer voltage and the states are numbered by the individual voltages (e.g. 12-6-9, 18-18-18, and 3-12-6). The measured input and output reflection coefficients are plotted in Figure 3-38. For reference, the 15 dB and 18 dB circles are plotted as well. Although the actual coverage occupies a much larger area, these points clearly demonstrate the achievable coverage. Each state shown in Figure 3-38 is suitable for complex conjugately match a set of impedances beyond 40 dB of input and output (or one of them), which can be found using optimization.

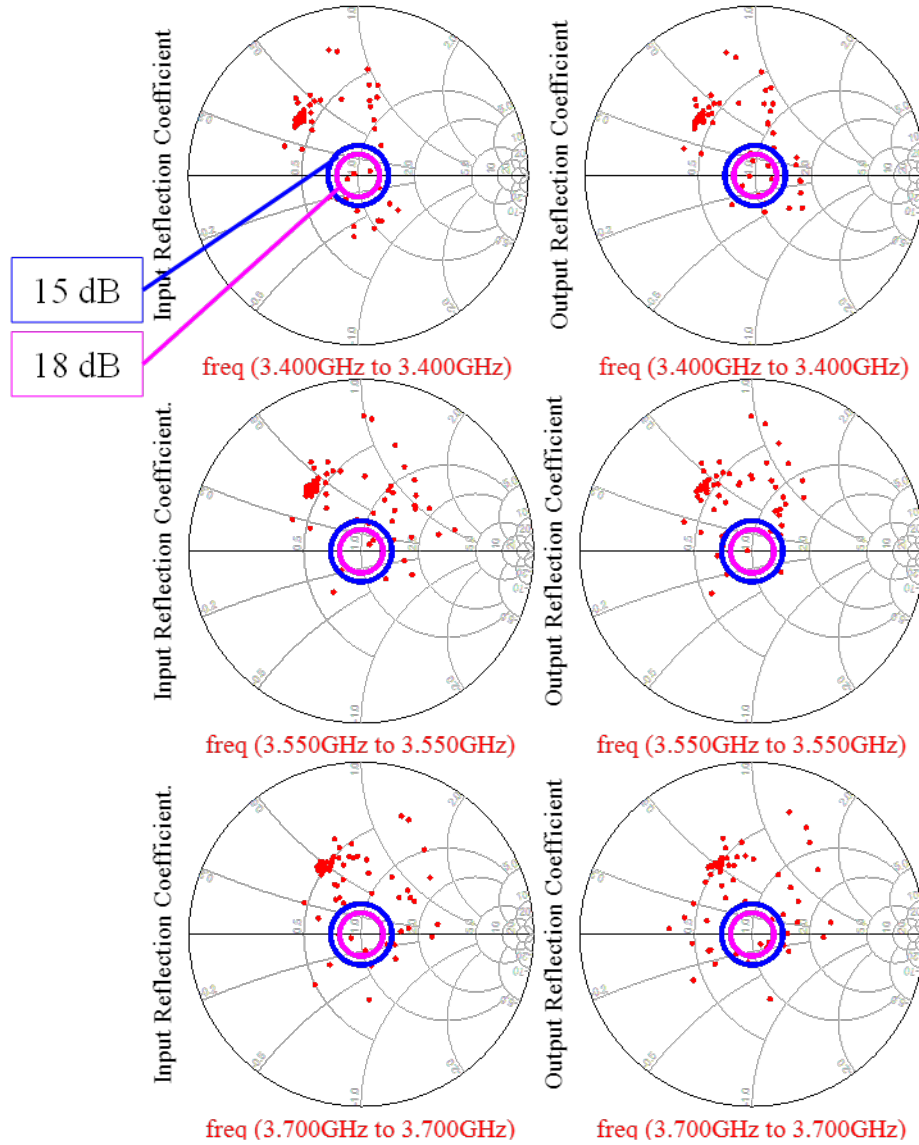


Figure 3-38 Measurement results for the BST DT-RIMN.

The performance of the device in terms of matched bandwidth and the insertion loss is studied by analysing a handful of states. The matching criteria is 40 dB of return loss for both the input and output. The state of the BST capacitors is not expected to determine the matched bandwidth. However, the insertion loss is determined by the quality factor of the BST capacitors. Here, lower bias voltages correspond to lower quality factors and higher losses for each BST. Therefore, the lowest insertion loss is associated with the 18-18-18 state, and the highest insertion loss is associated with the 2-2-2 state. Note that the 2V bias voltages are not needed to achieve the coverage and the 2-2-2 state is just studied as an example state with very high insertion loss. The optimized responses for these two states as well as a random state (i.e. 12-18-14) in between are depicted in Figure 3-39. As can be seen, the matched bandwidth is significantly wider

than 20 MHz, which makes the proposed design a suitable candidate for wideband applications. Moreover, an insertion loss of better than 0.6 dB is a reasonable expectation from this device. Note that the 2-2-2 state represents very good matching, but it shows about 5 dB of loss. This state does not correspond to capacitances within the design range. However, it is included in Figure 3-39 to highlight that although it is possible to get more than 40 dB matching, the insertion loss is determined by the quality factor of capacitors. The insertion loss of the RIMN predicts its use in practical applications.

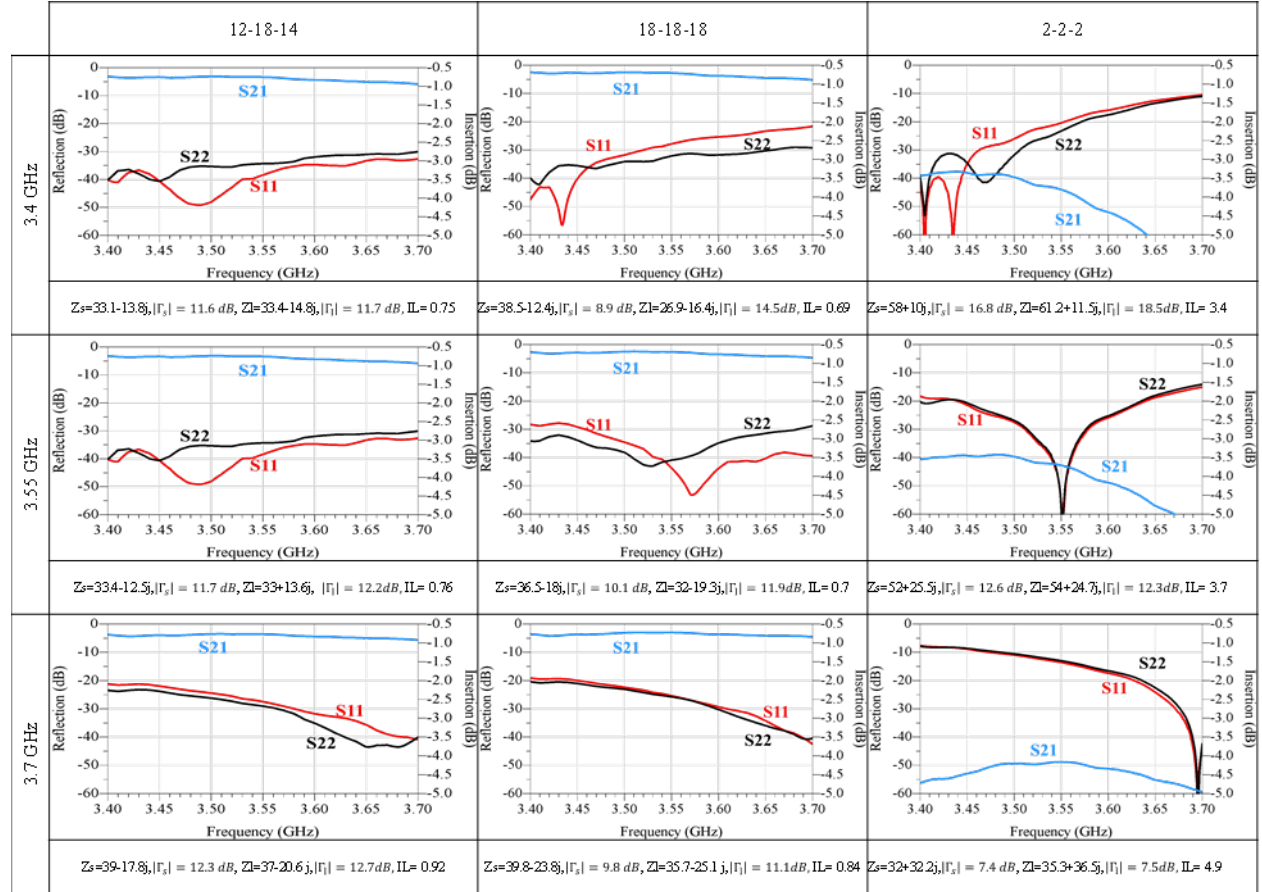


Figure 3-39 Optimized terminations for sample measured states. 12-18-14 is a random state, 18-18-18 is the low loss state, and 2-2-2 is an extreme case with exaggerated losses.

3.5 Improving the Circulator Isolation Using Doubly Terminated Reconfigurable Impedance Matching Networks

The network presented in the previous section is capable of suppressing the reflected signal from the antenna. However, the level of impedance matching between antenna and circulator does not improve the isolation between the ports of the circulator. In this section, we employ reconfigurable switched and continuously tunable networks to improve the effective isolation of circulators.

Figure 3-40 illustrates the circuit configuration to be studied. The main objective here are improving the isolation between transmit and receive ports. These ports are marked by Tx and Rx respectively. The main objective should be achieved while maintaining the input impedance matching at Tx port, as well as optimal insertion loss for Tx-to-antenna and antenna-to-Rx paths. The general specifications for this problem are as follows.

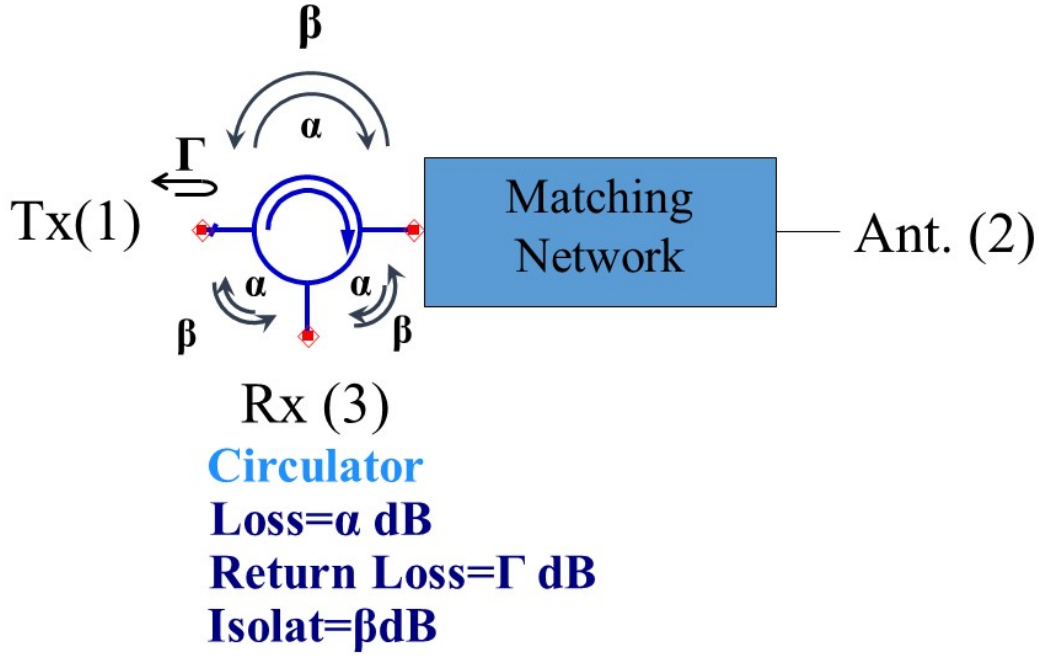


Figure 3-40 General blocks at the front-end

The circulator has performance parameters α , β , and Γ , where α is the circulator insertion loss, β is the circulator isolation, and Γ is the circulator port return. The phases of α , β , and Γ parameters are generally un-known and may differ on a device-by-device basis. Moreover, these parameters are frequency dependant. The S-parameters and reflection coefficients are illustrated in Figure 3-41. The scattering matrix for a circulator is expressed as:

$$\begin{bmatrix} v_1^- \\ v_2^- \\ v_3^- \end{bmatrix} = \begin{bmatrix} \Gamma & \beta & \alpha \\ \alpha & \Gamma & \beta \\ \beta & \alpha & \Gamma \end{bmatrix} \begin{bmatrix} v_1^+ \\ v_2^+ \\ v_3^+ \end{bmatrix}. \quad 3-18$$

In addition, the reflection coefficient of the RIMN is marked as $\gamma = \frac{v_2^+}{v_2^-}$, which is the transformed input impedance of the antenna by the RIMN. Conventional matching network design intends to achieve $\gamma = \Gamma^*$ for maximum power transfer. Note that a circulator is a non-reciprocal circuit element and therefore can be matched at all ports. However, [3], [50] show that this architecture can be modified for perfect cancellation

of the leakage path from Tx to Rx. For this purpose, assume the Rx is matched and the input is applied to the Tx port. This leads to:

$$v_2^- = \alpha v_1^+ + \Gamma v_2^+ \quad 3-19$$

$$v_3^- = \beta v_1^+ + \alpha v_2^+ = \beta v_1^+ + \alpha \gamma v_2^- \quad 3-20$$

Moreover, 3-19 can be rewritten as:

$$v_2^- = \frac{\alpha}{1-\Gamma\gamma} v_1^+ \quad 3-21$$

On the other hand, the isolation between Tx and Rx is defined as:

$$Iso = \frac{v_3^-}{v_1^+} = \frac{\beta v_1^+ + \alpha \gamma v_2^-}{v_1^+} \quad 3-22$$

Now substituting 3-21 in 3-22 leads to:

$$Iso = \beta + \frac{\gamma \alpha^2}{1-\Gamma\gamma} \quad 3-23$$

For perfect isolation between the Tx and Rx ports, we set 3-23 equal to zero, which leads to:

$$\gamma = \frac{\beta}{\beta\Gamma - \alpha^2} \quad 3-24$$

The presented calculations up to 3-24 is presented in [3], [50]. We can further extend the calculations by considering the fact that β and Γ are very small and their product will be negligible compared with α^2 . Therefore, 3-24 further simplifies to:

$$\gamma \cong \frac{-\beta}{\alpha^2} = \frac{-|\beta|}{|\alpha|^2} e^{-j(\angle\beta - 2\angle\alpha)} \cong -|\beta| e^{-j(\angle\beta - 2\angle\alpha)} \quad 3-25$$

In deriving 3-25, it is assumed that isolation is significantly smaller than the insertion loss and $|\alpha|^2$ roughly equals unity. This is a valid assumption for almost all commercially available over the counter circulators. Circulator datasheets usually contain the worst insertion loss and isolation values over the suggested frequency range of operation. Phase information are usually unavailable and are derived using characterization. For example, both the circulators presented in Figure 3-2 show an isolation better than 20 dB and an insertion loss of better than 0.274 dB. In the following experiments, we consider a 2dB margin for the isolation and considered 18 dB as the worst case. In addition, the insertion loss is rounded to 0.25 dB. Ensuring cancelation over the frequency range of operation requires meeting 3-25 for all β values below these values and for all phases. In other words, the device should demonstrate coverage for the entire $|\beta_b|$ circle, where β_b is the worst reported isolation value over the frequency range.

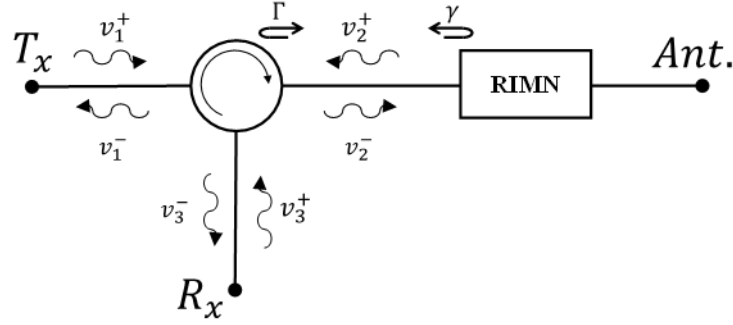


Figure 3-41 S-parameter representation for circulator isolation problem.

The 3-25 condition does not lead to perfect matching. However, its impact on the input impedance of Tx and Rx ports can be compared with the scenario in which all ports are complex conjugate matched. In this scenario when all ports of the circulator are matched, reflection coefficient for all the ports is written as:

$$\Gamma_{in} = \Gamma \Rightarrow Z_{in} = Z_0 \frac{1+\Gamma_{in}}{1-\Gamma_{in}} \quad 3-26$$

Since $\Gamma \neq 0$, some mismatch is inevitable. For example, for $\Gamma = 20\text{dB} = 0.1$, $Z_{in} = 1.22Z_0$ or $Z_{in} = 61\Omega$ for $Z_0 = 50\Omega$. As another example, for $\Gamma = 18\text{dB} = 0.124$, $Z_{in} = 1.28Z_0$ or $Z_{in} = 64\Omega$ for $Z_0 = 50\Omega$.

Now, when $\gamma = \frac{-\beta}{\alpha^2}$, input reflection coefficient at Tx and Rx ports can be calculated as follows. First, assume Rx port is matched. Then, we can write:

$$v_1^- = \Gamma v_1^+ + \beta v_2^+ \quad 3-27$$

$$v_2^+ = \gamma v_2^- = \frac{\gamma \alpha v_1^+}{1-\gamma \Gamma} \quad 3-28$$

Correspondingly,

$$v_1^- = \Gamma v_1^+ + \beta \frac{\gamma \alpha}{1-\gamma \Gamma} v_1^+ \quad 3-29$$

$$\Gamma_{in@Tx} = \Gamma + \frac{\beta \frac{-\beta}{\alpha^2} \alpha}{1-\Gamma \frac{-\beta}{\alpha^2}} = \Gamma - \frac{\beta^2 \alpha}{\alpha^2 - \Gamma \beta} \simeq \Gamma - \frac{\beta^2}{\alpha} \quad 3-30$$

A similar expression can be written for Rx port as well. In derivation of 3-30, the $\Gamma\beta$ term in the denominator is assumed negligible compared to α^2 and thus omitted. Note that the 3-25 condition will not significantly degrade the Tx and Rx matching, as long as Γ and β are small and α is close to unity.

As a proof of concept a test circuit is fabricated and assembled, which consists of an RIT between the output of a circulator and the antenna port. This is shown in Figure 3-42. The circuit uses the same RIT described in 3.4.2.

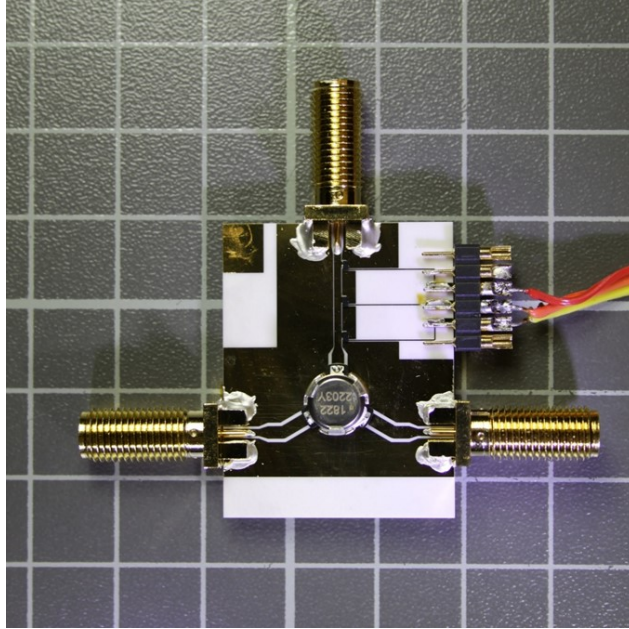


Figure 3-42 The circuit used to enhance the isolation between Tx and Rx

The circuit is terminated in a load impedance comprising a transmission line segment, an attenuator, and a wideband load termination. This configuration allows mimicking different antenna input impedances. a photo of an example load impedance, its input reflection coefficient (around 15 dB between 3.4 GHz and 3.7 GHz), and phase are shown in Figure 3-43. Figure 3-44 shows the measurement results for the return losses at Tx and Rx ports, as well as isolation between the two ports for two optimized states. These states are optimized to achieve maximum isolation at 3.40 GHz and 3.55 GHz respectively. As can be seen, the return loss of the Tx and Rx ports do not change significantly between these two states.

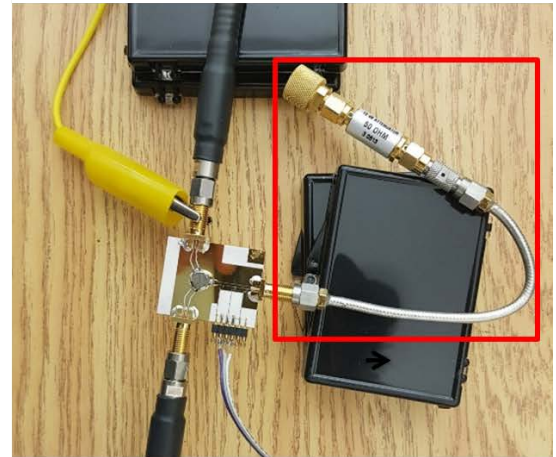
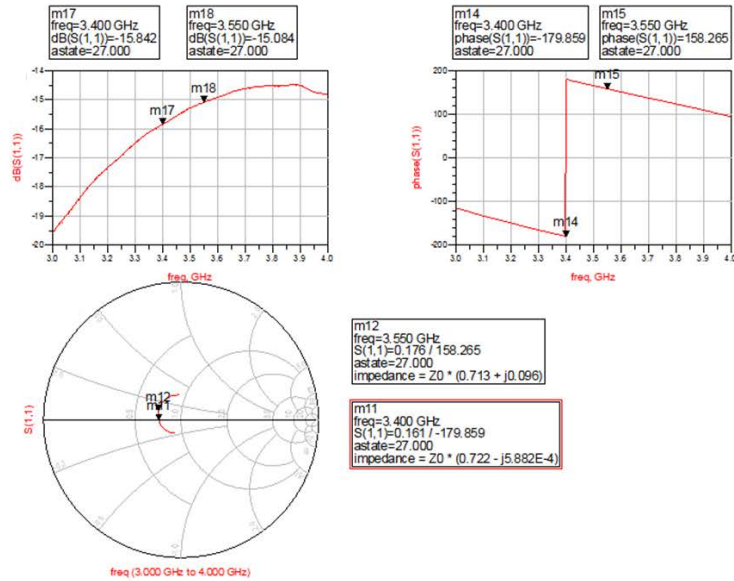


Figure 3-43 Terminating the circuit with a predetermined load impedance.

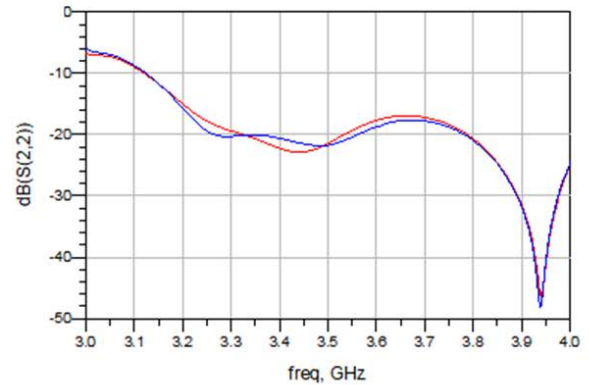
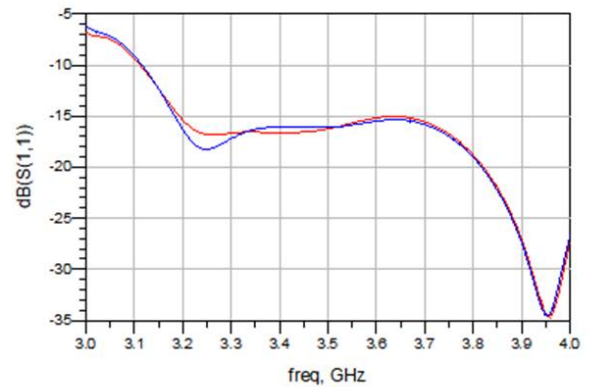
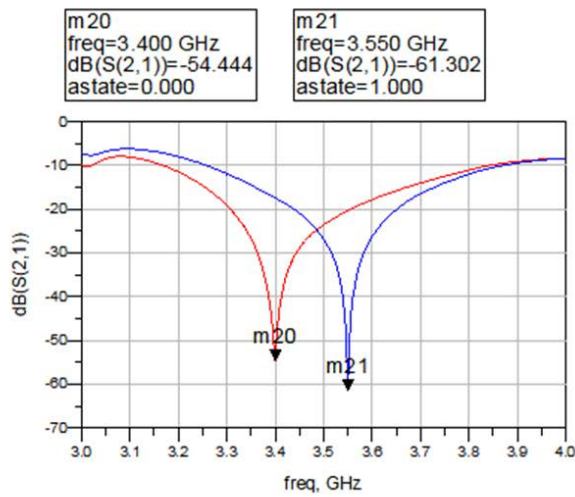


Figure 3-44 Measurement results for improving the isolation between Tx and Rx using a RIT.

3.5.1 Fully Integrated and Miniaturized Switched-stubs RITs Using PCM Switches

In this section, the analyses presented in 3.5 are used to develop a tuner for improving the isolation. For this purpose, both continuously tunable and switched designs can be employed. However, the operational frequency band for the continuously tunable designed is subject to the performance of the varactors. In other words, as the frequency increases, the quality factor of varactors decrease and the tuners become excessively lossy. Correspondingly, depending on the performance of the RF switches, switched designs may have a superior performance. In the section 3.3, SPST switches were used on stubs to create short circuit terminations at different lengths. However, this method requires quarter- and half- wavelength transformers, which not only results in limited bandwidth, but also makes these designs physically large. Alternatively, it is possible to employ multiport switches to select between different stubs instead of using SPSTs to terminate a stub at different lengths. This method allows for loading the main branch with independent impedances (capacitive and inductive) since these stubs can have different lengths and independent characteristic impedances. Moreover, it is possible to employ both lumped and distributed elements in this design. In this section, we investigate the performance of switched-stub designs and employ them for improving the isolation between Tx and Rx paths.

Die attach RF-switches can be conveniently assembled to circuits using flip-chip or wire bonds techniques. Another option is to integrate the entire circuit, which results in significant miniaturization. However, semiconductor RF-switches require an additional bias network circuitry, which significantly increases the design complexity and makes it susceptible to failure. Moreover, the bias networks usually take up a considerable area, which hinders the miniaturization purpose of using unpackaged switches in the first place. Recent breakthroughs regarding PCM technology suggest an alternative category of RF-switches that are extremely miniaturized, more linear than semiconductor counterparts, and very low loss. In addition, these switches are controlled using pulses and do not require DC bias networks. This gives them a considerable advantage over the semiconductor switches. Moreover, unlike RF-MEMS switches, they do not need high DC voltages for operation. Recently, Ge-Te based RF PCM switches with more than one million successful activation cycles have been reported [9]. In this section, these switches are employed in realizing compact, fully integrated, linear, passive, low-loss, wideband, and miniaturized RITs. The application of these circuits in improving the circulator isolation is also studied.

3.5.1.1 Realising Multi-branch Stubs Using Miniaturized PCM RF-switches

A significant advantage of the PCM switches is their miniaturized size. A PCM SP3T switch occupies an area of only $76\text{ }\mu\text{m}$ by $76\text{ }\mu\text{m}$, which includes the control pads, while a single SPST switch takes only 20

μm by $20\ \mu\text{m}$ of area. This extreme size adds a degree of freedom to the switched RIT design, which is the location of the switch along the stubs. Figure 3-45 depicts a transmission line model for multi-branch stubs, where (a) and (b) represent cases where the switch is located near the end and beginning of the stub length, respectively. Therefore, the length of the stub for each state comprises of a length of d , which is common among all states, the equivalent electrical length of the switch and additional extension lengths beyond the switches. The length of the different states are marked as l_1 , l_2 , and l_3 . Although these lengths are similar for case (a) and (b), case (a) has an advantage in terms of miniaturization because it takes less amount of space. Moreover, the circuit (a) require shorter bias lines. Furthermore, the bias lines are farther away from the main line, which results in lower perturbation for the main line. Figure 3-45 (c) shows a circuit model for any of theses states. The advantage of this model is that if d is close to l_i (i.e. the switch is located toward the end of the stub length), and the common TL is high impedance, then conductive losses can be greatly minimized. This leads to high quality factor capacitances if the stubs are open ended and their collective lengths, l_i , is less than 90° .

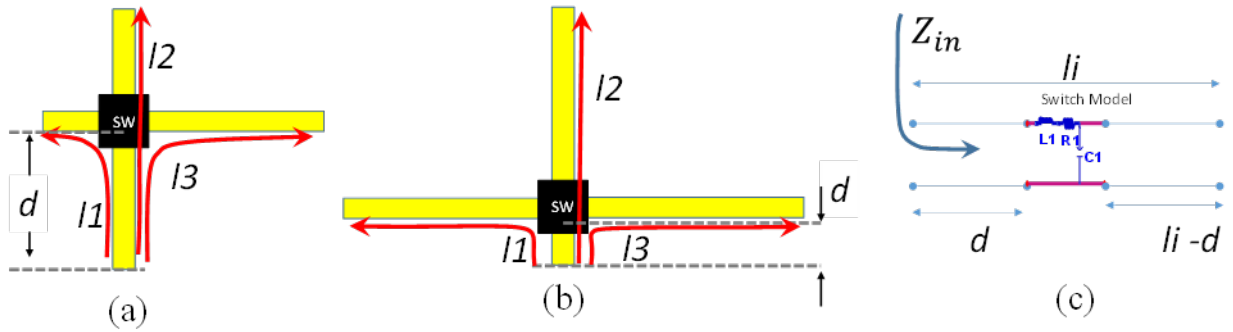


Figure 3-45 Location of the switch along a multi-branch stub, (a) longer length before the switch, (b) short length behind the switch, and (c) equivalent circuit.

Another advantage of the multi-branch stubs compared to the continuously tunable capacitors is associated with the capacitance/ quality factor relationship. The characterization of the BSTs shows that the quality factor decreases as the capacitance increases. However, this relationship can be averted if the switch can connect multiple branches to the common TL segment at the same time (e.g. if the switch in Figure 3-45 can connect all three stubs at the same time). In this case, if the input impedance of the branches are capacitive, their paralleled combination will have higher quality factor. This is because the total capacitance increases, while series resistance decreases for a parallel combination of capacitors. Figure 3-46 depicts the characterization results for a multi-state capacitor that is simulated using 450nm of gold thickness over 500 μm high resistivity silicon (HRSi). The Smith chart shows the reflection coefficient seen from the input port of the SP3T PCM switch. Measured S-parameters are used to model the SP3T switch.

The switched capacitor has 8 states, which are identified by variable $n1$. State $n1=1$ is OFF state for all switches, states 2, 3, and 4 are for when only one of the switches are ON, states 5, 7, and 8 are for when two of the switches are ON, and the state 6 is for when all are ON. As can be seen, state 1 has the lowest quality factor and capacitance, while state 6 has the highest quality factor and capacitance. The limited quality factor in these characterizations are associated with the degraded bulk conductivity of the thin gold layer [9], [43], [69]. This can be improved significantly by electroplating a thicker layer of gold on top of the fabricated devices.

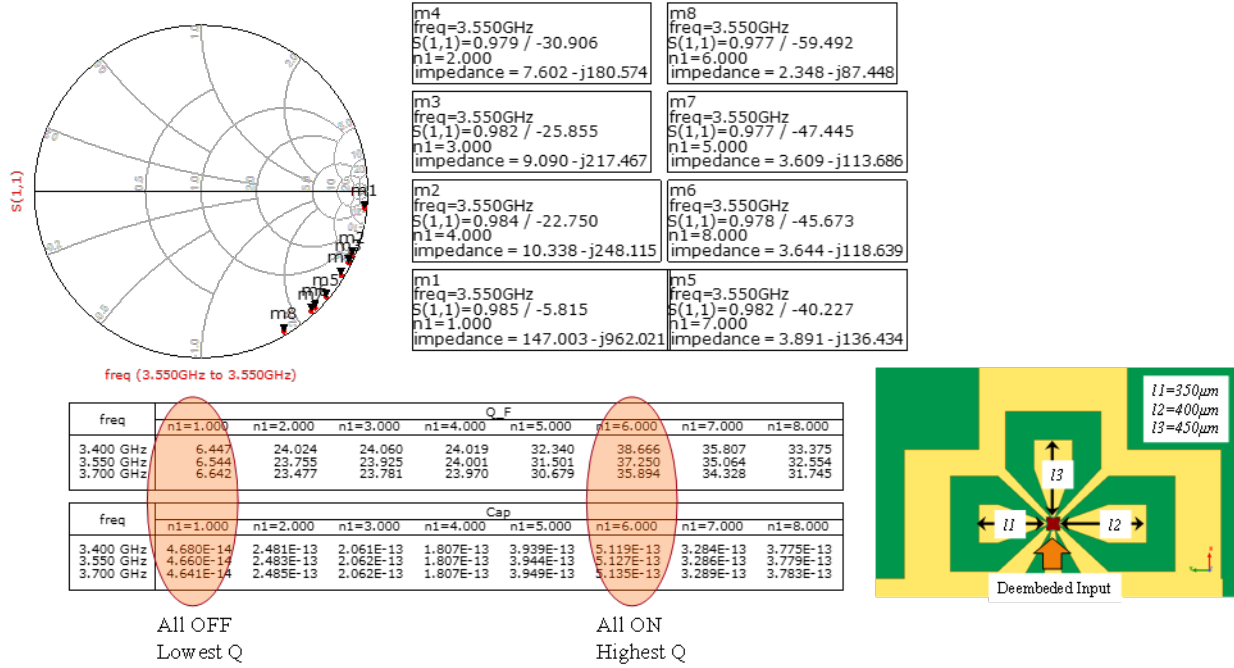


Figure 3-46 Characterization of a switched TL multi-state capacitor.

3.5.1.2 Simulation Results for Switched-stub PCM Based Loaded Lines

As studied in section 3.5, a loaded line circuit is designed to achieve coverage for the worst-case reported circulator isolation of 20 dB. Here, a 2 dB margin is considered and the circuit is designed for achieving coverage over 18dB circle. The proposed circuit is composed of a main TL line, which is loaded by PCM SP3T multi-state capacitors on both sides of the line. Figure 3-47 compares the achievable coverage for 4- and 6- tuning element loaded lines. It also includes a constant γ circle, which corresponds to $\gamma = -\beta/\alpha^2$ for $|\beta| = 18 \text{ dB} = 0.1259$ and $|a| = 0.25 \text{ dB} = 0.971$. These circuits are optimized to achieve a uniform coverage over the centre of the Smith chart. As can be seen the 4-element circuit cannot cover the constant γ circle over the frequency range of 3.4 GHz to 3.7 GHz. Note that the 4-element circuit can deliver $8^4 = 4096$ states, while the 6-element design can deliver $8^6 = 262144$ states. Generating coverage plots for the

6-element circuit in ADS takes an high amount of time and resources. As a result, only 25000 states are shown for the 6-element circuit in Figure 3-47 to demonstrate the coverage. The reader can rightfully speculate a much wider and uniform coverage.

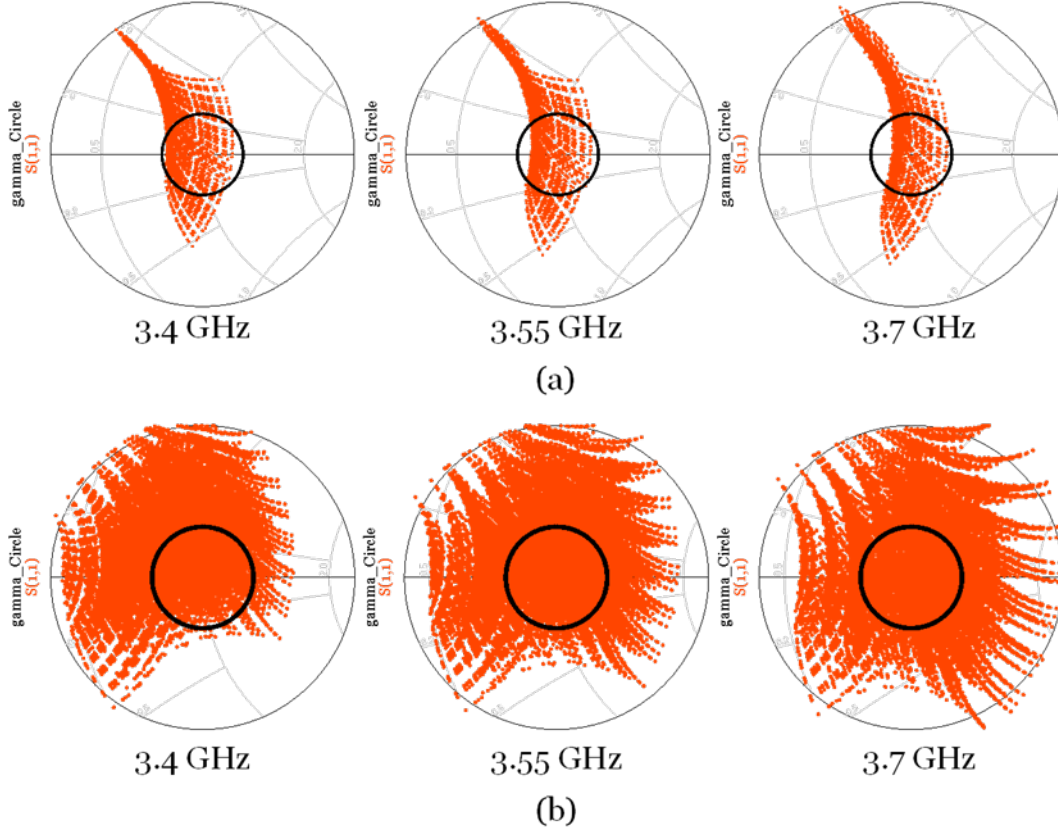


Figure 3-47 Comparing the achieved coverage for optimized circuits, (a) for 4 tuning element loaded line, and (b) for 6 tuning element loaded line. gamma_circle has a radius of β/α^2 .

The layout of the proposed PCM based loaded line using CPW lines is drawn in Figure 3-48 and its optimized design parameters are listed in Table 3-5. The circuit is composed of 3 pair of tuning elements that are separated by two TL segments. Each pair is identical and are attached to both sides of the main TLs. Note that the tuning element pairs near the input and out put are rotated. This is done so that a similar set of states can be achieved regardless of the choice of input and output ports. Accordingly, the state of operation is identified by a 6-digit number. Each digit can take any integer between 1 and 8.

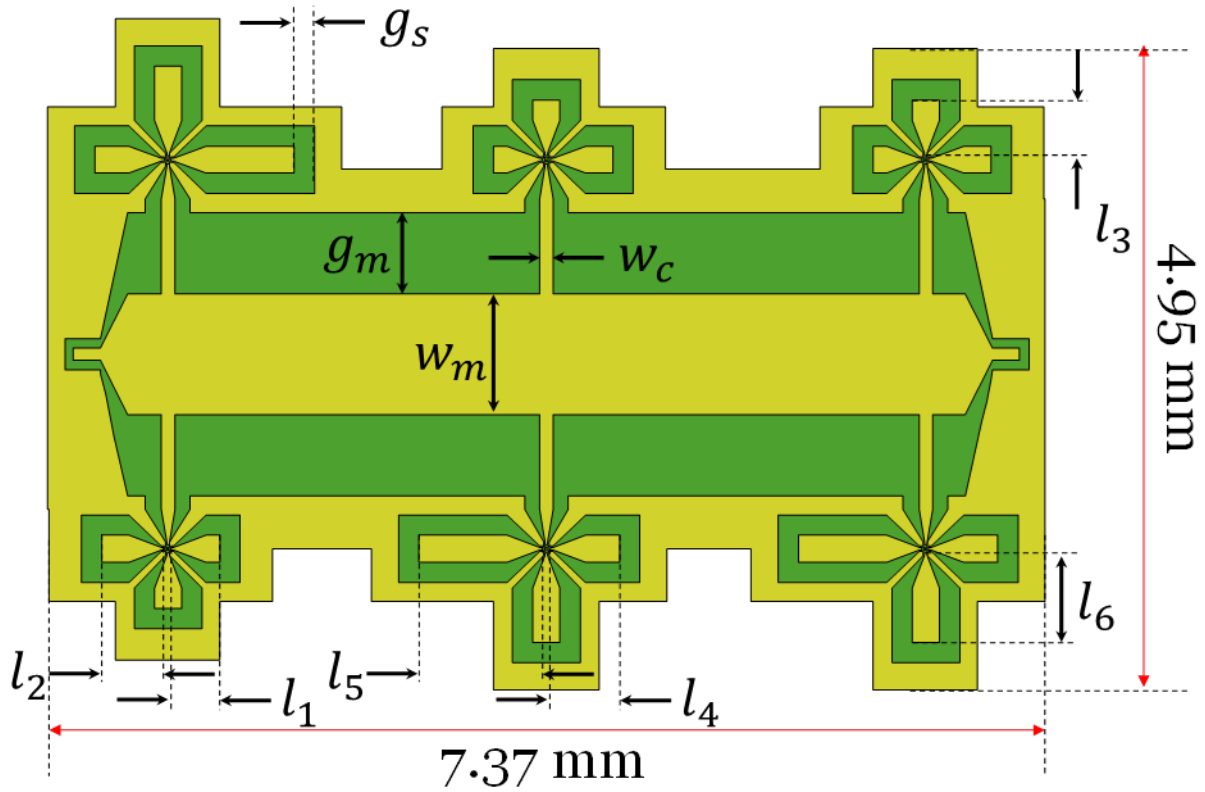


Figure 3-48 Schematic of the proposed switched-stub RIT based on PCM SP3Ts

Table 3-5 Optimized dimensions of the PCM based RIT.

Parameter	Value (μm)	Parameter	Value (μm)
w_m	900	l_2	400
g_m	600	l_3	450
g_s	150	l_4	500
w_c	100	l_5	650
l_1	350	l_6	900

The full-wave simulation results of the proposed RIT is used to demonstrate its capability in improving the isolation between Tx and Rx ports of a single antenna front-end. The circuit model used is depicted in Figure 3-49. The left side of the circuit is a circulator model. For simplicity, the parameters for this model (i.e. α , β , and Γ) are assumed constant over the frequency range and are set to the worst reported values by the manufacturer. The circuit on the right represents the designed RIT, which is terminated by an antenna with a 15 dB return loss. The internal ports 4 and 5 are used to ensure 3-25 condition is met. Then these ports are removed and the input of the RIT is connected to the output of the circulator and the S-parameters for the resulting 3-port network are reported.

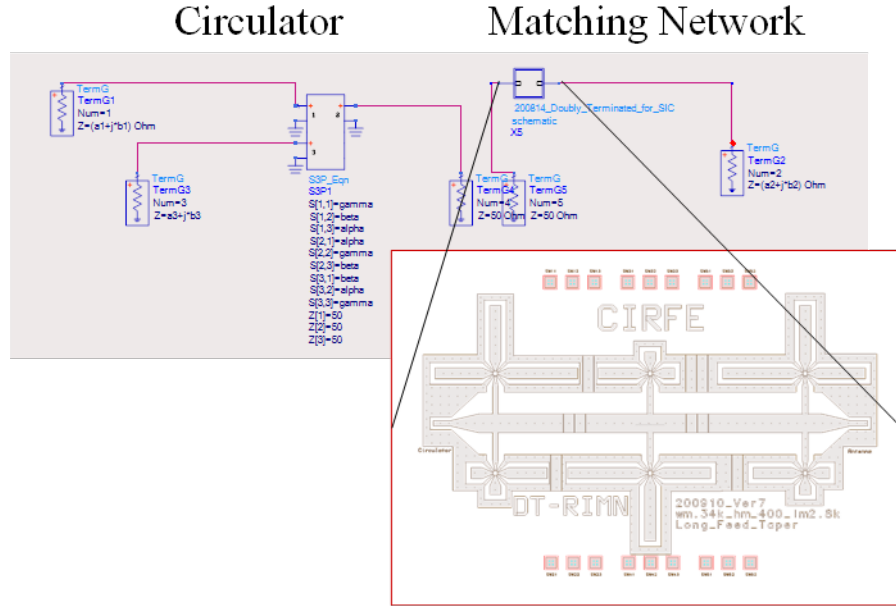


Figure 3-49 Circuit model to improve the isolation between Tx (port 1) and Rx (port 3) using the proposed PCM based switched RIT.

