

Spatiotemporal modulated multi-functional Non-reciprocal filtering phase shifter with continuously tunable frequency and isolator functionalities

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ABSTRACT

This paper presents RF co-design approach for a multi-function filtering phase shifter that integrate functionalities of a tunable bandpass filter, continuously tunable phase shifter, and isolator into a single circuit. To support detailed design, the analytical spectral S-parameters of the proposed multi-function non-reciprocal filtering phase shifter have been derived. The isolator functionality (*i.e.* $|S_{21}| \neq |S_{12}|$) can be achieved using time-varying capacitors modulated by progressive phase shift sinusoidal signal. The frequency and transmission phase can be simultaneously tuned by varying resonant frequency of resonators. For experimental validation, a microstrip line multi-function non-reciprocal filtering phase shifter is designed, fabricated and measured. The measurement demonstrates a phase shift range of 200° with forward insertion loss less than 4.9 dB and reverse isolation higher than 30 dB across all phase shift and frequency tuning states. The passband frequency is tuned from 1.42 GHz to 1.75 GHz, providing frequency tunability range of 330 MHz (*i.e.* tuning ratio: 1:1.2324), while maintaining phase shift range of 200° at each frequency tuning state. Furthermore, the measured input and output return losses are higher than 12 dB for each phase shift and passband frequency tuning states.

1. Introduction

Multi-functional and versatile microwave/mm wave circuits are highly demanded for next generation RF front-ends (RFFE) that can support multiple standards and operation modes while maintaining a smaller circuit footprint. As bandpass filters (BPFs) occupy significant portion of the circuit area in RFFEs, there is a growing interest in developing multi-functional circuits that integrate functionalities of BPF, variable phase shift, and isolator into single circuit. To integrate BPFs and tunable phase shifters into single circuit, various concepts and circuit topologies have been explored in [1–4]. Although numerous configurations of tunable phase shifter (TPS) [5–12] and bandpass filtering phase shifters [13–15] have been reported in literatures, integrated bandpass filter and tunable phase shifter (BPF-TPS) designs often struggle to preserve filtering characteristics while achieving large phase shift range (PSR). For instance, the work [16] integrates BPF-TPS by using passive vector summation method. While this configuration achieves bandpass filtering response along with 320° continuous phase tuning, it fails to preserve filtering characteristic due to the asymmetry of the signal paths. Similarly, integrated BPF-TPS is demonstrated in [17] by controlling external coupling network, however, tunable PSR in

this design is limited to 90° .

In recent years, BPFs with isolator functionality, also known as non-reciprocal bandpass filters (NR-BPF), have been investigated to facilitate the unidirectional signal transmission [18]. Various configurations of non-reciprocal circuits have been demonstrated using different approaches including magnetically-biased ferrite components [19,20] and exploiting the inherent non-reciprocity characteristics of transistors [21–23]. To eliminate the need of ferrite components and transistors in achieving non-reciprocity, the spatiotemporal modulation (STM) technique has been explored for designing magnetless circulators [24–26], and NR-BPFs [27,28] where non-reciprocity is achieved by modulating resonators with progressive phase shifted sinusoidal signals. Different topologies and configurations of lumped element and transmission line NR-BPFs as well as non-reciprocal filtering power dividers have been demonstrated in [29–36]. While these previously reported NR-BPFs and filtering non-reciprocal power dividers integrate functionalities of BPF and isolator, however, their transmission phase cannot be tuned. In subsequent studies [37,38], integrated non-reciprocal bandpass filter and tunable phase shifter/attenuator (NR-BPF-TPS/AT) functionalities have been demonstrated using power divider/combiner, filtering rat-race coupler, phase shifter, and time-modulated resonators. While

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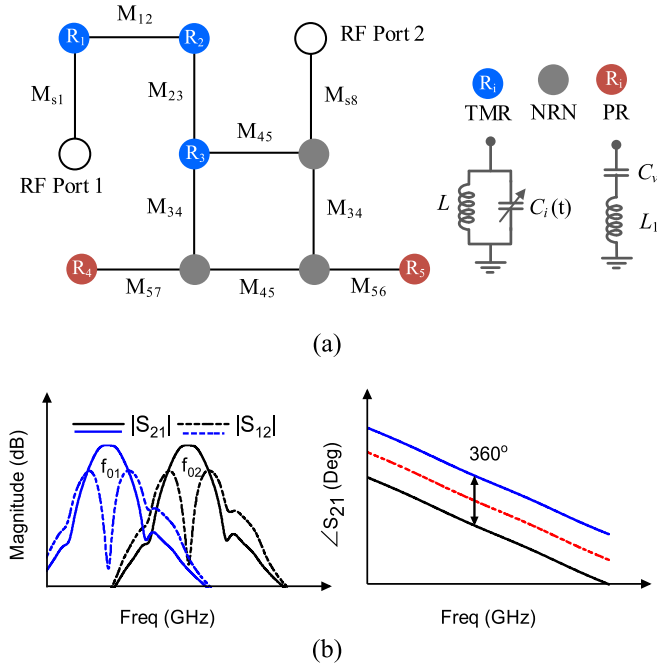


Fig. 1. (a) Coupling diagram and (b) conceptual frequency response of proposed multi-function non-reciprocal filtering phase shifter using two phase shift resonators.

these designs successfully integrate functionality of isolator, tunable phase shifter, and attenuator, these designs lack frequency tunability. Additionally, these approaches require a large number of tuning elements and time-varying capacitors for implementing tunable rat-race coupler and time-modulated resonators. As the number of time-modulated resonators increases, the required number of signal generators also increases, which significantly complicates the design of modulating AC signals circuitry.

Considering limitations of state-of-art, this work presents a detailed design methodology for a multi-functional phase shifter that simultaneously incorporates the functionalities of tunable BPF, continuously tunable phase shifter, and isolator. These capabilities help to reduce the footprint of next generation RFFE by integrating multiple functionalities into a single circuit and eliminating the need for separate modules typically required by RFFE. The proposed multi-functional circuit can be used in phased array antennas and MIMO antennas which demand beam-steering capabilities and high isolation between transmit and receive modules of RF transceiver.

2. Design theory

2.1. Type 1: Spectral S-parameters of non-reciprocal filtering phase shifter

Fig. 1 shows coupling diagram and conceptual frequency responses of the proposed multi-functional filtering phase shifter, which consists of input and output ports (RF port 1 and port 2), three time-modulated resonators (TMRs), two phase shifting resonators (PRs), and three non-resonating nodes (NRNs). It should be noted that the characteristics of both PR (i.e., value of C_v and L_1) are same. The TMR denoted with R_1 , R_2 , and R_3 consists of time-varying capacitor ($C_i(t)$) in parallel with inductor (L) whereas PRs (denoted by R_4 and R_5) consist of series inductor (L_i) and capacitor (C_v). To achieve isolator functionality, time-varying capacitors of TMR are modulated with progressive phase shift sinusoidal signal [27] as defined as (1).

$$C_i(t) = C_0[1 + m\cos\{2\pi f_m t + (i-1)\Delta\phi\}] \quad (1)$$

where f_m is modulation frequency, m is modulation index, $\Delta\phi$ is

progressive phase shift, and C_0 is nominal capacitance at static state (without modulation). When time-varying capacitor is modulated with progressive phase shift sinusoidal modulation signal, the intermodulation (IM) products [27,28] are generated. Considering number of harmonics (i.e. $k = -2, -1, 0, +1, +2$), the spectral admittance matrix of TMR can be written as (2).

