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**Unilateral and Multi-functional RF Front-end
Filtering Networks Using Spatiotemporally
Modulated Resonators Arrays**

Thesis presented by

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Doctor of Philosophy

in Electrical and Electronic Engineering

University College Cork

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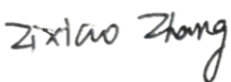
List of Acronyms

Acronym	Definition
6G	Sixth generation
AWG	Arbitrary wave generator
BER	Bit error rate
BPF	Bandpass filter
BPFC	Bandpass filtering circulator
BPFI	Bandpass filtering isolator
BW	Bandwidth
CMOS	Complementary metal oxide semiconductor
CRD	Coupling routing diagram
D	Directivity
DMC	Distributed modulated capacitors
DUT	Device under test
EVM	Error vector magnitude
FBW	Fractional bandwidth
FDTD	Finite-difference time-domain
IBFD	In-band full-duplex
IC	Integrated circuit
IL	Insertion loss
IoE	Internet-of-everything
JCAS	Joined communication and sensing
JJ	Josephson junction
MEMS	Micro-electro-mechanical system
MIMO	Multiple-input multiple-output
MMIC	Monolithic microwave integrated circuit
PCB	Printed circuit board
PD	Power divider
RBW	Reflectionless bandwidth
RF	Radio frequency
RFFE	RF front-end
RFQC	Reconfigurable filtering quadrature coupler
RL	Return loss

SIW	Substrate-integrated waveguide
SQUID	Superconducting quantum interference device
STM	Spatiotemporal modulation
STM-AM	Spatiotemporal modulation angular momentum
SWaP-C	Size, weight, power and cost
TDD	Time division duplex
TF	Transfer function
TZ	Transmission zero
UHF	Ultra high frequency
VA	Variable attenuator
VCO	Voltage-controlled oscillator
VNA	Vector network analyzer
VP	Variable phase shifter
YIG	Yttrium Iron Garnet

Declaration

This is to certify that the work I am submitting is my own and has not been submitted for another degree, either at University College Cork or elsewhere. All external references and sources are clearly acknowledged and identified within the contents. I have read and understood the regulations of University College Cork concerning plagiarism and intellectual property.

Signature of the candidate: 

Date: October 2nd, 2025

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Abstract

As wireless networks progress towards the sixth-generation (6G) of wireless, new communication technologies such as massive multiple-input-multiple-output (MIMO), in-band full-duplex (IBFD), and joint communication and sensing (JCAS) are emerging. These technologies impose stringent demands on radio-frequency front-ends (RFFE) requiring compact, reconfigurable, and multi-functional RF components that can reduce size, weight, power, and cost (SWaP-C) of the RFFE while maintaining high spectral efficiency and adaptability to a wide range of frequency bands, operating conditions, and application scenarios. To address these challenges, this thesis investigates spatiotemporally modulated (STM) resonators as the basis for a new class of RF filtering components and signal-routing networks. The co-designing of RF filters with other front-end functionalities can effectively reduce size and cost. At the same time, the use of STM resonators enables unilateral transmission while remaining low cost, compatible with IC integration and power efficient. First, a transversal STM-resonator architecture is introduced for the realization of bandpass filtering isolators (BPFIs) with high levels of reconfigurability in terms of center frequency, bandwidth (BW), number of passbands, out-of-band suppression and intrinsic switch-off capability. Second, this thesis demonstrates multi-functional STM based devices that integrate variable phase shifting (VP) and variable attenuation (VA) capabilities within the same BPF structure. They enable unilateral RF signal transmission with continuously phase and attenuation tuning without compromising the transfer function (TF) shape. Finally, the thesis extends the concept to more complex RF front-end architectures by introducing multi-port STM signal-routing networks. Through modulation-state control of the STM resonators, the proposed multi-port RF networks demonstrate versatile RF signal processing functions including RF switching, RF isolation and circulation, and power division with integrated RF filtering functionalities. Overall, this thesis advances the state-of-the-art by introducing three novel STM-resonator-based architectures that, respectively, demonstrate enhanced tunability in unilateral bandpass filtering, the co-integration of filtering with phase and amplitude control, and multi-functional multi-port RF signal routing enabled through modulation state control.

Chapter 1: Introduction

This chapter provides an overview of the motivation, background, and context for the research carried out in this thesis. It begins with an overview of the needs of compact, energy-efficient, cost-effective and multi-functional radio-frequency front-ends (RFFE) components for next-generation wireless systems and highlights the importance of designing multi-functional non-reciprocal filtering devices. Different approaches for achieving non-reciprocal transmission are reviewed and compared, positioning spatiotemporal modulation (STM) as a promising enabling technique. The chapter then presents a detailed literature review of the state-of-the-art multi-functional filtering devices, covering recent advances and limitations in three key areas: reconfigurable bandpass filtering isolators (BPFIs), multifunctional filters co-integrated with variable phase shifters (VPs) and variable attenuators (VAs), and multi-port filtering signal-routing networks. The chapter is concluded by outlining the research objectives and the organization of the remainder of the thesis.

1.1 Background and Motivation

The next generation of wireless networks (beyond 5G or 6G) will bring forward key RF technology-driven innovations including massive multiple-input–multiple-output (MIMO) antenna systems, in-band full-duplex (IBFD) communications, joint communication and sensing (JCAS) and quantum computing. Furthermore, they promise unprecedented levels of connectivity through the emergence of the Internet-of-Everything (IoE). To facilitate the deployment of 6G networks, highly-modular and highly-functional RF transceivers with ultra-small size, weight, power and cost (SWaP-C) will need to be developed. Furthermore, these transceivers must cover a wide range of frequency bands, from a few MHz up to 100s of GHz, while achieving ultra-high data rate transmission with low latency. However, these requirements are expected to impose stringent RF performance and integration specifications for the realization of their RFFEs, which must integrate multiple and diverse receiver and transmitter chains within the same RF module.

Modern RFFE design approaches go beyond the minimization of SWaP-C at the component level. They are focusing on holistic RF chain implementations approaches through the co-design and co-integration of multiple RF functions within a single multi-functional RF component. Among the various RFFE components, filters play a

key role in spectrum management and interference mitigation [1]–[4]. Therefore, co-design approaches that integrate filtering functionalities with other active or passive RFFE signal processing actions are increasingly being explored as an effective strategy to achieve miniaturization, performance improvement and multi-functionality. Such techniques are also being explored for the realization of non-reciprocal (also referred to as unilateral or directional) RF components, such as RF isolators and circulators, due to their pivotal role in a wide range of applications. These include: (i) in high-power microwave transmitters to protect power amplifiers from back-reflections from the antenna [5], (ii) in IBFD systems to enable simultaneous transmission and reception at the same frequency thus doubling the spectral efficiency [6]–[8], (iii) in JCAS [9]–[11], and (iv) in cryogenic quantum computing as part of the readout circuits for qubits [12]–[14].

Bringing together RF filtering and directionality, i.e., the function of an RF isolator or circulator within a single RF component by introducing non-reciprocity has led to the concept of non-reciprocal (also called unilateral or directional) RF filters. These filters provide simultaneously spectral selectivity and directional transmission and present a promising path towards more compact, multi-functional and flexible RF systems due to the combination of multiple RF functions in one. Beyond non-reciprocal filters, there is an increasing need for highly multi-functional RF components that also incorporate additional capabilities like VPs or VAs. These components are highly desirable in multi-antenna systems, such as phased arrays, where they facilitate high beam scanning resolution and low side lobe levels. They are also essential for adaptive wireless links, which require rapid adjustment of phase and amplitude to maintain reliable performance under varying channel or interference conditions. As a further requirement, next-generation wireless platforms will need scalable multi-port RF signal-routing networks that extend the functionality of individual devices into more complex small systems capable of dynamically connecting multiple antennas or signal paths within compact transceivers. The realization of such networks will not only facilitate signal routing between different ports but will also provide additional functionalities such as RF filtering, isolation, circulation or power division within the same RF component.

This thesis explores the development of multi-functional unilateral RFFE filtering components using spatiotemporally modulated (STM) resonator arrays.

Specifically, three distinct architectures are explored: (i) co-designed BPFIs based on STM transversal cells, featuring enhanced reconfigurability in terms of center frequency, bandwidth (BW), number of passbands, out-of-band rejection levels, and intrinsic RF switching-off capabilities; (ii) multi-functional BPFI architectures that integrate VP and VA functionalities, with scalable and flexible transfer function (TF) capabilities including reconfigurable selectivity, single- or dual-band responses and quasi-elliptic characteristics; and (iii) highly-flexible, scalable multi-port STM signal-routing networks with integrated multi-port switching, bandpass filtering, power division and unilateral transmission functionalities, realized through modulation-state control. Overall, this thesis advances the state of the art in STM-resonator-based BPF designs by enhancing tunability and by adding multifunctionality through novel coupling schemes, and further introduces modulation-state control of STM resonators as a new design technique for realizing complex multi-port signal routing.

In what follows, a comprehensive review of recent advances and emerging design strategies for multi-functional filtering devices is provided, with a special focus on STM-based unilateral/non-reciprocal RF filters. First, different integration approaches for incorporating non-reciprocity within a filtering network are presented with a specific emphasis on integration schemes using STM. Afterwards the discussion extends to state-of-the-art STM-based unilateral filter design approaches, multi-functional filters with co-integrated VP and VA functionalities, and multi-port multi-functional filtering routing networks.

1.2 Design and Integration Techniques for Unilateral/ Non-reciprocal RF Components

Non-reciprocal RF components are typically implemented using magnetically-biased ferrite materials such as Yttrium Iron Garnet (YIG), or composite ferrites. Y-junction circulators are among the most popular ones using these materials, comprising a magnetically-biased ferrite disk at the center of three symmetrically-spaced RF ports. The unilateral RF signal circulation results from the interference of counterrotating waves with different phase velocities under external magnetic biasing [15]–[17]. Commercially available Y-junction circulators exhibit low insertion loss ($IL < 1$ dB) and have up to 20 dB of isolation with fractional bandwidths (FBWs) between 20% to 40%. They are available over a wide frequency spectrum, spanning

from hundreds of MHz up to approximately 120 GHz and they have high power handling capacity up to several hundred watts [18]. However, they rely on permanent magnets placed above, below, or at the center of the ferrite disk, which increases their size and volume. The development of self-biased magnetic materials such as hexagonal ferrites or magnetic nanowire arrays has the potential to eliminate the need for magnets and enable integrated-circuit (IC)-based integration [19]–[21]. However, the integration of self-biased materials with standard CMOS and MMIC integration platforms remains a great challenge due to the difficulty of depositing thick, high-quality ferrite films on large-area semiconductor wafers without damaging the underlying circuitry due to the high processing temperatures, material incompatibility, and limitations in patterning techniques [22].

Early non-magnetic circulator/isolator implementation approaches exploited the inherent unilateral transmission properties of actively-biased transistors. Initial implementations were based on the connection of transistors in a loop for the realization of three-way circulators [23]. Subsequent advances in IC technologies enabled various architectures including symmetric three-port circulators [24]–[26] and quasi-circulators [27]–[31]. However, transistor-based solutions are inherently power-hungry (up to 900 mW in [25]), they have poor noise performance (up to 12.6 dB in [30]) and suffer from nonlinear distortion ($IP1dB = -3$ dBm in [30]) [32], [33]. In recent years, STM or parametric unidirectional circuits have emerged as an attractive alternative to facilitate unilateral transmission. STM achieves unilateral transmission by breaking the time-reversal symmetry through temporal modulation of various material parameters such as permittivity, permeability, or conductivity, using synchronized modulation signals [34], [35]. These approaches are summarized as follows:

Permittivity modulation: In this approach, the permittivity of the substrate is periodically varied along the signal propagation path. It was first introduced in 1950 in parametric amplifiers, where the semiconductor diode capacitance was modulated using pump signals [36], [37] within up-conversion and negative-resistance/reflection configurations. Initial unilateral devices employed phase-shifted pump signals along quarter-wave transmission lines forming the foundations of a concept that inspired contemporary microwave and photonic structures [38]. Specifically, Wang [39] and Qin [40] demonstrated distributed modulated capacitors (DMC) for traveling-wave

modulation and their application in RF circulator. As illustrated in Figure 1.1, varactors are loaded periodically along transmission lines with the transceiver located at one end of the line and the antenna at the other end. These varactors are modulated with a traveling-wave carrier (pump) signal. In the presence of the carrier, signals traveling in the same direction as the carrier (from antenna to receiver) are converted to the upper and lower sidebands, while signals propagating in the opposite direction (from transmitter to antenna) remain at their original frequency. However, this approach is inherently limited by the low modulation index of the varactors, which requires wavelength-scale transmission line lengths to achieve sufficient modulation. Additionally, high-power and high-frequency pump signals restrict their scalability at higher frequencies, and a duplexer is required for separating the transmitter and the receiver signals, increasing their size and degrading their bandwidth.

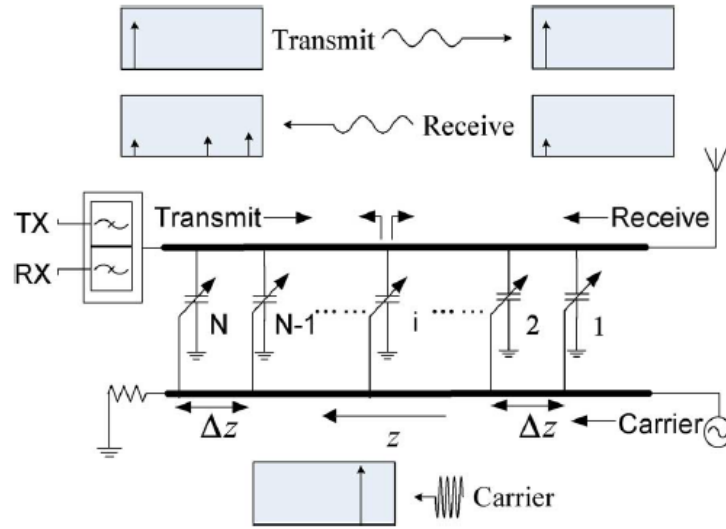


Figure 1.1 Distributed modulated capacitors concept utilized for the realization of an RF circulator in [40] © 2014 IEEE.

The aforementioned limitations in DMC approaches motivated the invention of the spatiotemporal modulation angular momentum (STM-AM) biasing technique that was first proposed by Sounas et al. [41]. It is based on phase-shifted permittivity modulation in resonant ring structures, creating distinct propagation paths by breaking degeneracy between clockwise and counterclockwise modes. This approach was later extended to RF applications by Estep et al. in [42]. Specifically, a circulator was demonstrated by connecting three ladder LC resonators in a delta topology and modulating their resonating frequencies with 120° phase-shifted signals. An

experimental prototype with a subwavelength footprint of $\lambda/75$ was demonstrated, exhibiting over 40 dB of isolation and a center frequency tuning range from 150 MHz to 210 MHz. Another topology was investigated in [43] by connecting series LC resonators in a wye architecture and in a delta configuration in [44] to improve matching and IL. As shown in Figure 1.2, the measured IL, return loss (RL) and reverse isolation of the prototype in [44] at 1000 MHz are 3.3 dB, -10.8 dB, 55 dB, respectively. Subsequently, several improvements and STM-based circuit topologies have emerged, including differential architectures [45], broadband implementation schemes [46], N-way configurations [47], and miniaturized CMOS implementations [48].

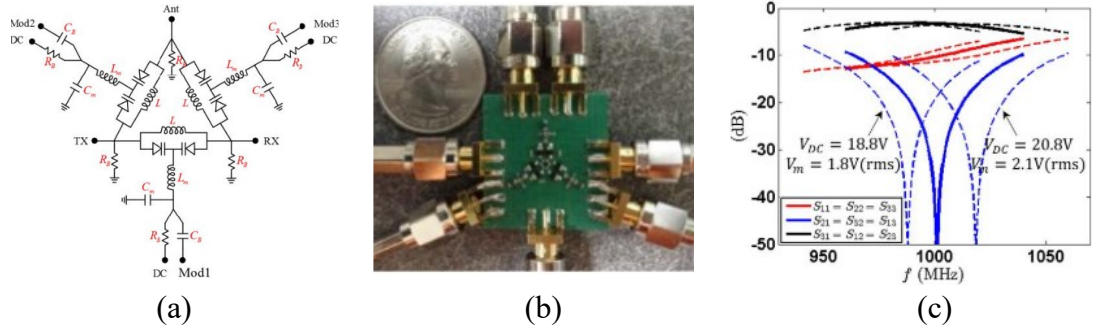


Figure 1.2. STM-AM concept in [44] © 2018 IEEE in which parallel STM LC tanks are connected in a delta topology: (a) Circuit schematic of the circulator; (b) PCB prototype; (c) Measured S-parameters.

In [49], Wu et al. demonstrated the integration of BPFs and RF isolators (BPFIs; also named non-reciprocal/unilateral filters) within a single RF component by applying STM to an inline coupled-resonator BPF, as shown in Figure 1.3 (a). Spectral domain analysis based on ABCD matrix approaches were proposed, providing a framework for optimizing the modulation parameters, such as frequency, amplitude, and phase shift. Furthermore, the method was validated via a three-pole unilateral BPF, demonstrating third order transfer function with low IL (1.5 dB) and high reverse isolation (20 dB) as shown in Figure 1.3 (c). To gain deeper physical insight, in [50] a more advanced coupling matrix representation was developed to model the time-varying filter as a network of coupled “harmonic resonators”. This method allowed the filter to be analyzed as an asynchronously tuned network, making it possible to apply the well-established coupling matrix theory for its analysis and design. This

approach was further developed to create bandpass filtering circulators (BPFCs) by incorporating unmodulated RF resonators into STM-AM structures to enhance their passband and stopband performance [51]. Specifically, the RF filtering circulator in [51] achieved a BPF-type TF with minimum in-band IL of 1.75 dB, in-band isolation better than 20 dB, and in-band RL better than 20 dB.

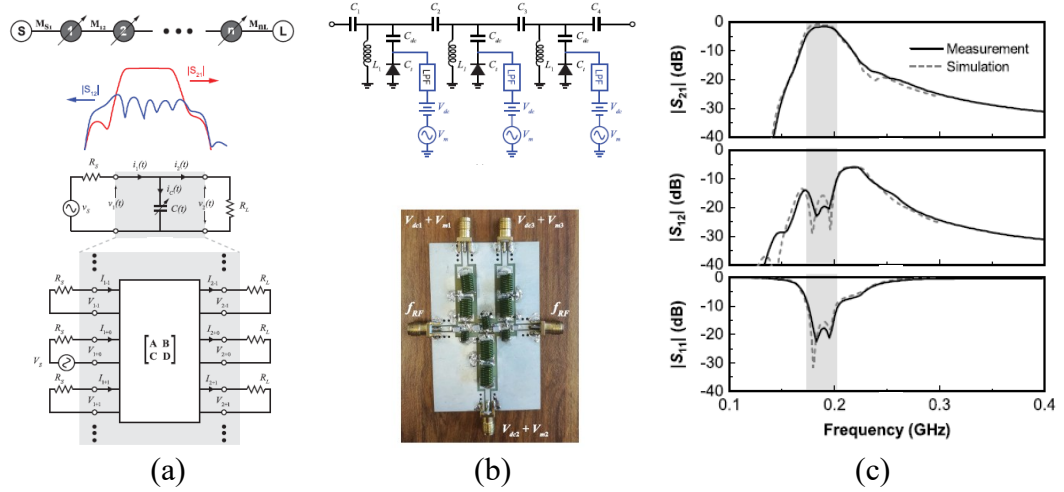


Figure 1.3. (a) Coupling routing diagram (CRD) and the conceptual S-parameters of the proposed in-line isolating filter in [49] © 2019 IEEE. The black circles with arrows are STM resonators with a successive phase shift. Below is the equivalent circuit of one STM resonator and its general spectral equivalent circuit with ports for all the intermodulation frequencies. (b) Circuit topology, photo of the prototype and the RF-measured S-parameters of the three-pole isolating filter in [49]. The STM resonators are implemented using a parallel connection of inductors and sinusoidally modulated varactors. (c) The RF-measured S-parameters of the three-pole isolating filter in [49].

Permeability modulation: Analogous to the capacitance modulation approach, permeability modulation is based on the modulation of inductance. While challenging to implement in integrated circuits, permeability modulation has been proven to be valuable in the quantum-computing community with the use of Josephson junctions (JJs) and superconducting quantum interference devices (SQUIDs) which serve as tunable inductive elements in cryogenic superconducting platforms. In 2017, Sliwa et al. demonstrated a non-magnetic, on-chip circulator using a field-programmable Josephson amplifier to parametrically couple three resonant modes. Signal circulating with over 20 dB of isolation was achieved by applying pump signals with carefully controlled phase difference [52]. In [53], an isolator was implemented by applying STM to the inductance of SQUID-based coupled resonators as shown in Figure 1.4,

resulting in a low forward IL of 1.3 dB, high reverse isolation of 25 dB, and center frequency tunability between 6 and 6.75 GHz.

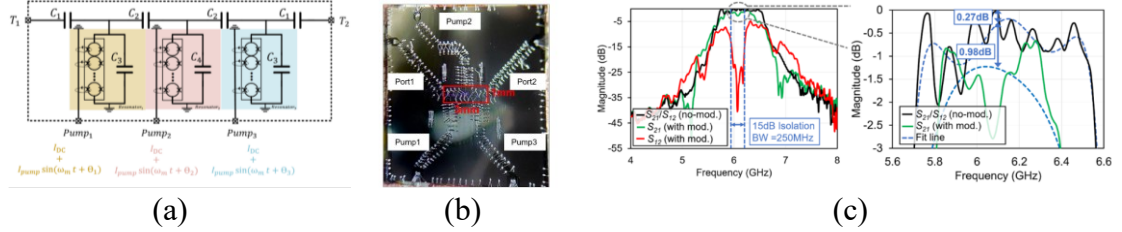


Figure 1.4. Permeability modulation concept for the realization of a circulator in [53] © 2023 IEEE. (a) Circuit schematic of the circulator. (b) Microphotograph of the chip. (c) RF-measured S-parameters.

Conductivity modulation: Conductivity modulation utilizing transistor-based switches benefits from substantially larger modulation ratios when compared to varactors or JJs. This concept originates from the N -path filter integration approach, where an input signal is down-converted, filtered by N identical linear time-invariant networks, and then upconverted back to the original frequency band using mixers driven by phase-shifted clocks. In [54]–[56], gyrators with non-reciprocal phase performances were demonstrated by staggering the clocks of the input and output switches in 2-port N -path filters. Embedding this gyrator approach in a loop of reciprocal transmission lines facilitates the realization of a circulator. The first implementation of such a fully-integrated CMOS circulator was presented in [56] as shown in Figure 1.5. It operated at 750 MHz, having an IL of 1.7 dB and a broadband isolation that is higher than 15 dB. Furthermore, its narrowband isolation can reach 50 dB with an external impedance tuner. However, the necessity for modulation frequencies equal to the operating frequency introduces substantial power consumption of over 59 mW and complexity. These limitations were surpassed by replacing the capacitors between the commutated switches with delay lines [57]–[60]. In this case a wideband response was achieved while having the increased bandwidth of the gyrator. Additionally, it decoupled the single frequency from the modulation frequency and decreased the modulation frequency by increasing the length of the transmission line of the gyrator. Circulators were also implemented in a millimetre-wave CMOS technology at 25 GHz with a modulation frequency of 8 GHz. Further research considered the use of quasi-transmission lines using lumped LC sections for

miniaturization [61] and switched transmission lines with BPFs to further increase the operation frequency to 60 GHz [62].

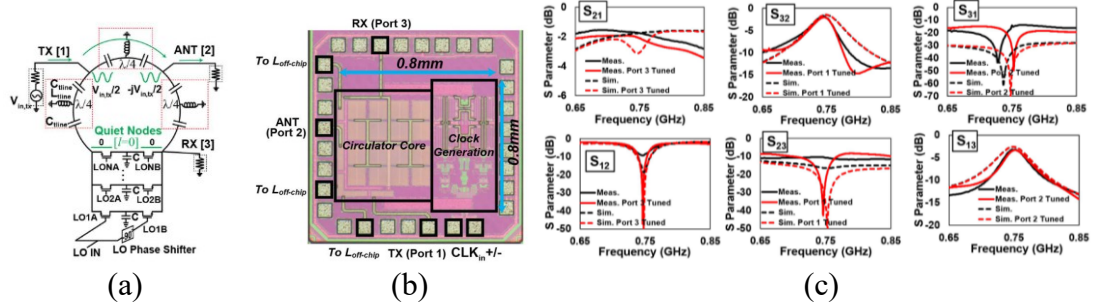


Figure 1.5. Conductivity modulation concept for the realization of a circulator in [56] © 2017 IEEE. (a) Circuit schematic of the circulator. (b) Microphotograph of the standalone 65-nm CMOS circulator. (c) RF-measured S-parameters.

Table 1.1 summarizes the key performance metrics of representative non-reciprocal devices based on ferrite, active, and STM-based approaches. As evidenced, ferrite-based circulators exhibit the lowest IL, the highest power handling, and the widest bandwidth. However, they are bulky, costly, and incompatible with standard IC technologies. On the other hand, active circulators are compact and compatible with IC integration, yet they are limited by low linearity, high noise figure, and have substantial DC power consumption. In contrast, STM-based circulators offer a good balance of performance metrics. They provide moderate IL, high reverse isolation, reasonable power handling, and are inherently compatible with IC technologies. Moreover, they exhibit significantly better noise performance and lower power consumption than active approaches. Among the three STM approaches, permittivity modulation stands out as particularly well-suited for multifunctional unilateral filter design. This is due to being compatible with coupled-resonator architectures commonly used in filters, without the need for advanced RF circuit elements such as JJs (superconducting devices working at cryogenic temperature required for permeability modulation) or complex switching networks (as in conductivity modulation). Additionally, their straightforward and simple modulation scheme does not disrupt the forward signal transmission, allowing seamless incorporation into existing co-designed filter architectures.

In summary, the STM approach offers the most promising route for realizing compact, reconfigurable, and multi-functional RFFE components. While recent

research has demonstrated the feasibility of STM-based multifunctional filters, existing devices remain constrained by limited scalability, restricted tunability, and a lack of integration across multiple functions and ports. These limitations motivate the following subsections, which review recent advances in STM-based unilateral filters, multi-functional filters, and multi-port signal-routing networks, and highlight how this thesis addresses these challenges.

Table 1.1. Comparison of non-reciprocal RF components, integration and design methodologies

Approaches	Y-junction ferrite		Active			STM						
						Permittivity modulation				Permeability modulation	Conductivity modulation	
Ref.	[23]	[24]	[30]	[35]	[36]	[49]	[51]	[53]	[56]	[58]	[59]	[67]
Device	C	C	C	C	C	C	C	C	FC	Unilateral BPF	C	C
Technology	Microstrip	GaN MMIC	PCB	90 nm CMOS	180 nm CMOS	PCB	PCB	180 nm CMOS	PCB	N/A	65 nm CMOS	45 nm SOI CMOS
f_{cen} (GHz)	2.7	5.7	1.8	40.5	24.38	1	1	0.91	0.825	6 ~ 6.75	0.75	60
IL(dB)	<0.5	2.7	2	3.5 ~ -4.3*	5.7	3.3	4.2	4.8	1.7	1.25	1.7	3.1 ~ 3.9
IS (dB)	>20	14	35	12 ~ 22	20	55	31	65	>20	25	40	40 (Tx-Rx)
BW(%)	40	N/A	11	132 (>13 dB IS)	2.8 (>15 dB IS)	3.4 [†]	14 [†]	2.4 [†]	8 [†]	4 (>15 dB IS)	4.3 [†]	14.5 (>25 dB Tx-Rx IS)
Size	14×14×5.5 mm ³	6×10 mm ²	N/A	0.97×0.96 mm ²	1.3×0.55 mm ²	143 mm ²	15×16 mm ²	0.2 mm ² (silicon area)	135×135 mm ²	3×1 mm ² (Core BPF)	5×5 mm ²	1.57×1.1 mm ² (Die area)
Max Power (Watts)	15	9.6	0.25	N/A	N/A	0.79	N/A	N/A	N/A	N/A	N/A	N/A
IP _{1dB} (dBm)	N/A	N/A	N/A	-3	9.5	29	23	N/A	N/A	-68	N/A	19.1
Noise Figure (dB)	N/A	N/A	5	7.1 ~ 12.6	10.1	4.5	N/A	5.2	N/A	N/A	4 (Rx)	3.2 (Rx)
IIP3	N/A	N/A	N/A	N/A	N/A	33	N/A	6	N/A	N/A	27.5 (Tx) 8.7 (Rx)	19
DC Power consumption	0	0	900 mW	41.8 mW	7.2 mW	0	0	64 mW	0	0	59 mW	41 mW

(IL: insertion loss at the center frequency, IS: isolation level in the reverse direction at the center frequency, C: circulator, FC: filtering circulator. *: negative values mean transmission gain. †: >20 dB isolation bandwidth.)

1.3 Multi-functional Filtering Devices Overview

1.3.1 STM-based unilateral filters

STM-based unilateral filters have attracted increasing attention as compact solutions that combine frequency selectivity with unilateral transmission. Following the first implementation of such devices in [49] based on STM in-line resonators, alternative integration schemes for STM-based BPFs have been presented using acoustic-wave resonators [63], [64], lumped-element configurations [65]–[68] and microstrip-based RF filtering components [50], [69]–[73]. However, the majority of the reported BPFs either work at low frequencies (e.g. up to 250 MHz in [67]), or have static TFs [49], [50], [69], [72], or have high IL (6.5 dB) [63] or low isolation (< 12 dB) [72] or their TF can only be tuned in terms of center frequency [65], [70], [71], [73], making them unable to meet the high levels of agility required in emerging communication systems.

Only a couple of tunable unilateral BPF topologies featuring multiple levels of RF tuning [66] have been demonstrated to date. In [66], a fully-reconfigurable unilateral BPF was introduced, offering tunable center frequency, tunable bandwidth, and switching-off capabilities. As shown in Figure 1.6(a), these functions are achieved by applying STM in a four-resonator array comprising a multi-configurable multi-resonant capable of generating one transmission pole and two transmission zeros. Center frequency tuning is accomplished by synchronously adjusting the resonant frequencies of all of its constituent resonators, while bandwidth tuning and switching-off functionalities are realized by asynchronously tuning the two resonators within the multi-resonant cell. Unilateral transmission is enabled by applying phase-shifted AC modulation signals to the resonators, as illustrated in Figure 1.6(a). A lumped-element PCB prototype centered at 300 MHz was manufactured and tested [see Figure 1.6(b)]. Its RF measured performance in Figure 1.6(c) and (d), demonstrated a 1.15:1 center frequency tuning range, a 2.77:1 bandwidth tuning range, and reverse isolation levels up to 30.9 dB.

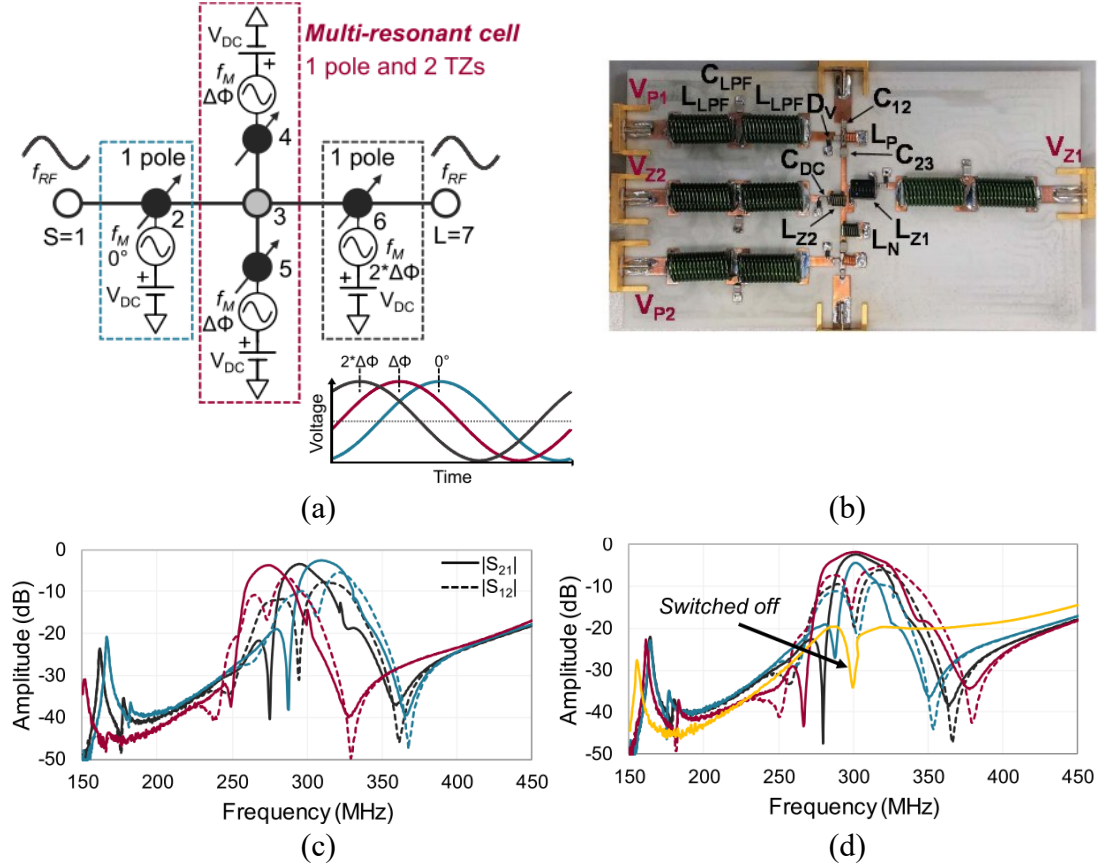


Figure 1.6. The fully reconfigurable unilateral BPF in [66] © 2020 IEEE. (a) The CRD. (b) Photograph of the prototype. (c) RF-measured center frequency tuning of the unilateral BPF. (d) RF-measured BW tuning and switching-off state of the unilateral BPF.

Further implementation schemes facilitated the realization of quasi-elliptic responses to improve selectivity [74], [75], or used cascaded multi-resonant cells [67], [68] or multimode resonators [76] to support multi-band operability. Additionally, balanced unilateral BPF configurations have been explored to suppress common-mode interference [77], [78], [79]. In 2022, Simpson et al. introduced a novel unilateral balanced BPF using STM to achieve a quasi-elliptic balanced response. The concept was validated through a microstrip prototype, demonstrating a remarkable reverse transmission isolation of 62.1 dB, and an IL of 3.2 dB in the forward direction. The measured common-mode suppression was measured over 45 dB across a bandwidth greater than one octave [79]. In summary, most state-of-the-art unilateral STM-based BPFs rely on inline resonator configurations, where tuning elements must be incorporated not only within the resonators but also in their impedance inverters to achieve fully tunable TFs. This requirement makes their realization complex, lossy, and challenging to optimize.

1.3.2 Multi-functional filtering variable phase shifters and variable attenuators

Beyond co-designing BPFs with isolators, recent research has also demonstrated multifunctional BPFs that incorporate VP and VA, particularly for phased array and MIMO applications. A number of co-designed bandpass filters and variable phase shifters (BPF-VPs) have been reported in [80]–[85]. However, they suffer from variable TF characteristics when large phase tuning ranges are implemented. Notable examples of this performance limitation can be seen in the BPF-VP in [82] that is based on passive vector summation method. The topology and the measured RF performances are presented in Figure 1.7. As shown in Figure 1.7 (a), the phase shifting function is co-designed with bandpass filtering using a pair of Wilkinson power dividers and a reconfigurable filtering quadrature coupler (RFQC). By adjusting the coupling between interstage resonators within the RFQC, the BPF-VP achieves a continuous phase-control range of 320° while maintaining a tunable second-order BPF response across 1.05–1.3 GHz. However, the asymmetry of the signal paths doesn't allow to preserve the TF filtering characteristics as can be observed from Figure 1.7 (c). The same TF degradation issue is also present in the BPF-VPs in [80]–[83]. In yet another approach, the BPF-VP in [85] demonstrated relatively good TF characteristics by employing tunable external coupling networks, however its phase tuning was limited to 90° .

A wide variety of co-designed bandpass filters and variable attenuators (BPF-VAs) have also been reported in [86]–[97]. They are based on on-chip implementations using transistors or graphene or utilizing the resistance change of PIN diodes under different biasing voltages. However, they are limited by a discrete number of tuning stages [86], [87], limited attenuation range (< 13 dB) [88]–[94], and have compromised RL or TF characteristics when their attenuation levels are being tuned [86] – [97]. The multi-functional RF component in [98] incorporates amplitude tunability into a BPF-VP. As shown in Figure 1.8 (a), it employs a trisection (three-resonator) filter network with tunable mixed electric/magnetic couplings, which inherently generates quadrature (I/Q) signals within the filter and enables continuous 0 – 360° phase tuning via a vector-sum mechanism. An independent resonator Q -factor tuning mechanism is introduced by adding an extra parallel branch consisting of one variable capacitor and one fixed resistance to provide ~ 10 dB of amplitude variation

while preserving the filter's frequency response. The measured prototype shown in Figure 1.8 (a) achieves a full 360° phase span and about 10 dB amplitude tuning over a 0.75–0.95 GHz center frequency range, with minimal phase/amplitude error ($< 8.45^\circ / < 2.58$ dB) and high linearity ($P_{1dB} \approx 11$ dBm). However, as indicated in the measured results in Figure 1.8 (b) and (c), the TFs characteristics change while tuning the phase or amplitude of the output RF signals. Overall, none of the reported BPF-VPs or BPF-VAs is scalable to higher order TFs or facilitates multi-band operability, limiting the application of these devices to systems requiring alternative filtering responses and different band coverages. Additionally, they fail to incorporate unilateral transmission characteristics.

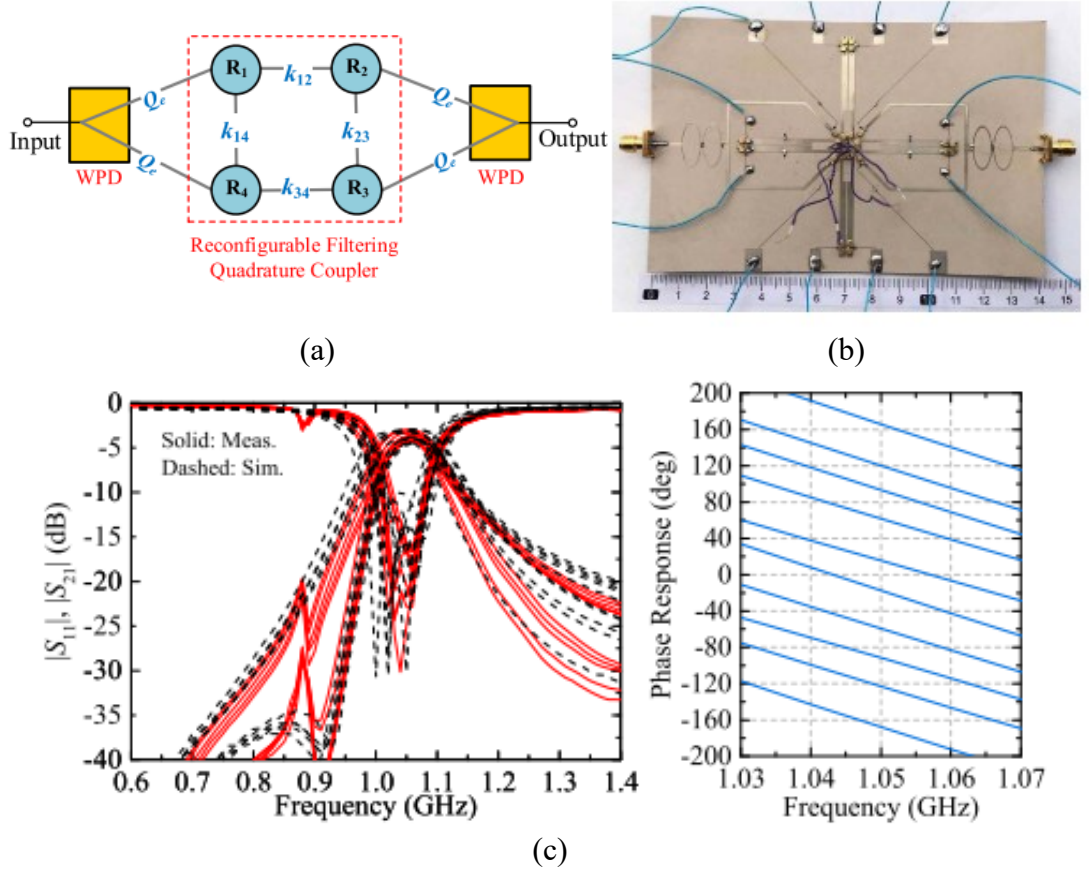


Figure 1.7. The vector-sum reconfigurable BPF-VP in [82] © 2022 IEEE. (a) The coupling schematic of the BPF-VP. (b) The experimental prototype. (c) The RF-measured amplitude and phase results.

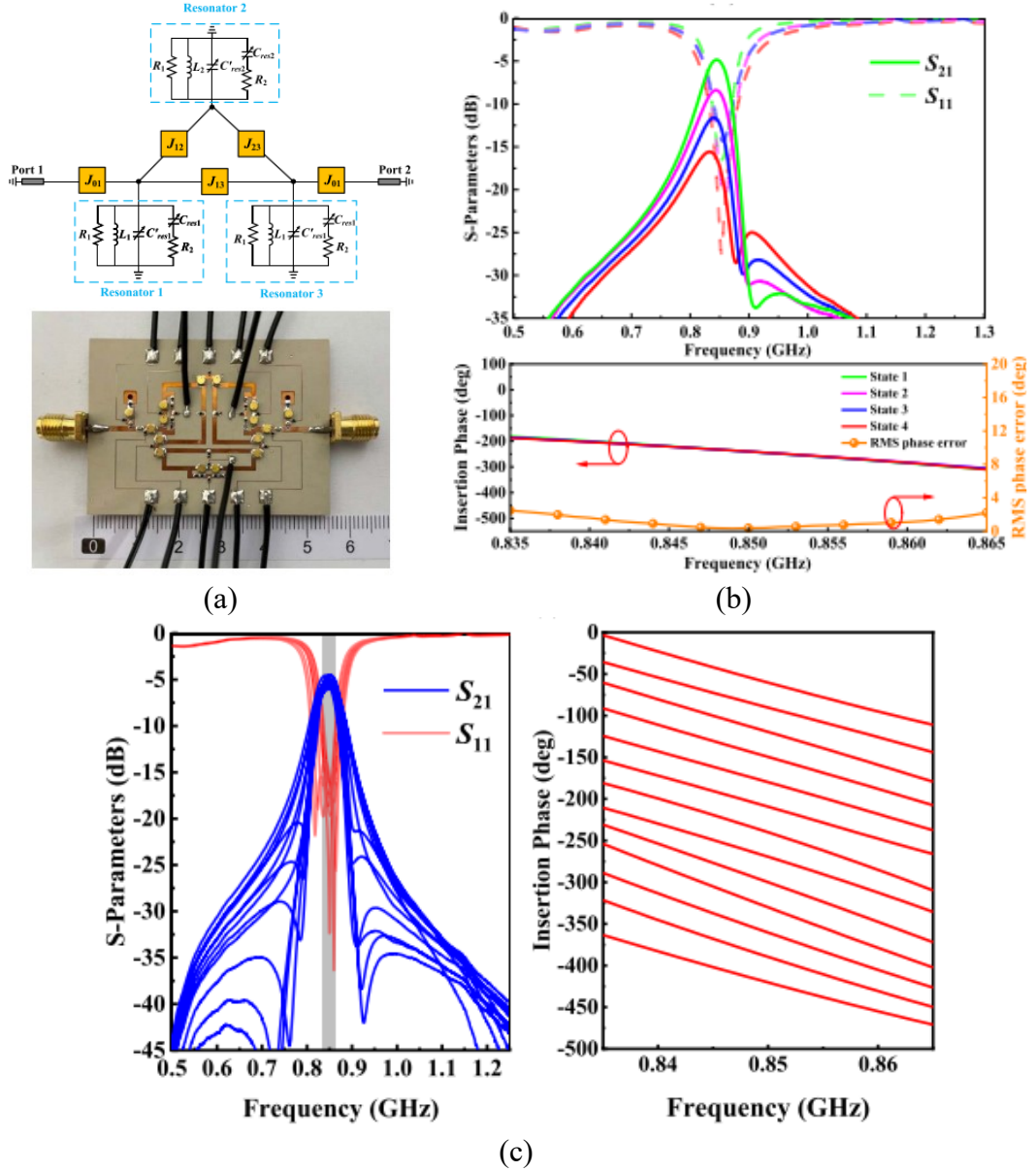


Figure 1.8. The reconfigurable amplitude-variable BPF-VP in [98] © 2022 IEEE. (a) Top: the circuit topology of the proposed amplitude-variable BPF-VP. Bottom: the experimental prototype. (b) The RF-measured amplitude and phase results when the output amplitude is tuned. (c) The RF-measured amplitude and phase results when the output phase is tuned.