The variables to demonstrate the performance of the proposed circuit are the antenna input impedance, circulator isolation, and its reflection coefficient. Accordingly, the magnitude of these variables are set to the worst-case scenario and their phases are swept. For each variable, the performance at 4 phases that are 90° apart are studied and the state of the RIT that results in better results and its results are reported. Collectively these results are reported in a tabular form using Figure 3-50 to Figure 3-56. The antenna impedance is swept similar to section 3.4.1 and each column of the table represents one of the four antenna input impedances. The values for γ are reported on the vertical axis and each row is 90° apart. α is assumed constant for all frequencies with a value of $0.25 \text{ dB} = 0.9715$. Phase of α is set to 0° because it does not influence γ . Moreover, $|\beta|$ is set to $18 \text{ dB} = 0.126$. Note that the phase of β is the only variable that determines the phase of γ . Figure 3-50 to Figure 3-53 demonstrate the achieved isolation and the corresponding state for different phases of Γ as can be seen the isolation of more than 40 dB is achieved for the majority of the states. The few cases where isolation target is not achieved is due to local minima in optimization algorithm and can be resolved using a different initial state. Figure 3-54 to Figure 3-56 demonstrate the achieved isolation at three frequencies for $|\beta| = 20 \text{ dB}$. As can be seen the 40 dB is achieved in all cases and the achieved isolation is higher than the cases with $|\beta| = 18 \text{ dB}$. The value of β is changed to demonstrate that 18 dB is the worst case and the results shown in Figure 3-50 to Figure 3-53 are the minimum improvement in isolation that can be achieved.

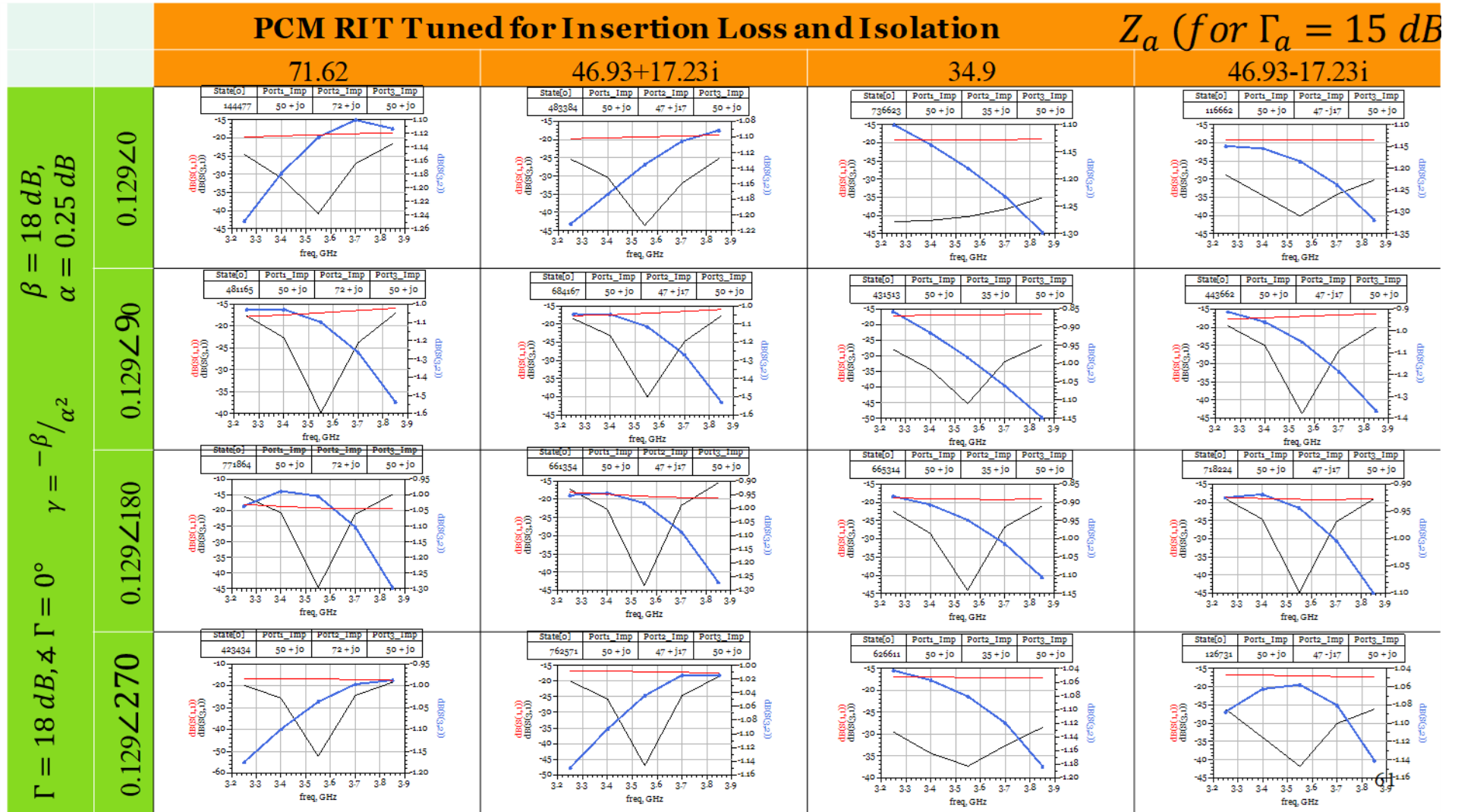


Figure 3-50 Optimized simulation results for $\alpha = 0.25$ dB, $\Gamma = 18$ dB, $\angle \Gamma = 0^\circ$, $|\beta| = 18$ dB, and $|\Gamma_a| = 15$ dB. Phases of antenna reflection coefficient and circulator isolation are swept.

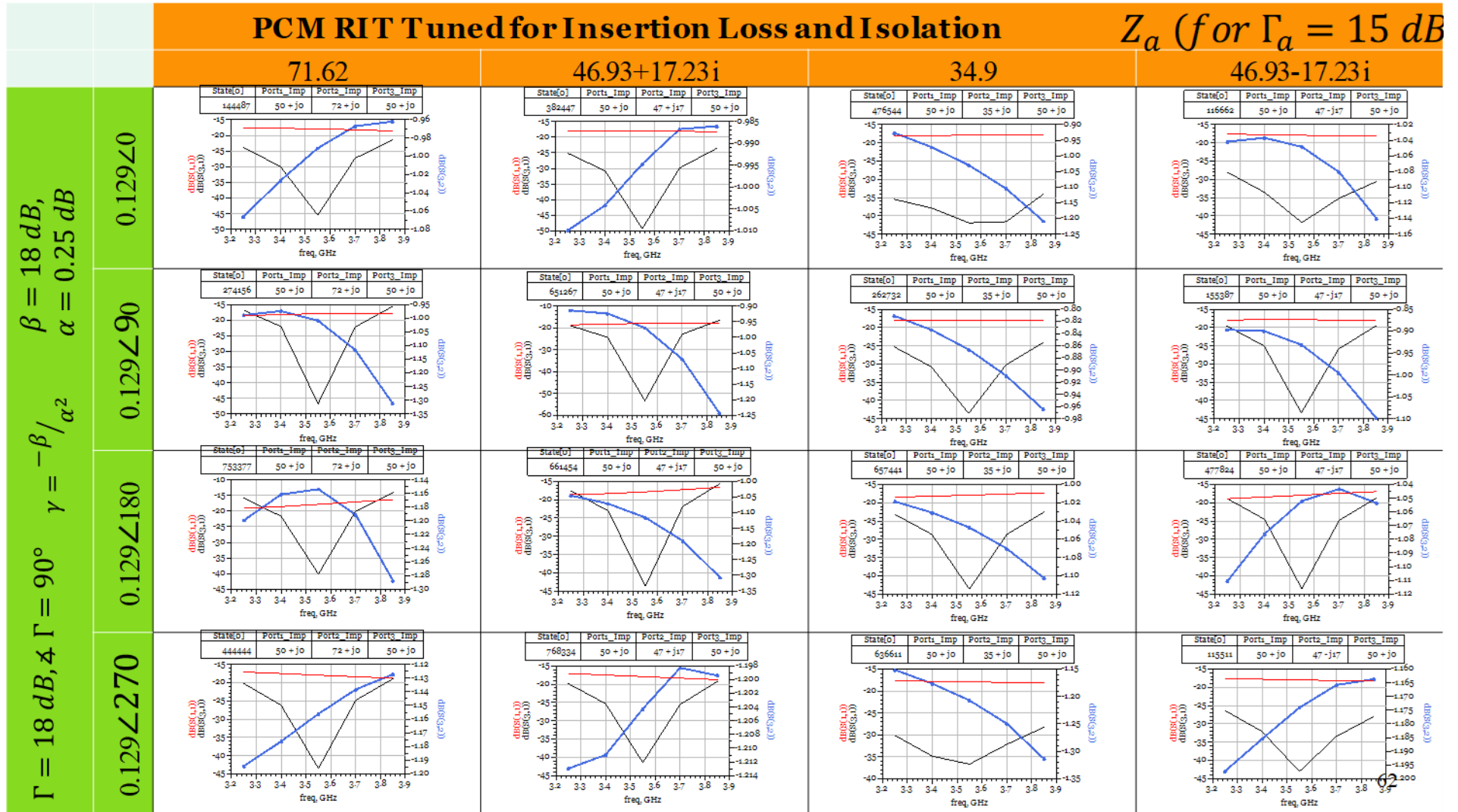


Figure 3-51 Optimized simulation results for $\alpha = 0.25$ dB $\Gamma = 18$ dB, $\angle \Gamma = 90^\circ$, $|\beta| = 18$ dB, and $|\Gamma_a| = 15$ dB. Phases of antenna reflection coefficient and circulator isolation are swept.

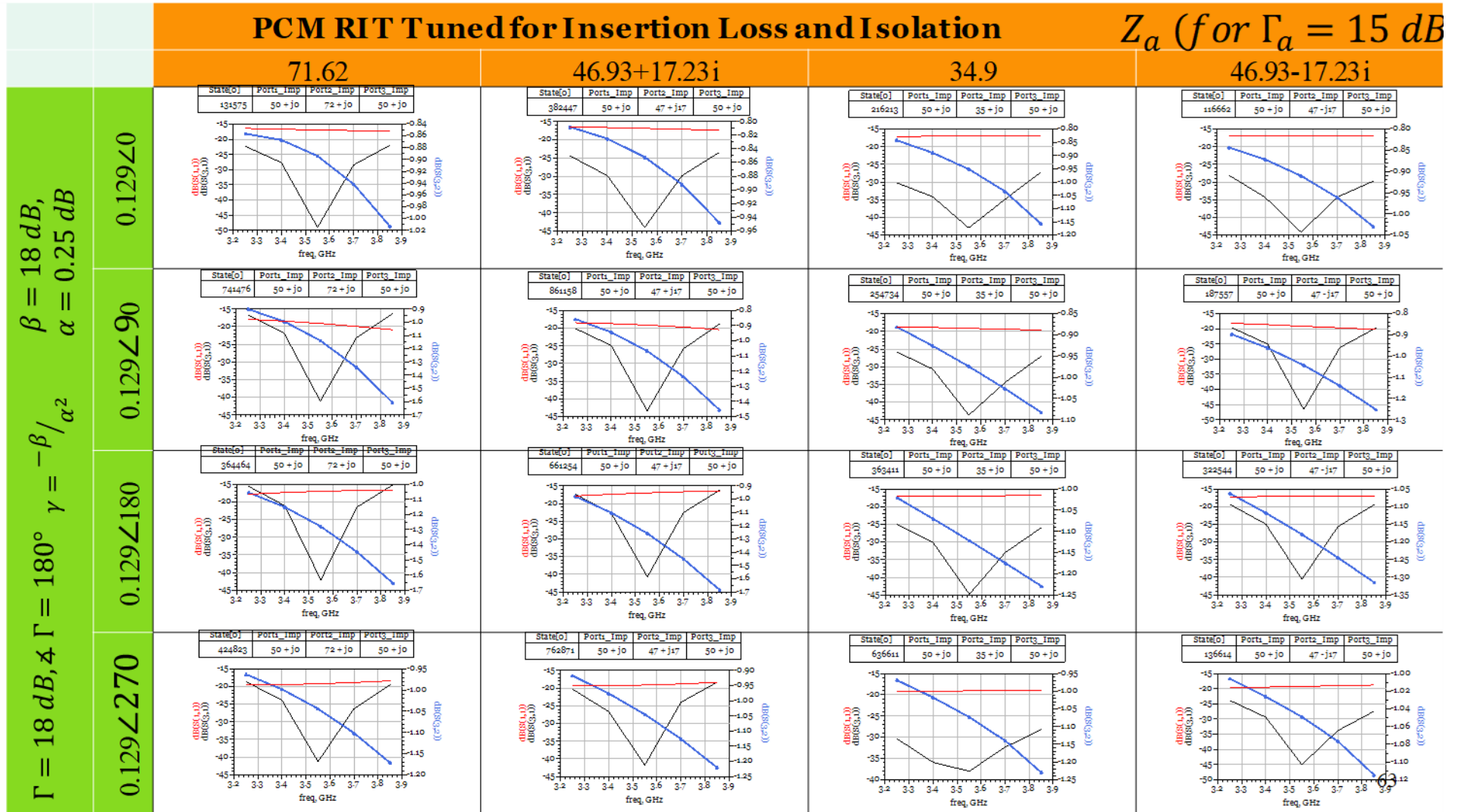


Figure 3-52 Optimized simulation results for $\alpha = 0.25$ dB $\Gamma = 18$ dB, $\angle \Gamma = 180^\circ$, $|\beta| = 18$ dB, and $|\Gamma_a| = 15$ dB. Phases of antenna reflection coefficient and circulator isolation are swept.

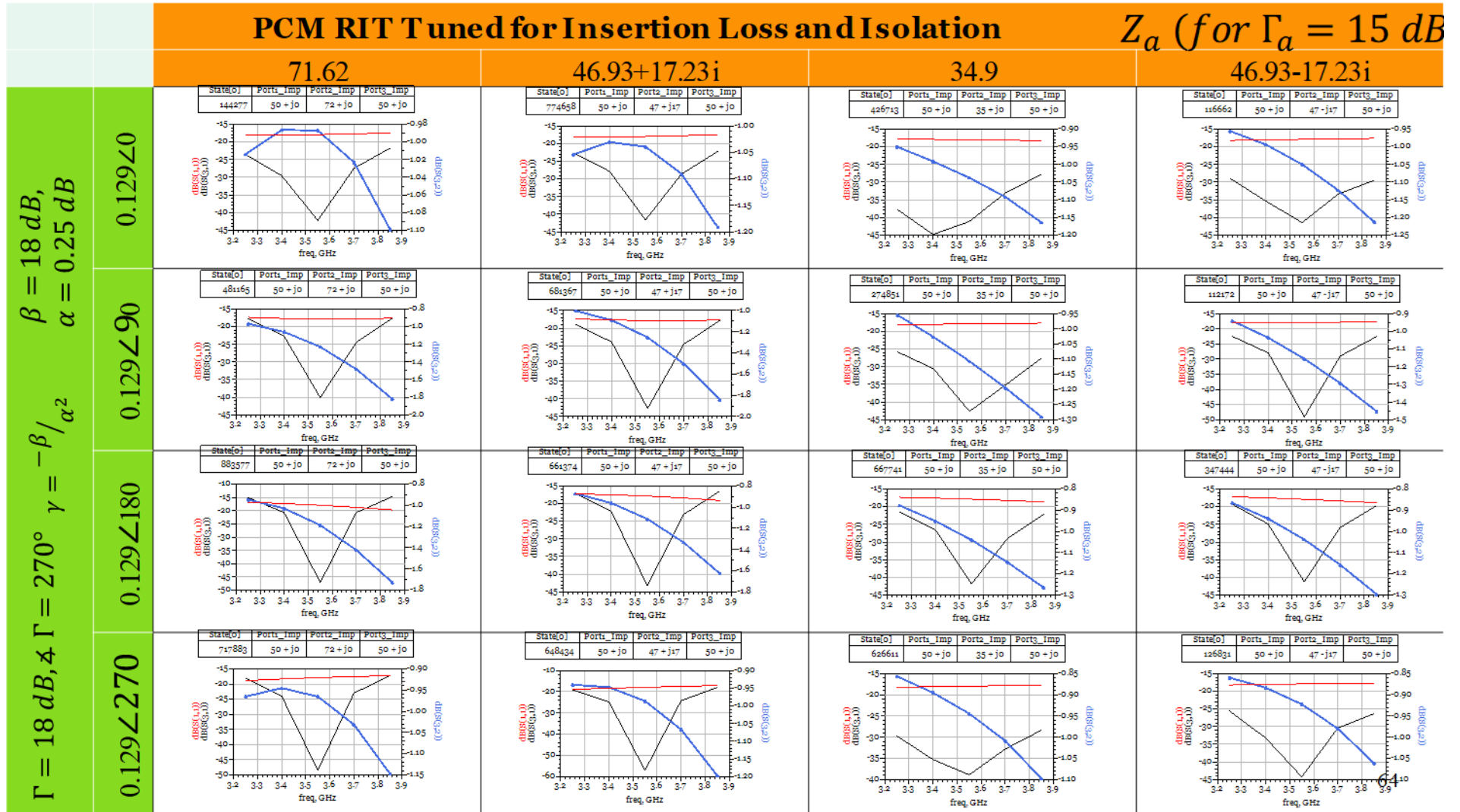


Figure 3-53 Optimized simulation results for $\alpha = 0.25$ dB $\Gamma = 18$ dB, $\angle \Gamma = 270^\circ$, $|\beta| = 18$ dB, and $|\Gamma_a| = 15$ dB. Phases of antenna reflection coefficient and circulator isolation are swept.

3.4 GHz

PCM RIT Tuned for Insertion Loss and Isolation

Z_a (for $\Gamma_a = 15 \text{ dB}$)

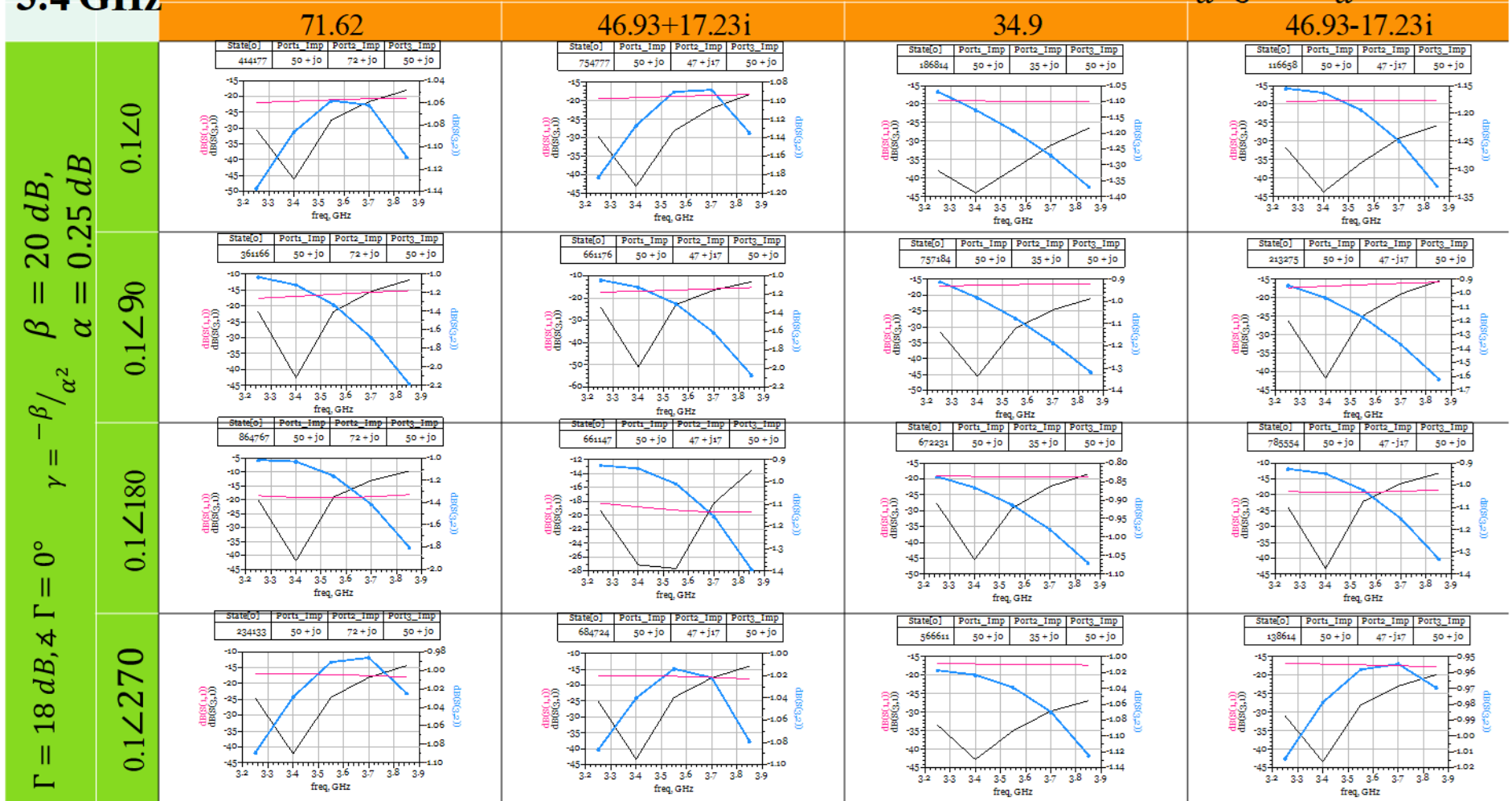


Figure 3-54 Optimized simulation results at 3.4 GHz for $\alpha = 0.25 \text{ dB}$, $\Gamma = 18 \text{ dB}$, $\angle \Gamma = 0^\circ$, $|\beta| = 20 \text{ dB}$, and $|\Gamma_a| = 15 \text{ dB}$. Phases of antenna reflection coefficient and circulator isolation are swept.

3.55 GHz

PCM RIT Tuned for Insertion Loss and Isolation

Z_a (for $\Gamma_a = 15 \text{ dB}$)

$$\Gamma = 18 \text{ dB}, \angle \Gamma = 0^\circ \quad \gamma = -\beta / \alpha^2 \quad \beta = 20 \text{ dB}, \quad \alpha = 0.25 \text{ dB}$$

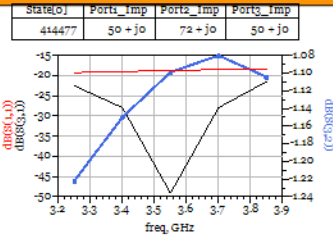
0.1 $\angle$ 0

0.1 $\angle$ 90

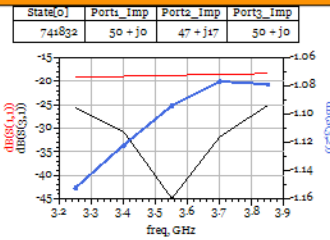
0.1 $\angle$ 180

0.1 $\angle$ 270

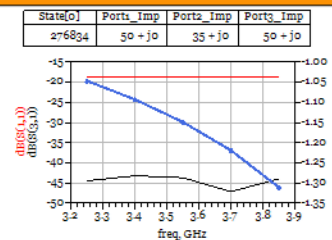
71.62



46.93+17.23i



34.9



46.93-17.23i

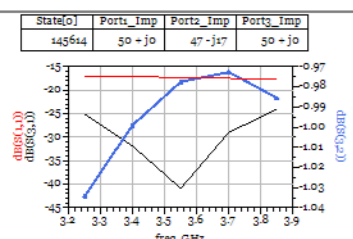
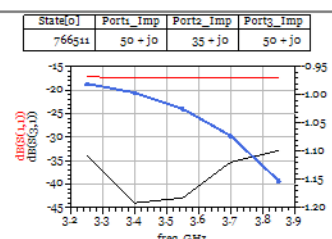
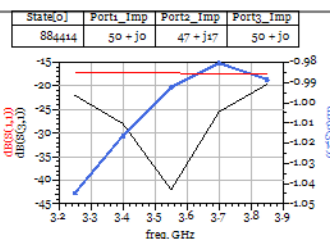
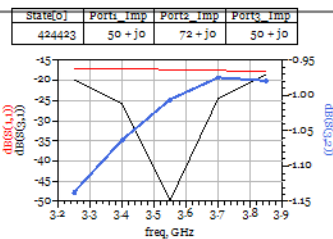
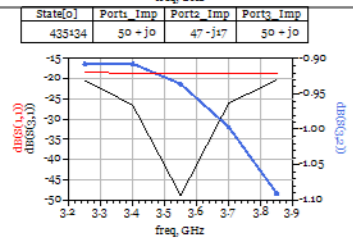
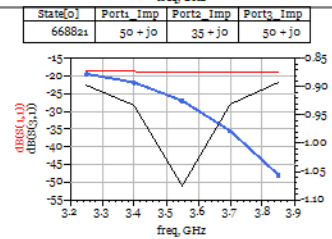
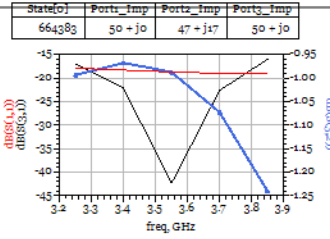
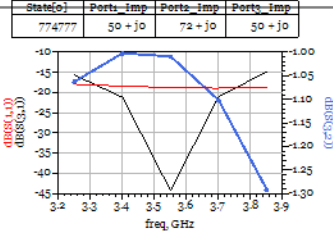
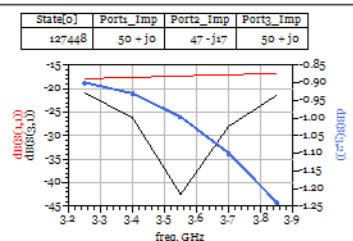
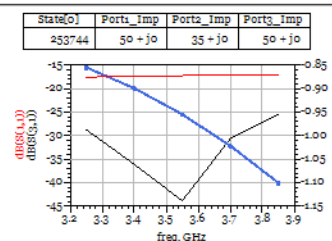
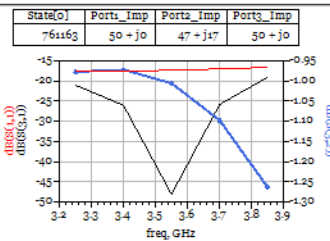
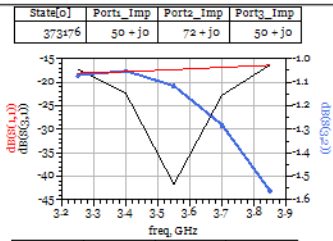
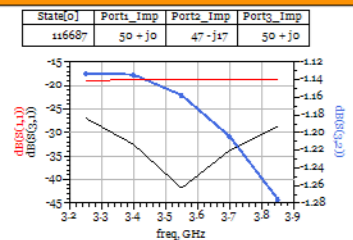


Figure 3-55 Optimized simulation results at 3.55 GHz for $\alpha = 0.25\text{ dB}$, $\Gamma = 18\text{ dB}$, $\angle\Gamma = 0^\circ$, $|\beta| = 20\text{ dB}$, and $|\Gamma_a| = 15\text{ dB}$. Phases of antenna reflection coefficient and circulator isolation are swept.

3.7 GHz

PCM RIT Tuned for Insertion Loss and Isolation

Z_a (for $\Gamma_a = 15$ dB)

$\Gamma = 18$ dB, $\angle \Gamma = 0^\circ$ $\gamma = -\beta/\alpha^2$ $\beta = 20$ dB, $\alpha = 0.25$ dB

0.1 $\angle 0$

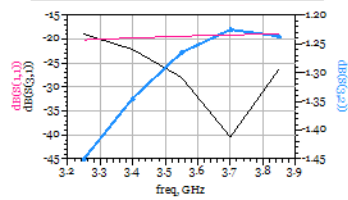
0.1 $\angle 90$

0.1 $\angle 180$

0.1 $\angle 270$

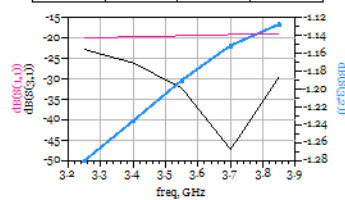
71.62

State[0]	Port1_imp	Port2_imp	Port3_imp
127565	50 + j0	72 + j0	50 + j0



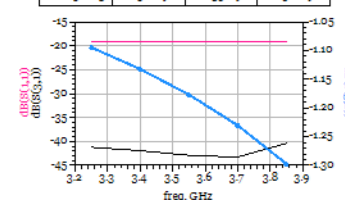
46.93+17.23i

State[0]	Port1_imp	Port2_imp	Port3_imp
474783	50 + j0	47 + j17	50 + j0



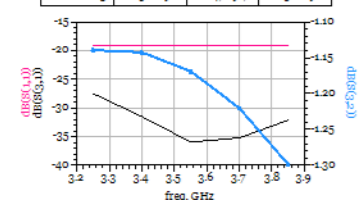
34.9

State[0]	Port1_imp	Port2_imp	Port3_imp
256623	50 + j0	35 + j0	50 + j0

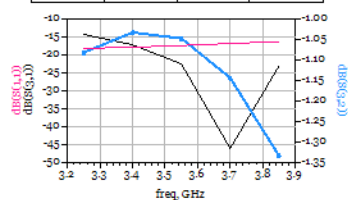


46.93-17.23i

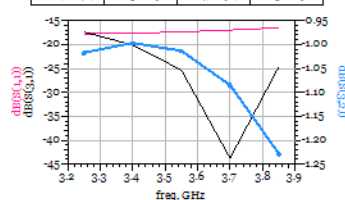
State[0]	Port1_imp	Port2_imp	Port3_imp
116663	50 + j0	47 - j17	50 + j0



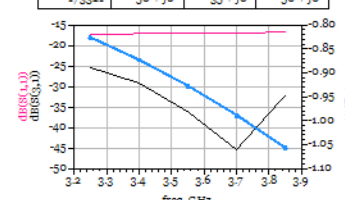
State[0]	Port1_imp	Port2_imp	Port3_imp
474768	50 + j0	72 + j0	50 + j0



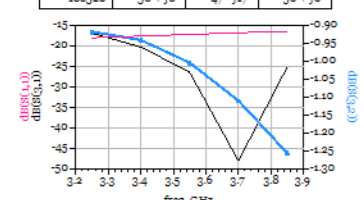
State[0]	Port1_imp	Port2_imp	Port3_imp
764487	50 + j0	47 + j17	50 + j0



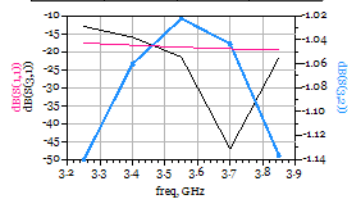
State[0]	Port1_imp	Port2_imp	Port3_imp
173321	50 + j0	35 + j0	50 + j0



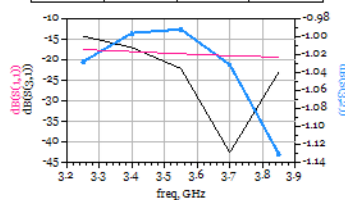
State[0]	Port1_imp	Port2_imp	Port3_imp
188328	50 + j0	47 - j17	50 + j0



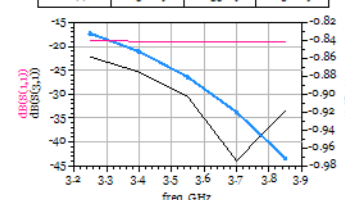
State[0]	Port1_imp	Port2_imp	Port3_imp
844853	50 + j0	72 + j0	50 + j0



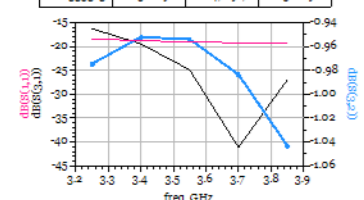
State[0]	Port1_imp	Port2_imp	Port3_imp
654442	50 + j0	47 + j17	50 + j0



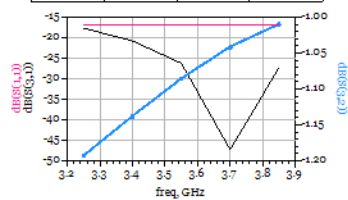
State[0]	Port1_imp	Port2_imp	Port3_imp
877111	50 + j0	35 + j0	50 + j0



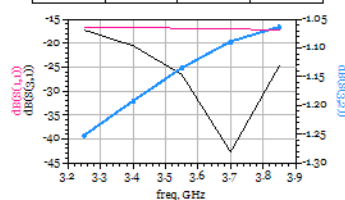
State[0]	Port1_imp	Port2_imp	Port3_imp
335313	50 + j0	47 - j17	50 + j0



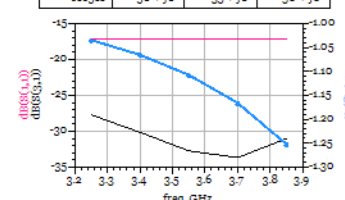
State[0]	Port1_imp	Port2_imp	Port3_imp
414421	50 + j0	72 + j0	50 + j0



State[0]	Port1_imp	Port2_imp	Port3_imp
857871	50 + j0	47 + j17	50 + j0



State[0]	Port1_imp	Port2_imp	Port3_imp
616511	50 + j0	35 + j0	50 + j0



State[0]	Port1_imp	Port2_imp	Port3_imp
335313	50 + j0	47 - j17	50 + j0

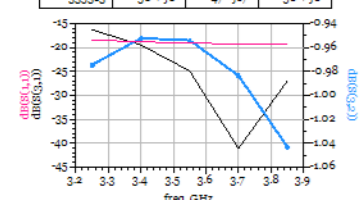


Figure 3-56 Optimized simulation results at 3.7 GHz for $\alpha = 0.25\text{ dB}$, $\Gamma = 18\text{ dB}$, $\angle\Gamma = 0^\circ$, $|\beta| = 20\text{ dB}$, and $|\Gamma_a| = 15\text{ dB}$. Phases of antenna reflection coefficient and circulator isolation are swept.

3.5.1.3 Fabrication and Measurement Results

The circuit is fabricated on a 20-mil HRSi substrate using UWPCM technology. To increase the bulk conductivity of the gold layer, an additional process of gold electroplating is performed to increase the thickness to 3 μm . The measured bulk conductivity of the gold is $2.8 \times 10^7 \text{ } \Omega/\text{m}$, which is measured using four-probe Van der Pauw resistivity measurement method. Figure 3-57 shows the layout of the fabricated circuit. It includes several variations in the input and output terminals, which are suitable for measurement by RF-probe pads, wire bonding, and attaching to SMAs. The layout also includes TL segments for characterization and calibration purpose. The layout of a prototype that is successfully fabricated is highlighted in Figure 3-57. As can be seen each tuning element has 4 bias lines. This is because the PCM SP3T switches are composed of 3 SPST switches that can be controlled independently and each SPST requires two bias pads to operate.

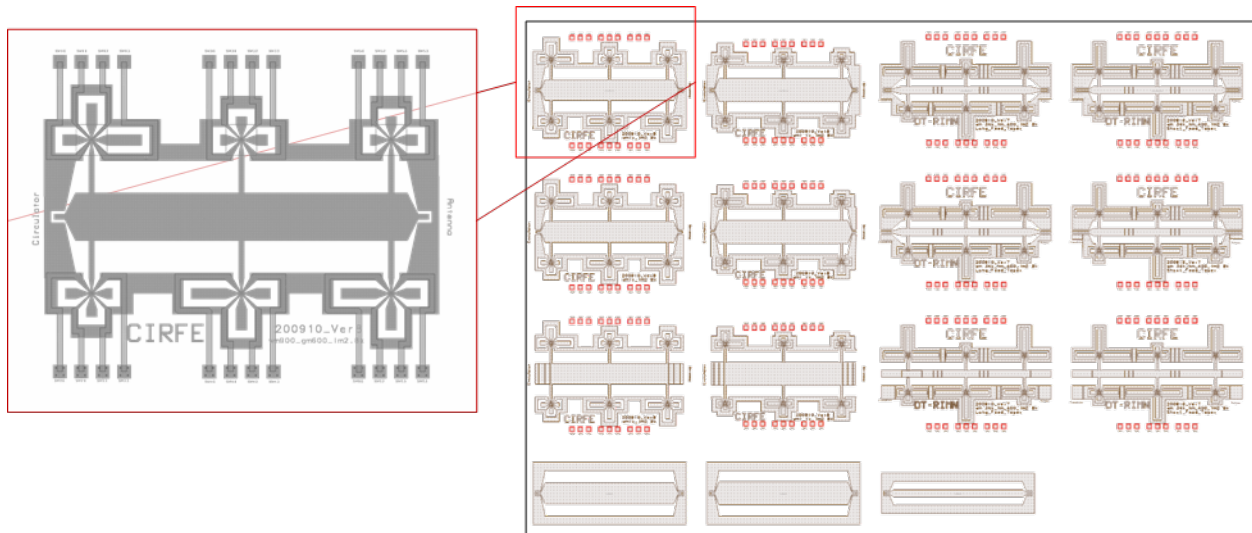


Figure 3-57 Layout of the fabricated PCM based switched RIT.

The fabricated and diced circuit is shown in Figure 3-58. The area of the diced device is 6mm by 7.3 mm, which includes the bias lines and pads. The thickness of the electroplated gold layer is 2.2 μm . After the release of the circuit and removing the photoresist material. The GeTe PCM switches are neither in a crystalline nor in an amorphous state. As a result, each switch acts as a resistor with a random value. Therefore, the state of the switches needs to be turned into either ON or OFF. This is done by applying pulses with suitable magnitudes and widths to each individual switch to turn their states into a known state. The correct pulse to switch state varies depending on a number of factors including but not limited to the

fabrication batch, length of the bias lines, and the heater properties. The suitable pulse properties is derived using characterization of the SPST and then gradually modify it for the used SP3Ts. It is highly likely that a few switches may be damaged during this cumbersome process.

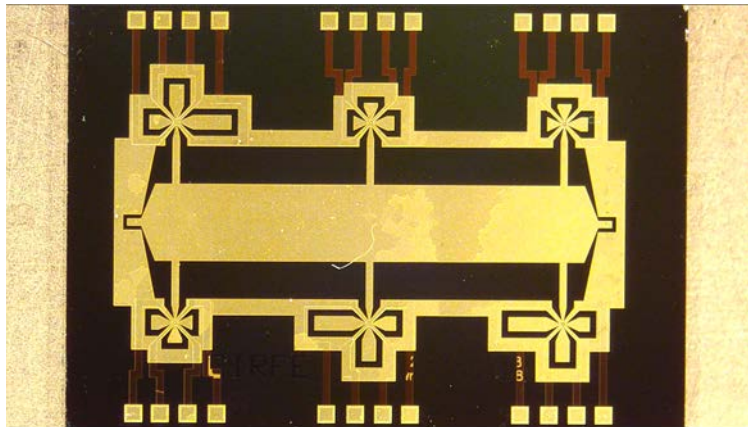


Figure 3-58 fabricated PCM based loaded line RIT

For the testing the devices, a DC offset voltage is added because of the limited amplitude of the pulse generator. The OFF pulse does not require a DC offset. A switch in its ON state behaves as a small resistor in series with the capacitive TL segment, which is equivalent to a high quality factor capacitor. A switch in its OFF state behaves as a large resistor in series with a capacitive TL segment, which is equivalent to a non-ideal open circuit termination. Accordingly, when a switch is not in an ON or OFF state, it behaves as a resistor attached in series with a capacitive TL segment. This is equivalent to a low quality factor capacitor, which degrades the insertion loss of the loaded line RIT. As a result, the RIT can provide impedance tuning, but with considerably high insertion loss. Figure 3-59 demonstrate the performance of the PCM RIT, when used as a doubly terminated matching network over the 3.4 GHz to 3.7 GHz band. As can be seen, input and output return losses of better than 40 dB is achieved over bandwidths wider than 50 MHz. However, the insertion loss varies from approximately 4.6 dB to 5.1 dB over the band of interest. This is associated with the performance of the switches when they are not in ON or OFF states. Figure 3-60 demonstrates the application of the PCM RIT for reducing the leakage between Tx and Rx paths. The isolation levels better than -40 dB is achieved over bands between 20 MHz (at 3.4 GHz) and 50 MHz (at 3.7 GHz). Here, the S11 (dB) is dictated by the circulator (i.e. in this case 18 dB). Moreover, antenna return loss of 12 dB is achieved. The insertion loss is approximately similar with the previous application as DT-RIMN. Hence, even if the proposed PCM RIT becomes damaged, it still capable of improving the isolation

between the Tx and Rx paths and acts as an absorptive attenuator. Therefore, the proposed device is a suitable candidate for applications where protection of the circuits on Rx path are a top priority.

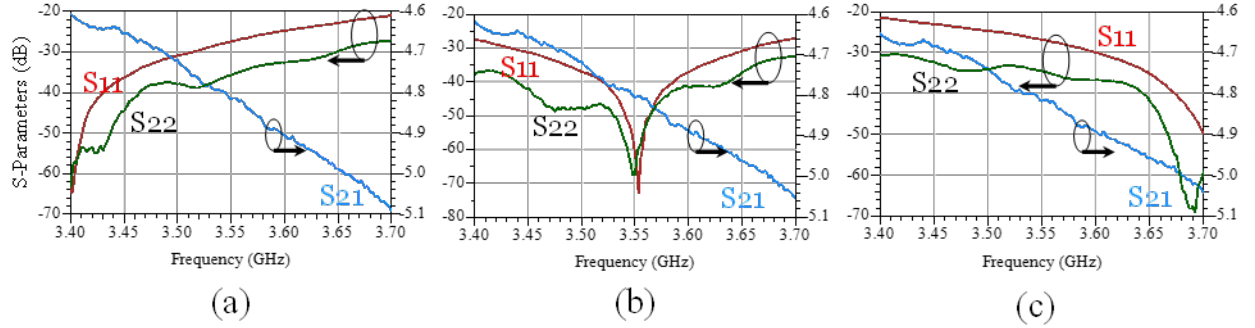


Figure 3-59 Demonstrating the performance of the proposed PCM RIT as a DT-RIMN, (a) at 3.4 GHz with $Z_A = 38.5 + j7 \Omega$ ($|\Gamma_A| = 16.4 \text{ dB}$) and $Z_C = 26 + j3 \Omega$ ($|\Gamma_C| = 9.9 \text{ dB}$), (b) at 3.55 GHz with $Z_A = 35.7 + j10 \Omega$ ($|\Gamma_A| = 13.9 \text{ dB}$) and $Z_C = 24.8 + j3.6 \Omega$ ($|\Gamma_C| = 9.4 \text{ dB}$), and (c) at 3.7 GHz with $Z_A = 32.5 + j11.5 \Omega$ ($|\Gamma_A| = 12 \text{ dB}$) and $Z_C = 24 + j5 \Omega$ ($|\Gamma_C| = 9 \text{ dB}$).

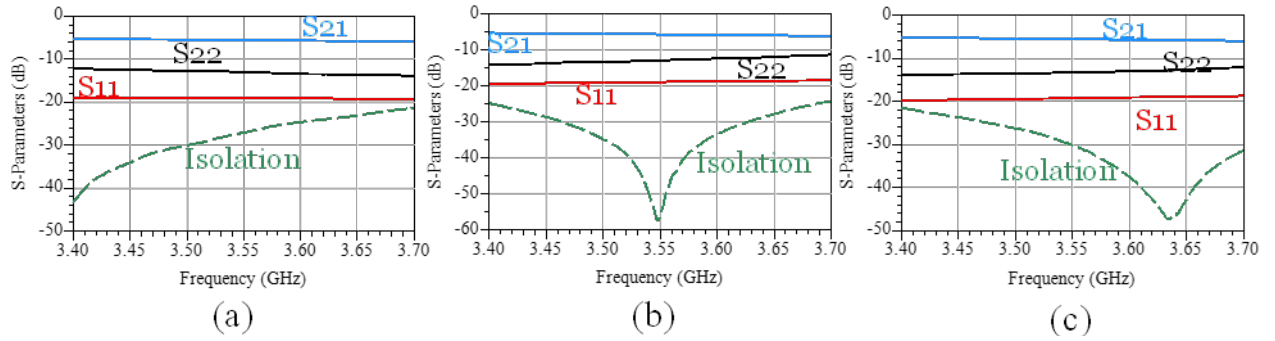


Figure 3-60 Demonstrating the performance of the PCM RIT in improving the isolation between Tx and Rx paths, (a) at 3.4 GHz with $Z_A = 42.8 + j7.88 \Omega$ ($|\Gamma_A| = 18.8 \text{ dB}$), $\angle \beta = 0^\circ$, and $\angle \Gamma = 0^\circ$ (b) at 3.55 GHz with $Z_A = 27.4 - j15.8 \Omega$ ($|\Gamma_A| = 9.1 \text{ dB}$), $\angle \beta = 74^\circ$, and $\angle \Gamma = -18^\circ$ and (c) at 3.7 GHz with $Z_A = 40 - j10 \Omega$ ($|\Gamma_A| = 15.6 \text{ dB}$), $\angle \beta = 47^\circ$, and $\angle \Gamma = 0^\circ$.