$$\lambda_i = \begin{bmatrix} A^{-2} & B^{-2} & 0 & 0 & 0 \\ D^{-1} & A^{+1} & B^{-1} & 0 & 0 \\ 0 & D_0 & A^0 & B^0 & 0 \\ 0 & 0 & D^{+1} & A^{+1} & B^{+1} \\ 0 & 0 & 0 & D^{+2} & A^{+2} \end{bmatrix} \quad (2)$$

where

$$A^k = 2\pi(f + kf_m)C_0 - \frac{1}{2\pi(f + kf_m)L} \quad (3a)$$

$$B^k = 2\pi(f + kf_m)\frac{mC_0}{2}e^{-j(i-1)\Delta\phi} \quad (3b)$$

$$D^k = 2\pi(f + kf_m)\frac{mC_0}{2}e^{j(i-1)\Delta\phi} \quad (3c)$$

$$C_0 = \frac{1}{2\pi \times BW}L = \frac{BW}{2\pi f_0^2} \quad (3d)$$

$$k = -2, -1, 0, +1, +2, \quad i = 1, 2, 3 \quad (3e)$$

and f_0 and BW are RF center frequency and bandwidth of filtering phase shifter at static state (without modulation), respectively. Using coupling diagram as shown in Fig. 1(a), the spectral S-parameters of the proposed circuit can be derived as (4).

$$S_{11} = \frac{\mathbf{M}_{12}^2 - \mathbf{M}_{s1}^2 \mathbf{M}_{23}^2 - \lambda_1 \lambda_2 + \mathbf{M}_{s1}^2 \lambda_2 \lambda_3 + j\mathbf{E}_1}{\mathbf{M}_{12}^2 + \mathbf{M}_{s1}^2 \mathbf{M}_{23}^2 - \lambda_1 \lambda_2 - \mathbf{M}_{s1}^2 \lambda_2 \lambda_3 + j\mathbf{E}_2} \quad (4a)$$

$$S_{22} = \frac{-\mathbf{M}_{12}^2 + \mathbf{M}_{s1}^2 \mathbf{M}_{23}^2 + \lambda_2 \lambda_1 - \mathbf{M}_{s1}^2 \lambda_3 \lambda_2 + j\mathbf{E}_3}{\mathbf{M}_{12}^2 + \mathbf{M}_{s1}^2 \mathbf{M}_{23}^2 - \lambda_2 \lambda_1 - \mathbf{M}_{s1}^2 \lambda_3 \lambda_2 + j\mathbf{E}_4} \quad (4b)$$

$$S_{21} = j \frac{2\mathbf{M}_{s1} \mathbf{M}_{12} \mathbf{M}_{23} \mathbf{E}_5}{\mathbf{M}_{12}^2 + \mathbf{M}_{s1}^2 \mathbf{M}_{23}^2 - \lambda_1 \lambda_2 - \mathbf{M}_{s1}^2 \lambda_2 \lambda_3 + j\mathbf{E}_2} \quad (4c)$$

$$S_{12} = j \frac{2\mathbf{M}_{s1} \mathbf{M}_{12} \mathbf{M}_{23} \mathbf{E}_5}{\mathbf{M}_{12}^2 + \mathbf{M}_{s1}^2 \mathbf{M}_{23}^2 - \lambda_2 \lambda_1 - \mathbf{M}_{s1}^2 \lambda_3 \lambda_2 + j\mathbf{E}_4}, \quad (4d)$$

where

$$\mathbf{E}_1 = \mathbf{M}_{23}^2 \lambda_1 - \mathbf{M}_{s1}^2 \lambda_2 + \mathbf{M}_{12}^2 \lambda_3 - \lambda_1 \lambda_2 \lambda_3 \quad (5a)$$

$$\mathbf{E}_2 = \mathbf{M}_{23}^2 \lambda_1 + \mathbf{M}_{s1}^2 \lambda_2 + \mathbf{M}_{12}^2 \lambda_3 - \lambda_1 \lambda_2 \lambda_3 \quad (5b)$$

$$\mathbf{E}_3 = \mathbf{M}_{23}^2 \lambda_1 - \mathbf{M}_{s1}^2 \lambda_2 + \mathbf{M}_{12}^2 \lambda_3 - \lambda_3 \lambda_2 \lambda_1 \quad (5c)$$

$$\mathbf{E}_4 = \mathbf{M}_{23}^2 \lambda_1 + \mathbf{M}_{s1}^2 \lambda_2 + \mathbf{M}_{12}^2 \lambda_3 - \lambda_3 \lambda_2 \lambda_1 \quad (5d)$$

$$\mathbf{E}_5 = \frac{4\pi^2 f^2 C_{v1} L_1 - 1 + j2\pi f C_{v1} Z_0}{4\pi^2 f^2 C_{v1} L_1 - 1 - j2\pi f C_{v1} Z_0} \quad (5e)$$

$$\mathbf{M}_{34} = \mathbf{M}_{56} = \mathbf{M}_{67} = \mathbf{M}_{88} = \mathbf{U}, \quad \mathbf{M}_{45} = \sqrt{2}\mathbf{U} \quad (5f)$$

and Z_0 is port impedance of 50 Ω and $\mathbf{U}$ is identity matrix. In (4) and (5), the frequency-independent couplings written in bold represent the spectral scattering at onset of progressive phase shift sinusoidal modulation i.e. $\mathbf{M}_{a,b} = M_{a,b} \mathbf{U}$.

Using (4a) and (4b), the relation among coupling matrix values $\mathbf{M}_{s1}$, $\mathbf{M}_{12}$, and $\mathbf{M}_{23}$ can be found as (6) by setting $S_{11} = S_{22} = 0$.