1.3.3 Multi-port multi-functional filtering routing networks

In contrast to the single-input/single-output multifunctional BPFs discussed in Sections 1.3.1 and 1.3.2, this subsection considers more complex multi-port signal-routing networks. These systems represent a further step in RF front-end development, where multiple ports are leveraged to flexibly route signals across different channels while maintaining integrated BPF functionality. In particular, the co-integration of

BPFs into multi-port devices has been reported. These devices include BPFs supporting directional transmission (e.g., circulation and isolation) [51], [99], [68], [100], [101], [102], power division [103]–[108] and switching functions such as single-pole multi-throw (SPMT) or dual-channel routing [109]–[121].

Among the different multi-port devices, STM based BPFCs have attracted considerable attention because they can facilitate FD radios. For instance, [42], [43] demonstrated BPFCs by modulating the resonant frequencies of three coupled resonators with 120° phase-shifted signals. However, they suffer from high IL of around 10 dB and poor in-band RL of around 8 dB. Later on these drawbacks were addressed by incorporating unmodulated resonators to improve passband transmission and stopband rejection [51], [99]. Dual-band BPFCs were also demonstrated by modulating all the resonators of in-parallel cascaded BPF branches with 120° progressively phase shifted signals as shown in Figure 1.9 (a) and (b) [68]. The reported circulator prototype shown in Figure 1.9 (c) operates over independently tunable passbands (the lower band from 185 to 206 MHz, and the upper band from 238 to 270 MHz) with minimum IL of 5.6 dB, and highest achievable isolation levels of 50 dB as demonstrated in Figure 1.9 (d). In another approach, BPFCs are implemented by modulating coupling capacitors in mixed-coupling LC tank circuits [100], [101]. Both concepts extend the BPFC concept by integrating extra functions. For example, [100] co-integrates an RF diplexer and single-pole double-throw switch within a single RF component, whereas [101] incorporates a BPF to realize two-port transmission. Despite their ability to incorporate multiple RF functions, these STM implementations still rely on external RF-switching circuitry for operational mode configuration, offer no power-division capability, and are difficult to scale to multi-port architectures.

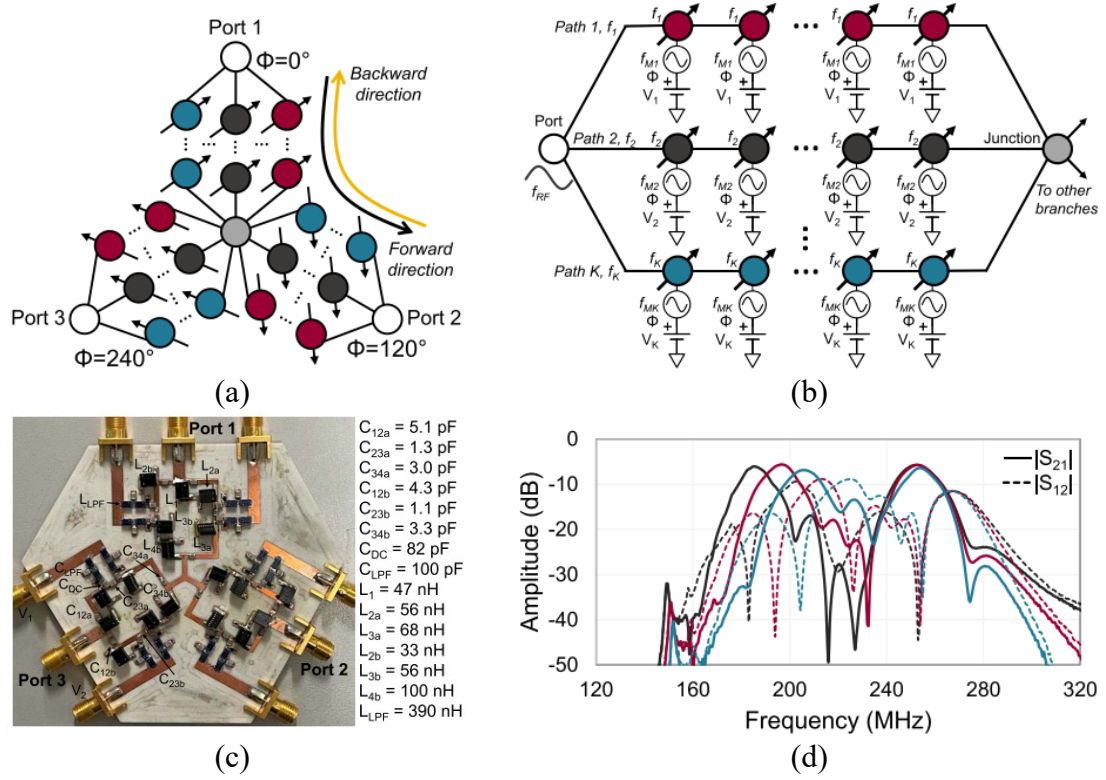


Figure 1.9. The multiband BPFC in [68] © 2023 IEEE. (a) Schematic of a K-band BPFC that comprises three N^{th} -order modulated BPF branches; colored circles: frequency tunable resonators, white circle: RF inputs, grey circle: junction, and black lines: impedance inverters. (b) Schematic and details of one of the three circulator branches. (c) The experimental prototype. (d) The RF-measured power transmission and reflection responses when the lower passband is tuned.

Power dividers (PDs) with embedded BPF isolators (PD-BPFIs) using STM resonator arrays have been explored. Unilateral transmission was achieved by applying progressive phase-shifted sinusoidal modulation to resonators embedded in conventional filtering power divider networks. Chaudhary et al. established systematic filtering PDs using time-modulated microstrip STM resonators, enabling unequal power divisions, arbitrary termination impedances, frequency tunability, dual-band operation, controlled transmission zeros, and broadband isolation [103], [106], [107], [108]. Zhao et al. demonstrated tunable PD-BPFIs based on substrate-integrated waveguide (SIW) cavities [105]. In this approach a common SIW cavity operating in mixed static and time-modulated degenerate modes is employed together with four triangular branch cavities to realize simultaneous power division, bandpass filtering, and nonreciprocal transmission as indicated in Figure 1.10 (a), (b) and (c). The RF-measured performance of the prototype at 2.4 GHz is presented in Figure 1.10 (d). It achieves a measured forward IL of 5.4 dB, a RL better than 13.5 dB, and output

isolation exceeding 21 dB across a 21 MHz nonreciprocal bandwidth. In addition, the amplitude imbalance between the two output ports is less than 0.6 dB and the phase difference is under 6.2° . Nevertheless, all the PD-BPFs rely on resistive elements for output isolation, lack the circulator and switching functionalities and do not support configurable input/output ports.

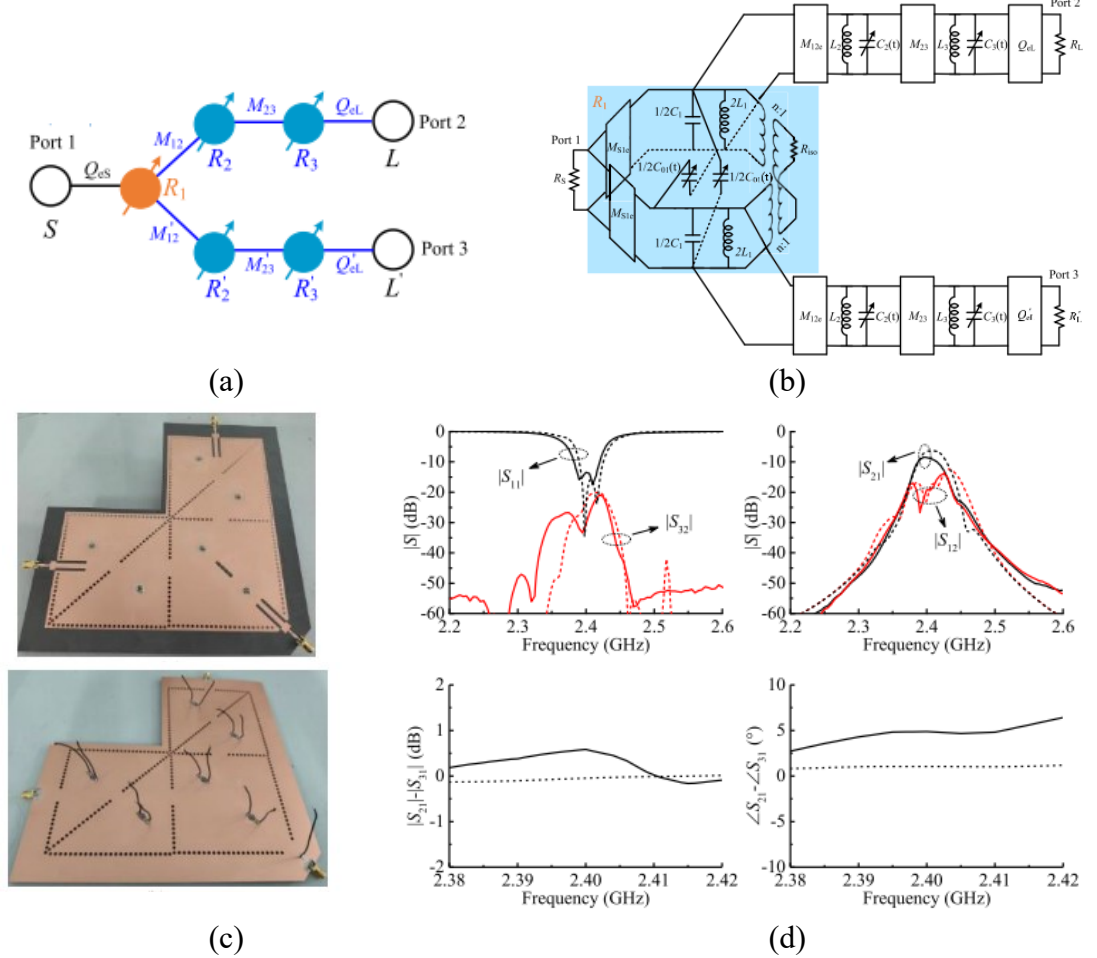


Figure 1.10. The frequency tunable PD-BPFI in [105] © 2024 IEEE. (a) CRD of the PD-BPFI using a common STM resonator and four branch STM resonators. (b) Equivalent circuit of the PD-BPFI. (c) The experimental prototype based on STM SIW resonators. (d) The simulated (dashed lines) and RF-measured results (solid lines) of the PD-BPFI with modulation at 2.4 GHz.

Single-pole multi-throw switches with integrated BPFs (SPMT-BPFs) have also been demonstrated [109]–[118]. They can be constructed by connecting multiple BPF branches. The switching function between different BPF branches can be achieved by detuning resonators or by controlling the inter-resonator couplings. PIN diodes are commonly used for switching purposes, which introduce additional loss and increase the power consumption. Additionally, they tend to be complex, lack isolator, PD or

circulator functionalities. In addition to SPMT-BPF switches that route the RF signal from the input port to different output ports, channel reconfigurable (CR) dual-channel routing networks with integrated BPF have also been explored in [119], [120], [121]. However, [120] and [121] can't facilitate power division and [119], [120], [121] fail to support unilateral signal transmission.

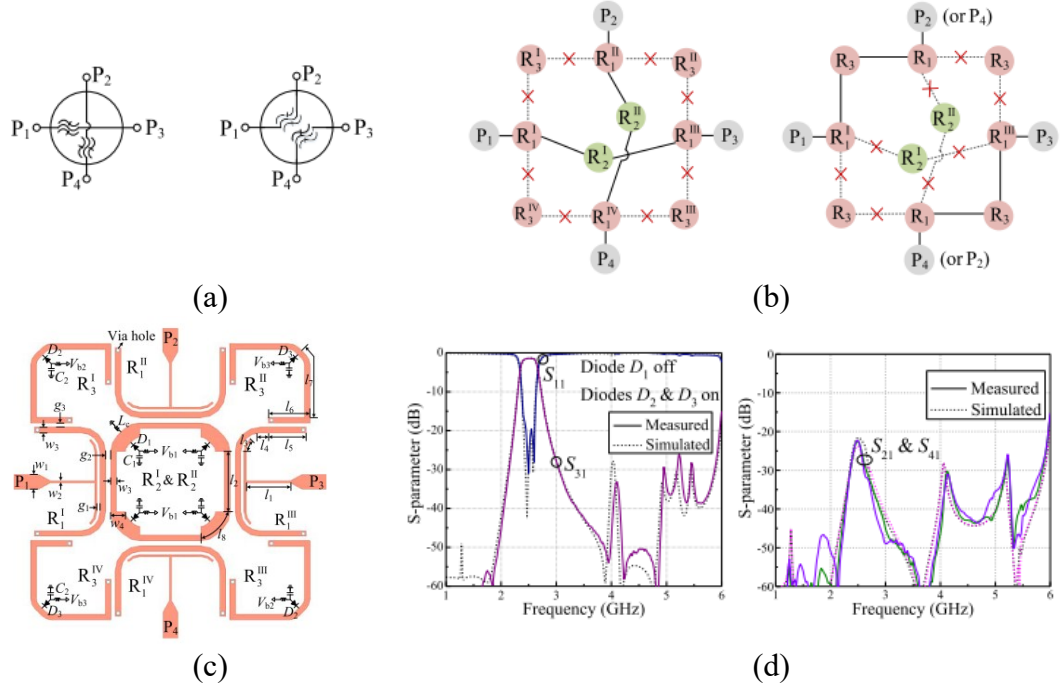


Figure 1.11. The T-Type filtering switch proposed in [120] © 2022 IEEE. (a) The operational states of the T-type switch integrated with bandpass responses with crossover-channel working state and adjacent-channel working state. (b) The CRD of the two different operational states. (c) The circuit configuration of the prototype. (d) The simulated (dotted lines) and RF-measured results (solid lines) of the T-type switch at crossover state.

1.4 Thesis Outline

STM devices have shown strong potential for realizing unilateral and multifunctional RF components. However, the state-of-the-art STM-based filtering devices remain constrained in several key aspects. Current implementations are often restricted to fixed TFs, offer limited tunability such as center frequency and BW. In addition, multifunctional integration such as combining unilateral filtering with VP or VA has not yet been achieved. Prior STM-based devices have primarily focused on unilateral transmission, overlooking the intrinsic switching off capability of STM

resonators, which inhibits the scalability and diverse routing capabilities of STM-based multiport networks. These constraints highlight the need for new STM-based architectures that facilitate higher reconfigurability, more functionalities, and more flexible signal routing. To address these limitations, this thesis explores novel non-reciprocal filtering architectures based on STM resonators to facilitate compact, reconfigurable, and multi-functional RF filtering components and networks. This thesis is organized as follows:

Chapter 2 presents a transversal-resonator-based design methodology that enables widely tunable transfer function characteristics, achieves high isolation between the forward and backward transmission, and supports multiple levels of single- and multi-band tunability. A detailed design methodology to synthesize the response of transversal resonator-based STM arrays is introduced for the first time, facilitating the synthesis of advanced RF filtering transfer functions based on transversal resonator arrays with incorporated unilateral transmission. The operating principles and scalability of the design method are demonstrated through the analysis of four distinct transversal resonator-based STM filtering topologies that facilitate the realization of high-order and highly-modular single-band and multi-band TFs with multiple levels of RF tuning including frequency tuning, bandwidth tuning, band controllability, and intrinsic switch-off capabilities. The concept has been validated at ultra high frequency (UHF) band through the manufacturing and testing of four lumped-element BPFs. The BPFs and contributions of this chapter have been presented in [122], [123].

Chapter 3 presents a novel RF co-design methodology for the realization of a new class of a multi-functional RF component that simultaneously incorporates the functionality of a single- or multi-band BPF, a VP with 360° continuously tunable phase shift, a VA with theoretically infinitely-tunable attenuation, an RF switch and an RF isolator that can be materialized within the volume of a single RF component. It is based on a power reconfigurable rat-race coupler, STM resonators and a switchable phase coupling element that controls three modes of operation. These include: (i) a co-designed single/multi-band reflectionless BPF, an RF isolator (enabled/disabled by turning ON/OFF STM) and a 360° -variable phase shifter (BPF/BPFI-VP), (ii) an RF switching mode and (iii) a single/multi-band BPF, an RF isolator (enabled/disabled by turning ON/OFF STM) and a variable attenuator

(BPF/BPFI-VA). A detailed theoretical framework is provided alongside a variety of single-band and multi-band design examples. The methodology is validated through the implementation of a single-band prototype and a dual-band prototype at 700 MHz. The multi-functional RF devices and contributions of this chapter have been presented in [124], [125], [126].

Chapter 4 demonstrates for the first time multi-port multi-functional RF signal routing networks to support multi-port RF switching (dynamically selecting signal paths), circulation and isolation (directional signal flow), and power division (splitting or combining signals across ports) within one device. They are based on reconfigurable STM resonator arrays that are arranged in star- and box-type configurations. Specifically, the star-type block supports three modes of operation: (i) a reconfigurable co-designed SPMT-BPFI, (ii) a PD-BPFI, and (iii) a BPFC. The box-type block supports two modes, namely: (i) a channel reconfigurable (CR) BPFI (CR-BPFI) mode that can support both concurrent two-channel or single-channel transmission, and (ii) a PD-BPFI mode. The proposed multi-port multi-functional STM-based array concept is demonstrated through circuit-level simulations and is validated by two experimental prototypes. The multi-port and multi-functional RF devices and contributions of this chapter have been presented in [127].

Chapter 5 summarizes the key contributions of this thesis and provides useful insights into further extensions of this work.

Chapter 2: Transversal-STM-Resonator-Based Bandpass Filtering Isolators with Tunable Transfer Function Characteristics

This chapter presents novel BPF designs using transversal-STM-resonators. By combining the tunability of transversal cells with the unilateral transmission enabled by STM, filtering and isolation are integrated within a single RF component. A detailed synthesis of transversal-STM-resonator-based arrays using spectrum network matrices is introduced for the analysis and design of reconfigurable TFs with directionality. Whereas spectral network parameter matrices of a time-varying resonator have been developed and applied in in-line BPF configurations [49], this method has not yet been applied to STM transversal cells, which serve as the building blocks for a wide variety of filters. The operating principles and scalability of the design method are demonstrated through the analysis and validation of four distinct transversal STM-resonator-based filtering topologies. These include a single-band BPF based on two cascaded transversal cells (Topology 1), a single-band fourth-order box-like BPF (Topology 2), a single-band sixth-order box-like BPF (Topology 3), and a six-resonator-based BPF using three series transversal cells (Topology 4). The BPFs and contributions of this chapter have been presented in [122], [123].

2.1 Analysis of STM-transversal-resonator-based BPFs

The details of the STM two-resonator transversal cell or doublet are illustrated in Figure 2.1 in terms of its CRD (Figure 2.1 (a)) and its circuit equivalent (Figure 2.1 (b)). The doublet, denoted by its ABCD matrix $\mathbf{N}_t$, can be considered as the parallel connection of two networks marked in blue and red, with their respective ABCD matrices being $\mathbf{N}_I$ and $\mathbf{N}_{II}$. The red (blue) network comprises a cascaded connection of an input inverter $M_{I,2}$ ($M_{I,3}$), an STM resonator R_1 (R_2) and an output inverter $M_{2,4}$ ($M_{3,4}$). The circuit equivalent of the STM doublet is provided in Figure 2.1 (b), where the STM resonators are implemented as parallel LC tanks whose capacitors are time-variant with capacitance $C_{0x}(t)$ as given in (2.1) and their inductors L_x (x is the index number of the resonator R_x) are time invariant. C_{0x} is the nominal capacitance, ξ_x is the modulation depth, f_m is the modulation frequency, and ϕ_x is the phase of the modulating waveform of the resonator R_x .

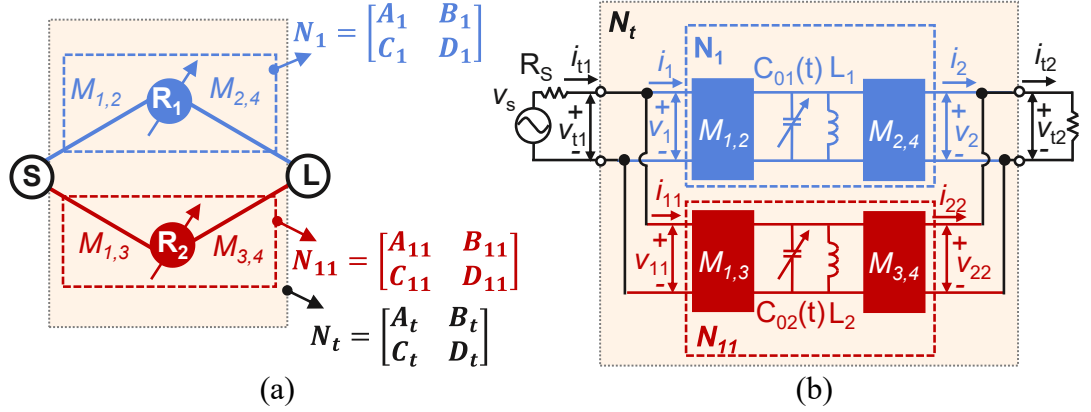


Figure 2.1 STM doublet concept. (a) CRD. Colored circles with arrows: STM resonators, white circles: non-resonating nodes. (b) Equivalent circuit where the STM resonators are represented by time-varying capacitors and time-invariant inverters and inductors. N_t , N_1 and N_{11} are the ABCD matrices of the doublet, the upper transversal path and the bottom transversal path.

The spectral admittance matrices $\mathbf{Y}_{C_x}$ and $\mathbf{Y}_{L_x}$ for the shunt time-varying capacitor and the time-invariant inductor are provided in (2.2). These matrices incorporate the energy conversion between different intermodulation frequencies $f \pm n f_m$ ($n=0, 1, 2$). As such, they should have theoretically infinite dimensions (i.e., n is infinite), but only a limited number of intermodulation frequencies need to be considered in practice. Using the expressions in (2.1) and (2.2), the ABCD matrix of a single time-varying resonator $\mathbf{N}_{R_x}$ and a time-invariant inverter $\mathbf{N}_{M_{a,b}}$ can be calculated using (2.3) and (2.4), where a and b are the indices of the inverters, $\mathbf{U}$ is the unity matrix and $\mathbf{0}$ is the zero matrix with the same dimension as the spectral admittance matrices $\mathbf{Y}_{C_x}$ and $\mathbf{Y}_{L_x}$. Since the blue and the red parts are the cascade connection of three two-port networks, $\mathbf{N}_1$ and $\mathbf{N}_{11}$ can be easily found by multiplying the ABCD matrices of the individual two-port subnetworks as shown in (2.5) and (2.6), respectively. $\mathbf{N}_t$ can be derived from $\mathbf{N}_1$ and $\mathbf{N}_{11}$ by considering the signal relations in (2.7)–(2.10) and its matrix elements are provided in (2.11)–(2.14). They have been obtained by substituting (2.7) and (2.8) in the relationships outlined in (2.9) and (2.10) that lead to V_{t1} , V_{t2} , I_{t1} and I_{t2} , which can be used to calculate the ABCD elements of $\mathbf{N}_t$. The S-parameters of the transversal cell $\mathbf{S}_t$ can be calculated from $\mathbf{N}_t$ using (2.15) where S_{11} , S_{21} , S_{12} and S_{22} represent the elements in $\mathbf{S}_t$ and A , B , C and D are the elements of $\mathbf{N}_t$ as provided in [49]. Once the ABCD and scattering matrices of a single STM two-resonator transversal cell are obtained, the scattering matrices of

$$C_{0x}(t) = C_{0x} + \zeta_x C_{0x} \cos(2\pi f_m t + \phi_x) \quad (2.1)$$

$$Y_{Cx} = j2\pi C_x \begin{bmatrix} f-2f_m & (f-2f_m)\frac{\zeta_x}{2}e^{-j\phi_x} & 0 & 0 & 0 \\ (f-f_m)\frac{\zeta_x}{2}e^{j\phi_x} & f-f_m & (f-f_m)\frac{\zeta_x}{2}e^{-j\phi_x} & 0 & 0 \\ 0 & f\frac{\zeta_x}{2}e^{j\phi_x} & f & f\frac{\zeta_x}{2}e^{-j\phi_x} & 0 \\ 0 & 0 & (f+f_m)\frac{\zeta_x}{2}e^{j\phi_x} & f+f_m & (f+f_m)\frac{\zeta_x}{2}e^{-j\phi_x} \\ 0 & 0 & 0 & (f+2f_m)\frac{\zeta_x}{2}e^{j\phi_x} & f+2f_m \end{bmatrix} \quad (2.2)$$

$$Y_{Lx} = 1/j2\pi L_x \begin{bmatrix} f-2f_m & 0 & 0 & 0 & 0 \\ 0 & f-f_m & 0 & 0 & 0 \\ 0 & 0 & f & 0 & 0 \\ 0 & 0 & 0 & f+f_m & 0 \\ 0 & 0 & 0 & 0 & f+2f_m \end{bmatrix}$$

$$\mathbf{N}_{Rx} = \begin{bmatrix} \mathbf{A}_{Rx} & \mathbf{B}_{Rx} \\ \mathbf{C}_{Rx} & \mathbf{D}_{Rx} \end{bmatrix} = \begin{bmatrix} \mathbf{U} & \mathbf{0} \\ \mathbf{Y}_{Lx} + \mathbf{Y}_{Cx} & \mathbf{U} \end{bmatrix} \quad (2.3)$$

$$\mathbf{N}_{M_{a,b}} = \begin{bmatrix} \mathbf{A}_{M_{a,b}} & \mathbf{B}_{M_{a,b}} \\ \mathbf{C}_{M_{a,b}} & \mathbf{D}_{M_{a,b}} \end{bmatrix} = \begin{bmatrix} \mathbf{0} & \frac{j}{\mathbf{M}_{a,b}}\mathbf{U} \\ j\mathbf{M}_{a,b}\mathbf{U} & \mathbf{0} \end{bmatrix} \quad (2.4)$$

$$\mathbf{N}_1 = \mathbf{N}_{M_{1,2}} \times \mathbf{N}_{R1} \times \mathbf{N}_{M_{2,4}} \quad (2.5)$$

$$\mathbf{N}_{11} = \mathbf{N}_{M_{1,3}} \times \mathbf{N}_{R2} \times \mathbf{N}_{M_{3,4}} \quad (2.6)$$

$$\mathbf{V}_1 = \mathbf{A}_1 \mathbf{V}_2 + \mathbf{B}_1 \mathbf{I}_2, \mathbf{V}_{11} = \mathbf{A}_{11} \mathbf{V}_{22} + \mathbf{B}_{11} \mathbf{I}_{22} \quad (2.7)$$

$$\mathbf{I}_1 = \mathbf{C}_1 \mathbf{V}_2 + \mathbf{D}_1 \mathbf{I}_2, \mathbf{I}_{11} = \mathbf{C}_{11} \mathbf{V}_{22} + \mathbf{D}_{11} \mathbf{I}_{22} \quad (2.8)$$

$$V_{t1} = V_1 = V_{11}, V_{t2} = V_2 = V_{22} \quad (2.9)$$

$$I_{t1} = I_1 + I_{11}, I_{t2} = I_2 + I_{22} \quad (2.10)$$

$$\mathbf{A}_t = (\mathbf{B}_1^{-1} + \mathbf{B}_{11}^{-1})^{-1}(\mathbf{B}_1^{-1}\mathbf{A}_1 + \mathbf{B}_{11}^{-1}\mathbf{A}_{11}) \quad (2.11)$$

$$\mathbf{B}_t = (\mathbf{B}_1^{-1} + \mathbf{B}_{11}^{-1})^{-1} \quad (2.12)$$

$$\mathbf{C}_t = \mathbf{C}_1 + \mathbf{C}_{11} + (\mathbf{D}_1 - \mathbf{D}_{11})(\mathbf{B}_1 + \mathbf{B}_{11})^{-1}(\mathbf{A}_{11} - \mathbf{A}_1) \quad (2.13)$$

$$\mathbf{D}_t = \mathbf{D}_{11} + (\mathbf{D}_1 - \mathbf{D}_{11})(\mathbf{B}_1 + \mathbf{B}_{11})^{-1}\mathbf{B}_{11} \quad (2.14)$$

$$\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \begin{bmatrix} U - 2[U + (\mathbf{A}\mathbf{R}_L + \mathbf{B})(\mathbf{C}\mathbf{R}_S\mathbf{R}_L + \mathbf{D}\mathbf{R}_S)^{-1}]^{-1} & 2[U + (\mathbf{A}\mathbf{R}_L + \mathbf{B})(\mathbf{C}\mathbf{R}_S\mathbf{R}_L + \mathbf{D}\mathbf{R}_S)^{-1}]^{-1} \times \\ & [\mathbf{A} - (\mathbf{A}\mathbf{R}_L + \mathbf{B})(\mathbf{C}\mathbf{R}_L + \mathbf{D})^{-1}\mathbf{C}] \\ 2\sqrt{\frac{R_S}{R_L}}[A + \frac{B}{R_L} + \mathbf{C}\mathbf{R}_S + \mathbf{D}\frac{R_S}{R_L}]^{-1} & U - 2[U + (\mathbf{A}\mathbf{R}_L + \mathbf{C}\mathbf{R}_S\mathbf{R}_L)^{-1}(\mathbf{B} + \mathbf{D}\mathbf{R}_S)]^{-1} \end{bmatrix} \quad (2.15)$$

filtering topologies based on STM transversal cells becomes a straightforward process, as they can be regarded as the cascading connection between single STM resonators, time-invariant inverters and STM transversal cells.

To evaluate the feasibility of the aforementioned design method for the synthesis of highly-selective filtering TFs, four fully-directional filtering topologies (Topology 1–4) are considered and are presented in the next section. Specifically, Topologies 1 and 2 demonstrate two different ways to realize 4th order TFs by arranging 4 STM resonators in either two cascaded doublets or in a box-like configuration. Furthermore, by adding STM resonators at the input and output ports of the Topology 2 CRD, a sixth order transfer function can be obtained as shown in Topology 3. To facilitate multiple levels of reconfigurability within a single RF filtering configuration, three series cascaded transversal cells need to be used as shown in the Topology 4 CRD.

2.1.1 Topology 1: fourth-order cascaded transversal cells BPFI

The CRD details of the Topology 1 are provided in Figure 2.2. In each transversal cell, the two resonators are asynchronously tuned and the inverters $M_{3,4}$ and $M_{5,7}$ need to have opposite polarity to obtain a four-pole bandpass TF. All the STM resonators (denoted by the black circles) consist of a parallel time-invariant inductor with the same inductance L_r and time-varying capacitors, whose capacitance are labeled in the boxes next to them. Specifically, all the time-varying capacitors have a nominal capacitance C_{0x} , and have the same modulation frequency f_m and modulation depth ξ . To enable directionality, the phase arrangement for the capacitors should follow the configuration labeled in the figure, i.e., 0° for R_2 and R_3 and φ° for R_6 and R_7 . For analysis purposes, the Topology 1 S-parameters can be calculated from its overall ABCD matrix that is equal to $\mathbf{N}_{t1} \times \mathbf{N}_{M4,5} \times \mathbf{N}_{t2}$ where $\mathbf{N}_{t1}$ is the ABCD matrix of the first doublet, $\mathbf{N}_{M4,5}$ is the ABCD matrix of the inverter and $\mathbf{N}_{t2}$ is the ABCD matrix of the last doublet.

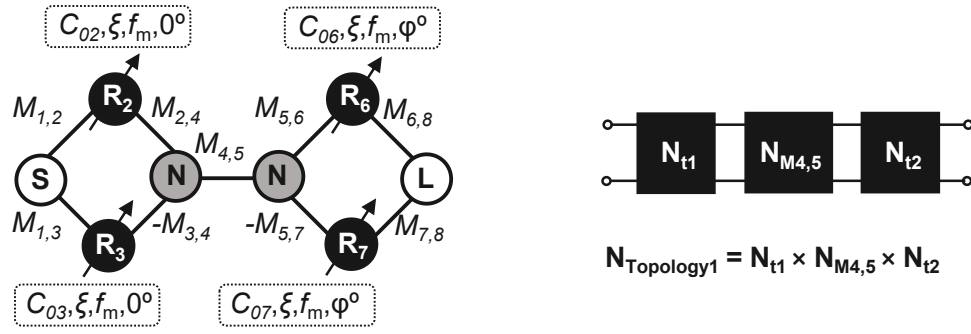


Figure 2.2. CRD of Topology 1.

For validation purposes, the synthesized S-parameters of Topology 1 are compared with their corresponding circuit-simulated ones (obtained in ADS from Keysight, utilizing ideal LC components and non-linear capacitors with the time-varying capacitance) and are provided in Figure 2.3 using the element values in the caption and considering an intermodulation index of 4 [intermodulation frequencies $f \pm nf_m$ ($n=0, 1, 2, 4$) is considered]. It should be noted that the coupling element values have been obtained by optimization. As shown, the BPF works at 700 MHz with a FBW of 10%. Furthermore, as observed in the left plot of Figure 2.3, when STM is OFF ($\xi = 0$), the filter response is reciprocal, i.e., $|S_{21}| = |S_{12}|$. When STM is ON ($\xi \neq 0$), really high IS can be created between the forward and backward transmission directions, as shown in the right plot of Figure 2.3. Moreover, a strong agreement between the synthesized results and the circuit simulated results can be observed, indicating the validity of the proposed synthesis method using STM transversal cells.

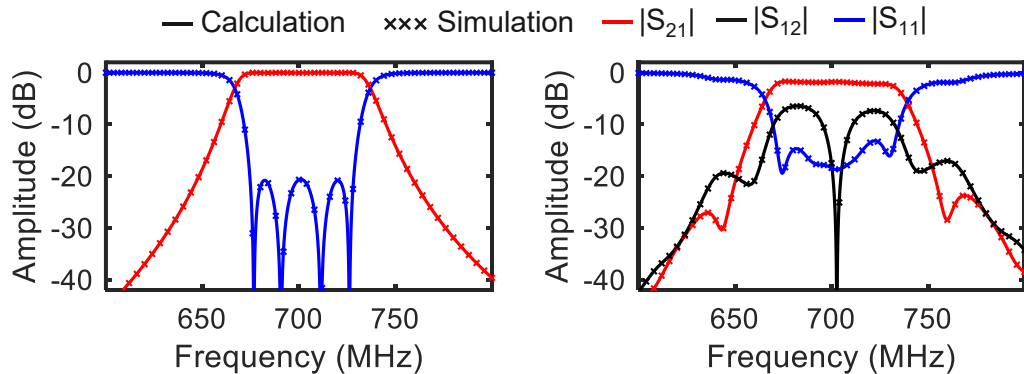


Figure 2.3. Synthesized and circuit-simulated S-parameters when STM is OFF (left) and ON (right). The coupling coefficients (obtained by optimization) are $M_{1,2} = M_{2,4} = M_{5,6} = M_{6,8} = M_{1,3} = M_{3,4} = M_{5,7} = M_{7,8} = 0.62$, $M_{4,5} = 1.56$, $M_{2,2} = M_{6,6} = 0.62$, $M_{3,3} = M_{7,7} = -0.67$ and the resonator element values are: $L_r = 2.4$ nH, $C_{02} = C_{06} = 23.06$ pF, $C_{02} = C_{06} = 20$ pF.

The tuning capabilities of Topology 1 are examined in Figure 2.4 through various design cases. As shown in Figure 2.4 (a), the center frequency of the filter can be tuned to lower (higher) frequencies by increasing (decreasing) the capacitance of all the four resonators at the same time whilst preserving high isolation between the forward and the backward direction. The BW tuning capabilities of the BPF are presented in Figure 2.4 (b). Wider (narrower) BW can be achieved by increasing (decreasing) the value of C_{02} and C_{06} while decreasing (increasing) the value of C_{03} and C_{07} . Furthermore, as shown in Case 4 in Figure 2.4 (b), the BPF can be intrinsically-switched off when C_{02} , C_{06} , C_{03} and C_{07} have the same value. For all of these cases the component element values are provided in the caption of Figure 2.3 and Figure 2.4. It is noted that the parameters of the time-varying capacitors should be adjusted accordingly in different tuning stages to achieve optimal isolation levels.

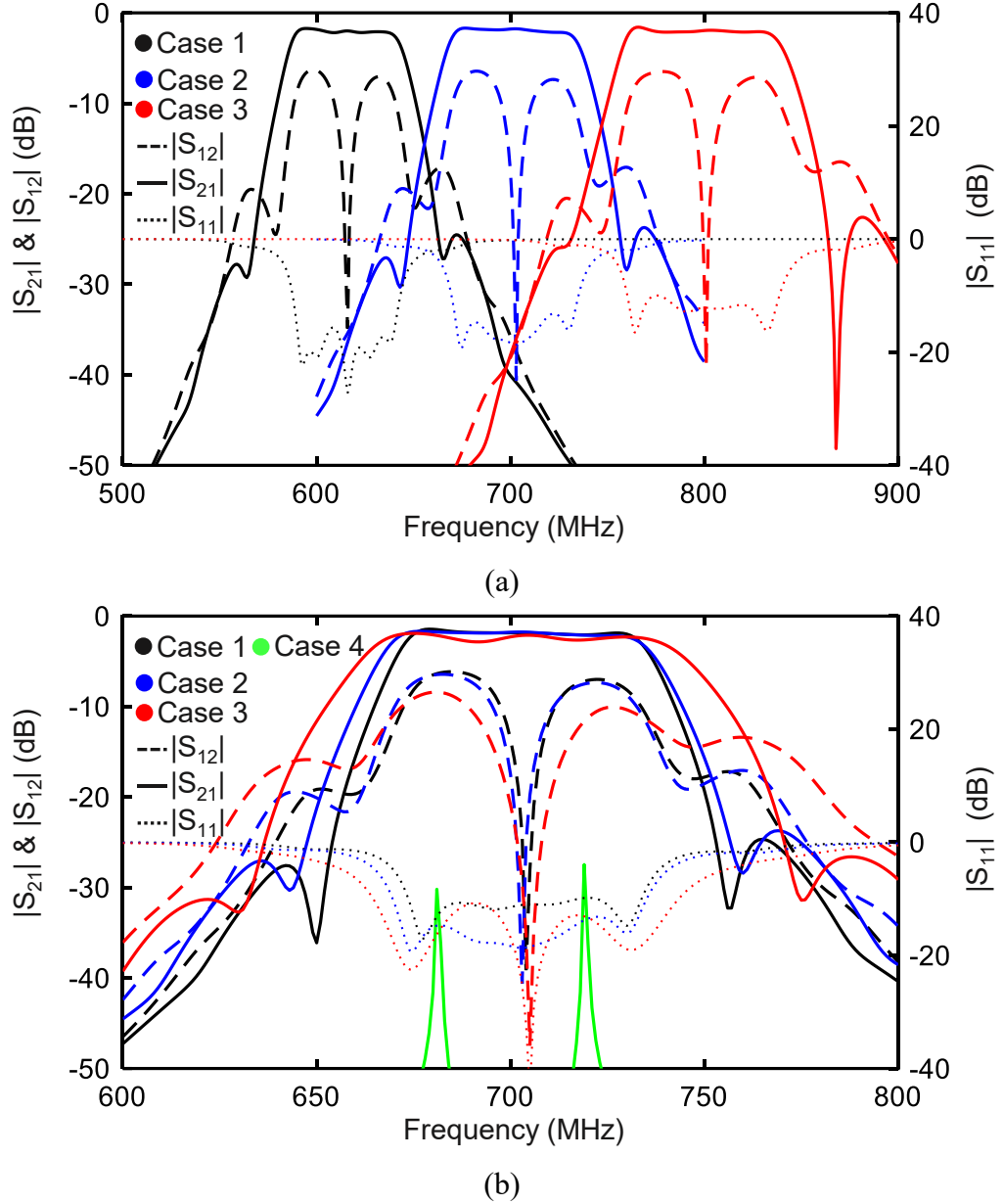


Figure 2.4. Circuit-simulated tuning capabilities for Topology 1. (a) Center frequency tuning. The parameters are: Case 1 : $C_{02}=C_{06}=30.05$ pF, $C_{03}=C_{07}=26.04$ pF, $f_m=26$ MHz, $\xi=0.068$, $\varphi=53$; Case 2: $C_{02}=C_{06}=23.06$ pF, $C_{03}=C_{07}=20$ pF, $f_m=30$ MHz, $\xi=0.067$, $\varphi=53$; Case 3: $C_{02}=C_{06}=15.44$ pF, $C_{03}=C_{07}=20$ pF, $f_m=30$ MHz, $\xi=0.067$, $\varphi=53$. (b) BW tuning. The parameters are: Case 1 : $C_{02}=C_{06}=22.74$ pF, $C_{03}=C_{07}=20.1$ pF, $f_m=26.5$ MHz, $\xi=0.061$, $\varphi=53$; Case 2: $C_{02}=C_{06}=23.06$ pF, $C_{03}=C_{07}=20$ pF, $f_m=30$ MHz, $\xi=0.067$, $\varphi=53$; Case 3: $C_{02}=C_{06}=23.46$ pF, $C_{03}=C_{07}=19.56$ pF, $f_m=33$ MHz, $\xi=0.088$, $\varphi=74$. Case 4: $C_{02}=C_{06}=C_{03}=C_{07}=21.54$ pF, $\xi=0$, $\varphi=60$.

2.1.2 Topology 2: fourth-order box-like BPF

The details of the Topology 2 CRD are provided in Figure 2.5 (a). It demonstrates an alternative way to functionalize a fourth order transfer function by

connecting two synchronously tuned resonators R_2 and R_5 , directly to a single transversal cell, forming a box-like topology. A single-band bandpass transfer function can be obtained when resonators R_2 and R_5 are set to resonate at the center frequency of the BPF, while R_3 and R_4 resonate at lower and higher frequencies. Using as a basis the design methodology mentioned above, a fourth order BPF was designed at 700 MHz with a fractional BW of 9% using the parameters specified in the captions of Figure 2.5 (coupling element values have been obtained by optimization). Furthermore, R_2 is modulated with an AC signal having a phase of 0° , R_3 and R_4 with an AC signal having a phase of φ° and R_5 with a phase of $2\varphi^\circ$. Figure 2.5 (b) and Figure 2.5 (c) demonstrate a comparison between the synthesized S-parameters and the ones obtained by circuit simulations which appear to be in excellent agreement. The tuning capabilities of Topology 2 are shown in Figure 2.6 in terms of center frequency (Figure 2.6 (a), Case 1-3), BW tuning (Figure 2.6 (b), Case 1-3)) and intrinsic switching off (Figure 2.6 (b), Case. 4). They have been obtained in a similar manner to the ones of Topology 1.