3.5.2 Comparing the Performance of RITs Employed as DT-RIMNs and for Improving the Tx-Rx Isolation

As shown in section 3.5, the demonstrated coverage of the BST DT-RIMN is sufficient for improving the isolation between Tx and Rx ports. Therefore, the same device can be studied for comparing the two

applications. Accordingly, the measured states are used in 3 different cases. First, the model is terminated with the antenna and circulator input impedances. Second, the model is attached to the circulator on its circulator side and terminated with an antenna input impedance on the other side, and is tuned for the complex conjugate matching between the RIMN and the antenna, as well as for the best insertion loss performance. Finally, it is attached to the circulator in the same fashion as the second case, but is tuned for improving the isolation. The criteria for all three cases is set for a performance better than 40 dB of either return loss or isolation, depending on the case. Figure 3-61 shows the circuit model used for the later two cases.

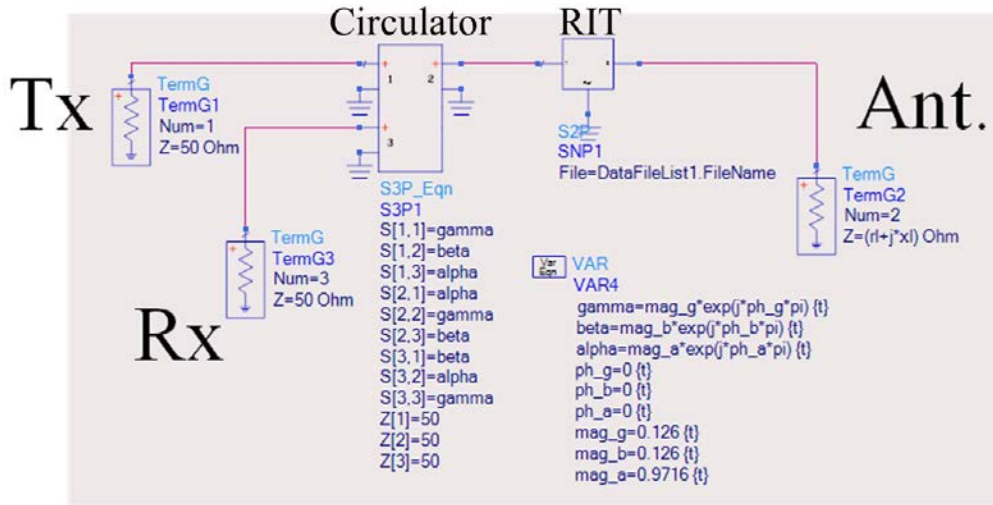


Figure 3-61 Circuit model used to compare the performance of the RITs for the two applications.

The measured state used for this study is the one when all BSTs are connected to a 19V bias voltage. This case corresponds to the highest quality factor for the capacitors. Therefore, the circuit itself shows the best insertion loss among the measured states. For brevity, only this state is used and the comparison is performed only at 3.4 GHz. However, the results are valid for other states and frequencies.

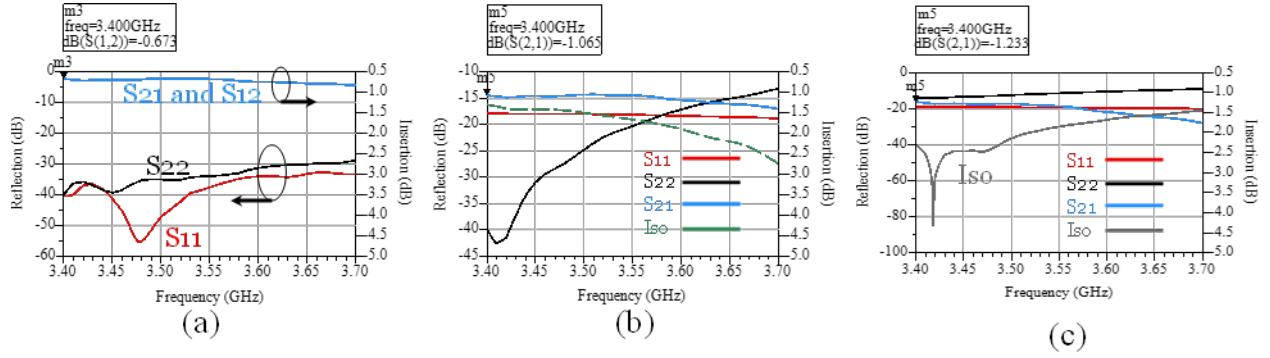


Figure 3-62 Comparing the performance of the RITs in the three cases, (a) optimized for source and load matching with $Z_S = 33.6-j15.9\Omega$ ($|\Gamma_S| = 11.4 \text{ dB}$), and $Z_L = 31.7-j16.6\Omega$ ($|\Gamma_L| = 10.6 \text{ dB}$), (b) optimized for load matching and low insertion loss with $Z_L = 18-j1.86\Omega$ ($|\Gamma_L| = 5.6 \text{ dB}$), and (c) optimized for improved isolation with $Z_L = 19.2-j11.6\Omega$ ($|\Gamma_L| = 6.5 \text{ dB}$).

The optimized results are shown in Figure 3-62. In all three cases, a bandwidth of more than 20 MHz is achieved. Note that for the third case the isolation is more than -40 dB for over an 80 MHz of bandwidth. This may be associated with the assumption of constant phase for β over frequency. However, since isolation dips more than -80 dB, it is expected that a linear phase for β , as opposed to the constant phase assumption, would still result in very wide band isolation improvement. In terms of insertion loss, the first case represents the best results among the studied cases with an insertion loss as low as 0.67 dB. The other studied cases carry an inherent 0.25 dB insertion loss associated with the circulator model. Accordingly, when complex conjugate matching is targeted (i.e. second case), a 1.06 dB of insertion loss is observed, which is negligibly 0.15 dB higher than the anticipated value. Interestingly, when the RIT is used for optimized for isolation, the insertion loss degrades by only another 0.17 dB for improving the isolation by 23 dB. Therefore, the cost of each 10 dB improvements in the isolation is only 0.07 dB of insertion loss. This value is compared among other measured states and a mean of 0.1 dB of degradation in insertion loss for each 10 dB improvements in isolation can be used as a rule of thumb for the proposed devices.

Chapter 4

Dual-band Impedance Matching Networks

In this chapter, a new methodology for designing dual-band matching networks based on filters and phase shifters is presented. The proposed Dual-band Cascade Matching Network (DBCMN) is capable of independently match load impedances at two frequencies. The theory behind the proposed concept is presented along with detailed design procedure for realizing alternative circuit implementations. Prototype circuits are fabricated and tested to verify the proposed concept. Two types of implementation are proposed for the DBCMN. The first type, TYPE-I employs single band phase shifters and a set of High Pass Filters (HPFs) and Low Pass Filters (LPFs). The second type, TYPE-II uses a Band Pass Filter (BPF) and dual band phase shifters.

4.1 TYPE-I: Implementation Using HPF/LPF and Phase-shifters

The Type-I DBCMN is composed of a LPF, an HPF, and two phase-shifters in a cascade configuration. A block diagram of the proposed device is illustrated in

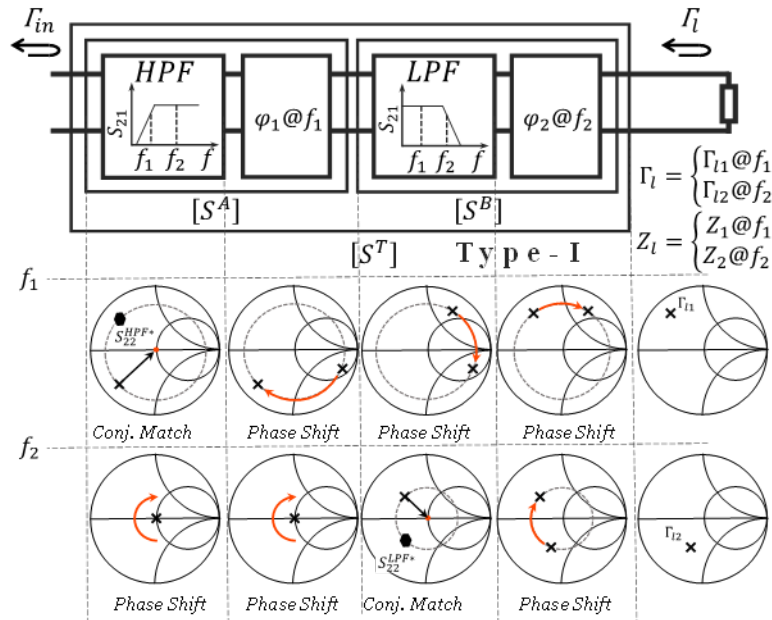


Figure 4-1 Block diagram of the Type-I DBCMN and the performance breakdown for the filters and the phase shifters.

The HPF and the first phase shifter form the S^A block and the LPF and the second phase shifter form the S^B block. The concept can be implemented interchangeably by connecting either the S^A to the dual band load or, the S^B to the dual-band load. These two configurations lead to different outcomes in terms of the matched bandwidths and insertion loss. In [70], an example is presented where S^A is connected to a dual band load. In this paper, we show the same example, where S^B is instead connected to the load first. The performance of these two configurations is compared to determine guidelines for selecting appropriate configuration.

Figure 4-1 also illustrates the performance of the Type-I DBCM by showing the input impedance of the circuit after each block looking toward the load. The device is composed of ideal and lossless elements and is terminated by a hypothetical dual-band load. The dual-band load is defined by its reflection coefficients at a lower frequency, f_1 , and a higher frequency, f_2 , hence $f_1 < f_2$. For the reference, the reflection coefficients of Γ_{l1} and Γ_{l2} are associated with the impedances of Z_1 and Z_2 respectively. Note that f_1 and f_2 are located over the passbands of the LPF and HPF respectively, where these ideal filters would only add phase to a passing signal. Consequently, at f_1 , the phase shifters and the LPF will only add phase to the Γ_{l1} . The HPF has an output return loss with the same magnitude as the Γ_{l1} . Therefore, complex conjugate matching is achieved by modifying the phase shift of the phase shifters and the LPF. Similarly, at f_2 , the LPF has an output return loss with the same magnitude as the Γ_{l2} . Therefore, the second phase shifter adds phase to Γ_{l2} to achieve complex conjugate matching and the consequent phase shifts by S^A would not influence the matching. Since S^B is responsible for matching at f_2 , the required phase shift at f_1 is achieved using the first phase shifter.

For simplicity, each phase shifter is realized by a specified length of a 50Ω transmission line. The ideal LPF and HPF are initially designed to have their cut-off frequencies near the two matching frequencies. The cascade scattering matrix of the HPF and the first phase shifter is presented as $[S^A]$, and the cascade of the LPF and the second phase shifter is presented as $[S^B]$ matrices. The overall scattering parameters for this arrangement can be written as [13]:

$$S_{11}^T = S_{11}^A + \frac{S_{12}^A S_{11}^B S_{21}^A}{1 - S_{22}^A S_{11}^B} \quad 4-1$$

$$S_{12}^T = \frac{S_{12}^A S_{12}^B}{1 - S_{22}^A S_{11}^B} \quad 4-2$$

$$S_{21}^T = \frac{S_{21}^A S_{21}^B}{1 - S_{22}^A S_{11}^B} \quad 4-3$$

$$S_{22}^T = S_{22}^B + \frac{S_{21}^B S_{22}^A S_{12}^B}{1 - S_{22}^A S_{11}^B} \quad 4-4$$

Ideally, at f_1 , $|S_{11}^B| = |S_{22}^B| = 0$, $S_{21}^B = S_{12}^B = 1 \angle (\angle S_{21}^{LPF} + \varphi_2)$. At f_2 , $|S_{11}^A| = |S_{22}^A| = 0$ and $S_{21}^A = S_{12}^A = 1 \angle (\angle S_{21}^{HPF} + \varphi_1)$. The reciprocity imposes that $S_{21}^A = S_{12}^A$ and $S_{21}^B = S_{12}^B$. Therefore, the equations can be simplified further to:

$$[S^T] = \begin{bmatrix} S_{11}^A & S_{21}^A \angle (\angle S_{21}^{LPF} + \varphi_2) \\ S_{21}^A \angle (\angle S_{21}^{LPF} + \varphi_2) & S_{22}^A \angle 2(\angle S_{21}^{LPF} + \varphi_2) \end{bmatrix} \quad 4-5$$

at f_1 and

$$[S^T] = \begin{bmatrix} S_{11}^B \angle 2(\angle S_{21}^{HPF} + \varphi_1) & S_{21}^B \angle (\angle S_{21}^{HPF} + \varphi_1) \\ S_{21}^B \angle (\angle S_{21}^{HPF} + \varphi_1) & S_{22}^B \end{bmatrix} \quad 4-6$$

at f_2 . Therefore, the filters would not modify the magnitude of the total s-parameters over their passbands. The conditions for the impedance matching at the two frequencies are $\Gamma_{l1} = S_{22}^{T*}$ at f_1 and $\Gamma_{l2} = S_{22}^{T*}$ at f_2 . Moreover, the source reflection coefficients are zero at both frequencies. Consequently, the input reflection coefficient, Γ_{in} , can be written similarly at both frequencies as:

$$\Gamma_{in} = S_{11}^T + \frac{S_{12}^T S_{21}^T S_{22}^{T*}}{1 - S_{22}^T S_{22}^{T*}} = S_{11}^T - \frac{S_{12}^T S_{12}^{T*} S_{11}^T}{1 - S_{22}^T S_{22}^{T*}} = 0 \quad 4-7$$

The impedance matching condition for dual-band matching then becomes:

$$|S_{22}^{HPF}(f_1)| \angle (2\angle S_{21}^{LPF}(f_1) + 2\varphi_2(f_1) + \angle S_{22}^{HPF}(f_1) + 2\varphi_1(f_1)) = \Gamma_{l1}^* \quad 4-8$$

$$|S_{22}^{LPF}(f_2)| \angle (\angle S_{22}^{LPF}(f_2) + 2\varphi_2(f_2)) = \Gamma_{l2}^* \quad 4-9$$

where φ_1 and φ_2 provide two degrees of freedom for simultaneous matching. These two equations show that for any angle of S_{22} of the filters, which ranges between 0-360°, there is at least one combination of φ_1 and φ_2 with values between 0-180° that leads to conjugate matching. However, the bandwidth of matching depends on the phase response of the phase shifters, as well as order, type, and the cut-off frequency of the HPF/LPF filters. These factors are studied in the following section.

The matched bandwidth refers to the bandwidth over which the complex conjugate matching is valid and the return loss is better than a set value. To study the matched bandwidth of the DBCMN, a cascade

combination of phase shifters and ideal filters is analyzed. One factor that influences the matched bandwidth is the slope of the phase response for the phase shifters. A transmission line segment is a basic phase shifter. As the length of a transmission line increases, the value and slope of phase response increase. As a result, the valid bandwidth of complex conjugate matching to a frequency independent load decreases. This is demonstrated in Figure 4-2, where matched bandwidths are shown for source and load pairs with the same magnitude of the reflection coefficient (i.e. $|\Gamma_s| = |\Gamma_l|$), but different phases. The matching is achieved by a $50\ \Omega$ transmission lines with varying lengths. As shown in [70] and here in equation 4-9, the electrical length of the transmission line needs to be half the phase difference between the source and the load. As can be seen, the matched bandwidth decreases as the electrical length of transmission line increases. Moreover, lower magnitudes of reflection coefficient (i.e. closer to 0dB) lead to narrower matched bandwidths.

The matched bandwidth of the proposed DBCM is determined by the complex conjugate matching between the output impedance of the filters and the input impedance of corresponding phase shifters. Both of these impedances are frequency dependent and their phases increase with the frequency since they are composed of non-dispersive transmission lines. This enforces that the complex conjugate matching condition holds over narrow frequency bands. This can be studied by considering the block diagram shown in Figure 4-3. For a perfectly matched source, Γ_s is equal to 0. Therefore, the conjugate matching needs to take place between S_{22} of the filters and Γ_{in} of the phase shifter, both of which are frequency dependent. Consequently, the matched bandwidth not only depends on the phase shifter, but also on the S_{22} , which is determined by the filter types (e.g. Butterworth, Chebyshev, etc.) with different properties (order, ripple, etc.). The following analysis is performed for a single band matching but the results are valid for individual bands of the dual-band matching network.

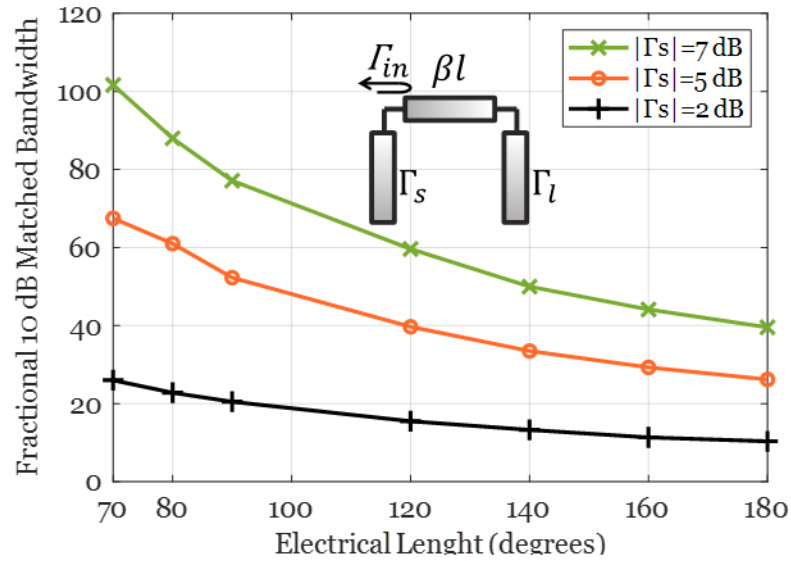


Figure 4-2 The fractional matched bandwidth versus the electrical length of a 50 Ω transmission line connecting two impedances with the same return loss.

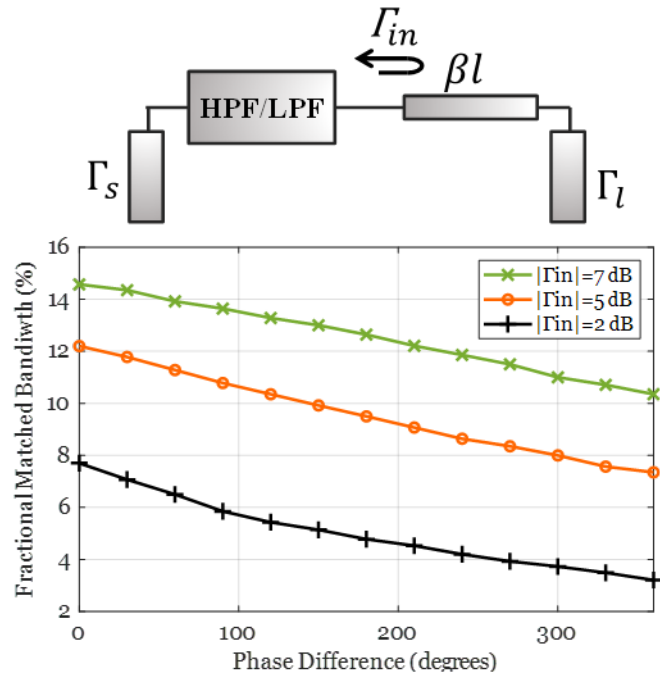


Figure 4-3 The fractional matched bandwidth for a cascade connection between a filter and a transmission line phase shifter.

Figure 4-3 shows the matched bandwidth between ideal third order Chebyshev filters and microstrip line segments with respect to the phase difference between the load and the output reflection coefficient of the filter. This is done for different magnitudes of the load's reflection coefficient. For generalization, only the phase difference between the load and filter is considered and the initial phase of the filter's reflection coefficient is disregarded. Moreover, the fractional matched bandwidth is reported for frequency independent analysis. For the same purpose, the cut-off frequency of the filter is not reported. As a result, these results are valid for both frequencies of a dual-band matching network. As can be seen, the bandwidth decreases as the phase shift and/or the magnitude of the load reflection coefficient increases.

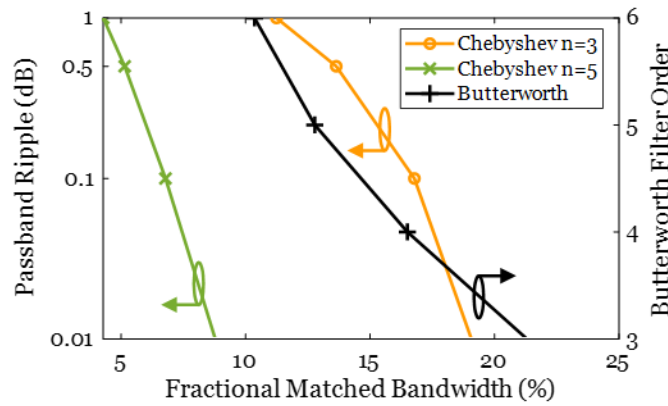


Figure 4-4 The fractional matched bandwidth for different filters.

Figure 4-4 shows the matched bandwidths for Chebyshev filters of order 3 and 5 with varying ripple coefficients, as well as for different orders of the Butterworth filters. For the reference, the phase difference between the load and the output reflection coefficients of the filter is set to 50° . As can be seen, employing lower order filters results in wider matched bandwidths for both butter worth and Chebyshev filters. Furthermore, lower ripple levels for Chebyshev filters leads to wider bandwidths. Interestingly, employing third order Butterworth filters results in widest matched bandwidth among the studied cases. Moreover, third order Chebyshev filters can achieve wider matched bandwidths than Butterworth filters with higher than 4th order and 5th order Chebyshev filters.

The conditions 4-8 and 4-9 can be met either analytically or empirically. The analytical method uses the expressions for S-parameters of the filters and close form representation of phase shifters to determine the initial values for the components needed, while the empirical method takes advantage of circuit simulators

and optimizers for deriving such component values. In both cases, the design procedure includes modifying the cut-off frequencies of the two filters so that S_{22}^{LPF} with φ_2 conjugate matches Γ_{l2} at f_2 according to (9), and S_{22}^{HPF} with φ_1 matches to the Γ_{l1} at f_1 according to (8). Accordingly, the analytical approach differs based on the choice of filters and phase shifters and may include cumbersome calculations depending on the load. However, the analytical calculations remain simple for low-order filters, transmission line based phase shifters, and the dual-band loads that are considered constant over each band. For example, Chebyshev filters have characteristic polynomials of [13]:

$$S_{21}(s) = \frac{P(s)}{\varepsilon E(s)}, \quad S_{11}(s) = \frac{F(s)}{E(s)} \quad 4-10$$

$$P(s) = 1, \quad F(s) = T_N(s), \quad \text{and} \quad E(s) = \prod_{k=1}^{k=N} (s - s_k) \quad 4-11$$

$$s_k = \pm \sinh\left(\frac{1}{n} \sinh^{-1} \frac{1}{\varepsilon}\right) \sin\left(\frac{\pi(2k-1)}{2n}\right) + j \cosh\left(\frac{1}{n} \sinh^{-1} \frac{1}{\varepsilon}\right) \cos\left(\frac{\pi(2k-1)}{2n}\right) \quad 4-12$$

where $T_N(s)$ is the Nth order Chebyshev polynomial and ε is the passband ripple for a LPF prototype with cutoff frequency of 1 and normalized impedance of 1. Transferring these polynomials to the desired frequency and impedance for the two filters and satisfying 4-8 and 4-9 allows for determining the filter components.

However, the analytical method does not account for the coupling between different circuit elements and needs to be accompanied by a full-wave simulation and optimization. In contrast, the empirical method, which employs a cascaded network model of parametrized blocks, is less complex, is demonstrated visually, and can be accompanied by the space mapping technique to reduce the duration of the final optimization of the full-wave simulation. In the following, the empirical method is demonstrated through a design example.

4.2 Design and Fabrication of TYPE-I DBCMN

The proposed design methodology and discussions are demonstrated through a design example with improved bandwidth performance with respect to that presented in [70]. The load impedances are kept to $150+j100$ and $100+j80$ at 1 GHz and 1.8 GHz, respectively, which correspond to $\Gamma_{l1} = 0.635 \angle 18.43^\circ$ and $\Gamma_{l2} = 0.555 \angle 29.9^\circ$. The frequency ratio between these two frequencies is called $r = f_2/f_1$, and is equal to 1.8. These loads are chosen for comparison with [70], and because they cannot be matched by a conventional T-network [18], [20], [52].

Figure 4-5 illustrates the microstrip layout of the proposed DBCMN. The proposed circuit is implemented using distributed microstrip line elements on FR-4: TG170 substrate ($\epsilon_r = 4.4$, $\tan \delta = 0.02$, thickness of 1.6 mm, and 1 oz. copper cladding). For simplicity, the components are realized using either 25 Ω , 50 Ω , or 100 Ω segments. The LPF is a third order Butterworth filter, while the HPF is a third order Chebyshev with 0.05 dB ripple. The LPF is realized using stepped impedance microstrip segments, while the HPF is realized using two shunt and short-circuited stubs, and a series lumped chip capacitor, which is used for size reduction. Moreover, the length of l_5 is modified to account for the shorting end via. The phase shifts are achieved using 50 Ω microstrip line segments. Furthermore, the circuit dimensions are tuned to meet the conditions 4-8 and 4-9. The optimized circuit properties are presented in Table 4-1.

The microstrip segments and the S-parameter model for the capacitor are co-simulated and optimized in ANSYS HFSS and KEYSIGHT ADS environments. For simplicity, the dual band load is assumed to have constant reflection coefficients of Γ_{l1} and Γ_{l2} below and above the mid-band frequency, respectively. The circuit occupies an overall area of 95mm $\times$ 25mm. The circuit area is dominated by the phase shifters. Further miniaturization can be achieved by using meander lines, higher dielectric constant substrates, multilayer implementation, and by employing lumped components.

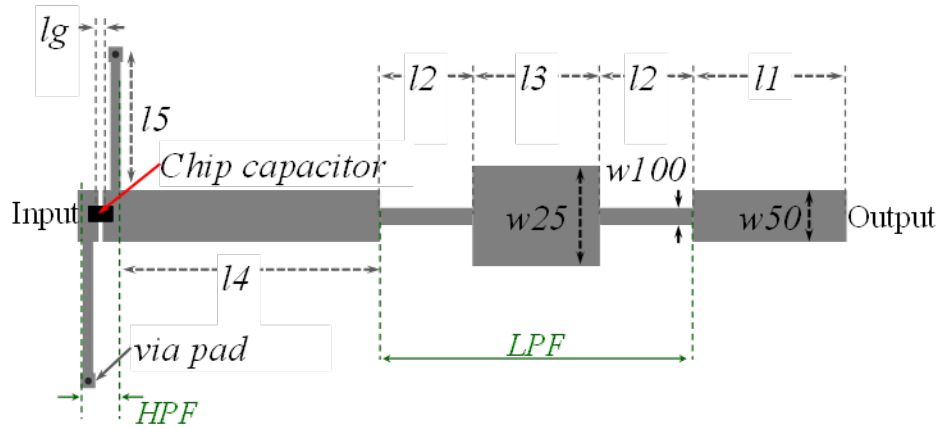


Figure 4-5 Microstrip layout for the proposed Type-I DBCMN

Table 4-1 The design values for the PCB board

PCB Board Dimensions (mm)			
Board size	95mm $\times$ 25mm	l_1	18
w_{50}	3.04	l_2	7.9

w_{25}	8.34	l_3	9.6
w_{100}	0.675	l_4	49
l_g	0.3	l_5	7.5

Figure 4-6 depicts the tuned return loss of the filters, where the intersection of the output return loss of the filters with the constant $|\Gamma_{l1}|$ and $|\Gamma_{l2}|$ circles are depicted. The optimized filter responses slightly differ from the circuit models to account for the phase shifter losses, circuit imperfections, use of the lumped capacitor, microstrip discontinuity, and input/output connectors. The simulation results for the circuit and the full-wave models are presented in Figure 4-7. The fractional bandwidths are 8% and 19% around 1 GHz and 1.8 GHz, respectively. The higher bandwidth is considerably wider than the 13% reported in [70], which is associated with the third order Butterworth. Moreover, the long transmission line segment between the filters has narrowed the lower bandwidth.

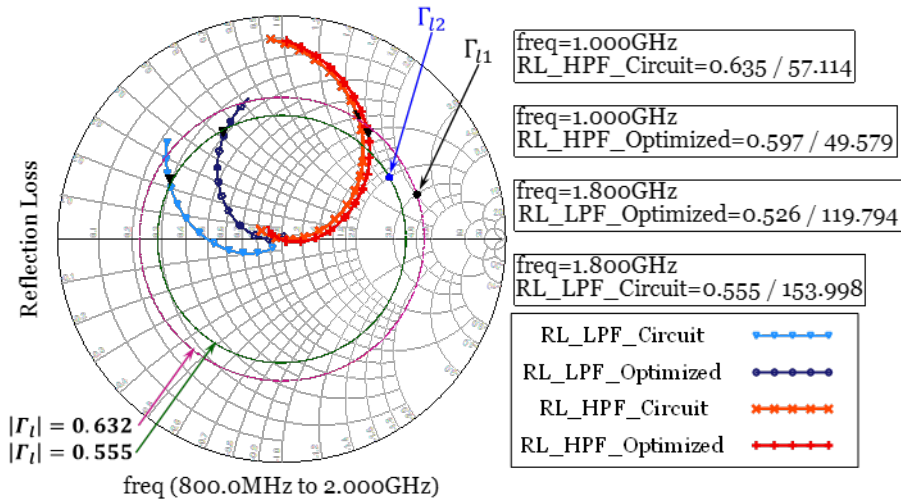


Figure 4-6 Return loss of the LPF and HPF over frequency

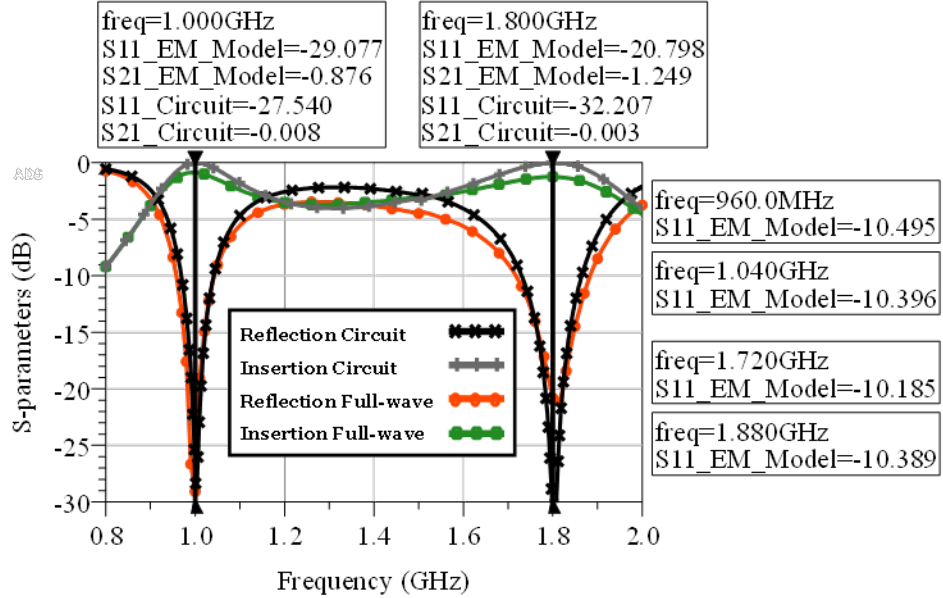


Figure 4-7 Simulated results for the prototype circuit.

The assembled device under test condition is shown in Figure 4-8. The measurements were performed at the University of Waterloo, CIRFE labs. The measured S-parameters were terminated with the dual-band load in ADS, which allows for two-port analysis of the measured results and deriving the insertion loss of the circuit. The SMA connectors and the connector bends are de-embedded to increase the accuracy. The measurement results are presented in Figure 4-9. The lower band has a fractional bandwidth of 6.5% and an average insertion loss of 1.52 dB. The higher band has a fractional bandwidth of 11% and an average insertion loss of 1.48 dB. The discrepancies between simulation and measured results are due to variation in dielectric constant, fabrication tolerances, and the variation in capacitance value for the chip capacitor. The chip capacitor tolerance has caused the HPF to have a lower cutoff frequency, which has reduced the higher bandwidth.

The results presented in this section demonstrate that the bandwidth and insertion loss are degraded by using long transmission line phase-shifters. Therefore, 4-8 and 4-9 imply that if the HPF is placed closer to the dual-band load, the matched bandwidth at f_1 would be wider than at f_2 due to the unavoidable added phase associated with the HPF and the phase shifter at f_2 . By the same observation, if the LPF is located near the dual-band load, the matched bandwidth at f_2 is wider due to the unwanted added phase at f_1 . Moreover, if the type and order of filters are kept the same, then the arrangement which requires shorter

phase shifters performs better in terms of insertion loss. Additionally, a comparison between the prototype design presented here and that presented in [70] shows that both arrangement may require transmission line segments as long as half wavelength at f_1 . The dominant size of the phase shifters causes uncertainty in terms of the required area for this component for different loads. Finally, the series capacitor in the HPF does not permit full integration of this matching network, except for high frequency applications or for small values of capacitors where the capacitors can be realized using multilayer and interdigital capacitors. These shortcomings may hinder the applications of the proposed Type-I DBCM. In the following section, a novel and significantly miniaturized design is presented.

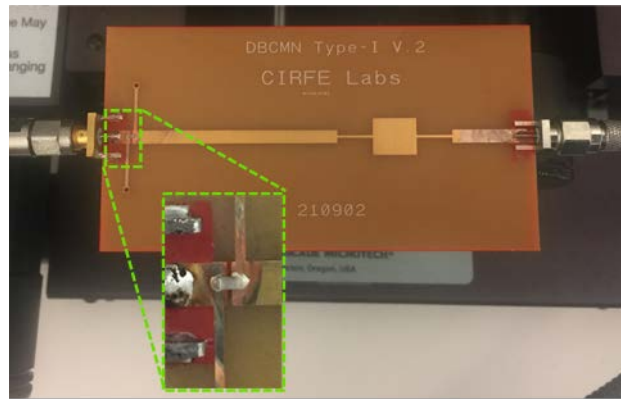


Figure 4-8 Fabricated and assembled device under test.

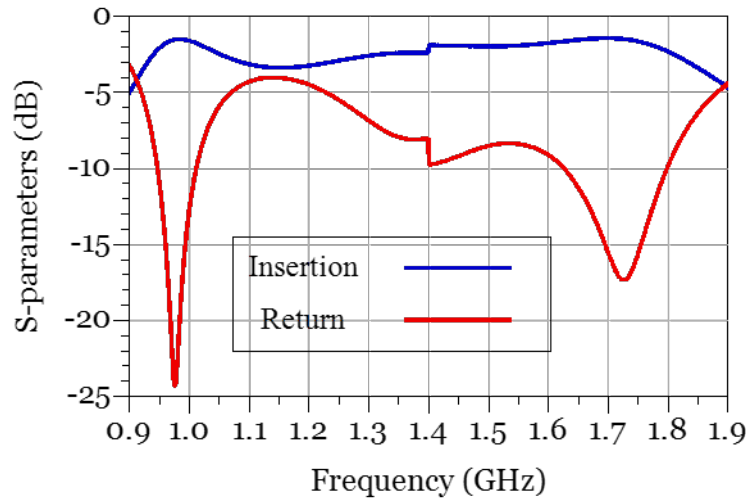


Figure 4-9 Measurement results for the fabricated DBCM.

4.3 TYPE –II: Implementation Using a BPF and a Dual-band Phase-shifter

The block diagram of the Type-II DBCM is composed of two cascaded blocks instead of the four blocks topology of the Type-I. This is done by replacing the LPF and HPF in Type-I together to form a BPF with the passband edges near the two frequencies of interest. Moreover the phase shifters in Type-I are replaced by a dual band phase shifter located between the BPF and the dual-band load. The performance of the proposed design is described in this section.

The proposed topology is shown in Figure 4-10, where $[S^A]$ and $[S^B]$ represent the S-parameters of the BPF and the dual band phase shifter, respectively. Figure 4-11 demonstrates the performance of the proposed design for an arbitrary load using the Smith chart. As illustrated, the complex conjugate matching takes place between the BPF and the dual-band phase shifter at the two frequencies. To achieve this, the center frequency and bandwidth of the BPF is modified so that its response intersects with $|\Gamma_{l1}|$ and $|\Gamma_{l2}|$ circles at f_1 and f_2 , respectively. The complex conjugate matching is then achieved by adding the independent phase shifts to the load reflection coefficients at f_1 and f_2 . Alternatively, this can be expressed as the following impedance matching conditions:

$$|S_{22}^{BPF}(f_1)|\angle(\angle S_{22}^{BPF}(f_1) + 2\varphi_1) = \Gamma_{l1}^* \quad 4-13$$

$$|S_{22}^{BPF}(f_2)|\angle(\angle S_{22}^{BPF}(f_2) + 2\varphi_2) = \Gamma_{l2}^* \quad 4-14$$

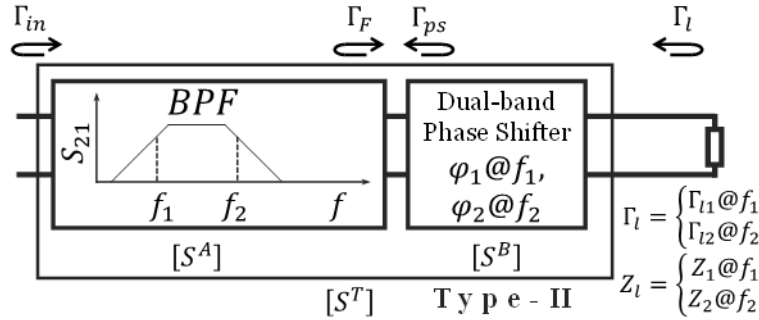


Figure 4-10 Block diagram of the Type-II DBCM

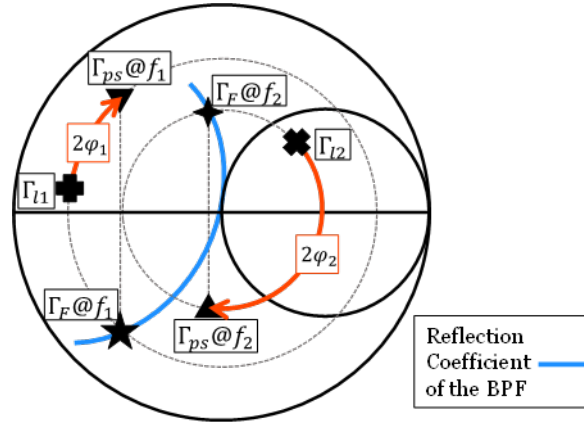


Figure 4-11 The performance breakdown for the Type-II DBCMN at the two frequencies

4.4 Design and Fabrication of TYPE-II DBCMN

In this section, the performance of the proposed DBCMN is evaluated by a design example. A potential schematic to meet 4-13 and 4-14 is illustrated in Figure 4-12. Although it is possible to demonstrate the performance through designing for a handful of loads, for example see [20], this work is focused on demonstrating the coverage throughout the design procedure. Therefore, the Type-II DBCMN is designed to match the same dual-band frequency dependant load used to demonstrate the performance of Type-I. For the purpose of miniaturization, the proposed design is based on 4μm thin-film Gold electroplated on Alumina (99.6 % purity Al_2O_3) substrate with a thicknesses of 10 mil. In the following, the considerations for designing the BPF and the dual band phase shifter are presented.

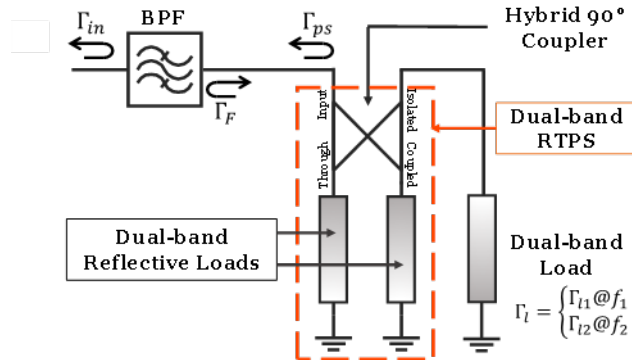


Figure 4-12 Schematic of the proposed Type-II DBCMN.

4.4.1 BPF Design

The BPF is realized using the ultra-wideband BPF design presented in [71], which is composed of coupled resonators. The resonators are realized using a pair of shunt open- and short-circuit terminated stubs, which are coupled by transmission line inverters. This filter topology is capable of achieving very wide passband bandwidths, exceeding 70% [72]. The bandwidth and the center frequency are adjusted by modifying the length and characteristic impedance of the stubs to set the magnitude of the return loss near the cut-off frequencies equal to the magnitude of the reflection coefficient of the load at f_1 and f_2 .

The order and type of the BPF are determined based on the value of r (i.e. $\frac{f_2}{f_1}$), magnitude of the load reflection coefficient at the two frequencies, and the frequency dependency of the load impedance. For high r -values and loads with the magnitude of reflection coefficient closer to one, higher order filters or filter types with steeper roll-off rates (Chebyshev, inverse Chebyshev, or even responses with transmission zeros) are more suitable. In any case, a lower order filter is preferred because of the size and simplicity of design. Figure 4-13 experimentally characterizes the relationship between r -value and fractional bandwidth for Butterworth filter with different orders and a special case where $|\Gamma_{l1}|$ and $|\Gamma_{l2}|$ are both equal to 0.55 (corresponding to 5dB reflection coefficient). Similar diagrams can be used to decide the design and order of the filter.

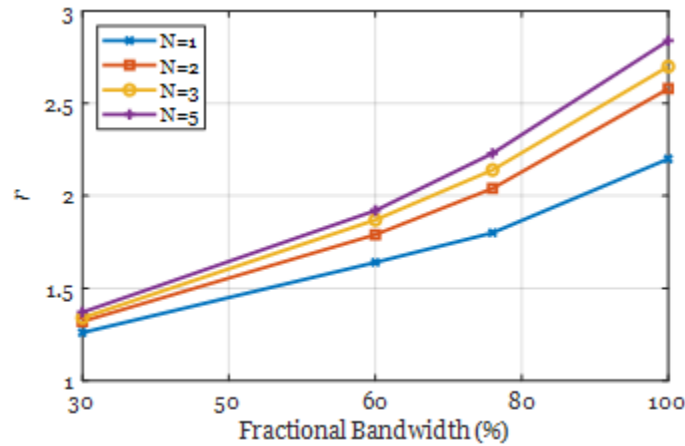


Figure 4-13 Required fractional bandwidth to achieve frequency ratios for Butterworth filter of different orders

Since the load impedances at f_1 and f_2 are assumed constant, a maximally flat filter response is used because of its low roll-off rate. Moreover, the value of r allows for the use of a first order BPF, which is realized using an open circuit terminated and a short circuit terminated stubs to form the resonator. The filter schematic and its response in accordance with $|\Gamma_{l1}| = 0.635$ and $|\Gamma_{l2}| = 0.555$ are presented in Figure 4-14.

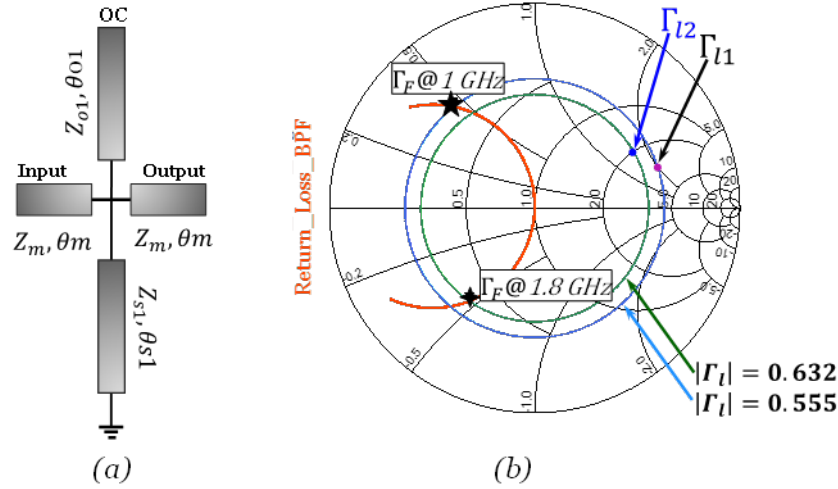


Figure 4-14 BPF Design, (a) schematic of the filter using ideal transmission line stubs, (b) return loss on Smith chart for 50Ω transmission line segments with electrical lengths of $\theta_m = 2.5^\circ$, $\theta_{o1} = 53^\circ$, and $\theta_{s1} = 31.5^\circ$, all at 1.4 GHz.