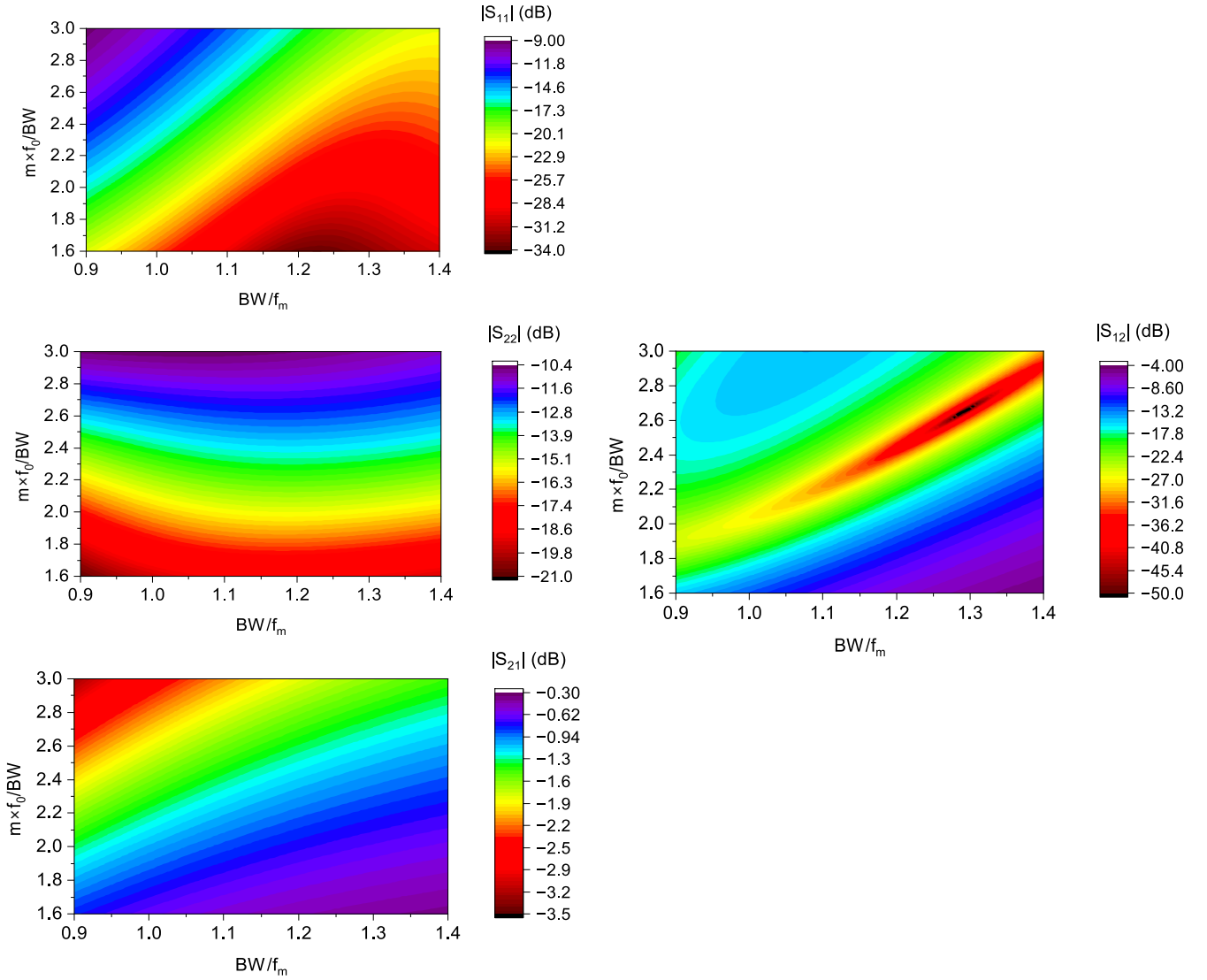


Fig. 2. Calculated S-parameters of proposed filtering phase shifter at center frequency according to modulation frequency (f_m) modulation index (m) and progressive phase shift $\Delta\phi = 60^\circ$. Coupling matrix values: $M_{12} = 1.70$, $M_{23} = 0.95$, $M_{s1} = M_{12}/M_{23}$, $M_{45} = M_{57} = M_{s8} = 1$, and $M_{34} = 1.4142$. $L_1 = 2$ nH, and C_v is varied from 0.4 pF to 10 pF. Phase shift range: $> 210^\circ$.

$$\mathbf{M}_{s1} = \frac{\mathbf{M}_{12}}{\mathbf{M}_{23}} \quad (6)$$

As observed from (6), the value of $\mathbf{M}_{s1}$ can be easily determined if $\mathbf{M}_{12}$ and $\mathbf{M}_{23}$ are known. Furthermore, the transmission phase of the proposed circuit can be written as (7) using (4c).

$$\angle S_{21} = \tan^{-1}\left(\frac{\mathbf{E}_2}{\mathbf{E}_6}\right) + 2\tan^{-1}\left(\frac{2\pi f C_{v1} Z_0 \mathbf{U}}{4\pi^2 f^2 C_{v1} L_1 - 1}\right) \quad (7)$$

where

$$\mathbf{E}_6 = \mathbf{M}_{12}^2 + \mathbf{M}_{s1}^2 \mathbf{M}_{23}^2 - \lambda_1 \lambda_2 - \mathbf{M}_{s1}^2 \lambda_2 \lambda_3 \quad (8)$$

From (7) and (8), the phase shift can be changed by varying value of C_v from $C_v^{\min}$ to $C_v^{\max}$. Therefore, the PSR of the proposed circuit can be written as (9).

$$PSR_{type-1} = \angle S_{21}|_{C_{v1}^{\max}} - \angle S_{21}|_{C_{v1}^{\min}} = \left\{ \begin{array}{l} 2\tan^{-1}\left(\frac{2\pi f C_{v1}^{\max} Z_0 \mathbf{U}}{4\pi^2 f^2 C_{v1}^{\max} L_1 - 1}\right) \\ -2\tan^{-1}\left(\frac{2\pi f C_{v1}^{\min} Z_0 \mathbf{U}}{4\pi^2 f^2 C_{v1}^{\min} L_1 - 1}\right) \end{array} \right\} \quad (9)$$

As can be seen from (9), transmission phase of non-reciprocal filtering phase shifter can be continuously tuned by varying capacitance from $C_{v1}^{\min}$ to $C_{v1}^{\max}$, while center frequency of circuit can be tuned by varying value of C_o of time-varying capacitor.

2.2. Modulation parameter selection for achieving non-reciprocal response

To achieve low insertion loss (IL) in forward direction and high isolation in backward direction while maintaining the larger PSR, the modulation parameters (i.e. f_m , m , and $\Delta\phi$) of TMR should be chosen appropriately. For this purpose, heatmaps data visualization results are shown in Fig. 2, where color bar represents S-parameter magnitudes (such as $|S_{11}|$, $|S_{21}|$, $|S_{12}|$ and $|S_{22}|$). In these figures, x-axis represents

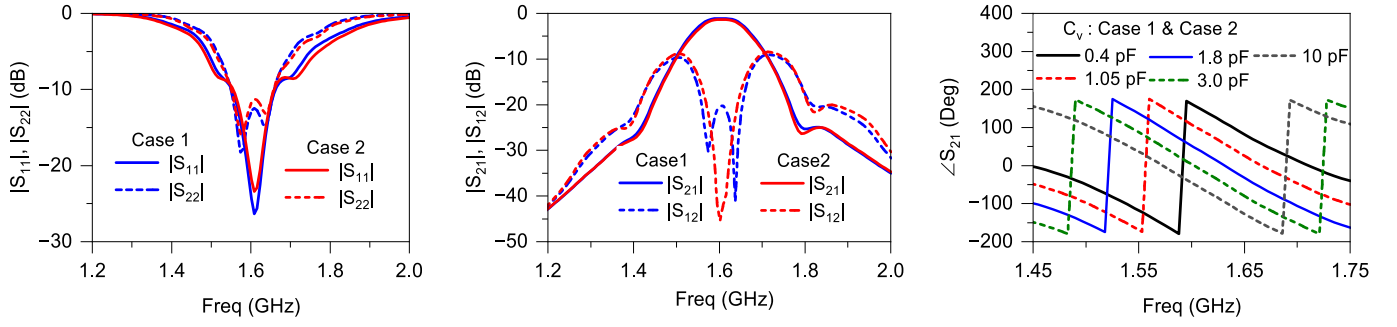


Fig. 3. Simulated S -parameters of proposed multi-function non-reciprocal filtering phase shifter according to different modulation parameters: Case 1: $f_m = 1.25 \times BW$, $m = 2.38 \times BW/f_0$, $\Delta\phi = 60^\circ$ and Case 2: $f_m = 1.25 \times BW$, $m = 2.62 \times BW/f_0$, $\Delta\phi = 60^\circ$. Coupling matrix values: $M_{12} = 1.70$, $M_{23} = 0.95$, $M_{s1} = M_{12}/M_{23}$, $M_{45} = M_{57} = M_{s8} = 1$, $M_{34} = 1.4142$, $BW = 100$ MHz, $f_0 = 1.60$ GHz. $L_1 = 3.5$ nH, and C_v is varied from 0.4 pF to 10 pF. Phase shift range: $0^\circ \sim 212^\circ$.

