

## A Miniaturized Tunable Bandpass Filter with Constant Fractional Bandwidth

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**Abstract**—This paper presents a miniaturized tunable bandpass filter, consisting of two coaxial dielectric resonators and a pair of parallel-coupled lines. A coaxial dielectric resonators and a microstrip line form a new step-impedance resonator (SIR), which is different from a conventional SIR. Varactor diodes are connected to SIRs to tune the center frequency. The gap between parallel-coupled lines controls the inter-stage coupling coefficient. Lumped inductors used for coupling to I/O ports can reduce design complexity. The variations of coupling coefficient and external quality factor with tuning frequency are analyzed using HFSS software. A appropriate coupling coefficient which satisfies with constant fractional bandwidth within the tuning range is available. A tunable filter has been made of dielectric ceramics with dielectric constant of 38, fabricated on dielectric substrate and measured using Networks analyzer. Center frequencies vary from 0.43 GHz to 0.78 GHz, 3 dB fractional bandwidth from 6.4% to 6.8% when bias voltages are applied from 0 V to 10 V. The measured results validate the approach and agree with the simulation.

### 1. INTRODUCTION

Tunable bandpass filters (BPF) are key devices for multiband communication and radar systems due to their potential to greatly reduce system volume and complexity. Tunable bandpass filters with constant fractional bandwidth (CFBW) or absolute bandwidth (CABW) have attracted a lot of attention [1–20]. The design approach, which is satisfied with the requirement of keeping constant bandwidth across tuning range, is to control the inter-stage coupling coefficient and external quality factor, e.g., for filters with CFBW, the inter-stage coupling coefficient and external quality factor keeping unchanged with tuning frequencies, for filters with CABW, the inter-stage coupling coefficient varying inversely with tuning frequencies and external quality factor being proportional to tuning frequencies. Introducing a mixed electric and magnetic coupling between open-loop resonators is effective to realize constant bandwidth [1–7], because open-loop resonators can easily realize various couplings by changing the position, at which resonators couple to each other. Comblines or modified combline resonators are also used to realize tunable filter with CABW [8–15]. Wang et al. [9] introduced a ring resonator to improve the passband characteristics of the BPF. Kim and Yun [10] used step-impedance microstrip lines to control coupling. Park and Rebeiz [11] designed low-loss tunable filters with three different fractional bandwidth variations including CFBW. Zhao et al. [12] mixed combline and split-ring resonators to obtain tunable filter with CABW. El-Tanani and Rebeiz [13] proposed microstrip corrugated coupled-line to maintain a nearly CABW across the tuning range. Zhao et al. [14] mixed microstrip combline and pseudo-combline resonators to design tunable filters with CABW. Besides coupled microstrip lines, lumped elements LC and microstrip LC were used to design tunable BPF [16–18]. Lee and Sarabandi [19]

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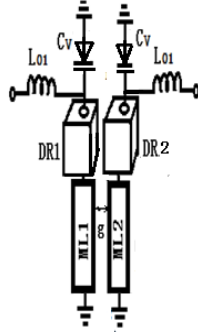
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and Kapilkovich [20] employed step-impedance planar resonators and the equalizing circuits of L- and T-types lumped capacitances to construct a tunable BPF with CABW. Chaudhary et al. [21, 22] used dual-mode microstrip resonators to design two tunable dual-band bandpass filters, one with CFBW and broad harmonic suppressed characteristics and the other with independently tunable center frequencies and bandwidths. Tunable microstrip line BPFs are fabricated on a dielectric substrates. Substrates have low dielectric constant ( $\epsilon_r$ ), usually less than 10. The lengths of microstrip lines are longer than that of coaxial dielectric resonators, because coaxial dielectric resonators are made of high  $\epsilon_r$  dielectric ceramics, in which  $\epsilon_r$  is in the range of  $10 \sim 100$ . However, tunable coaxial dielectric filters are seldom reported in literatures.