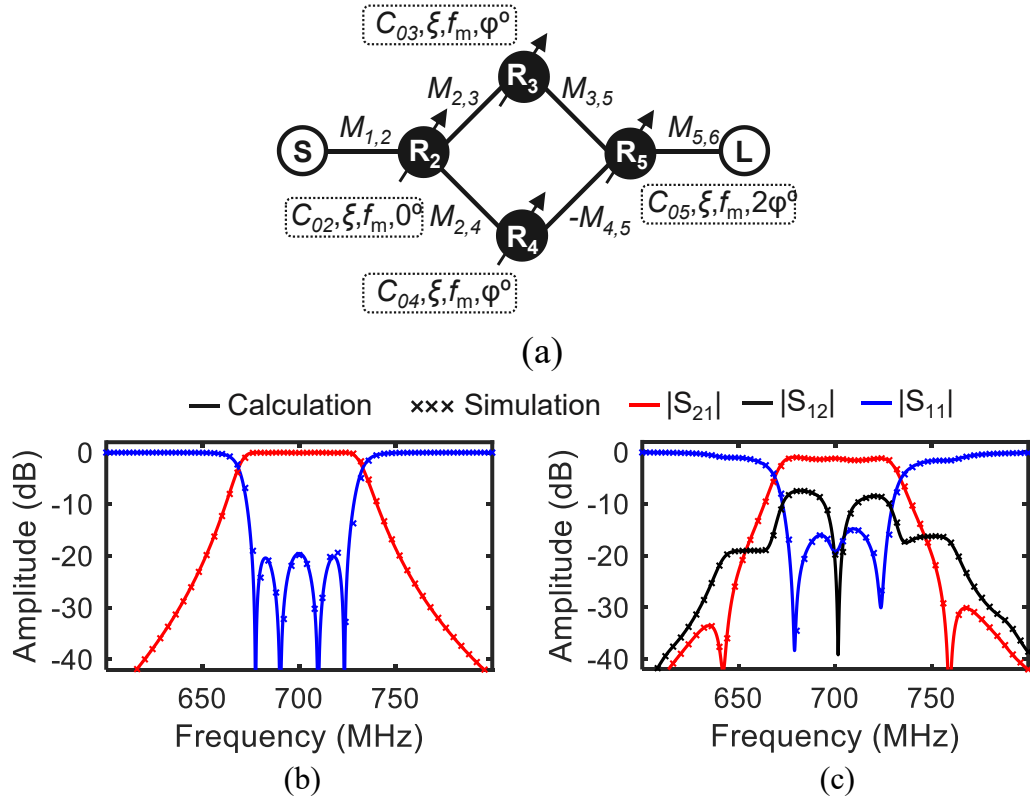
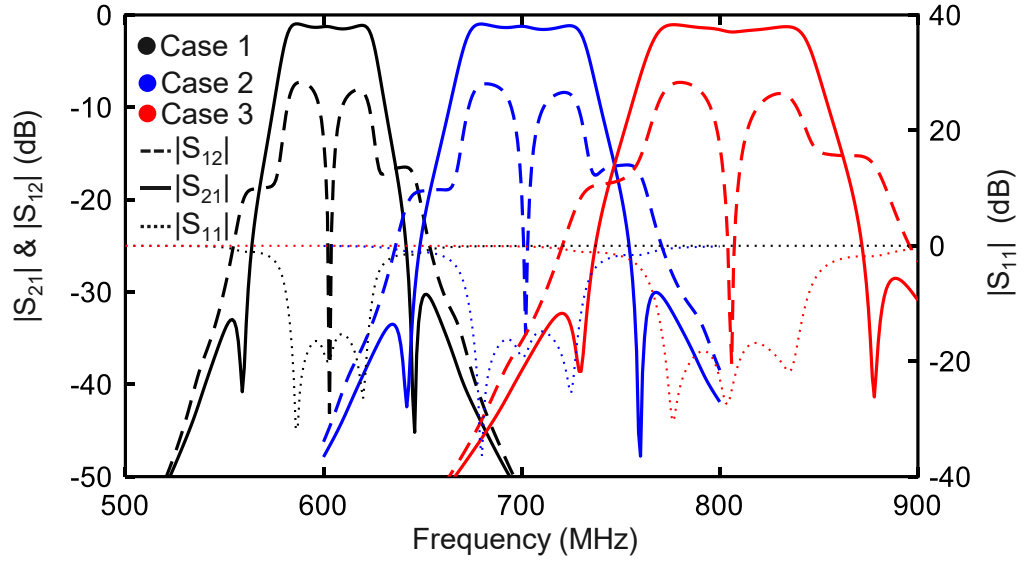
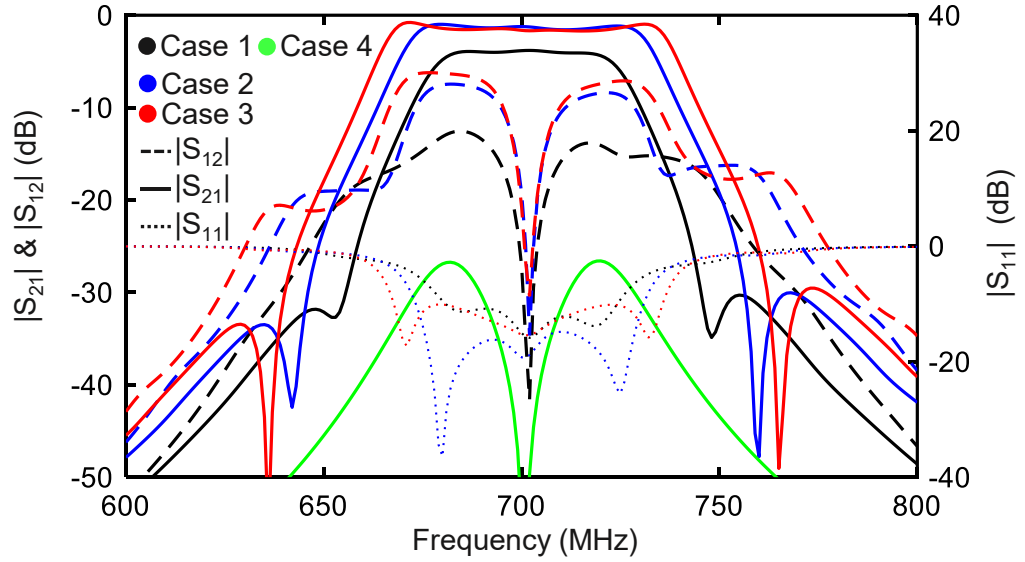


Figure 2.5. (a) Topology 2. (b), (c) Synthesized and circuit-simulated S-parameters of Topology 2 when STM is OFF (left) and ON (right). The coupling coefficients obtained by optimization are $M_{1,2}=M_{5,6}=0.936$, $M_{2,3}=M_{3,5}=M_{2,4}=M_{4,5}=0.52$, $M_{2,2}=M_{5,5}=0$, $M_{3,3}=0.562$, $M_{4,4}=-0.562$ and the resonator element values are: $L_r=7.74$ nH, $C_{02}=C_{05}=6.68$ pF, $C_{03}=7.02$ pF, $C_{04}=6.35$ pF.



(a)



(b)

Figure 2.6. Circuit-simulated tuning capabilities for Topology 2. (a) Center frequency tuning. The parameters are: Case 1: $C_{02}=C_{05}=9.035$ pF, $C_{03}=9.43$ pF, $C_{04}=8.65$ pF, $f_m=22$ MHz, $\xi=0.055$, $\varphi=60$; Case 2: $C_{02}=C_{05}=6.68$ pF, $C_{03}=7.02$ pF, $C_{04}=6.35$ pF, $f_m=30$ MHz, $\xi=0.065$, $\varphi=60$; Case 3: $C_{02}=C_{05}=5.08$ pF, $C_{03}=5.38$ pF, $C_{04}=4.8$ pF, $f_m=37$ MHz, $\xi=0.073$, $\varphi=60$; (b) BW tuning. The parameters are: Case 1: $C_{02}=C_{05}=6.68$ pF, $C_{03}=6.848$ pF, $C_{04}=6.497$ pF, $f_m=23.5$ MHz, $\xi=0.06$, $\varphi=60$; Case 2: $C_{02}=C_{05}=6.68$ pF, $C_{03}=7.02$ pF, $C_{04}=6.35$ pF, $f_m=30$ MHz, $\xi=0.065$, $\varphi=60$; Case 3: $C_{02}=C_{05}=6.68$ pF, $C_{03}=7.151$ pF, $C_{04}=6.26$ pF, $f_m=32$ MHz, $\xi=0.065$, $\varphi=60$; Case 4: $C_{02}=C_{05}=C_{03}=C_{04}=6.68$ pF, $f_m=30$ MHz, $\xi=0$, $\varphi=60$.

2.1.3 Topology 3: six-order box-like BPF

The Topology 3 CRD in Figure 2.7 demonstrates scalability to higher-order transfer functions by cascading two additional resonators in the box-like Topology 2 configuration, resulting in a sixth-order single-band bandpass transfer function. As depicted in Figure 2.7 (a), to achieve directionality, the modulating AC signals need to be applied with the following phase progression: 0° for the signal applied to R_2 , φ° for R_3 , $2\varphi^\circ$ for R_4 and R_5 , $3\varphi^\circ$ for R_6 , and $4\varphi^\circ$ for R_7 . The comparison between the synthesized S-parameters and the circuit-simulated ones is presented in Figure 2.7 (b) and Figure 2.7 (c). They have been obtained with the same method as above and appear to be in an excellent agreement, successfully validating the proposed synthesis method. The tuning capabilities of Topology 3 are similar to the ones obtained for Topology 2, thus omitted here for brevity.

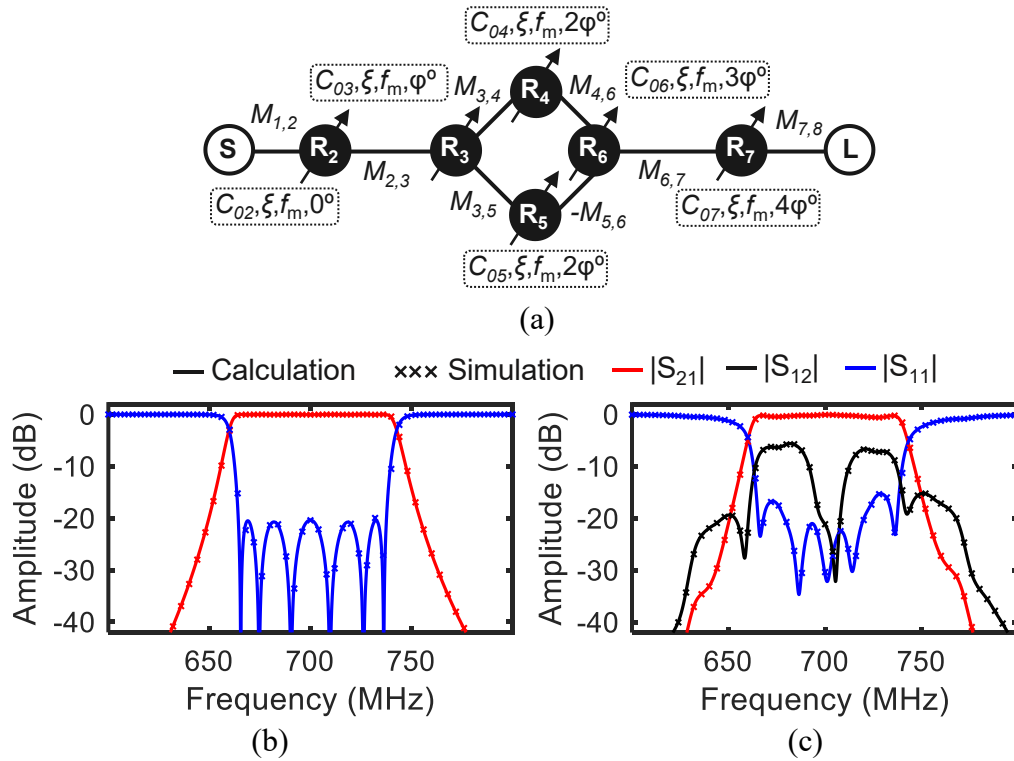


Figure 2.7. (a) Topology 3. (b), (c) Synthesized and circuit-simulated S-parameters of Topology 3 when STM is OFF (left) and ON (right). The coupling coefficients are obtained by optimization and are $M_{1,2} = M_{7,8} = 0.769$, $M_{2,3} = M_{6,7} = 0.484$, $M_{3,4} = M_{4,6} = M_{3,5} = M_{5,6} = 0.244$, $M_{2,2} = M_{3,3} = M_{6,6} = M_{7,7} = 0$, $M_{4,4} = 0.356$, $M_{5,5} = -0.326$ and the resonator element values are: $L_r = 6$ nH, $C_{02} = C_{03} = C_{06} = C_{07} = 8.6$ pF, $C_{04} = 9.16$ pF, $C_{05} = 8.1$ pF.

2.1.4 Topology 4: multi-configurable cascaded transversal cell BPFI

Topology 4 facilitates the realization of multi-configurable TFs by only tuning the resonant frequencies of its constituent resonators. Specifically, it enables various levels of tunability, such as center frequency, bandwidth, intrinsic switching off, reconfigurability of the number of passbands through band merging and cancellation and tuning of the out-of-band rejection by transmission zero (TZ) reallocation. To achieve these functionalities, all inverters share the same polarity, and the phases of the two resonators within the same STM transversal cell are identical. However, a successive phase difference is required for adjacent transversal cells, as illustrated in Figure 2.8.

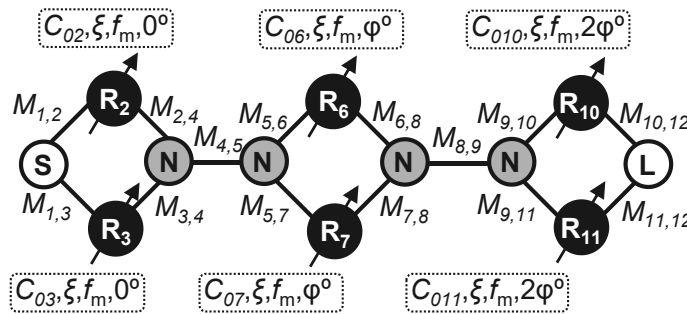


Figure 2.8. Topology 4. The coupling coefficients (obtained by optimization) are $M_{1,2} = M_{5,6} = M_{6,8} = M_{10,12} = 0.837$, $M_{1,3} = M_{5,7} = M_{7,8} = M_{11,12} = 0.86$, $M_{2,4} = M_{9,10} = 1$, $M_{3,4} = M_{9,11} = 1.028$, $M_{8,9} = M_{4,5} = 1.364$, and the resonator inductances are: $L_r = 2.4$ nH.

Table 2.1. Parameters for time-varying capacitance for the different tuning cases of Figure 2.9

	1	2	3	4	5	6
C₀₂ (pF)	20.236	21.5	15.6	15.5	15.5	18.055
C₀₃ (pF)	23	21.5	21.5	21.5	21.5	26.552
C₀₆ (pF)	20.236	21.5	14.6	21.5	21.5	22.1
C₀₇ (pF)	23	21.5	21.5	21.5	21.5	22.02
C₀₁₀ (pF)	20.236	21.5	26.6	30.7	21.5	7.568
C₀₁₁ (pF)	23	21.5	21.5	21.5	21.5	63.73
f_m(MHz)	20	33	19.5	21	24	20
ξ	0.05	0.085	0.052	0.053	0.07	0.05
$\phi(^{\circ})$	65	65	65	65	65	65

A summary of the tuning capabilities of Topology 4 is provided in Figure 2.9, using the component element values in the caption and in Table 2.1. The coupling coefficients for all 6 tuning states are the same as listed in Figure 2.8. As shown in Figure 2.9 (a), a third order dual-band transfer function can be generated when resonators R_2 , R_6 and R_{10} are set to the same frequency that coincides with that of the higher passband while resonators R_3 , R_7 and R_{11} resonate at the lower passband. In another configuration [see Case 2 in Figure 2.9 (b)], an equi-ripple response can be generated by setting all the resonators to resonate at the same frequency. Furthermore, when placing three resonators inside the passband (resonators R_3 , R_7 and R_{11}), and the rest are outside of the passband, transfer functions with three TZs can be generated, as demonstrated in Figure 2.9 (c). Figure 2.9 (d) and Figure 2.9 (e) show the TFs that can be obtained when placing resonators R_3 , R_6 , R_7 and R_{11} inside the passband, facilitating quasi-elliptic transfer functions with two symmetrically allocated TZs around the passband or with just one TZ. The filter can also be intrinsically switched-off [see Case 6 in Figure 2.9 (f)] when setting resonators R_6 and R_7 to resonate at the same frequency while offsetting the other resonators from the center frequency of the filter. Although not shown here, the center frequency in all the aforementioned cases can be freely tuned through altering all the resonators' resonating frequency at the same time as it will be shown in the experimental validation of this concept. Furthermore, the TZs in Cases 3 to 5 can be placed at either side of the passband. Their location can be altered by tuning the corresponding resonators. As it can be observed from Figure 2.9, large isolation levels can be achieved in all the tuning states, demonstrating the effectiveness and versatility of the proposed topologies.

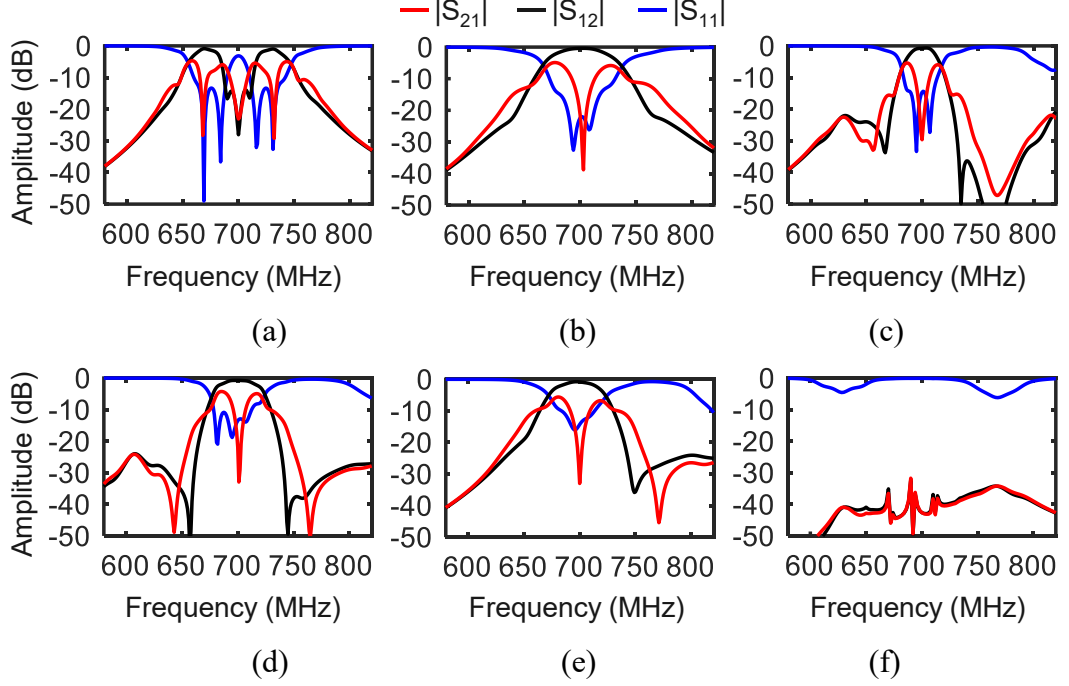


Figure 2.9. Circuit-simulated tuning capabilities of Topology 4 using the component values in Table I. (a) Case 1, (b) Case 2, (c) Case 3, (d) Case 4, (e) Case 5, (f) Case 6.

2.2 STM Parameters Selection and Practical Realization Aspects

2.2.1 STM parameter selection

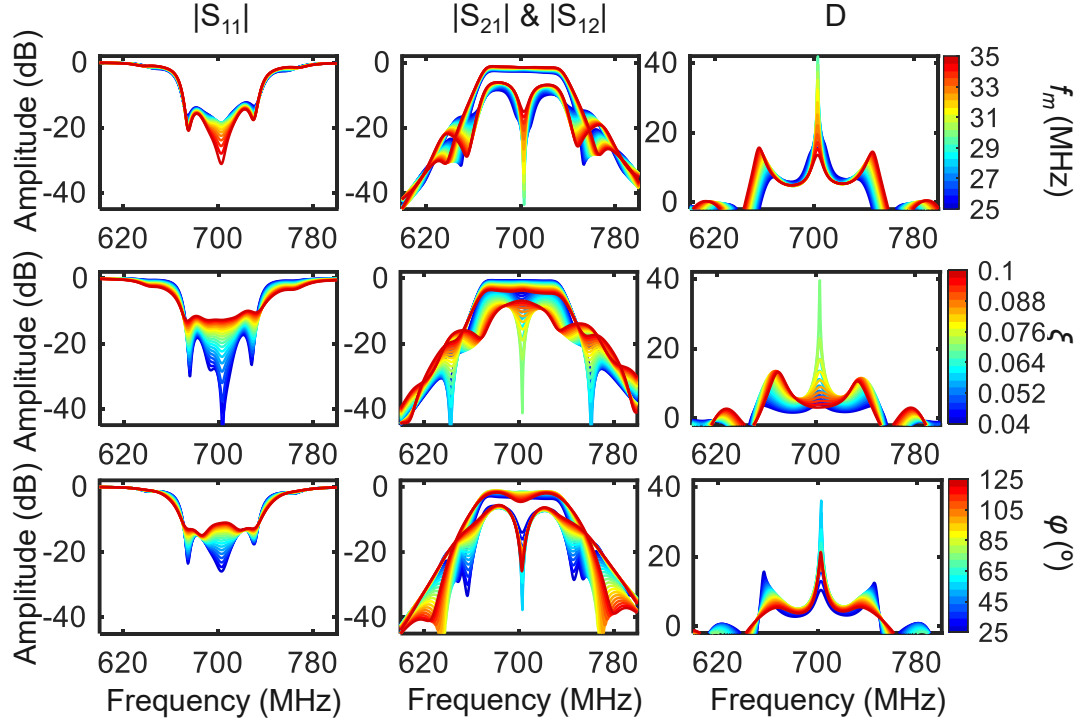
Directionality in the transversal-based BPFs can be achieved when the capacitance of the constituent resonators follows the sinusoidal variation described in (2.1). Depending on the selected modulation parameters ξ , f_m and φ , alternative S-parameter profiles can be obtained. Thus, parametric studies need to be performed to determine the desired transfer function characteristics as shown for the example cases of Topology 1 and Topology 4 in Figure 2.10 and Figure 2.11. Similar studies can be performed for the rest of the filtering topologies, but they have been omitted for brevity. In order to effectively present the parametric sweeping results, two data visualization methods have been adopted, namely line plots and heatmaps. These two complementary data visualization techniques provide a comprehensive view of the parametric sweeping results, and their characteristics are summarized as follows:

- **Line plots** : In Figure 2.10 (a) and Figure 2.11 (a), the amplitudes of the S-parameters ($|S_{11}|$, $|S_{21}|$, and $|S_{12}|$) along with the directivity ($D=|S_{21}|-|S_{12}|$) are plotted on the y-axis against the frequency on the x-axis. Curves representing different modulation parameters are plotted in the same coordinate system using

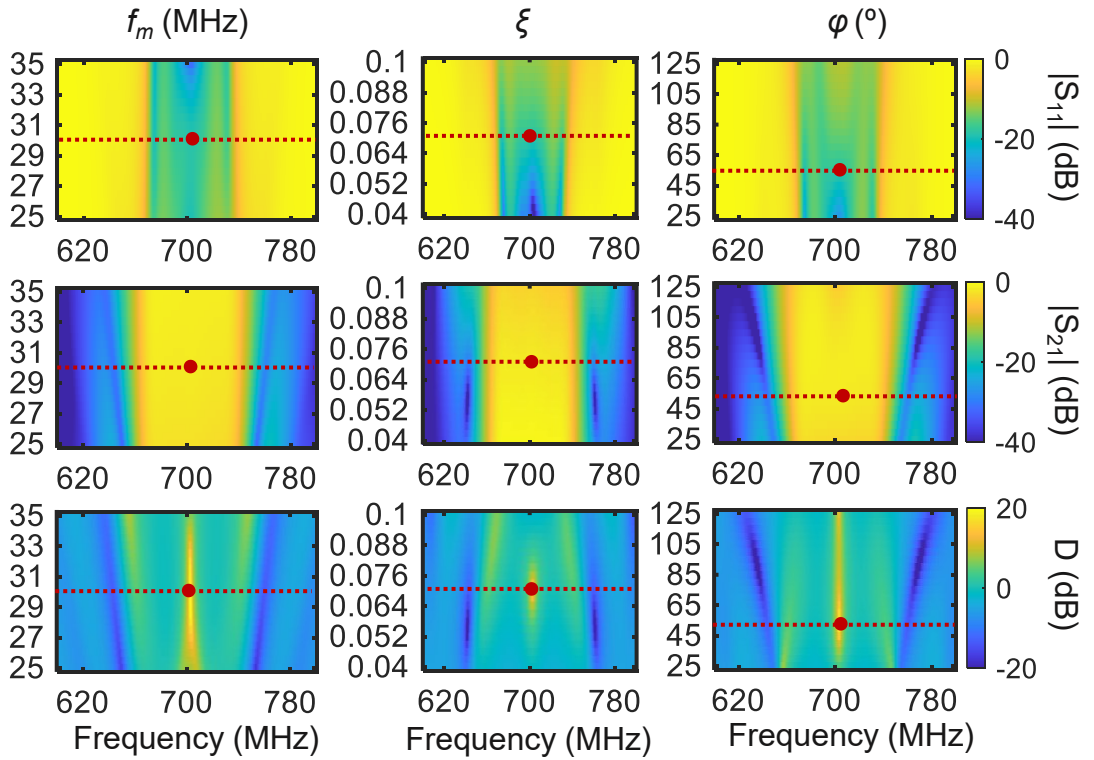
varying colors, as indicated by the color bar on the right side of each row. This visualization method allows for convenient observation of how the S-parameter shapes change with the modulation parameters.

- **Heatmaps:** Figure 2.10 (b) and Figure 2.11 (b) employ a heatmap approach to display the S-parameters as colors in a two-dimensional matrix. One axis represents the frequency range, while the other axis represents the modulation parameters. The color intensity in each cell indicates the value of the corresponding S-parameter.

Using as a basis these performance visualization methods, the Topology 1 performance trade-off metrics are provided in Figure 2.10. Specifically, a $f_m = 30$ MHz results in the highest D and an acceptable IL. Larger f_m values lead to lower IL but also lower D, while too small f_m values lead to high IL and low D as well. Regarding the modulation index ξ , the optimal value is chosen when $\xi = 0.07$. Although lower ξ values yield lower IL, the D is also lower. Too high ξ values cause both lower D and higher IL. Finally, an optimal value of 60° can be observed for φ , which achieves high D with low IL while maintain good transfer function shapes. Figure 2.11 illustrates the trade-offs analysis study for Topology 4 when operating in a dual-band mode of operation. The optimal modulation parameters of the dual-band BPF are chosen as $f_m = 20$ MHz, $\xi = 0.05$ and $\varphi = 65^\circ$ for high D for both bands while maintaining an acceptable IL and good transfer function shapes. Using as a reference these plots, optimal STM parameters can be selected for the desired S-parameters for Topology 1 and 4 and are demonstrated in Figure 2.4 (b) and Figure 2.9 (c), respectively.

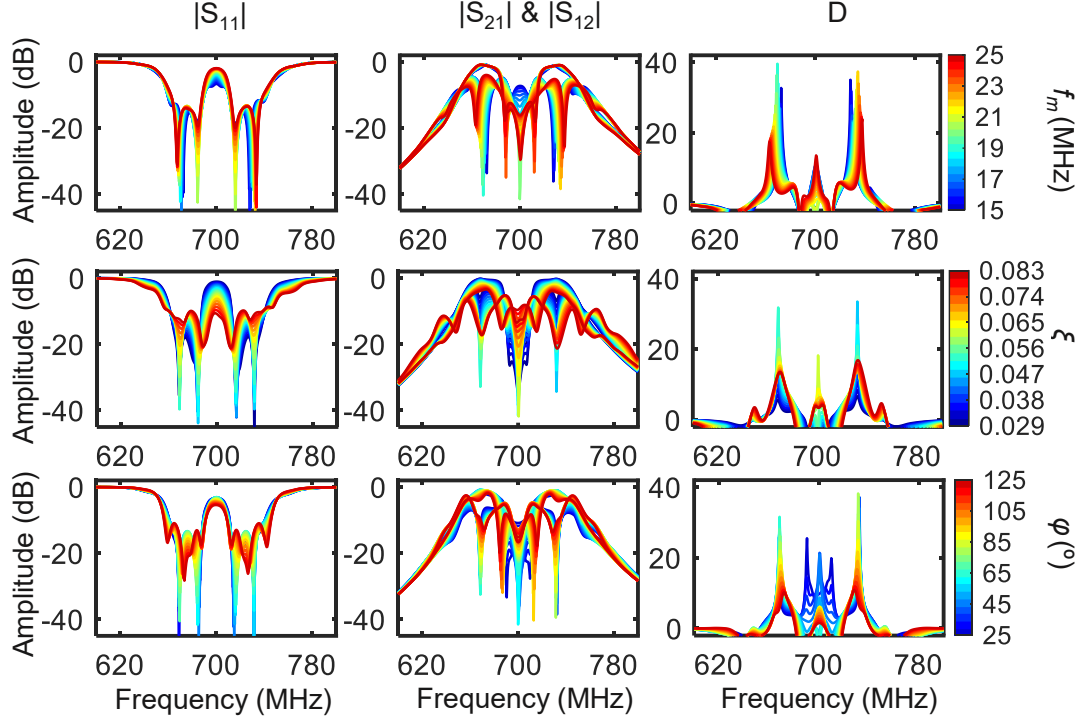


(a)

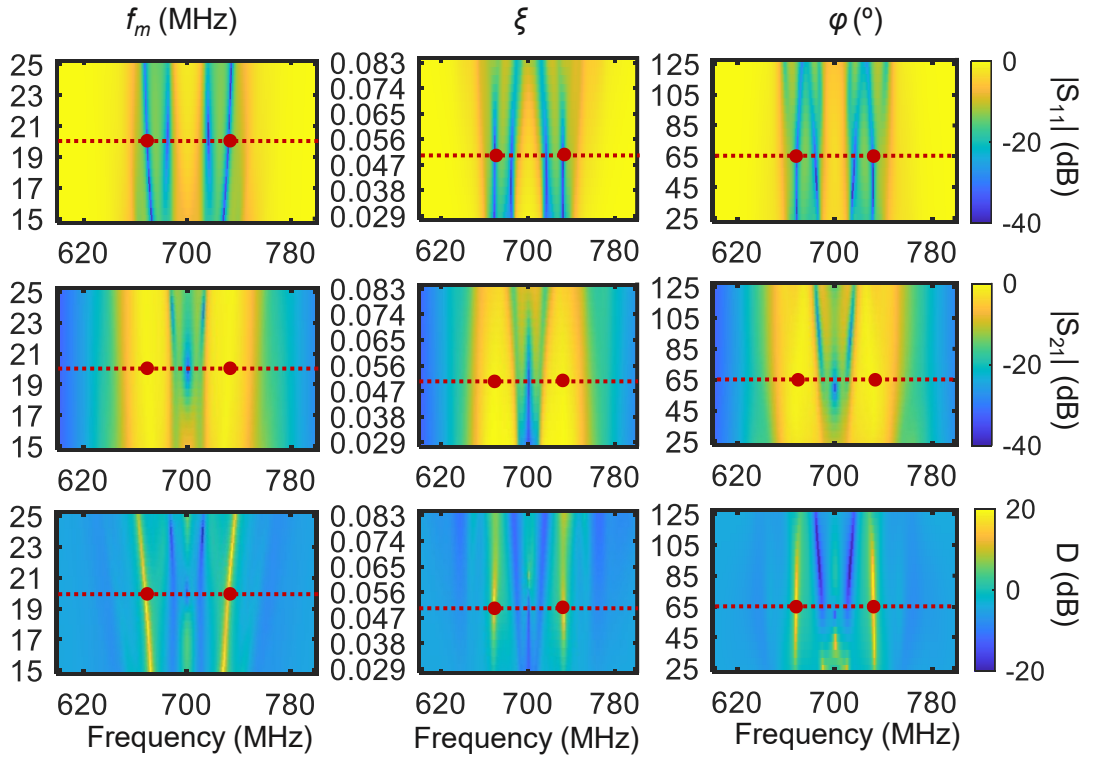


(b)

Figure 2.10. S-parameters of Topology 1 when varying f_m , ζ and φ . (a) Line plots. (b) Heatmap plots.



(a)



(b)

Figure 2.11. S-parameters of Topology 4 operating in the dual band mode when varying f_m , ζ and ϕ . (a) Line plots. (b) Heatmap plots.

2.2.2 Practical realization aspects

Based on the CRDs and the coupling coefficients provided in the previous sections, the four multi-functional filtering topologies can be implemented using LEs and conventional filter design techniques to specify the LE element values as for example the ones described in [128]. Next, the DC biasing on the varactor is selected so that it gives the desired resonator capacitance. Subsequently, the AC modulation parameters f_m , V_M , and ϕ are obtained through parametric simulations according to the desired IL in the forward direction and the desired IS in the reverse direction. Lastly, the tuning capabilities are obtained by altering the DC and AC biasing.

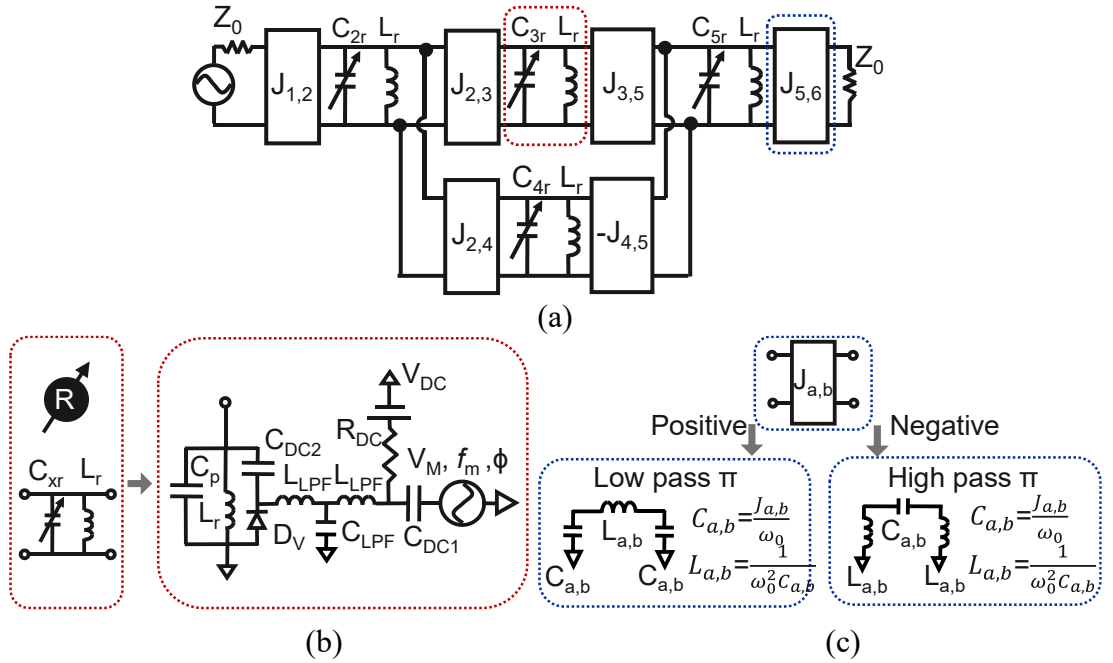


Figure 2.12. (a) Circuit equivalent of the CRD of Topology 2. (b) STM resonator implementation using lumped components. (c) Inverter implementation using lumped components.

Figure 2.12 shows a design example when implementing the CRD of Topology 2 in Figure 2.6 (a). Firstly, a reciprocal filter is designed based on the CRD and the coupling coefficients and the equivalent circuit is given in Figure 2.12 (a), where C_{xr} ($x=2,3,4,5$) is the capacitance L_r is the inductance of the resonators, and $J_{a,b}$ is the inverter value after impedance scaling and bandpass transformation to a center frequency of ω_0 and a fractional bandwidth of FBW. The values can be calculated using (2.16). The inductance of the resonators is set to the same value of L_r in all of the designing cases.

After acquiring all the parameters of the equivalent circuit in Figure 2.12 (a), the STM resonators and the static inverters can be implemented using lumped elements as shown in Figure 2.12 (b) and Figure 2.12 (c), respectively. Specifically, a parallel inductor L_r , a static capacitor C_r and a varactor diode D_v constitute the inductance and capacitance in each resonator. DC biasing V_{DC} is applied through a $10\text{ M}\Omega$ resistor R_{DC} to set the nominal resonant frequency of each resonator. All the resonators are modulated by low frequency AC signals with frequency f_m , amplitude V_m and phase ϕ so that the capacitance of the varactor satisfies the sinusoidal variation as the AC signal applied to it. A low-pass filter is added after each AC source to allow for the modulating AC signals to be applied on the resonators while protecting the AC source from the high frequency RF signals. C_{DC1} and C_{DC2} are used as DC blocks to prevent the DC signals from entering the AC sources and the RF circuit. Positive coupling coefficients are materialized with first order low-pass pi-networks whereas the negative ones with first order high-pass networks as shown in Figure 2.12 (c) using (2.17). To reduce the number of circuit elements required, the parallel capacitors of the inverters are combined with the ones of the adjacent resonators.

$$L_x = \frac{1}{\omega_0 \left(\frac{1}{FBW} - \frac{M_{xx}}{2} \right)}, C_x = \frac{\left(\frac{1}{FBW} + \frac{M_{xx}}{2} \right)}{\omega_0}, \omega_x = \frac{1}{\sqrt{L_x C_x}}$$

$$J_{1,2} = M_{1,2} \sqrt{\frac{L_2}{Z_0 L_r}}, J_{5,6} = M_{5,6} \sqrt{\frac{L_2}{Z_0 L_r}} \quad (2.16)$$

$$J_{a,b} = M_{a,b} \sqrt{\frac{L_a L_b}{L_r L_r}} \quad (a = 2, 3, 4).$$

$$C_{a,b} = \frac{J_{a,b}}{\omega_0}, L_{a,b} = \frac{1}{\omega_0^2 C_{a,b}} \quad (2.17)$$

To evaluate the effect of tolerances on the performance of the non-reciprocal RF filtering components, a sensitivity analysis can be conducted for the tolerances which are typically 2%. For the sake of illustration, the circuit architectures of Topology 1 and Topology 2 are analyzed through Monte Carlo simulations. The components marked in yellow in Figure 2.13 (a) and Figure 2.14 (a) are meant to be static and as such are chosen to have a 2% tolerance. The simulated results for Topology 1 and Topology 2 are presented in Figure 2.13 (b) and Figure 2.14 (b), respectively. The solid lines denote the simulation results when all the components have the nominal values while the shaded ones indicate how the S parameters vary with a 2% tolerance

for the chosen components in 1600 iterations. It should be noted that for all the 1600 iterations, the same modulation parameters f_m , ξ and φ are adopted. As shown, the effect of the tolerances is almost unnoticeable with the RL being always below 10 dB and isolation levels remaining >20 dB when STM is turned on.

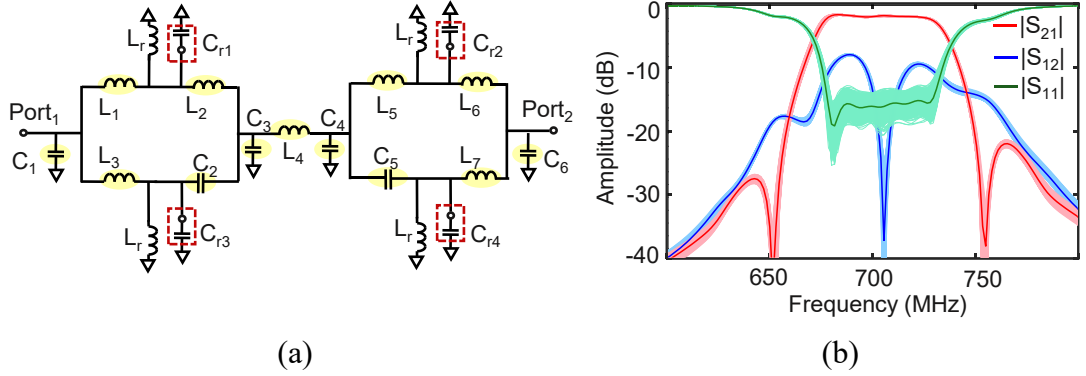


Figure 2.13. Sensitivity analysis of Topology 1. (a) Detailed circuit schematic. (b) Monte Carlo simulation results with 1600 iterations and 2% tolerance for the static components marked in yellow when STM is ON.

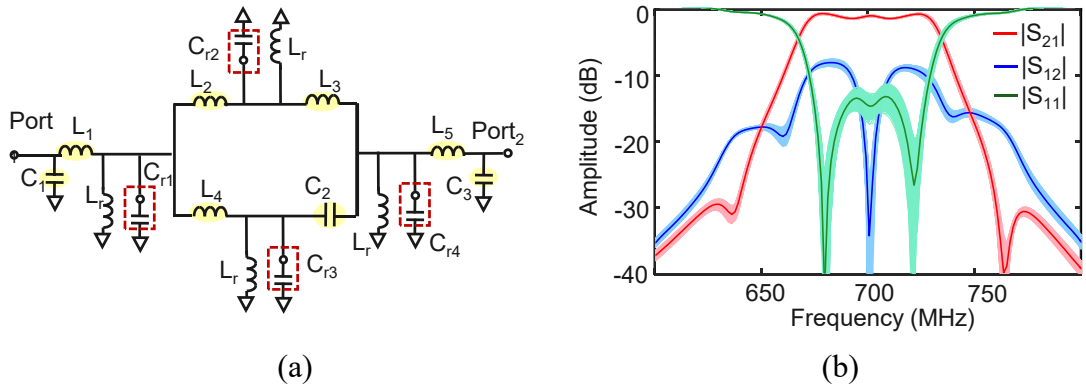


Figure 2.14. Sensitivity analysis of Topology 2. (a) Detailed circuit schematic. (b) Monte Carlo simulation results with 1600 iterations and 2% tolerance for the static components marked when STM is ON.

2.3 Experimental Validation

For proof-of-concept validation purposes of the transversal resonator-based BPF1 concept, four filter prototypes (Filters 1-4) implementing the CRD Topologies 1-4 were designed, and manufactured on a RO4003C dielectric substrate with a height of 1.5 mm. The S-parameters for the four filters, including their tuning capabilities, were characterized using a Keysight PNA-X N5244A network analyzer.

2.3.1 Filter 1: fourth-order cascaded transversal cells BPFI

The photograph of the manufactured prototype and the utilized lumped element components are provided in Figure 2.15 (a). A comparison of the RF measured and EM simulated S-parameters for one tuning state is shown in Figure 2.15 (b). They appear to be in a good agreement successfully validating the proposed concept. The measured 20 dB IS BW (the frequency range when $IS > 20$ dB) is 647.5 MHz to 659.5 MHz (12 MHz). The frequency band is marked in light purple. The RF measured tuning capabilities for Filter 1 are presented in Figure 2.16. Although a finite number of tuning states is shown, the filter exhibits continuous tuning range. Specifically, its center frequency can be tuned between 571 MHz to 724 MHz (1.27) with an IL ranging from 4.8 dB to 6.5 dB and IS between the forward and the backward direction from 21 dB to 33 dB. Figure 2.16 (b) demonstrates the BW tuning characteristics of the filter which exhibits a tunable BW between 39.5 MHz to 73.9 MHz (1.87) with IL ranging from 5.9 dB to 6.8 dB and IS between its forward and its backward direction between 25 dB to 35 dB. The intrinsic switch-off capability of Filter 1 is also demonstrated in Figure 2.16 (b) and is represented by the green traces in Case 4.

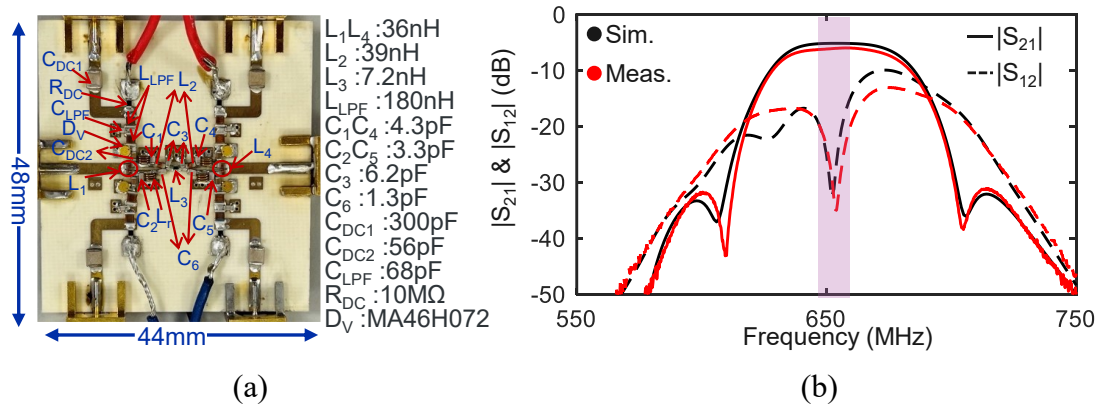


Figure 2.15. (a) Manufactured prototype of Filter 1 based on the Toplogy 1 CRD. (b) Comparison of RF-measured and EM-simulated S-parameters for one tuning state of Filter 1. The purple band marks the frequency range when $|S_{12}| < -20$ dB.

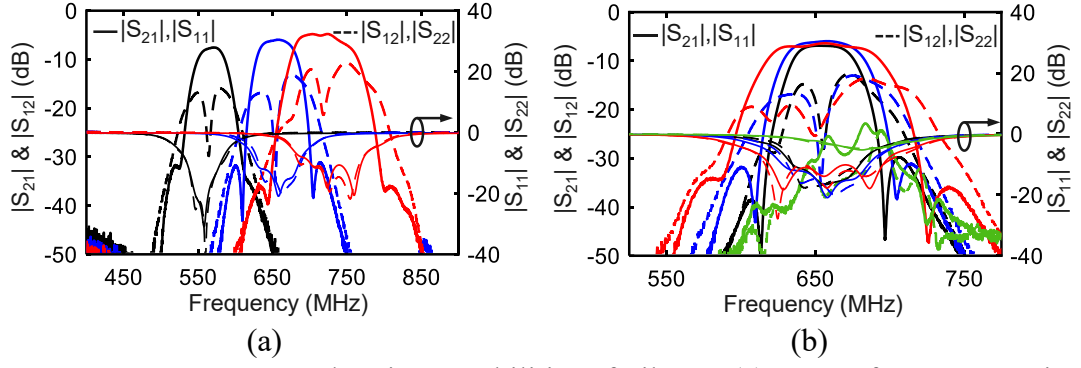


Figure 2.16. RF measured tuning capabilities of Filter 1. (a) Center frequency tuning. (b) BW tuning and intrinsic switching-off (green trace).

2.3.2 Filter 2: fourth-order box-like BPF

Figure 2.17 (a) depicts the photograph of the manufactured prototype of Filter 2 and a comparison between the RF-measured and the EM-simulated S-parameters for one tuning state are presented in Figure 2.17 (b). For this specific tuning state, the measured 20 dB IS BW is 667 MHz to 705 MHz (38.4 MHz). As can be seen from Figure 2.18 (a), the center frequency of Filter 2 can be tuned between 630 MHz to 735 MHz (1.17). For these states IL varies between 4.8 dB to 6.7 dB and the IS between the forward and the backward direction between 20 dB and 30 dB. Continuous BW tuning between 24.8 MHz to 58.1 MHz (2.34) is shown in Figure 2.18 (b). For these states, IL varies between 5.4-6.8 dB and the IS between the forward and the backward direction between 29-40 dB. Filter 2 can also be switched-off intrinsically as shown in Case 4 state in Figure 2.18 (b).