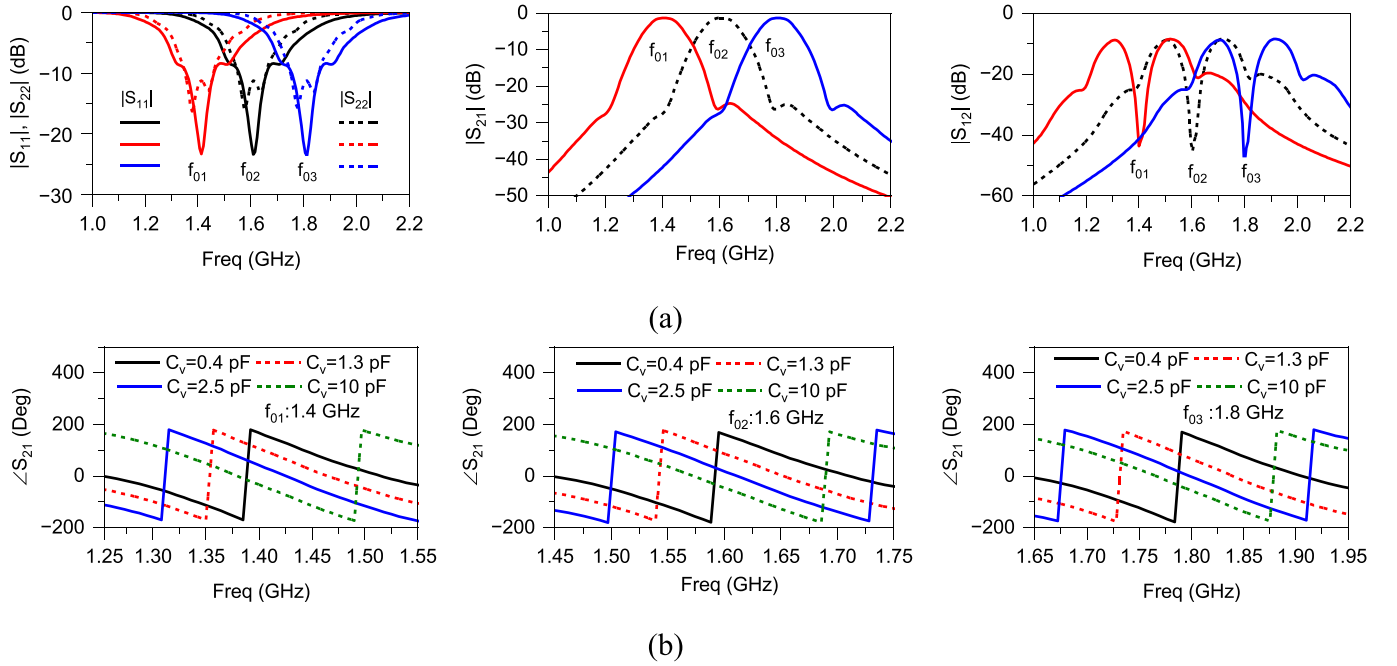


Fig. 4. Calculated tunable frequency response of proposed filtering phase shifter with isolator functionality: (a) S -parameters and (b) transmission phase characteristics ($\angle S_{21}$). Coupling matrix values: $M_{12} = 1.70$, $M_{23} = 0.95$, $M_{s1} = M_{12}/M_{23}$, $M_{45} = M_{57} = M_{s8} = 1$, $M_{34} = 1.4142$, $BW = 100$ MHz, $f_0 = 1.60$ GHz. $L_1 = 3.5$ nH and C_v is varied from 0.4 pF to 10 pF. Phase shift range (PSR): $0 \sim 212^\circ$.

ratio of BW and f_m whereas y-axis represents ratio of product of $m \times f_m$ and BW . For these results, the value of progressive phase shift (*i.e.* $\Delta\phi$) is assumed to be 65° . As seen from these results, when BW/f_m ratio increases, the ratio of $m \times f_m/BW$ should be also increased to get higher reverse isolation (IX). Larger BW/f_m ratio exhibits higher IX, however forward IL and output return loss ($|S_{22}|$) are degraded. Therefore, trade-off performances of the proposed phase shifter should be considered. It should be noted that PSR is maintained around 210° while sweeping modulation parameters for data visualization approaches.

Based on observed trends from parametric studies shown in Fig. 2, empirical relationship between desired design specifications of the proposed non-reciprocal phase shifter (such as center frequency, bandwidth, and reverse isolation (IX)) and modulation parameters can be established as follows:

$$f_m = 1.25 \times BW \quad (10a)$$

$$m = \begin{cases} \frac{2.38 \times BW}{f_0} & \text{for IX} > 20 \text{ dB \& two null} \\ \frac{2.62 \times BW}{f_0} & \text{for IX} > 40 \text{ dB \& single null} \end{cases} \quad (10b)$$

$$\Delta\phi = 60^\circ \sim 65^\circ, \quad (10c)$$

where f_0 and BW are RF center frequency and bandwidth of filtering phase shifter at static state (without modulation), respectively. These empirical relationships will provide design guidelines for the proposed non-reciprocal phase shifter.

To illustrate the trade-off performances, Fig. 3 shows the calculated S -parameters and forward transmission phase responses of the proposed filtering shifter using (2)–(8). Based on previously described data visualization approaches, the modulation parameters are chosen as $f_m = 1.25 \times BW$, $m = 2.38 \times BW/f_0$, $\Delta\phi = 60^\circ$ for case 1 and $f_m = 1.25 \times BW$, $m = 2.62 \times BW/f_0$, $\Delta\phi = 60^\circ$ for case 2. Assuming $f_0 = 1.60$ GHz and $BW = 100$ MHz, calculated S -parameters and forward transmission phase response are shown in Fig. 3. As seen from these results, larger m exhibits