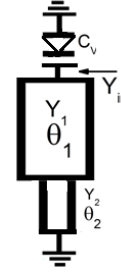
In this paper, we propose a miniaturized tunable combline bandpass filter composed of coaxial dielectric resonators mixed with parallel-coupled microstrip lines. A coaxial dielectric resonator and microstrip line form a SIR, which greatly reduces the size. The tuning of the filter passband is implemented using varactors in series with coaxial dielectric resonators. The coupling between SIRs satisfies the requirement of keeping fractional bandwidth constant across the tuning range. The lumped inductors used for coupling to I/O ports can reduce design complexity.

## 2. DESIGN PROCEDURE

Figure 1 shows a schematic diagram of the tunable bandpass filter, consisting of a pair of parallel-coupled microstrip lines and two coaxial dielectric resonators loaded with varactors. Coaxial dielectric resonators, whose inner radius is 0.75 mm and side length 5 mm, are half wavelength with two ends open-circuited.



**Figure 1.** Proposed tunable filter.



**Figure 2.** The structure of a single resonator.

### 2.1. Resonant Frequency

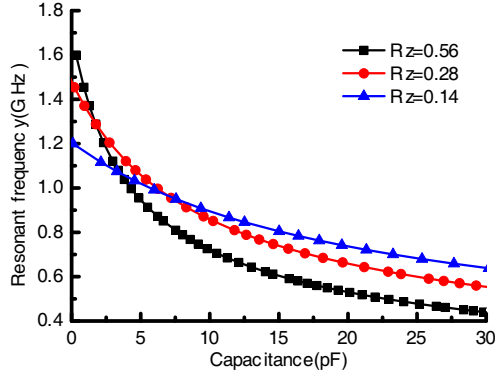
For convenience, the open-ended  $\lambda g/2$  coaxial dielectric resonator is simplified to a transmission line with characteristic admittance  $Y_1$ , electrical length  $\theta_1$  and physical length  $l_1$ . The parameters of a microstrip line are denoted by  $Y_2$ ,  $\theta_2$  and  $l_2$ , respectively, as shown in Figure 2.

The resonant frequency of a single resonator can be estimated by the following formula,

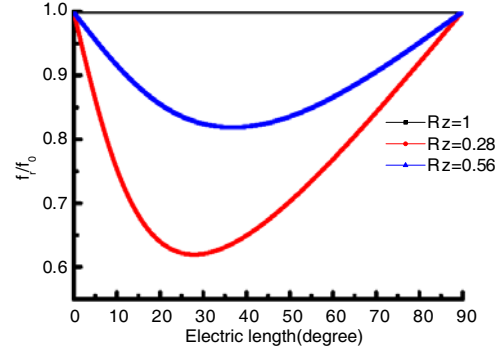
$$Y_{in} = j\omega C_V + jY_2 \frac{-Y_1 \cot \theta_1 + Y_2 \tan \theta_2}{Y_2 + Y_1 \cot \theta_1 \cdot \tan \theta_2} = 0 \quad (1)$$

Figure 3 shows the variation of resonant frequency with the capacitance of  $C_v$  when  $R_z$  gets three values,  $R_z$  is a ratio of  $Y_2$  to  $Y_1$ . It is found that the smaller is  $R_z$ , the lower is the resonant frequency of the resonator with capacitance zero, and the higher is the linearity of the variation of resonant frequency with capacitance  $C_v$ .

In fact, the resonator shown in Figure 2 is a step-impedance resonator (SIR) with  $R_z$  less than one. Figure 4 shows normalized resonant frequency ( $f_r/f_0$ ) of the resonator loaded without varactor, and horizontal ordinate represents the electrical length of a microstrip line,  $\theta_2$ . The normalized resonant



**Figure 3.** Variation of resonant frequency with capacitance.



**Figure 4.** Relationship between normalized resonant frequency and electric length of a short-ended microstrip line.

frequency attains a minimum value when  $\theta_2$  is equal to  $\theta_1$ . The minimum total physical length ( $l_1 + l_2$ ) can be expressed as follows [23],

$$(l_1 + l_2)_{\min.} = (\tan^{-1} \sqrt{R_z}) \times c / (\pi \times f_0 \times \sqrt{\varepsilon_{ref}}) \quad (2)$$

where  $f_0$  is the resonant frequency of a uniform characteristic impedance resonator,  $c$  the velocity of light, and  $\varepsilon_{ref}$  the effective dielectric constant of a microstrip line.