4.4.2 Dual-band RTPS and Reflective Load Design

For the dual-band phase shifter, dual-band RTPSs are employed because of their wideband performance, compact size, and low insertion loss [73]. A dual band RTPS is composed of a quadrature hybrid coupler terminated by two dual-band reflective loads at its through and coupled ports. This is depicted in Figure 4-12. The input port of the hybrid coupler is connected to the BPF and its isolated port is terminated with the dual-band load to be matched. The phase shift of the RTPS is determined by the phase of the reflection coefficient of the reflective loads [74].

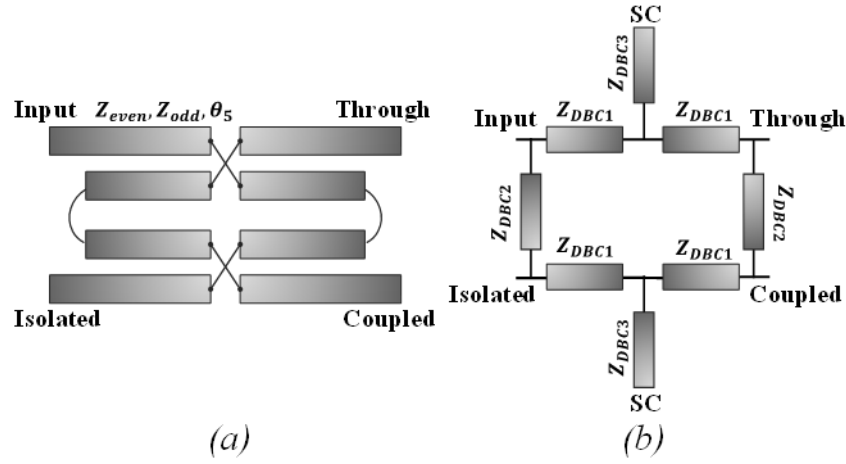


Figure 4-15 Dual-band hybrid couplers, (a) tandem design $Z_{even} = 77.3 \Omega$, $Z_{odd} = 31.8 \Omega$, and $\theta_5 = 45^\circ$ at 1.4 GHz (b) branch-line loaded with T-section $Z_{DBC1} = 24.8 \Omega$, $Z_{DBC2} = 56 \Omega$, $Z_{DBC3} = 14.7 \Omega$, and $\delta = \frac{r-1}{r+1} = 0.286$.

Two alternative designs are employed for realizing the RTPS. For the first design, called design-A, a wideband tandem coupler [75] is used, which is illustrated in Figure 4-15 (a). It is composed of two -8.3 dB couplers, which are connected in a tandem configuration to form the -3 dB coupler. The transmission line segments and the coupling gaps for tandem couplers do not need to be as narrow as other conventional hybrid couplers (e.g. Lange couplers), and they perform over wider bandwidths [75]–[77]. The tandem coupler is composed of four coupled line segments connected by crossovers, where each segment is 45° at the center frequency. To achieve dual band performance, the coupling at the center frequency between f_1 and f_2 is increased to simultaneously achieve the -3 dB coupling at f_1 and f_2 . This is done by modifying the even and odd impedances of the coupled line segments. The proposed hybrid coupler for design-A is realized using coupled CPW lines with wire bonds connecting the traces at the crossover, as well as ground planes on both sides of the hybrid coupler to avoid the excitation of slot-line CPW modes [78]. The wire bonds also contribute to the coupling between the lines in the tandem coupler. Consequently, the actual length of the coupler is slightly less than 45° at the center frequency. The dual band response of the hybrid coupler is reported on Figure 4-16 (a). The layout of the CPW tandem coupler is shown in Figure 4-17 (a).

For the second design, design-B, a dual-band branch-line coupler is used. Dual-band hybrid coupler performance can be achieved by replacing the arms of conventional branch-lines with either π - or T-

networks. However, feasibility of such transformation depends on the achievable range of characteristic impedances, which is limited by the fabrication constraints. For example, π -network designs are not suitable for r -values smaller than 1.5 because they require higher than 100Ω stubs, which add significant losses to the hybrid couplers [79].

For the design example at hand, a T-network dual-band hybrid coupler design requires less than 56Ω characteristic impedances, which is suitable for microstrip realization of the circuit. It is formed by replacing the transmission line segment between input and through ports, and between the coupled and isolated ports with T-networks, so that the electrical lengths of these networks are supplementary at the two frequencies (i.e. θ_b at f_1 and $\pi - \theta_b$ at f_2 , where θ_b is equal to $\pi/(1+r)$) [79], [80]. This is depicted in Figure 4-15 (b), where the electrical lengths of individual segments are set to $\theta_b = 64.3^\circ$ at 1 GHz. The characteristic impedance of these line segments are calculated using equations 17-21 from [41]. The simulated S-parameters of the phase shifter for design-B are presented in Figure 4-16 (b), and the microstrip layout of the dual-band branch-line coupler is illustrated in Figure 4-17 (b).

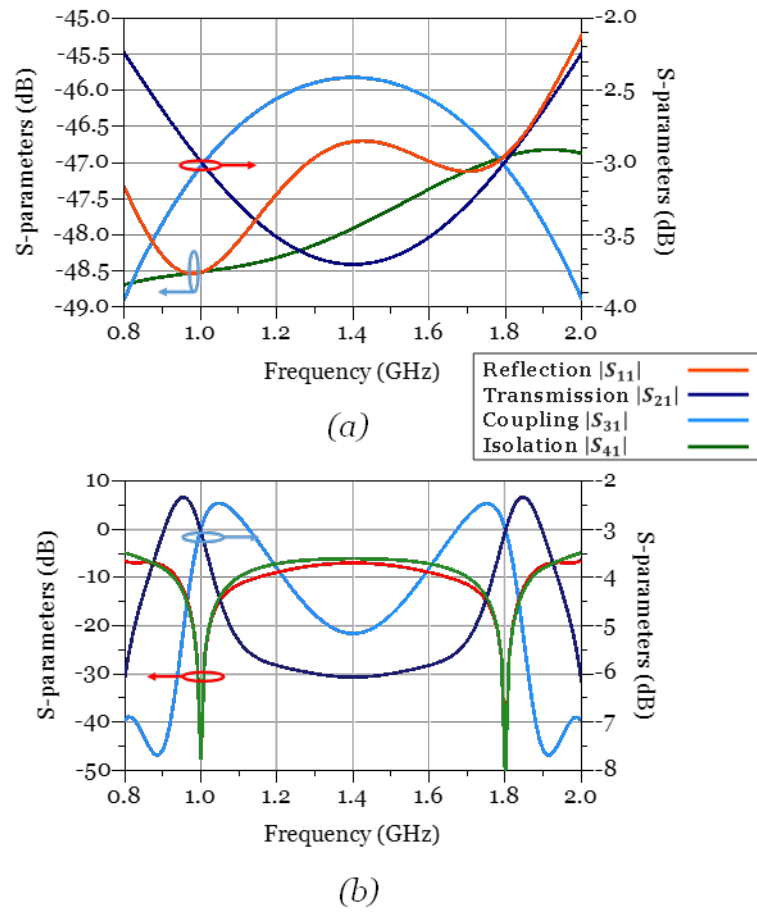


Figure 4-16 Response of the quadrature couplers, (a) tandem coupler, and (b) dual-band branch-line coupler.

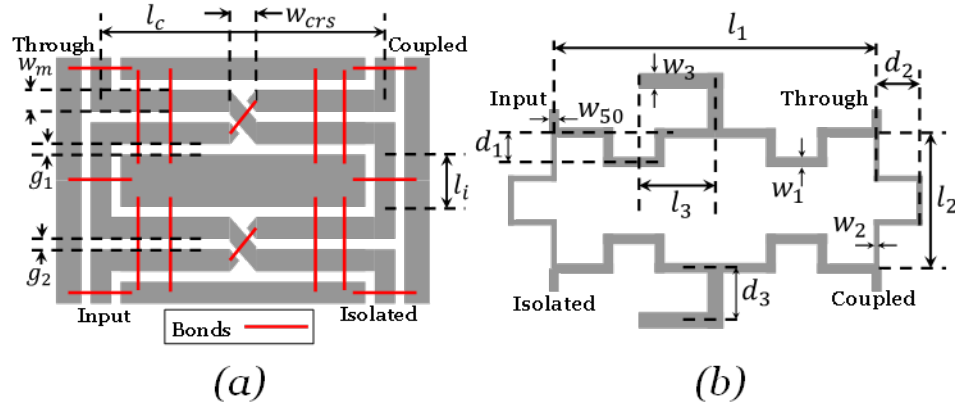


Figure 4-17 Schematic of the dual-band hybrid couplers (a) tandem coupler, and (b) T-network loaded branch-line hybrid coupler.

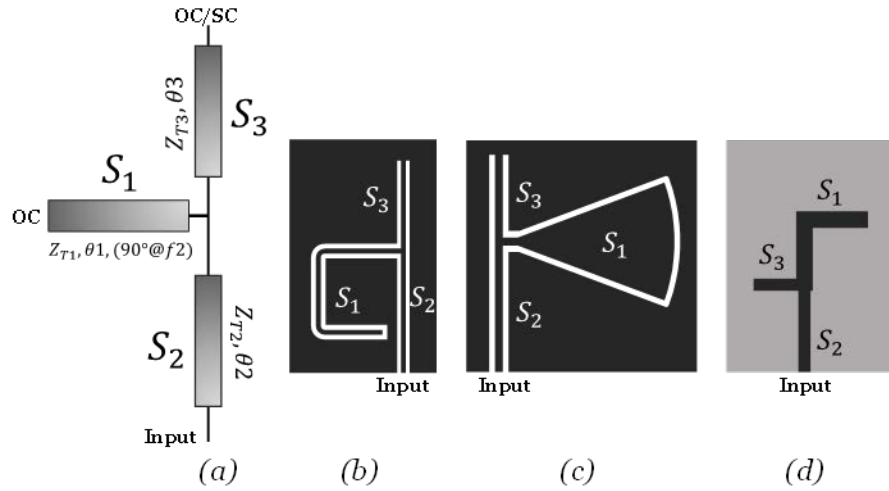


Figure 4-18 The dual-band reflective loads (a) circuit model, (b) CPW with folded stub, (c) CPW with a radial stub, (d) folded microstrip.

Figure 4-18 depicts a realization method for dual-band reflective loads using stubs. The circuit model is composed of two vertical transmission line segments, s_2 and s_3 , which are separated by a T-junction and an open circuit terminated stub called s_1 . s_1 has an electrical length of 90° at f_2 , which acts as a short circuit termination for s_2 , rendering s_3 ineffective at f_2 . Therefore, the input impedance is determined by s_2 around f_2 and by s_2 and s_3 around f_1 . Figure 4-18 (b)-(d) illustrate the realization of the reflective loads using CPW

and a folded s_1 , a radial s_1 , and bent microstrip line segments, respectively. The optimized stub dimensions are expressed in Table 4-2.

Table 4-2 Design Parameters for the Type-II DBCMNs Prototype

	Variable		Circuit-based Calculated Values UNITS: (μm)		
			design-A, folded stubs	Design-A, radial stub	Design-B
BPF	w_m, w_{oc}, w_{sc}		700	700	239
	g_m, g_{oc}, g_{sc}		188, 300, 100	188, 300, 100	-
	l_m		1350	1350	1100
	l_{oc}		15503	14324	11430
	l_{sc}		14130	14805	7121
Dual-band Reflective Loads	s_1	Width	700	-	714
		Gap	300	188	-
		Length	23306	17774	14662
		Angle	-	41°	-
	s_2	Width	700	700	239
		Gap	188	188	-
		Length	12207	13255	924
	s_3	Width	700	700	239
		Gap	188	188	-
		Length	1732	1389	5080
Tandem Coupler		$w_m = 700, g_1 = 129, g_2 = 120, w_{crs} = 1000, l_c = 23500, l_i = 500$			
T-network Loaded Branch-line Coupler		$w_{50} = 239, w_1 = 802, w_2 = 190, w_3 = 1603, d_1 = 3802, d_2 = 3000, d_3 = 6797, l_1 = 28364, l_2 = 15960, l_3 = 1382$			

4.4.3 Phase coverage for the dual-band RTPS

The electrical length of the stubs s_2 and s_3 , θ_2 and θ_3 , determine the achieved phase shift. According to 4-13 and 4-14, the maximum phase shift needed to achieve full coverage is 180° for φ_1 at f_1 and φ_2 at f_2 . To investigate the required range for θ_2 and θ_3 , a circuit level study is performed on a hybrid coupler terminated by the proposed reflective loads. For reference, electrical lengths at 1.8 GHz are considered. Figure 4-19 depicts the input/output phase shift. Figure 4-19 (a) shows the phase shift when θ_3 is kept constant and θ_2 is swept from 0° to 180° . This leads to a 360° shift at 1.8 GHz and a $360^\circ/r = 200^\circ$ shift at 1 GHz. Moreover, Figure 4-19 (b) shows the phase shift when θ_2 is kept constant and θ_3 is varied by a 180° , which has negligible influence on the phase shift at 1.8 GHz due to the short circuit termination by

s_1 . Therefore, the phase shift can be modified independently at the two frequencies. Likewise, a maximum phase shift of 180° at both frequencies requires a maximum length of $\theta_2 + \theta_3 = 90^\circ \times (1 + f_2/f_1)$ for the stubs.

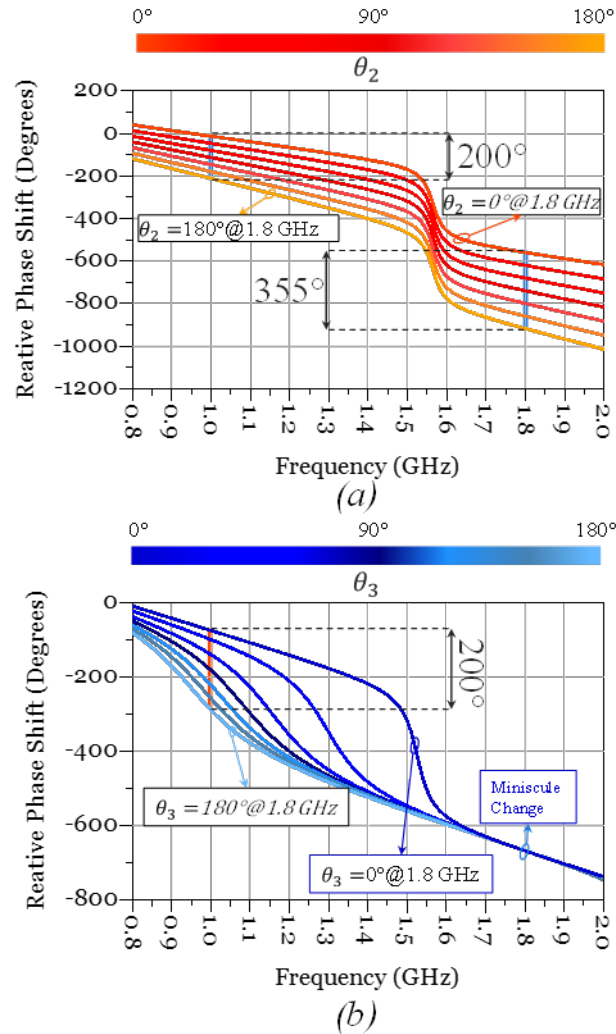


Figure 4-19 Phase shift achieved at the two frequencies by (a) tuning the electrical length of S3, and (b) tuning the electrical length of S4.

4.4.4 Simulation Results

Figure 4-20 illustrates the layouts for the proposed Type-II design-A and design-B circuits. Both design-A circuits are realized using CPW technology. The wire bonds are used for connecting the ground planes, as

well as connecting the signal lines at the crossovers. The optimized hybrid response is shown in Figure 4-21. As can be seen, the input signal is split equally between the reflective loads at f_1 and f_2 , with a coupling coefficient of -3.4 dB and -3.2 dB, respectively. Figure 4-22 shows the tuned reflection coefficients between the BPF and the dual-band RTPS. Achieving complex conjugate at both frequencies requires fine-tuning the reflective loads and the BPF. This is due to the non-ideal behaviors of the filter and the phase shifter, which are caused by the circuit losses and the coupling between the RTPS and BPF stubs. As can be seen, $\Gamma_{ps} = \Gamma_F^* = 0.5424 \angle -118^\circ$ at 1 GHz, and $\Gamma_{ps} = \Gamma_F^* = 0.5074 \angle 129^\circ$ at 1.8 GHz. Accordingly, $|\Gamma_{ps}|$ and $|\Gamma_F|$ at 1 GHz and 1.8 GHz are smaller than $|\Gamma_{l1}|$ and $|\Gamma_{l2}|$.

Figure 4-23 shows the simulated results for the Type-II DBCMN design-A with bent stubs. It has a bandwidth of 8% and an insertion loss of 0.22 dB at 1 GHz. Moreover, it has a bandwidth of roughly 4% and an insertion loss of 0.17 dB at 1.8 GHz. The parallel connection of s_1 and s_3 resonates at 1.53 GHz, which behaves as a radiative element. Nevertheless, this resonance can be shifted by either using radial stubs, modifying the characteristic impedance of the stubs, or extending the length of s_1 and adjust the reflective load accordingly. An example design using radial stubs is shown in Figure 4-20 (b). Figure 4-24 shows a 40 MHz shift in the resonance frequency without palpable change of behavior around the matched bands.

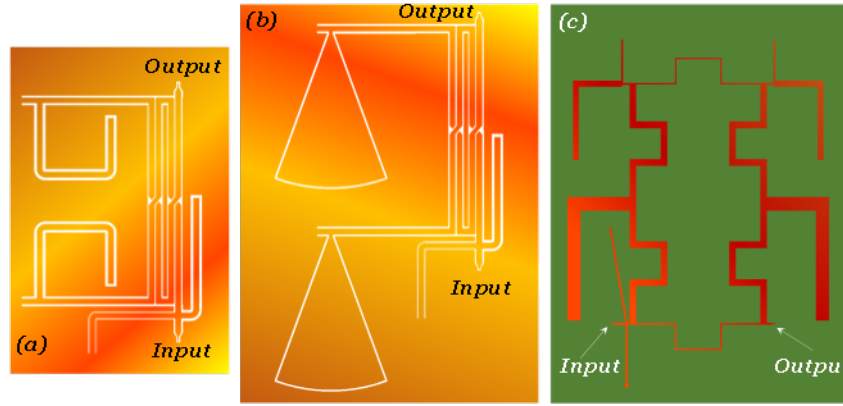


Figure 4-20 Realization of the proposed Type-II DBCMN, using tandem coupler and (a) folded shunting stub, (b) radial shunting stub, and (c) using dual-band branch-line couplers.

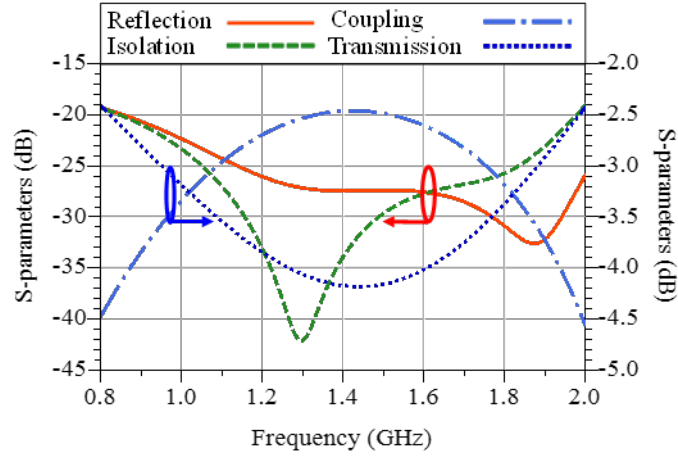


Figure 4-21 Optimized S-parameters for the tandem coupler used in the Type-II DBCM design-A designs.

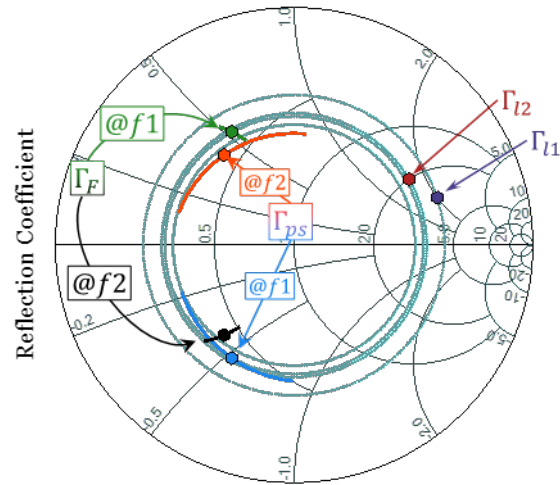


Figure 4-22 Tuned reflection coefficients to achieve complex conjugate matching at both frequencies for the Type-II DBCM design-A designs.

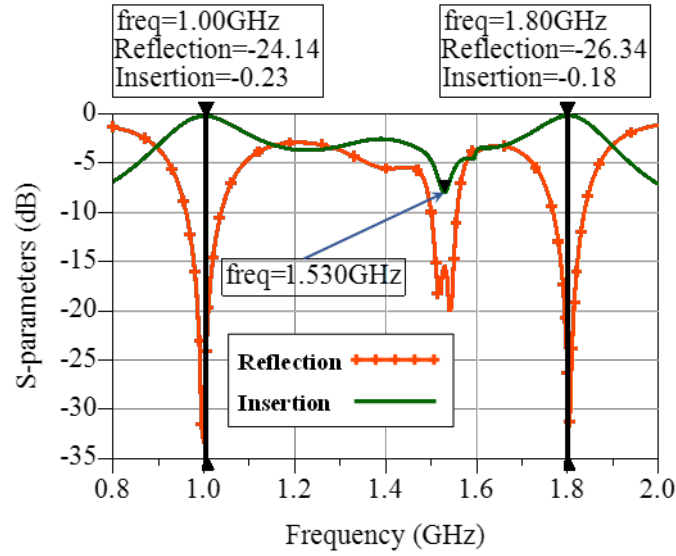


Figure 4-23 Simulated S-parameters for the Type-II DBCM design-A with bent CPW lines.

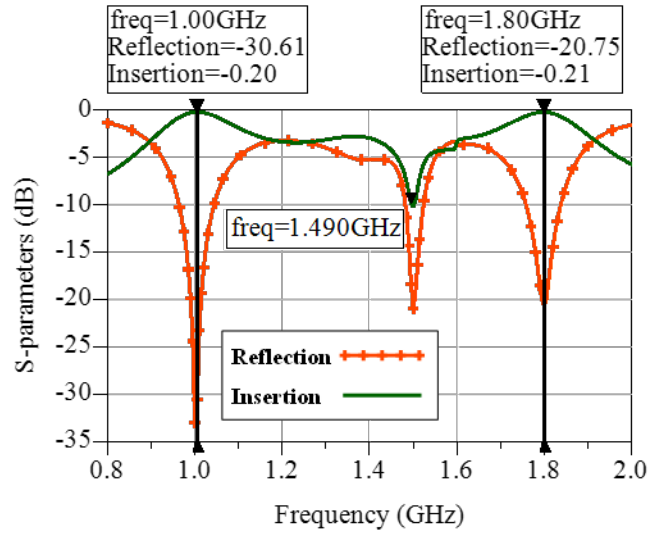


Figure 4-24 Simulated S-parameters for the Type-II DBCM design-A with radial stubs.

Figure 4-20 (c) shows the layout for the design-B. Microstrip technology is chosen for design-B due to the fabrication consideration and monolithic realization. High-impedance transmission lines are avoided to reduce the losses and short-circuited stubs are chosen for the perpendicular stub of the branch-line coupler. Each section is 64.3° at 1 GHz, except for the short-ended perpendicular stub, which is 128.5° . Moreover, the reflective loads and the coupler are folded for size reduction. Furthermore, the open-ended stub of the

BPF is rotated by 10° to reduce coupling with the branch-line segments. The optimized hybrid response is shown in Figure 4-25. The coupling coefficient is -3.5 dB at 1 GHz and 1.8 GHz. Figure 4-26 shows the tuned reflection coefficients, which are $\Gamma_{ps} = 0.6214 \angle -125^\circ$ at 1 GHz and $\Gamma_{ps} = 0.4554 \angle 127^\circ$ at 1.8 GHz. Figure 4-27 shows the simulated S-parameters for the design-B. The transmission zero around 1.4 GHz is associated with short-circuited stub of the branch-line.

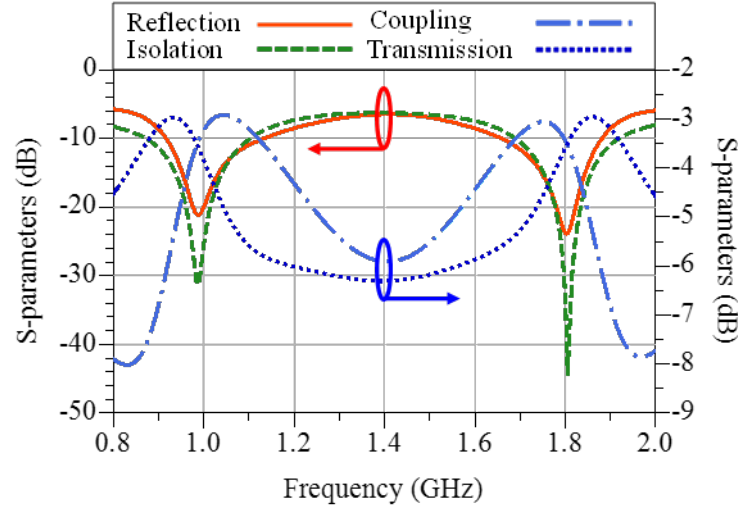


Figure 4-25 Optimized S-parameters for the loaded branch-line hybrid employed in the Type-II DBCMN design-B design.

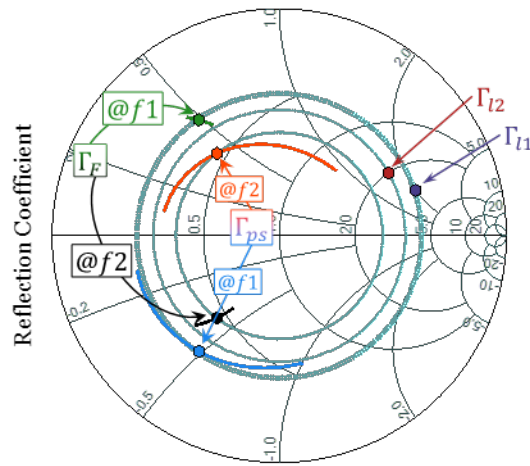


Figure 4-26 Tuned reflection coefficients for the Type-II DBCMN design-B design.

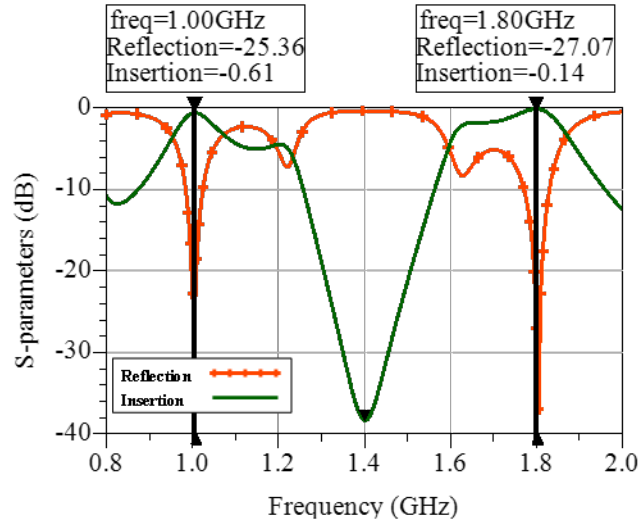


Figure 4-27 Simulated S-parameters for the Type-II DBCMN design-B.

4.4.5 Fabrication and Measurement Results

The Type-II DBCMN prototypes are fabricated on 10 mil Alumina (Al_2O_3) wafers with more than 96 percent purity. The conducting layer is a 4 μm gold, which is electroplated over a 100 μm titanium-tungsten (TiW) seed layer. The fabricated circuits are shown in Figure 4-28. The thickness of the conductor layer is characterized by DektakXT profilometer. The gold layer has an average thickness of 3.8 μm with around 500 nm surface roughness. The measurements are performed from 0.8 GHz to 2 GHz. The measured S-parameters are shown in Figure 4-29. The matched bandwidths for the Type-II DBCMN design-A with folded shorting stubs are 9% and 3% for the lower and higher bands, respectively. The insertion loss are 1.46 dB and 1.68 dB at the center frequencies of the lower and higher bands, respectively. In addition, the matched bandwidths for the Type-II DBCMN design-A with radial stubs are 9% and 5% for the lower and higher bands, respectively. The insertion loss are 1.47 dB and 1.58 dB at the center frequencies of the lower and higher bands, respectively. Finally, the matched bandwidths for the Type-II DBCMN design-B are 4% and 3% for the lower and higher bands, respectively. The insertion loss are 1.76 dB and 1.32 dB at the center frequencies of the lower and higher bands, respectively.

The simulation and measurement results agree well over for the matched bandwidths and the measurement results for both of the design-A variations are slightly wider than the simulations. The narrow matched bandwidth of the Type-II DBCMN is associated with the use of the dual band phase shifters.

Figure 4-30 shows the simulated unwrapped phase of the design-A and design-B designs. As can be seen, the two designs induce roughly the same phase shift around 1 GHz and 1.8 GHz. However, the phase shift is at least 90° at both frequencies due to the use of the dual-band hybrid couplers. Moreover, Figure 4-22, Figure 4-26, and Figure 4-30 show that the phase response of the proposed dual-band RTPSs changes rapidly with the frequency. Hence, the bandwidths of the design examples of Type-II DBCMN are narrower than the Type-I counterparts.

Moreover, the discrepancies between simulation and measurements are caused by a fabrication and assembly errors associated with the low conductivity of the top gold layer, error in circuit dimensions, and use of wire bonds. The conductivity is measured by Keithley Model 4200-SCS Semiconductor Characterization System with four source-measure units (SMUs) using Van der Pauw method. The measured volume conductivity is $2.18 \times 10^7 \text{ } \Omega/\text{m}$, which is lower than the conductivity of gold ($4.1 \times 10^7 \text{ } \Omega/\text{m}$) [11]. Simulations using this conductivity value adds roughly 1.2 dB to the insertion loss. Moreover, the errors in dimensions of the circuit elements has caused a slight frequency shift for all the prototypes. The phase shifters are sensitive to dimension errors, which has resulted in variations in return and insertion losses. The frequency shifts in the measured S-parameters are also caused by the manual assembly and difference in length/height of the wire bonds. Figure 4-31 shows a few assembled wire bonds. These bonds can be replaced by air bridges [78] or assembled using automated wire bonders, which result in consistent sizes and predictable behavior. The loss is also attributed to fabrication errors such as some via walls not fully covered by gold.

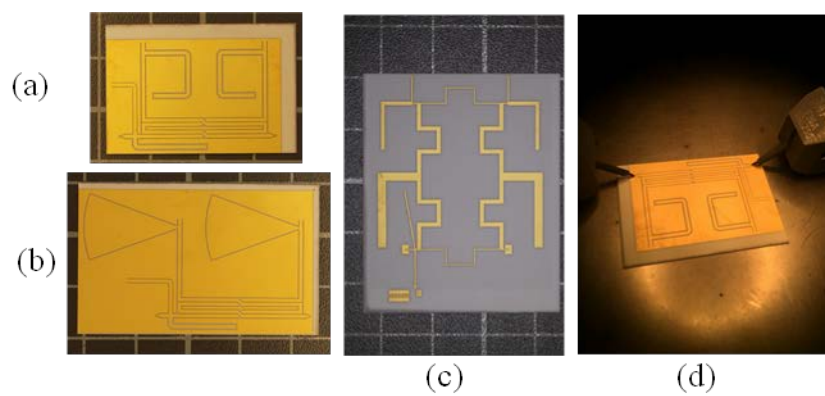


Figure 4-28 Fabricated Type-II DBCMNs, (a) using tandem couplers and folded shorting stubs, (b) radial shorting stubs, (c) using dual-band branch-line couplers, and (d) device under test using RF probe station.

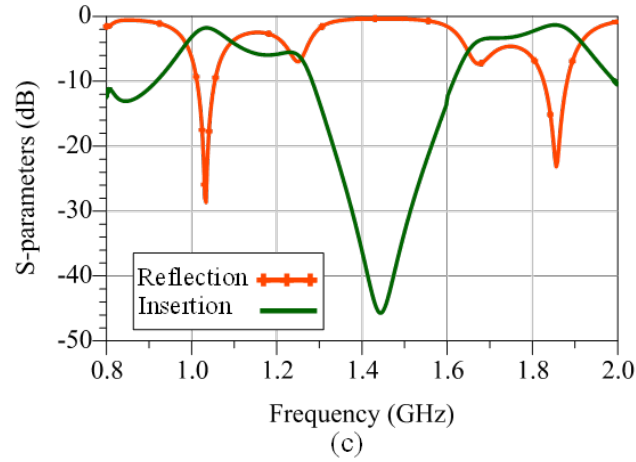
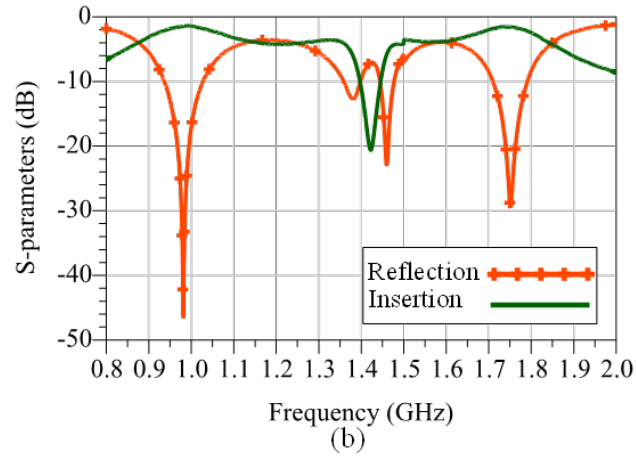
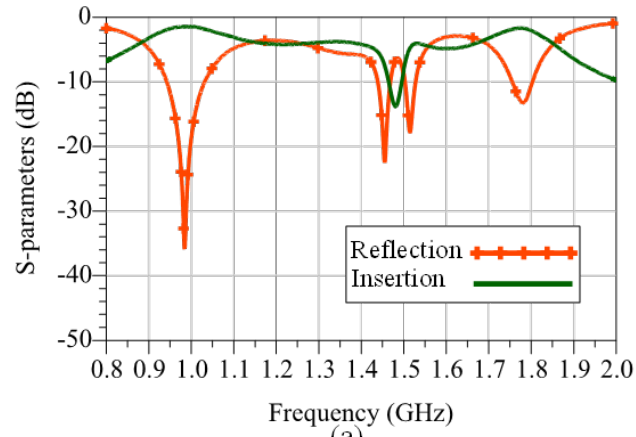


Figure 4-29 Measured S-parameters for the Type-II DBCMNs (a) design-A with folded shorting stubs, (b) design-A with radial stubs, and (c) design-B.

4.5 Comparison between the Proposed DBCMN Types

The proposed types of the DBCMNs can be compared in terms of their matched bandwidth, insertion loss, and size. The following points can be used to determine the appropriate type for different possible applications:

- 1) Compared with Type-I DBCMN, Type-II design has a narrower matched bandwidth around the two frequencies. This is associated with the use of hybrid couplers in dual-band RTPSs, which have a minimum 90° phase shift associated with the structure of the hybrid couplers.
- 2) The presented examples show that Type-II DBCMN presents slightly better insertion loss. This is due to the losses from four cascaded components in Type-I, in contrast with only two components in Type-II circuit. Moreover, the use of lumped series capacitor in HPF for the Type-I further degrades the insertion loss. Correspondingly, improved insertion loss is expected with fully integrated implementation (i.e. by employing series distributed or MIM capacitors).
- 3) Type-II designs are smaller than Type-I counterparts because of the area of the phase shifters. As demonstrated, the phase shifters in Type-I need to provide a maximum of 180° phase shift at each of the frequencies, while the dual-band reflective loads in Type-II need to provide a maximum of 90° phase shift. In addition, the hybrid couplers can be miniaturized by using tandem or folded branch-line designs. Further miniaturization is possible by using lumped on-chip hybrid design [81].
- 4) The use of series capacitors (and hence HPFs) in RF circuits is not preferred due to fabrication and assembly considerations. Moreover, the Type-I designs need some level of compromising due to the discrete range of values of the lumped capacitors. Hence, Type-II is preferred in terms of fabrication and assembly.

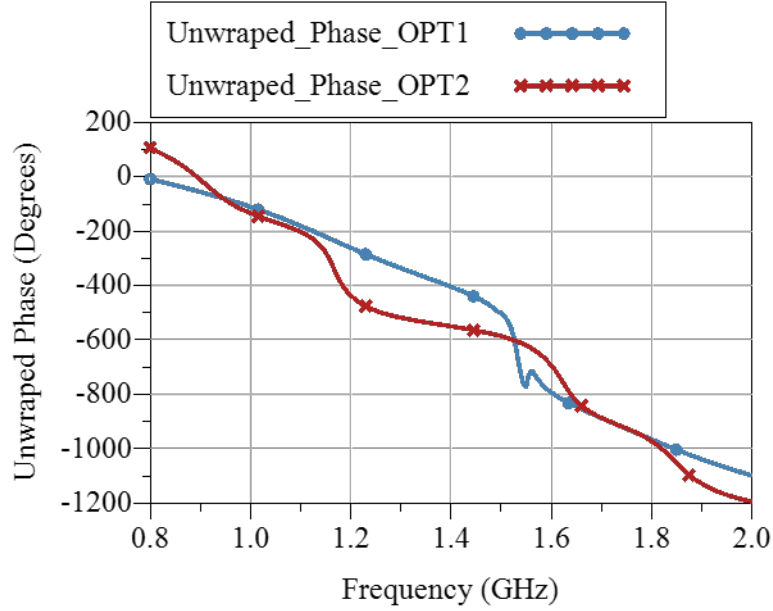


Figure 4-30 Simulated phase shift for the dual band phase shifters.

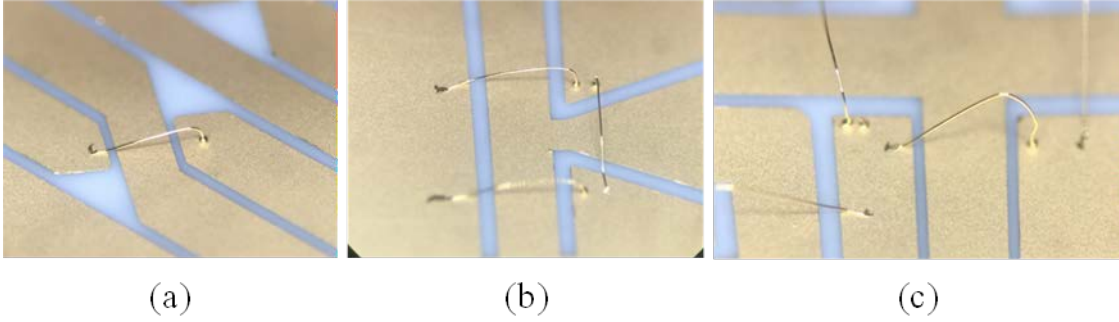


Figure 4-31 Different hight and shape of wire bonds (a) centre of tandem coupler, (b) stub T-junction, (c) filter cross-junction.

4.6 Conclusion

This chapter has presented a novel methodology for designing dual-band matching networks based on a cascaded network of filters and phase shifters. The focus has been presenting multiple implementation techniques for the proposed concept. Accordingly, solutions using HPF/LPF and single band phase shifters, and BPF and dual-band RTPSs are proposed, studied in detail, and implemented with comparable outcomes in terms of matched bandwidths, losses, physical size, fabrication considerations, and potential applications.

The shortcomings of the designed circuits can be resolved using different implementations of the filters and phase shifters in future contributions. Potential applications for the concept may include concurrent matching for dual band PAs and self-interference cancelation devices, to name a few.

Chapter 5

Study of DB-RIMNs

The results from the experiments presented in previous chapters are utilized to develop Dual-Band Reconfigurable Impedance Matching Networks (DB-RIMNs). In this chapter, the tunability of dual-band matching networks is discussed first, followed by the requirements for employing the concept of filters and phase shifters in designing dual-band tunable configurations. Accordingly, different combinations of filters, tuning elements and phase shifters can be used to realize dual-band RIMNs. The remaining part of the chapter deals with such designs and the appropriate application for them. For the reference, the intention is to design dual-band RIMNs for Sub-6 GHz band, where the centre frequency of the two bands can vary in addition to load variation at each band. We assume 5 channels for each of the lower and higher bands. The lower bands cover the 2.3 GHz to 3.8 GHz frequency range and the higher bands cover the 4.4 GHz to 5.8 GHz. The choice of the bands is made based on the 5G bands.

5.1 Dual-band Tunability

A handful of design criteria are defined in this section, which are critical for deciding the topology for the dual-band RIMNs. These criteria are predefined for different applications and can be derived from the design specifications. They include the following:

- Frequency Ratio (FR), is the difference between the frequency of the two bands it is either defined as $r = \frac{f_2}{f_1}$ or $\delta = \frac{r-1}{r+1}$, as defined in chapter 5. A dual-band RIMN requires a variable FR, which may be caused by changing the frequency of one of the bands or both bands at the same time. In this work, FR and r -value have been used interchangeably to refer to the same measure.
- Matched bandwidth is the bandwidth over which impedance matching is achieved for both bands. Dual-band RIMNs require certain bandwidths for each of the bands, which can be variable.
- Coverage range is the range of impedances that a dual-band RIMN must be able to match. This may be defined by the lowest return loss that need to be covered over each band. It might as well be a set of predetermined impedances or a specific region on the Smith chart (e.g. a range of loads on the real-axis or third quarter).

- Phase shift range is particular for the proposed method of using filters and phase shifters and refers to the range of relative phase shift needed to achieve the required coverage. The phase shift range depends on the topology of the matching network as well as the type of the phase shifter used (i.e reflection type or direct phase shifters).

5.1.1 Circuit typologies and implementation of DB-RIMNs

In this section, possible circuit architectures for realizing DB-RIMNs using filters and phase shifters are discussed. This is followed by design examples using different combination of filter types and phase shifters. We propose two circuit architectures for designing the DB-RIMNs. These circuit architectures are presented in Figure 5-1. These networks are composed of Single Band Impedance Tuning Elements (SBITEs) as their building blocks. Each SBITE is responsible for impedance matching for one of the bands and is composed of a cascade connection of a microwave filter and a phase shifter. The difference between these two architectures is at how their building blocks are connected. The cascade architecture is composed of two SBITEs connected in cascade. For this architecture, the signal at both frequencies pass through both SBITEs. For the manifold architecture, a set of manifold structures separates the path of signals. The role of the manifold structure is to allow passage of one path at each frequency band, while the other paths should appear as open circuit. The technique requires that the return loss of each branch look as open circuit termination for the other branch(es) at their matching frequencies. The choice of the architecture depends on the frequency ratio of the two bands. The cascade architecture is suitable for the cases where the bands are close to each other or moderately separated. This is because the filters and phase shifters need to be wideband enough for both frequencies. The manifold structure on the other hand, is suitable for the cases where the frequencies are far apart. This is because the manifold needs to suppress the loading between the SBITEs at different tuning states, which becomes challenging when the frequencies are close to each other [82], [83]. Note that the cascade architecture requires fewer components. However, the dependency of the performance of the SBITEs restricts the application of the cascade architecture, which will be studied in the following section. For brevity, only the cascade architecture is studied in this chapter. Two block diagrams can be studied for the cascade architecture based on the choice of tunable phase shifters. These are shown in Figure 5-2, where filters and phase shifters are tunable. In the first case, the phase shifters are wideband to cover both frequencies. In the second case the phase shifter is not required to pass the signal

at frequencies between the two frequencies, but is required to tune the phase shift at the two frequencies. In the next section, the requirements for these components are presented.