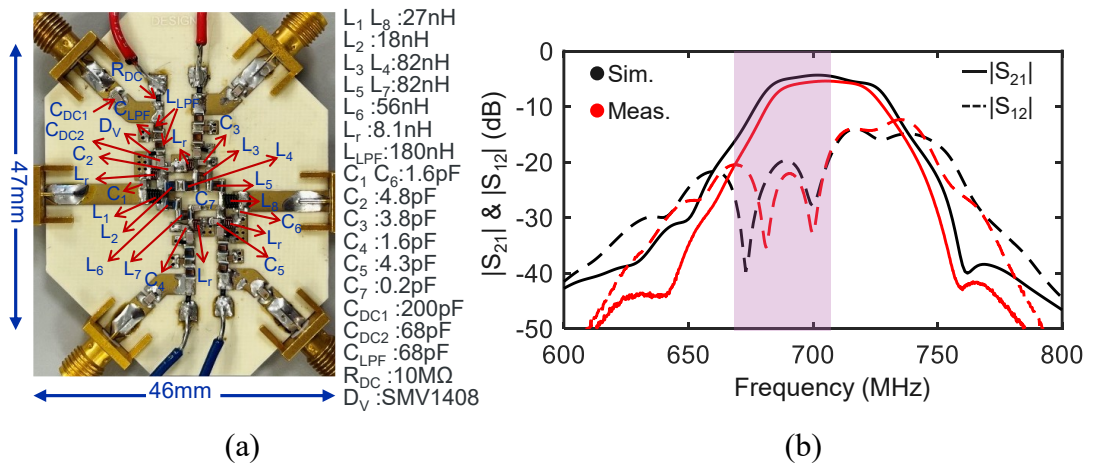


Figure 2.17. (a) Manufactured prototype of Filter 2 based on the Topology 2 CRD. (b) Comparison of RF measured and EM-simulated S-parameters for one tuning state of Filter 2. The purple band marks the frequency range when $|S_{12}| < -20$ dB.

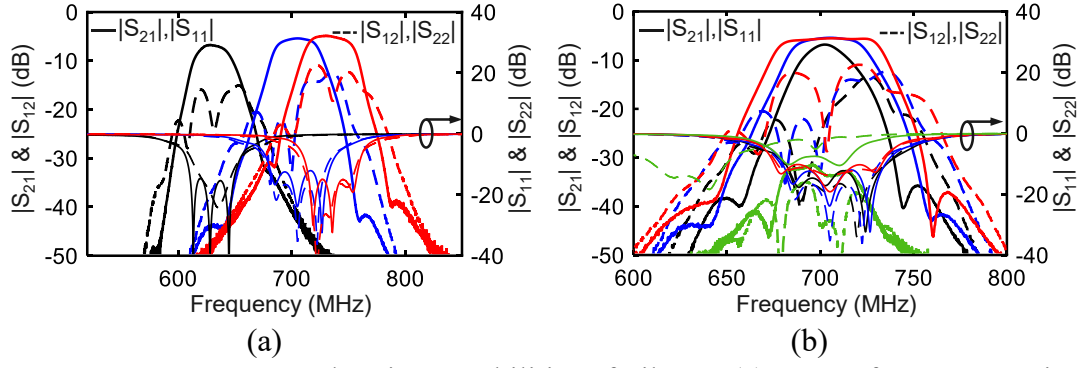


Figure 2.18. RF measured tuning capabilities of Filter 2. (a) Center frequency tuning. (b) BW tuning and intrinsic switching-off (green trace).

2.3.3 Filter 3: six-order box-like BPF

The experimental prototype of Filter 3 is provided in Figure 2.19 (a). Figure 2.19 (b) showcases the good agreement between its EM simulated and RF measured performance for one tuning state further supporting the effectiveness of the transversal resonator-based BPF concept. For this tuning state in Figure 2.19 (b), the measured 18 dB IS BW is 670 MHz to 718 MHz (48 MHz). As shown in Figure 2.20 (a), Filter 3 allows for center frequency tuning between 655 MHz and 760 MHz (1.16). For these states, the in-band passband IL varies between 4.5 dB and 5.7 dB and the IS between the forward and the backward direction between 30 dB and 45 dB. As illustrated in Figure 2.20 (b), the BW tuning range of Filter 3 was measured between 52.9 MHz and 84 MHz (1.57), with IL ranging from 4.9 MHz to 6.7 dB and IS from 27 dB to 34 dB. Filter 3 can also be intrinsically switched-off as demonstrated in Case 4 in Figure 2.20.

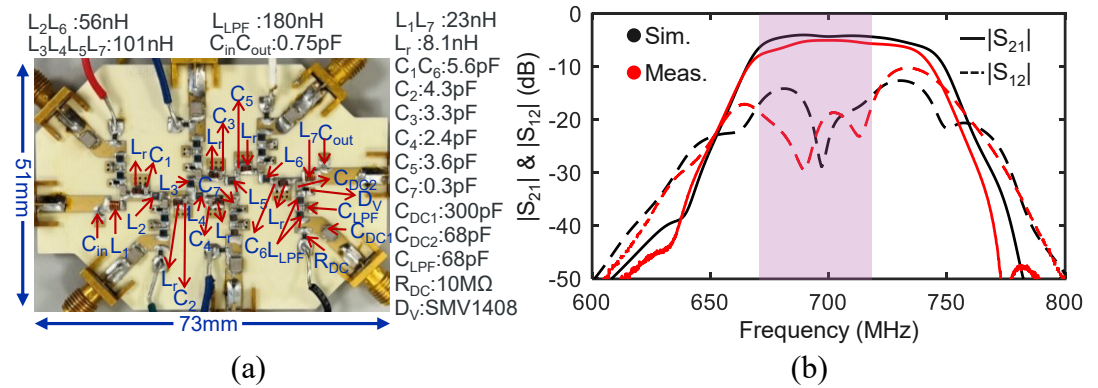


Figure 2.19. (a) Manufactured prototype of Filter 3 based on the Topology 3 CRD. (b) Comparison of RF measured and EM-simulated S-parameters for Filter 3. The purple band marks the frequency range when $|S_{12}| < -18$ dB.

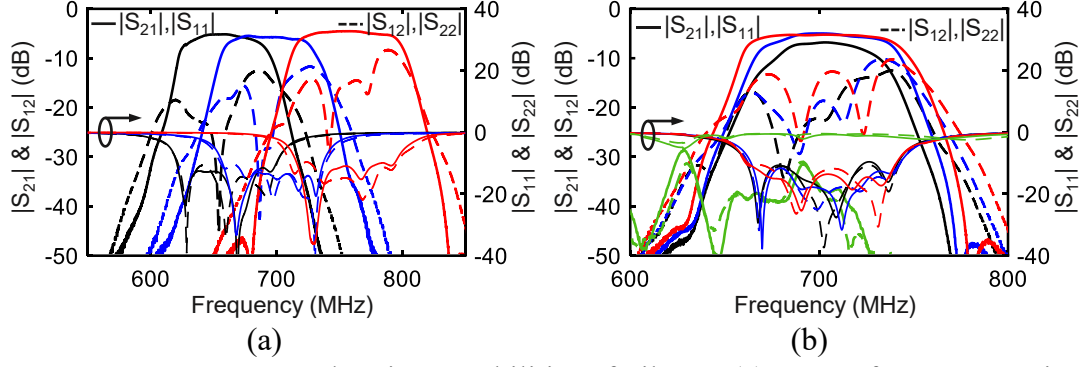


Figure 2.20. RF measured tuning capabilities of Filter 3. (a) Center frequency tuning. (b) BW tuning and intrinsic switching-off (green trace).

2.3.4 Filter 4: multi-configurable cascaded transversal cell BPF

Figure 2.21 shows the manufactured prototype of the multi-configurable Filter 4. Figure 2.22 (a) and Figure 2.22 (b) provide the comparison between the RF-measured and the EM-simulated S-parameters when operating at its dual-band and its single-band mode of operation which appear to be in a good agreement. The measured 20 dB IS BW for the dual-band mode is 653 MHz to 657 MHz (4 MHz) for the lower band and 733 MHz to 740 MHz (7 MHz) for the upper band. The measured 20 dB IS BW for the single-band mode is 681 MHz to 691 MHz (10 MHz).

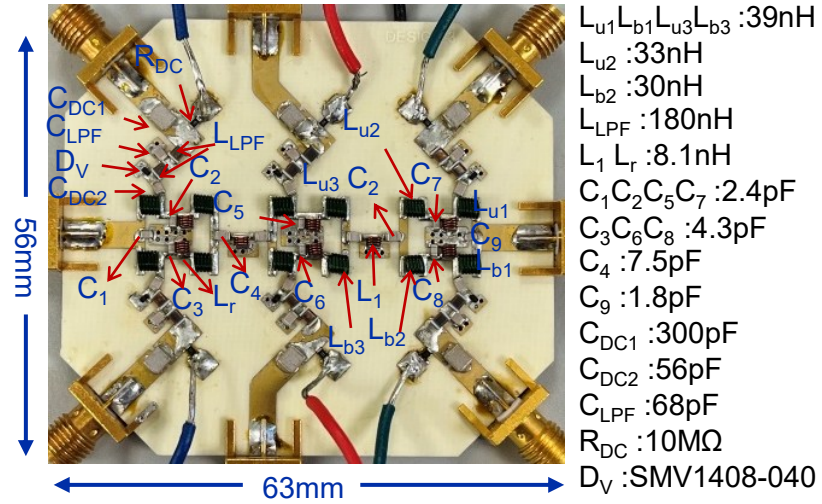


Figure 2.21. Manufactured prototype of Filter 4 based on the Topology 4 CRD.

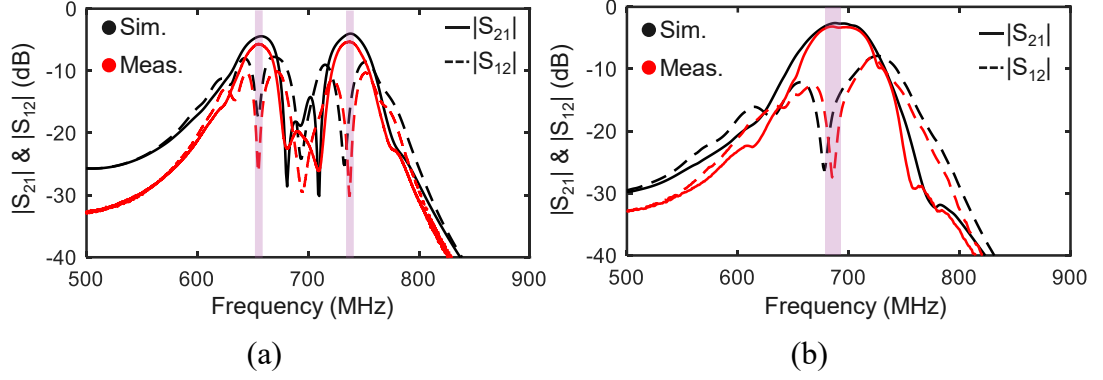


Figure 2.22. Comparison of RF-measured and EM-simulated S-parameters for two example tuning states of Filter 4. The purple band marks the frequency range when $|S_{12}| < -20$ dB. (a) Dual-band state. (b) Single-band state.

The tuning capabilities of Filter 4 are summarized in Figure 2.23 to Figure 2.27, and show multiple levels of reconfigurability. Specifically, when Filter 4 operates in its single-band mode of operation, the out-of-band response can be reconfigured to have three TZs, two TZs, one TZ and no TZ as shown in Figure 2.23 (a)-(d). The center frequency of the four states are 700 MHz and the passband BW is 26 MHz, 34 MHz, 44 MHz and 56 MHz respectively and their in-band IL is 4.46 dB, 3.55 dB, 3.46 dB and 3.2 dB. The measured TFs demonstrate that $IS > 25$ dB can be obtained for all the tuning states shown in Figure 2.23. Figure 2.24 shows center frequency tuning when the filter is operating as a single band BPF. Its passband can be tuned from 659-724 MHz with IL between 3.6-4.1 dB and $IS > 20$ dB.

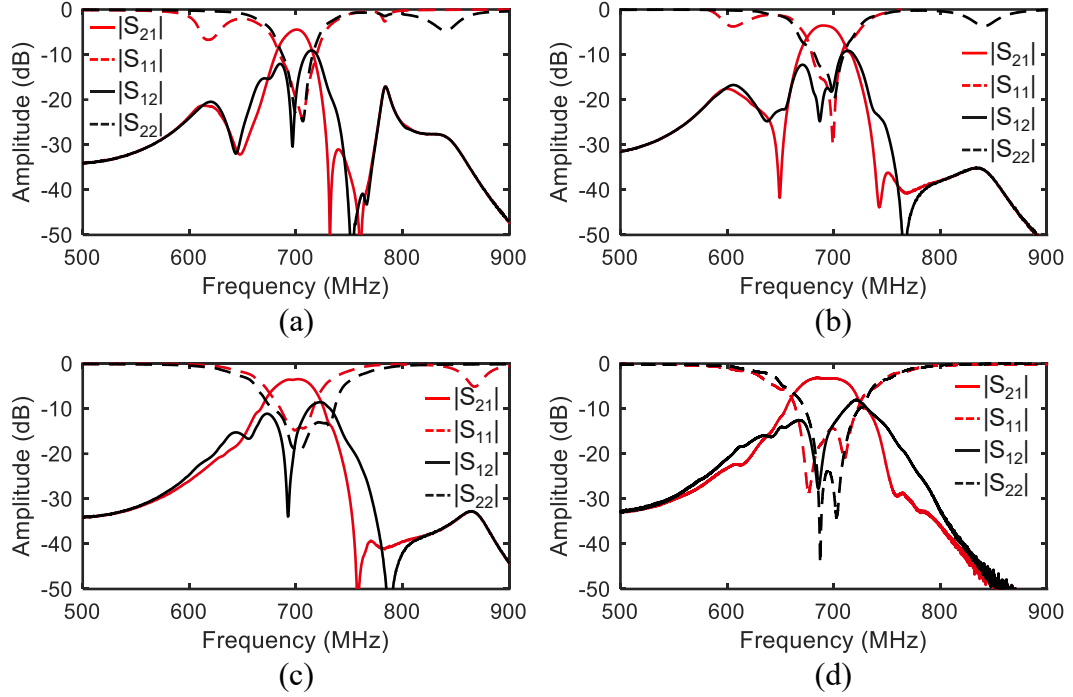


Figure 2.23. RF measured TZ reconfiguring capabilities of Filter 4 when operating at its single-band mode. (a) Three TZs. (b) Two TZs. (c) One TZ. (d) No TZ.

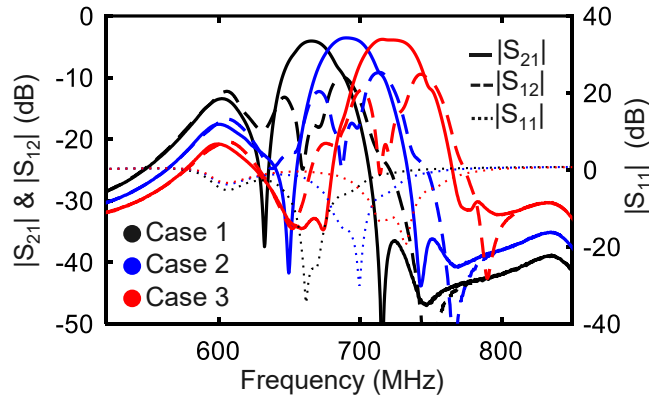


Figure 2.24. RF measured f_{cen} tuning capabilities of Filter 4 when operating at its single-band mode.

The dual band state tuning capabilities are demonstrated in Figure 2.25 and Figure 2.26. Specifically, when reconfiguring the sign of the phase progression of the resonators in each band, the direction of propagation can be tuned as shown in Figure 2.25 (a) and (b) by altering θ from 65° to -65° . Furthermore, Figure 2.26 (a) and (b) show how the two passbands can be tuned independently. Specifically, the lower passband can be tuned between 600-680 MHz with IL between 5.6-6 dB and IS between 23-26 dB. The upper band can be tuned between 716-831 MHz with IL

between 4.8-6.2 dB and IS between 25-31 dB. Furthermore, Filter 4 can be intrinsically switched-off as demonstrated in Figure 2.27. RL for all prototypes is better than 10 dB throughout the entire tuning range.

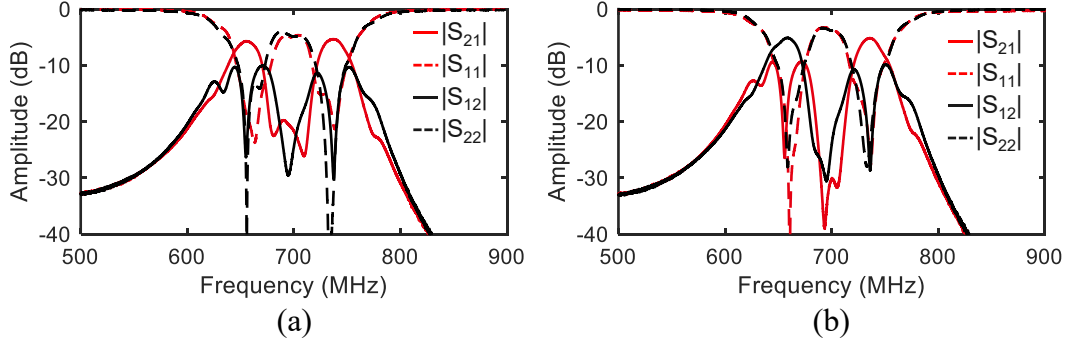


Figure 2.25. RF measured reconfigurable capabilities of Filter 4 when operating at its dual-band mode. (a) Both bands transmit in the same direction. (b) Each band is transmitting in a different direction.

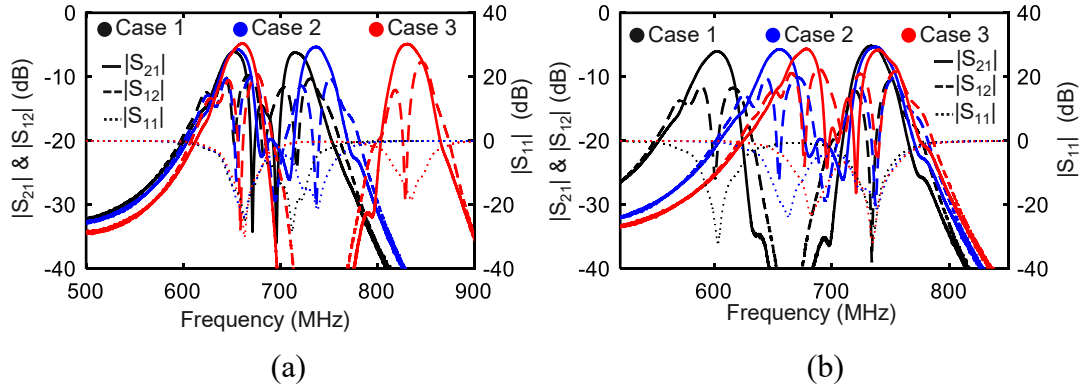


Figure 2.26. RF measured f_{cen} tuning capabilities of Filter 4 when operating at its dual-band mode of operation. (a) Upper band f_{cen} tuning. (b) Lower band f_{cen} tuning.

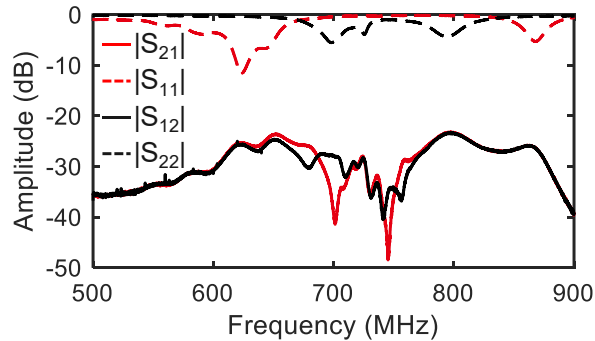


Figure 2.27. RF measured intrinsic switching off tuning state of Filter 4.

A comparison of the transversal-cell-based BPF concept with state-of-the-art STM-based non-reciprocal BPFs is provided in Table 2.2. As shown, Filters 1, 2, and 3 showcase competitive performance in terms of tuning range, isolation, and RL when compared to the state-of-the-art. Filters 1 to 3 also exhibit higher order when compared to many existing STM-based BPFs [54], [71]–[73], [74]–[78], having higher selectivity and out-of-band rejection. Moreover, Filter 4 shows the highest levels of reconfigurability among all the existing BPFs. It can support both single- and dual-band transfer functions and offers independent tuning of two passbands. Moreover, it can reconfigure its out-of-band response with various TZ configurations, providing improved flexibility in adapting to different system requirements and operating conditions. Furthermore, the proposed filters can be intrinsically switched-off, and can be exploited to reduce the number of RF switches in the RF front-end.

Table 2.2. Comparison with the state-of-the-art BPFs

Ref.	f_{cen} MHz	BW MHz	IL dB	IS dB	IIP3 dBm	OFF
[66]	270~310 (1.14)	15~41.5 (2.8)	1.7~4.3	6.9~30.9	N/A	Yes
[67]	188~202 (1.07) 237~257 (1.08)	14.5, 14	3.7, 5.5 (min)	26.1, 35.9 (max)	N/A	No
[68]	198~224 (1.13) 249~291 (1.17)	13, 21.8	2.9, 2.6 (min)	8.4~44.8, 8~44.3	N/A	No
[50]	900	57	4.4	18	N/A	No
[73]	885~1031 (1.16)	42	3.9~4.6	20~30	11.8	No
Filter 1	570~724 (1.27)	37.5~73.9 (1.87)	4.8~6.8	21~35	-1~13	Yes
Filter 2	630~735 (1.17)	24.8~58 (2.34)	4.8~6.8	20~40	-4~13	Yes
Filter 3	655~760 (1.16)	53~84 (1.57)	4.5~6.7	27~45	0~13	Yes
Filter 4 (Single-band)	665~723 (1.09)	26, 34, 44, 56	3.2~4.5	>25	6.4~10.3	Yes
Filter 4 (Dual-band)	600~680 (1.13), 716~831 (1.16)	25, 29	5.6~6, 4.8~6.2	23~26, 25~31	0~10, 5~13	Yes

2.4 Summary

This chapter reported on a comprehensive design methodology for reconfigurable BPFs using STM transversal cells. The proposed method was applied to the design of four filter topologies and demonstrated the potential to achieve single- and multi-band unilateral transfer functions with enhanced reconfigurability. Multi-levels of RF tuning can be obtained for the four topologies using a simple RF tuning mechanism where only the resonant frequencies of the resonators are altered. Four proof-of-concept prototypes were manufactured and measured and validated the effectiveness of the transversal cell-based BPF concept in achieving wider tuning ranges, high isolation levels, and superior reconfigurability compared to state-of-the-art STM-based BPFs. These results underline the potential of using STM transversal cells in designing multi-functional RF components for emerging RF communication systems.

Chapter 3: Multi-functional STM BPFs with Integrated Variable Phase Shifter and Attenuator Functionalities

This chapter introduces a novel RF co-design methodology that allows to co-integrate the functionality of a single-/multi-band BPF, an RF switch, a VP, a VA, and an RF isolator within a single multi-functional RF component. It is based on a power reconfigurable rat-race coupler, STM resonators and a switchable phase coupling element that controls three modes of operation. These include: (i) a co-designed single/multi-band reflectionless BPF, an RF isolator (enabled/disabled by turning ON/OFF STM) and a 360°-variable phase shifter (BPF/BPFI-VP), (ii) an RF switching mode and, (iii) a single/multi-band BPF, an RF isolator (enabled/disabled by turning ON/OFF STM) and a variable attenuator (BPF/BPFI-VA). A detailed theoretical framework is provided alongside a variety of single-band and multi-band design examples. The methodology is validated through the implementation of a single-band prototype (Prototype 1) and a dual-band prototype (Prototype 2) at 700 MHz. The multi-functional RF devices and contributions of this chapter have been presented in [124], [125], [126].

3.1 Theoretical Foundations

The block diagram of the proposed multi-functional RF component concept is illustrated in Figure 3.1 alongside the conceptual S-parameters of its three reconfigurable modes of operation. It consists of two power dividers, each followed by two coupling elements with static phase (φ_1) or reconfigurable phase (φ_2) and an RFRC as shown in Figure 3.1 (a). The multi-functional RF component functionalizes different modes of operation which can be obtained as follows:

(i) When $\varphi_2 - \varphi_1 = 90^\circ$, the BPF/BPFI-VP mode can be obtained while having a symmetric-input-reflectionless response as shown in the conceptual S-parameters in Figure 3.1 (b). In this case, the insertion phase can be tuned between 0°–360° by altering the RFRC's coupling strength.

(ii) When $\varphi_2 - \varphi_1 = 90^\circ$, the couplings between P₁, P₄ and P₂, P₃ inside the RFRC are set to zero while the couplings between P₁, P₃ and P₂, P₄ are the same, the multi-functional component operates in the switching-off mode of operation as indicated in Figure 3.1 (c).

(iii) When $\varphi_2 - \varphi_1 = 0^\circ$, the multi-functional component operates in the BPF/BPFI-VA mode and can achieve an attenuation range from 0 to ∞ dB, as demonstrated in Figure 3.1 (d).

(iv) For both the BPFI-VP and the BPFI-VA modes of operation, the isolator functionality can be enabled/disabled by turning ON/OFF the STM to reconfigure the signal transmission mode from uni-directional to bi-directional.

The characteristics of each of the elements of the multi-functional RF component alongside the operating principles of each mode of operation are detailed in the next sections through analytical equations and a wide range of design examples.

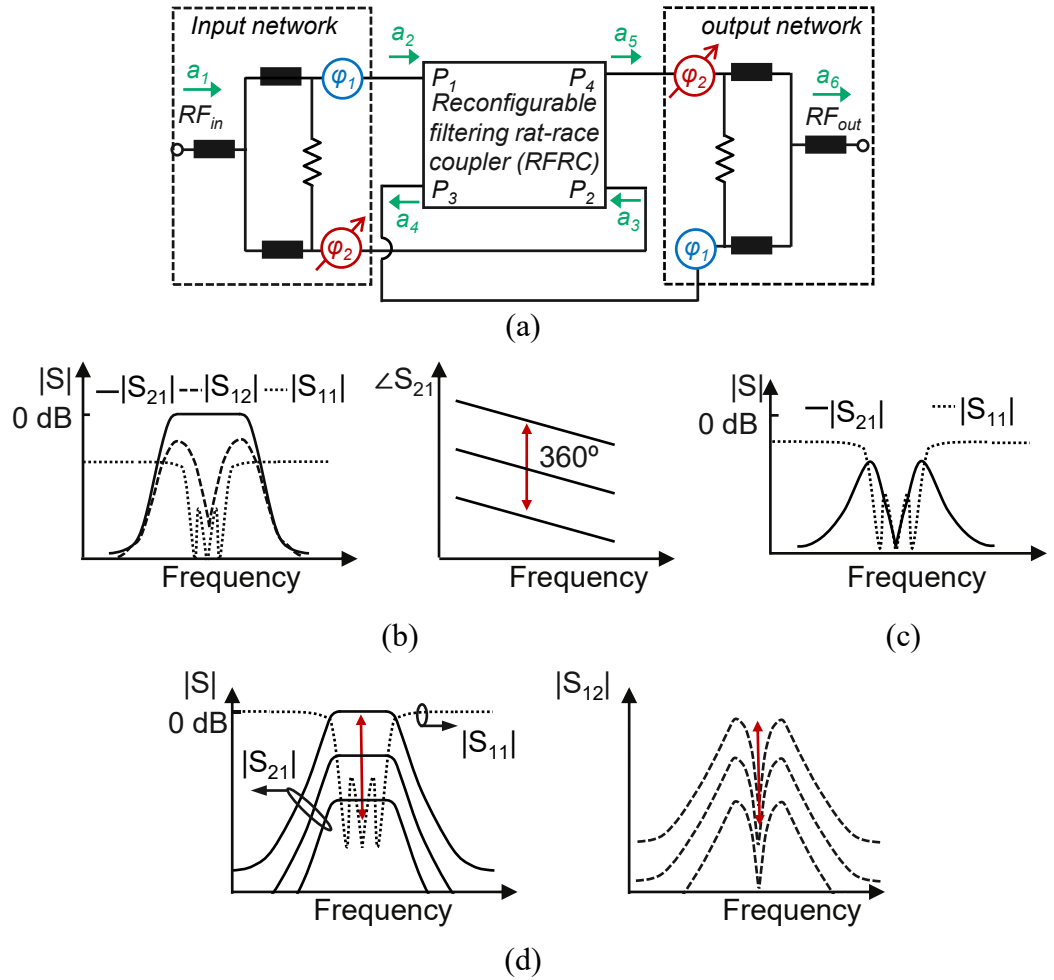


Figure 3.1. Block diagram of the proposed multi-functional RF component with three reconfigurable modes of operation: 1) a BPF/BPFI-VP mode, 2) an intrinsically-switched-off mode and 3) a BPF/BPFI-VA mode. Reconfigurability between the three modes is achieved through the phase variation couplings φ_1 and φ_2 as well as the couplings in RFRC. (b) Reflectionless BPF/BPFI-VP mode of operation when $\varphi_2 - \varphi_1 = 90^\circ$ with a 0-360° of phase shift by altering the coupling ratio of the RFRC. (c)

Intrinsic switching-off mode of operation when $\varphi_2 - \varphi_1 = 90^\circ$ and the couplings between P₁, P₄ and P₂, P₃ are zero while the couplings between P₁, P₃ and P₂, P₄ are the same. (d) BPF/BPFI-VA mode of operation when $\varphi_2 - \varphi_1 = 0^\circ$ with a continuously tunable attenuation level between 0 and ∞ dB by altering the coupling ratio of the RFRC.

3.1.1 Reconfigurable filtering rat-race coupler (RFRC)

Figure 3.2 (a) demonstrates the CRD of the RFRC which is the most important building block of the multi-functional RF component. It consists of four synchronously tuned resonators R₁ to R₄—denoted by black circles. The resonators are coupled through reconfigurable couplings, denoted as xM_m and yM_n , with x and y indicating the coupling polarity and are limited to values of +1 or -1. M_l indicates the external coupling which is connected to the phase coupling elements φ_1 and φ_2 . Due to the RFRC being symmetrical, it can be analyzed using even and odd mode analysis as in [129], [130] which leads to the S-parameters that outlined in equations (3.1) to (3.4). As shown in (3.2) and (3.3), M_m and M_n determine the output power distribution and this four-port network functions as a rat-race coupler, characterized by a power division ratio (k), which is described in (3.5). Specifically, the RF power from P₁ is divided based on k towards P₃ and P₄, while P₂ remains isolated.

$$S_{11} = S_{22} = S_{33} = S_{44} = \frac{s^2 + M_m^2 + M_n^2 - M_1^4}{s^2 + 2sM_1^2 + M_m^2 + M_n^2 + M_1^4} \quad (3.1)$$

$$S_{14} = S_{41} = S_{23} = S_{32} = \frac{2jM_1^2M_m}{s^2 + 2sM_1^2 + M_m^2 + M_n^2 + M_1^4} \quad (3.2)$$

$$S_{13} = S_{31} = -S_{24} = -S_{42} = \frac{2jM_1^2M_n}{s^2 + 2sM_1^2 + M_m^2 + M_n^2 + M_1^4} \quad (3.3)$$

$$S_{12} = S_{21} = S_{34} = S_{43} = 0 \quad (3.4)$$

$$k = \frac{M_m^2}{M_n^2} \quad (3.5)$$

$$M_m^2 + M_n^2 = M_c^2 \quad (3.6)$$

From a filtering perspective, the RFRC facilitates the realization of a second-order filtering TF, whose characteristics can be determined by comparing the polynomial expressions in (3.1)–(3.3) with those of a second-order BPF. Setting $M_m^2 + M_n^2$ to a constant value M_c^2 as defined in (3.6) ensures that the input reflection coefficient (S_{11}) and the denominators of S_{41} and S_{31} stay constant across variations of

k , thus maintaining the TF shape while allowing to alter the power division ratio between P_3 , P_4 whose response can be considered equivalent to the one obtained by the second-order CRD in Figure 3.2 (b).

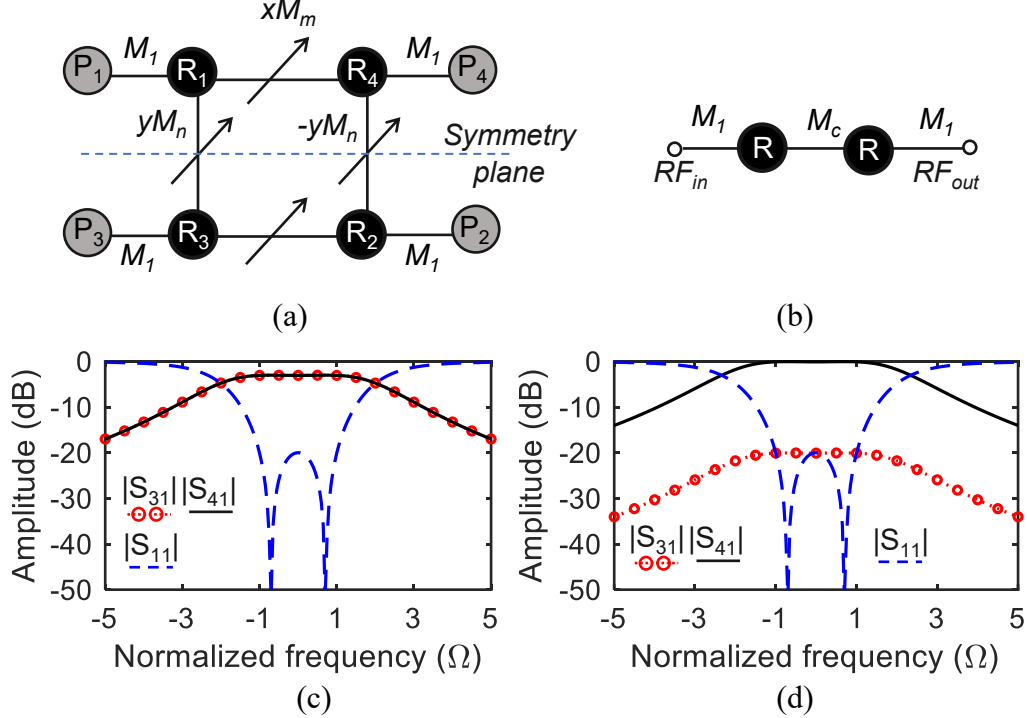


Figure 3.2. (a) CRD of a RFRC functionalizing second-order TFs at its output ports P_3 , P_4 . (b) CRD of the equivalent filtering network for each of its RF signal propagation paths. (c) Synthesized scattering parameters of the RFRC when $k=1$ and $M_l=1.226$, $M_c=1.662$, $M_m=M_n=1.1752$. (d) Synthesized scattering parameters of the RFRC when $k=100$ and $M_l=1.226$, $M_c=1.662$, $M_m=1.6538$, $M_n=0.1654$.

To demonstrate this functionality, the RFRC has been designed for a second order Chebyshev TF with a 0.043 dB ripple level (RL of 20 dB) and its responses are shown in Figure 3.2 (c), (d) for different $k=1$ and 100, respectively. They have been obtained by calculating M_l, M_2, M_c of the CRD in Figure 3.2 (b) as in any conventional second-order BPF as described in [128]. Then M_n and M_m are calculated from (3.5) and (3.6). As indicated in Figure 3.2 (c), the input power is equally coupled to P_3 and P_4 when k equals to 1. When $k=100$, the power delivered to P_4 is 20 dB larger than that of P_3 as indicated in Figure 3.2 (d). Notably, second order Chebyshev responses are formed for both cases and the RL remains the same when varying k .

The RFRC can be scaled to accommodate higher-order, quasi-elliptic as well as multi-band TFs as it is further illustrated in Figure 3.3 to Figure 3.6 through its corresponding CRDs in Figure 3.3 (a), Figure 3.3 (c), Figure 3.4 (a), Figure 3.5 (a)

and Figure 3.6 (a). By applying the same design methodology to these configurations and utilizing the coupling coefficients specified in the captions, various types of TFs can be realized, as demonstrated in Figure 3.3 (b), Figure 3.3 (d), Figure 3.4 (b), Figure 3.5 (b) and Figure 3.6 (b). Specifically, the topology depicted in Figure 3.3 illustrates a method for scaling the filter order, showcasing how third-order and fourth-order TFs can be obtained. Additionally, Figure 3.4 presents a method for implementing a quasi-elliptic transfer function, highlighting the flexibility of the proposed structure. In these cases, all the black resonators are resonating at the same frequency and create the poles of the TF. The blue and red resonating nodes, along with the grey non-resonating node, form a multi-resonant cell. This cell introduces two TZs and one additional pole. The TZs are located at the resonant frequencies of the blue and red nodes. The additional pole occurs at the frequency where the two paths in the multi-resonant cell exhibit equal magnitude but opposite sign admittances [66].

To showcase the versatility and practicality of the proposed synthesis approach, two distinct CRDs commonly employed for the realization of multi-band TFs are explored in Figure 3.5 and Figure 3.6. The CRD in Figure 3.5 is based on a split-type resonator approach [131], whereas the one in Figure 3.6 is using multi-resonant resonators [132]. Whereas both topologies lead to identical multi-band responses, the CRD in Figure 3.5 based on split-type resonators uses less elements, whereas the one based on multi-resonant resonators in Figure 3.6 allows for bands that can be tuned independently. Notably, in all the synthesized instances for these five different topologies shown in Figure 3.3 to Figure 3.6, the desired responses are achieved with well-preserved transfer function at the output ports and $|S_{11}|$ remaining constant as k is adjusted. This characteristic facilitates the constant TF characteristics of the proposed multi-functional RF component when its phase or amplitude is being tuned as it will be discussed in the next sections.

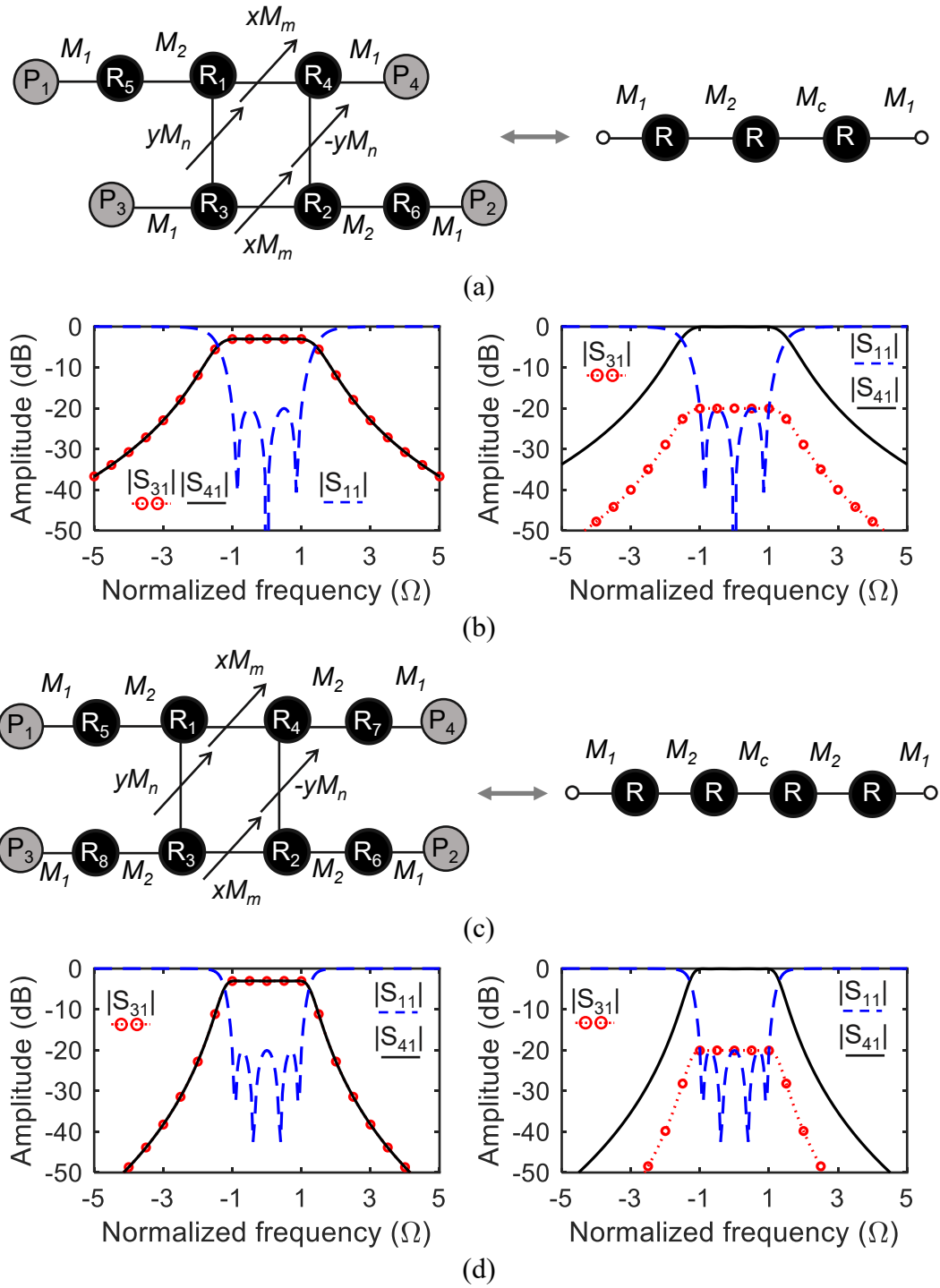


Figure 3.3. Order scalability of the proposed RFRC. (a) CRD of a RFRC functionalizing third-order TFs at its output ports P_3 , P_4 and its equivalent filtering CRD for each of its RF signal propagation paths. (b) Synthesized scattering parameters of the RFRC when $k=1$ or 100 . When $k=1$: $M_m=M_n=0.7295$. When $k=100$: $M_m=1.0266$, $M_n=0.1027$. The rest of the coupling coefficients are: $M_l=1.083663$, $M_2=M_c=1.0317$. (c) CRD of a RFRC functionalizing fourth-order TFs at its output ports P_3 , P_4 and its equivalent filtering CRD for each of its RF signal propagation paths. (d) Synthesized scattering parameters of the RFRC when $k=1$ or 100 . When $k=1$: $M_m=M_n=0.4953$. When $k=100$: $M_m=0.697$, $M_n=0.0697$. The rest of the coupling coefficients are: $M_l=1.0362$, $M_2=0.9116$, $M_c=0.7005$.

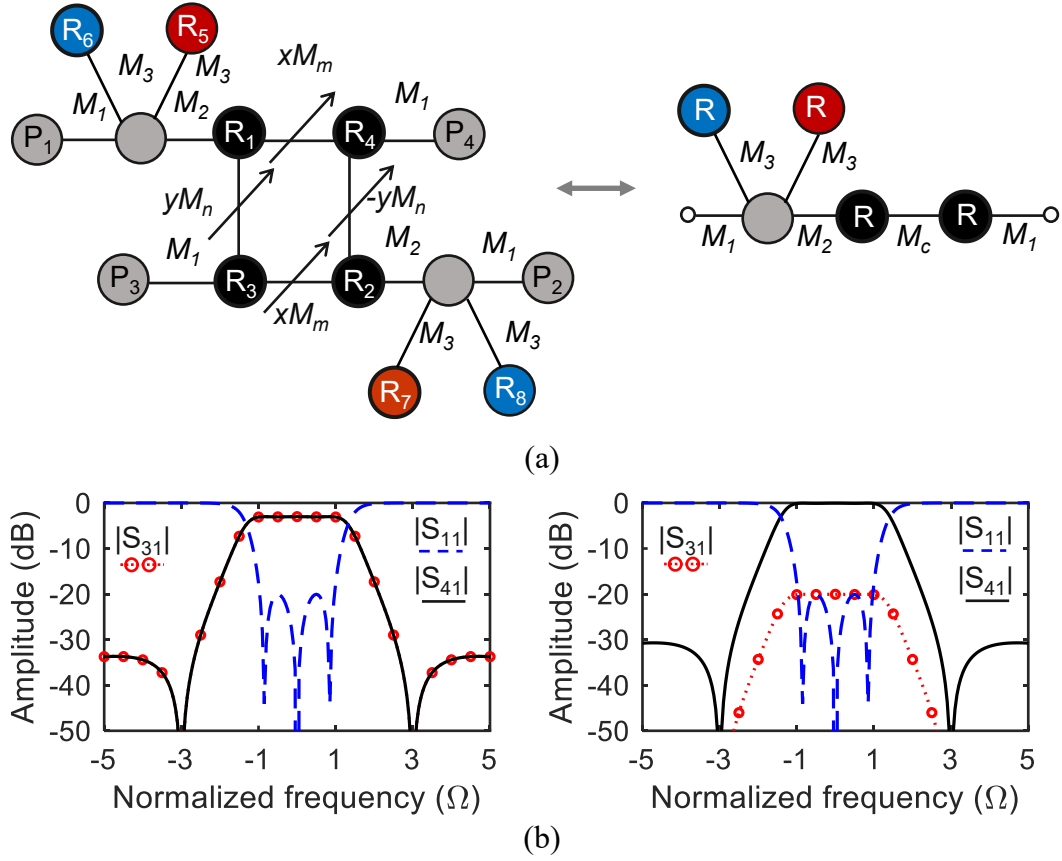


Figure 3.4. CRD of the RFRC functionalizing quasi-elliptic third order TFs with two zero at its output ports P_3 , P_4 and its equivalent filtering CRD for each of its RF signal propagation paths. The nodes in the same color have the same self-resonating coefficients. The grey nodes are non-resonating nodes. (b) Synthesized scattering parameters of the RFRC for $k=1, 100$. When $k=1$, $M_m=M_n=0.6998$. When $k=100$, $M_m=0.9848$, $M_n=0.0985$. The rest of the coupling coefficients are: $M_l=1.05614$, $M_2=0.98972$, $M_3=2.0512$, $M_c=0.98972$, $M_{66}=M_{88}=3$, $M_{55}=M_{77}=-3$.