This paper proposes a new method to greatly reduce the physical length of a tunable bandpass filter. The low impedance line is realized by using a coaxial dielectric resonator made of ceramics with high dielectric constant ( $\varepsilon_r = 38$ ). The high impedance line is a microstrip line formed on a substrate with dielectric constant of 2.2 and thickness 0.5 mm, as shown in Figure 2. Compared to a conventional SIR made of microstrip lines, Table 1 lists the calculated results, assuming that the resonant frequency is 1.2 GHz and that the characteristic impedances of microstrip line and dielectric resonator are  $50 \Omega$  and  $14 \Omega$ , respectively.

Table 1 indicates that the physical length of the proposed SIR is much shorter than that of a conventional SIR when both electrical lengths are nearly the same.

**Table 1.** Physical length and electric length of a SIR.

| $R_z = 0.28$     | $\theta_1$ (deg.) | $\theta_2$ (deg.) | $l_1$ (mm) | $l_2$ (mm) |
|------------------|-------------------|-------------------|------------|------------|
| This work        | 53.26             | 5.92              | 6          | 3          |
| conventional SIR | 27.89             | 27.89             | 14.14      | 14.14      |

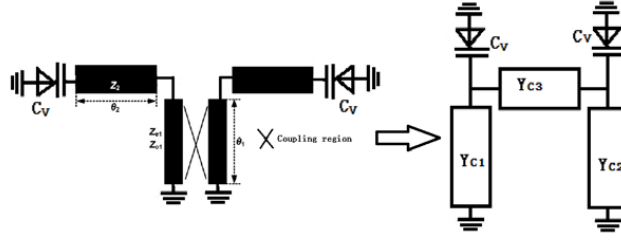
## 2.2. Coupling Coefficient

In Figure 1, a pair of parallel microstrip lines couple to each other, but two coaxial dielectric resonators have no coupling. Figure 5 shows the equivalent circuit of coupled resonators.

In Figure 5,  $Z_{0e}$  and  $Z_{0o}$  are even- and odd-mode characteristic impedances of parallel-coupled microstrip lines, respectively.  $Y_{c1}$  and  $Y_{c2}$  are equivalent admittances of SIRs, and  $Y_{c3}$  is admittance of coupling between a pair of coupled resonators. Referring to [10],  $Y_{c1}$ ,  $Y_{c2}$  and  $Y_{c3}$  are given as the following,

$$Y_{c1} = Y_{c2} = Y_2 \frac{-Y_{1e} \cot \theta_{1e} + Y_2 \tan \theta_2}{Y_2 + Y_{1e} \cot \theta_{1e} \cdot \tan \theta_2} \quad (3)$$

$$Y_{c3} = \frac{jY_2^2 \csc^2 \theta_2}{2} \left( \frac{1}{Y_{1o} \cot \theta_{1o} + Y_2 \cot \theta_2} - \frac{1}{Y_{1e} \cot \theta_{1e} + Y_2 \cot \theta_2} \right) \quad (4)$$



**Figure 5.** Equivalent circuit of a pair of coupled resonators.

where  $Y_{1o}$  and  $Y_{1e}$  are odd- and even-mode admittances of coupled microstrip lines, and  $\theta_{1o}$  and  $\theta_{1e}$  are the odd- and even-mode electrical lengths of coupled microstrip lines.

