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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 CFBW, 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. Combline 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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and Kapilkvich [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 (ε_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 ε_r dielectric ceramics, in which ε_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.

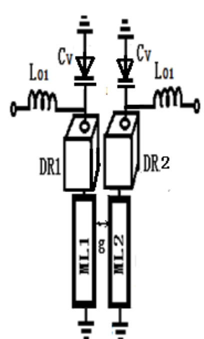


Figure 1. Proposed tunable filter.

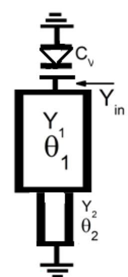


Figure 2. The structure of a single resonator.

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 θ_1 and physical length l_1. The parameters of a microstrip line are denoted by Y_2, θ_2 and l_2, respectively, as shown in Figure 2.

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

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

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, θ_2. The normalized resonant

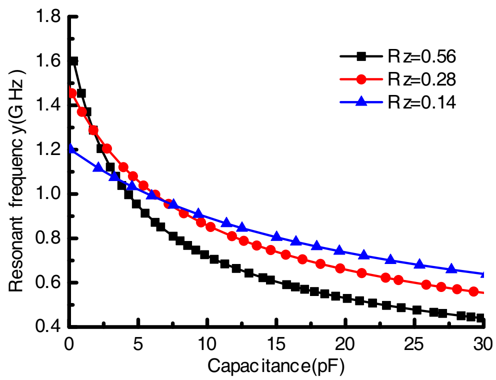


Figure 3. Variation of resonant frequency with capacitance.

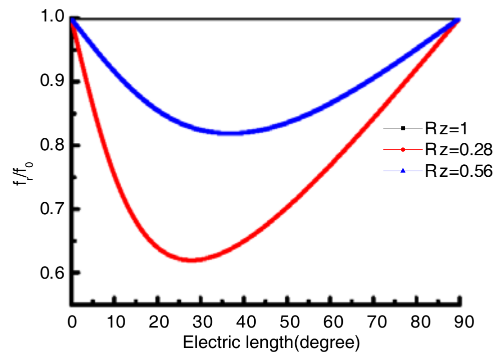


Figure 4. Relationship between normalized resonant frequency and electric length of a short-ended microstrip line.

frequency attains a minimum value when θ_2 is equal to θ_1. The minimum total physical length $(l_1 + l_2)$ can be expressed as follows [23],

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

where f_0 is the resonant frequency of a uniform characteristic impedance resonator, c the velocity of light, and ε_{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 Ω and 14 Ω, 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.

$R_z = 0.28$	θ_1 (deg.)	θ_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,

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

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

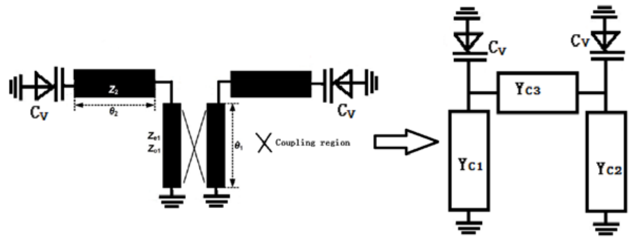


Figure 5. Equivalent circuit of a pair of coupled resonators.

where Y_{1o} and Y_{1e} are odd- and even-mode admittances of coupled microstrip lines, and θ_{1o} and θ_{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],

$$k_{12} = \frac{|jY_{c3}|}{b} \tag{5}$$

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

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

It is difficult to solve (5) because θ_{1o}, θ_{1e}, θ_2 and ω are variable when C_v varies. The coupling coefficient can be calculated through eigenmode solution of HFSS simulated model [24],

$$k_{12} = \frac{f_e^2 - f_m^2}{f_e^2 + f_m^2} \tag{7}$$

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],

$$k_{12} = \frac{BW}{f_0\sqrt{g_1 g_2}} \approx cons\tan t \tag{8}$$

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 (Qe) can be calculated by [10],

$$Q_e = \frac{b}{J_{01}^2/G_A} \tag{9}$$

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).

$$J_{01} = \frac{1}{\omega L_{01}} \tag{10}$$

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.