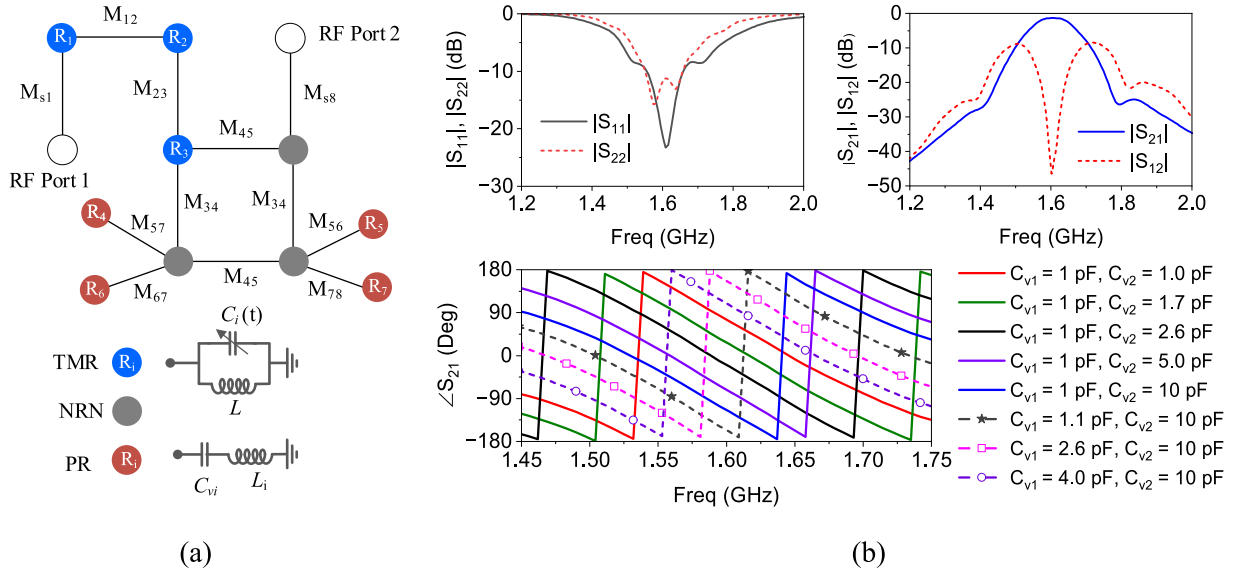


Fig. 5. (a) Coupling diagram of the proposed multi-functional non-reciprocal filtering phase shifter type 2 using four phase shift resonators and (b) simulated results of multi-function non-reciprocal filtering power divider at $f_0 = 1.60$ GHz with $BW = 100$ MHz using four phase shift resonators. Coupling matrix values: $M_{12} = 1.70$, $M_{23} = 0.95$, $M_{s1} = M_{12}/M_{23}$, $M_{45} = M_{56} = M_{67} = M_{57} = M_{78} = M_{s8} = 1$, and $M_{34} = 1.4142$, $L_1 = 0.1$ nH, $L_2 = 4.90$ nH, C_{v1} : 1 to 4 pF and C_{v2} : 1 to 10 pF. Modulation parameters: $f_m = 1.25 \times BW$, $m = 2.62 \times BW/f_0$, $\Delta\phi = 60^\circ$. Phase shift range: $> 360^\circ$.

high IX (> 45 dB @ f_0) with single null, while smaller m exhibits slightly smaller IX (> 20 dB @ f_0) and wider isolation bandwidth with two nulls. At f_0 , the forward transmission IL is less than 0.94 dB while input and output return losses are higher than 24 dB and 11 dB, respectively. Although the proposed circuit is analyzed without lossy circuits, power conversion into IM products contributes small forward transmission IL because these powers are not fully recovered back to fundamental RF frequency signal [27,28]. The forward transmission PSR of 212° is achieved by varying C_v from 0.4 pF to 10 pF for both cases. Furthermore, the magnitude response (*i.e.* $|S_{11}|$, $|S_{22}|$, $|S_{21}|$, and $|S_{12}|$) remained same while tuning transmission phase characteristics.

Fig. 4 illustrates the frequency tunable response of the proposed

filtering phase shifter. The frequency tunable response can be achieved by tuning the resonant frequency of the TMR. For these results, the modulation parameters of circuit are chosen as $f_m = 1.25 \times BW$, $m = 2.62 \times BW/f_0$, $\Delta\phi = 60^\circ$ to achieve forward transmission IL < 0.94 dB, IX > 45 dB, input return loss > 25 dB, and output return loss > 11 dB at f_0 . As observed from figure, the passband center frequency is tuned from 1.40 GHz to 1.80 GHz by adjusting resonant frequencies of TMR. The forward transmission phase is continuously tuned around 212° for each passband tuning state by varying C_v from 0.4 pF to 10 pF.

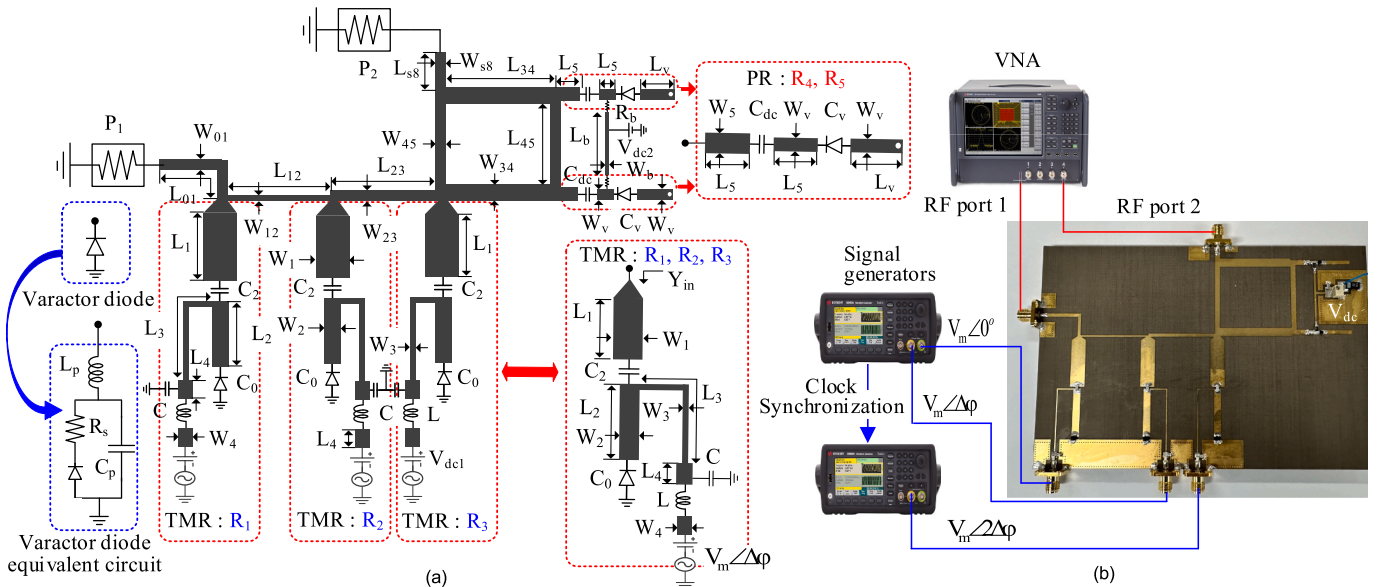


Fig. 6. (a) Microstrip line implementation of proposed non-reciprocal filtering phase shifter and (b) photograph of fabricated circuit with experimental setup. Physical dimensions: $W_{01} = 1.48$, $W_{12} = 0.68$, $W_{23} = 1.48$, $W_{45} = 2.38$, $W_{34} = 3.9$, $W_{s8} = 2.38$, $L_{01} = 26.5$, $L_{12} = 31.1$, $L_{23} = 30.7$, $L_{34} = 31.8$, $L_{45} = 28.6$, $L_{s8} = 6$, $L_4 = 5.5$, $W_1 = 6.6$, $W_2 = 3.3$, $W_3 = 0.5$, $W_4 = 2.38$, $W_v = 1.8$, $L_1 = 22$, $L_2 = 22$, $L_3 = 28$, $L_v = 14$, $C_2 = 3.9$ pF, $W_b = 0.4$, $L_b = 10$, $C_{dc} = 33$ pF, $L = 90$ nH, and $R_b = 10$ k Ω . C_0 : varactor SMV 1233-079LF, C_v : varactor SMV1231-011LF. Varactor diode parasitic values: $C_p = 0.5$ pF, $L_p = 1.9$ nH, and $R_s = 1.2$ Ω . Physical dimension unit: millimeter (mm).