The coupling coefficient between a pair of coupled resonators can be calculated [10],

$$k_{12} = \frac{|jY_{c3}|}{b} \quad (5)$$

where  $b$  is the slope parameter of SIR loaded varactor, which can be derived from Figure 5 [10],

$$b = \frac{\omega}{2} \frac{d(jY_{c1} + \omega C_V)}{d\omega} \bigg|_{\omega=\omega_0} \quad (6)$$

It is difficult to solve (5) because  $\theta_{1o}$ ,  $\theta_{1e}$ ,  $\theta_2$  and  $\omega$  are variable when  $C_v$  varies. The coupling coefficient can be calculated through eigenmode solution of HFSS simulated model [24],

$$k_{12} = \frac{f_e^2 - f_m^2}{f_e^2 + f_m^2} \quad (7)$$

where  $f_e$  and  $f_m$  are odd- and even-mode resonant frequencies, respectively. Figure 6 presents the variation of coupling coefficient with frequency. In Figure 6,  $g$  is the gap between a pair of parallel-coupled lines, and the smaller is  $g$ , the stronger is the coupling coefficient. It is found that the variation of coupling coefficient with frequency is smaller and that coupling coefficient basically keeps constant and is satisfied with the following equation [25],

$$k_{12} = \frac{BW}{f_0 \sqrt{g_1 g_2}} \approx \text{const} \quad (8)$$

where  $BW$  is fractional bandwidth, and  $g_1$  and  $g_2$  are the first and second normalized element values of a second-order lowpass prototype filter, respectively.

### 2.3. External Quality Factor

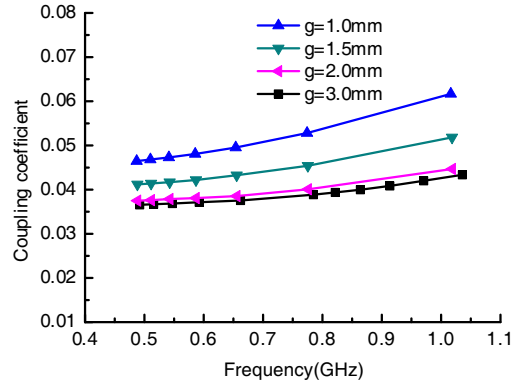
Lumped inductors can be used for input and output coupling structures, as illustrated in Figure 1. The external quality factor ( $Q_e$ ) can be calculated by [10],

$$Q_e = \frac{b}{J_{01}^2 / G_A} \quad (9)$$

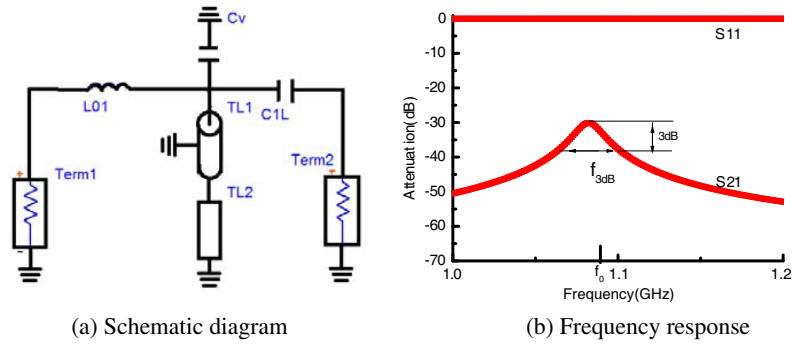
where  $b$  is the slope parameter of the proposed resonator, given as (6),  $G_A$  a conductance of source,  $0.02S$ , and  $J_{01}$  a admittance of input port, given as (10).