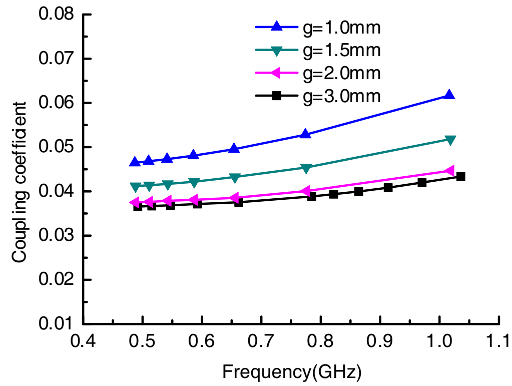
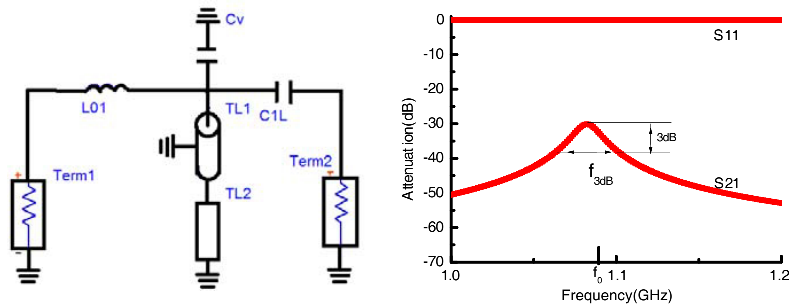


Figure 6. Variation of coupling coefficient with tuning frequency.



(a) Schematic diagram (b) Frequency response

Figure 7. Q_e measurement.

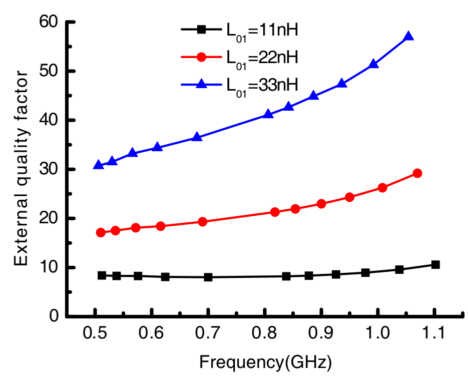


Figure 8. External quality factor with tuning frequency.

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

$$Q_e = Q_{loaded} = \frac{f_0}{\Delta f_{3\,\mathrm{dB}}} \tag{11}$$

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. Qe is unchanged with tuning frequency when L_{01} is 11 nH, which satisfies the requirement of constant fractional bandwidth, as expressed in (12) [25]

$$Q_e = \frac{f_0 g_0 g_1}{BW} = cons\tan t \tag{12}$$

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 ε_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 ($(Zr_{0.8}Sn_{0.2})TiO_4$) with $\varepsilon_r = 38$, $Q \cdot f \geq 40000$, temperature coefficient of resonant frequency less than 5 ppm/°C. Two 100 kΩ 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,

$$f_0 = (f_1 + f_2)/2 \tag{13}$$

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], nad coupling coefficient and external quality factor are demanded as following: $k_{12} = 0.046$, $Qe = 51.7$. 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
θ_1 at f_0	26.63 deg.	l_2	3 mm
θ_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.

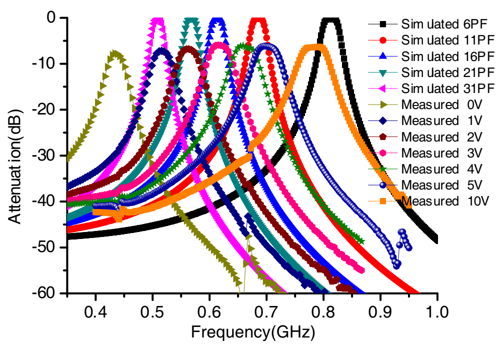


Figure 9. Simulated and measured transmission coefficients.