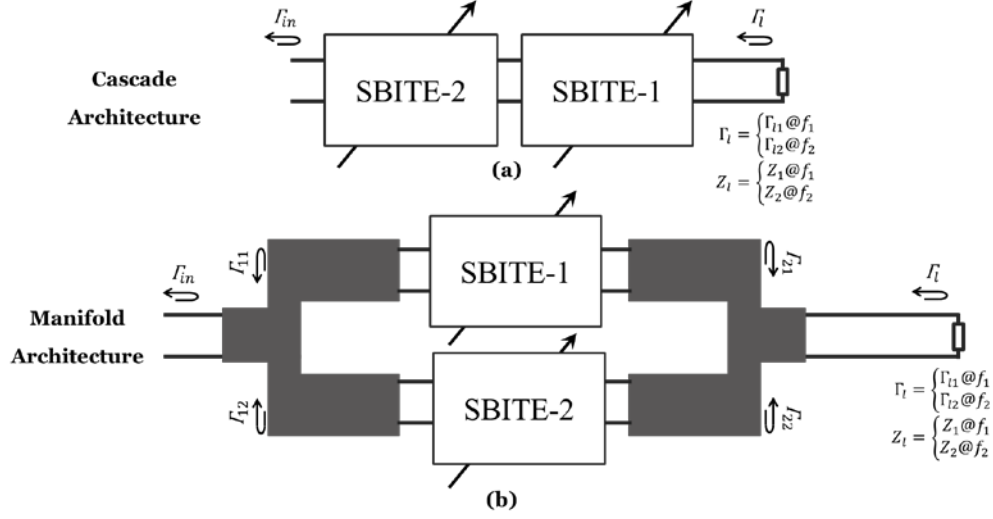


Figure 5-1 Proposed circuit architectures for the DB-RIMNs, (a) Cascade architecture and (b) Manifold architecture.

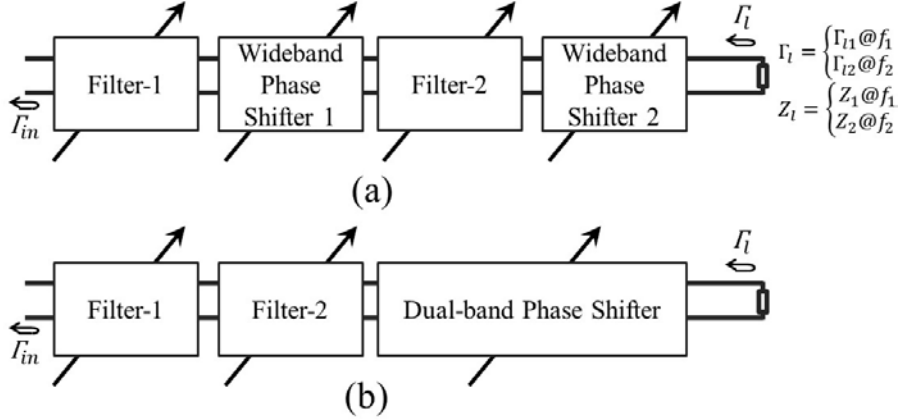


Figure 5-2 The block diagram of the proposed DB-RIMN (a) cascade architecture with wideband phase shifters and (b) cascade architecture with a dual-band phase shifter.

5.1.2 Requirements for the DB-RIMN Tunable Components

In this section, some design considerations and challenges associated with the use of filters and phase shifters in realizing dual-band RIMNs are studied. First, the interaction between the employed filters need to be addressed in the design stage. Technically, the filter response of each filter should have a minimal impact on the response of the other filters. For the cascade combination, as demonstrated in the previous chapter, the filters must have their passbands covering the cut-off edge of each other. This requirement significantly influences the circuit topology employed. For example, consider the case where LPF and HPF are employed in a cascade configuration. If r value is very small and the two frequencies are close to each other, then the order of the filters needs to be high enough to support a sharp roll-off rate for the insertion loss. This has two drawbacks. First, it is harder to tune the cut-off frequency of a high order filter while maintaining its passband response constant. Second, high order filters often have higher insertion loss over their passband, which would degrade the performance of the RIMN. In this case, the use of Band-Stop Filters (BSFs) is more convenient. It is more convenient to employ Band-Pass Filter (BPFs) instead of wideband LPF and HPF filters in the cascade topology because they can be realized using tapped resonator structures [84], which provides the designer with more flexibility over the passband performance. On the other hand, for large r values, the cascade topology is no longer suitable because LPFs and HPFs with wide passbands are impractical and BPF filters have limited band width (i.e. wideband reported performance of 72% [71]). For this case, a manifold structure is preferred. This allows the RIMNs for each band to perform independently with a minimum impact on each other. Based on this requirement, different topologies will be presented and studied in this chapter.

Second, phase-shifters must have a wide enough band width(s) to cover both frequencies. Such bandwidth is the frequency range over which the return loss is better than a set value. In this study 10 dB threshold is chosen. Simultaneously, as shown in the previous chapter, the phase shifters of a DBCMN are required to provide a maximum of 180° of phase shift at the two frequencies. These conditions can be achieved using wideband phase shifters that cover both frequencies and the bandwidth between them, or by dual band phase shifters that cover only the vicinity of the two bands. The phase shifters are studied in the following section in more detail.

Third, demonstrating the performance of the dual-band RIMNs is challenging because the demonstration must simultaneously show the coverage and independent performance over each band. To address this

challenge, the coverage at each frequency is depicted by plotting the output reflection coefficient of the network on a Smith chart. If the output return loss of all these states fall near the centre of the Smith chart at the other frequency, then the independent performance of the RIMN at these two frequencies are verified. This demonstration method must be demonstrated for both frequencies.

Finally, an intuitive yet challenging requirement for the filters and phase shifters is that they need to be low loss. This is because of the number of components involved in this type of RIMNs. Therefore, the losses of the filters and phase shifters add-up and may render the feasibility of the network. Therefore, components with lower loss and possibly dual band performance are preferred because such components reduce the overall number of employed components and lead to possibly lower insertion loss. For example, a dual band phase shifter is preferred over two single band phase shifters.

Accordingly, the following conditions need to be met for the cascade architecture to be operable, which include:

1. The responses of each filter should have minimal impact on each other. Although the added phase shifts are inevitable, the magnitude of the S-parameters must be addressed accordingly. For the cascade combination, the filters must have their passbands covering the cut-off edge of each other.
2. The phase-shifters must be able to provide a maximum of 180° of phase shift at their respective frequencies. Moreover, they must have wide enough bandwidths to cover both frequencies. Such bandwidths are the frequency range over which the return loss is better than a set value.
3. Each SBTIE shall have at least one partially matched state, for which tuning the other SBTIE provides coverage at the other frequency. This condition will be presented through demonstrating the coverage for the implementation examples in sections 5.3 and 5.4.
4. The filters and phase shifters need to be low loss. This is because of the number of components involved. Therefore, the losses of the filters and phase shifters add-up and may render the feasibility of the network.

In the following, implication of r -value for the cascade architecture is studied through implementation examples. These examples employ tunable BPFs, but can be modified to employ BSFs, depending on the r -value. In the reminder of this chapter, solutions to address these requirements are presented and studied. This includes designs for phase shifters, as well as circuit topologies that combine different filters with phase shifters. The tunable filters have been investigated in detail. However, addressing second and fourth

conditions is challenging, the first part of the reminder of this chapter is dedicated to designing phase shifters that are suitable for the proposed DB-RIMNs.

5.2 Phase Shifter Design

In this section, several phase shifter designs are presented. These designs are suitable for utilization in the proposed dual-band RIMN architectures. First, a single band direct phase shifter is presented, which employs loaded lines at its core. Although the phase shifters are not required to provide 360° for the proposed dual-band RIMNs, the proposed device covers an entire 360° of phase shift to maintain the generality of the design. Second, a 180° RTPS is presented, which employs wideband hybrid couplers. Finally, dual-band reconfigurable phase shifters are studied. These later discussions mainly revolve around the choice of hybrid coupler and the reconfigurable reflective loads.

5.2.1 Single-band Tunable Direct Phase Shifters

In this section a hybrid design, which employs continuously tunable and switched designs, is presented for realizing tunable phase shifters. Although it was determined in the previous chapter that a phase shift of less than 180° is only needed for the matching networks, the aim here is to demonstrate 360° phase shift capability for maintaining the generality of the design. The design is scalable and can be modified to achieve different phase shifts. The proposed design uses BST variable capacitors as continuously tunable tuning elements and semiconductor RF- switches or RF-MEMS switches as switching elements.

The proposed design is illustrated in Figure 5-3. It is composed of a continuously tunable 90° phase shifter and two pairs of RF-switches that select between four predetermined phase shifts of 0° , 90° , 180° , and 270° . Each pair of switches are identified by two states of up (u) and down (d). The d state in both cases corresponds to the reference phase. The continuously tunable section of the circuit adds an additional phase of up to 90° to its reference phase. The combination of switch states and continuous tuning are listed in Table 5-1.

Table 5-1 The combination of states and the resulting phase shifts

States	Phase Shift
d-d	0° - 90°
u-d	90° - 180°
d-u	180° - 270°

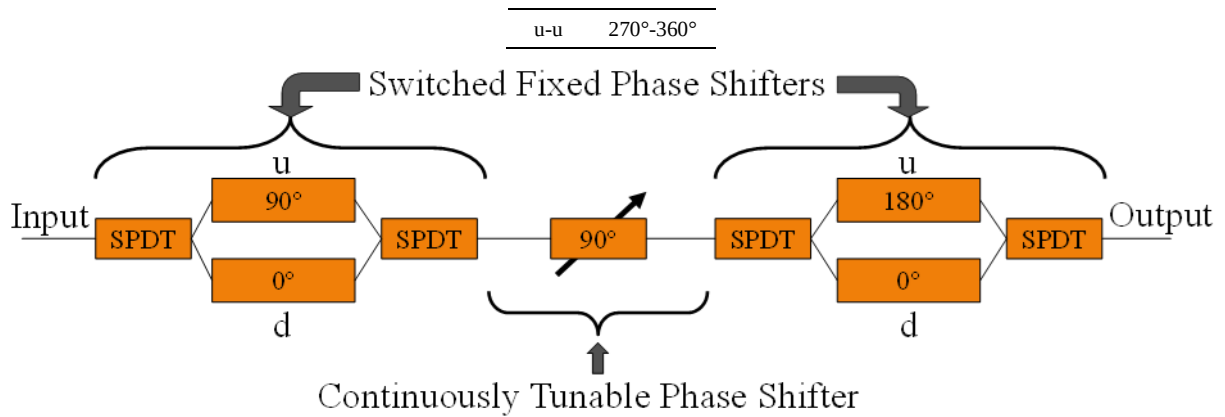


Figure 5-3 The architecture of the proposed 360° phase shifter.

5.2.1.1 Design and Implementation of Continuously Tunable and Low-loss 90° Phase Shifters Based on Loaded Lines

The design used for the continuous phase shifter is a similar loaded line architecture to the continuously tunable matching network, which employs BST tunable capacitors as tuning elements. The model is depicted in Figure 3-34. However, the main line in this design has a high impedance, the main signal line is narrow, and the tuning capacitors are attached using additional transmission line segments. In addition, the tunable capacitors are similarly biased so that the capacitors have similar values. This is because unequal capacitance values for different capacitor produces significant mismatch without significant improvement in phase shift capability [85], which is useful for designing matching networks but not phase shifters. Moreover, the ground plane for this design is not defected. This means that the connector TL segments in Figure 3-34 are removed and the high impedance TL segments are extended from the main branch and disconnected from the ground plane by a gap. This gap is the location where the tunable capacitor is effectively inserted. This combination is illustrated in Figure 5-4 and the circuit model is sketched in Figure 5-5. The high impedance sections add an inductive impedance between the tuning elements and the main transmission line. Therefore, for low capacitor values, the main line is loaded by inductive impedances. Correspondingly, when the capacitor values increase, the impedance becomes capacitive.

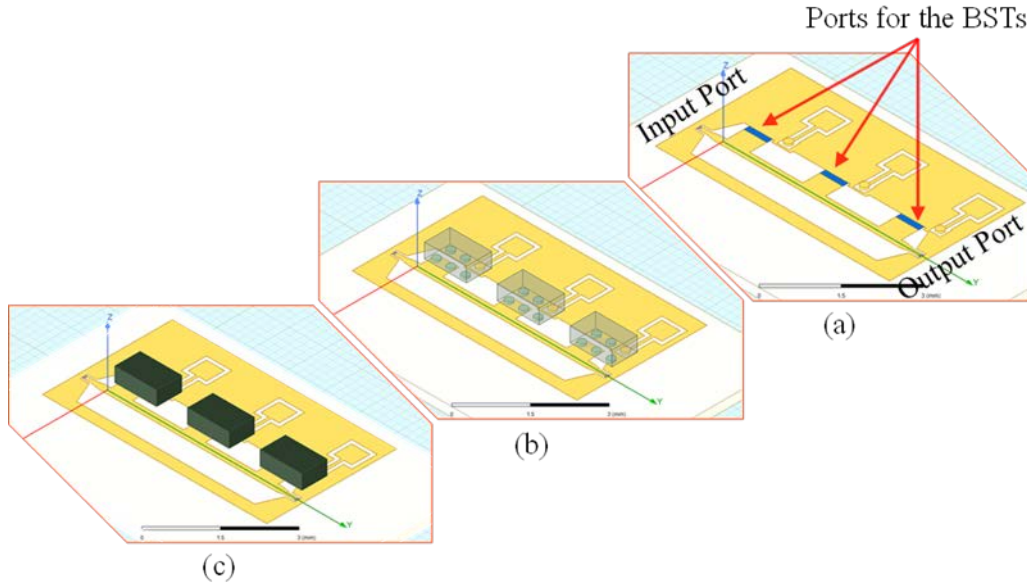


Figure 5-4 HFSS model of the proposed continuously tunable phase shifter, (a) the capacitors are modeled as lumped ports, (b) adding a transparent BST model to show the location of the leads, and (c) and opaque BST model.

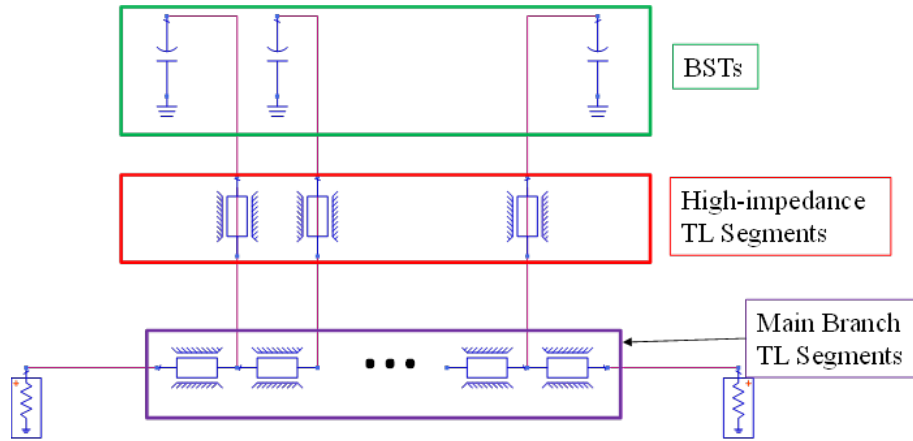


Figure 5-5 Circuit model for the reconfigurable loaded-line phase shifters.

The performance of the loaded line phase shifter is determined in terms of achievable range of phase shift, insertion loss, return loss, as well as bandwidth of operation. These indicators depend on numerous design parameters, including the dimensions of TL segments, number of tuning elements, capacitance range of the tuning elements, independent or simultaneous operation of the tuning elements, fabrication

limitations, as well as the quality factor of the tunable capacitors. However, the choice of tuning elements and fabrication technology essentially set the achievable limits for the performance indicators. In the following, BST variable capacitors in combination with a high dielectric constant substrate are selected. Realization using thin-film gold technologies, with and without additional gold electroplating to increase the conductor thickness, are compared alongside each other.

The characterization of the BST varactors is presented in section 3.4.1.2. Accordingly, circuit level study of the phase shifter for the number of tuning elements is performed. For simplicity, only three cases is considered, where 2, 3, and 4 BST capacitors are used. In all cases, the CPW lines on 25 mil Alumina substrate and 2 μ m gold thickness are considered. Although the nominal capacitance ratio for the BST varactors are 4.65, this ratio in practice is hindered by lower quality factor for higher capacitance values and self-resonance frequencies. The frequency range of 3 GHz to 4 GHz is chosen for this study because this range represent the higher end of operation frequency range for these capacitors. Moreover, a constant quality factor of 30 is assumed for all the capacitance values. The gap of the main branch dictates the length of the high impedance sections. The circuit parameters are optimized to achieve better than 10 dB of return loss, less than 0.8 dB of insertion loss, while maximizing the phase shift range. The circuit parameters are presented in Table 5-2. The simulated S-parameters and achieved phase shifts are presented in Figure 5-6, Figure 5-7, and Figure 5-8.

The reflection figures indicate that as the value of the loading capacitors increase, the return loss initially decreases before increasing. Since the main branch lines have higher than 75 Ω impedances, loading the line by capacitors decreases the characteristic impedance of the equivalent line. Therefore, as the equivalent impedance reaches 50 Ω , the return loss improves, and as it goes below 50 Ω , and then the return loss decreases again. The phase shift is calculated by subtracting the highest and the lowest phase shift from each other, which are associated with the highest and the lowest capacitance values, respectively.

Table 5-2 Optimized circuit parameters for the continuously tunable phase shifter

Parameter	Values		
Number of tuning elements	2 BST Capacitors	3 BST Capacitors	4 BST Capacitors
Minimum capacitance (pF)	0.3	0.3	0.22
Maximum capacitance (pF)	1.15	1.15	1
Main line (w-g-l) (μ m)	350-580-3620	200-600-1800	200-400-1770
Main Line Characteristic Impedance (Ω)	77	89	78
High-z lines (w-g-l) (μ m)	600- ∞ -455	600- ∞ -300	600- ∞ -350

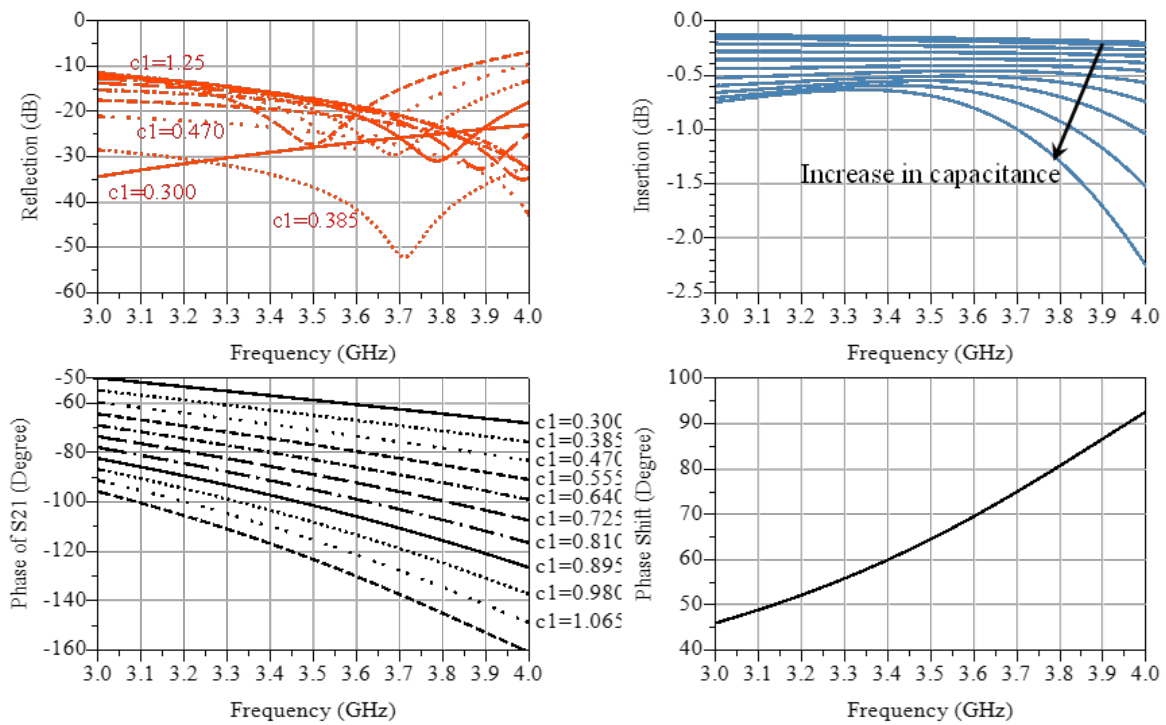


Figure 5-6 Simulated results for the 2-tuning element circuit model.

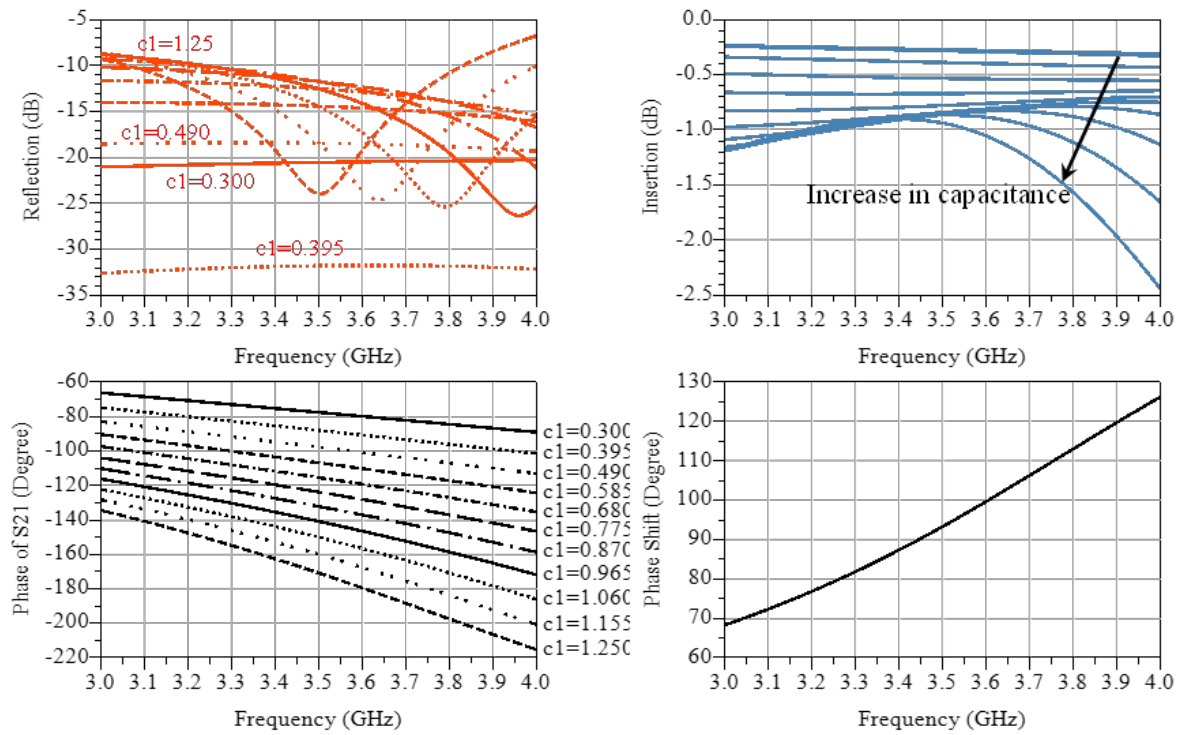


Figure 5-7 Simulated results for the 3-tuning element circuit model.

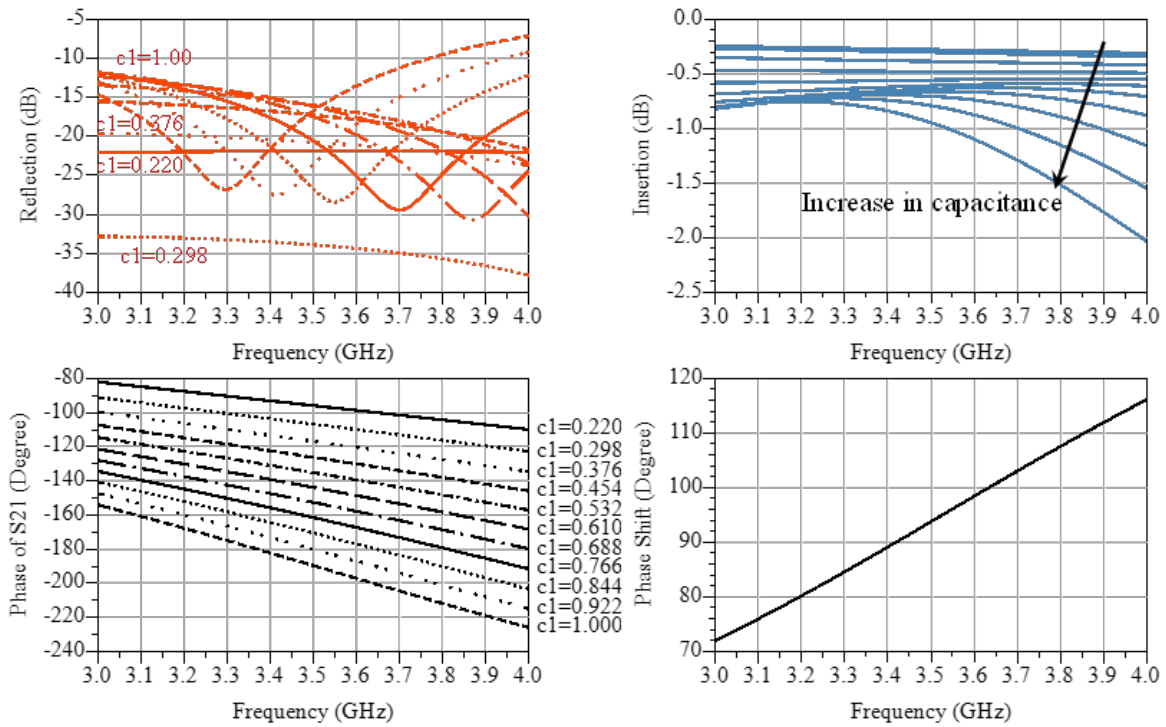


Figure 5-8 Simulated results for the 4-tuning element circuit model.

The 2- and 3- tuning element circuits are optimized for the highest phase shift. The 4-element circuit is optimized to achieve at least 90° phase shift and improved insertion loss. As can be seen, each tuning element can be used to add another 30° of phase shift. Moreover, for the same phase shift, higher number of tuning elements can be added to improve insertion loss. The lowest insertion losses are 0.15dB, 0.25dB, and 0.28dB, for the 2-, 3-, and 4- element devices, respectively. Therefore adding more elements not only increases the size of the circuit, but also increases the minimum insertion loss. Accordingly, the 3-element circuit is selected for realization because it represents a compromise between insertion loss, size, and achievable phase shift.

Figure 5-9 shows a schematic of the tunable phase shifter circuit. The width of the high-impedance TLs are set to be as wide as the BST capacitors. Moreover, transitions to 50Ω lines with $150\mu\text{m}$ pitch are added to input and output for RF-probe measurements. The circuit, including the RF- and DC-probe pads, have an area of 3.5 mm by 6.2 mm. The full-wave model for the phase shifter is simulated using HFSS. The HFSS model is shown in Figure 5-10. The location of the capacitors are modeled by lumped ports and the

s5p simulation results are exported to ADS environment where differential s5p block is used to terminate these ports in accurate BST models and tune for different capacitance values.

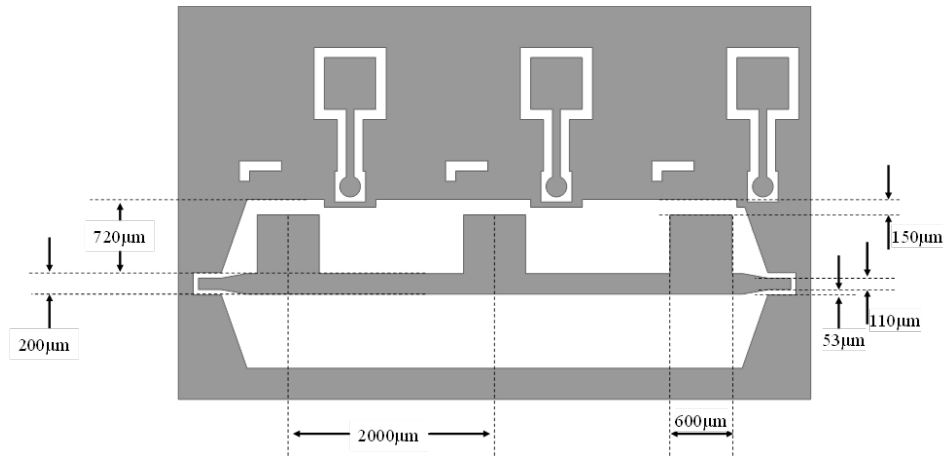


Figure 5-9 Dimensions of the continuously tunable 90° phase shifter circuit.

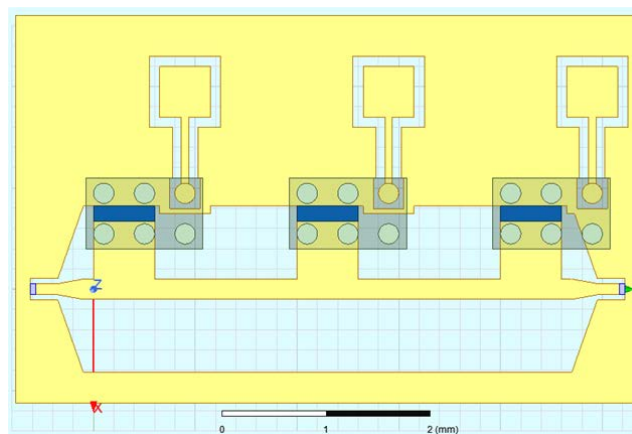
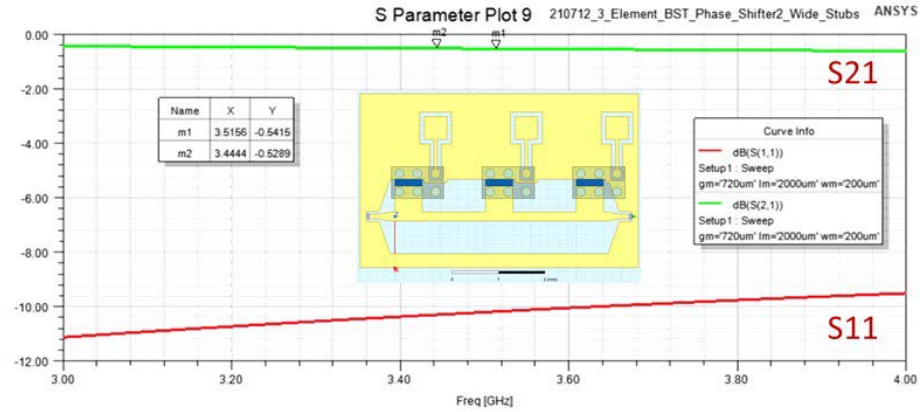
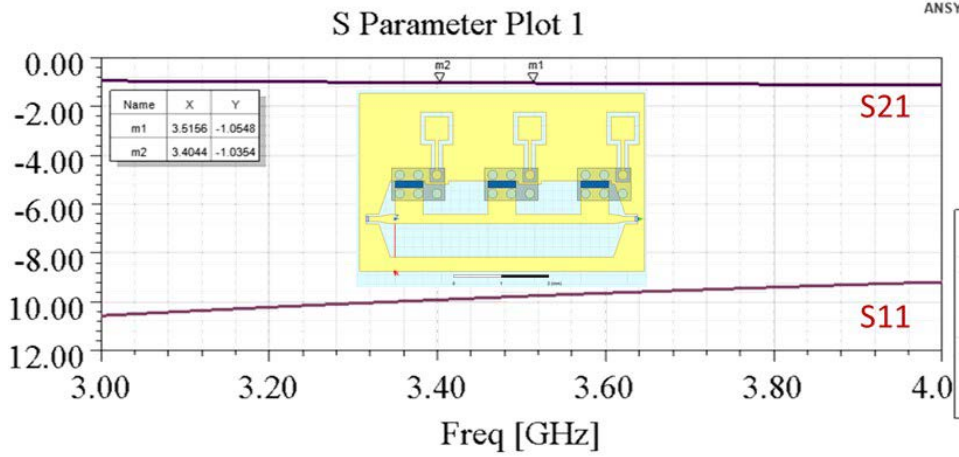


Figure 5-10 Full-wave model of the tunable phase shifter in HFSS.

Two models are simulated to study the role of gold thickness. The first circuit has a gold thickness of only 150 nm, which corresponds to the thickness achieved with only e-beam evaporation. The second circuit has a thickness of 2 μm, which corresponds to added thickness using gold electroplating. Figure 5-11 shows the S-parameters for the unloaded circuits. For generating these figures, the capacitor lumped ports are terminated with 50 MΩ impedances. As can be seen, the thin circuit has almost twice the insertion loss of the increased thickness design, while both have almost identical return losses.



(a)



(b)

Figure 5-11 Full-wave simulation results for the unloaded circuits, (a) 2 μ m gold thickness, (b) 150nm gold thickness.

Figure 5-12 shows the simulation results after the s5p models are loaded with BST models in ADS. The capacitance is swept from 0.5 pF to 1.5 pF. As can be seen, a phase shift of more than 90° over the band 3.4 GHz to 3.7 GHz band is achieved in both cases. In addition, the achieved phase shift increases almost linearly with frequency and a phase shift of 90° at the higher end of the band can be achieved by sweeping the capacitance from 0.5 pF to 1.3 pF. Moreover, the return loss is better than 10 dB for both circuits and for all capacitance values. The insertion loss given in Figure 5-12 indicate that the loss associated with the

thinner circuit is almost twice the loss for the thicker device. The marker on the insertion loss figures show the loss associated with the state that gives 90° phase shift at 3.7 GHz.

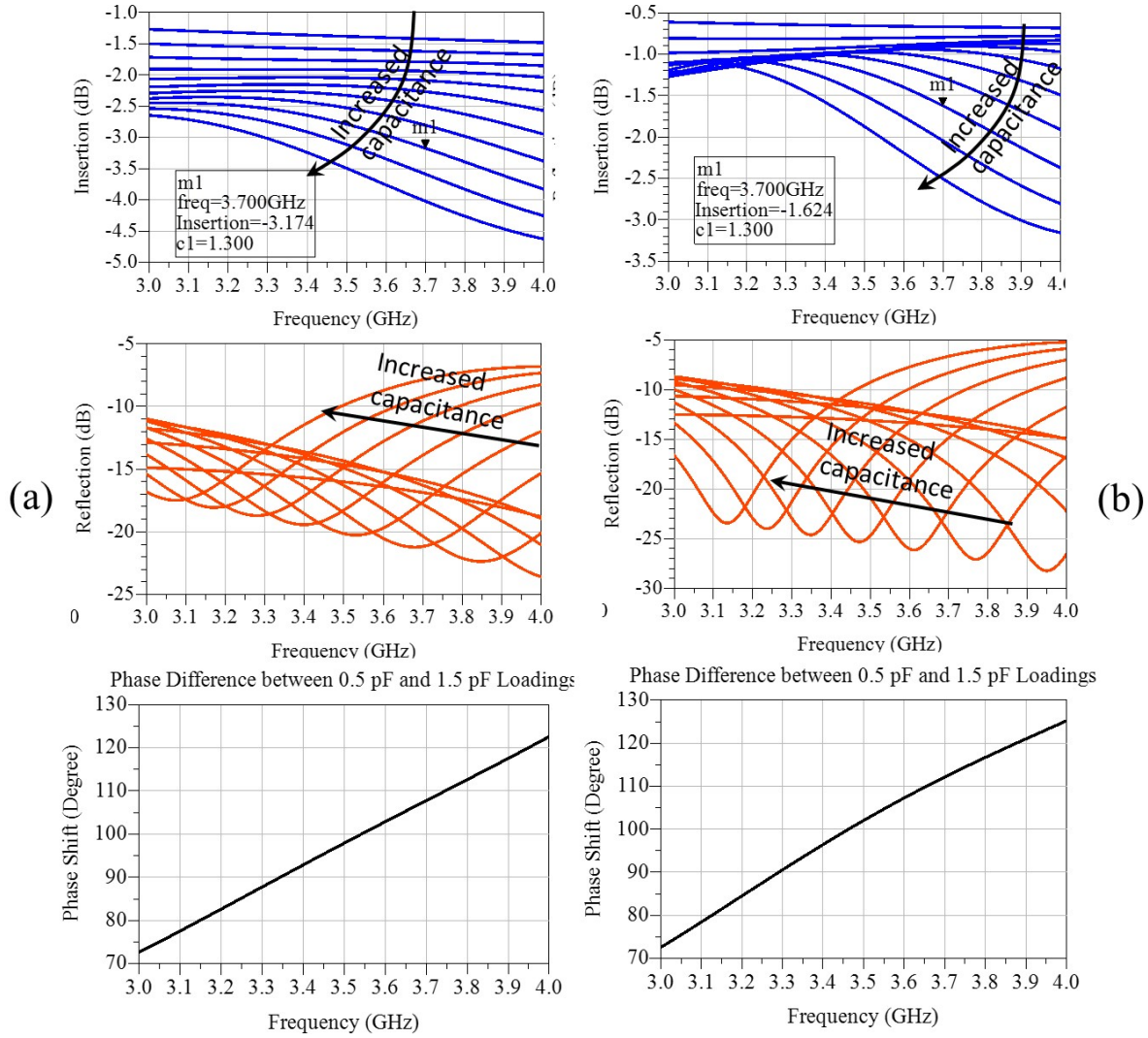


Figure 5-12 Simulation results for continuously tunable 90° phase shifter, (a) 150nm thin gold circuit, (b) $2\mu\text{m}$ thick gold circuit.

5.2.1.2 Design of the Digital Fixed Phase Shifters

The constant phase shifter segments are composed of pairs of 50Ω TL segments. The first pair is composed of two TL segments that have an electrical length difference of 90° , and the second pair is composed of two lines with a phase difference of 180° at 3.4GHz. The TL pairs are simulated and optimizing in HFSS. The

simulated full-wave models are shown in Figure 5-13. The HFSS models include wire bonds that are necessary for the performance of the CPW lines and avoid excitation of slot-line modes [78]. In mass production applications, these bonds are implemented as either air bridges [86] or automated machinery. The wire bonds are especially important near the bends to compensate the difference in the length of the CPW line slots. The height of the wire bonds are set to $200\mu\text{m}$ in the simulations. Moreover, an additional set of bonds are added for the 180° phase shifter due to its longer length. The simulated S-parameters and phase responses are presented in Figure 5-14.

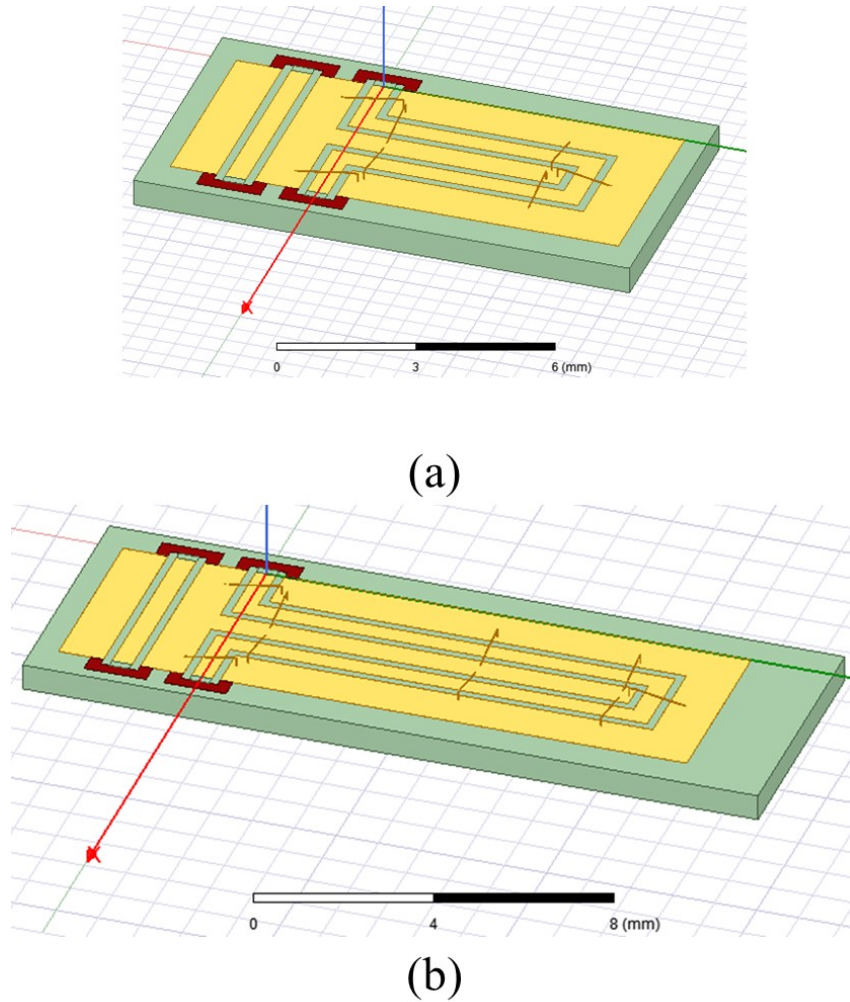
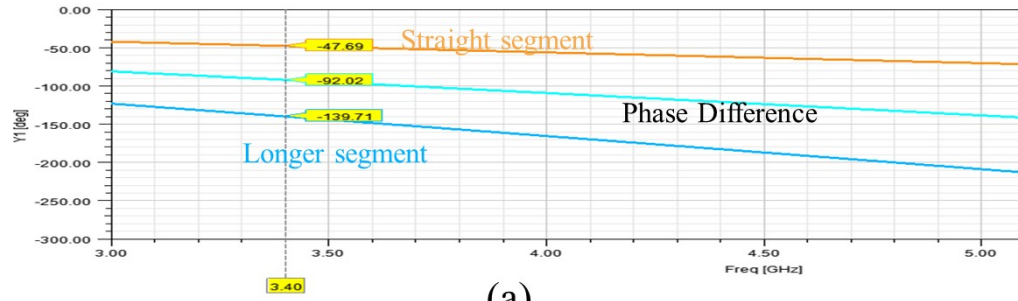
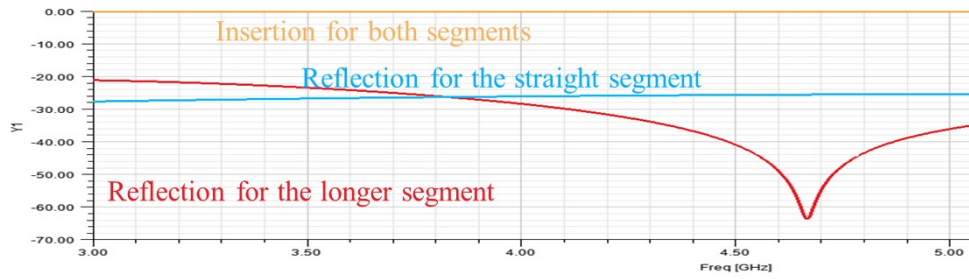


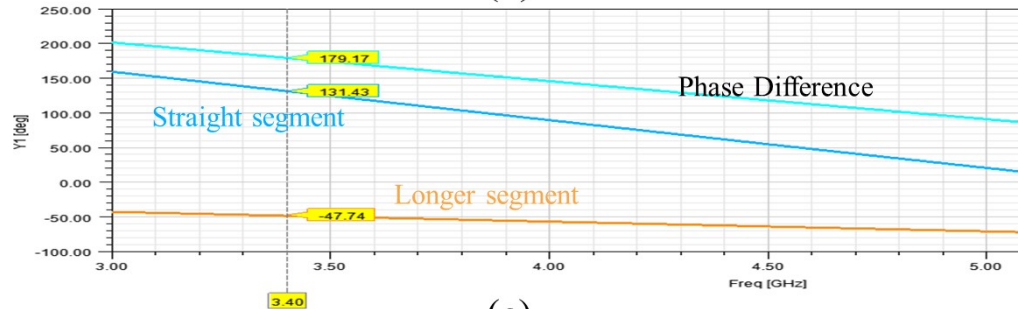
Figure 5-13 Fixed TL phase shifters with wire bonds, (a) 0° and 90° , and (b) 0° and 180° .