$$J_{01} = \frac{1}{\omega L_{01}} \quad (10)$$

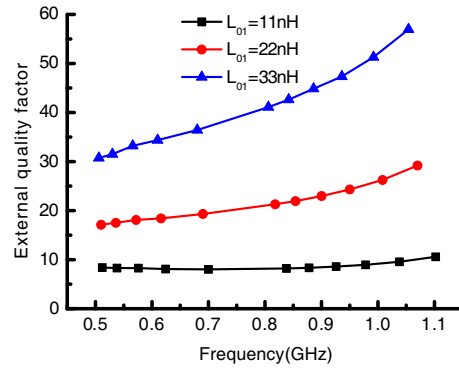
It is difficult to calculate  $Q_e$  based on (9), as  $k_{12}$  described above. The external  $Q$  value can be attained by the single resonator circuit shown in Figure 7. Here, the resonator is coupled to the I/O port by a desired inductor but can also be coupled by a capacitor with very little capacitance, such as 0.01 pF, which allows us to do the transmission measurement shown in the right in Figure 7. The influence of a capacitor can be neglected if the peak of the resonance characteristic is kept below 25 to 30 dB.



**Figure 6.** Variation of coupling coefficient with tuning frequency.



**Figure 7.**  $Q_e$  measurement.



**Figure 8.** External quality factor with tuning frequency.

The external coupling is found by measuring the 3 dB bandwidth of the resonance curve-denoted  $\Delta f_{3dB}$  and resonant frequency-denoted  $f_0$ . The external  $Q$  value is then found by [25],

$$Q_e = Q_{loaded} = \frac{f_0}{\Delta f_{3dB}} \quad (11)$$

Figure 8 shows frequency variation of the external quality factor, which is a simulated curve using HFSS13.0.  $Q_e$  value changes from 10.6 to 57 at 1.1 GHz when  $L_{01}$  varies from 11 nH to 33 nH.  $Q_e$  is unchanged with tuning frequency when  $L_{01}$  is 11 nH, which satisfies the requirement of constant fractional bandwidth, as expressed in (12) [25]

$$Q_e = \frac{f_0 g_0 g_1}{BW} = \text{constant} \quad (12)$$

where  $g_0$  is the source resistance or the source conductance, equal to one, and  $g_1$  is the first normalized element values of a second-order lowpass prototype filter.

### 3. DESIGN EXAMPLE

A tunable bandpass filter using SIRs, which consists of two coaxial dielectric resonators and a pair of parallel-coupled microstrip lines, has been designed. Tuning frequency ranging from 400 MHz to 800 MHz is required. 3 dB fractional bandwidth is about 6%, and second-order is selected to meet stopband attenuation.

As tuning devices, SMV1236-079LF varactors by Skyworks are used. The substrate used for parallel-coupled lines is Rogers RT/duroid 5880 with  $\varepsilon_r$  2.2 and thickness 0.5 mm. Coaxial dielectric resonators, whose cross section is square with inner diameter 1.5 mm and length of a side 5 mm, are made of microwave ceramics  $((\text{Zr}_{0.8}\text{Sn}_{0.2})\text{TiO}_4)$  with  $\varepsilon_r = 38$ ,  $Q \cdot f \geq 40000$ , temperature coefficient of resonant frequency less than 5 ppm/°C. Two 100 k $\Omega$  resistors are used for DC biasing to reduce the RF-signal leakage through the bias networks. Two chip inductors (0603 type, 33 nH) are connected to input and output ports, and two chip capacitors (ATC600S, 47 pF) are in series with diode varactors to block DC.

The design procedure is explained as follows. At first, the center frequency is determined, and the frequency is selected to be a arithmetic average value of  $f_1$  and  $f_2$ ,

$$f_0 = (f_1 + f_2)/2 \quad (13)$$

where  $f_1$  and  $f_2$  are the lower and upper frequencies of tuning range. Next, the resonator electrical length  $(\theta_1 + \theta_2)$  at  $f_0$  is preset, and loaded capacitance ( $C_v$ ) desired is obtained by making use of (1). The minimum capacitance ( $C_{\min}$ ) and maximum capacitance ( $C_{\max}$ ) of the loaded capacitor can be calculated for given lower and upper frequencies in the tuning range. Finally, second-order bandpass filter is designed by using lowpass prototype filter and admittance transformation [25], and coupling coefficient and external quality factor are demanded as following:  $k_{12} = 0.046$ ,  $Q_e = 51.7$ . Based on Figure 6 and Figure 8, physical dimensions of the filter are available and optimized by HFSS13.0. The design parameters and dimensions are shown in Table 2.