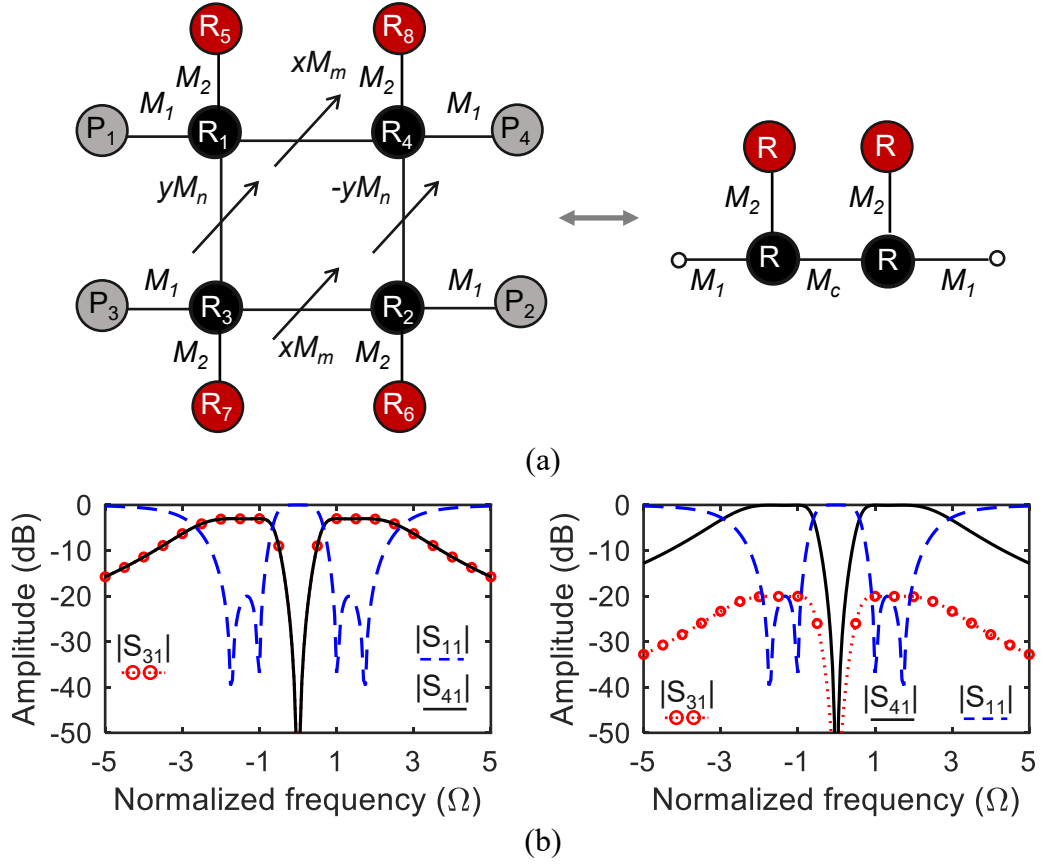


Figure 3.5. (a) CRD of the RFRC functionalizing a dual-band second-order TFs at its output ports P₃, P₄ using split-type resonators and its equivalent filtering CRD for each of its RF signal propagation paths. (b) Synthesized scattering parameters of the RFRC when $k=1$ or 100 . When $k=1$: $M_m=M_n=1.1752$. When $k=100$: $M_m=1.6538$, $M_n=0.1654$. The rest of the coupling coefficients are: $M_l=1.226$, $M_2=1.326$, $M_c=1.662$.

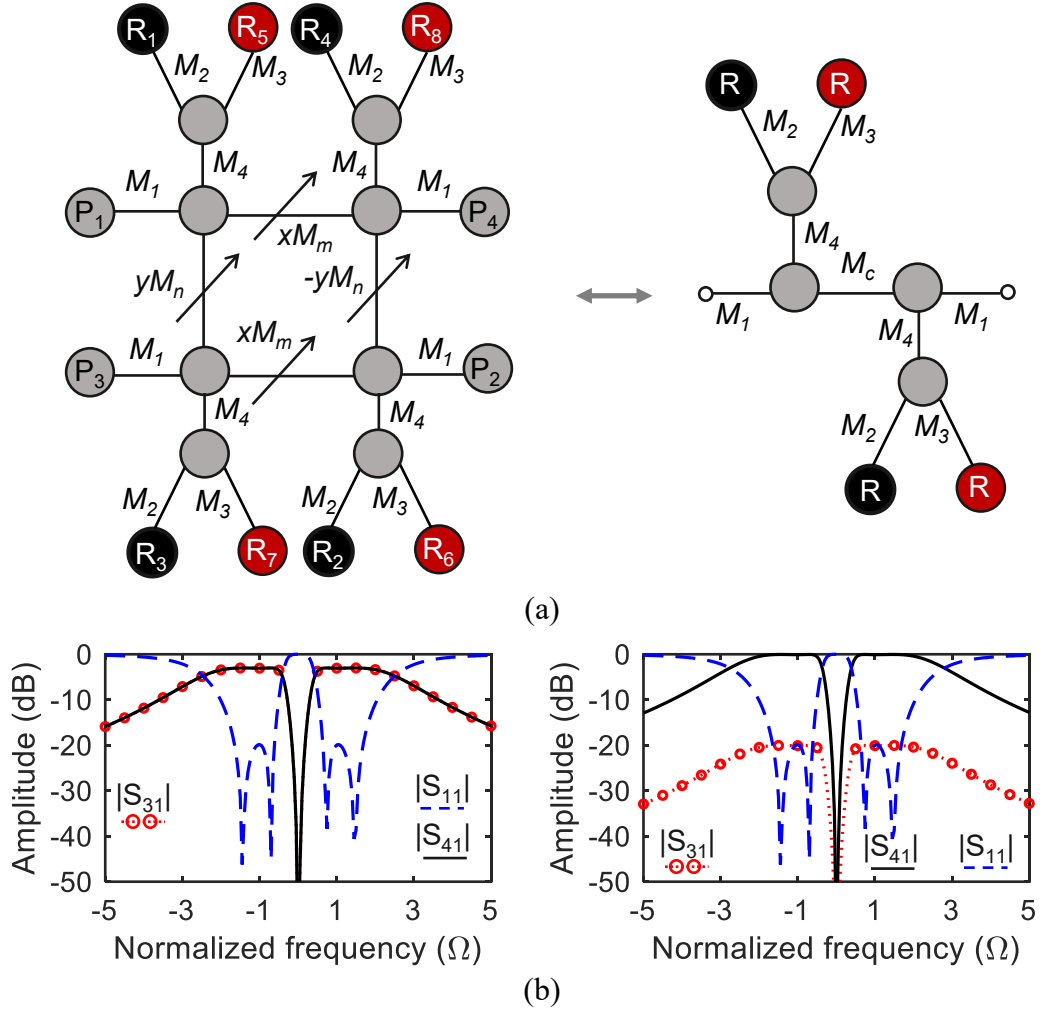


Figure 3.6. (a) CRD of the RFRC functionalizing a dual-band second-order TFs at its output ports P₃, P₄ using multi-resonant resonators and its equivalent filtering CRD for each of its RF signal propagation paths. The black nodes have the same self-resonating coefficients and the red ones have the same self-resonating coefficients. The grey nodes are non-resonating nodes. (b) Synthesized scattering parameters of the RFRC for $k = 1, 100$. When $k = 1$, $M_m = M_n = 0.5841$. When $k = 100$, $M_m = 0.8219$, $M_n = 0.0822$. The rest of the coupling coefficients are: $M_1 = 0.8635$, $M_2 = 0.78567$, $M_3 = 0.7823$, $M_4 = 0.771$, $M_c = 0.826$, $M_{11} = M_{22} = M_{33} = M_{44} = -1.04$, $M_{55} = M_{66} = M_{77} = M_{88} = 1.0085$.

3.1.2 Mode 1: RF co-designed bandpass filter, isolator and variable phase shifter (BPF/BPFI-VP).

The BPFI-VP mode of operation is obtained when the STM is ON and $\varphi_2 - \varphi_1 = 90^\circ$, facilitating a filtering response in the direction of propagation with a 360° continuously adjustable phase shift and RF signal cancellation in the reverse direction as depicted in Figure 3.1 (b). In this mode of operation, the isolator function can be enabled/disabled by turning ON/OFF the STM. To best analyze its behavior, we first

consider the case when STM is OFF, leading to a BPF-VP function whose response can be described by the relationships listed in (3.7) to (3.13). Specifically, (3.7) quantifies the incident power at P₁ and P₂ of the RFRC, with θ_{pd} being a fixed phase delay induced by the input network. Subsequently, (3.8) details the scattering matrix (S_c) of the RFRC at the center frequency, and (3.9) details the power relationship within the RFRC. Combining (3.7)–(3.9), the output power wave a_4 at P₃ can be expressed in relation to the input power a_1 . Then using (3.10), the total output power wave a_6 can be correlated with a_4 , and consequently with a_1 , leading to (3.11). Using (3.11), the amplitude $|S_{21}|$ and the phase $\angle S_{21}$ of the BPF-VP mode of operation (STM is OFF) can be calculated using (3.12) and (3.13). Notably, $|S_{21}|$ maintains a value of 1, while the insertion phase varies as a function of x , y , and k , as specified in (3.14) to (3.17).

$$a_2 = \frac{a_1}{\sqrt{2}} e^{j\theta_{pd}}, a_3 = a_2 e^{-j\frac{\pi}{2}} \quad (3.7)$$

$$[S_c] = \frac{-j}{\sqrt{k+1}} \begin{bmatrix} 0 & 0 & y \times 1 & x\sqrt{k} \\ 0 & 0 & x\sqrt{k} & -y \times 1 \\ y \times 1 & x\sqrt{k} & 0 & 0 \\ x\sqrt{k} & -y \times 1 & 0 & 0 \end{bmatrix} \quad (3.8)$$

$$a_4 = S_{c,31}a_2 + S_{c,32}a_3 \quad (3.9)$$

$$a_6 = \sqrt{2}e^{j\theta_{pd}}a_4 \quad (3.10)$$

$$\angle S_{21} = \angle a_6 - \angle a_1 = 2\theta_{pd} + \arctan\left(\frac{-x\sqrt{k}-jy}{\sqrt{k+1}}\right) \quad (3.11)$$

$$|S_{21}| = \frac{|a_6|}{|a_1|} = 1 \quad (3.12)$$

$$\angle S_{21} = \angle a_6 - \angle a_1 = 2\theta_{pd} + \arctan\left(\frac{-x\sqrt{k}-jy}{\sqrt{k+1}}\right) \quad (3.13)$$

$$-180^\circ \leq \angle S_{21} \leq -90^\circ, x = 1, y = 1 \quad (3.14)$$

$$90^\circ \leq \angle S_{21} \leq 180^\circ, x = 1, y = -1 \quad (3.15)$$

$$90^\circ \leq \angle S_{21} \leq 180^\circ, x = -1, y = -1 \quad (3.16)$$

$$-90^\circ \leq \angle S_{21} \leq 0^\circ, x = -1, y = 1 \quad (3.17)$$

It should be noted that for each x and y combination, adjusting k from 0 to ∞ yields a 90° phase tuning, and all combinations result in an overall 360° phase tuning. Additionally, due to the 90° phase difference of the phase coupling elements φ_1 and φ_2 , the RF signals reflected from the RFRC outside of the passband at P_1 and P_2 have equal amplitude and 90° phase difference. When these signals encounter the phase coupling elements again, they undergo an additional 90° phase shift, resulting in a total phase difference of 180° between them. Consequently, by the time these signals reach the output ports of the Wilkinson power divider, they have equal amplitudes but a 180° phase difference. Consequently, the two signals cancel each other out, preventing any reflection back to the input and ensuring a reflectionless response.

To showcase the continuous phase tuning capability of the multi-functional RF component, circuit simulations were carried out using the CRD in Figure 3.2 and ideal phase couplings with $\varphi_2 - \varphi_1 = 90^\circ$. The amplitude and phase responses of S_{21} for four distinct phase tuning cases are provided in Figure 3.7. In this case, k is set equal to 1 while x and y are varied across the four different combinations, as specified in (3.14) to (3.17). As shown, a constant filtering TF is maintained for all variable phase tuning states.

When the STM is turned ON, the isolator function can be enabled, and the device is working at its BPFI-VP mode. The STM is enabled by modulating the resonators of the RFRC in time and in space with progressively phase-shifted low-frequency AC signals as described in Chapter 2. To further demonstrate the operating principles of STM inside the RFRCs, two design examples are considered in Figure 3.8. Specifically, Figure 3.8 (a) depicts the CRD of a STM RFRC that exhibits a single-band third order TF, while Figure 3.8 (b) showcases a topology that facilitates a third order dual-band TF. The coupling coefficients are determined using the design methodology in Section 3.1.1 and the STM details of the applied AC signals are noted on the CRDs in Figure 3.8.

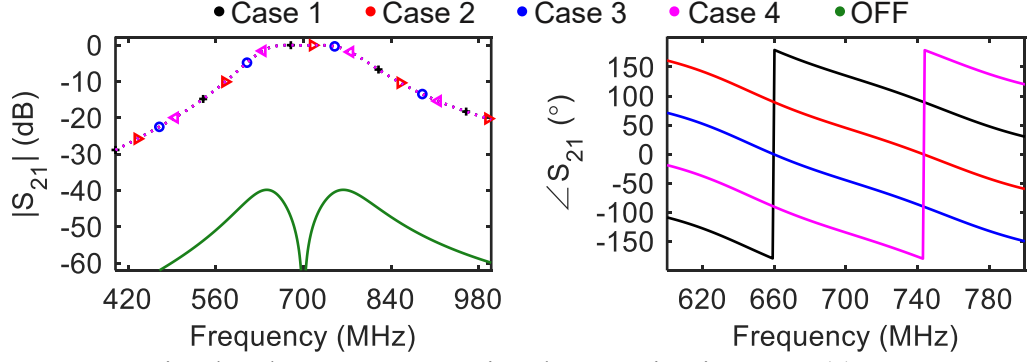


Figure 3.7. Simulated S-parameters using the CRD in Figure 3.2 (a) at BPF-VP mode. Case 1 : $x=-1, y=+1$, Case 2: $x=y=+1$, Case 3: $x=+1, y=-1$, Case 4: $x=y=-1$ and $k=1$ for all the four cases.

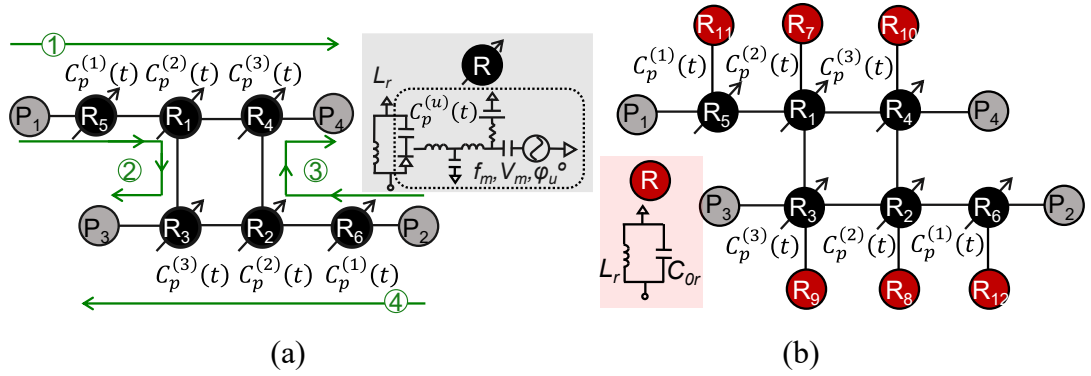
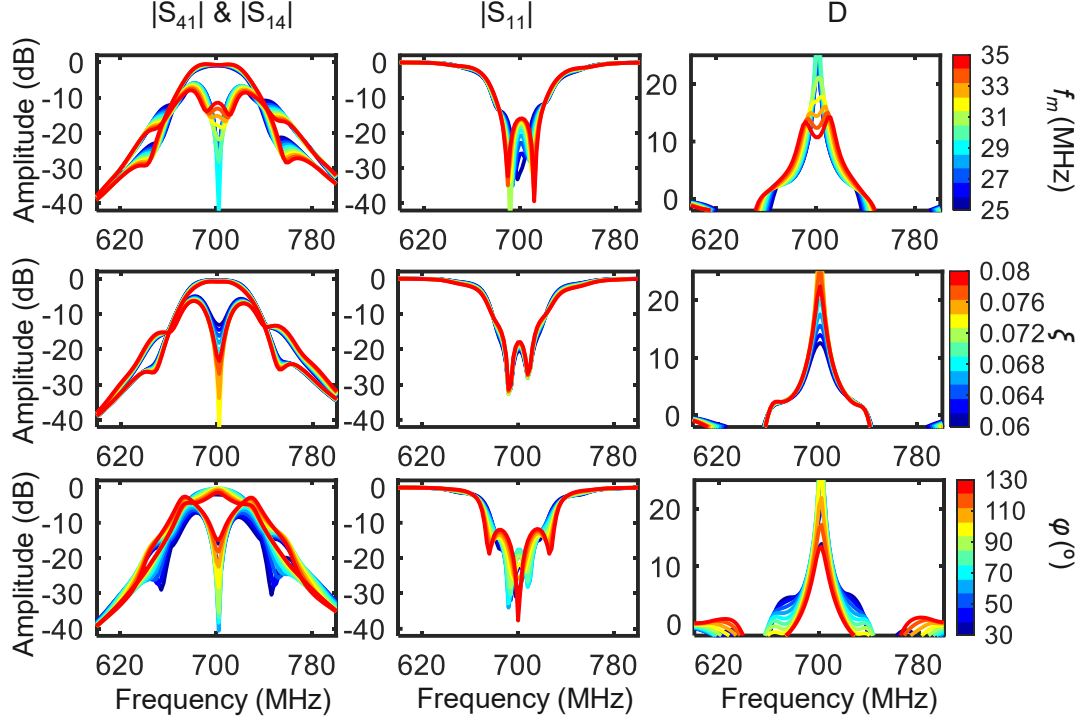


Figure 3.8. STM RFRC concepts allow to incorporate the isolator functionality. (a) CRD for the realization of a third order single-band response. (b) CRD for the realization of a third order dual-band response. In these CRDs, the black circles represent STM resonators and the red circles denote static resonators. The inductance for all of the resonators is set equal to $L_r = 5.17$ nH. The capacitance of the static resonators and the static capacitance of the STM resonators are $C_{0r}=10$ pF.

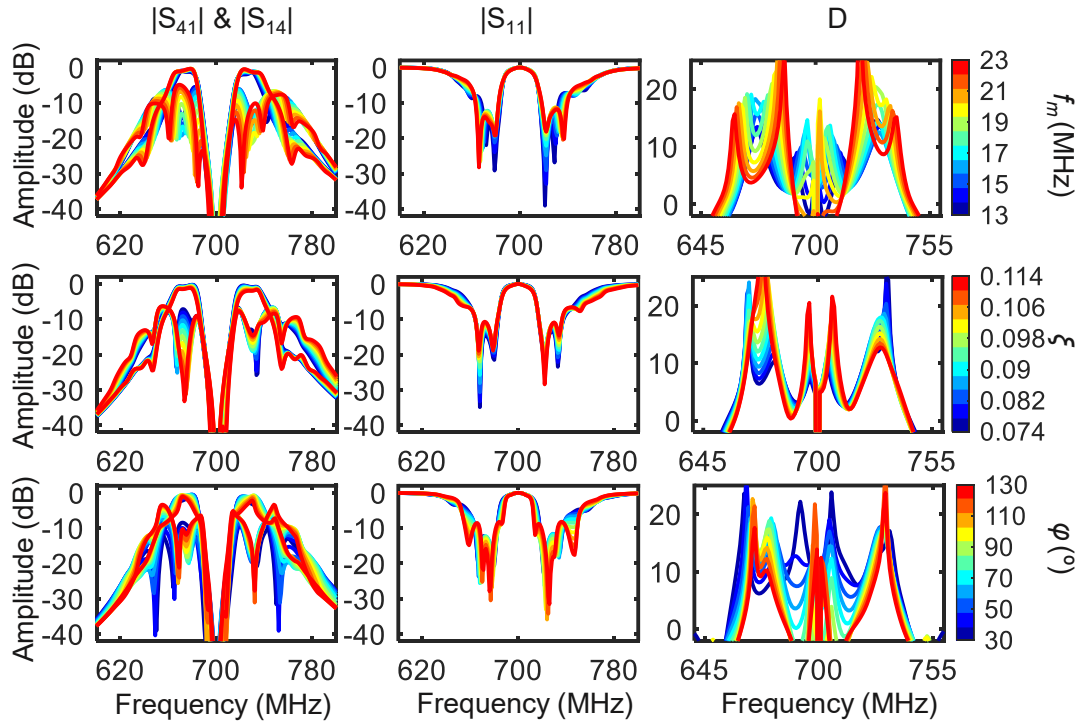
As presented in Figure 3.8 (a), the STM resonators is implemented by a parallel connection of LC network with a static inductance L_r and a time-varying capacitances $C_p^{(u)}(t)$ as discussed in Chapter 2. The capacitances of the resonators should satisfy $C_p^{(u)}(t) = C_{0r}[1 + \xi \cos(2\pi f_m t + \varphi_u)]$, where C_{0r} denotes the static capacitance, f_m is the AC signal frequency, ξ is the modulation index controlled by the amplitude of the AC signals and φ_u is the modulation phase. In contrast to the BPFIs discussed in Chapter 2, the AC signals for the STM resonators inside the RFRCs need to be carefully introduced to adjacent resonators and along each propagation path, so that they maintain a consistent phase difference φ . For instance, in Figure 3.8 (a), when power is introduced at P_1 and P_2 , it is subsequently directed to P_3 and P_4 , forming four distinct signal paths within the topology as marked and labeled in green. This

arrangement results in a sequential phase shift across all four paths. In the RFRC in Figure 3.8 (b), only the resonators depicted in black need to be modulated, as they form the passband of the filter. Conversely, the resonators marked in red create the notch between the two bands, and thus do not need to be modulated. By carefully selecting the aforementioned modulation parameters, unilateral transmission can be achieved, as it is conceptually illustrated in Figure 3.1 (b) and Figure 3.1 (d) by the broken dashed lines.

The STM parameters f_m , ζ and φ result in alternative performance trade-offs for the BPF in terms of IL, directivity (D), and RL as illustrated in Figure 3.9. They have been obtained using schematic circuit simulations in ADS Keysight using ideal lumped LC components and non-linear capacitors for the realization of STM resonators. Their element values are provided in the caption of Figure 3.8. In these cases, the power ratio of the RFRCs is set to 100 and the polarity of the reconfigurable couplings is chosen as $x = 1, y = 1$. The obtained performances are provided in Figure 3.9, where the amplitudes of the S-parameters ($|S_{11}|, |S_{14}|, |S_{41}|$) and the directivity ($D = |S_{41}| - |S_{14}|$) are plotted against frequency for different modulation parameters. As it is shown in Figure 3.9 (a), as f_m decreases, D increases, whereas IL decreases. Furthermore, when looking at the first row and third column, it can be observed that a larger f_m results in a wider 10 dB isolation BW (red curve) compared to a smaller f_m (blue curve), which provides higher D at the center frequency. Higher ζ enhances D but at the cost of increased IL. Large φ distorts the passbands and a φ of 65° provides the largest D . For the third order dual-band responses shown in Figure 3.9 (b), performance trade-offs can also be observed. Specifically, the increasing of f_m results in lower D for both bands and different values of f_m can be observed to obtain the highest D for the two bands (16 MHz for the lower band and 19 MHz for the upper band). Larger ζ leads to larger D for the lower band but smaller D at the upper band. While D increases for the upper band with an increase in φ , it first increases and then decreases for the lower band at the center frequency. Additionally, excessively large values of φ can distort the TF shapes for both bands. It can be noticed that the selection of STM parameters is application dependent. In this thesis, they are optimized to achieve high D within the passbands, while maintaining acceptable in-band IL.



(a)



(b)

Figure 3.9. S-parameters of the RFRCs in Figure 3.8 when varying f_m , ζ and ϕ . (a) Third order single-band response using the CRD in Figure 3.8 (a). (b) Third order dual-band response using the CRD in Figure 3.8 (b).

3.1.3 Mode 2: RF switching mode

The proposed multi-functional RF component is also able to operate as an RF switch when $\varphi_2 - \varphi_1 = 90^\circ$. Specifically, it can be intrinsically switched-off by setting the coupling element M_m to zero, while ensuring the couplings between R_1 and R_3 , and R_2 and R_4 , are identical as indicated in Figure 3.10. In this configuration, the input RF signal reaches the output after being equally divided between two distinct signal paths (marked by green dotted and solid lines) that introduce a 180° phase difference and are subsequently cancelled. The intrinsically switched-off capabilities of the concept are demonstrated in Figure 3.10 (green trace) for the same circuit configuration that was studied in the previous section.

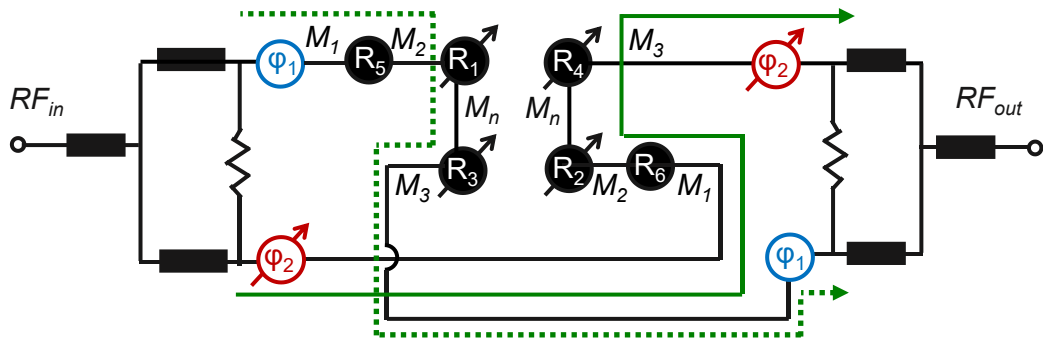


Figure 3.10. RF signal flow in the multi-functional RF component when $\varphi_2 - \varphi_1 = 90^\circ$, $M_m=0$ (i.e., Mode 2). In this configuration it functionalizes an intrinsically switched-off response.

3.1.4 Mode 3: RF co-designed bandpass filter, isolator, variable attenuator (BPF/BPFI-VA)

The BPF-VA mode of operation can be obtained when $\varphi_2=\varphi_1$, as shown in Figure 3.1 (c). Its operational behavior can be described using equations (3.18)–(3.21). Specifically, as detailed in (3.20), the output power wave a_6 can be expressed in terms of the input power wave a_1 , and can be adjusted from 0 to 1 by varying the power division ratio k from 0 to ∞ . This is illustrated in Figure 3.11 for the example case of the RFRC in Figure 3.2 with three different tuning states. Specifically, when $k=100$, 1 and 0.01 the attenuation is altered between 0, 3, 20 dB while the filtering TF and phase response remain unaltered. The same parametric studies can be performed at BPF/VA

mode to obtain the desirable modulation parameters and they have been omitted for brevity.

$$a_2 = a_3 = \frac{a_1}{\sqrt{2}} e^{j\theta_{pd}} \quad (3.18)$$

$$a_4 = S_{c,31}a_2 + S_{c,32}a_3, a_5 = S_{c,41}a_2 + S_{c,42}a_3 \quad (3.19)$$

$$a_6 = \frac{\sqrt{2}}{2} e^{j\theta_{pd}} (a_5 + a_4) \quad (3.20)$$

$$|S_{21}| = \frac{|a_6|}{|a_1|} = \left| \frac{\sqrt{k}}{\sqrt{k+1}} \right| \quad (3.21)$$

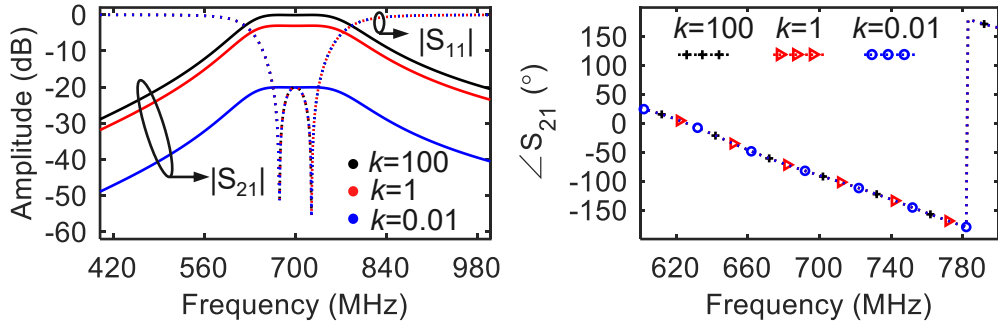


Figure 3.11. Simulated S-parameters using the CRD in Figure 3.2 at BPF-VA mode for different values of k .

3.2 Experimental Validation

To validate the multi-functional RF component concept, a single-band (Prototype 1) and a dual-band (Prototype 2) multi-functional prototypes were designed, manufactured, and characterized using as a reference the RFRCs in Figure 3.8 and two wideband Wilkinson dividers. Specifically, static and reconfigurable phase coupling elements φ_1 and φ_2 were employed in Prototype 1 to facilitate the realization of three reconfigurable modes of operation: (i) single-band BPF/BPFI-VP, (ii) intrinsic RF switching ON/OFF and (iii) single-band BPF/BPFI-VA. Prototype 2 has been implemented with a constant phase difference $\varphi_2 - \varphi_1$ equal to 90° allowing to realize two modes of operation, namely: (i) a dual-band BPF/BPFI-VP mode and (ii) an RF switch mode. The two prototypes were implemented for a center frequency of 700 MHz and a fractional bandwidth of 10% and 15% using a dual-core multi-layer

stack up comprising of two 0.508 mm-thick RO4003C substrates due to the need to realize a crossover as indicated in Figure 3.1 (a).

The manufactured prototypes are illustrated in Figure 3.12 and Figure 3.13 for the single-band and the dual-band prototype respectively. As it can be seen in Figure 3.12, to facilitate the VP/VA reconfigurability in Prototype 1, a static phase coupling element ϕ_1 (in blue dashed boxes) and a switchable phase coupling element ϕ_2 (in red dashed boxes) are used to allow for a reconfigurable phase difference between 0° and 90° between the two output ports by controlling the four PIN diodes D_1 to D_4 . Specifically, when D_1 and D_2 are ON while D_3 and D_4 are OFF, a phase difference of 90° is enabled between ϕ_2 and ϕ_1 , and the device is operating at the VP mode. On the contrary, the VA mode is enabled when D_1 and D_2 are OFF while D_3 and D_4 are ON. Prototype 2 adopts two static phase coupling elements ϕ_1 and ϕ_2 to facilitate a constant 90° phase difference at the two output ports. The detailed circuits of the RFRCs for the two prototypes are provided in the black dashed boxes when using lumped elements. Static inverters are implemented through their first-order π -type equivalent networks. Additionally, the circuits for the STM resonators and tunable couplings are distinctly highlighted in red and blue dotted boxes, respectively, located on the upper right side of Figure 3.12. The RF performance of the two prototypes were characterized using a Keysight N5244A Network Analyzer. The required STM low-frequency RF signals were generated using arbitrary waveform generators.

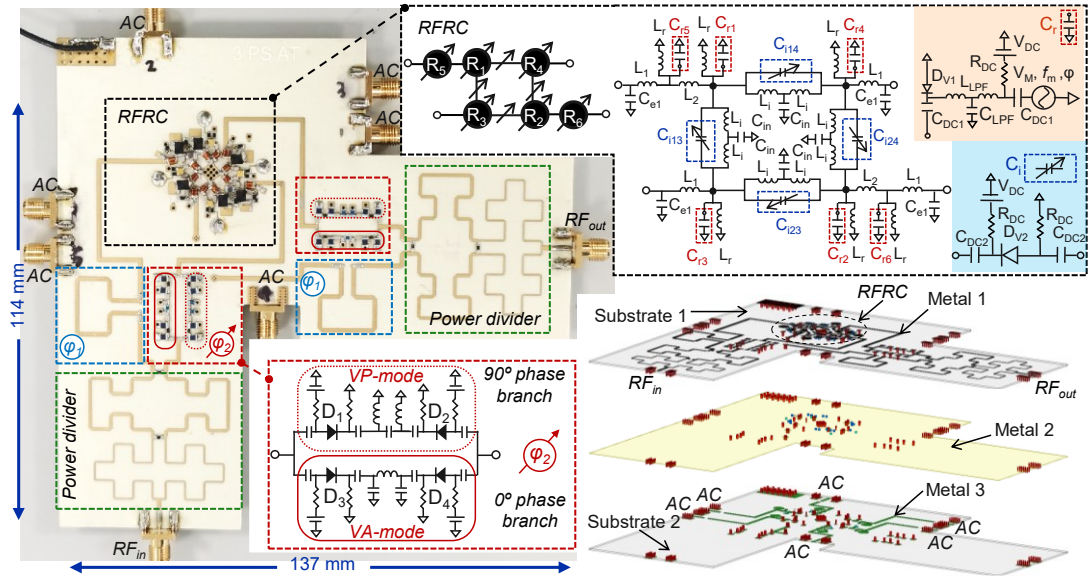


Figure 3.12. Manufactured Prototype 1: multi-functional component functionalising a single-band third order TF.

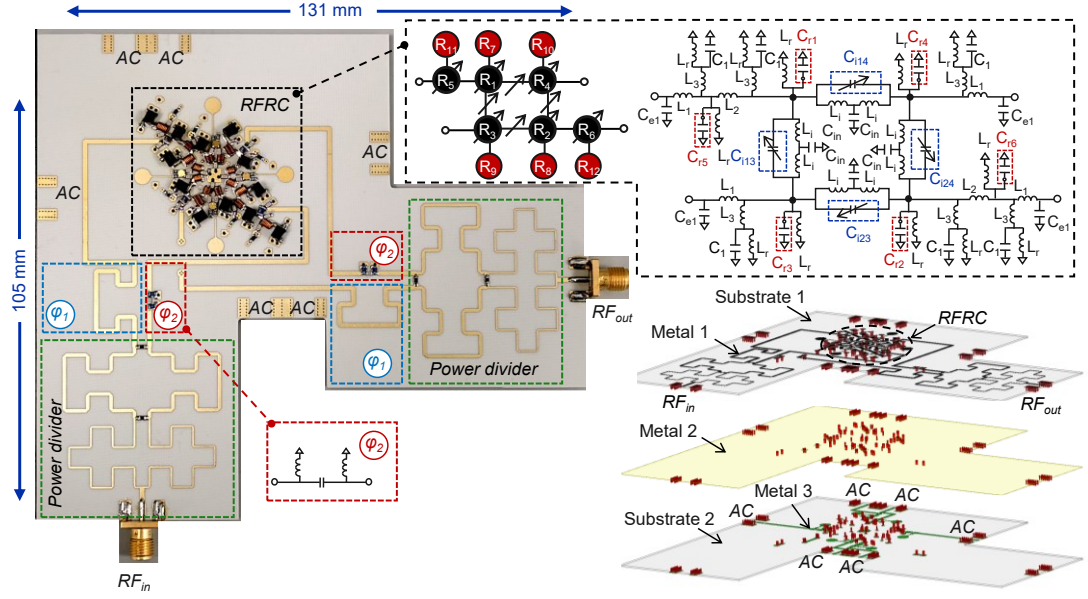


Figure 3.13. Manufactured Prototype 2: multi-functional component functionalising a dual-band third order TF.

The measured S-parameters for Prototype 1 are detailed in Figure 3.14 to Figure 3.16 for the three modes of operation. Specifically, Figure 3.14 shows the comparison between simulated and measured results at one tuning state, demonstrating good agreement. Figure 3.15 (a) depicts the performance of the multi-functional component when STM is OFF and operates in a BPF-VP mode of operation allowing for a full 360° phase tuning capability. Although nine distinct states are shown, its phase shift is continuously tunable. For these states, the IL was measured between 4 dB and 5.6 dB, alongside an input-reflectionless bandwidth (RBW) spanning from 375 MHz to 1093 MHz. The intrinsic switch-off mode of operation is also included in this figure (green trace). When STM is ON, the multi-functional component operates as a BPF-VP, as depicted in Figure 3.15 (c), (d) for the exact same states as in Figure 3.15 (a), (b). As evidenced, its measured D is >20 dB at the center frequency for all tuning states while exhibiting a 360° phase variation range as observed in Figure 3.15 (d). The measured 20 dB isolation BW (the frequency range when $|S_{21}| > 20$ dB) vary from 11 MHz to 14 MHz. It can also be observed that when STM is ON, the insertion phase is offset for all states due to STM introducing both phase and amplitude non-reciprocity, a phenomenon has also been discussed in Figure. 3 of [49]. However, this phase shift remains constant when the modulation parameters are unchanged, ensuring that the STM does not alter the overall phase tuning range of the device.

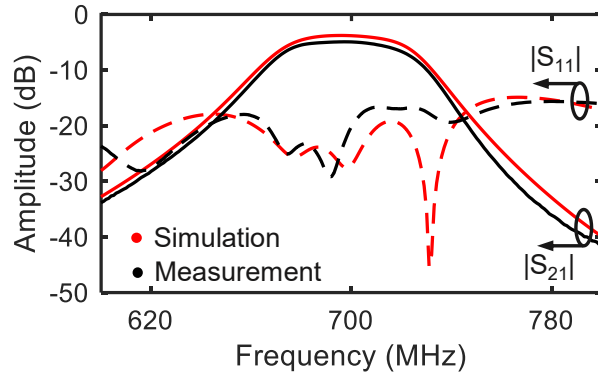
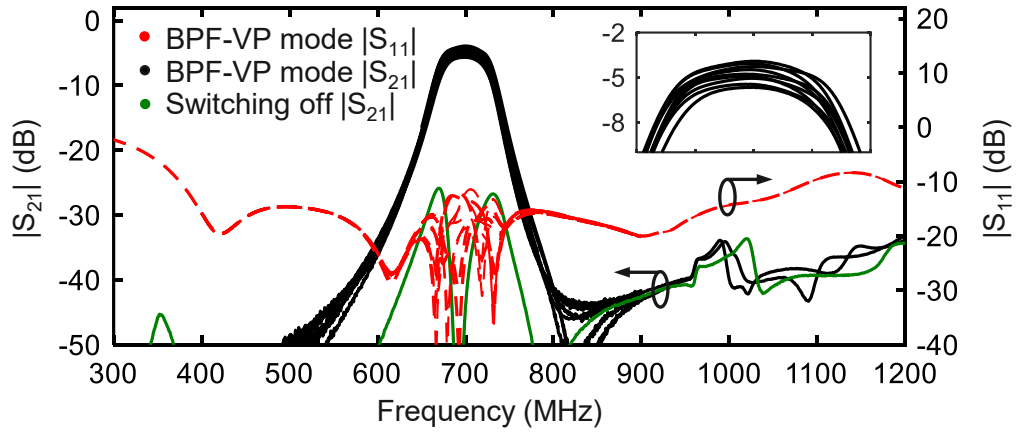
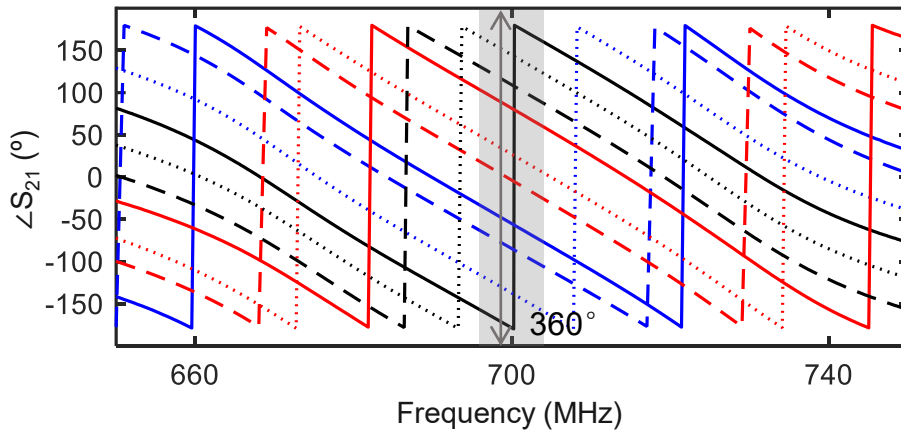


Figure 3.14. Comparison between simulated and measured results for Prototype 1 at one tuning state.



(a)



(b)

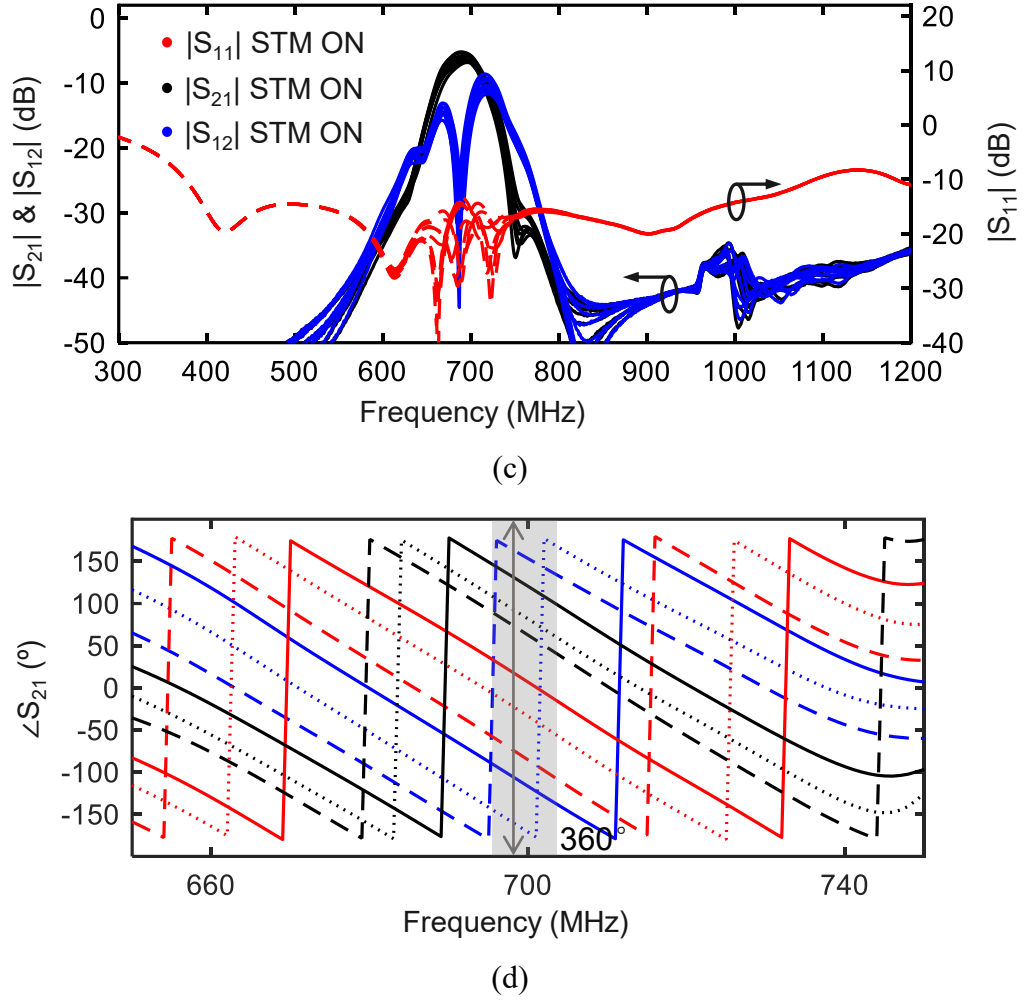


Figure 3.15. RF-measured performance of the single-band prototype when configured in its BPF/BPFI-VP mode. (a), (b) Amplitude and phase responses when STM is OFF and is set in its BPF-VP mode of operation. (c), (d) Amplitude and phase responses when STM is ON and is set in its BPFI-VP mode of operation.