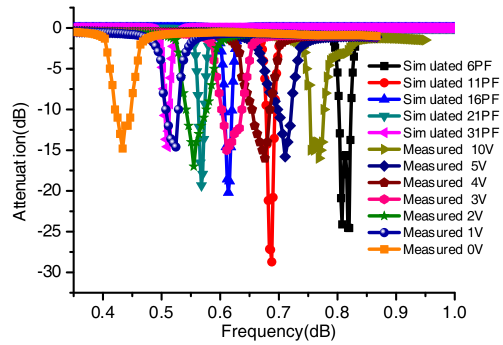


Figure 10. Simulated and measured reflection coefficients.

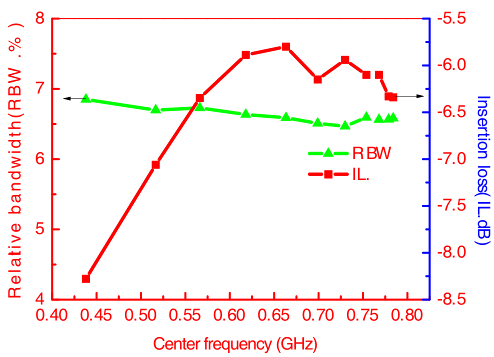


Figure 11. Variations of insertion loss and passband bandwidth with tuning frequency.

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^2)
[1]	0.78 ∼ 1.085 (39.1%)	50 MHz	0.5–30	Microstrip line	∼ 48 × 50 ($0.125\lambda_0 \times 0.13\lambda_0$)
[2]	0.60 ∼ 0.94 (56.7%)	63 ± 7 MHz	0–30	Microstrip line	∼ 52 × 50 ($0.1\lambda_0 \times 0.1\lambda_0$)
[5]	0.68–1.0 (47%)	80 ± 3.5 MHz	×	Microstrip line	50 × 34 ($0.1\lambda_0 \times 0.077\lambda_0$)
[6]	0.71–0.89 (25.3%)	35–37 MHz	1–5	Microstrip line	∼ 30 × 30 ($0.071\lambda_0 \times 0.071\lambda_0$)
[11]	0.85–1.4 (64.7%)	5.5 ± 0.3%	2.4–22	Microstrip line	24.7 × 13.6 ($0.070\lambda_0 \times 0.038\lambda_0$)
[21]	0.85–1.2 (41%)	13%	1.5–15	Microstrip line	40 × 35 ($0.11\lambda_0 \times 0.099\lambda_0$)
[17]	0.481–0.686 (42.6%)	19–21 MHz	×	LC	56 × 54 ($0.090\lambda_0 \times 0.086\lambda_0$)
[18]	0.45–0.75 (66.7%)	147 ± 1 MHz	0–12.5	LC	33 × 15 ($0.05\lambda_0 \times 0.022\lambda_0$)
This work	0.43–0.78 (81.4%)	6.4–6.8%	0–10	Mixed resonator	24 × 22 ($0.034\lambda_0 \times 0.032\lambda_0$)

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.