In Table 2,  $C_{V0}$ ,  $C_{V\min}$  and  $C_{V\max}$  are the series capacitances of varactor ( $C_V$ ) and chip capacitor (47 pF).

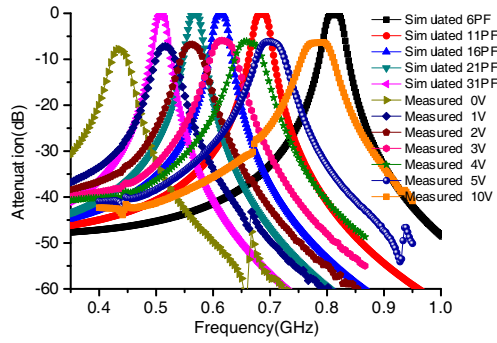
Figure 9 and Figure 10 show simulated responses. The frequencies vary from 0.52 GHz to 0.81 GHz with loaded capacitance changing from 31 pF to 6 pF.

The filter was fabricated on a Rogers RT/duroid 5880 substrate, measured by Networks analyzer E5701B. Measured performance of the proposed filter is presented in Figure 9 and Figure 10, and tuning frequencies vary from 0.43 GHz to 0.78 GHz when bias voltages are applied from 0 V to 10 V, which is a little different from the simulated results because of simulation neglecting parasitic effects of chip components and fabrication producing tolerances.

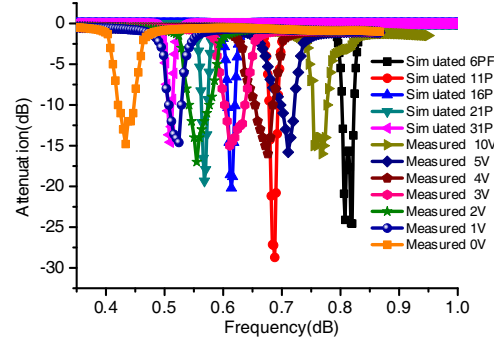
**Table 2.** Design parameters and dimensions.

| Symbol              | Quantity   | Symbol                          | Quantity |
|---------------------|------------|---------------------------------|----------|
| $f_0$               | 600 MHz    | $l_1$                           | 6 mm     |
| $\theta_1$ at $f_0$ | 26.63 deg. | $l_2$                           | 3 mm     |
| $\theta_2$ at $f_0$ | 2.96 deg.  | Width of parallel-coupled lines | 1.5 mm   |
| $C_{V0}$            | 12.26 pF   | $g$                             | 3 mm     |
| $C_{V\min}$         | 6.15 pF    | Width for feedlines             | 1.5 mm   |
| $C_{V\max}$         | 30.83 pF   |                                 |          |

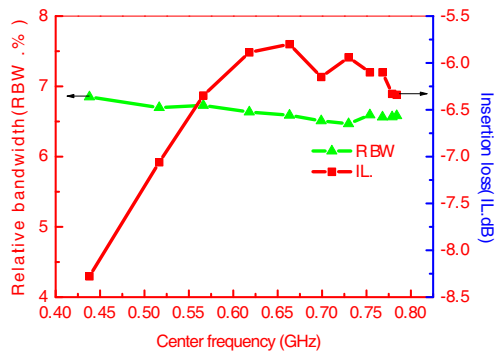
The 3-dB fractional bandwidth and insertion loss of the fabricated filter are plotted in Figure 11. The insertion loss varies from  $-8.28$  dB to  $5.8$  dB, and fractional bandwidth varies from 6.4% to 6.8%, keeping constant within tuning range. Insertion loss is  $-8.28$  dB at 0 V due to low quality factor of varactors at 0 V or lower bias voltages. Figure 12 shows a photo of the proposed tunable filter. The volume of filter containing bias networks is  $24 \times 22 \times 6 \text{ mm}^3$  ( $0.034\lambda_0 \times 0.032\lambda_0 \times 0.0086\lambda_0$ ). Performance comparison of the proposed work with state-of-art is described in Table 3.