(a)



(b)



(c)



(d)

Figure 5-14 Simulated results of the switched fixed phase shifters, (a) phase response for the 0° - 90° pair, (b) S-parameters for the 0° - 90° pair, (c) phase response for the 0° - 180° pair, and (d) S-parameters for the 0° - 180° pair.

5.2.1.3 Fabrication of the 360° Phase Shifter

As Figure 5-3 illustrates, the continuous and the digital phase shifters are connected using SPDT switches. Semiconductor RF- switches and RF-MEMS switches can be used to realize the proposed phase shifter. For semiconductor switches, two commercially available switches from MACOM [87] and Analog Devices [88] are employed. For RF-MEMS switches, prototype SPDT switches from RADANT are used [89]. The properties of these switches are compared in Table 5-3. A handful of different layouts are designed and fabricated, which defer based on the use of straight or bent transmission lines, MEMS or semiconductor switches, as well as using phase inverters [10] (presented in Figure 5-15) or 180° transmission line segments.

Table 5-3 Properties of the RF-switches employed.

Model	IL @ 3.5 GHz	IP3	Bias Voltage	Switch Type	Technology	Package	Size
HMC544AE	< 0.7 dB	60 dBm	0V/+5V	Reflective	GaAs PHEMT	SOT26	3mm x 3mm
MASW-008853	< 0.7 dB	56 dBm	0V/+2.7V	Reflective	GaAs PHEMT	SC70	2mm x 2.1mm
RMSW220HP	< 0.33 dB	>65 dBm	0V/±100V	Reflective	MEMS	die	1.4mm x 1.45mm

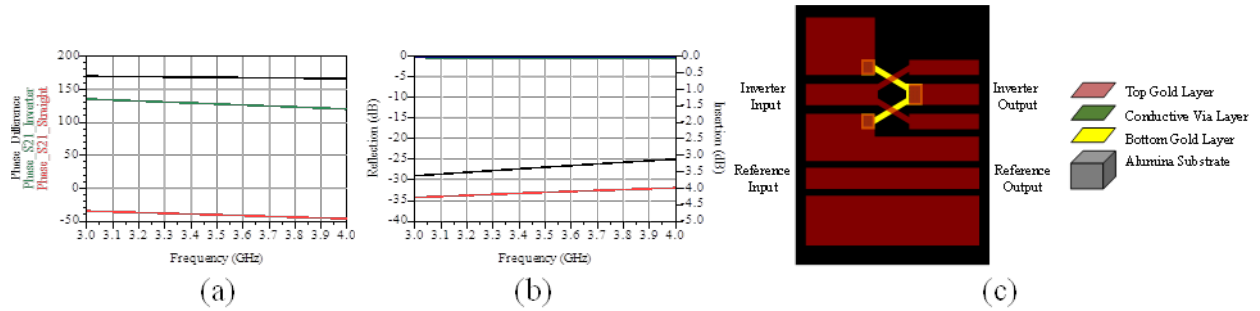


Figure 5-15 Simulated results for the phase inverter, (a) phase responses, (b) S-parameters, (c) Layout and layer map.

The overall layout is presented in Figure 5-16. The fabricated circuit uses 25 mil alumina substrates. The fabricated board has a thin layer of 150nm Gold on top of 40nm of chrome seed layer, which is deposited using E-beam evaporation. An additional 2μm thick gold layer is deposited using gold electroplating. The average thickness is measured using Bruker DektakXT profilometer. The measured thickness is 3μm and the purity of gold is 99.99%. Photographs of the fabricated circuits are shown in Figure 5-17.

- Breakdown of the circuits Phase shifters:
- ▤ 4 circuits using **HMC544AE** switches using 1.5pF BSTs
 - ▤ 4 circuits using **MASW-008853 (MACOM)** switches using 1.5pF BSTs
 - ➡ 1 circuit using Radant SPDT switches using 1.5pF BSTs
 - ▤ 4 copies of the continuous part of the phase shifters using 1.5pF BSTs

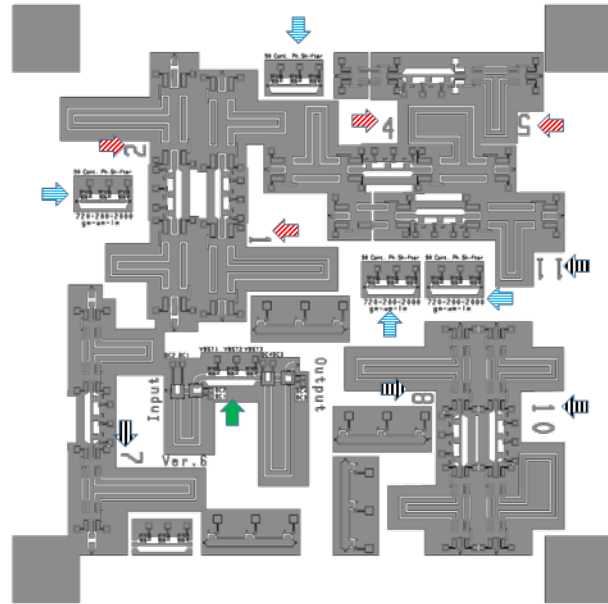
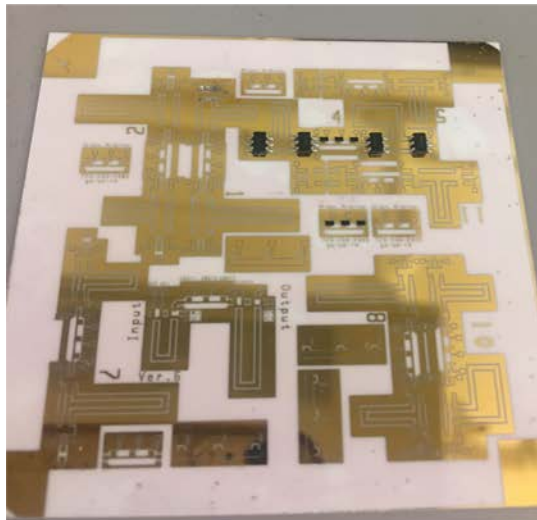
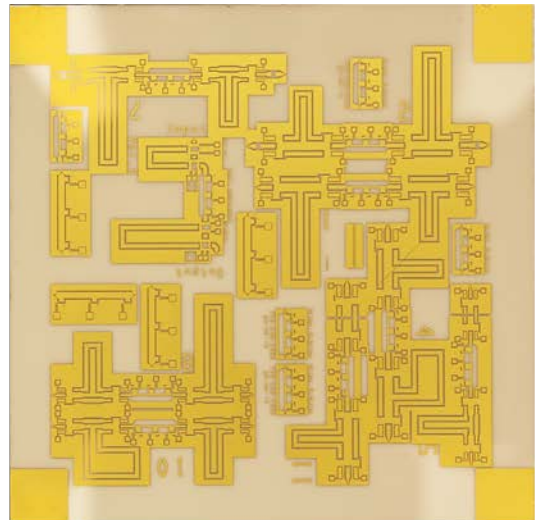


Figure 5-16 Layout description for the proposed 360° phase shifter.



(a)



(b)

Figure 5-17 Fabricated circuits, (a) a circuit after assembly, and (b) bare circuit after electroplating.

5.2.1.4 Assembly and Measurement Results for a 360° Phase Shifter Using BST Tunable Capacitors and MEMS Switches

In this section, measurement results for the fabricated circuits are presented. The BST capacitors and RF-switches are attached to the circuits using silver epoxy and wirebonds. Wire bonds are attached manually. The measurements are done using on-wafer ground-signal-ground (GSG) RF probes with 150 μ m pitch RF-probes and KEYSIGHT PNA Network Analyzer model N5227B. The measurements are performed from 2.5 GHz to 4.5 GHz.

The assembled phase shifters are shown in Figure 5-18, which includes the continuous 90° BST phase shifter as well as 360° phase shifter using semiconductor and MEMS switches. As can be seen, the bias pads for the BST capacitors are connected together using wire bonds. In addition, the RF-switched are paired by connecting their control voltages so that they can be switched simultaneously, as well as reducing the number of DC probes needed during measurement. A total number of 4 DC probes are used for the measurements. The measurements are recorded for 4 states of the digital phase shifter (i.e. switched RF-switch pairs) and for each of these states, 5 bias voltage points are set for the continuous phase shifter.

Figure 5-19 depicts the measured S-parameters for the continuous 90° phase shifter. The bias voltage is swept between 4V and 26V. The return loss is below 10 dB for all states over the frequency range of interest. Moreover, the insertion loss for the 26 V bias voltage is 0.425 dB at 3.4 GHz and 3.7 GHz. It degrades to 1.37dB at 3.4GHz and 1.4dB at 3.7GHz, when the bias voltages are reduced to 4V and 5V, respectively. The measured phase of S21 and the achieved phase shift is presented in Figure 5-20. The phase shift is calculated by subtracting the phases of S21 associated with different bias voltages. As can be seen the 90° phase shift over the frequency band of 3.4GHz to 3.7GHz is achieved by varying the bias voltage between 26V and 4V~5V, depending on the frequency. Figure 5-21 shows the measured group delay between the input and output ports, where the measured data is averaged over 40 consequent points to alleviate the measurement noise. Accordingly, decreasing the bias voltage from 26V to 4 V adds a group delay of 164ps at 3.4 GHz and 158ps at 3.7 GHz. Moreover, for the 4V bias voltage the group delay fluctuates by 9ps over the frequency range, and for the 26V bias voltage it fluctuates by 15ps over the frequency range.

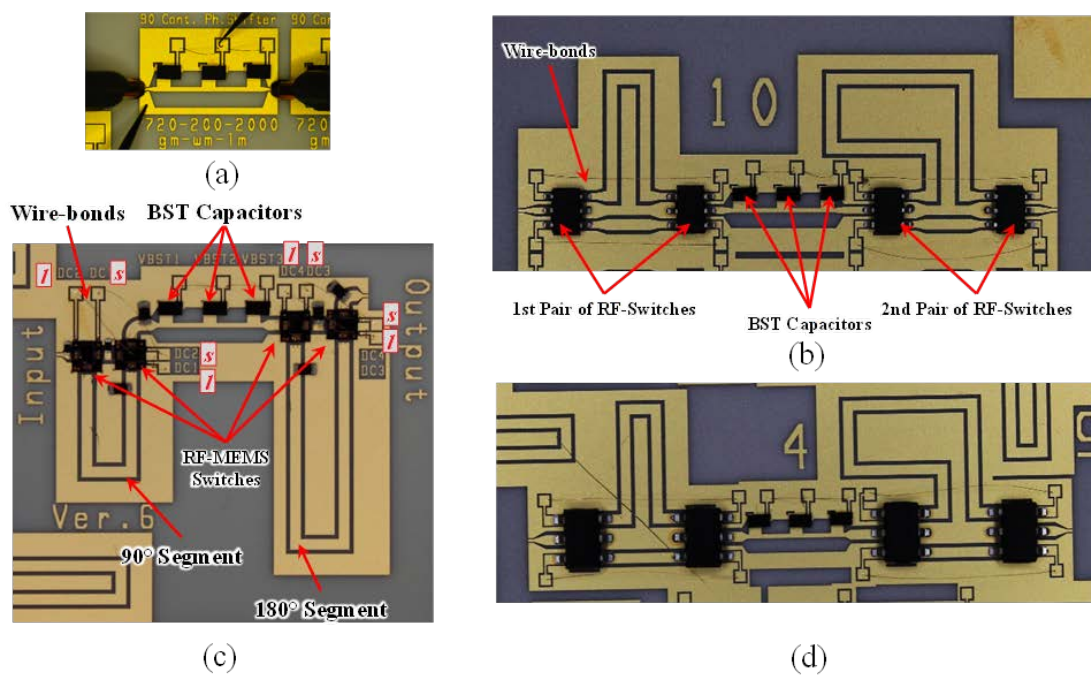


Figure 5-18 Photos of the assembled 360° phase shifters on electroplated gold circuits, (a) continuous 90° phase shifter, (b) entire phase shifter using MASW-008853 and folded 180° section, (c) entire phase shifter using RADANT RMSW220HP RF-MEMS switches, and (d) using HMC544AE and folded 180° sections.

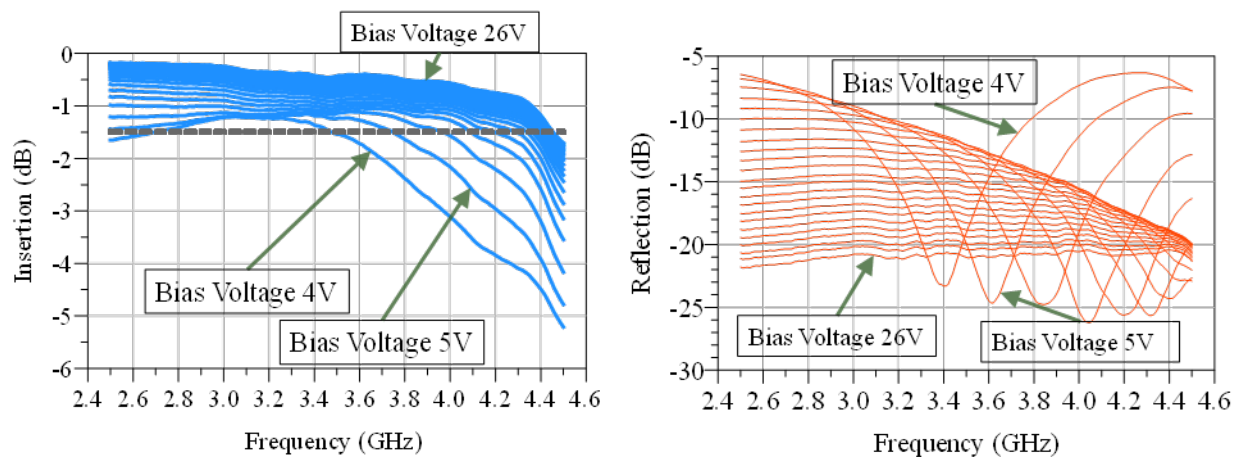


Figure 5-19 Measured S-parameters for the continuous 90° phase shifter on the electroplated circuit.

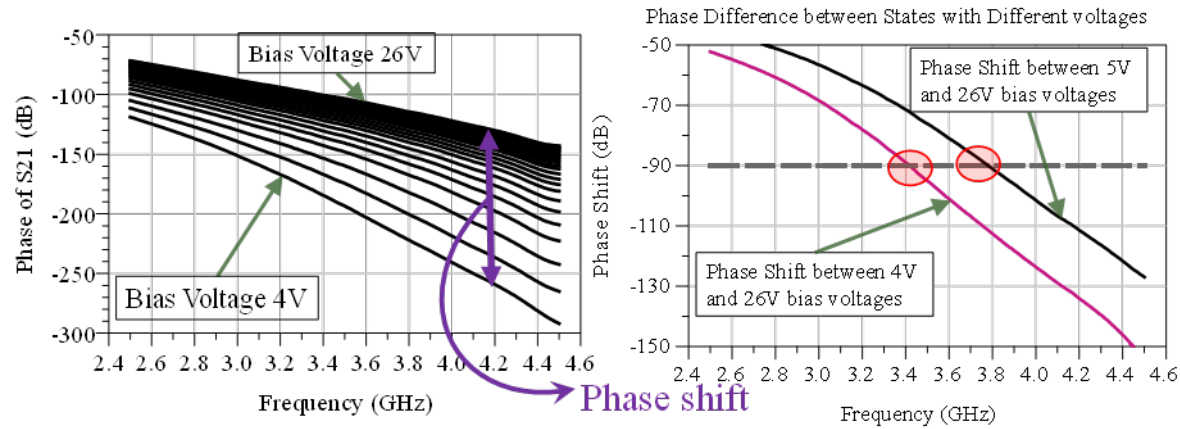


Figure 5-20 Measured phase shift for the continuous 90° phase shifter.

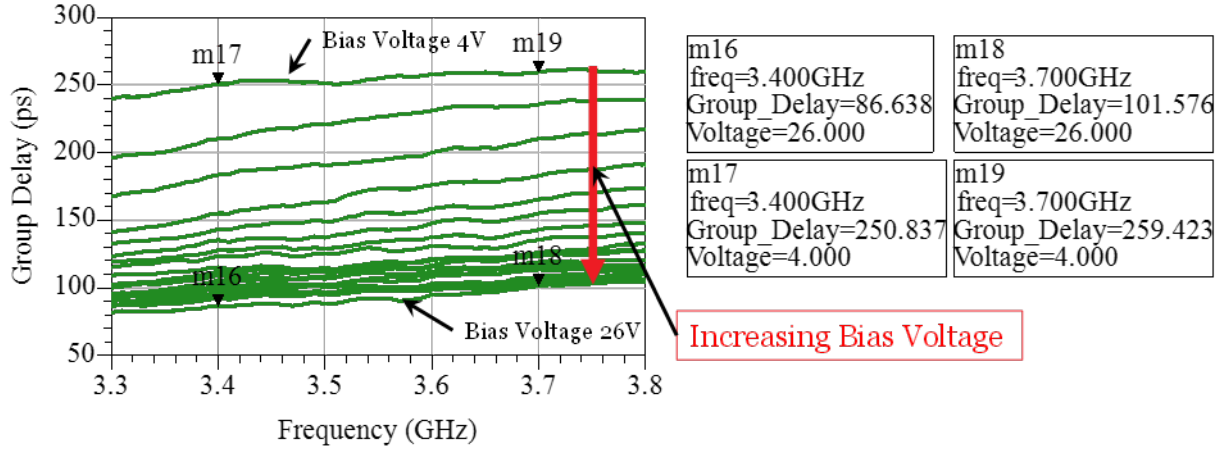


Figure 5-21 Measured group delay for the continuous 90° phase shifter on the electroplated circuit.

The measurement results for the circuit with MEMS switches is presented in the following due to the miniature size of the switches as well as their low loss performance. These switches are very low loss and are used as a prototype for demonstrating a future fully integrated design using MEMS switches. Only one circuit is assembled due to the availability of these switches. A picture of this circuit is presented in Figure 5-18 (c). Each pair of switches has two state of short (s) or long (l) which are highlighted in the photo. The bias pads for these pads are connected and these switches are controlled in pairs. Moreover, the switches need a resistor of between 4k Ω and 10k Ω connecting each port to the ground for MEMS bias paths.

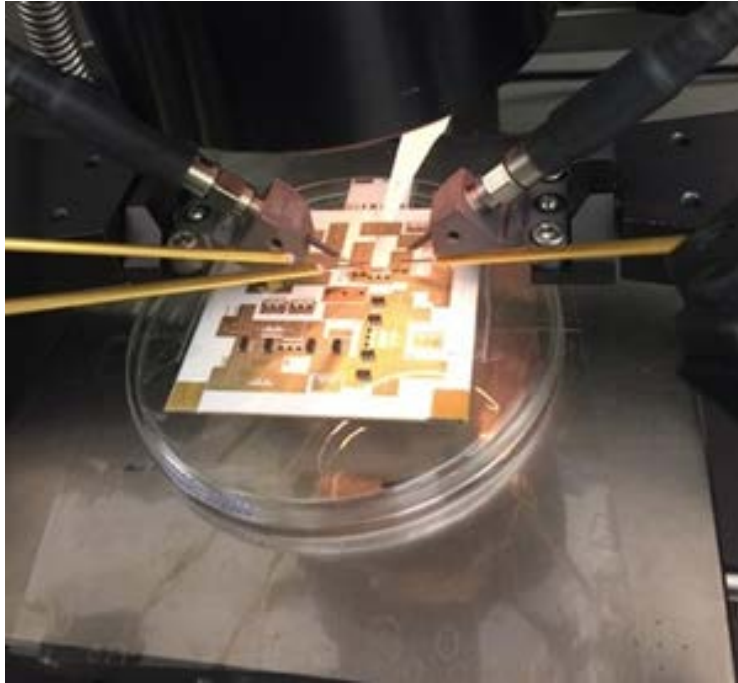


Figure 5-22 Circuit under test.

The device under test using probe station and DC probes is shown in Figure 5-22. The measured S-parameters for the RADANT 360° phase shifter with wire bonds attached over the fixed phase shifters is presented in Figure 5-23. The input and output return loss are better than 12 dB for all states. It is worth noting that the majority of states have return losses of better than 15 dB over the frequency range. Moreover, the recorded insertion losses are between 1 dB and 2 dB for almost all states. A few states have an insertion loss of worst than 2 dB on the higher end of the frequency range. These states are not needed to achieve the 360° phase shift at higher frequencies.

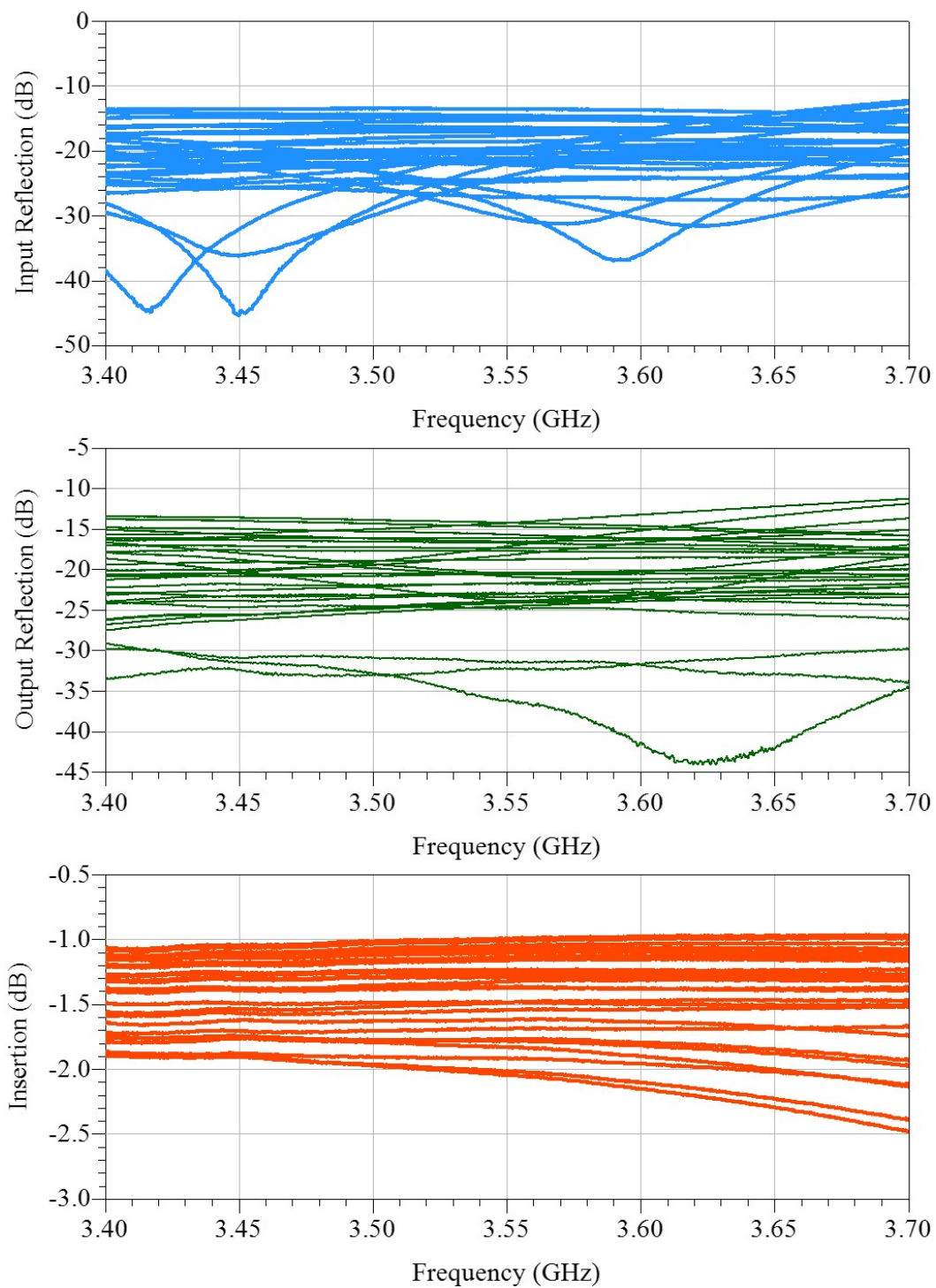


Figure 5-23 Measured S-parameters for the lifted 360° phase shifter.

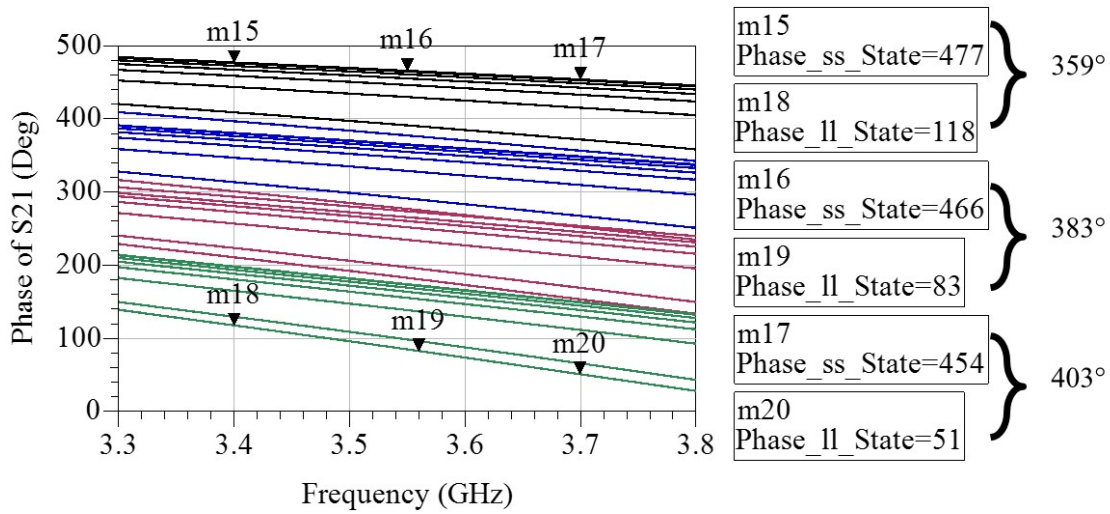


Figure 5-24 Measured phase shift for the 360° phase shifter using RF-MEMS switches on electroplated gold.

The measured phase shift is presented in Figure 5-24. The IF bandwidth during the measurements is set to 10MHz. The phase shift 359° is achieved at 3.4 GHz. Moreover, the achieved phase shift is more than 360° over the band. The measured wideband group delay and narrowband group delays over 20 MHz bandwidths around 3.4 GHz, 3.55 GHz, and 3.7 GHz are shown in Figure 5-25. The state 1 corresponds to the *ll* state with the bias voltage of 4V while the state 32 corresponds to the *ss* state with bias voltage of 18V. To maintain the accuracy of results, the group delay responses are drawn with minimal averaging over only 3 consecutive samples to account for the trace noise. The ripples in the group delay responses are associated with the unavoidable measurement noise as well as calibration error. As can be seen the group delay variance over the frequency range is the lowest for state 32, where it varies by 12 ps. The highest variance is recorded for the state 1 where it varies by 55 ps. Moreover, the group delays show very flat narrowband behaviour, except for the results around 3.4 GHz, which is due to the calibration error.

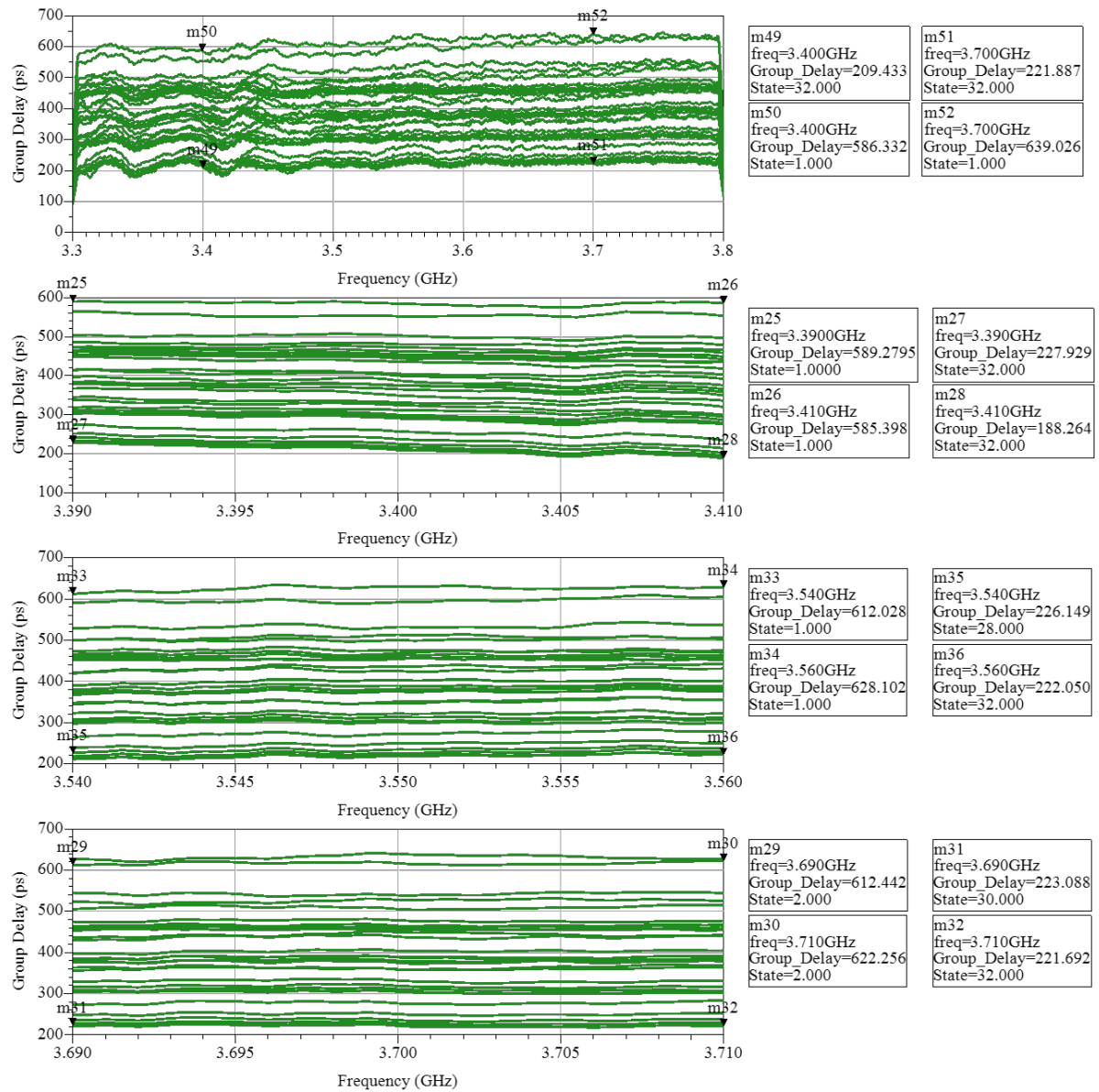


Figure 5-25 Measured group delay for the 360° phase shifter.

5.2.2 Wideband Reconfigurable RTPS Designs

Although low loss performance was demonstrated for the direct phase shifter in the previous section, its physical size, high number of switches used, and the required bias circuitry hinder its largescale application.

In this section, an alternative design is presented, which uses a single pair of tuning elements to achieve 180° phase shift. Moreover, the alternative design has a smaller size, compared to the direct phase shifter.

The intention of this section is to modify the single band fixed RTPS from the previous chapter to make it tunable. The goals for this design is to develop compact and wideband hybrid couplers so that it covers the two bands. Moreover, it is intended that a minimum number of tuning elements be used to tune the phase shifter for the sake of simplicity of design as well as minimizing the losses. Furthermore, it is desired for the phase shifter to operate using a single tuning voltage signal.

5.2.2.1 Design of Wideband Hybrid 90° Couplers

In this section, several design options are studied. The goal is to design a wideband coupler that operates over a frequency range that covers all of the bands of the dual-band RIMN (i.e. from 2.5 GHz to 5.7 GHz). This is extremely challenging and commercial planner hybrid couplers that cover this band are not available.

Other monolithic hybrid couplers include coupled line hybrid couplers, branchline couplers, and tandem couplers. Coupled line couplers can achieve ultra-wideband performance. However, such couplers need a thin gap between two long strips, which is very challenging to fabricate. For example, the simulation results for the coupler shown in Figure 5-26. It is designed on 25 mil Alumina substrate, the strips are $545\ \mu\text{m}$ wide and $9964\ \mu\text{m}$ long. However, the gap between them is only $15\ \mu\text{m}$, which impractically needs to be extended for roughly 10 cm of length. Moreover, the branch-line couplers are inherently narrowband and their CPW implementation need air bridges.

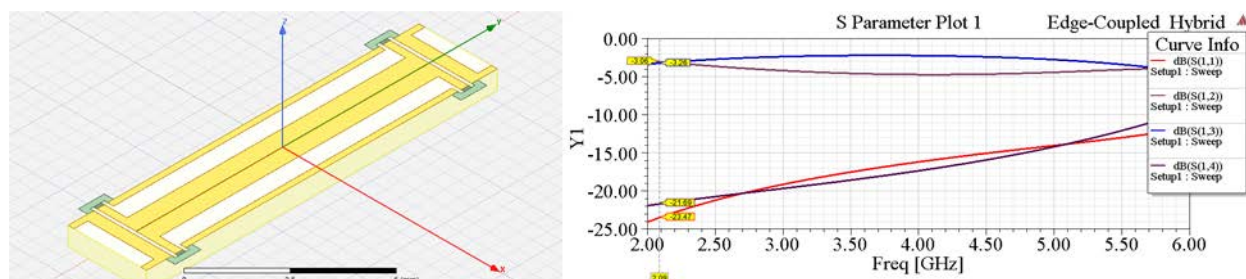


Figure 5-26 An example of coupled line hybrid 90° coupler on 25 mil Alumina substrate.

In this case, tandem couplers offer a compromise between the size and ease of fabrication. These couplers require lower coupling between the strips. The design steps include designing a three section coupled line

hybrid coupler and replace the centre section, which has the highest coupling factor with a tandem coupler which has a wider gap between its traces [90]. A tandem 3 section hybrid coupler using CPW lines is shown in Figure 5-27. The overall size of the circuit is 24 mm by 3.6 mm. for convenience all the trace widths are set to $268\text{ }\mu\text{m}$. The gap between the traces for the centre section is $65\text{ }\mu\text{m}$ and the gap between the traces and the ground plane is $120\text{ }\mu\text{m}$. The two side couplers are identical and have a gap of 1mm between the traces and $40\text{ }\mu\text{m}$ between the traces and the ground plane. The simulated S-parameters and the differential phase response are depicted in Figure 5-28 and Figure 5-29. The S-parameters indicate the usable bandwidth is between 2 GHz and 6 GHz. The phase response indicates that a $90^\circ \pm 2^\circ$ can be achieved over the usable bandwidth.

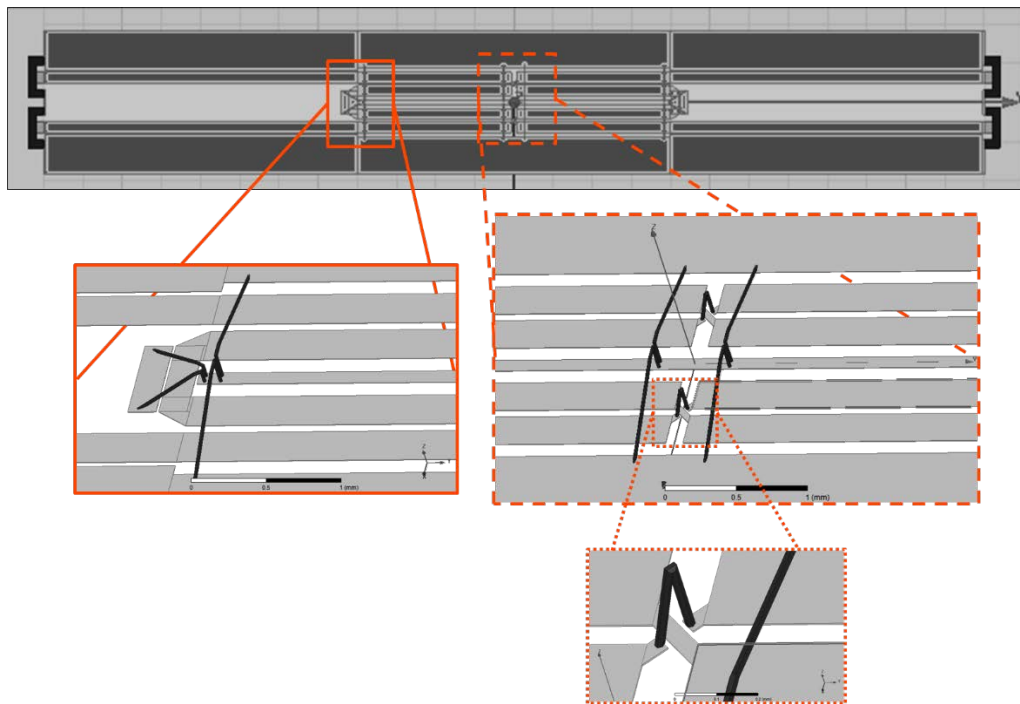


Figure 5-27 A three section tandem coupler.

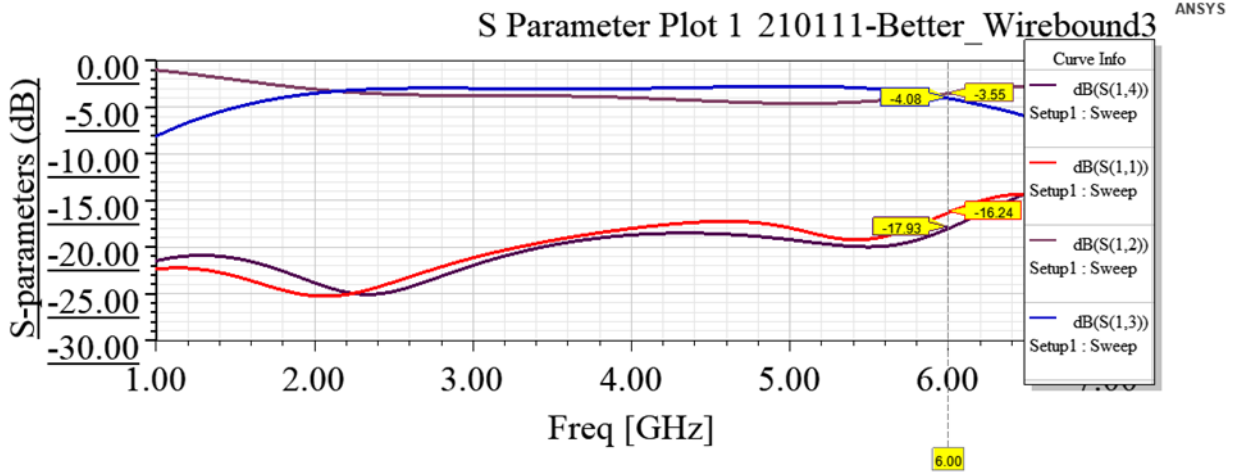


Figure 5-28 The simulated S-parameters for the three section wide band hybrid coupler

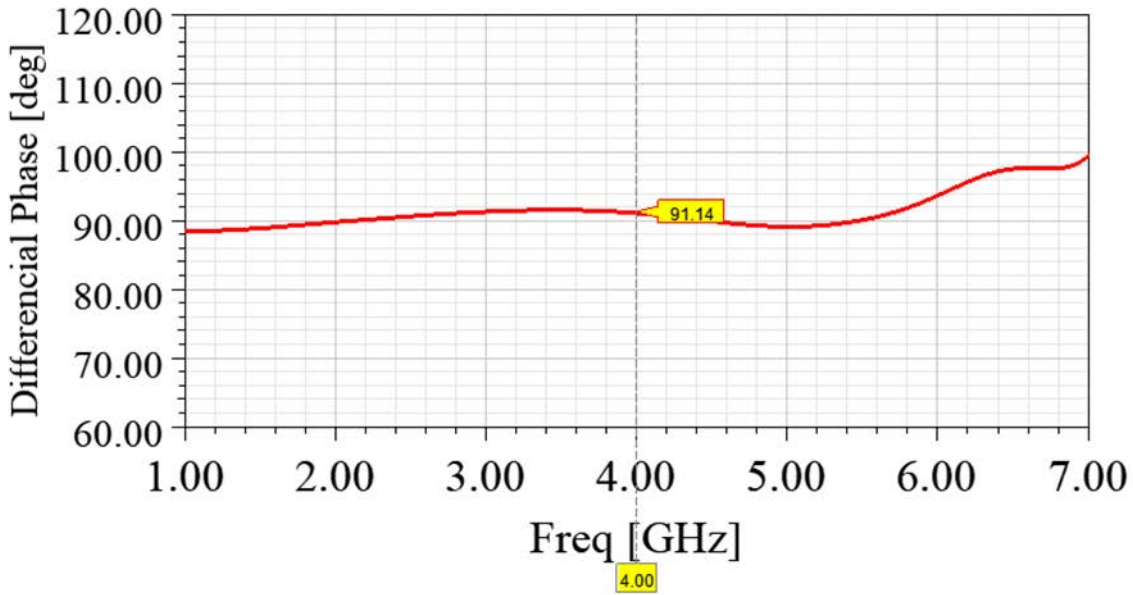


Figure 5-29 The simulated differential phase shift for the three segment hybrid coupler.

The performance of the phase shifter is first studied over a 1 GHz bandwidth (i.e. 3 GHz to 4 GHz) using a single segment tandem hybrid quadrature coupler in the following as the proof of the concept. The proposed reflective load includes a series combination of an inductive TL segment and the tunable capacitor, connected in parallel with another inductive TL segment. This combination is then connected to the ports of the Hybrid coupler using a low-impedance TL segment. The circuit schematic of the reflective

load is presented in Figure 5-30. Here the reflective load is realised using microstrip TL segments. Moreover, a microstrip T-junction model and bend are considered to increase the accuracy of the circuit model. The tuning element is a BST variable capacitor with assumed constant quality factor of 30. Figure 5-31 shows the simulated results for the single band RTPS. As can be seen, the proposed reflective load allows for at least 180° of phase shift over the 3 GHz to 4 GHz range. Moreover, the insertion loss for all the states is 0.95 ± 0.28 dB. The return loss for all the states is better than 15 dB.