Figure 3.16 details the performances of Prototype 1 in its BPF/BPFI-VA mode when $\varphi_2 - \varphi_1 = 0^\circ$. When STM is OFF and is configured in the BPF-VA mode of operation, a large attenuation tuning range can be obtained with its attenuation being continuously tunable between 4.2 dB to 20 dB as indicated in Figure 3.16 (a). The transmission zeros are generated by the mixed coupling within the filtering rat-race couplers. Adjusting these couplings to change the power division ratio also alters the frequency response of the equivalent network, thereby shifting the location of the TZs. For all these states, the measured ripple levels are less than 0.1 dB for attenuation levels less than 13.7 dB. At higher attenuation levels, it can be observed that the ripple levels increase to 0.4 dB due to the dispersion of the lumped element-based coupler and manual tuning of varactors. As can be seen in Figure 3.16 (b), the insertion phase

exhibits low phase imbalance to only 5° at the center frequency. Furthermore, when STM is ON, a $D > 20$ dB can be obtained for all tuning states as depicted in Figure 3.16 (c). The TF shape for all these tuning states remain well-preserved. For all these states, the IIP3 was measured between 14 dBm to 20 dBm.

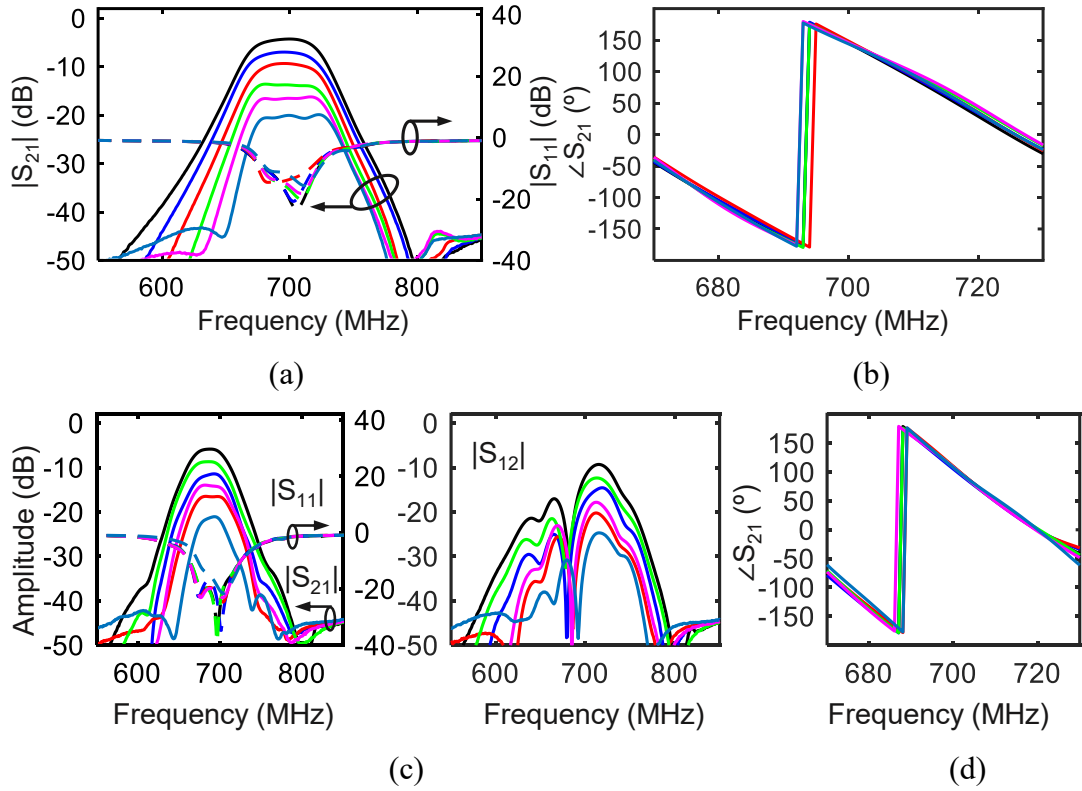


Figure 3.16. RF-measured performance of the single-band prototype when configured in its BPF/BPFI-VA mode. (a), (b) Amplitude and phase responses when STM is OFF and is set in its BPF-VA mode of operation. (c), (d) Amplitude and phase responses when STM is ON and is set in its BPFI-VA mode of operation.

The measurement results of Prototype 2 are concluded in Figure 3.17 and Figure 3.18. Figure 3.17 shows the comparison between simulated and measured TFs at one tuning state, which are in good agreement. The degradation in the isolation slope, isolation level between the two bands, and band flatness is mainly due to the tolerances of the lumped components. In the topology shown in Figure 3.8 (b), the static resonators (in red) are used to create the notch between the two bands and must resonate at the same center frequency. However, tolerances in the corresponding capacitors and inductors result in slight variations in the resonating frequencies of these resonators. When STM is OFF, it works as BPF-VP with the measured amplitude

responses provided in Figure 3.18 (a) and phase responses in Figure 3.18 (b). It can be observed from Figure 3.18 (a) that third order dual-band transfer functions are generated and the continuously phase tuning is achieved for both bands as indicated in Figure 3.18 (b), resulting a 360° phase tuning range in total. IL for both bands varies from 4.3 dB to 5.5 dB with a large RBW ranging from 368 MHz to 1200 MHz. The measured performance of its switching off mode is also included in Figure 3.18 (a) and is represented as the green trace with an isolation level of > 30 dB. When STM is ON, the multi-functional device works at its BPFI-VP mode with uni-directional transmission for both bands. Figure 3.18 (c) showcases enabled uni-directional transmission for both bands. In this case, the 18 dB isolation BW for the lower band varies from 6 MHz to 14 MHz. For the upper band, D is more than 10 dB for all the tuning cases at the center frequency of 733 MHz. A degradation of the IL and isolation can be observed for the upper band due to its higher sensitivity to variations in the capacitance and inductance values of the notch resonators. It is noted that across all displayed states for both BPF/BPFI-VP modes, the TFs for both bands are well preserved. The measured IIP3 across all these tuning states ranges from 17 dBm to 24 dBm.

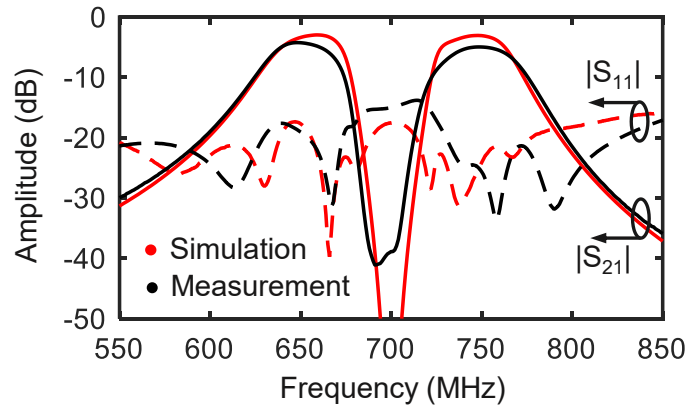
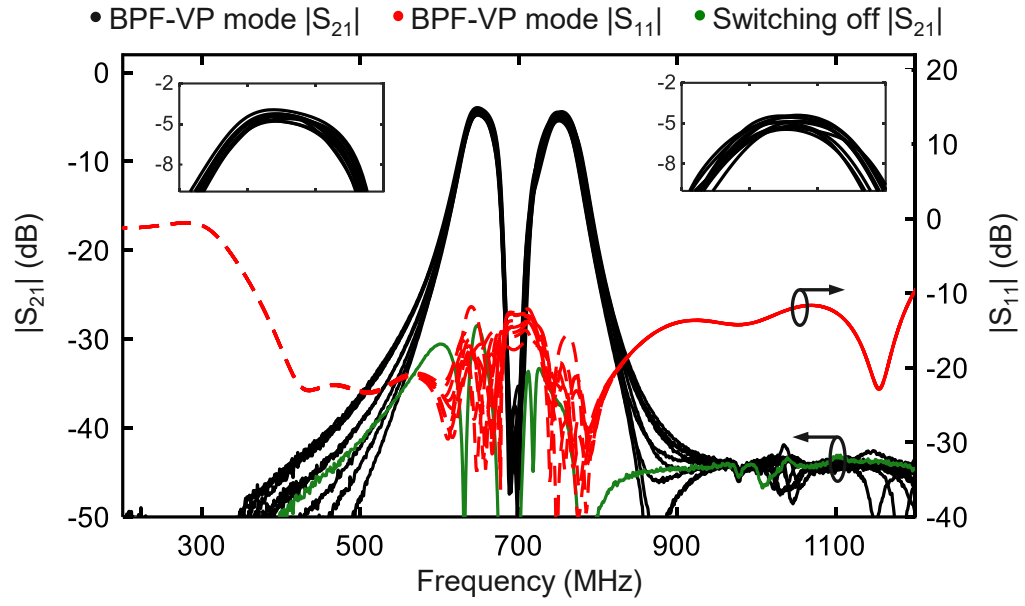
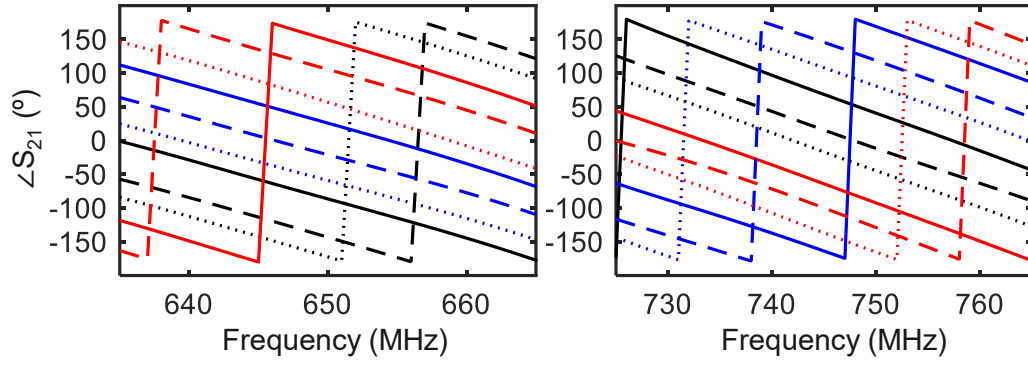


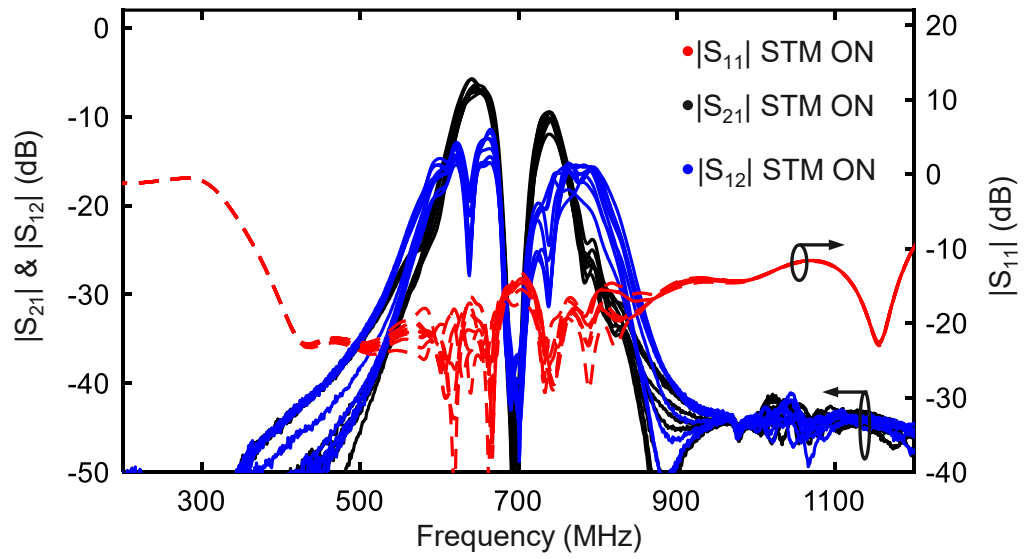
Figure 3.17. Comparison between simulated and measured results for Prototype 2 at one tuning state.



(a)



(b)



(c)

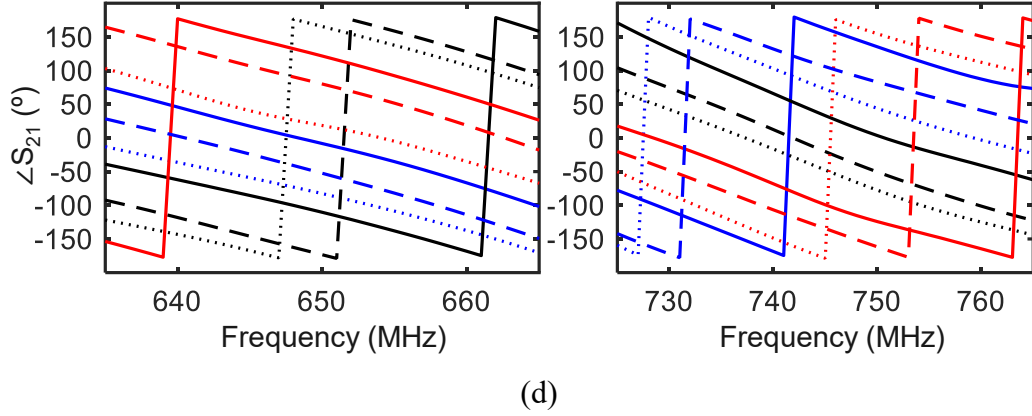


Figure 3.18. RF-measured performance of the dual-band prototype when configured in its BPF/BPFI-VP mode. (a), (b) Amplitude and phase responses when STM is OFF and is set in its BPF-VP mode of operation. (c), (d) Amplitude and phase responses when STM is ON and is set in its BPFI-VP mode of operation.

A comparison of the proposed multi-functional BPF concept with SOA co-designed BPFs, VPs/VAs and ISs is provided in Table 3.1. As evidenced, Prototype 1 achieves the largest phase tuning range with well-preserved transfer functions as opposed to the VPs in [82], [83], [84], [85] that have lower phase tuning range and distorted TF. Furthermore, through the manufacturing and testing of Prototype 2, this thesis demonstrates for the first time the potential to functionalize a dual-band TF exhibiting three multi-functional modes of operation as a unique advantage to be highlighted in relation to the SOA. In addition, this concept facilitates the realization of a reflectionless filtering TF, an additional RF switching-off mode of operation and also incorporates the functionality of an isolator with $D > 20$ dB. These operational features haven't been demonstrated in any of the existing VP topologies in Table 3.1. When comparing the performance of the proposed multi-functional circuit with the VAs in [93] and [97], Prototype 1 exhibits a larger attenuation tuning range. While the VAs in [95] offer a wider attenuation tuning, they experience considerable variations in the TF shape and RL during the tuning process. Additionally, the proposed BPFI-VA is also the only one to integrate an isolator function to a VA, achieving a D of more than 20 dB in all tuning states. Although [98] allows for both phase and attenuation adjustments, its attenuation range is limited, and tuning its phase or attenuation significantly alters the TFs. When comparing with the BPFIs that enable uni-directional transmission for single band [73] or multi-band [123] TFs, the proposed multi-functional device supports more functionalities and can work as a VP

or VA. The relatively high IL in the proposed multi-functional component is due to the use of lossy LEs having moderate quality factors (e.g, 80-200 for inductors and 50-75 for varactors) as also observed in LE-based RF component implementations. However, given the multi-functional nature of this circuit, integrating four functionalities, the IL is reasonable and comparable to other co-designed RF filters, such as those in [98], with IL ranging from 5.2 to 6.6 dB. When compared to state-of-the-art multi-functional filters, the dimensions of the proposed component are similar to those in [82], [83], and [95]. Although some other devices have slightly smaller sizes, the proposed device offers additional functionalities.

Table 3.1. Comparison with state-of-art co-designed multi-functional filters

Ref.	Mode	f_{cen} (GHz)	TF	PTR ($^{\circ}$)	ATR(dB) /PI($^{\circ}$)	RBW (GHz)	OFF	CTF	D (dB)	IL @ f_{cen} (dB)	Size ($\lambda_g * \lambda_g$)
T.W: P1	IS+VP	0.7	3 rd BPF	360	No	0.72	Yes	Yes	>20	4~5.6 (STM OFF) 5.4~6.8 (STM ON)	0.6*0.5
	IS+VA	0.7	3 rd BPF	No	15.8/<5	No			>20	4.2~20 (STM OFF) 5.8~21 (STM ON)	
T.W: P2	IS+VP	0.65; 0.745	3 rd dual-band	360	No	0.83	Yes	Yes	>10	4.3~5.5 (STM OFF) 6.5~12 (STM ON)	0.58*0.46
[82]	VP	1.05~1.3	2 nd BPF	320	No	No	No	No	No	4.1~5 @ 1.05 GHz	0.9*1.1
[83]	VP	0.6~0.71	4 th BPF	175	No	No	No	Yes	No	1.9~3	0.67*0.37
[84]	VP	0.65~0.86	3 rd BPF	360	No	No	No	No	No	2.4~4.3 @ 0.65 GHz	0.14*0.14
[85]	VP	1.07~1.36	3 rd BPF	90	No	No	No	Yes	No	0.8~1.4	0.31*0.27
[93]	VA	1	3 rd BPF	No	8/<1.5	No	No	Yes	No	2~10	0.145*0.29
[95]	VA	1.01~1.19	2 nd BPF	No	18/N/A	No	No	No	No	3.3~21.5	1.02*0.72
[97]	VA	2	5 th BPF	No	14/N/A	No	No	Yes	No	2.76~16.99	0.36*0.31
[98]	VA+VP	0.75~0.95	2 nd BPF	360	10/0~8	No	No	No	No	5.3~6.6(VP) 5.3~15(VA)	0.48*0.4
[73]	IS	0.88~1.03	3 rd BPF	No	No	No	No	N/A	>20	3.9 (STM OFF) 5.2 (STM ON)	0.26*0.1
[123]	IS	0.6~0.68; 0.72~0.83	3 rd dual-band	No	No	No	No	N/A	>20	4.8~6.2 (STM ON)	0.28*0.24

(T.W: This work, P1: Prototype 1, P2: Prototype 2, TF: transfer function, PTR: phase tuning range, ATR: attenuation tuning range, PI: phase imbalance at center frequency, RBW: reflectionless bandwidth, OFF: intrinsic switching off capability, CTF: constant transfer function, D: directivity at center frequency, IL: insertion loss, IS: isolator, VA: variable attenuator, VP: variable phase shifter.)

3.3 Conclusion

This chapter presented an innovative RF co-design method for designing multi-function devices capable of serving as BPFs with single-/dual-band TFs, VPs, VAs, RF switches and ISs. The methodology was validated at 700 MHz through two

prototypes, showcasing the ability to obtain both single- and dual-band bandpass TFs. Specifically, Prototype 1 demonstrates three distinct modes of operation: (i) BPF/BPFI-VP mode, (ii) RF switch mode, and (iii) BPF/BPFI-VA mode. In the BPF-VP mode, it features a third order bandpass TF with a 360° phase tuning range and a broad RBW spanning 375 to 1093 MHz. The BPF-VA mode offers a third-order bandpass TF with an attenuation range from 4.2 dB to 20 dB as well as a low phase imbalance of $<5^\circ$. Both BPFI-VP mode and BPFI-VA mode exhibit unidirectional transmission with a D exceeding 20 dB. Prototype 2 is designed to facilitate a third-order dual-band bandpass TF with BPF/BPFI-VP mode and RF switching off mode. A continuously tunable phase shift of 360° with a RBW spanning from 368 MHz to 1200 MHz is demonstrated for both bands at BPF/BPFI-VP working mode as well as a $D > 10$ dB at BPFI-VP mode. It should be noted that the filtering TFs across all operational states for both prototypes are well preserved.

Chapter 4: Multi-port Multi-functional RF Routing Networks Using STM Resonator Arrays

This chapter introduces a new type of reconfigurable multi-port multi-functional RF signal-routing network based on STM resonator arrays. They are based on the reconfigurable STM resonator arrays that are arranged in star- and boxtypes configurations. Specifically, the star-type block supports three modes of operation: (i) a reconfigurable SPMT co-designed BPFI; (ii) a PD-BPFI; and (iii) a BPFC. The box-type block supports two modes: (i) a CR-BPFI mode that can support both concurrent two-channel or single-channel transmission and (ii) a PD-BPFI mode. The proposed multiport multifunctional STM-based array concept is demonstrated through circuit-level simulations and is validated by two experimental prototypes. This approach differs significantly from prior STM-based RF devices by exploiting all these three modulation states within a unified multi-port architecture to realize a broader set of reconfigurable RF functions. The multi-port and multi-functional RF devices and contributions of this chapter have been presented in [127].

4.1 Theoretical Foundations

4.1.1 Multi-port multi-functional RF signal-routing network

The proposed reconfigurable multi-port multi-functional RF signal-routing network was conceptually illustrated in Figure 4.1. It consists of an array of reconfigurable STM resonators denoted by black circles with arrows. Each resonator operates in one of three states, determined by the modulation index of its time-varying capacitance. Specifically, as the modulation index increases from zero, the resonator transitions from an unmodulated state (black circle) to a moderately-modulated state (blue circle with arrow), and then to an over-modulated state (red circle with arrow). As such, the RF signal propagation between a set of interconnected resonators can be altered from a bilateral BPF response to a unilateral BPFI response and to an intrinsically switched-off response as it will be explained in the next section.

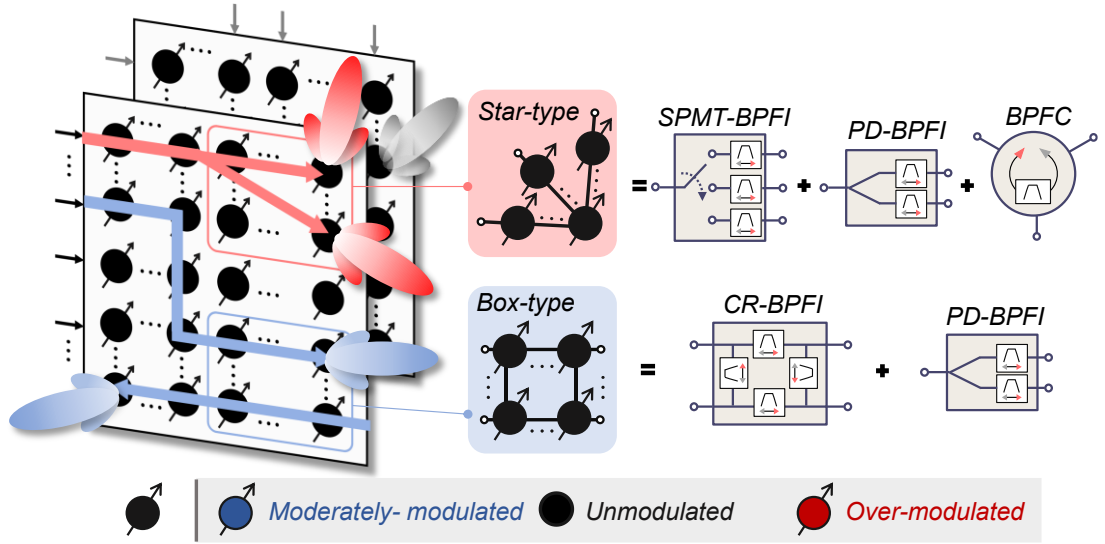


Figure 4.1. Conceptual architecture of the proposed reconfigurable multi-port multi-functional RF signal-routing network based on STM resonator arrays. Each resonator is configured in one of three states: unmodulated (black), moderately-modulated (blue), or over-modulated (red), enabling reconfigurable signal paths.

By adjusting the modulation states of resonators across the array, the RF signal flow between the ports can be dynamically controlled, enabling diverse RF signal-processing functionalities as shown in Figure 4.1. These functionalities are obtained through two reconfigurable multi-functional building blocks—a star-type (red) and a box-type (blue)—whose representative RF signal flows are indicated by the corresponding arrows in Figure 4.1. Each block can be reconfigured to realize multiple alternative functions as summarized in the right side of Figure 4.1. Specifically, the star-type block exhibits three modes of operation: (i) a reconfigurable single-pole multi-throw BPFI (SPMT-BPFI), (ii) a PD-BPFI, and (iii) a BPFC. The box-type block provides two modes of operation: (i) a channel reconfigurable BPFI (CR-BPFI), which can operate in either concurrent dual-channel or single-channel transmission configurations, and (ii) a PD-BPFI. All operational modes are realized by solely controlling the modulation states of the resonators, eliminating the need for additional RF switches or control elements.

4.1.2 Operational states of the star- and box- type blocks

Figure 4.2 (a) shows the CRD of the proposed star-type block in its SPMT-BPFI mode of operation. It consists of N BPF branches, each comprising k reconfigurable STM resonators (represented by black circles with arrows) and they are all connected

to a shared central resonator, R_0 . In total, the block contains $(k \times N) + 1$ reconfigurable resonators. Each STM resonator can be modeled as a parallel LC tank consisting of an inductor L_r and a time-varying capacitor $C_{0r}(t) = C_{0r}[1 + \xi \cos(2\pi f_m t + \varphi)]$, where C_{0r} is the nominal capacitance, f_m is the modulation signal frequency, ξ is the modulation index and φ is the modulation phase. Depending on the value of ξ , each resonator can operate in one of three states: unmodulated ($\xi=0$, represented by a black circle), moderately-modulated ($\xi < 0.2$, represented by a blue circle with arrow), or over-modulated ($\xi > 0.2$, represented by a red circle with arrow). The capacitance variation of the time-varying capacitors is shown in Figure 4.2 (b) for the three representative modulation states. In the unmodulated state ($\xi = 0$), the capacitance remains constant, and the resonator behaves as a bilateral resonant element. In the moderately-modulated state ($\xi < 0.2$), the capacitance varies sinusoidally with a small modulation index, enabling directional transmission when used in an in-line STM resonator array. In the over-modulated state ($\xi > 0.2$), the larger capacitance swing transfers a significant portion of the RF energy to higher-order products, suppressing the fundamental tone and enabling intrinsic switching-off without the need for additional RF switches. By configuring the modulation state of each resonator inside the star-type block in Figure 4.2 (a), any two ports can be selected as the RF input/output and a BPFI response of order $2k+1$ can be obtained in either direction, while the remaining ports are isolated.

Figure 4.2 (c) and (d) illustrate an example switching state, where ports 1, N are set as input/output ports. In this configuration, the resonators along the active transmission path are moderately-modulated to support the BPFI mode of operation. The remaining paths are disabled using over-modulated resonators that suppress the RF signal propagation as it will be explained in the next section. The transmission direction of the active paths is defined by the spatial phase progression of the moderately-modulated resonators, as discussed in Chapter 2 and Chapter 3. The corresponding conceptual S-parameters of the star-type CRD in Figure 4.2 (c) are illustrated in Figure 4.2 (d) when the signal transmission direction is from Port 1 to Port N. As shown, the active path exhibits a forward bandpass-type transfer function with reverse attenuation (black curves), while all other ports remain isolated (brown curves).

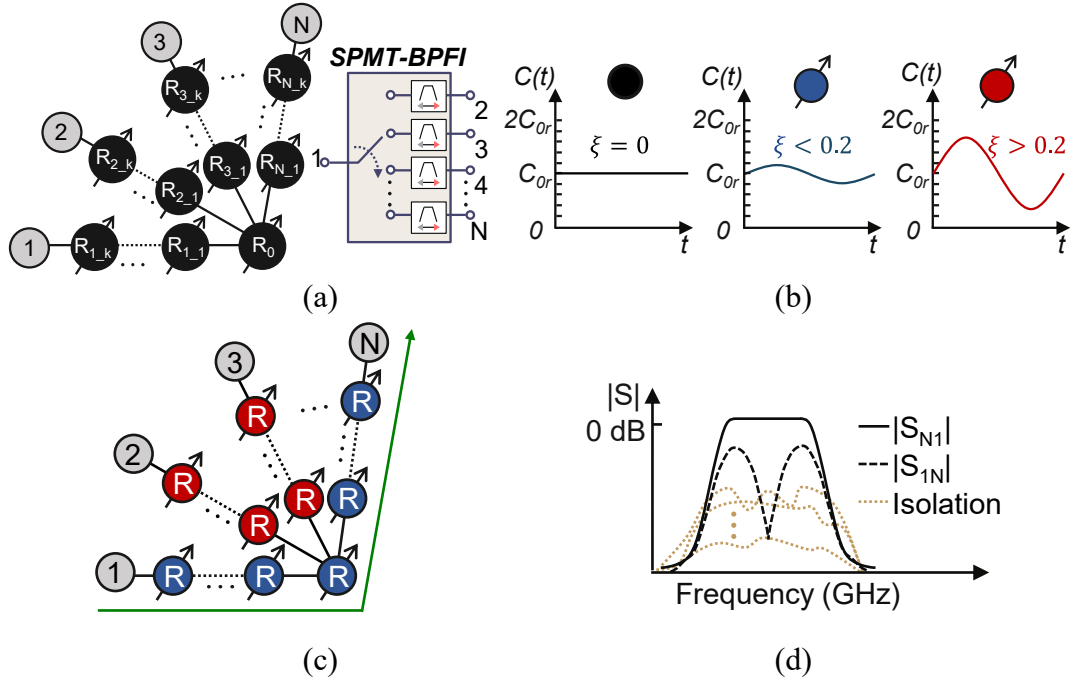


Figure 4.2. (a) Generalized CRD of the proposed star-type SPMT-BPFI. (b) The capacitance variation of the time-varying capacitors for the three representative modulation states. (c), (d) Modulation scheme and conceptual S-parameters of the generalized star-type SPMT-BPFI when Port 1 and Port N are configured as the input and output ports.

In addition to SPMT-BPFI, the star-type block also supports two other operational modes (i.e., the PD-BPFI and the BPFC) as indicated in Figure 4.3 (a). These modes are enabled within a three-port configuration with each branch containing at least two resonators ($k \geq 2$). Furthermore, the reconfigurable resonator R_0 needs to be altered between a resonating node and a non-resonating node, offering sufficient degrees of freedom for modulation-state control. Figure 4.3 (b) shows an example of the PD-BPFI mode, where the signal is inserted at port 1 and is equally split into ports 2 and 3. In this configuration, the central resonator R_0 is configured as a resonating node, while the resonators along the signal paths are moderately-modulated to establish symmetric transmission paths. Figure 4.3 (c) illustrates the BPFC mode, where resonator R_0 is configured as a non-resonating node. The first resonators connected to R_0 (i.e., $R_{1,1}$, $R_{2,1}$ and $R_{3,1}$) are moderately-modulated while the others remain unmodulated to enable clockwise signal circulation among ports 1 to 2, 2 to 3 and 3 to 1 as indicated by the conceptual S-parameters.

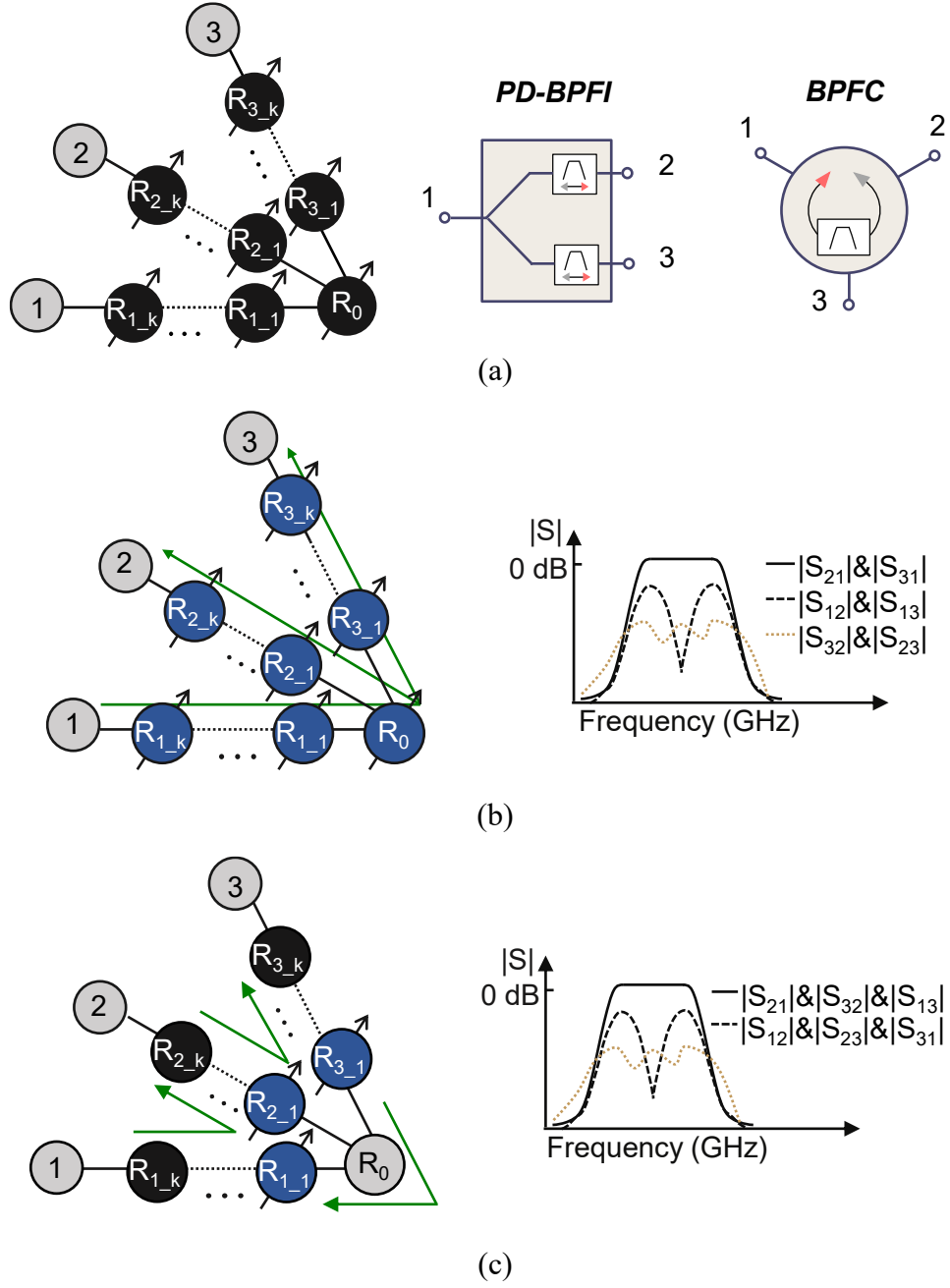


Figure 4.3. (a) CRD of the proposed star-type block with PD-BPFI and BPFC modes. (b) Modulation scheme and conceptual S-parameters of the star-type PD-BPFI when the RF power is equally divided. (c) CRD and modulation scheme of the proposed star-type BPFC.

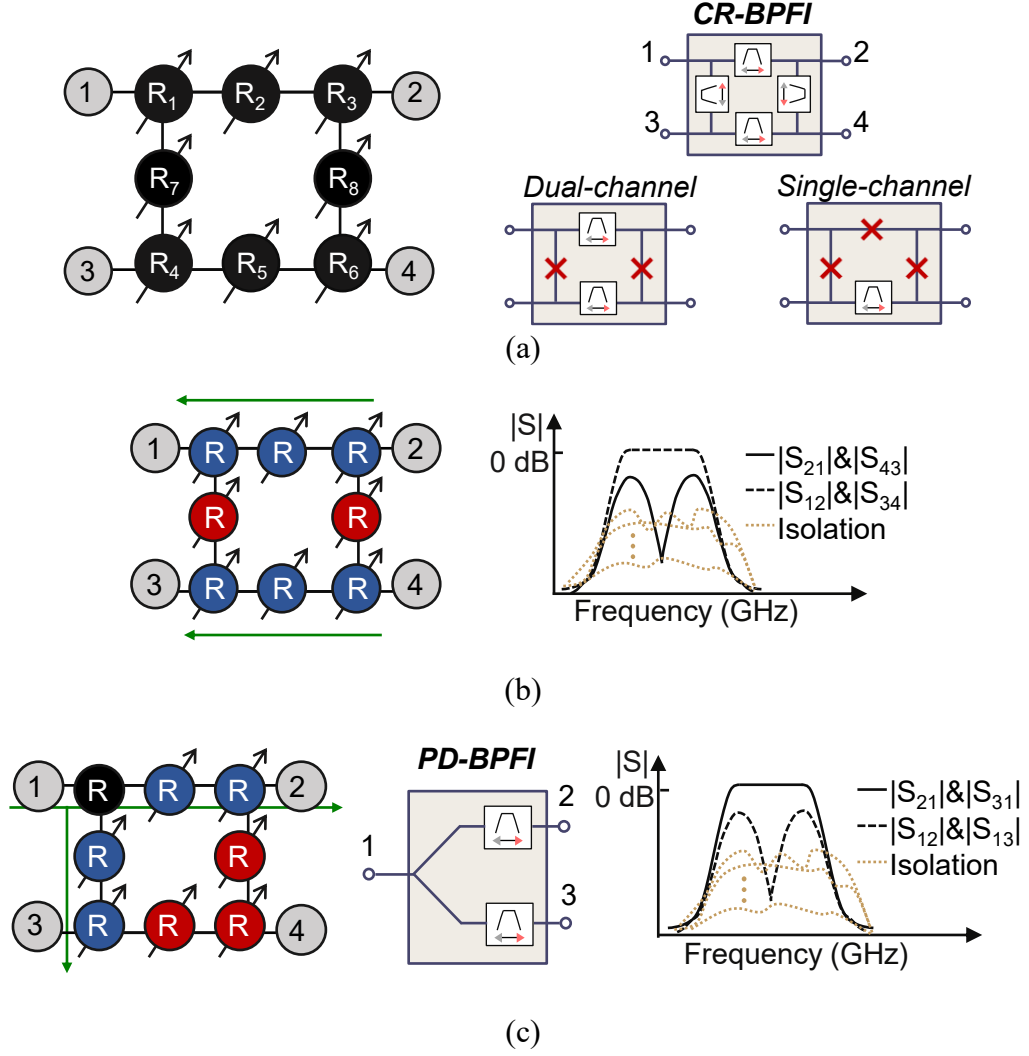


Figure 4.4. (a) CRD of the proposed box-type topology and its CR-BPFI mode. (b) Modulation scheme of the CR-BPFI mode when supporting a dual-channel transmission and its conceptual S-parameters. (c) Modulation scheme of the PD-BPFI mode and the conceptual S-parameters.

While the star-type block achieves flexible single-path reconfiguration, certain signal routing cases in multi-antenna systems demand simultaneous dual-channel transmissions with minimal interference. To address this requirement, a box-type STM block is proposed next to support concurrent and reconfigurable routing across multiple channels. Figure 4.4 (a) illustrates the CRD of the proposed box-type network, which consists of eight reconfigurable resonators arranged in a box configuration. This structure supports a CR-BPFI mode, which facilitates a dual-channel or single-channel BPFI response, as conceptually indicated in Figure 4.4 (a). In both single- and dual-channel cases, the RF signal channels can be dynamically

selected by using moderately-modulated resonators along the activated paths, while applying over-modulation to all other resonators to ensure isolation. Figure 4.4 (b) presents an example of a dual-channel transmission, where the RF signals propagate simultaneously from Port 2 to Port 1 and from Port 4 to Port 3. Figure 4.4 (c) illustrates the PD-BPFI mode, enabled by having an unmodulated resonator at the input, moderately-modulated resonators along the two signal-dividing paths, while the resonators connected to isolated ports are over-modulated to suppress unintended signal propagation.

4.1.3 Spatiotemporal modulation as a reconfiguration mechanism

Unlike the state-of-the-art and the BPFIs presented in Chapters 2 and 3, which primarily focus on achieving unilateral transmission using STM resonator arrays, this section expands the application of STM to multi-port networks by investigating the different modulation states as a reconfiguration mechanism. A simple three-resonator inline resonator array is considered to explain how STM can be used as a reconfiguration mechanism in the star-type and box-type multi-functional blocks. Three different modulation schemes are studied: (i) unmodulated resonators array in Figure 4.5 (a), (ii) moderately-modulated resonators array in Figure 4.5 (b) and (iii) mixed unmodulated and over-modulated resonators array in Figure 4.6 and Figure 4.7 and their details are summarized as follows:

Unmodulated resonators array: Figure 4.5 (a) shows the CRD of the three-resonator in-line network in which resonators R_1 – R_3 are all unmodulated. The inter-resonator couplings are implemented by ideal couplings and the resonators are realized by a parallel connection of an inductor L_r and a capacitor C_{0r} using the component values in Figure 4.5. They have been specified for a three-pole Chebyshev response with 0.1 dB ripple and a 5% FBW at 1.2 GHz as shown in the unmodulated S-parameter response in Figure 4.5 (a).

Moderately-modulated resonators array: As explained in Chapter 2 and Chapter 3, STM can be introduced to the three-resonator array by modulating the capacitances of all three resonators R_1 – R_3 using low-frequency AC signals with spatially progressive phase shifts to facilitate unilateral transmission. As conceptually illustrated in Figure 4.5 (b), each node comprises a fixed inductor L_r and a time-

varying capacitor with their modulation parameters labelled on top. Using the modulation parameters in the caption of Figure 4.5, unilateral signal propagation can be obtained as evidenced in Figure 4.5 (b). A forward IL of approximately 0.55 dB is observed, which results from energy transfer to the intermodulation (IM) products as also discussed in [49]. This behavior is further examined through harmonic balance simulations in Figure 4.5 (c) and Figure 4.5 (d) that present the output spectra when a single-tone signal is applied at Port 1 and Port 2, respectively. As it can be observed in Figure 4.5 (c), when -10 dBm excitation at 1.2 GHz is applied at Port 1, the output spectrum at Port 2 shows spectral components at the fundamental frequency f_0 , and at $f_0 \pm f_m$, $f_0 \pm 2f_m$, and $f_0 \pm 3f_m$. Higher-order IM products are below -60 dBm and are thus neglected. The received power at f_0 aligns with the simulated $|S_{21}|$ in Figure 4.5 (b). In reverse transmission, the output power at Port 1 is significantly attenuated to approximately -36 dBm at f_0 , as shown in Figure 4.5 (d), confirming the unilateral signal propagation with a directivity ($D = |S_{21}| - |S_{12}|$) of 26 dB.

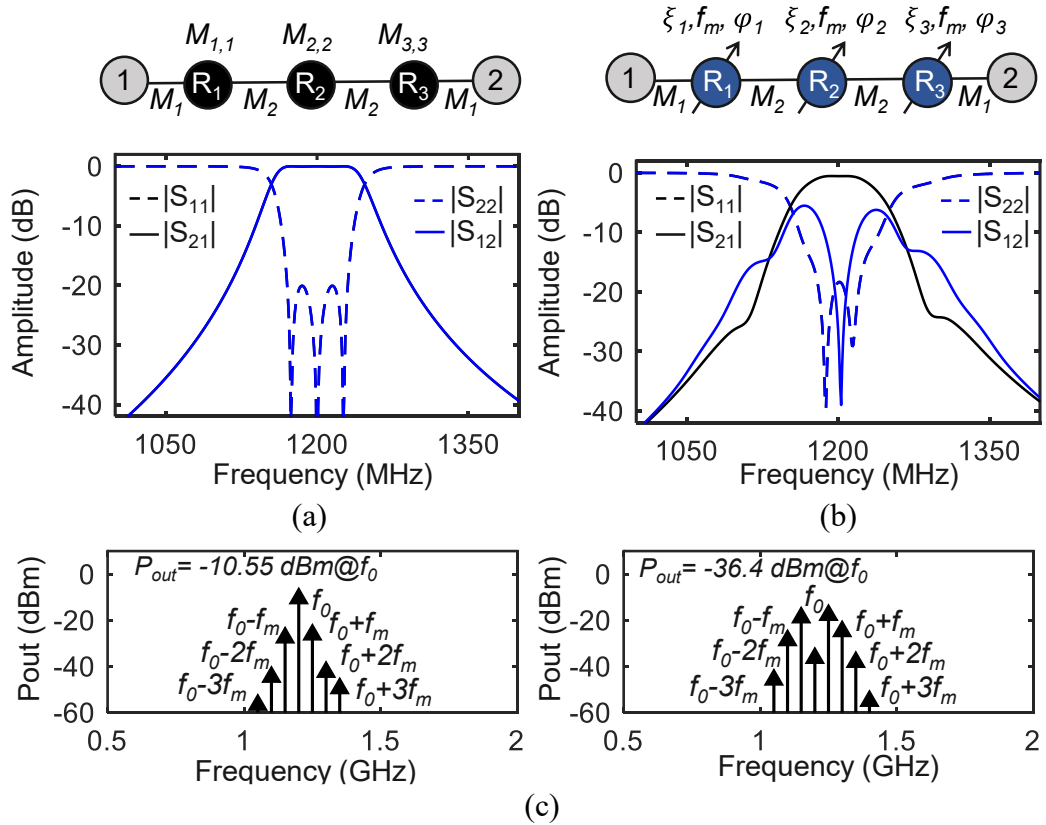


Figure 4.5. CRDs and circuit-simulated S-parameters of a three-resonator array. The coupling coefficients of the CRD are: $M_1=1.0836$, $M_2=1.0317$, $M_{11}=M_{22}=M_{33}=0$, and the nominal resonator element values are: $L_r=1.76 \text{ nH}$, $C_{0r}=10 \text{ pF}$. (a) Unmodulated resonators array. (b) Resonator array having moderately-modulated resonators (blue circles with arrow) with $f_m=50 \text{ MHz}$, $\phi_3-\phi_2=\phi_2-\phi_1=60^\circ$ and modulation index $\xi_1=\xi_2=$

$\xi_3=0.072$. (c), (d) Output spectra under -10 dBm excitation at Port 1 and Port 2, respectively.