**Figure 9.** Simulated and measured transmission coefficients.



**Figure 10.** Simulated and measured reflection coefficients.



**Figure 11.** Variations of insertion loss and passband bandwidth with tuning frequency.



**Figure 12.** Photo of the fabricated filter.

**Table 3.** Performance comparison among tunable filters.

|           | Frequency Tunability (GHz) | 3 dB or 1 dB Bandwidth | Bias voltage (V) | Filter Structure | Area (mm <sup>2</sup> )                                     |
|-----------|----------------------------|------------------------|------------------|------------------|-------------------------------------------------------------|
| [1]       | 0.78 ~ 1.085 (39.1%)       | 50 MHz                 | 0.5–30           | Microstrip line  | ~ 48 × 50<br>(0.125λ <sub>0</sub> × 0.13λ <sub>0</sub> )    |
| [2]       | 0.60 ~ 0.94 (56.7%)        | 63 ± 7 MHz             | 0–30             | Microstrip line  | ~ 52 × 50<br>(0.1λ <sub>0</sub> × 0.1λ <sub>0</sub> )       |
| [5]       | 0.68–1.0 (47%)             | 80 ± 3.5 MHz           | ×                | Microstrip line  | 50 × 34<br>(0.1λ <sub>0</sub> × 0.077λ <sub>0</sub> )       |
| [6]       | 0.71–0.89 (25.3%)          | 35–37 MHz              | 1–5              | Microstrip line  | ~ 30 × 30<br>(0.071λ <sub>0</sub> × 0.071λ <sub>0</sub> )   |
| [11]      | 0.85–1.4 (64.7%)           | 5.5 ± 0.3%             | 2.4–22           | Microstrip line  | 24.7 × 13.6<br>(0.070λ <sub>0</sub> × 0.038λ <sub>0</sub> ) |
| [21]      | 0.85–1.2 (41%)             | 13%                    | 1.5–15           | Microstrip line  | 40 × 35<br>(0.11λ <sub>0</sub> × 0.099λ <sub>0</sub> )      |
| [17]      | 0.481–0.686 (42.6%)        | 19–21 MHz              | ×                | LC               | 56 × 54<br>(0.090λ <sub>0</sub> × 0.086λ <sub>0</sub> )     |
| [18]      | 0.45–0.75 (66.7%)          | 147 ± 1 MHz            | 0–12.5           | LC               | 33 × 15<br>(0.05λ <sub>0</sub> × 0.022λ <sub>0</sub> )      |
| This work | 0.43–0.78 (81.4%)          | 6.4–6.8%               | 0–10             | Mixed resonator  | 24 × 22<br>(0.034λ <sub>0</sub> × 0.032λ <sub>0</sub> )     |

#### 4. CONCLUSION

A miniaturized tunable bandpass filter with constant fractional bandwidth has been designed and fabricated. A new SIR structure which consists of a coaxial dielectric resonator and a microstrip line can greatly reduce the size. A parallel-coupled microstrip lines only provide inter-stage coupling. The desired coupling coefficient keeping constant with tuning frequency can be attained by optimizing the gap of parallel-coupled lines. Inductors are used for input/output coupling. Insertion loss of the fabricated filter can be improved by using high  $Q$  varactors and high  $Q$  chip capacitors and inductors, such as GaAs varactors MA46H200 Series. The structure of the proposed tunable filter will be expected to be used in more than two poles filters and VHF/UHF filters for multi-band communication systems.

#### ACKNOWLEDGMENT

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