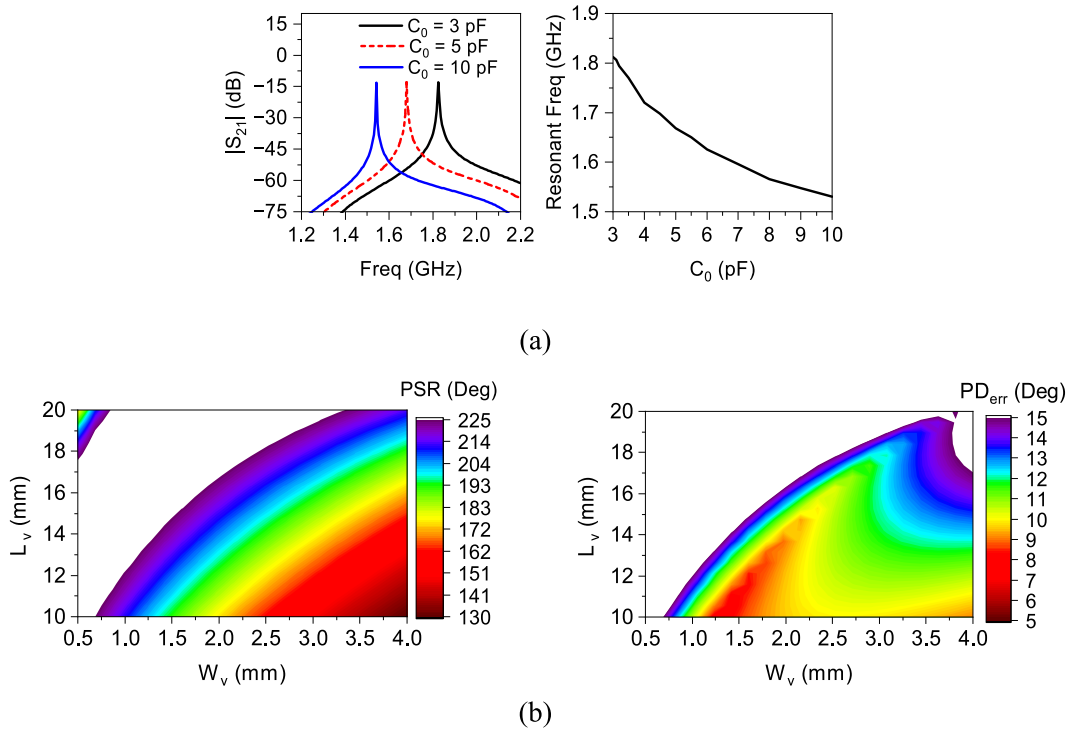


Fig. 7. (a) Resonant frequency characteristics of time modulated resonator for different value of C_0 and $W_1 = 6.6$, $W_2 = 3.3$, $W_3 = 0.5$, $W_4 = 2.38$, $L_1 = 22$, $L_2 = 22$, $L_3 = 28$, $L_4 = 5$, $C = 33$ pF, $L = 90$ nH and (b) forward transmission phase shift range (PSR) and phase deviation error (PD_{err}) within bandwidth of 300 MHz according to physical parameters of phase shift resonator. Taconic substrate with $\epsilon_r = 2.20$, $h = 0.787$ mm, and tangent loss = 0.0009 and C_v is varied from 0.4 pF to 10 pF.

2.3. Type 2: Spectral S-parameters of non-reciprocal filtering phase shifter

To further enhance PSR, Fig. 5(a) depicts coupling diagram of type-2 multi-function non-reciprocal phase shifter. Whereas the number of TMR resonators (i.e. R_1 , R_2 , R_3) remained same as type-1 structure (refer to Fig. 1(a)), the number of phase shift resonators has been increased from 2 to 4. The increase in phase shift resonators provides additional flexibility to enhance PSR of forward transmission. In this structure, value of PS resonator 4 and 5 (i.e. R_4 and R_5) are same as PS resonator 6 and 7 (i.e. R_6 and R_7). The expression of analytical S-parameters remains same as (4) and (5) except value of (5e), which is given as (10).

$$\mathbf{E}_5 = \frac{E_7 + jE_8}{E_7 - jE_8}, \quad (11)$$

where

$$E_7 = \omega^4 C_{v1} C_{v2} L_1 L_2 - \omega^2 (C_{v1} L_1 + C_{v2} L_2) + 1 \quad (12a)$$

$$E_8 = Z_0 \{ \omega^3 C_{v1} C_{v2} (L_1 + L_2) - \omega (C_{v2} + C_{v1}) \} \quad (12b)$$

Therefore, forward transmission phase of type-2 structure can be given as (11).

$$\angle S_{21} = \tan^{-1} \left(\frac{E_2}{E_6} \right) + 2 \tan^{-1} \left(\frac{E_8 U}{E_7} \right) \quad (13)$$

As shown in (11)–(12), type-2 structure offers greater flexibility for enhancing PSR of the proposed multi-functional non-reciprocal phase shifter by allowing different values of C_{v1} , L_1 , C_{v2} and L_2 . Since TMRs are same as in the type-1 structure, the modulation parameters of type-2 design can be selected as discussed in previous section.

To demonstrate the frequency response of type-2 structure, Fig. 5(b) presents the simulated frequency response of multi-function non-reciprocal phase shifter at $f_0 = 1.60$ GHz. The circuit parameters and modulation parameters used in the simulation are also shown in Fig. 5(b). As seen from the figure, the proposed structure achieves the 360° PSR while

maintaining low transmission IL and high IX. It should be noted that PSR of this structure is enhanced from 212° to 360° because increasing number of phase shift resonators provide more degree of freedom to control transmission phase of multi-function non-reciprocal phase shifter. However, it can increase the circuit size and complexity of circuit.

2.4. Microstrip line implementation of non-reciprocal phase shifter

Fig. 6(a) shows the overall microstrip line implementation of non-reciprocal filtering phase shifter, which consists of three TMRs and two-phase shift resonators. As depicted in figure, the coupling M_{s1} between the first TMR (R_1) and port 1 is implemented using TL with a width and length of W_{01}/L_{01} . Similarly, the coupling M_{12} between first and second TMRs (i.e. R_1 and R_2) is implemented using TL with a width and length of W_{12}/L_{12} whereas the coupling M_{23} between second and third TMRs (i.e. R_2 and R_3) are implemented through TL of W_{23}/L_{23} . Likewise, the coupling M_{45} and M_{34} are implemented using TL with widths and lengths of W_{45}/L_{45} and W_{34}/L_{34} , respectively. Finally, the coupling M_{s8} is implemented with W_0/L_0 . The couplings that provide desired the filtering characteristic can be achieved by controlling these physical parameters. It should be noted that TL length for these coupling should be designed as quarter-wavelength ($\lambda_g/4$) at midpoint of the tunable center frequency. As depicted in figure, TMR consists of series of TLs with width and lengths of W_1/L_1 , W_2/L_2 , W_3/L_3 , dc block capacitor (C_2), varactor diode with capacitance of C_0 . The progressive phase shift modulation signal is supplied through TL (W_3/L_3), series inductor L , and shunt capacitor C . The resonant frequency of time-modulated resonator can be found by equating $\text{imag}(Y_{in}) = 0$ as shown in figure.