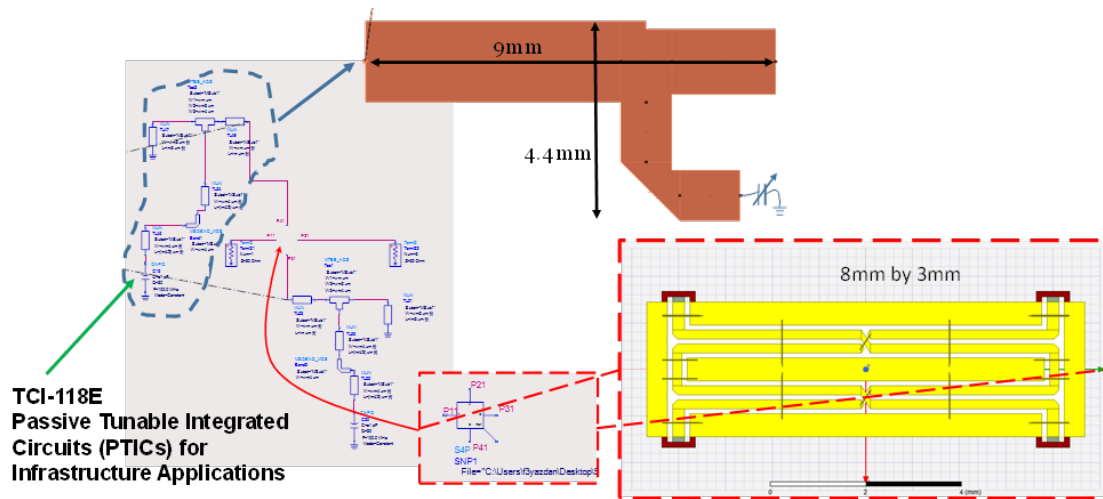


Figure 5-30 Circuit schematic of the proposed single band RTPS

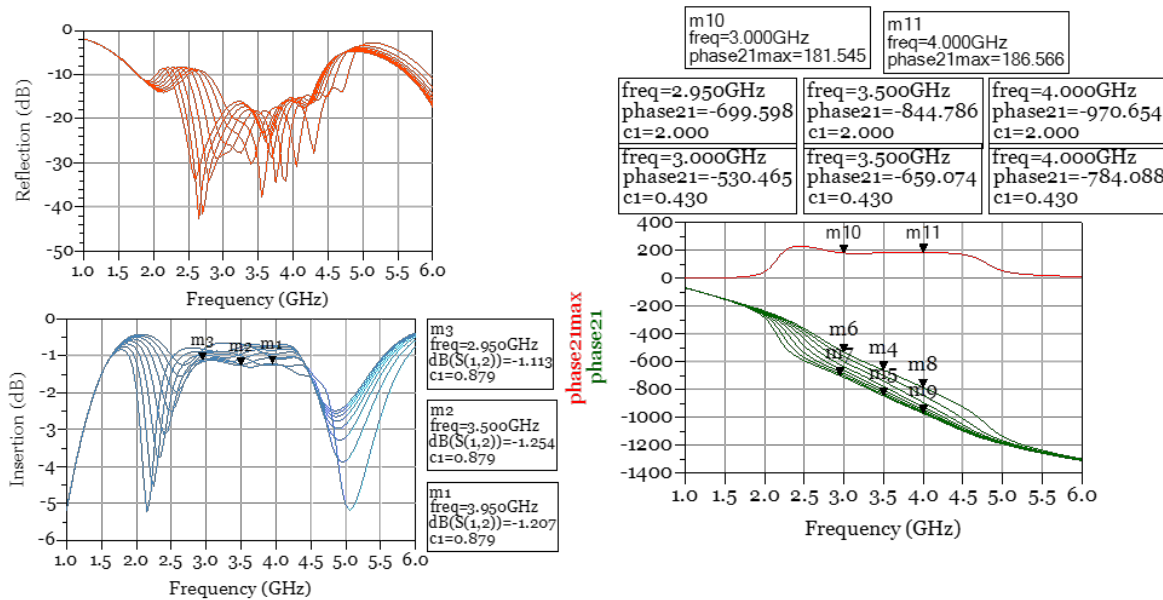


Figure 5-31 The simulated S-parameters and phase response of the circuit model for the wideband phase shifter

The results shown in Figure 5-31 meets the requirement for wideband and low loss 180° phase shift over a 1 GHz bandwidth. However, it requires cumbersome fine-tuning to achieve the wideband performance. Therefore, microstrip lines are not suitable for the implementation of these phase shifters due to the impacts of vias as short circuit terminations. Therefore, CPW and GCPW lines are used to design the reflective loads. Furthermore, to ensure a reliable coverage for the proposed DBRIMN, the designed phase shifters should achieve more than 180° of phase shift, as a reliability margin. As stated in [77], [91], and [74], the range of phase shift achieved using single tuning element reflective loads is less than 180° . As a result, more than one tuning element for each reflective load is considered in developing the proposed wideband phase shifters. Accordingly, a 3 tuning element reflective load using CPW lines and a 2 element reflective load based on GCPW lines is used. A total of 4 phase shifters are developed, which employ either the tandem hybrid coupler or a commercial hybrid coupler, XC3500M-03S model from Anaren [92], as a redundant design. The optimized layout of these phase shifters are presented in Figure 5-32.

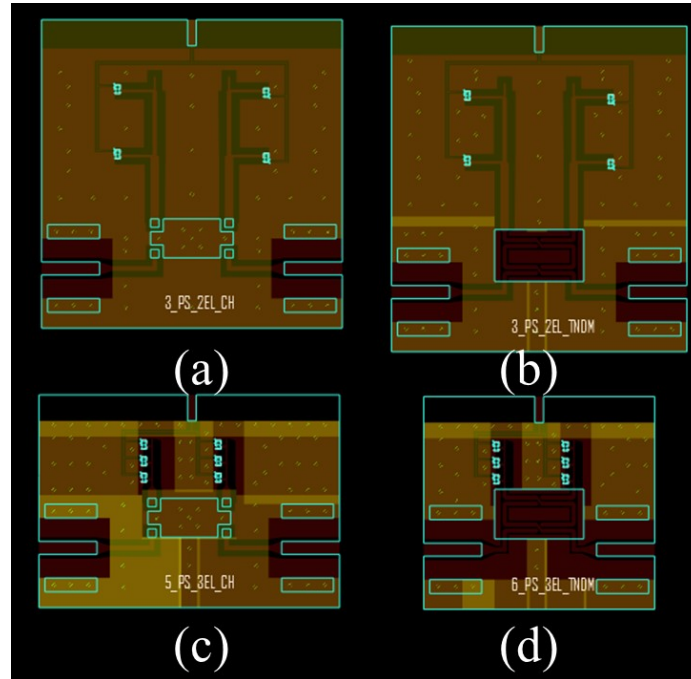


Figure 5-32 the optimized layouts of the proposed wideband phase shifters: (a) 2 element GCPW phase shifter using a chip hybrid coupler, (b) 2 element GCPW phase shifter using a tandem hybrid coupler, (c) 3 element CPW phase shifter using a chip hybrid coupler, and (d) 3 element CPW phase shifter using a tandem hybrid coupler.

The simulated results for the full-wave model are imported in ADS as a s4p file and terminated by capacitors with constant quality factor of 30. The simulated results are shown in Figure 5-33 and Figure 5-34. As can be seen, a phase shift of at least 180° is achieved, while the insertion and return losses are better than 2 dB and 15 dB throughout the frequency range of 3 GHz to 4 GHz, respectively in both cases.

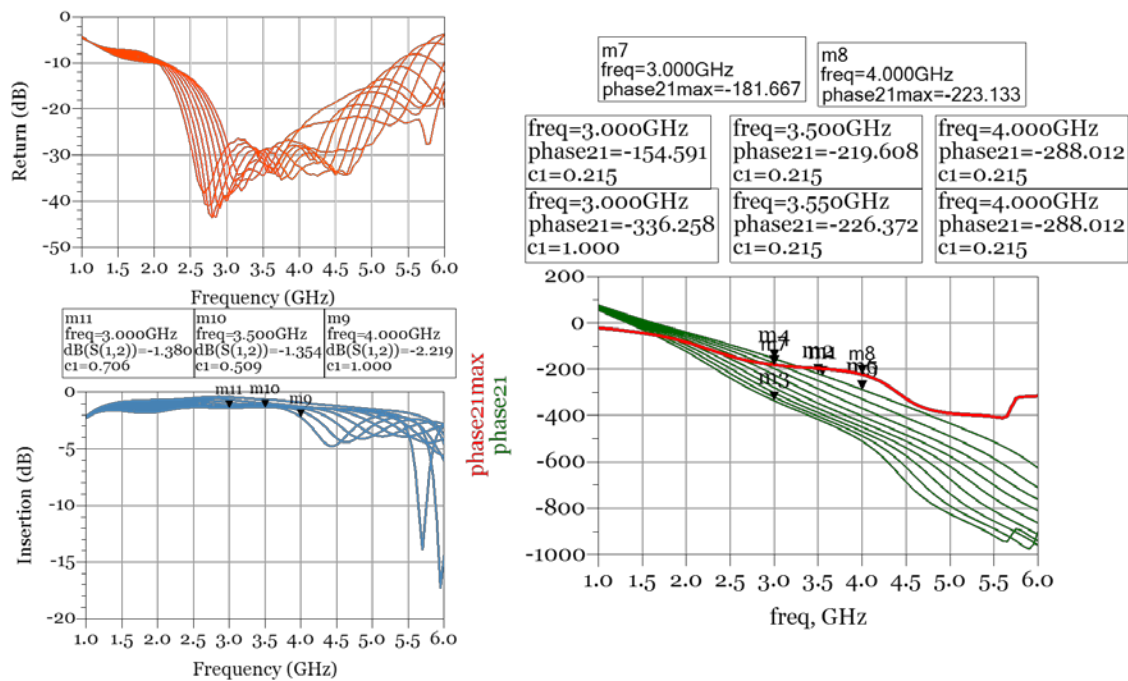


Figure 5-33 Simulated results for the 3 element CPW phase shifter

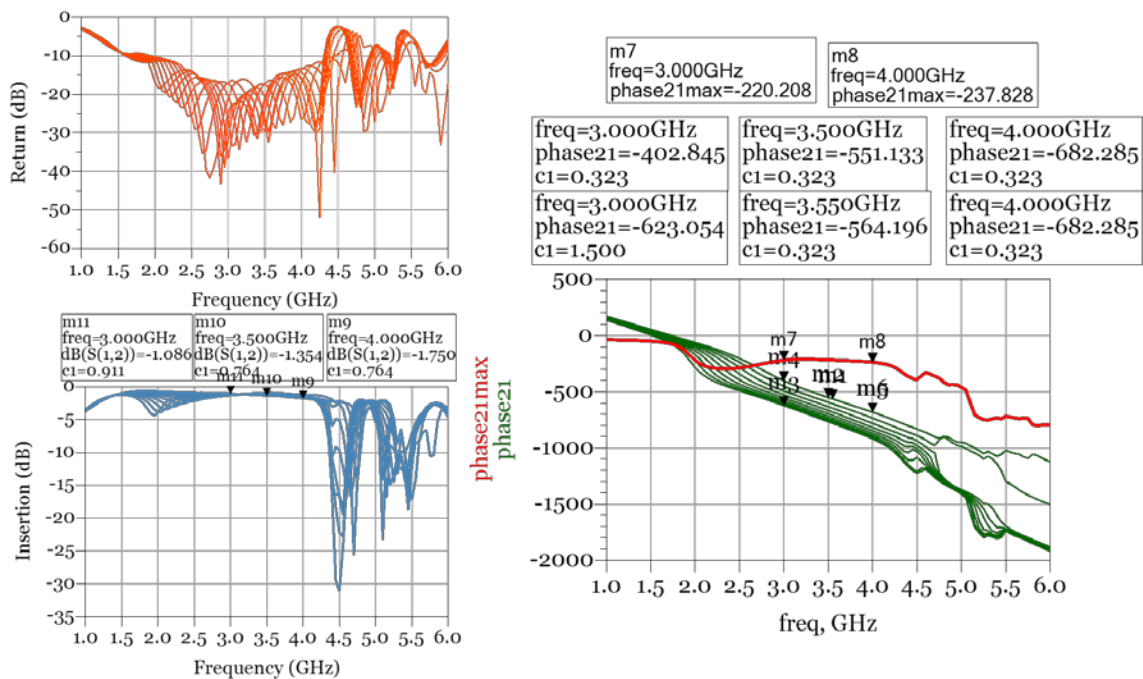


Figure 5-34 Simulation results for the 2 element GCPW phase shifter

5.2.3 Dual-band Reconfigurable Phase Shifter Designs

The bandwidth of the dual-band phase shifter is critical for determining the topology of the dual-band RIMNs. For cascade topologies, the phase shifters must pass both frequencies with minimal insertion losses. For RTPSs, the operational bandwidth of the phase shifter is determined by the bandwidth of the hybrid couplers. Accordingly, the highest limit of the r -value for the cascade dual-band RIMNs is determined by the choice of the hybrid couplers.

Current state of the art dual-band and reconfigurable phase shifter designs fall short of meeting the requirements presented in the beginning of this chapter for the following reasons. While designs are presented in the form of fixed designs [93], [94] and switched designs [95], these designs are not able to provide high-resolution phase shift and their loss often increases when their number of states increases. MMIC solutions are presented in [96]. These designs have limited quality factor and their implementation in lower frequencies becomes large and lossy. In the following, the fixed phase shifter used in chapter 4 is modified to achieve a dual-band reconfigurable phase shifter that is suitable for the proposed dual-band RIMNs.

5.2.3.1 Design of Reconfigurable Dual-band Reflective Loads

The reflective loads presented in chapter 4 can become tunable by loading each segment with a shunt capacitor in the middle. However, the lengths of each segment must be tuned in accordance with the value and tuning range of the capacitor used. This equivalent circuit model of the tunable and dual-band reflective load is depicted in Figure 5-35.

The proposed reflective load is composed of two reconfigurable segments that are terminated in the middle by a reconfigurable shunt shorting-stub. Note that the reflective load operates at two frequencies of f_1 and f_2 , while $f_1 < f_2$. The shorting stub behaves as a short circuit termination at f_2 , while acting as a shunt impedance to Impedance 2 at f_1 . Accordingly, at f_1 , the load is composed of the parallel combination of the shorting stub and the Impedance 2, which is in series with the Impedance 1. However, at f_2 , the impedance seen from the input comprises only from the Impedance 1 and the Impedance 2 is removed. Each of the mentioned impedances are tunable for frequency. Moreover, the phases of Impedance 1 and Impedance 2 are tunable so that the phase of the total reflective load can be modified.

The tunable shorting stub delivers a short circuit termination to the end of the impedance 1 at f_2 . Therefore, it only need to represent a tunable quarter wavelength segment over the higher band. On the other hand, the phase of the reflection coefficient for the impedances 1 and 2 needs to be tuned by only 90° . Since the tuning range for these segments is limited, each segment is tuned using a single tunable capacitor in the middle of its length.

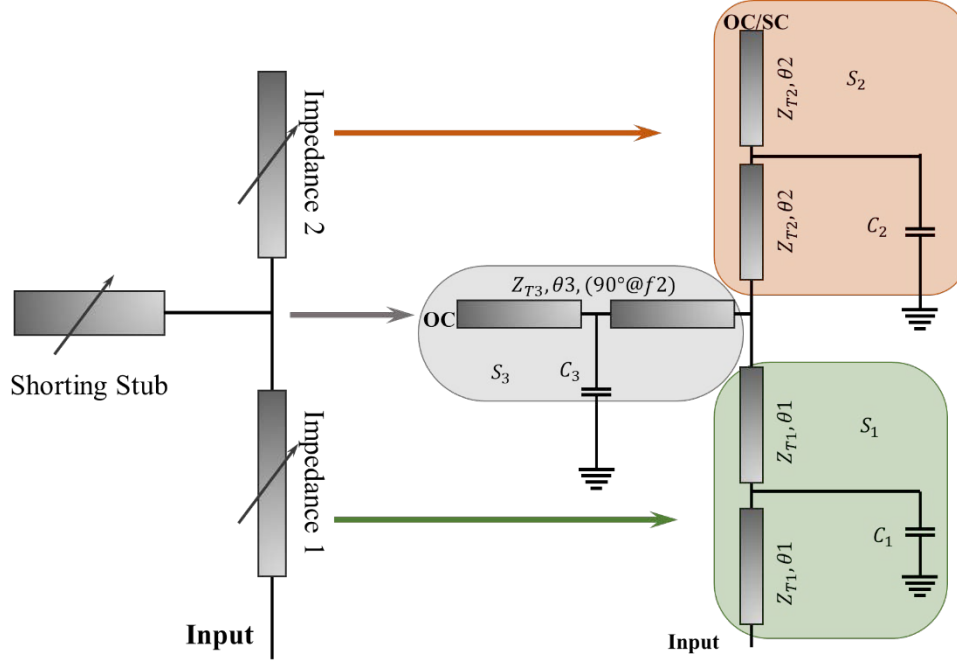


Figure 5-35 Equivalency between dual-band reflective loads and circuit model

To study the requirements for the first segment (i.e. impedance 1), we assume that it is terminated in a short circuit to the ground. For simplicity, we assume that the TL segments are 50Ω . Moreover, a ratio of 4.56 is assumed between C_{max} and C_{min} for the BST tunable capacitors, which determines the Γ_{inmax} and Γ_{inmin} of the load, respectively. This is depicted in Figure 5-36, where the difference between the phases of the input reflection coefficient is plotted for different electrical lengths. Essentially, in both cases of loading the line with C_{min} and C_{max} , the wave travelling the overall circuit is slowed. If C_{min} could reach zero, then the design would require finding a combination of electrical length and C_{max} so that loading the unloaded lines by C_{max} would result in 180° phase difference. However, derivation of these curves becomes useful in practice since the value of C_{min} in practice dependent on the C_{max} (i.e. depending on the technology used to realize the variable capacitor). As can be seen, the 45° lines cannot

achieve 180° , no matter the choice of the capacitor. Moreover, the 40° line segment design hardly achieves 180° for the capacitor with highest value of 2.7 pF. As can be seen, the 35° line can provide a phase shift of 180° for 2.2 pF and 2.7 pF BST capacitors. Note that larger BST capacitors tend to have lower self-resonance frequencies. As a result, it is more preferred to choose smaller option for higher frequency applications. For example, a 20° can achieve more than 200° of phase shift using 4.7 pF. However, if intended for 3 GHz to 4 GHz applications, it will have very lossy performance due to the low quality factor of the capacitor; according to Figure 3-31.

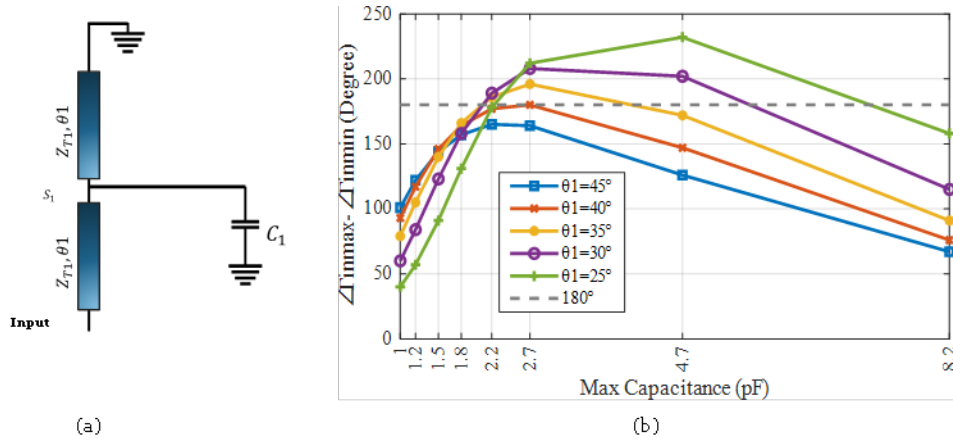


Figure 5-36 Study of the first segment for the proposed dual-band reflective loads, (a) circuit model, and (b) phase difference for the input reflection coefficient of the first segment.

The shorting stub needs to resemble an open-ended TL segment that tunes between 90° at 4 GHz and 6 GHz. This means that the imaginary part of the input impedance for this segment needs to cross the zero value over the higher band. Figure 5-37 shows the simulated imaginary part of the input impedance for a single tuning element loaded line. The TL segment on the left is open circuited. The TL segments are less than 45° at 6 GHz. Consequently, as the value of C_3 increases, the resonance frequency of the loaded line decreases. Since the value of the BST capacitors is bounded by C_{min} and C_{max} , then the range of the resonant frequencies is also limited. The corresponding range of resonant frequencies for two combinations of electrical lengths and capacitor values are shown in Figure 5-37 as well.

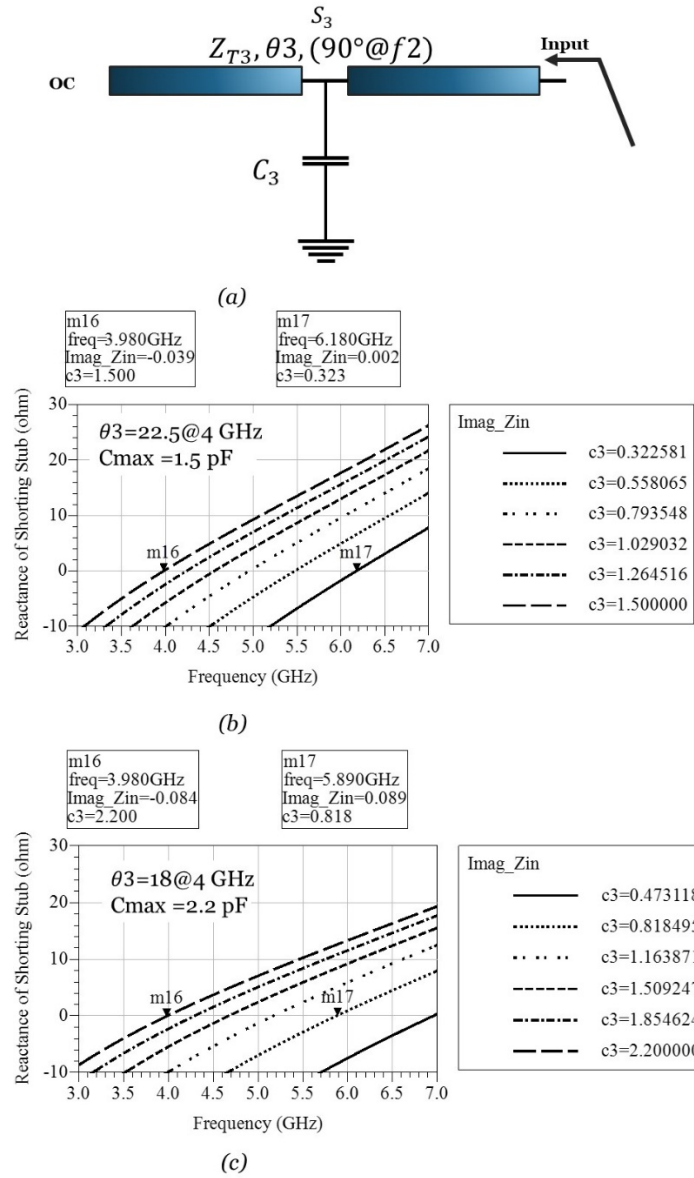


Figure 5-37 Simulation results for the input impedance of the tunable shorting stub, (a) circuit schematic of the tunable shorting stub, (b) $\theta_3 = 34^\circ$ at 6 GHz and $C_{\text{max}} = 1.5 \text{ pF}$, and (c) $\theta_3 = 27^\circ$ at 6 GHz and $C_{\text{max}} = 2.2 \text{ pF}$.

A disadvantage of using a tunable shorting stub is that it behaves as a shunt impedance for the Impedance 2 at the lower frequency band. As a result, the real part of the input impedance of the shorting stub influences the losses for the total reflective load at the lower frequency. This is illustrated in Figure 5-38, where a 90°

unloaded TL segment and a loaded line with 2.2 pF BST capacitor model are used to deliver a short circuit termination between impedance 1 and 2 at 4 GHz. The rest of the circuit elements are identical in two cases. As can be seen replacing the quarter wavelength TL segment with the tunable loaded line degrades the quality factor of the reflective load beyond usability.

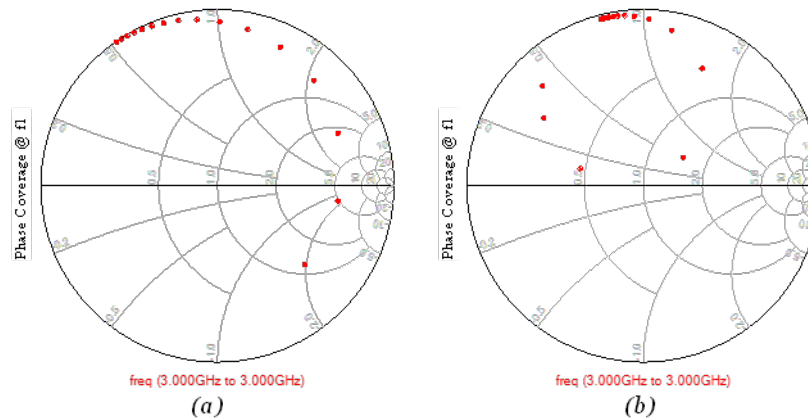


Figure 5-38 Comparing the reflection coefficient of the tunable reflective load, (a) where shorting stub is an unloaded 90° (at 4 GHz) TL segment and (b) where shorting stub is composed of two 18° (at 4 GHz) segments and a 2.2 pF BST shunt capacitor in the middle.

Alternatively, since the number of the higher frequency bands is limited to only 5 bands, the tunable shorting stub can be realized using SP4T switches. For this purpose, a 90° TL segment at 6 GHz is terminated with a switch and the output ports of the switch are terminated by additional TL segments. The electrical length of the SP4T switch (including its all OFF state) and the TL segments extends the electrical length of the 90° segment to realize shorting stubs at 5.7 GHz, 5.3 GHz, 4.9 GHz, 4.7 GHz, and 4.5 GHz. A schematic of the proposed switched-length shorting stub is depicted in Figure 5-39.

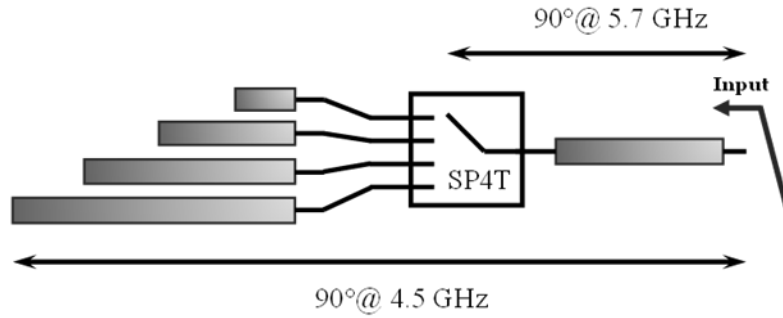


Figure 5-39 Proposed switched length shorting stub.

An example of the proposed dual-band reflective load and its input reflection coefficient at the two frequencies is depicted in Figure 5-40. The electrical lengths of the first segment and the shorting stub are expressed at 4.5 GHz. The electrical length of the second stub is expressed at 3.7 GHz. These two bands are the highest band of the lower bands and the lowest band of the higher bands. Figure 5-40 also shows the achieved phase shifts at the two frequencies. The states at 3.7 GHz are drawn for the highest and lowest value of the C_1 . As can be seen, a phase shift of more than 200° at 4.5 GHz is achieved. Moreover, for the highest and lowest values of C_1 (i.e. 2.7 pF and 0.47 pF) phase shift of more than 320° and 290° are achieved at 3.7 GHz by sweeping the values of C_2 , respectively. Note that the electrical lengths of the TL segments for the second segment are derived experimentally and depend on the value of C_2 . For example, better performance at 3.7 GHz in terms of the quality factor can be achieved when 83° segments and a capacitor with maximum value of 1 pF are employed for the second segment, which shows a trade-off between the size and the quality factor performance of the second segment.

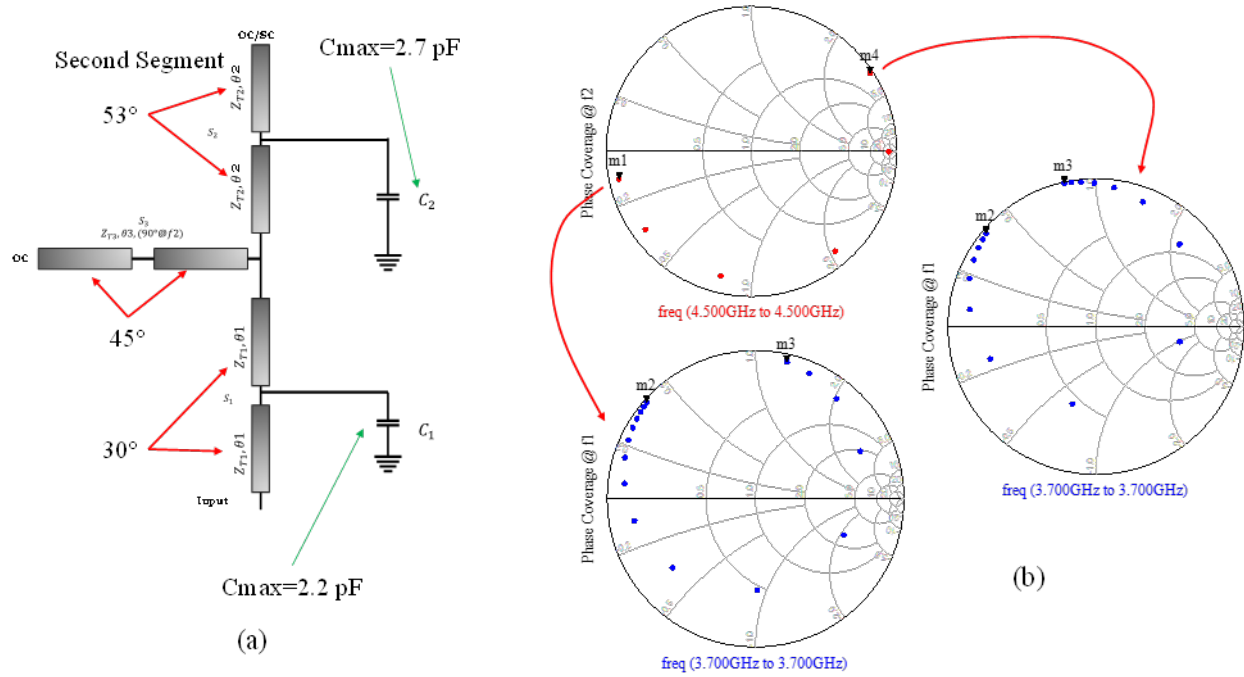


Figure 5-40 An example of the proposed reflective dual-band load and its performance at 3.7 GHz and 4.5 GHz.

5.3 Design and Implementation of Dual-band Reconfigurable Impedance Matching Networks Using BPFs and Phase Shifters

This section presents a novel methodology for designing Dual-band Reconfigurable Impedance Matching Networks (Db-RIMNs) using a cascade network of tunable filters and phase shifters. The proposed design can be used to match frequency dependent and impedance variable dual-band loads. The theory of operation, the design considerations, and the design procedure are provided. Moreover, a prototype circuit is designed for a dual-band variable load at sub-6 GHz bands. The performance of the proposed design is not restricted by the frequency ratio between the two bands or by the range of realizable characteristic impedances. Moreover, the achievable impedance coverage and frequency ratio of the two bands are not restricted by the circuit fabrication limitations. Finally, the performance in terms of matched bandwidth and losses are investigated. The simulation and measurement results for a fabricated prototype are in close agreement. The proposed design has a wide range of anticipated applications in design of concurrent dual

band power amplifiers, multiband antennas, phased arrays, to name a few. The cascade architecture of tunable filters and phase shifters is the sole published design for Db-RIMNs in the literature.

5.3.1 Introduction

Novel communication systems, including 5G cellular networks and software defined/cognitive radio technologies, call for compact, multi-band, and reconfigurable circuit components. Specifically, multi-band Power Amplifiers (PAs) and multiband antennas are widely employed in 5G systems. Simultaneously, these devices need to be tunable to cover numerous bands. Impedance tuners are fundamental microwave components that allow maximum power transfer between different components. An increasing number of multiband components such as antennas, PAs, power dividers, and baluns, among others are being introduced, which require multiband matching networks. As a result, the desired matching networks should be able to simultaneously address the variable frequency bands (shift in frequency) and variable impedance of the different components [97], [98]. In more detail, the requirements for the multiband impedance tuners include varying frequency ratios, wide impedance coverage at each frequency, low loss, and high power performance. Moreover, it is desirable for the design of tuners for each band to be independent from the other bands.

A handful of literature present dual band matching networks using multi-section transmission line circuits. For example, a parallel and series stub solution is presented in [99], and a three section series transmission line segment solution is presented in [100]. However, these solutions are suitable for fixed load impedances and they cannot be used as tunable solutions because they rely on the characteristic impedance of transmission lines, which cannot be conveniently tuned. Moreover, there is a limited range of transmission line characteristic impedance that can be practically realized. This limitation limits the range of load impedances that can be matched using such solutions.

A handful of works use switches to select the matching frequencies. In [101], a front-end architecture is proposed, which is composed of two tunable single band impedance tuners. A pair of switches select either of the impedance tuners and thus the operation band. In [102], semiconductor switches are used on transmission line transformers to achieve matching at different frequencies. In [103], a T-network matching network is loaded by two switched capacitors to provide matching at 2.4 GHz or 5 GHz. However, these solutions operate at a single frequency at a time and cannot provide simultaneous matching at different bands.

A novel design methodology for matching networks is presented in [104] based on HPF and LPF filters and transmission line phase shifters. Although this method is employed for matching a constant load impedance, its independence to the characteristic impedance makes it a suitable candidate for designing tunable matching networks. This paper tends to generalize the idea using tunable filters and phase shifters to develop Dual-band Reconfigurable Impedance Matching Networks (Db-RIMNs) for variable load impedances and variable frequency ratios. To our knowledge, this work serves as the sole design methodology and implementation technique for Db-RIMNs.

The structure of the paper is as follows. First, the tunability of dual-band matching networks is discussed by setting the requirements for employing the concept of tunable filters and phase shifters in designing Db-RIMNs. Accordingly, the design procedure and arguments regarding the different combinations of filters, tuning elements and phase shifters are presented. The remainder deals with a design example and measurements in the sub-6 GHz region, where the center frequency of the two bands can vary in addition to load variation at each band. The design example is presented for a frequency ratio as low as 1.09. Bands with such low frequency ratios cannot be matched using current fixed dual-band matching networks due to the TL impedance constraints [100]. The design example is intended to operate at sub-6 GHz region for 5G networks. Moreover, the frequencies of 3.4 GHz and 3.7 GHz are chosen as an extreme case in terms of the performance of the tuning elements.

5.3.2 Analysis of the Proposed Concept

5.3.2.1 Network Architecture

The proposed concept is intended to provide matching at the two frequencies independent from each other. This is accomplished by employing two Single Band Impedance Tuning Elements (SBITEs). Each SBITE provides impedance tuning at one of the bands, but has a limited impact on the matching of the other band. Each SBITE can be realized by a cascade combination of a microwave filter and phase shifter as shown in Figure 5-41(a), where arrows denote the tunability of these components. For an arbitrary SBITE terminated in a load with a reflection coefficient Γ_l , the input reflection coefficient can be written as

$$\Gamma_{in}^{SBITE} = S_{11}^{SBITE} + \frac{S_{12}^{SBITE} S_{21}^{SBITE} \Gamma_l}{1 - S_{22}^{SBITE} \Gamma_l}. \quad 5-1$$

Using unitary conditions for lossless and reciprocal networks, it can be shown that

$$\Gamma_{in}^{SBITE} = 0, \text{ if } S_{22}^{SBITE} = \Gamma_l^*, \text{ and} \quad 5-2$$

$$\Gamma_{in}^{SBITE} = |\Gamma_l|(\angle \Gamma_l + 2\angle S_{12}^{SBITE}), \text{ if } S_{11}^{SBITE} = S_{22}^{SBITE} = 0. \quad 5-3$$

Equation 5-2 refers to a condition where the SBITE is complex conjugate matched to Γ_l at its output, which in return leads to no power reflected back from its input port. A SBITE in this state is called a “matching” SBITE. On the other hand, 5-3 refers to a condition where the SBITE is matched to Z_0 . In this case, the input reflection coefficient has the same magnitude as the termination load, but with a modified phase. A SBITE in this state is called a “passing” SBITE.

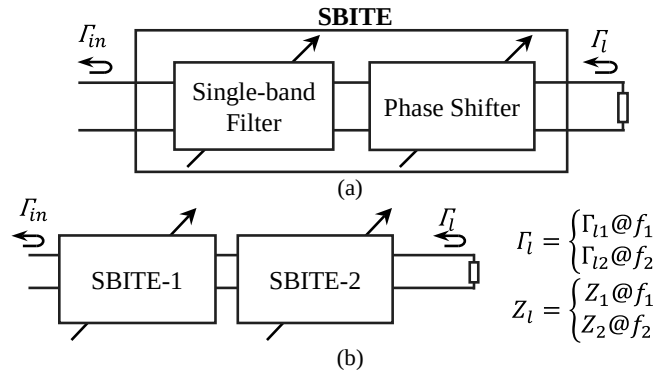


Figure 5-41 Proposed circuit architectures for the DB-RIMNs, (a) a SBITE and (b) Cascade architecture.

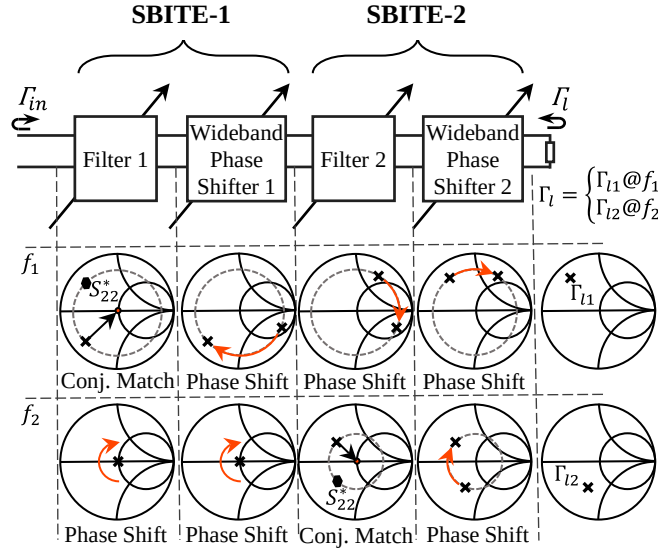


Figure 5-42 Block diagram of the proposed Db-RIMN and its step-by-step performance breakdown.

This work builds on the method presented in [104] for fixed dual-band loads, and makes a case for designing Db-RIMNs for impedance and frequency variable loads. The circuit architecture of the proposed Db-RIMN is illustrated in Figure 5-41(b). It is composed of two SBITEs connected in cascade configuration. The termination is a dual-band variable load defined by reflection coefficients Γ_{l1} and Γ_{l2} at two frequencies f_1 and f_2 , respectively. Note that any of the reflection coefficients and frequencies can be variable. For the reference $f_1 < f_2$. Each SBITE is tuned to operate as a matching SBITE at one of the bands, and as a passing SBITE at the other. Moreover, the two SBITEs should have opposite states at each band, i.e. one is passing while the other is matching and vice versa.

As illustrated in Figure 5-42, the circuit performance can be demonstrated by plotting the reflection coefficient after each block looking toward the load. At f_1 , the SBITE-1 and SBITE-2 are in matching and passing states, respectively. At f_2 , their states are swapped. To satisfy 5-2 and 5-3, the phase shifters need to pass the signal at both bands. This can be achieved using wideband phase shifters or dual-band phase shifters. Moreover, Filter 1 and Filter 2 must have transition band at f_1 and f_2 , with output reflection coefficients with the same magnitudes as Γ_{l1} and Γ_{l2} , respectively. In addition, the Filter 1 and Filter 2 must have passbands over f_2 and f_1 , respectively. Accordingly, the matching conditions can be written as:

$$|S_{22}^{Filter-1}(f_1)| \angle \left(2\angle S_{21}^{Filter-2}(f_1) + 2\varphi_2(f_1) + \angle S_{22}^{Filter-1}(f_1) + 2\varphi_1(f_1) \right) = \Gamma_{l1}^*, \text{ and} \quad 5-4$$

$$|S_{22}^{Filter-2}(f_2)| \angle \left(\angle S_{22}^{Filter-2}(f_2) + 2\varphi_2(f_2) \right) = \Gamma_{l2}^*, \quad 5-5$$

where, φ_1 and φ_2 represent the phase shifts of phase shifters 1 and 2, respectively. Therefore, the following conditions need to be met for the cascade architecture to be operable:

1. The filters used in the passing SBITEs shall have a passband over the transition band of the filters used in the matching SBITEs.
2. The phase-shifters shall provide a maximum of 180° of phase shift at their respective frequencies.
3. The phase-shifters shall be operable at both frequency bands. This can be achieved by wideband or dual-band phase shifters.
4. The tuning of the matching SBITE shall not result in significant degradation in the return loss of the passing SBITE.

In the following, the design considerations for tunable filters are studied in more detail. This will be followed by a design example as a proof of concept.

5.3.2.2 Filter Requirements

Figure 5-43 illustrates the return loss of a hypothetical LPF. This filter is used as Filter-2 in SBITE-2 that has a passing state over a lower band between f_{1L} and f_{1U} . The SBITE-2 also has a matching state over a higher band between f_{2L} and f_{2U} . The reflection coefficient of the load varies between Γ_{l2min} and Γ_{l2max} over the higher band. The return loss is also intended to be better than Γ_{f1} over the lower band to ensure the passing state for SBITE-2 over the lower band. The return loss is depicted when the filter is tuned to have a cutoff frequency shift between f_{cL} and f_{cU} , without changing the order or type. This shift in cutoff frequency allows the return loss magnitude to cover the values between $|\Gamma_{l2min}|$ and $|\Gamma_{l2max}|$ over the higher band, while the SBITE-2 maintains a passing state over the lower band. This figure can be used to determine the order of the filter and its tuning range.

The order of the filter can be determined using the slope of the return loss over the transition band. The goal is to determine the lowest order of the LPF with a return loss equal to $|\Gamma_{f1}|$ and $|\Gamma_{l2min}|$ at f_{1U} and f_{2L} , respectively. Increasing the order of the LPF beyond this minimum value would result in a better

performance at f_{1U} , but also in excessive circuit complexity. Note that the slope of return loss is in inverse relation with the frequency ratio f_{1U}/f_{2L} . Moreover, the roll-off slope is directly proportionate to the ratio of $|\Gamma_{l2min}|$ and $|\Gamma_{f1}|$.

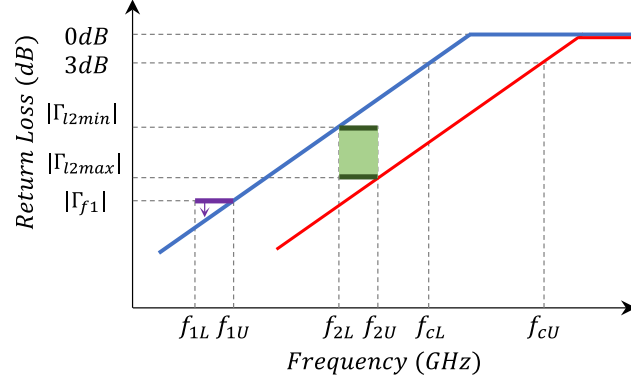


Figure 5-43 Filter requirements for a tunable LPF to provide matching at the higher band.

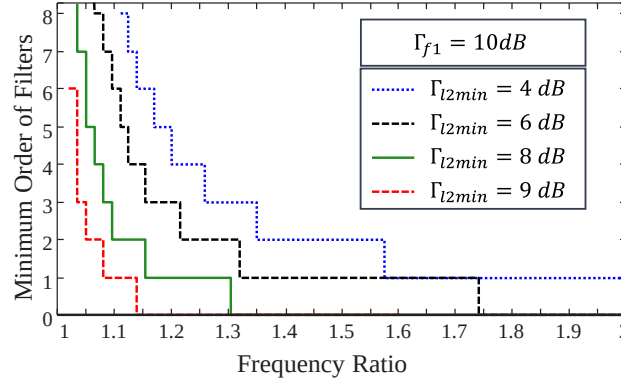


Figure 5-44 Minimum required order of a maximally flat LPF.