Mixed unmodulated and over-modulated resonators array: Considering that the RF signal power spreads to the harmonics when the modulation is applied, the potential to use over-modulation as an RF switching mechanism is explored in Figure 4.6 for the case of increasing ξ for R_1 , while keeping R_2 and R_3 unmodulated. Specifically, Figure 4.6 (a) shows the simulated S-parameters as ξ increases from 0 to 0.6, with f_m fixed. As it can be observed, both $|S_{21}|$ and $|S_{12}|$ decrease with the increase of ξ , reaching -16.5 dB at 1.2 GHz for $\xi=0.6$, indicating high rejection in both directions due to the energy redistribution across a wide range of IM frequencies as shown in the spectral responses of the three-resonator network in Figure 4.6 (b) and Figure 4.6 (c). Specifically, when a -10 dBm excitation is applied at Port 1 in Figure 4.6 (b), a significant drop in the fundamental tone amplitude is observed followed by an increased IM content at the IM frequencies when compared to the moderately modulated STM case in Figure 4.5 (c). A similar trend is observed in Figure 4.6 (c) when Port 2 is excited. As evidenced, when increasing ξ from 0.2 to 0.6, the order of the IM products increases from 6 to 15. As it can be seen in Figure 4.6 (a) symmetric attenuation occurs in both directions. This is due to the fact that only one resonator is being modulated in this case as opposed to conventional STM approaches where multiple resonators need to be modulated with progressively shifted RF signals as also discussed in [49].

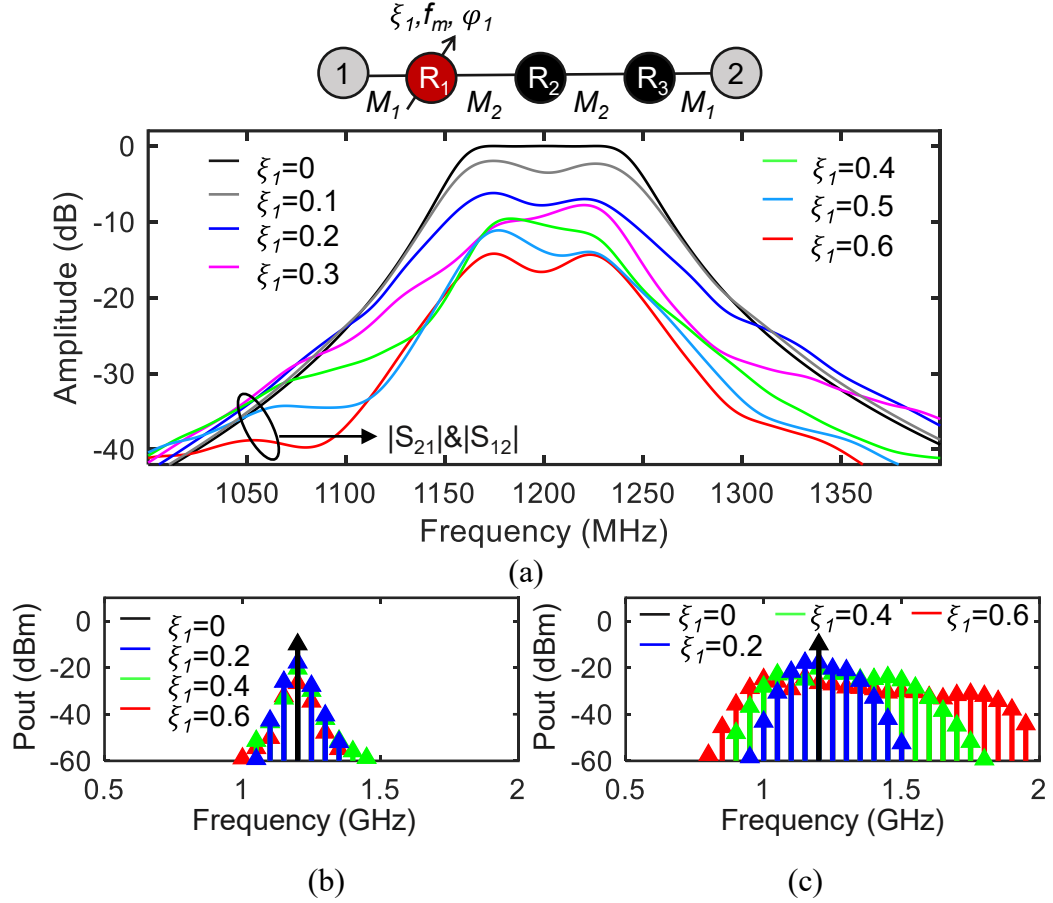


Figure 4.6. Three-resonator array having one over-modulated resonator (red circle with arrow) and two unmodulated resonators (black circles). (a) CRD and simulated S-parameters with increasing modulation index ξ_l between 0–0.6. (b), (c) Output spectra under -10 dBm excitation at Port 1 and Port 2, respectively while altering ξ_l from between 0–0.6.

Next, an enhanced switching configuration is demonstrated by over-modulating additional resonators within the three resonator-network. Figure 4.7 shows the CRD and the S-parameter response when R_1 and R_2 are over-modulated ($\xi_l = \xi_2 = 0.24$) and R_3 is unmodulated. Compared to the case when only one resonator is over-modulated [see Figure 4.6 (a)], the isolation improves to over 30 dB [see Figure 4.7 (a)]. This is due to the multi-path destructive interference enabled with the modulation of R_2 . The simulated spectrum in Figure 4.7 (b) and Figure 4.7 (c) also verify the high isolation at the center frequency and the increase of order to 8 for the IM products. To maintain symmetric isolation in both directions (i.e., $|S_{21}| = |S_{12}|$), the same modulation phase is applied to R_1 and R_2 to avoid unilateral transmission introduced by spatial modulation. As expected, by modulating additional resonators in the in-line resonator network, a larger isolation can be obtained using a smaller modulation index. This is important

for the realization of the over-modulated STM resonators due to realistic varactors having a limited modulation depth.

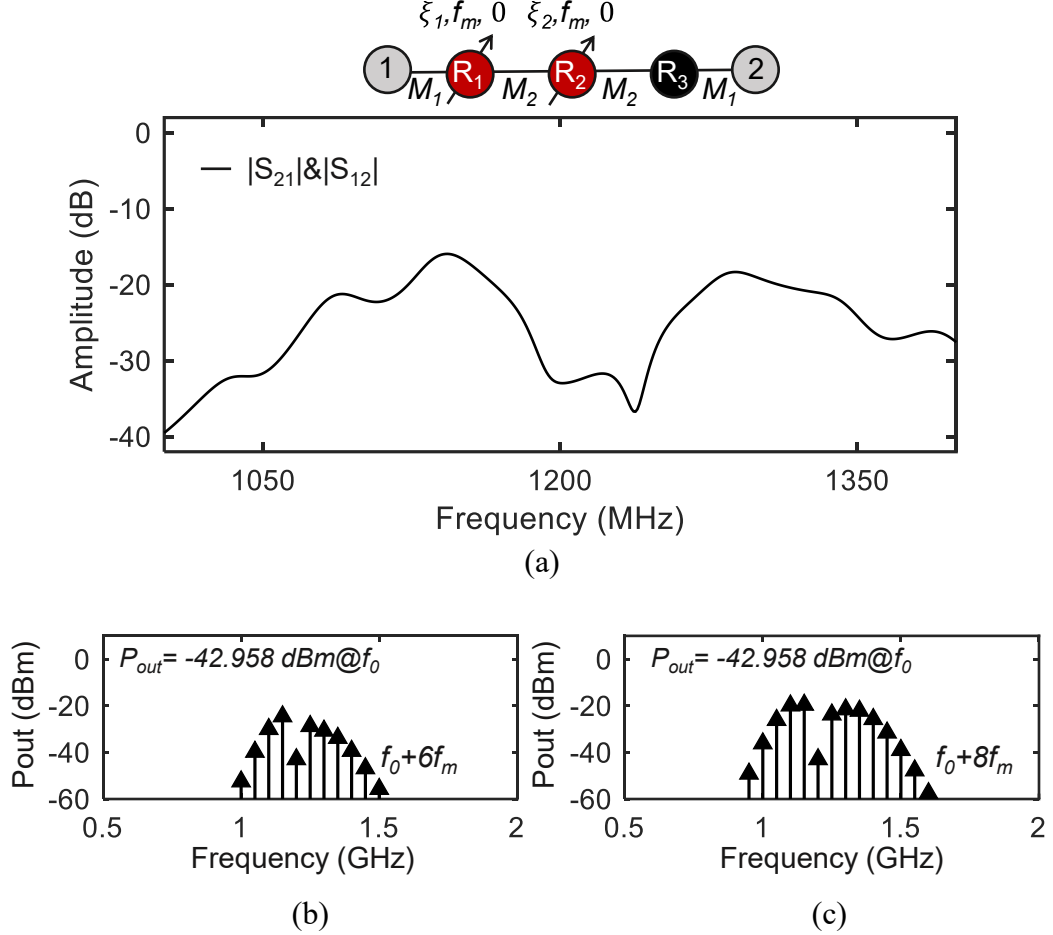


Figure 4.7. (a) CRD and simulated S-parameters when R_1 and R_2 are over-modulated and R_3 is unmodulated. (b), (c) Output spectra under -10 dBm excitation at Port 1 and Port 2, respectively.

4.1.4 Design examples of star-type multi-port signal routing block

This section introduces design examples for the star-type multi-port signal routing block, and their responses are presented in Figure 4.8, Figure 4.9 and Figure 4.10. Specifically, Figure 4.8 examines cases where each transmission path contains three resonators ($k = 1$), while Figure 4.9 and Figure 4.10 investigate configurations with five resonators per path ($k = 2$).

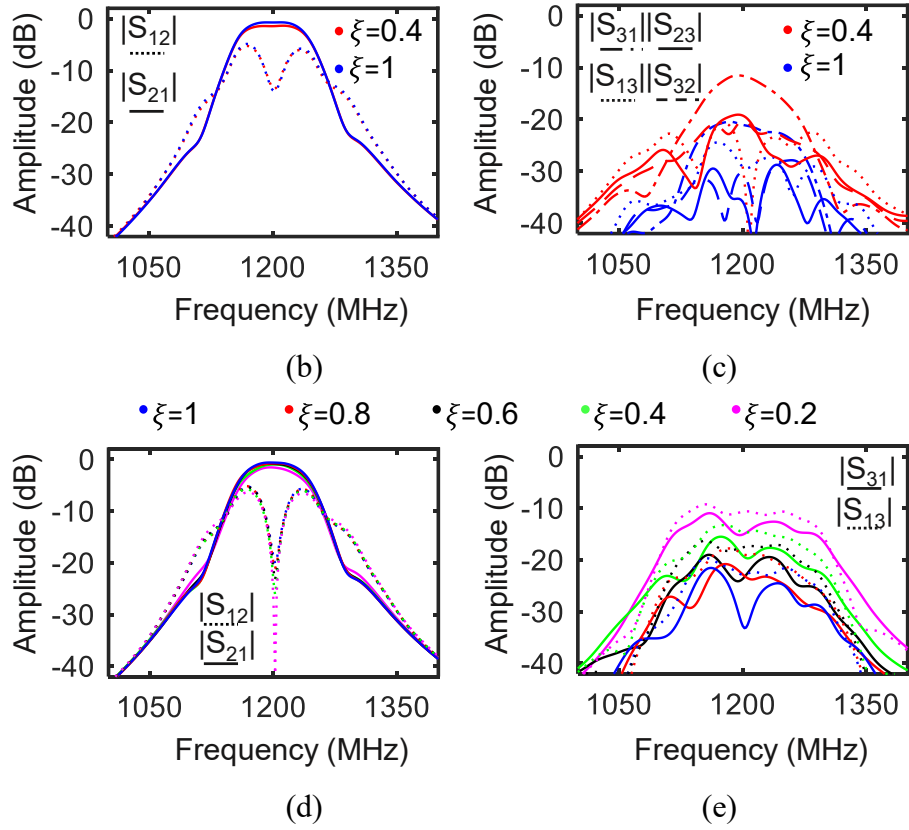
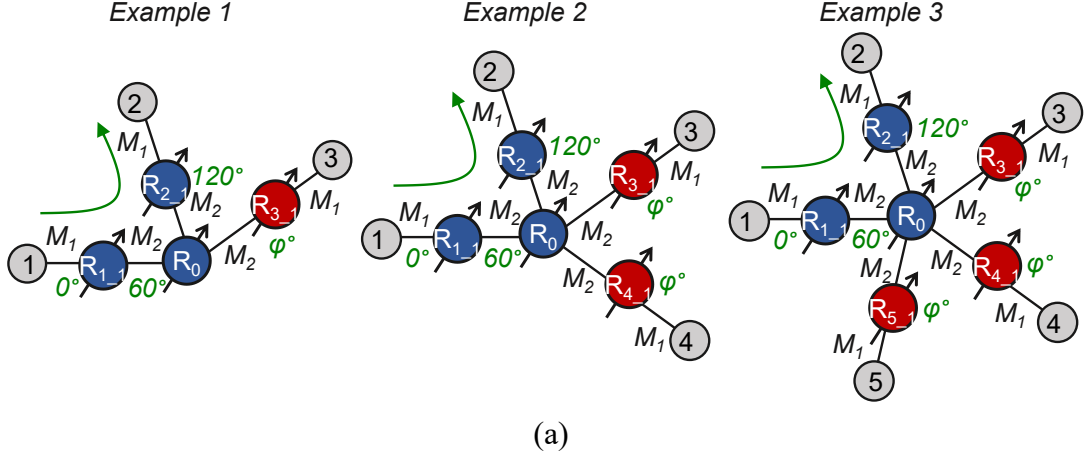
Three examples (Examples 1–3) are analyzed in SPMT-BPFI working mode using the CRDs shown in Figure 4.8 (a), with port numbers $N=3,4$, and 5 , respectively. In all cases, Port 1 and Port 2 are selected as the active transmission pair, while the

remaining ports are configured for isolation. To enable transmission from Port 1 to Port 2 (indicated by the green arrow), resonators R_{1_1} , R_0 , and R_{2_1} are moderately-modulated with phases of 0° , 60° , and 120° , respectively. The modulation parameters of the transmission path are chosen by parametric sweeping as discussed in Chapter 2 and Chapter 3 aiming to obtain higher isolation at the center frequency with a reasonable IL.

Example 1 in Figure 4.8 (a) illustrates the case with $k=1$, $N=3$, forming a three-port network with four resonators. Resonator R_{3_1} , which connects to Port 3, is over-modulated to suppress unwanted signal propagation. However, due to the unilateral nature of STM, the inactive paths involving Port 3 (i.e., Port 1 and Port 3 and Port 2 and Port 3) also exhibit directionality. If the modulation phase of R_{3_1} is not properly selected—for instance, if it continues the phase progression from R_0 (e.g., φ set to 120°)—constructive phase alignment may occur along the unintended path from Port 1 to Port 3, leading to insufficient isolation and an elevated $|S_{31}|$. Figure 4.8 (b) and Figure 4.8 (c) show the simulated S-parameters for $\varphi = 120^\circ$, with modulation indices $\xi=0.4$ and $\xi=1$, respectively. As observed in Figure 4.8 (c), $|S_{31}|$ is noticeably higher than the other isolation paths due to the constructive STM phase alignment, although increasing ξ to 1 can reduce $|S_{31}|$ to approximately -20 dB. However, achieving such high modulation depth is often impractical in varactor-based implementations and can cause additional power leakage into higher-order IM products.

To avoid constructive STM progression in any isolation path, the phase of the over-modulated resonators φ need to be identical or differ by 180° from the phase of the central resonator R_0 . In this configuration, the resulting S-parameter symmetry satisfies $|S_{13}| = |S_{32}|$ and $|S_{31}| = |S_{23}|$, ensuring that no unintended directional transmission is favored. Figure 4.8 (d) and Figure 4.8 (e) present the simulated S-parameters when $\varphi = -120^\circ$ and ξ varies from 0.2 to 1. The results show that IL from Port 1 to Port 2 decreases as ξ increases, while isolation improves across all undesired paths. When $\xi=0.6$, isolation levels greater than 20 dB are achieved in all isolated paths, confirming effective suppression of unwanted transmission. Finally, Figure 4.8 (f) and Figure 4.8 (g) show the simulated S-parameters of all three examples from Figure 4.8 (a), using the same modulation configuration for all cases: $f_m=48$ MHz, $\varphi=-120^\circ$. The modulation index for the moderate-modulated resonators and over-modulated resonators are 0.06 and 0.6, respectively. As shown in Figure 4.8 (f), clear unilateral

transmission is achieved with $D > 20$ dB at the center frequency. Across all configurations, the simulated isolation exceeds 17 dB in every isolation path as indicated in Figure 4.8 (g), demonstrating the scalability in port count of the proposed star-type blocks.



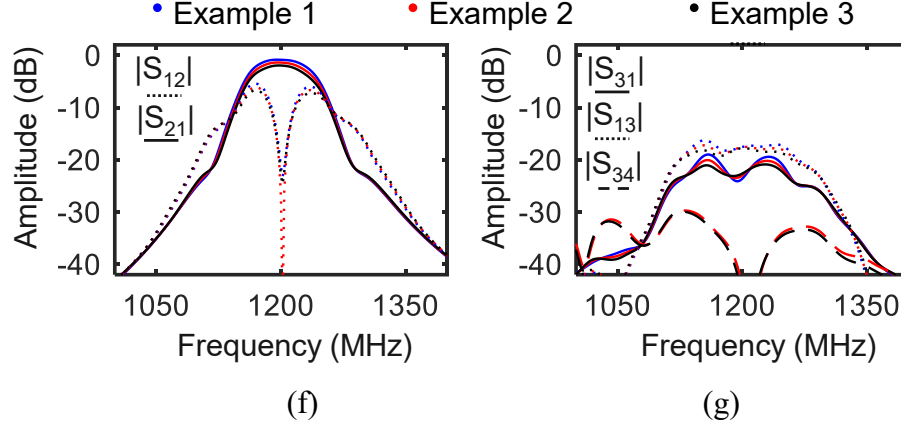


Figure 4.8. Parametric study of star-type blocks with $k = 1$. The coupling coefficients of the CRD are: $M_1=1.0836$, $M_2= 1.0317$. (a) CRDs for Examples 1–3 with $N = 3, 4, 5$. (b), (c) Simulated S-parameters between the two active ports and the isolated ports, respectively in Example 1 when $\varphi=120^\circ$. (d), (e) Simulated S-parameters between the two active ports and the isolated ports, respectively in Example 1 when $\varphi=-120^\circ$. (f), (g) Simulated S-parameters between the two active ports and the isolated ports, respectively in Examples 1, 2 and 3 when $\varphi=-120^\circ$.

Figure 4.9 presents a design example with $k=2$ and $N=3$, which realizes a SPDT switch exhibiting a fifth-order BPF response. As discussed in section 4.1.3, increasing the number of the over-modulated resonators in the isolated branches can improve the signal suppression due to enhanced intermodulation-induced power scattering and interference. The simulated results in Figure 4.9 (b) show a clear fifth-order unidirectional BPF response between Port 1 and Port 2, with an IL of 0.7 dB and D of 30 dB, based on the modulation parameters provided in the figure caption. The corresponding isolation performance is shown in Figure 4.9 (b), where the S-parameters of the isolation paths indicate more than 20 dB of suppression across all directions. It is also noted that compared to the $k=1$ examples in Figure 4.8, the increased number of over-modulated resonators in this configuration allows equivalent or improved isolation to be achieved using a lower modulation index.

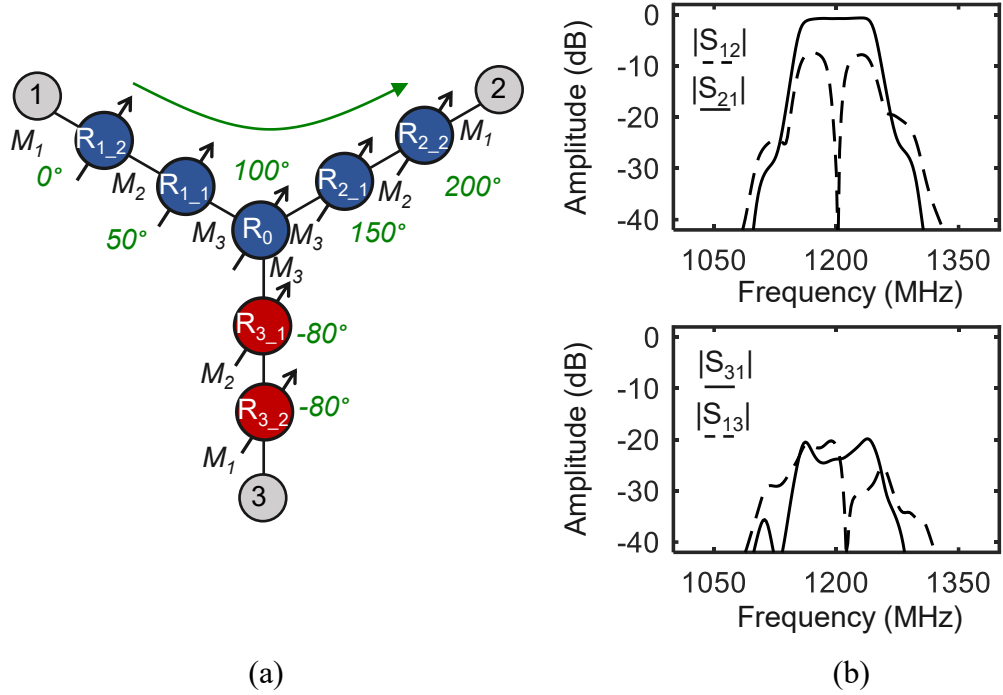


Figure 4.9. Star-type network with $k = 2$ and $N = 3$ operating as a SPDT-BPFI. The coupling coefficients and modulation parameters are $M_1=1.0146$, $M_2=0.866$, $M_3=0.636$, $\text{FBW}=0.07$, $f_m=52$ MHz. The modulation index for the moderately-modulated resonators and over-modulated resonators are 0.042 and 0.42, respectively. (a) CRD and modulation scheme. (b) Simulated S-parameters between the two active ports and the isolated ports, respectively.

The CRDs and modulation schemes of the PD-BPFI and the BPFC modes are shown in Figure 4.10 (a) and Figure 4.10 (b), respectively. The coupling coefficients in these examples are identical to those in Figure 4.9. To realize the PD-BPFI function [Figure 4.10 (a)], the modulation phases are configured to follow a consistent spatial progression along both output paths (indicated by green arrows). The phase progression between Port 2 and Port 3 decreases and then increases, isolating the transmission in both directions. This technique effectively suppresses signal transmission between the ports, eliminating the need for resistive terminations as typically used in conventional power dividers. As shown in Figure 4.10 (c), the power from Port 1 is evenly distributed to Ports 2 and 3 with a 5th-order BPFI response and an IL of 3.8 dB and $D > 25$ dB inside passband. Furthermore, the isolation between Port 2 and Port 3 is more than 20 dB.

To implement the BPFC mode of operation [Figure 4.10 (b)], the central resonator R_0 is deactivated (non-resonating), effectively separating the three active branches. By applying progressive modulation phases of 0° , 120° , and 240° to the first

resonator in each path (R_{1_1} , R_{2_1} , and R_{3_1}), a clockwise circulation behavior is established. The static resonators in each path not only increase the filter order for enhanced selectivity but also act as band-stop elements at IM frequencies, preventing leakage of IM power to the RF ports and thereby reducing the loss. As shown in the simulation results, a low IL of 0.4 dB is achieved in the transmission paths, whereas D is above 30 dB.

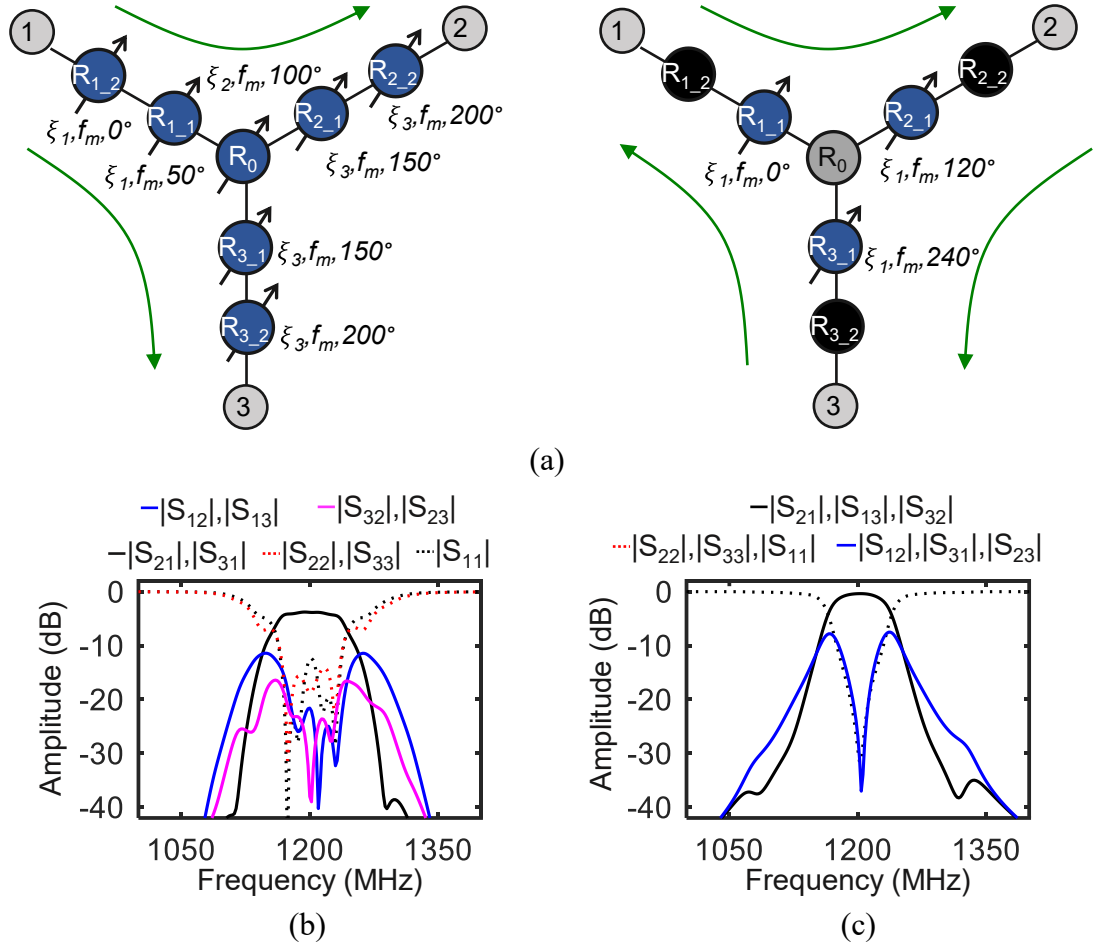


Figure 4.10. (a) CRD and modulation scheme for the PD-BPFI mode of operation. The modulation parameters are: $f_m=30$ MHz, $\xi_1=0.058$, $\xi_2=0.088$ and $\xi_3=0.072$. (b) CRD and modulation scheme for the BPFC mode of operation. The modulation parameters are: $f_m=90$ MHz, $\xi_1=0.124$. (c), (d) Simulated S-parameters in the PD-BPFI and BPFC mode of operation, respectively.

4.1.5 Design examples of the box-type multi-port signal routing block

The two working modes of the proposed box-type block are analyzed, and the modulation schemes and the simulated results are presented in Figure 4.11 and Figure 4.12. Figure 4.11 (a) and (b) shows the CRDs and modulation scheme in CR-BPFI

mode to enable the horizontal dual-channel transmissions and single-channel horizontal transmission, respectively. The simulated results for two selected horizontal channels are presented in Figure 4.11 (c) and Figure 4.11 (d). By overmodulating R_7 and R_8 , the two channels become independently-reconfigurable, enabling unilateral transmission from Port 1 to Port 2 and from Port 3 to Port 4. The modulation parameters are: $f_m=36$ MHz, $\zeta_1=\zeta_2=\zeta_3=\zeta_4=\zeta_5=\zeta_6=0.054$, $\zeta_7=\zeta_8=0.4$, and $\varphi_1=\varphi_4=0^\circ$, $\varphi_2=\varphi_5=60^\circ$, $\varphi_3=\varphi_6=120^\circ$, $\varphi_7=\varphi_8=0^\circ$. As shown in Figure 4.11 (c), the simulated forward IL is 0.7 dB and $D>25$ dB. As shown in Figure 4.11 (d), the isolation levels between the two channels are greater than 20 dB across the passband. Additionally, by over-modulating all of the resonators between Port 3 and Port 4, the entire branch is deactivated, as demonstrated by the CRD in Figure 4.11 (b). The modulation parameters are: $f_m=36$ MHz, $\zeta_1=\zeta_2=\zeta_3=0.054$, $\zeta_4=\zeta_5=\zeta_6=0.25$, $\zeta_7=\zeta_8=0.4$, and $\varphi_1=0^\circ$, $\varphi_2=60^\circ$, $\varphi_3=120^\circ$, $\varphi_4=\varphi_5=\varphi_6=\varphi_7=\varphi_8=0^\circ$. Forward IL is 0.8 dB with $D>22$ dB as indicated in Figure 4.11 (e). As shown in Figure 4.11 (f), all the isolated channels between Port 1 and Port 3, Port 1 and Port 4, Port 2 and Port 3, Port 2 and Port 4, Port 3 and Port 4, show rejection levels exceeding 17 dB across the passband.

Furthermore, PD-BPFI can be easily achieved in the box-type block following the modulation scheme in Figure 4.12 (a). The modulation parameters are: $f_m=41$ MHz, $\zeta_1=0$, $\zeta_2=\zeta_3=\zeta_7=\zeta_4=0.076$, $\zeta_5=\zeta_6=\zeta_8=0.41$, and $\varphi_2=\varphi_6=0^\circ$, $\varphi_3=\varphi_8=60^\circ$, $\varphi_7=180^\circ$, $\varphi_4=\varphi_5=240^\circ$. Under this configuration, the input power from Port 1 can be equally divided to Port 2 and Port 3 with an IL of 4.2 dB and $D>20$ dB as indicated by the simulated S-parameters presented in Figure 4.12 (b). The simulated isolation levels are larger than 24 dB as shown in Figure 4.12 (c).

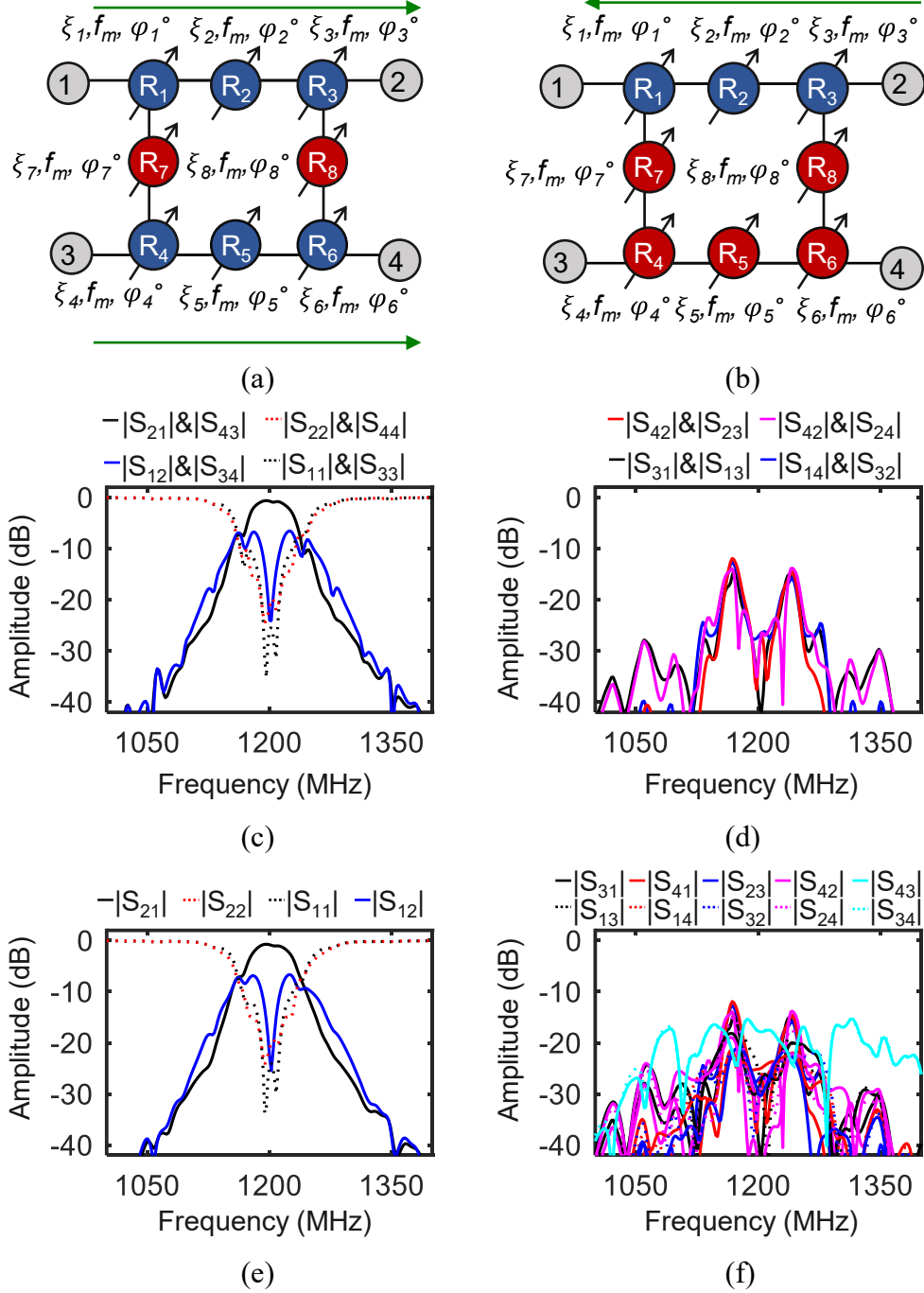


Figure 4.11. (a), (b) CRD and modulation scheme to enable the dual-channel horizontal transmissions and one channel horizontal transmission, respectively. The coupling coefficients are: $M_1=1.0836$, $M_2= 1.0317$. FBW=4% (c), (d) Simulated S-parameters when two transmission paths are configured as Port1 to Port 2 and Port 3 to Port 4. (e), (f) Simulated S-parameters when one transmission path is configured as Port 1 to Port 2.

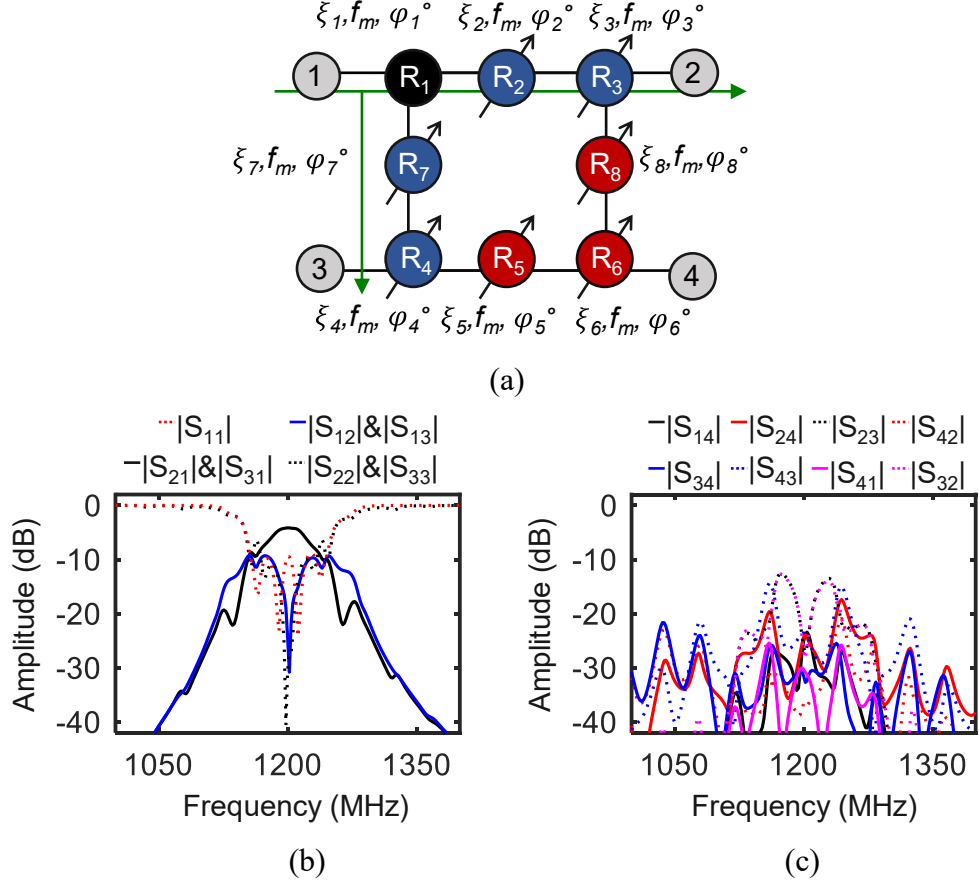


Figure 4.12. (a) Modulation scheme to enable the PD-BPFI mode. (b), (c) Simulated S-parameters when power division is configured from Port 1 to Port 2 and Port 3.

4.2 Experimental Validation

To verify the proposed multi-port multi-functional RF signal routing network concept, two prototypes (Prototype 1, 2) were designed using circuit and electromagnetic co-simulations in Keysight ADS. They were manufactured in-house on a Rogers 5880 substrate (relative dielectric constant: 2.2, loss tangent: 0.0009, thickness: 1.57 mm) and they were experimentally tested using a Keysight PNA-X 4 port analyzer. These include a star-type block (Prototype 1) exhibiting the following three modes of operation: (i) SPDT-BPFI, (ii) PD-BPFI, (iii) BPFC and a box-type block (Prototype 2) enabling: (i) a CR-BPFI and (ii) a PD-BPFI modes of operation.

4.2.1 Prototype 1: star-type block

Prototype 1 implements the star-type block with $k=2$ and $N=3$, as shown in Figure 4.13. The central resonator R_0 is an RF-switched resonator that enables three different modes of operation: (i) SPDT-BPFI, (ii) PD-BPFI and (iii) BPFC. All

peripheral resonators (R_{1_1} , R_{1_2} , R_{2_1} , R_{2_2} , R_{3_1} , R_{3_2}) are realized using half-wavelength varactor-loaded transmission lines, with DC and AC biasing networks indicated in the yellow boxes in Figure 4.13. Specifically, each biasing branch includes a low-pass filter (L_{LPF} and C_{LPF}) after the AC source, which allows the modulating AC signals to reach the varactors while preventing high-frequency RF leakage into the modulation source. The cutoff frequency of this low-pass filter is chosen to be higher than the modulation frequency (to pass the AC biasing signal) but lower than the RF frequency (to provide isolation between modulation and RF paths). In addition, C_{DC1} and C_{DC2} act as DC blocks to prevent DC from entering the RF circuitry and the AC sources. The DC bias sets the nominal capacitance, while the AC modulation signals induce the required time-varying capacitances. R_0 is implemented using a lumped LC tank (grey box in Figure 4.13) and controlled via two PIN diodes, D_1 and D_2 . When D_1 is ON and D_2 is OFF, R_0 functions as a resonating node, supporting the SPDT-BPFI and PD-BPFI modes of operation. Conversely, when D_1 is OFF and D_2 is ON, R_0 becomes a non-resonating node, enabling the BPFC functionality. The prototype was designed at 1.12 GHz with a FBW of 7%. The detailed geometry of the prototype and its photograph are presented in Figure 4.13. The peripheral resonator dimensions and coupling coefficients M_1 and M_2 are extracted using EM-parameter simulations using conventional filter design techniques as in [128]. The M_3 couplings are implemented using first-order high pass π -networks whose inductance absorbed in the resonators R_{1_1} , R_{2_1} and R_{3_1} and a central shared inductor L_{m3} .

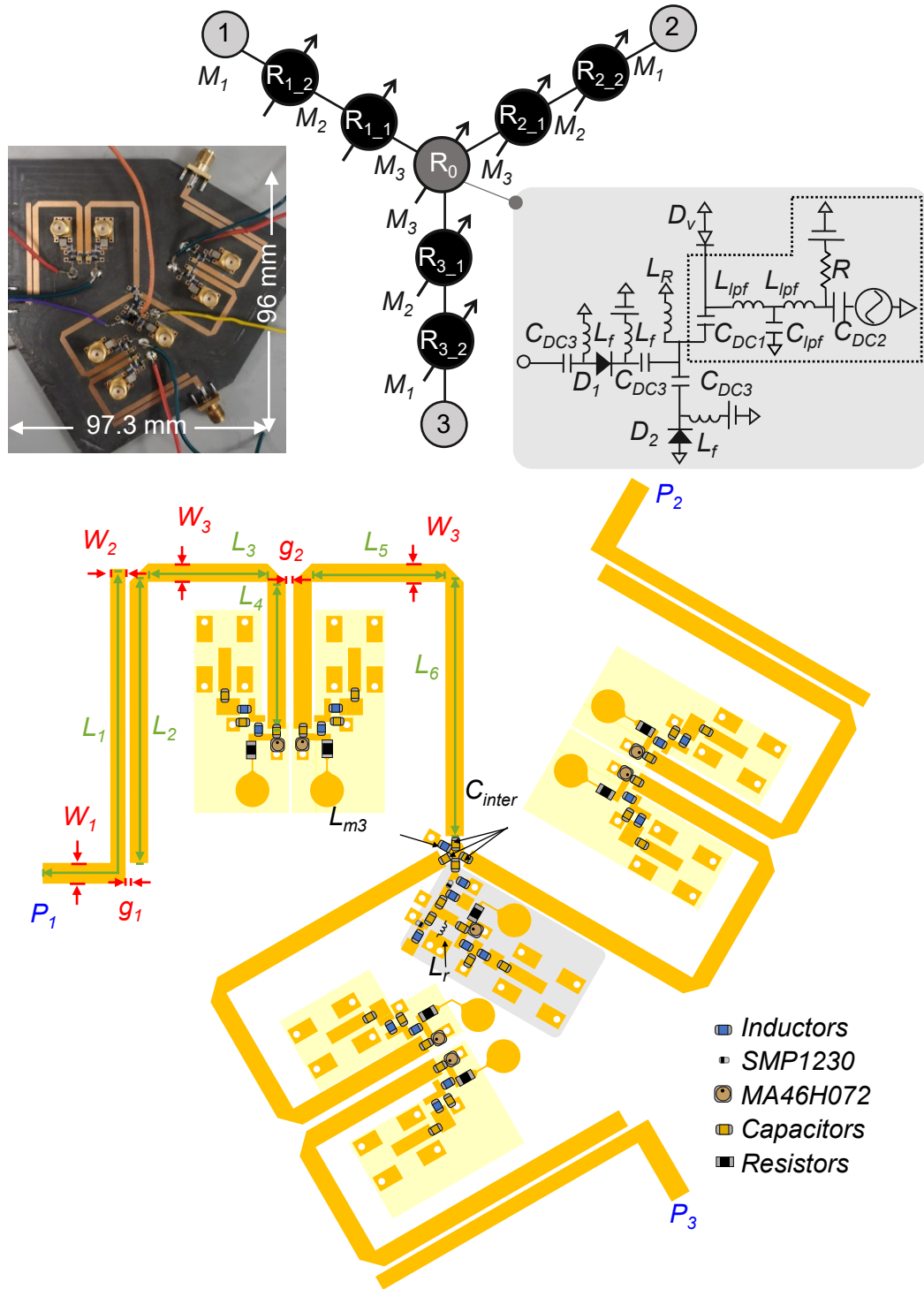


Figure 4.13. Layout and photograph of the star-type block-Prototype 1. The dimensions (all in mm) and the lumped elements are selected as follows: $W_1=2.452$, $W_2=1.796$, $W_3=2.17$, $L_1=44.78$, $L_2=33.2$, $L_3=14.093$, $L_4=17.33$, $L_5=15.76$, $L_6=30$, $g_1=0.5$, $g_2=0.9$, $L_f=100$ nH, $L_{lpf}=50$ nH, $L_{m3}=18$ nH, $L_r=2.5$ nH, $C_{lpf}=27$ pF, $C_{DC1}=47$ pF, $C_{DC2}=680$ pF, $C_{DC3}=50$ pF, $C_{inter}=0.2$ pF.

The RF measured S-parameters for the three modes of operation are shown in Figure 4.14 – Figure 4.16. First, for the SPDT-BPFI working mode, the measured results are provided in Figure 4.14 with three different switching states, Port 1 to Port 2, Port 1 to Port 3, and Port 3 to Port 2. The active transmission paths (Port 2 to 1, Port 1 to 3, and Port 3 to 2) exhibit a fifth-order BPFI response with a passband IL between 5.18–5.5 dB and $D=30$ dB. The isolation for the corresponding isolated paths is greater than 20 dB across the passband frequencies. The RF measured performance when operating in its PD-BPFI mode with Port 1 as the input is shown in Figure 4.15. The transmission from Port 1 to both output Ports 2 and 3 shows a fifth-order BPFI response with a measured IL of 4.7 dB and $D=18$ dB. The isolation between Port 2 and Port 3 is better than 30 dB at the center frequency. Finally, for the BPFC mode of operation, the corresponding RF measured performance is provided in Figure 4.16. As it can be seen, it exhibits a clear filtering circulating behavior with a minimum in-band IL of 4~4.2 dB in all transmission directions and a D of more than 27 dB in the reverse direction at center frequency.