Fig. 7(a) illustrates the resonant frequency characteristics of time-modulated resonator for different value of C_0 . The other parameters of resonator are depicted in the figure. As observed from figure, the resonant frequencies can be tuned from 1.40 GHz to 1.82 GHz by varying value of C_0 from 3 pF to 10 pF. Similarly, phase shift resonators consist of TLs with widths and lengths of W_5/L_5 , W_v/L_5 , W_v/L_v , dc block capacitor

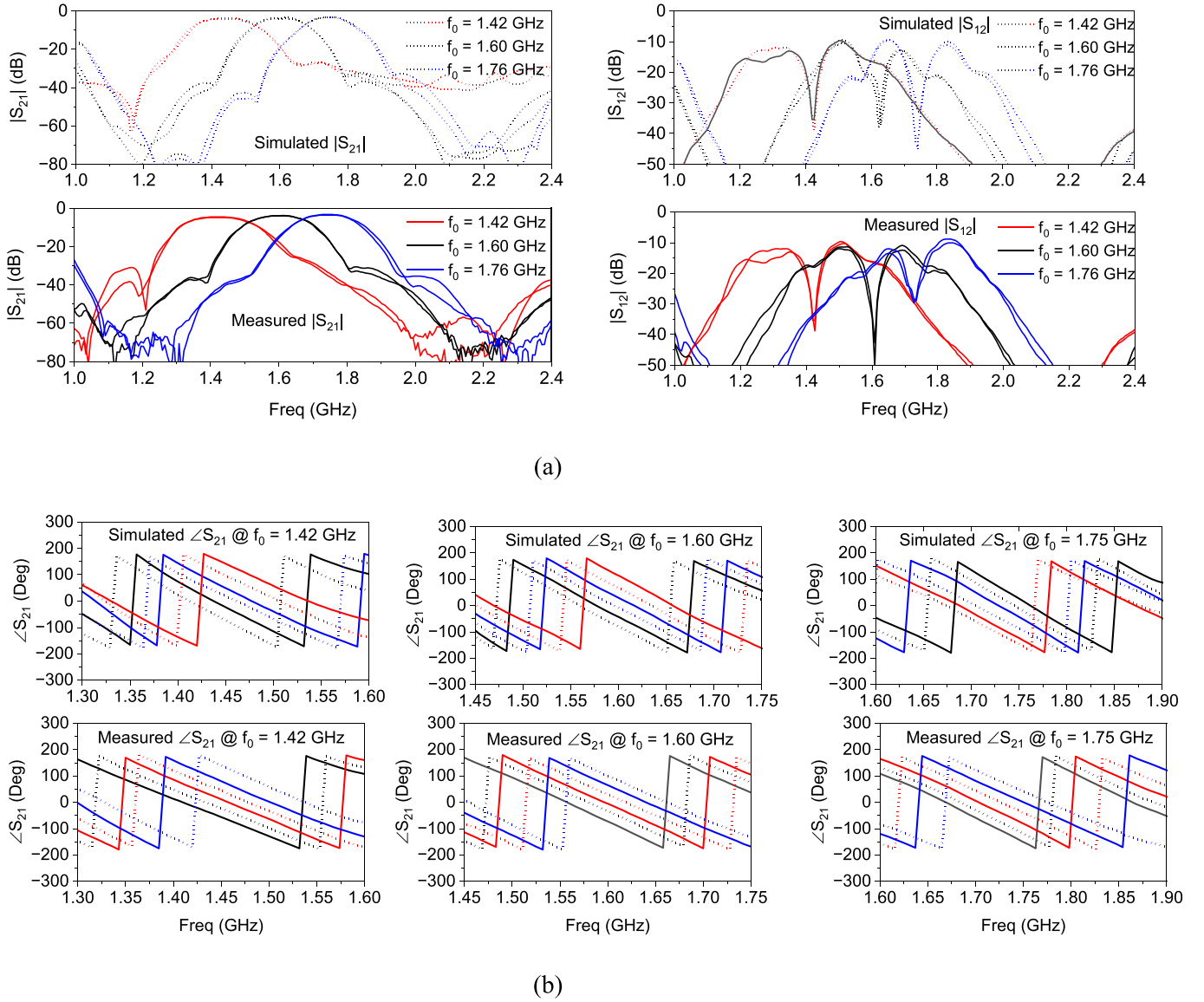


Fig. 8. Simulated and measured frequency response of proposed multi-function non-reciprocal filtering phase shifter: (a) forward and reverse transmission S-parameter magnitudes and (b) transmission phase response.

C_{dc} and varactor diode (C_v). By controlling W_v , L_v , and C_v , different PSR can be obtained as shown in Fig. 7(b).

To achieve flat phase shift characteristics within passband frequency,

Table 1

Summary of measurement results of proposed multi-function non-reciprocal filtering phase shifter.

		Passband center frequency		
		$f_0 = 1.42$ GHz	$f_0 = 1.60$ GHz	$f_0 = 1.75$ GHz
Modulation parameters	f_m (MHz)	120	120	120
	V_m (V)	3.20	3.30	3.40
	$\Delta\phi$ (Deg)	65	66	64
	V_{dc1} (V)	0	2.40	5.10
	V_{dc2} (V)	0–15	0–15	0–15
Measured results	RL_{min} (dB)	> 11.83	> 10.75	> 10.25
	IL (dB)	< 4.56	< 4.10	< 3.77
	IX (dB)	27.05–38.77	29.40–34.76	19.88–25.28
	PSR	204.18	194.21	193.31
	(Deg)			

the phase deviation error (PD_{err}) is limited to $\pm 15^\circ$ and results are shown in Fig. 7(b). As observed from figure, the proposed circuit can provide maximum PSR of 225° if PD_{err} is within $\pm 15^\circ$. It should be noted that the proposed circuit can provide PSR higher than 225° , but PD_{err} is worse ($> 40^\circ$). If small PD_{err} is desired, the PSR will be decreased. Therefore, tradeoff exists between PSR and PD_{err} .

3. Experimental results

For proof-of-concept, a microstrip line multi-function non-reciprocal filtering phase shifter is designed and manufactured using Taconic substrate TLY-5 substrate, which has dielectric constant of 2.2 and thickness of 0.787 mm. The goals of type 1 structure are to achieve PSR higher than 200° , forward IL less than 4 dB, IX exceeding higher than 30 dB, while tuning frequency from 1.40 GHz to 1.80 GHz. In this work, time-varying capacitors ($C_i(t)$) are implemented by modulating varactor diode SMV 1233-079LF from Skyworks Inc., while variable capacitors for PS resonators (*i.e.* R_4 and R_5) are implemented using the varactor diode SMV 1231-011LF from Skyworks Inc., which provides variable capacitance ranging from 0.45 pF to 4 pF. The layout and physical

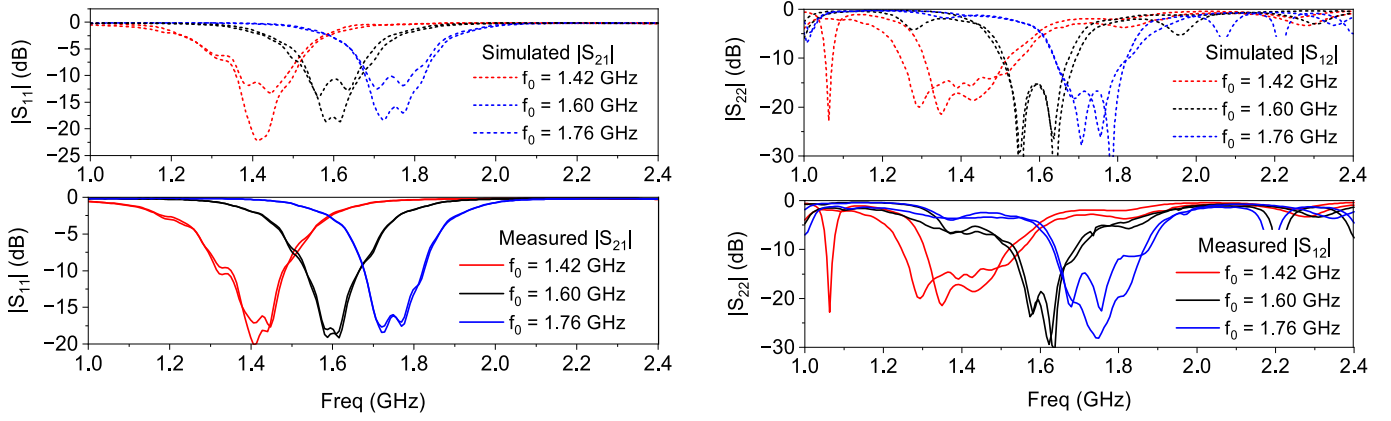


Fig. 9. Simulated and measured input and output return losses of proposed multi-functional non-reciprocal filtering phase shifter.