The reflection coefficient of a filter is associated with its power loss ratio, P_{LR} , through [11]

$$|\Gamma|^2 = \frac{P_{LR}^{-1}}{P_{LR}}. \quad 5-6$$

For a maximally flat LPF with 3dB insertion loss at f_{cL} , the power loss ratio is

$$P_{LR} = 1 + \left(\frac{f}{f_{cL}}\right)^{2N}, \quad 5-7$$

$$|\Gamma|^2 = \frac{f^{2N}}{f_c^{2N} + f^{2N}}, \quad 5-8$$

$$f_{cL}^{2N} = \frac{f_{2L}^{2N}}{|\Gamma_{l2min}|^2} (1 - |\Gamma_{l2min}|^2) \quad 5-9$$

and,

$$f_{cL}^{2N} = \frac{f_{1U}^{2N}}{|\Gamma_{f1}|^2} (1 - |\Gamma_{f1}|^2). \quad 5-10$$

Equating 5-9 and 5-10 and using logarithms to solve the equation, one can write

$$N = \frac{\frac{\log\left(\frac{1-|\Gamma_{f1}|^2}{1-|\Gamma_{l2min}|^2}\right)}{2} + \log\left(\frac{|\Gamma_{l2min}|}{|\Gamma_{f1}|}\right)}{\log\left(\frac{f_{2L}}{f_{1U}}\right)}. \quad 5-11$$

To determine the tuning range for the filter, one can write

$$f_{cU}^{2N} = \frac{f_{2U}^{2N}}{|\Gamma_{l2max}|^2} (1 - |\Gamma_{l2max}|^2). \quad 5-12$$

Dividing 5-10 by 5-12 and using logarithms to solve the equation, one can write

$$\log\left(\frac{f_{cL}}{f_{cU}}\right) = \log\left(\frac{f_{2L}}{f_{2U}}\right) + \frac{\log\left(\frac{|\Gamma_{l2max}|}{|\Gamma_{l2min}|}\right)}{N} + \frac{\log\left(\frac{1-|\Gamma_{l2min}|^2}{1-|\Gamma_{l2max}|^2}\right)}{2N}. \quad 5-13$$

If $f_{2L} = f_{2U}$ (i.e. higher band is a single frequency), then 5-13 simplifies to

$$\log\left(\frac{f_{cL}}{f_{cU}}\right) = \frac{\log\left(\frac{|\Gamma_{l2max}|}{|\Gamma_{l2min}|}\right)}{N} + \frac{\log\left(\frac{1-|\Gamma_{l2min}|^2}{1-|\Gamma_{l2max}|^2}\right)}{2N}. \quad 5-14$$

Moreover, if the magnitude of the reflection coefficient of the load does not change (i.e. load impedance is located on a constant VSWR circle on the smith chart), then 5-14 further simplifies to

$$\log\left(\frac{f_{cL}}{f_{cU}}\right) = \log\left(\frac{f_{2L}}{f_{2U}}\right). \quad 5-15$$

Similar expressions can be derived for a HPF used in a SBITE that has a matching state over the lower band and passing state over the higher band. In this case, the return loss is intended to cover the values between $|\Gamma_{l1min}|$ and $|\Gamma_{l1max}|$ over the lower band and be better than Γ_{f2} over the higher band. The minimum order of the HPF can be written as

$$N = \frac{\frac{\log\left(\frac{1-|\Gamma_{f2}|^2}{1-|\Gamma_{l1min}|^2}\right)}{2} + \log\left(\frac{|\Gamma_{l1min}|}{|\Gamma_{f2}|}\right)}{\log\left(\frac{f_{2L}}{f_{1U}}\right)}, \quad 5-16$$

which follows the same trajectory as the LPF, which is shown in **Error! Reference source not found.**. The tuning range can be written as

$$\log\left(\frac{f_{cL}}{f_{cU}}\right) = \log\left(\frac{f_{1L}}{f_{1U}}\right) + \frac{\log\left(\frac{|\Gamma_{l1max}|}{|\Gamma_{l1min}|}\right)}{N} + \frac{\log\left(\frac{1-|\Gamma_{l1min}|^2}{1-|\Gamma_{l1max}|^2}\right)}{2N}. \quad 5-17$$

Figure 5-44 depicts the minimum order of a maximally flat LPF. Note that since the filter order is an integer, the graph takes a staircase form. The minimum filter values can be plotted versus the frequency ratio f_{1U}/f_{2L} , for a given value for $|\Gamma_{f1}|$ and a few values for $|\Gamma_{l2min}|$. For instance, Figure 5-44 is plotted for $|\Gamma_{f1}| = 10 \text{ dB}$ and $|\Gamma_{l2min}|$ between 4 dB and 9dB. Note that the return loss is steepest over the transition band for $|\Gamma_{l2min}| = 4 \text{ dB}$ and flattest for $|\Gamma_{l2min}| = 9 \text{ dB}$. Hence, a higher order LPF is needed for $|\Gamma_{l2min}| = 4 \text{ dB}$ compared to the other values and for a given frequency ratio. Moreover, the minimum order of the filter dramatically increases for small frequency ratios. As given in the Appendix, similar expressions can be derived for HPF filters used as Filter 1 in the SBITE-1 by substituting the frequencies and reflection coefficients. The expressions for other types of filters (e.g. Chebyshev filters) can be derived using a similar process.

Equations 5-11 and 5-13 provide theoretical support for the proposed design method. However, for a tunable filter the passband ripple and roll-off rate would change from state to state. This is because continuous tuning is done only using tunable capacitors. Therefore, the analytical expressions can only serve as a rough estimate and the filter design would require further optimization based on the tuning elements used.

The filters should have steep roll-off while maintaining low-loss and low-ripple passband. This can be realized by a pair of HPF and LPFs. However, the order of these filters need to increase as the frequency ratio between the two bands decreases. Moreover, employing tunable HPF filters adds another challenge associated with the use of series variable capacitors [104]. Alternatively, BPF filters can be used by tuning their center frequencies. This is specifically convenient for wideband BPFs (as presented in [72]). Tuning

of the center frequency can be achieved by applying a single bias voltage to identical tuning elements attached to the end of open circuited stubs.

5.3.2.3 Design Procedure

The design steps are shown in Figure 5-45 and their description is as follows:

- 1) Characterize the dual-band frequency and impedance variable load and calculate the frequency ratio and reflection coefficient range at the two bands. The required coverage may occupy specific region of the smith chart (e.g. for optimum impedance matching of a dual-band power amplifier), or cover a ring around the center of the smith chart (e.g. for generic multi-purpose chip design, antenna matching, or high-power applications).
- 2) Decide the order and type of the filters based on the frequency ratio and the reflection coefficients range of the load.

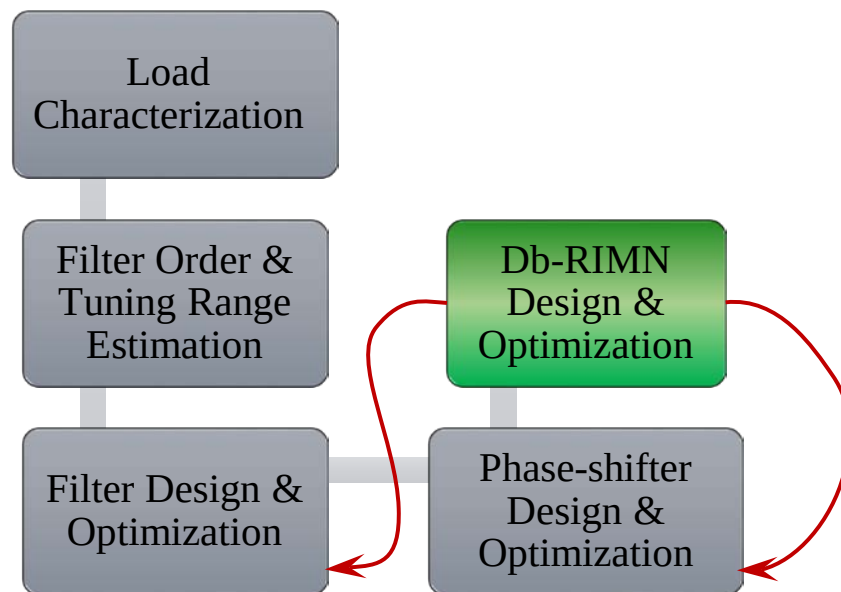


Figure 5-45 Design procedure for the proposed Db-RIMN.

5.3.3 Circuit Design

In this section, the performance of the proposed Db-RIMN is demonstrated through a design example. The two bands are at a vicinity of 3.4 GHz and 3.7 GHz. The load is assumed to be an antenna that is required to operate over two bands. The antenna impedance varies with frequency over each band with a reflection coefficient between 5dB and 10 dB and with any phase value.

For tuning Packaged Barium Strontium Titanate (BST) tunable capacitors are considered because of their relatively higher quality factors compared to semiconductor varactors. . The used tuning elements have a capacitance ratio of 4.65. The average quality factor is 30 from experimental characterization of the BST varactors. Only one bias voltage is intended for each of the filters and phase shifters to simplify the tuning process. Accordingly, the Db-RIMN circuit sub-components are developed based on Rogers RO3010 substrates with dielectric constant of 10, thickness of 25 mil, loss tangent of 0.0022, and 17 μ m copper metal layers.

5.3.3.1 Tunable BPFs

The schematic of the tunable filters is depicted in Figure 5-46. The filter is composed of two resonators that are coupled to each other using a 90° transmission line segment. Each resonator is composed of a short circuited and an open circuited stubs, which together form a parallel LC resonator. The open-ended segments for the two resonators are terminated by identical tunable BST capacitors. The filters are implemented using CPW lines and the ground planes are connected using wirebonds. Their dimensions are presented in Table 5-4 Dimensions of the TL segments for the BPFsTable 5-4.

The filters are synthesized initially without the tunable capacitors. Consequently, a full-wave HFSS model is developed where tunable capacitors are replaced by matched ports. Then, the simulated S-parameters of the model are exported to Keysight ADS, where the capacitor ports are terminated with tunable capacitors. The TL segments are optimized to mimic the behavior of the initial synthesized model and improve the insertion loss. The BSTs used for tuning Filter 1 and Filter 2 have the maximum value of 3 pF and 1 pF, respectively. Note that the entire tuning range of the capacitors are not needed. Filter 1 and the Filter 2 only need values between 0.8 pF and 1.6 pF, and 0.2 pF and 0.5 pF, respectively. The simulation results for the designed filters are plotted in Figure 5-47. Figure 5-47(a) shows the return losses of the Filter 1 loaded by a 3 pF capacitor (i.e. maximum value of the BST), and Filter 2 when it is loaded by a 0.05 pF

capacitor (i.e. miniscule value of the BST). Note that the return loss of Filter 1 shifts to higher frequencies by reducing the value of the BST. In contrast, the Filter 2 response shifts to lower frequencies by increasing the value of the BST capacitor. Figure 5-47 (b) shows the return loss response of Filter 1 for the BST values between 0.8 pF and 1.6 pF. As can be seen, the return loss of Filter 1 covers reflection coefficients of better than 4 dB at 3.4 GHz, while maintaining a passband with better than 12 dB return loss at 3.7 GHz. Similarly, Figure 5-47 (c) shows the response of Filter 2 for the BST values between 0.2 pF and 0.5 pF. Filter 2 has a return loss of better than 15 dB around 3.4 GHz, while covering reflection coefficients of better than 3 dB at 3.7 GHz.

5.3.3.2 Wideband tunable phase shifter

Current state of the art reconfigurable phase shifter designs fall short of meeting the requirements presented in the beginning of this chapter for the following reasons. Firstly, non-tunable designs [93], [94] and switched designs [95], are not able to provide high-resolution phase shift. Secondly, their loss often increases when their number of states increases. Thirdly, MMIC solutions [96] have limited quality factor and their implementation in lower frequencies becomes large and lossy.

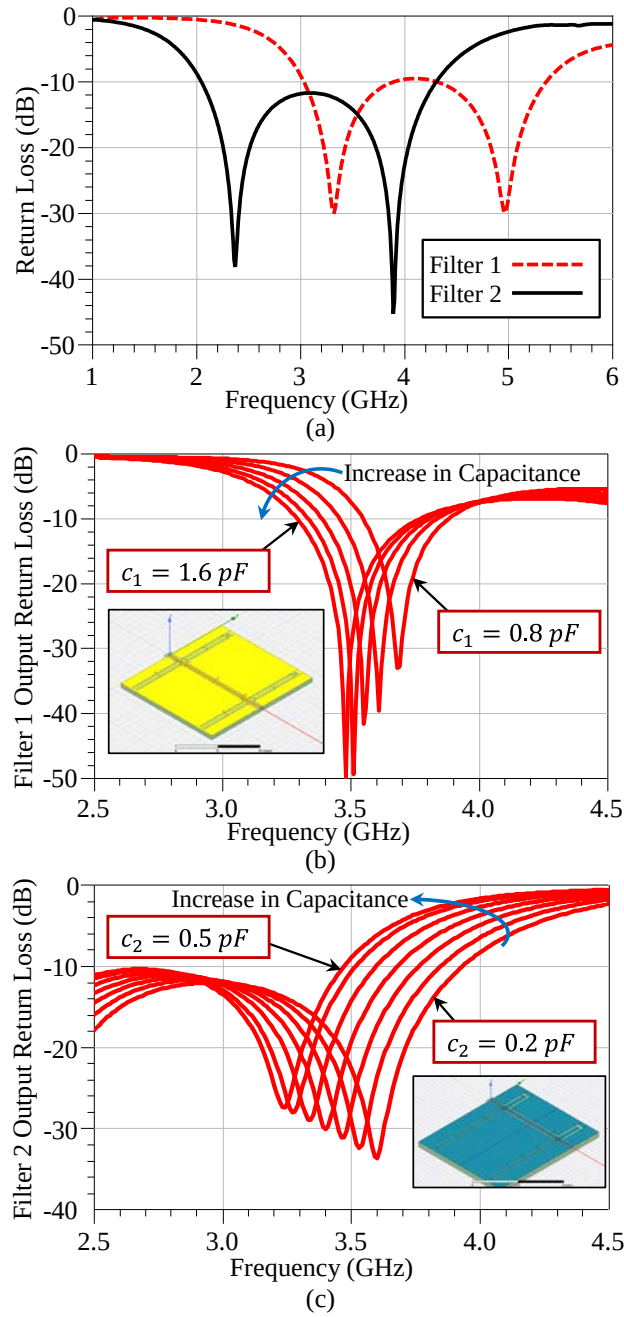


Figure 5-47 Simulated return loss of the filters, (a) filter responses without tuning, (b) Filter-1 loaded by c_1 with values ranging from 0.8 pF to 1.6 pF, (c) Filter-2 loaded by c_2 with values ranging from 0.2 pF to 0.5 pF.

Reflection type phase shifters (RTPSs) are employed because of their compact size compared to direct phase shifters (e.g. loaded line). Moreover, they use fewer tuning elements. A tunable RTPS is composed of two tunable reflective loads connected to the coupled and through ports of a quadrature hybrid coupler. The phase shifters are intended to provide at least 180° of continuous phase shift to address (4) and (5). A schematic of the proposed RTPS and its dimensions are depicted in Figure 5-48. As stated in [77], [91], and [74], the range of phase shift achieved using single tuning element reflective loads for RTPSs is less than 180° . Accordingly, a 2-element reflective load is used. It is designed using GCPW lines to avoid further wirebonds. Each reflective load is composed of a three TL segment main branch and two extension stubs, which are terminated in the variable capacitors. The extensions stubs are connected to the main branch using T-junctions. Tandem couplers are used because they require less fabrication tolerance compared to traditional designs, which allows PCB fabrication of the proposed circuit. In addition, tandem couplers have wider bandwidths compared to branch-line couplers [90].

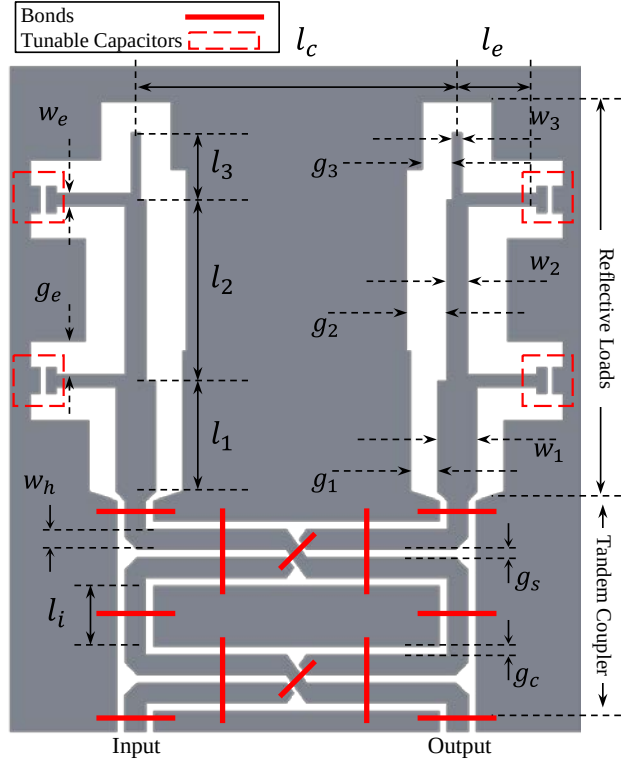


Figure 5-48 Schematic of the proposed RTPS. Dimensions in μm : $l_c = 7977, l_e = 3973, l_1 = 7747, l_2 = 7500, l_3 = 1650, l_i = 1500, g_1 = 670, g_2 = 1000, g_3 = 755, g_c = 206, g_s = 200, g_e = 823, w_1 = 980, w_2 = 523, w_3 = 230, w_h = 500, w_e = 311$.

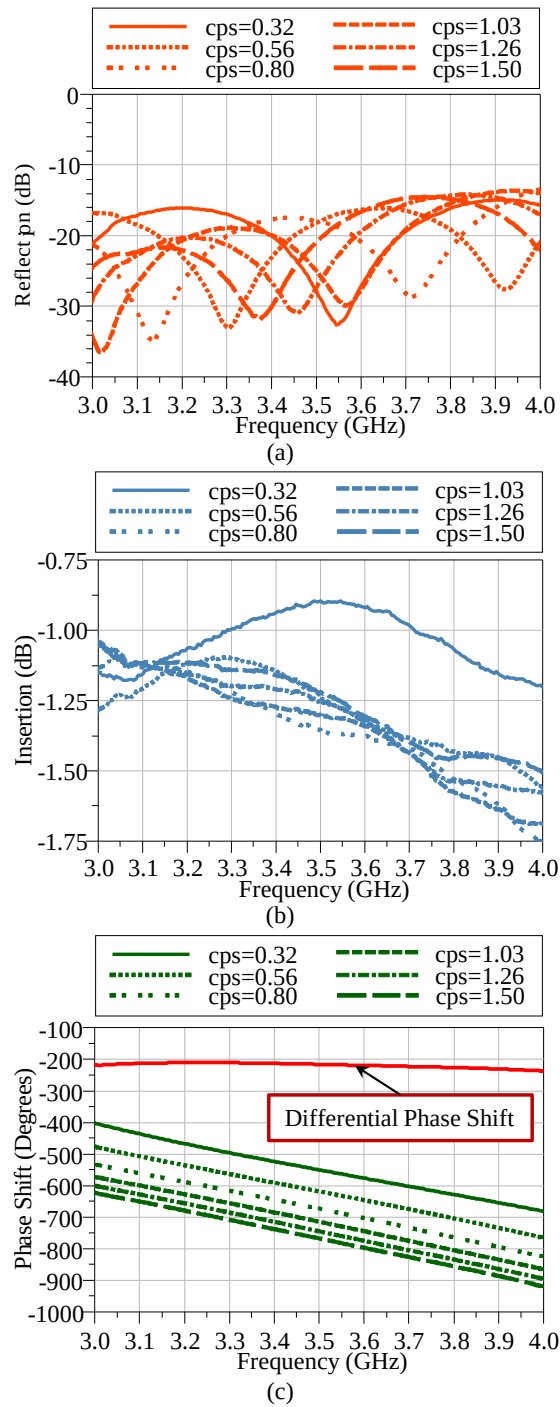


Figure 5-49 Simulated results for the RTPS sub-component when loaded by cps with values ranging from 0.32 pF to 1.5 pF (a) return loss, (b) insertion loss, and (c) phase shift.

The BST used for the proposed RTPS have a maximum capacitance of 1.5 pF with an average quality factor of about 30. The circuit shown in Figure 5-48 is optimized in HFSS. The extracted S-parameters are then terminated in measured BST values. The simulation results are shown in Figure 5-49. As can be seen, a phase shift of at least 200° is achieved by modifying the capacitor value from 0.32 pF to 1.5 pF. The insertion and return losses are better than 1.75 dB and 14 dB throughout the frequency range of 3 GHz - 4 GHz. Figure 5-50 investigates the impact of the capacitor imperfection on the insertion loss of the proposed RTPS. As can be seen, using a tuning element with a higher Q values can considerably reduce the insertion loss of the phase shifter.

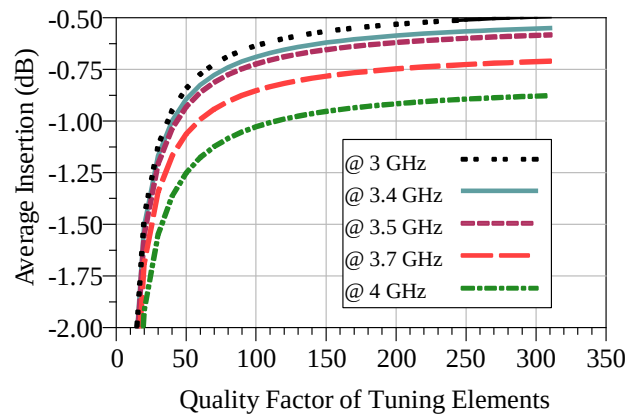


Figure 5-50 Simulated average insertion loss for the RTPS sub-component based on the quality factor of BST tunable capacitor.

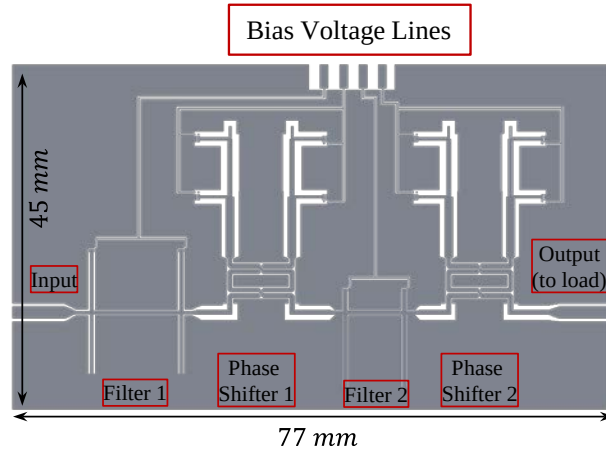


Figure 5-51 Scaled layout of the proposed Db-RIMN, where sub-components are highlighted. SBITE-1 is composed of Filter-1 and Phase Shifter 1, and SBITE-2 is composed of Filter-2 and Phase Shifter 2.

5.3.3.3 Dual-band Impedance Coverage

The final layout of the proposed Db-RIMN is presented in Figure 5-51. One way to demonstrate the dual-band impedance coverage is by demonstrating the independent coverage at each band. First, the circuit is simulated using a broadband 50Ω load and the tuning capacitors are tuned to achieve matching at both frequencies. Then the coverage at each band is demonstrated by tuning the SBITE-1 and SBITE-2 separately. **Error! Reference source not found.**(a) shows the coverage achieved by tuning only the SBITE-2. As can be seen, the output reflection coefficients remain inside the 10 dB circle at 3.4 GHz, while coverage is demonstrated for the area within the 5 dB circle at 3.7 GHz. In addition, **Error! Reference source not found.** (b) shows the coverage achieved by tuning only the SBITE-1. Consistent coverage for the impedances within the 5 dB circle at 3.4 GHz is shown in this figure as well. **Error! Reference source not found.**(a) and **Error! Reference source not found.**(b) together show that the two bands can be matched independently.

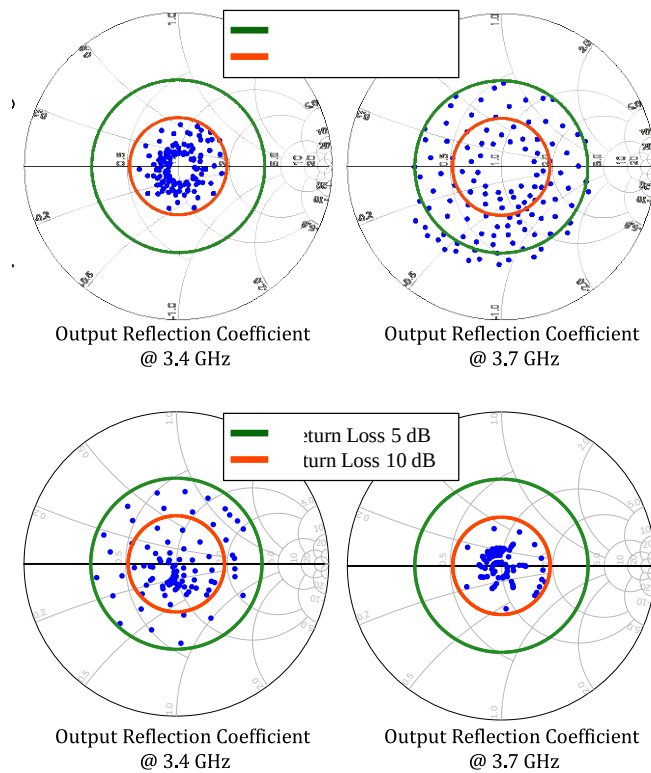


Figure 5-52 Simulated load impedance coverage (a) at the higher band when lower band for $c_1 = 1.5$ pF, $cp_1 = 0.9$ pF, c_2 sweep from 0.22 pF to 0.55 pF in 0.05 pF steps, and cp_2 sweeps from 0.45 pF to 2.2 pF in 20 steps, (b) at the lower band when lower band for $c_2 = 0.2$ pF, $cp_2 = 0.6$ pF, c_1 sweep from 0.5 pF to 1.6 pF in 0.2 pF steps, and cp_1 sweeps from 0.45 pF to 2.2 pF in 13 steps.

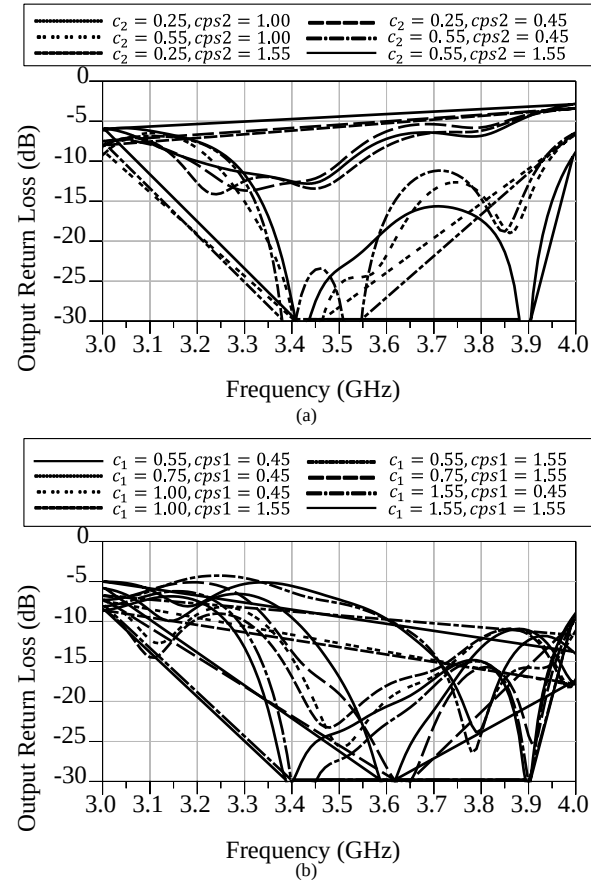


Figure 5-53 The simulated S-parameters for achieving coverage over (a) higher band and (b) lower band.

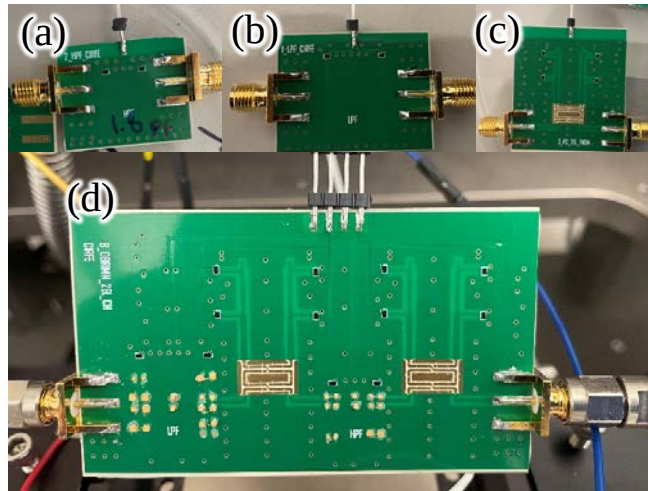
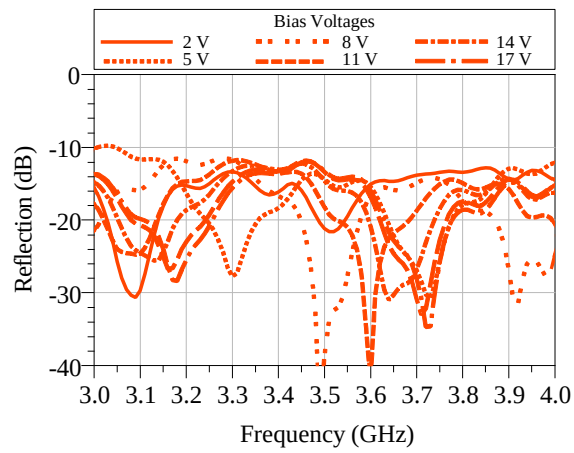
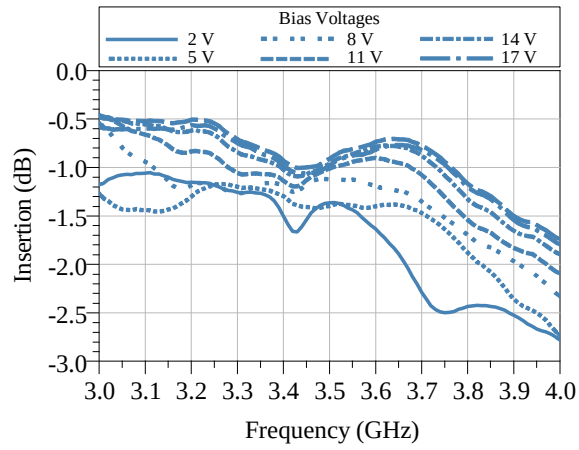


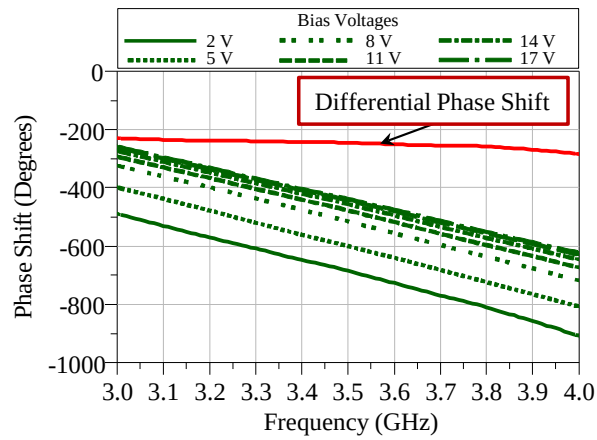
Figure 5-54 Fabricated circuits (a) Filter-1, (b) Filter-2, (c) RTPS, and (d) the proposed Db-RIMN.



(a)



(b)



(c)

Figure 5-55 Measured results for the RTPS sub-component when biased by 2V to 17V (a) return loss, (b) insertion loss, and (c) phase shift.

Figure 5-53 plots the return loss over frequency for a selection of states. For Figure 5-53(a), the SBITE-1 is in passing state and SBITE-2 is matching. As can be seen, the output return loss remains better than 12 dB at 3.4 GHz. The results at 3.7 GHz, illustrates that loads with reflection coefficient between 5 dB and 15 dB can be matched. Note that the return loss remains better than 10 dB from 3.27 GHz to 3.5 GHz. Therefore, this band can be used as the lower band of the proposed Db-RIMN. Similarly, Figure 5-53(b) is plotted by keeping the SBITE-2 in a passing state and tuning the SBITE-1. Here, the output return loss remains better than 15 dB at 3.7 GHz. The results shown at 3.4 GHz illustrates that loads with reflection coefficient better than 5 dB can be matched. Likewise, the return loss remains better than 10 dB between 3.62 GHz and 3.96 GHz. However, the quality factor of the BSTs degrade dramatically beyond 3.8 GHz. Therefore, a higher band of between 3.6 GHz to 3.8 GHz can be used by the proposed Db-RIMN.

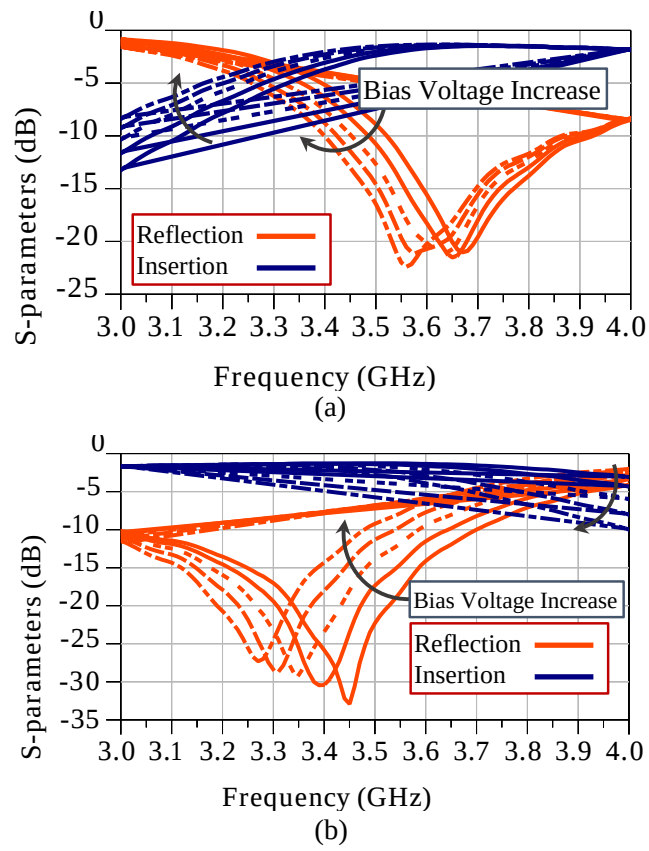


Figure 5-56 Measurement results for (a) Filter-1 and (b) for Filter-2.

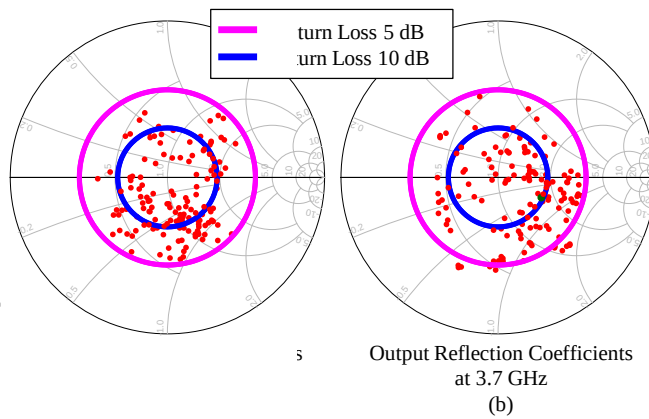


Figure 5-57 Measurement coverage achieved (a) at 3.4 GHz and (b) at 3.7 GHz.

5.3.4 Measurement Results

Prototypes of the proposed Db-RIMN are fabricated on RO3010 substrates. The Gold Immersion surface finish is used to allow wirebonding as well as attachment of the BSTs. Figure 5-54 shows the photos of the fabricated and assembled circuits. The diced filters and phase shifters are measured separately for verification and loss measurements. The measurements are performed using Keysight PNA Network Analyzer model N5227B.

A bias voltage of between 2V and 17V are applied to the phase shifter and the measurement results are plotted in Figure 5-55. The measurement results agree well with the simulations. The variation in insertion loss is due to the variation in quality factor of the BSTs. Note that higher bias voltages lead to smaller capacitor values and higher quality factor than the assumed 30. Moreover, the 2V bias is associated to larger than 1.5 pF. This leads to a significantly small quality factor and an insertion loss of more than 1 dB. Since the achieved phase shift is more than 220° , the lossy states are not needed. The measurement results for the filters are shown in Figure 5-56. Here the bias for Filter 1 varies between 4V and 10V. For Filter 2, it varies between 10V and 17 V. as can be see, each filter degrades the insertion loss by at least 1 dB over their passbands, which is mainly due to the addition of the wirebonds and limited quality factor of the BSTs.

For measuring the performance of the entire Db-RIMN, the two ports are calibrated to 50Ω broadband load. Then the measured results are terminated with variable load impedances in ADS environment instead of terminating the circuit with physical load impedances. This method has the advantages of being able to evaluate the insertion loss of the circuit by modifying the impedance of the output port. 4 independent bias voltages are applied. The bias voltage range for the filters is similar to the individual filter measurements. For the phase shifters the voltage is between 4V and 17 V to avoid excessive losses. A total of 110 random measurement points are collected and the coverage is depicted in Figure 5-57.

The matched results in terms of output reflection coefficient and the insertion loss, for a few example loads are depicted in Figure 5-58. As can be seen, different bandwidths and insertion losses are recorded at each band. The poor insertion loss performance is consistent with the sub-component measurements and is mainly due to the limited quality factor of the tuning elements and post processing of the circuit.

Finally, the relationship between reflection coefficient of the load, the matched bandwidth and insertion loss are investigated in Figure 5-59. To derive these trends, a handful of load impedances with similar magnitude of the reflection coefficient (i.e. with different phase). Then their bandwidths and insertion losses

are averaged separately. The results are shown for each band. The insertion loss degrades roughly by 2 and 3 dB at 3.4 GHz and 3.7 GHz, respectively. Moreover, the bandwidth is reduced by a factor of 2 and 3 at 3.4 GHz and 3.7 GHz, respectively. Accordingly, loads with worse reflection coefficients tend to have narrower bandwidth and worse insertion loss.

5.3.5 Conclusion

A novel design theory and realization technique for designing Dual-band Reconfigurable Impedance Matching Networks (Db-RIMNs) was presented. The sub-component choices were addressed. The impacts of imperfections of the tuning elements and circuit choices were also investigated. The performance of the proposed circuit can be significantly improved at higher frequencies, where integrated solutions provide high quality factor tuning elements, and at lower frequencies, where packaged solutions offer better quality factors. To the knowledge of the authors, this work is the only theoretical and experimentally proven method for designing Db-RIMNs. A wide array of applications is anticipated for the proposed Db-RIMN, which may include concurrent PAs, multiband transceivers, millimeter wave transceiver, and high-power transmitters.

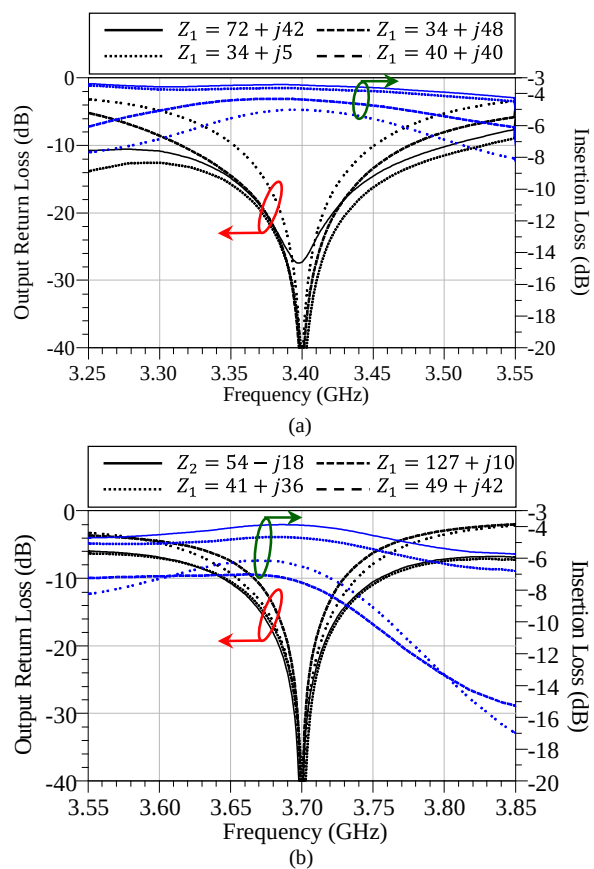


Figure 5-58 Measured S-parameters for a selection of load impedances (a) over lower band and (b) over the higher band.

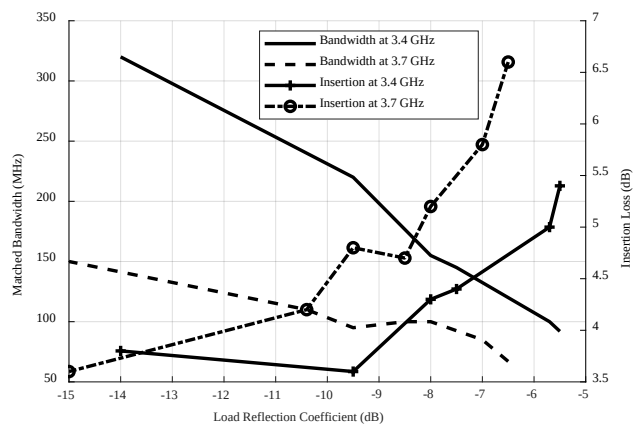


Figure 5-59 Average bandwidth and insertion loss based on the magnitude of the reflection coefficient of the load.

Chapter 6

Conclusion and Future Work

6.1 Contributions

This research has been focused on developing novel impedance tuners and matching networks for the next generation of wireless networks. More specifically, this includes doubly terminated impedance tuners for self-interference cancelation in propagation domain, and fixed and tunable dual-band matching networks.

Chapter 3 was focused on developing high precession and low loss reconfigurable doubly terminated matching networks. First, a method for incorporating packaged RF MEMS switches in DT-RIMNs was presented. This method was used to achieve better than 30 dB of matching between generic variable source and load impedances described by 18dB and 15 dB of RL, respectively. The use of switched stubs allowed for realizing a π -network where individual section could be switched between inductive and capacitive values. The matched bandwidth for the tested prototype varied between 20 MHz and 170 MHz and the IL varied by less than 0.5 dB between the different states. Additionally, a loaded line RIMN using BST varactors was designed, where the junctions and connecting TL segments were studied in much detail. The measurement results indicated 40 dB of matching over the 3.4 GHz to 3.7 GHz frequency range. The matched bandwidth for each state was at least 20 MHz and the IL was better than 0.6 dB. Furthermore, it was analytically shown that for the majority of existing circulators, the isolation between Tx and Rx channels can be significantly improved by using a DT-RIMN between the antenna and the circulator, with negligible penalty on the matching and insertion loss. Finally, a switched loaded line was designed, which employed 6 SP3T PCM switches to achieve the coverage needed for increasing the isolation between Tx and Rx channels. The resulting circuit was extremely miniaturized and offered very fine coverage over the designed band. The measurement results indicated an at least 40 dB of isolation between the two channels, while the transmitter matching of about 18 dB and antenna matching of 12 dB were maintained. A further comparison between the case uses of the DB-RIMNs demonstrated that the cost of every 10 dB improvement in isolation between the Tx and Rx channels was less than 0.1 dB of insertion loss.

Chapter 4 was focused on the theory of dual-band impedance matching networks. It also featured various realizations of this method for fixed dual-band loads. The realization efforts were narrowed to only cascade combinations of filters and phase shifters. Accordingly, two types of DBCMNs were developed. The first one, TYPE-I included HPF and LPF filters and TL segments as phase shifters. The implications of order of these subcomponents for the matched bandwidth and size of the circuit was studied and the matched

bandwidths of 6.5% and 11% was reported for the fabricated prototype. The second type used a BPF and a dual-band RTPS. Different designs using CPW lines and microstrip lines, tandem hybrid couplers, and modified dual-band branch-line coupler were studied. The achieved bandwidths for the three fabricated prototypes were 9%-3%, 9%-5%, and 4%-3% for the lower and higher bands, respectively. The two types of DBCMNs were compared based on the matched bandwidth, IL, physical size, and fabrication/assembly considerations. Accordingly, guidelines for selecting proper type depending on the application were presented. These matching networks are expected to find application in designing dual-band PAs, and high power transmission systems.

Chapter 5 introduced the concept of tunable dual-band RIMNs based on the analysis presented in the previous chapter. The requirements for the tunable filters and tunable phase shifters were investigated. Accordingly, three low-loss tunable phase shifters were designed. The first one had a continuously tunable 90° π -network. It was accompanied by two digital single bit 90° and 180° phase shifters to achieve 360° of phase shift. The second phase shifter was a single tuning element RTPS for achieving 180° of phase shift. The last one included a novel dual-band tunable reflective load. Moreover, different circuit topologies for these matching networks were presented based on the frequency ratio. Finally, prototypes were designed and implemented using BPFs. To our knowledge, this was the first time DB-RIMN were introduced. A wide range of applications were anticipated for these components, which included handheld devices, concurrent PA design, multi-band antenna design, beam steering, to name a few.

6.2 Future Work

The investigations presented in this thesis evoke many open questions worthy of investigation in the future works, a few of which are briefly described in the following.

1. Improving the reliability of the PCM based DT-RIMN by using similar bias lines for all the switches
2. Further miniaturization of the DT-RIMNs by incorporating the PCM switches on multilayer structures such as Low Temperature Co-fired Ceramics (LTCCs)
3. Leveraging the flexibility of the DBCMNs by designing non-planar high-power matching networks for space communication or base station applications.

4. Enhancing the matched bandwidth of the dual-band matching networks by using alternative phase shifters
5. Investigating the implications of using delay lines in-lieu-of or in addition to the phase shifters to increase the matched bandwidth of the proposed dual-band matching networks
6. Further investigation and implementation of the manifold architecture for the dual-band matching networks
7. Fabrication and testing the proposed DB-RIMNs (This will be done before the defense date)
8. Implementation of the proposed dual-band phase shifter
9. Implementing the DB-RIMNs using PCM switches
10. Improving the matched bandwidth of the DB-RIMNs
11. Investigating the implications of using other filter types and possibly 3D structures in designing the DB-RIMNs
12. Implementing the filter based matching networks at higher frequencies using fully integrated circuits
13. Implementing the filter based matching networks at lower frequencies using lumped elements

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