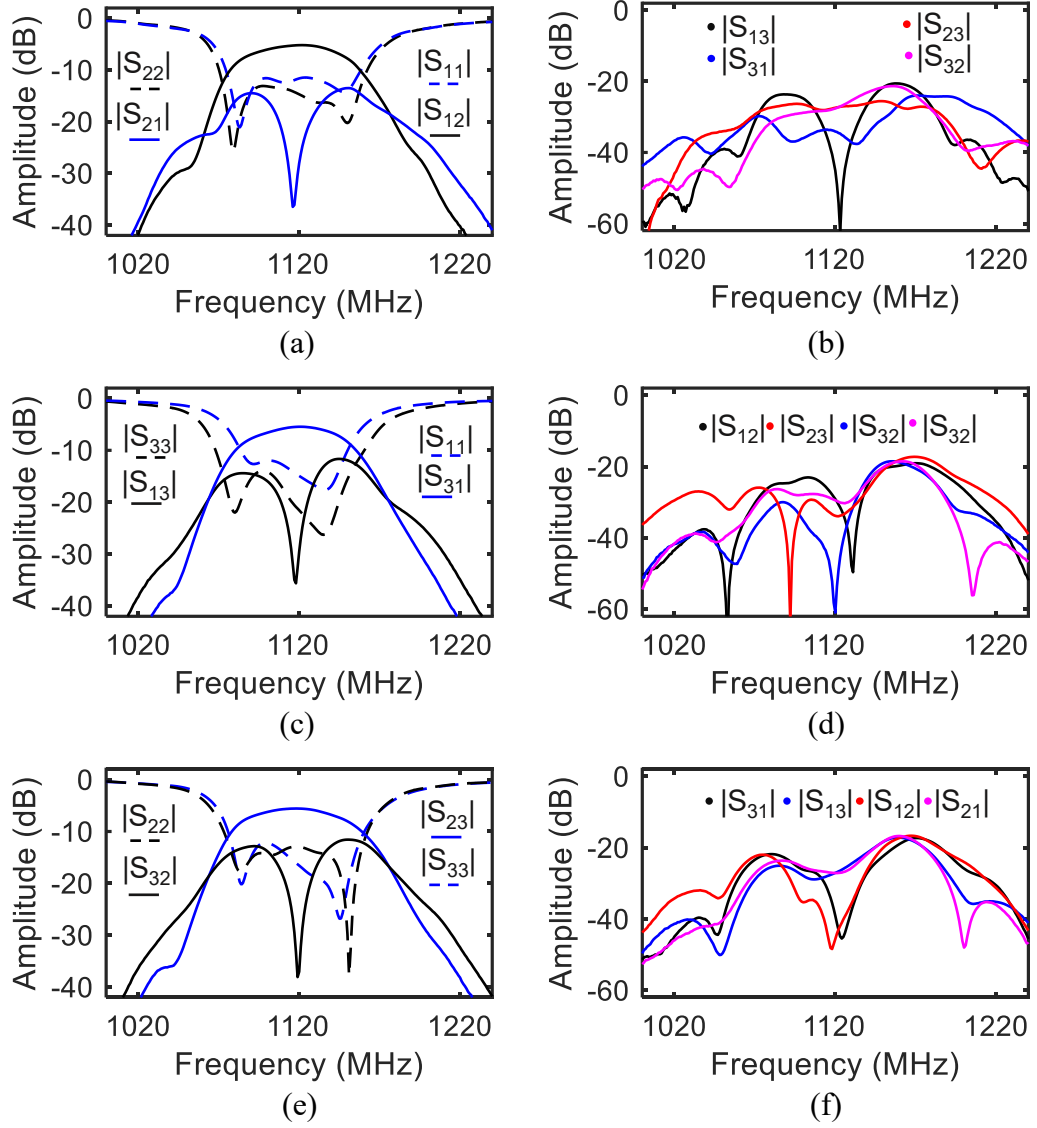


Figure 4.14. RF-measured S-parameters of Prototype 1 when configured in the SPDT-BPFI mode of operation. (a), (b) Signal transmission is enabled from Port 2 to Port 1. (c), (d) Signal transmission is enabled from Port 1 to Port 3. (e), (f) Signal transmission is enabled from Port 3 to Port 2.

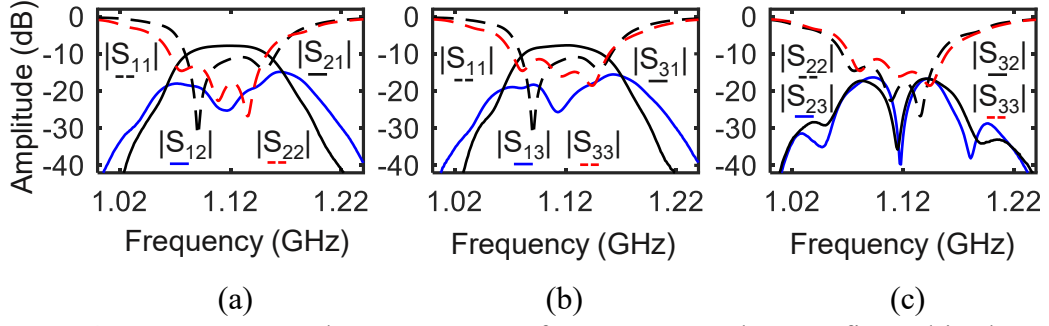


Figure 4.15. RF-measured S-parameters of Prototype 1 when configured in the PD-BPFI mode of operation. (a) Signal transmission between Port 1 and Port 2. (b) Signal transmission between Port 1 and Port 3. (c) Signal transmission between Port 2 and Port 3.

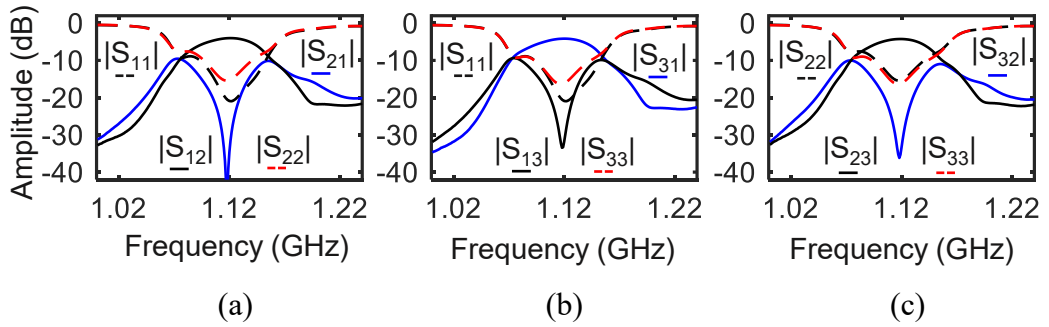


Figure 4.16. RF-measured S-parameters of Prototype 1 when configured in the BPFC mode of operation. (a) Signal transmission between Port 1 and Port 2. (b) Signal transmission between Port 1 and Port 3. (c) Signal transmission between Port 2 and Port 3.

4.2.2 Prototype 2: box-type block

Prototype 2 implements the box-type block and was designed for a center frequency of 1.2 GHz with a 7% FBW. The physical layout and the manufactured prototype are shown in Figure 4.17. Resonators R_1 , R_3 , R_4 , and R_6 are implemented as half-wavelength varactor-loaded microstrip lines, while R_2 , R_5 , R_7 , and R_8 are constructed using half-wavelength resonators with varactors loaded at both ends. A similar biasing scheme as the one used in Prototype 1 is also used in this case. For the double-loaded resonators, as shown in the grey box of Figure 4.17, the DC signals are fed directly through resistors R to the varactors on both sides of the resonator, with C_{DCI} blocking DC from the RF path. The AC modulation is injected at the resonator's center through an inductor L_{feed} . This inductor presents high impedance at the RF frequency, preventing RF leakage into the modulation port, while offering low

impedance to the modulation frequency so that the AC signals can be applied symmetrically to the varactors on both sides of the resonator.

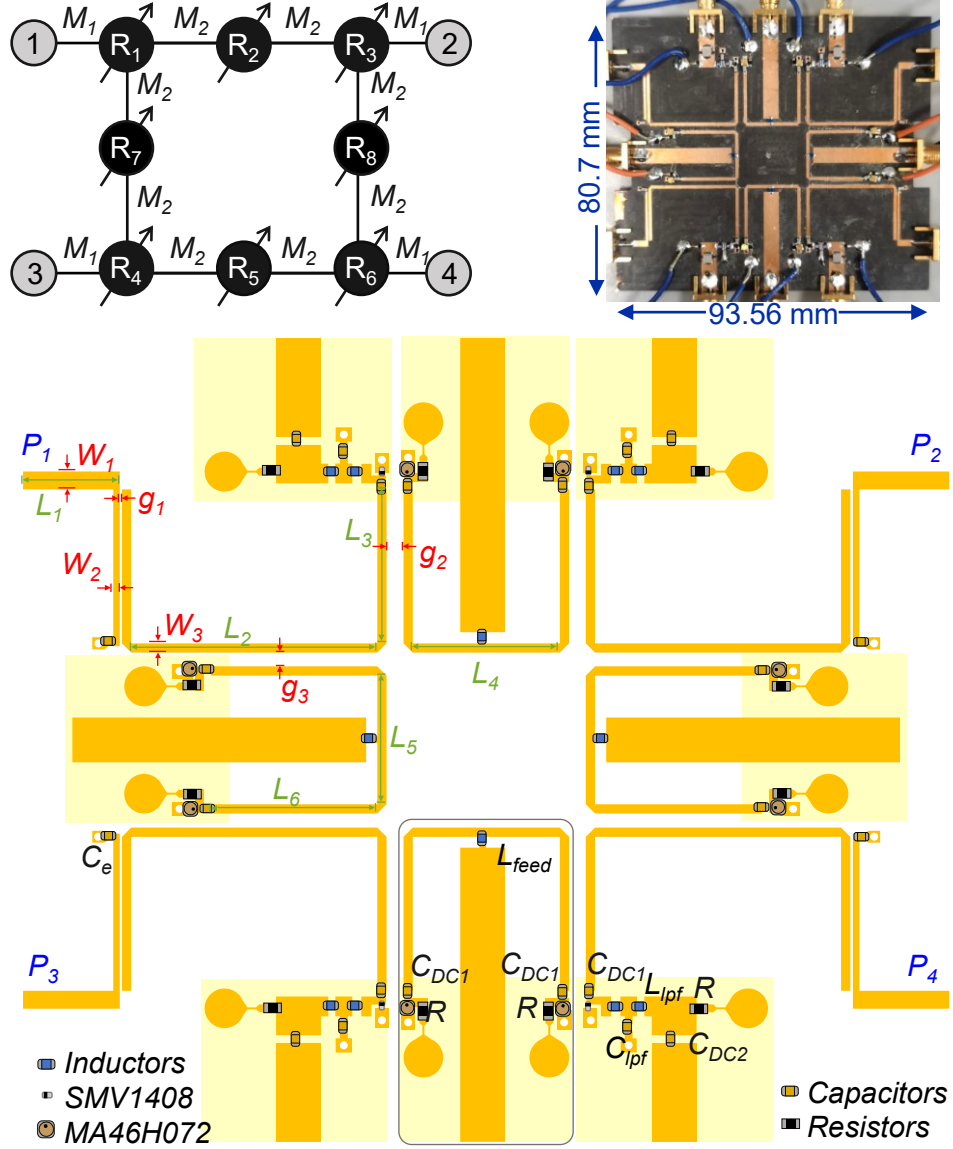


Figure 4.17. Layout and photograph of Prototype 2. The dimensions (all in mm) and the lumped elements are selected as follows: $W_1=1.795$, $W_2=0.65$, $W_3=0.929$, $L_1=9.04$, $L_2=24.716$, $L_3=15.45$, $L_4=14.77$, $L_5=12.9$, $L_6=16.716$, $g_1=0.25$, $g_2=1.7$, $g_3=1.4$, $L_{lpf}=50$ nH, $L_{feed}=180$ nH, $C_{lpf}=27$ pF, $C_{DC1}=47$ pF, $C_{DC2}=680$ pF, $C_e=1.1$ pF, $R=10$ MOhm.

The two functional modes of the box-type prototype i.e., CR-BPFI and PD-BPFI, were experimentally characterized. In the CR-BPFI mode of operation, concurrent signal transmission is enabled along two horizontal paths—from Port 2 to Port 1 and from Port 4 to Port 3—by over-modulating R_7 and R_8 with their measured S-

parameters shown in Figure 4.18 (a) and Figure 4.18 (b). The RF-measured results in Figure 4.18 (a) indicate in-band ILs of 4.1 dB (from Port 2 to Port 1) and 3.8 dB (from Port 4 to Port 3) in the forward direction and a D of 27 dB for the two transmission channels. The measured isolation levels between the two channels are greater than 22 dB as indicated in Figure 4.18 (b). When facilitating a single-channel transmission between Port 3 and Port 4 in the CR-BPFI mode of operation, resonators R_1 , R_2 , and R_3 are also over-modulated, effectively deactivating the RF signal channel between Port 1 and Port 2. As shown in Figure 4.19 (a), the measured IL from Port 4 to Port 3 is 3.8 dB with a D of 26 dB and the transmission between Port 1 and Port 2 is less than -30 dB. Additionally, Figure 4.19 (b) shows that the transmission between Port 1 and Port 3, Port 1 and Port 4, Port 2 and Port 3, Port 2 and Port 4 are less than -30 dB. Figure 4.20 presents the measured RF performances of the PD-BPFI mode, where Port 4 is isolated and the input power at Port 1 is equally divided between Ports 2 and 3. As shown in Figure 4.20 (a) and Figure 4.20 (b), the measured ILs from Port 1 to Ports 2 and 3 are 4.3 dB and 4.8 dB, respectively, with D being greater than 22 dB. The measured isolation levels between Port 2 and Port 4, Port 2 and Port 3, Port 3 and Port 4, Port 1 and Port 4 are more than 20 dB as illustrated in Figure 4.20 (c).

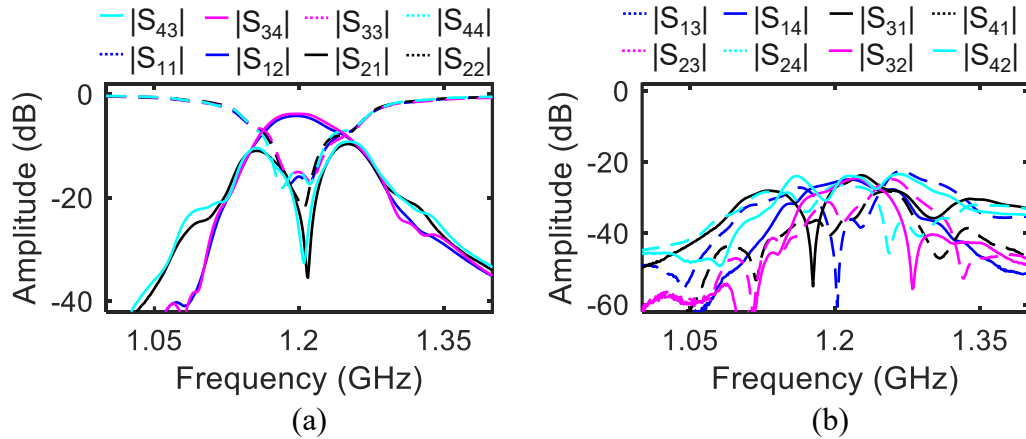


Figure 4.18. RF-measured S-parameters of Prototype 2 when working at CR-BPFI mode and dual-channel signal transmission is enabled from Port 2 to Port 1 and from Port 4 to Port 3. (a) Signal transmission of the two horizontal paths. (b) Signal transmission of the isolated paths.

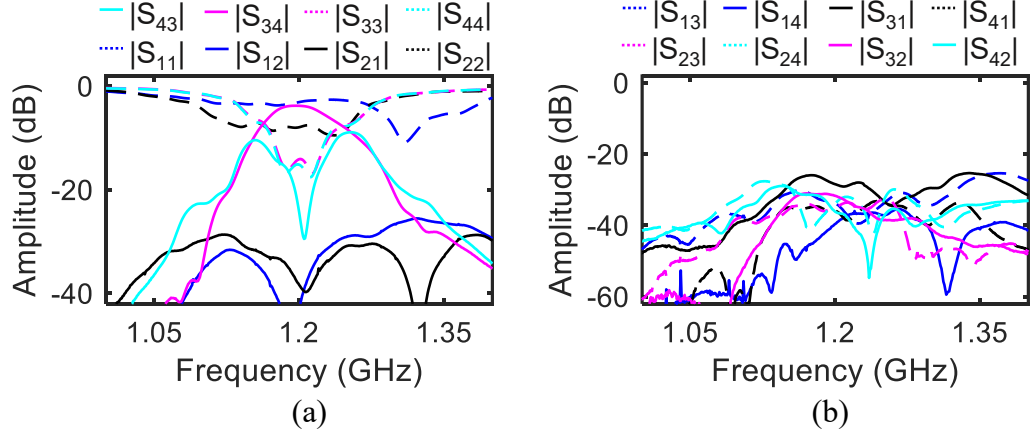


Figure 4.19. RF-measured S-parameters of Prototype 2 when working at CR-BPFI mode and single-channel signal transmission is enabled from Port 4 to Port 3. (a) Signal transmission of the two horizontal paths. (b) Signal transmission of the isolated paths.

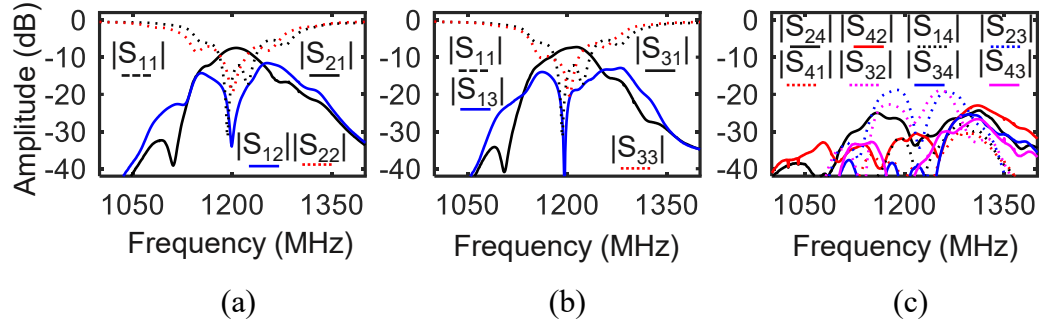


Figure 4.20. RF-measured S-parameters of Prototype 2 in PD-BPFI working state. (a) Signal transmission between Port 1 and Port 2. (b) Signal transmission between Port 1 and Port 3. (c) Signal transmission between Port 2 and Port 4, Port 2 and Port 3, Port 3 and Port 4, Port 1 and Port 4.

4.2.3 Comparison with state-of-the-art multi-functional RF components

A comparison of the proposed multi-port multi-functional RF signal routing network concept with state-of-the-art co-designed multi-functional RF components is provided in Table 4.1. Both of the proposed prototypes (star-type, Prototype 1 and box-type, Prototype 2) exhibit high levels of multi-functionality and flexibility, demonstrating the highest number of reconfigurable modes of operation in a single RF component. The reconfiguration between the multi-functional operational modes is based on a new type and fairly-simple RF switching mechanism using over-modulated STM resonators that can be also exploited to incorporate directionality. The proposed RF switching mechanism doesn't require additional external switching elements or

matching circuits, significantly simplifying the implementation of the proposed multi-functional multi-port RF networks.

Table 4.1. Comparison with state-of-art co-designed multi-functional RF components

Ref.	Mode	FBW	f_{cen} (GHz)	TF	IL (dB)	IS (dB)	TD	Channels	Size ($(\lambda_g * \lambda_g)$)
T.W: P1	ISO+SPDT	7%	1.118	5 th BPF	5.18~5.5	>20	Uni	\	0.54*0.54
	ISO+PD		1.118	5 th BPF	4.7	>25	Uni		
	C		1.118	3 rd BPF	4~4.2	>30	Uni	\	
T.W: P2	ISO+DC	7%	1.2	3 rd BPF	3.8~4.1	>22	Uni	2 or 1	0.55*0.48
	ISO+PD		1.2	3 rd BPF	4.3~4.8	>20	Uni	\	
[113]	SPDT	10%	1	3 rd BPF	0.97	>20	Bi	\	0.54*0.5
[114]	SP5T	4.5% $\pm 0.2\%$	1.86~2.25	3 rd BPF	2.45	>36	Bi	\	0.56
[115]	SPST	9%	0.78	2 nd BPF	1.8	>28	Bi	\	0.072*0.066
[116]	SPDT	123%	2.2	N/A	1.18	>20	Bi	\	0.3132
[103] Prototype 1	ISO+PD	N/A	1.66~1.98	3 rd BPF	3.7~4	>20	Uni	\	N/A
[104]	ISO+PD	7.3%	2.2~2.55	3 rd BPF	5.9	>30	Uni	\	1.46*0.89
[105]	ISO+PD	1.6%*	2.38~2.42	3 rd BPF	4.8~5.9	>20	Uni	\	2.72*2.72
[106]	ISO+PD	9%	1.39~1.76; 2.54	3 rd dual band	3.59~3.92; 2.87	>20	Uni; Bi	\	N/A
[68]	C	6.9%; 6.5%	0.185~0.206; 0.238~0.27	3 rd dual band	5.6; 5.5	>30	Uni	\	N/A
[99]	C	15.7%	0.825	3 rd BPF	1.7	>20	Uni	\	N/A
[101]	C	N/A	0.88~0.93	2 nd BPF	5.5~6.9	15~23	Uni	\	0.43*0.44
[100]	SPDT	N/A	0.85~1.28	4 th BPF	3.6~6.4	>36	Bi	\	0.578*0.578
	C		1.06~1.16	2 nd BPF	5.2~6.2	>12.47	Uni	\	
[119] Filter 1	DC	4.1%~ 4.7%	1.88~2.81	2 nd BPF	2.27~3.29	>19.8	Bi	2	0.28*0.28
	PD	N/A	1.7~2.8	2 nd BPF	1.44~2.35	>10	Bi		
[120]	DC	12.4% ~12.8%	2.5	3 rd BPF	<1.54	>22.3	Bi	2	0.69*0.69

(T.W: This work, P1: Prototype 1, P2: Prototype 2, TF: transfer function, IL: insertion loss, IS: isolation of the inactive paths, TD: signal transmission direction, ISO: isolator, PD: power divider, C: circulator, DC: dual-channel, Bi: bi-directional, Uni: uni-directional. *:10 dB FBW.)

More specifically, the star-type block (Prototype 1) provides distinct advantages across multiple operation modes. For example, when operating as a SPDT-BPFI, its measured isolation is comparable to that of SPDT filtering switches in [113] and [116]. Although [114] and [115] achieve higher IS, they rely on additional PIN diodes as switching elements. Besides, it supports higher-order filtering transfer functions in its active states, outperforming [113]–[115] in selectivity and port reconfigurability. While [116] exhibits better selectivity, it requires a large number of PIN diodes, limiting integration and scalability. Both [114] and the proposed star-type concept allow flexible port assignment; however, port scaling in [114] requires additional PIN diodes, increasing circuit complexity and power consumption. Notably, the proposed star-type concept is the only RF component that enables unilateral transmission between arbitrarily selected ports. Furthermore, when compared with PD-BPFIs such as the ones in [103]–[106], the proposed star-type block has lower IL than [104] and [105] and exhibits higher IS than [103], [105], and [106]. It also uniquely supports arbitrary port selection for input/output ports. When configured in its BPFC mode of operation, the proposed star-type block achieves lower IL than [68] and [101] and achieves the same IS as [68] and better IS than [99] and [101]. Moreover, none of the prior works integrate all three functions within a unified structure. Specifically, [100] demonstrates both SPDT switch and BPFC modes, but it has lower IS in BPFC mode, lacks power division functionality and only supports bidirectional transmission in its RF switch mode of operation.

The proposed box-type block (Prototype 2) demonstrates higher levels of flexibility when compared to state-of-the-art channel-reconfigurable dual-channel BPFs such as the ones reported in [119] and [120]. It achieves comparable isolation levels while providing superior reconfigurability by enabling both single-channel and dual-channel transmissions. Although [119] supports a power division function, its isolation between the two output ports is limited to 10 dB. Furthermore, neither [119] nor [120] support unilateral signal transmission. The measured IL in the proposed multi-functional RF component is mainly attributed to the low quality factors of the varactors (50–75), however it remains within a reasonable range and is comparable to that of recent co-designed multi-functional filters [104], [68], [101], [100], which report maximum IL values between 5.5 dB and 6.9 dB. Considering that the circuit

consolidates five distinct RF functions within a single platform, the obtained IL is relatively small and not a particular limitation of the proposed approach.

4.3 Conclusion

This chapter presented novel multi-port multi-functional RF signal routing networks capable of simultaneously integrating multi-port RF switching, bandpass filtering, power division, and isolation within unified structures. The proposed networks employ reconfigurable STM resonator arrays arranged into star- and box-type configurations. The star-type block supports three modes: (i) a SPMT-BPFI mode, (ii) a PD-BPFI mode, and (iii) a BPFC mode. The box-type block supports two modes: (i) a CR-BPFI mode, enabling concurrent dual-channel or single-channel transmission, and (ii) a PD-BPFI mode. The architecture is experimentally verified through two distinct prototypes. Prototype 1 implements the star-type block at 1.12 GHz with 7% FBW, demonstrating three operating modes with ILs ranging from 4.0 to 5.5 dB. Isolation performance exceeds 20 dB in all modes and surpasses 30 dB in BPFC mode. Prototype 2 implements the box-type block at 1.2 GHz with 7% FBW, achieving ILs between 3.8 to 4.8 dB in both modes. Additionally, the isolation levels are greater than 20 dB for all isolated paths.

Chapter 5: Conclusion and Future Work

5.1 Summary and Contributions

With the development of 6G wireless systems, emerging applications such as massive MIMO, IBFD, JCAS, and the Internet of Everything (IoE) are placing unprecedented performance and integration demands on RFFE. Consequently, future transceivers must not only meet low SWaP-C requirements but also integrate multiple RF functionalities while allowing for adaptability to a wide range of frequency bands, operating conditions, and application scenarios. This thesis has addressed these challenges by developing new multi-functional and multi-configurable RF circuit architectures and design methodologies based on STM resonator arrays to design compact, reconfigurable, and multi-functional RFFE components. Specifically, the key contributions of this thesis are as follows:

- Chapter 2 introduced the STM transversal-resonator-based BPF as a new RFFE architecture that overcomes the complexity and limited tunability typically associated with inline STM filters. In this chapter, a complete design methodology based on spectral network parameter matrices was developed, and four topologies based on transversal cell were analysed. The proposed topologies can be uniquely reconfigured in terms of: (i) center frequency, (ii) BW, (iii) out-of-band transmission zero location, (iv) propagation direction and (v) intrinsically switching off. The proposed concept has been validated by four BPFs, demonstrating wider tuning ranges in terms of center frequency and BW, enhanced isolation levels, and superior reconfigurability compared to the state-of-the-art BPFs based on inline topologies. The BPF contributions of this chapter have resulted in one IEEE conference and one IEEE journal publications that are now published in [122], [123].
- Chapter 3 builds upon the STM BPF concept by realizing multi-functional architectures that co-integrate VP and VA functionalities into the STM-based filter topologies. While prior work on filter co-design demonstrated integration with either VP or VA functionalities, very few implementations demonstrated both VP and VA in the same devices. Additionally, state-of-the-art filtering VPs or VAs often have compromised transfer function

shapes while the phase or amplitudes are tuned and can't facilitate unilateral transmission. In this chapter, a novel multi-functional BPFI was demonstrated, capable of simultaneously integrating the function of an RF isolator, a VP, a VA, and an RF switch. A detailed and rigorous analysis of the signal flow inside the multi-functional device was presented, demonstrating its scalability to facilitate various filtering transfer functions. The concept was validated at 700 MHz through two prototypes, showcasing the ability to obtain both single- and dual-band bandpass TFs. Measurement results showed a third order bandpass TF with continuous 360° phase tuning, 16 dB attenuation control with preserved filter responses, broad reflectionless BW ranging from 375 MHz to 1093 MHz, high directivity exceeding 20 dB and inherent switch-off capability. The third order dual-band prototype demonstrated a continuously tunable phase shift of 360° with a reflectionless BW spanning from 368 MHz to 1200 MHz, as well as a $D > 10$ dB. The multi-functional RF device contributions of this chapter have resulted in one IEEE conference and two IEEE journal publications that have been published in [124], [125], [126].

- Chapter 4 expanded the STM resonator concept further to address multi-port systems, where the requirements of next-generation wireless architectures extend beyond single-input/single-output components to include dynamic signal routing across multiple ports. In this chapter, STM resonator arrays were developed into star- and box-type blocks that can be dynamically configured through modulation-state control (unmodulated, moderately-modulated, and over-modulated), eliminating the need for separate RF switches. This approach enabled the same network to operate in multiple modes, including a single-pole multi-throw filtering switch mode, a channel-reconfigurable dual-path BPF mode, an RF isolator and circulator mode, and an RF power divider mode. The two blocks were experimentally validated through two microstrip line STM resonator-based prototypes. The prototypes showed that across all these modes the active paths demonstrate unilateral transmission with $IL < 5.5$ dB and $D > 18$ dB, the isolation exceeded 20 dB in the inactive paths. The proposed modulation control of STM resonators represents a substantial leap beyond conventional

moderately-modulated STM resonator architectures, showcasing how STM can facilitate reconfigurability and integration of RF signal routing functions in one compact structure. The multi-port and multi-functional RF devices contributions of this chapter have resulted in one IEEE conference and one IEEE journal publications that have been published in [127], [133].

5.2 List of Publications

Publications from this thesis:

- Z. Zhang and D. Psychogiou, “Multi-configurable bandpass filters with tune-all single-, dual-band transfer functions and reconfigurable directionality”, in 2023 IEEE/MTT-S International Microwave Symposium (IMS), San Diego, CA, USA, 2023.
- Z. Zhang and D. Psychogiou, “Non-reciprocal RF co-designed filtering phase shifters with continuously tunable phase shift”, in 2023 53rd European Microwave Conference (EuMC), Berlin, Germany, 2023, pp. 102-105.
- Z. Zhang and D. Psychogiou, “Incorporating directionality in transversal-resonator-based bandpass filters with tunable transfer function characteristics”, IEEE Transactions on Circuits and Systems I: Regular Papers, vol. 70, no. 12, pp. 5194-5207, Dec. 2023.
- Z. Zhang and D. Psychogiou, “Multifunctional bandpass filter with codesigned tunable attenuator and reflectionless phase shifter functionalities”, IEEE Microwave and Wireless Technology Letters.
- Z. Zhang and D. Psychogiou, “Multi-functional single/multi-band bandpass filters with co-integrated RF isolator, variable phase shifter or variable attenuator functionalities”, IEEE Transactions on Circuits and Systems I: Regular Papers, vol. 71, no. 9, pp. 4032-4045, Sept. 2024.
- Z. Zhang and D. Psychogiou, “Unilateral single-pole multi-throw filtering switch using spatiotemporally modulated resonator arrays”, in 2025 IEEE/MTT-S International Microwave Symposium (IMS), San Francisco, CA, USA, 2025.

- Z. Zhang and D. Psychogiou, “Multi-port multi-functional RF signal routing networks using spatiotemporally modulated resonator arrays”, IEEE Transactions on Microwave Theory and Technology.

Other Publications:

- A. Kumar, Z. Zhang, D. Sarkar, S. Nikolaou, P. Vryonides, and D. Psychogiou, “FDTD modeling of time-modulated resonators-based bandpass filters using modified telegrapher’s equations,” IEEE MTT-S Int. Microw. Symp. Dig., pp. 612–615, 2024.

5.3 Future Work

This thesis has laid the foundations for new classes of multi-configurable, multi-functional and multi-port RF components with multiple levels of adaptivity and advanced RF functionality using as a basis tunable STM-based resonator arrays. Several challenges remain unaddressed creating opportunities for future research:

5.3.1 Automated measurement and characterization platforms

One important research direction lies in the development of automated testing platforms for STM-based devices. At present, the characterization of STM-based RF circuits requires extensive manual tuning of bias voltages and modulation signals, which is time-consuming and may not always reach the optimal operating point predicted in EM simulations. Since the simulation results are typically obtained under carefully optimized trade-offs among IL, isolation, and transfer-function characteristics, reproducing these conditions in practice demands very precise parameter adjustment. As demonstrated in [134] and [135], software-aided automated setups can greatly enhance this process by efficiently exploring the tuning parameters and achieving the same optimal conditions as in the EM simulation environment. Figure 5.1 illustrates a proposed conceptual setup where the STM device under test (DUT) is simultaneously connected to an arbitrary waveform generator (AWG) and a vector network analyzer (VNA). The AWG provides the required modulation signals, while the VNA records the S-parameters of the DUT. These measurements are fed into a computer that performs optimization control, adjusting the DC biasing voltages and modulation parameters such as modulation frequency, modulation amplitude

phase progression. Specifically, the proposed optimization control module in Figure 5.1 can extract the parameters of the network model of the DUT from the measured S-parameter data and decide the parameters need to be tuned for achieving the target performances. Various algorithm can be applied such as gradient-based method in [134], or machine learning techniques such as reinforcement learning in [135], or artificial neural network as in [136]. Such an automated platform can significantly accelerate device characterization, improve repeatability, and automatically identify the optimal responses in terms of IL and directivity.

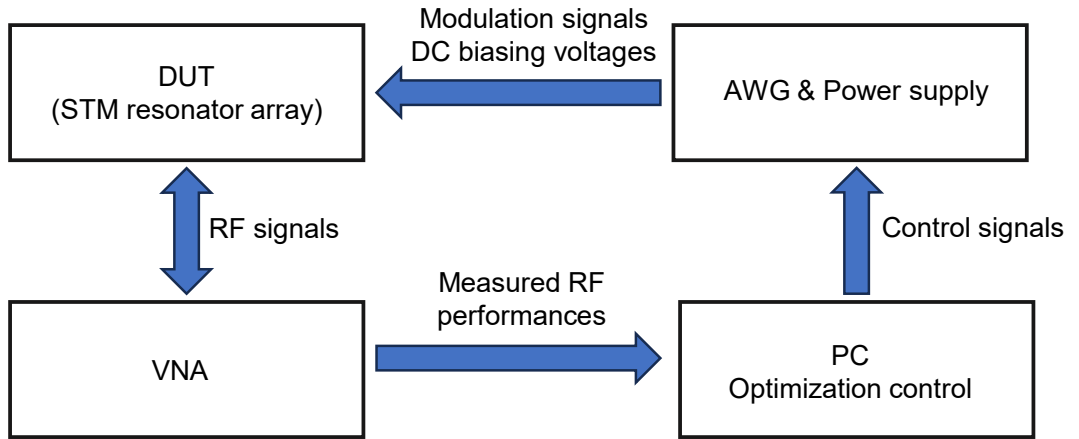


Figure 5.1. Conceptual setup of the automatic RF measurement and optimization system.

5.3.2 Alternative modulation elements

The majority of STM-based devices demonstrated to date rely on varactor diodes as the modulation and tuning elements. However, varactors impose significant limitations on the overall performance and scalability of STM architectures. Their moderate quality factor restricts achievable IL, while their breakdown voltage and power handling capability limit device operation at higher power levels. In addition, parasitic effects make it challenging to extend varactor-based implementations to higher frequencies such as millimeter-wave ranges. Future research should therefore focus on alternative modulation technologies that can overcome these limitations. One possible approach is to replace varactors with high-speed transistor-based switching of capacitors as in [48], [137] to enable operation at higher frequencies while also to achieve higher power handling performances. Another promising way is using RF

micro-electro-mechanical system (MEMS) and mechanically modulated resonators as indicated in [138]. It proposed a MEMS-only STM circulator and achieved an IL of 1.8 dB at 2.68 GHz by eliminating discrete varactors and using the resonator's own mechanical modulation. This mechanical approach improved the linearity and power handling.

5.3.3 Miniaturization and IC integration

Another important future direction is the miniaturization of STM resonator arrays through monolithic integration. Currently, the STM-based devices are built with distributed or lumped components on PCBs, where the modulation circuitry (bias circuits, AC sources) occupy far more volume than the core STM-resonator circuit. As shown in Figure 2.2, nearly 88% of the board's volume is taken up by SMA connectors and the biasing network for providing the two RF and six modulation signals to the STM resonators. Thus integrating the modulation waveform generation circuitry fully onto an IC platform has the potential to substantially reduce the size and enhance integration. Initial work in this direction has already shown in [62] where a complete magnetless circulator in which all of the resonators, switches and modulation signals are realized on-chip using CMOS technology. Future research should build on these implementations to develop IC-compatible unilateral filtering components. Figure 5.2 presents a conceptual block diagram of a chip-scale STM-based RF filtering device. In this architecture, the STM resonator array is directly driven by an on-chip voltage-controlled oscillator (VCO), which generates the modulation signals at the desirable frequency. A phase and amplitude control network distributes the appropriate modulation states to each resonator, while the DC biasing network provides the necessary DC biasing voltages. By embedding all these elements within the same IC platform, the bulky external circuits can be eliminated.

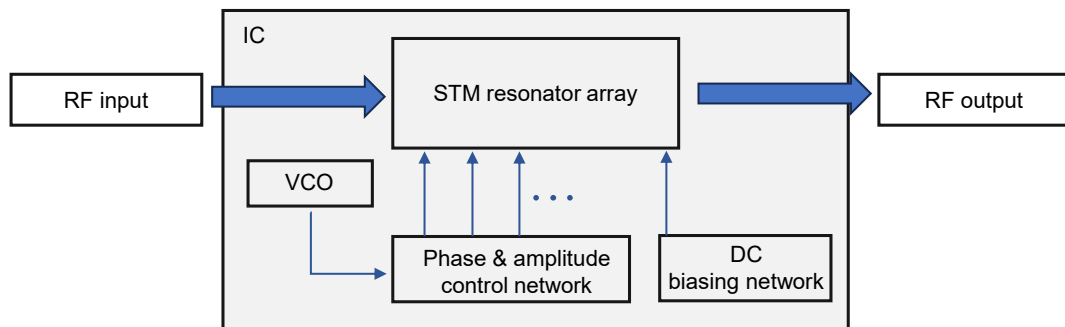


Figure 5.2. Conceptual block diagram of an IC-integrated STM-based RF filtering device.

5.3.4 Advanced theoretical and computational models

The theoretical analysis of STM filtering devices poses significant challenges. Existing numerical methods often involve substantial computational complexity. Developing a generalized theoretical framework that provides accurate approximate models and simplified analytical formulations is essential for gaining deeper insights into STM system. Current analysis methodologies primarily employ frequency-domain methods, considering only a limited number of IM products to maintain computational simplicity. Although they are effective at low modulation ratios, these methods lose accuracy and encounter convergence issues under large modulation conditions, necessitating consideration of more IM products. Recent efforts based on time-domain analyses such as the finite-difference time-domain (FDTD) method can account for all frequency components without truncation [139], [140].

As an initial step, [139] introduced a FDTD modelling technique based on modified Telegrapher's equations for STM-based BPFs. This preliminary work demonstrated that time-domain solvers can directly capture intermodulation products and non-reciprocal behaviour without harmonic truncation, offering a powerful alternative to frequency-domain methods. However, [139] was limited to inline BPF topologies and did not account for general multi-port interconnections or transmission line (TL) junctions. As a potential extension of this work, FDTD-based framework can be developed for the analysis and design of complex multi-connected topologies, such as transversal cell based BPFs, as well as multi-port networks including non-reciprocal power dividers, circulators and reconfigurable couplers. This generalized time-domain solver can not only supports arbitrary interconnections of STM resonators but also allows direct extraction of both frequency-domain (S-parameters) and time-domain metrics (waveforms, spectra, and constellation diagrams) within the same simulation environment. Future work should also leverage these capabilities to evaluate STM circuits under realistic communication signals, such as quadrature phase-shift keying (QPSK), where constellation distortion, error vector magnitude (EVM), and bit error rate (BER) can be directly evaluated.

5.3.5 System-level validation and applications

System-level validation of STM-based unilateral filtering components in various application scenarios remains an important step. One target application is FD wireless radios, where the transmitter and the receiver operate simultaneously using the same frequency. FD has the potential to double the spectral efficiency but requires extremely high isolation between the transmitter and receiver to cancel self-interference. Notably, conductivity-modulation-based STM approaches have already achieved system-level demonstrations as shown in [55], [141], [142], highlighting both the feasibility and the benefits of STM-enabled FD transceivers. System-level demonstrations of such concepts will be critical to quantifying their real-world impact. As an example, an STM-based multi-functional module such as the star-type block in Chapter 4 can be utilized for connecting the shared antenna, receiver chain and the transmitter chain as indicated in Figure 5.3. FD and half duplex applications can be realized using the same architecture by configuring the working states of the STM-based multi-functional module. Specifically, when it functions as a circulator, it allows the received signal to transmit to the LNA and transmit the signal from the PA simultaneously. Alternatively, time division duplex (TDD) wherein the signals are received and transmitted using different time slots, can be achieved when the module functions as a single-pole multi-throw switch. Metrics such as gain, noise figure, linearity, self-interference suppression, and power consumption should be evaluated in practical environments.

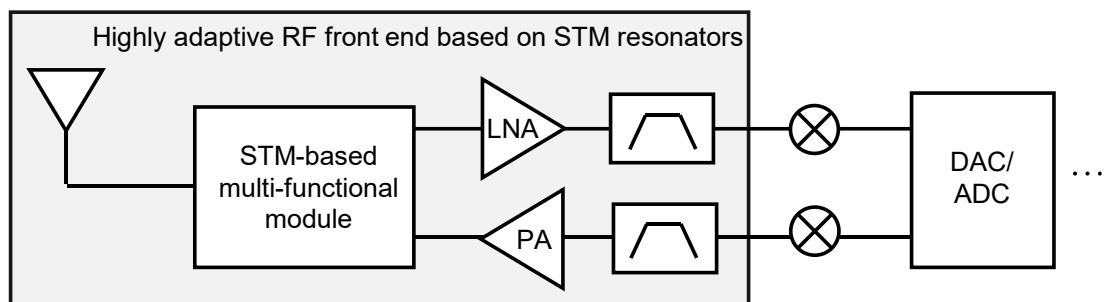


Figure 5.3. Conceptual architecture of STM-based highly adaptive RF transceivers.

Beyond FD, MIMO and beamforming systems would also benefit from the multi-functional RF components based on STM resonator array. Large antenna arrays currently require complex switching networks to steer beams or reconfigure

connectivity. Reconfigurable STM-based networks could simplify the structure by combining switching, filtering, and isolation in one device. Future work can involve deploying STM-based multi-port routing networks (such as the star- and box-type blocks of this thesis) in a prototype phased array or MIMO system, to facilitate functions like dynamic beam selection or steering. As shown in Figure 5.4, the multi-port STM networks can be connected to antenna arrays to facilitate various beam configurations. For example, the two arrays in Figure 5.4 can be independently selected and controlled by changing the dynamic beam forming network based on STM resonator arrays. Moreover, both arrays can facilitate non-reciprocal transmission or reception of RF signals as indicated by the red and blue beams. The two arrays can also be configured as a whole enabling constructive interference of the transmit/ receive signal as indicated by the grey beam.

Moreover, integrating non-reciprocal filtering functionality directly into antenna systems is a promising path to further reduce system size and cost. STM-based non-reciprocal filtering antennas have been demonstrated in [143] and [144], which simultaneously provide directionality, spectral control, and radiation. Future research should therefore explore the co-design methodologies for integrating antenna within the multi-port STM resonator array. This could enable flexible beam-steering antennas whose transmit and receive patterns differ, providing adaptive interference suppression in MIMO systems. These system-level integrations with FD and MIMO systems highlights the potential of STM-enabled multi-functional architectures in next-generation wireless networks.

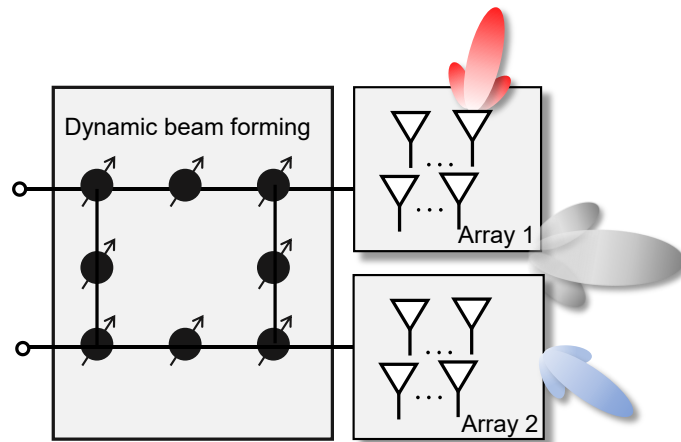


Figure 5.4. Conceptual diagram of dynamic beam forming network based on STM multi-port network.

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