Table 2

Performance comparison of proposed non-reciprocal phase shifter with state-of-art multi-functional circuits.

Ref	Functionality	Frequency (GHz)	FTR (MHz)	PSR (Deg)	IL@ f_0 (dB)	IX @ f_0 (dB)	TVR	Circuit size (λ_g^2)	FT
[3]	BPF + TPS	0.6–0.71	110 (1:1.1833)	320	1.90–3	NA	NA	0.196×0.098	Y
[4]	BPF + TPS*	0.65–0.86	210 (1:1.323)	360	2.20–4.30	NA	NA	0.177×0.177	Y
[5]	TPS**	2.45	NA	360	3	NA	NA	NA	N
[6]	TPS	1.50	NA	280	1–2.6	NA	NA	0.26×0.37	N
[9]	TPS	2.50	NA	146	1.28	NA	NA	0.50×0.50	N
[10]	TPS	10	NA	190	< 2.40	NA	NA	NA	N
[16]	BPF + TPS	1.05–1.30	250 (1:1.2380)	320	3.2–4.10	NA	NA	0.9×1.10	Y
[17]	BPF + TPS	1.20	NA	90	<1.40	NA	NA	0.31×0.27	N
[27]	BPF + IS	0.2	NA	NA	<1.50	20	3	NA	N
[28]	BPF + IS	1.0	NA	NA	<4.40	18	3	0.167×0.51	N
[29]	BPF + IS	0.15	NA	NA	<1.50	18	3	NA	N
[31]	BPF + IS	0.57–0.724	154 (1:2701)	NA	4.8–6.8	21–35	4	0.104×0.095	Y
[32]	BPF + IS	1.64–1.97	330 (1:2021)	NA	3.94–4.92	20	3	0.36×0.36	Y
[33]	BPF + IS	1.62–1.94	320 (1:1975)	NA	4.30–5.40	> 20	4	0.44×0.40	Y
[37]	BPF + IS+TPS	0.69	NA	340	5.90–6.90	17	4	0.145×0.145	N
[38]	BPF + IS+TPS	0.70	NA	360	5.40–6.80	25.4–26.8	6	0.266×0.32	N
This work	BPF + IS +TPS + FT	1.42–1.75	330 (1:1.2324)	198–204	3.77–4.56	19.88–38.77	3	0.79×0.53	Y

BPF: bandpass filter, TPS: tunable phase shifter, IS: isolator functionality, PSR: phase shift range, IL: forward transmission insertion loss, IX: reverse isolation, TVR: number of time-varying resonators, FTR: frequency tunability range, *: Lumped element implementation, **: CMOS implementation, λ_g : guided wavelength at operating center frequency.

dimensions of fabricated multi-function non-reciprocal filtering phase shifter are shown in Fig. 6(a) and a photograph of fabricated circuit along with the measurement setup is shown in Fig. 6(b).

To obtain simulated results, a multi-port touchstone file is firstly generated by performing 3D electromagnetic (EM) simulation using ANSYS HFSS. This multi-port touchstone file is then exported to Keysight advanced design system (ADS) where SPICE models of varactor diodes [39] are combined with EM results. The equivalent SPICE model of varactor diodes is shown in Fig. 6(a). Finally, the simulated S-parameters are obtained through harmonic balance simulation in ADS.

Fig. 8 depicts the simulated and measured forward and reverse transmission S-parameter magnitudes ($|S_{21}|$ and $|S_{12}|$) and transmission phase ($\angle S_{21}$) results of type 1 multi-function non-reciprocal filtering phase shifter. A summary of the measured results is provided in Table 1. As seen from these results, the measured results are well agreed with simulated results. Although the frequency is continuously tuned, the figure highlights the results of three specific states of tunable frequency characteristic centered at 1.42 GHz, 1.60 GHz, and 1.76 GHz are shown in figure. The tunable frequency characteristics with non-reciprocal response is achieved by varying dc-bias voltage of varactor diode and applying progressive phase shift modulation signals. Based on the experimental results, center frequency is tuned from 1.42 GHz to 1.76 GHz achieving frequency tunable range (FTR) of 20.82% (or 1:1.2324). Throughout the measured tuning results, the forward IL remained

within 3.77 dB to 4.56 dB, while IX exceeds 38.77 dB. The PSR is higher than 194.4° for each center frequency. Similarly, input/output return losses are better than 10.25 dB for each frequency tuning state as shown in Fig. 9.

Table 2 shows the performances comparison between the proposed multi-function non-reciprocal filtering phase shifter and state-of-art co-designed circuits including BPF, TPS, and isolator. As shown in table, the works [3,4,13,16] achieve a continuously tunable phase shift along with BPF response. However, these designs lack isolator functionality. On the other hand, works [26–28,30–32] demonstrate a BPF response with isolator functionality, but they do not provide continuous tunable transmission phase characteristics. While works [36,37] offer tunable phase shifting and BPF responses combined with isolator functionality, these circuits do not support the frequency tunability. Additionally, these designs [36,37] require a larger number of time-modulated resonators, which increases complexity in modulation circuitry and number of required signal AC signal sources. In contrast, the proposed filtering phase shifter achieves superior performances by offering larger PSR, frequency tunability range, and well-preserved filtering characteristics, while using smaller number of time-modulated resonators (only three time-modulated resonators). Moreover, the proposed multi-functional filtering phase shifter integrates the isolator functionality with directivity ($|S_{21}|/|S_{12}|$) exceeding 20 dB in all tuning states, insuring higher isolation capability.

4. Conclusion

This work demonstrates a compressive design methodology and experimental validation of multi-function non-reciprocal filtering phase shifter that can provide not only tunable phase characteristics, but also high non-reciprocity in reverse direction. As result, the proposed circuit can integrate functionalities of phase shifter, tunable bandpass filter and isolator within same circuit. To insight non-reciprocity within tunable phase shifter, the analytical *S*-parameters have been derived and validated through simulation results. For proof-of-concept, a microstrip line multi-function non-reciprocal phase shifter is designed, manufactured, and characterized. The proposed circuit can incorporate with phased array antennas for simultaneous control of transmission and reception beamforming characteristics at same operating frequency which can open doors for new applications in radar, sensing and full-duplex systems.

CRedit authorship contribution statement

Girdhari Chaudhary: Writing – original draft, Validation, Methodology, Investigation, Formal analysis. **Yongchae Jeong:** Writing – review & editing, Visualization, Supervision, Project administration, Funding acquisition, Conceptualization.

Declaration of competing interest

The authors declare that they have no known competing financial interests or personal relationships that could have appeared to influence the work reported in this paper.

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Data availability

Data will be made available on request.

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