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REFERENCES

1. Liu, B., F. Wei, Q. Y. Wu, and X. W. Shi, "A tunable bandpass filter with constant absolute bandwidth," *Journal of Electromagnetic Waves and Applications*, Vol. 25, No. 11–12, 1596–1604, 2011.
2. Huang, X.-G., Q.-Y. Feng, Q.-Y. Xiang, and D.-H. Jia, "Constant absolute bandwidth tunable filter using varactor-loaded open-loop resonators," *Microwave and Optical Technology Letters*, Vol. 56, No. 5, 1178–1181, 2014.
3. Jia, D.-H., Q.-Y. Feng, X.-G. Huang, and Q.-Y. Xiang, "A two-pole tunable filter with constant fractional-bandwidth characteristics," *International Journal of Electronics*, Vol. 101, No. 7, 983–993, 2014.
4. Athukorala, L. and D. Budimir, "Open-loop tunable resonators and filters with constant bandwidth," *IET Microw. Antennas Propag.*, Vol. 6, No. 7, 800–806, 2012.
5. Zhang, H.-L., X. Y. Zhang, and B.-J. Hu, "Tunable bandpass filters with constant absolute bandwidth," *9th International Symposium on Antennas Propagation and EM Theory (ISAPE)*, 1200–1203, 2010.
6. Yu, F. L., Y. B. Zhang, X. Y. Zhang, B. J. Hu, and X. Y. Wang, "Tunable bandpass filters with constant absolute bandwidth and high linearity," *International Conference on Microwave and Millimeter Wave Technology (ICMMT)*, 1–3, 2012.
7. Makimoto, M. and M. Sagawa, "Varactor tuned bandpass filters using microstrip-line ring resonators," *IEEE MTT-S Int. Microwave Symp. Dig.*, 411–414, 1986.
8. Hunter, I. C. and J. D. Rhodes, "Electronically tunable microwave bandpass filter," *IEEE Trans. Microwave Theory Tech.*, Vol. 30, 1354–1360, 1982.
9. Wang, Y.-Y. F. Wei, B. Liu, H. Xu, and X.-W. Shi, "A tunable bandpass filter with constant absolute bandwidth based on one ring resonator," *Journal of Electromagnetic Waves and Applications*, Vol. 26, No. 11–12, 1587–1593, 2012.
10. Kim, B.-W. and S.-W. Yun, "Varactor-tuned combline bandpass filter using step-impedance microstrip lines," *IEEE Trans. Microwave Theory Tech.*, Vol. 52, No. 4,1279–1284, 2004.
11. Park, S.-J. and G. M. Rebeiz, "Low-loss two-pole tunable filters with three different predefined bandwidth characteristics," *IEEE Trans. Microwave Theory Tech.*, Vol. 56, No. 5, 1137–1148, 2008.
12. Zhao, Z., J. Chen, L. Yang, and K. Chen, "Three-pole tunable filters with constant bandwidth using mixed combline and split-ring resonators," *IEEE Microwave and Wireless Components Letters*, 1–3, 2014.
13. El-Tanani, M. A. and G. M. Rebeiz, "Corrugated microstrip coupled lines for constant absolute bandwidth tunable filters," *IEEE Trans. Microwave Theory Tech.*, Vol. 58, No. 4, 956–963, 2010.

14. Zhao, Z.-Y., J. Chen, L. Yang, and K.-H. Chen, "Three-pole tunable filters with constant bandwidth using mixed combline resonators," *International Journal of Electronics*, 1–15, 2015.
15. Jia, D., Q. Feng, X. Huang, and Q. Xiang, "Tunable 3-pole bandpass filter with constant absolute bandwidth and > 60 dB rejection at desired location of Stopband," *Journal of Electromagnetic Waves and Applications*, 1–14, 2015.
16. Xiang, Q., Q Feng, X. Huang, and D. Jia, "Electrical tunable microstrip LC bandpass filters with constant bandwidth," *IEEE Trans. Microwave Theory Tech.*, Vol. 61, No. 3, 1124–1130, 2013.
17. Lee, J.-H., J.-W. Choi, X.-G. Wang, and S.-W. Yun, "Design of tunable bandpass filter using PIN diode with constant absolute bandwidth," *Asia-Pacific Microwave Conference Proceedings*, 191–193, 2013.
18. Kholodnyak, D., V. Turgaliev, and A. Baskakova, "A method to design lumped-element tunable bandpass filters with constant absolute bandwidth," *Proceedings of the 44th European Microwave Conference*, 335–338, 2012.
19. Lee, J. and K. Sarabandi, "An analytic design method for microstrip tunable filters," *IEEE Trans. Microwave Theory Tech.*, Vol. 56, No. 7, 1699–1706, 2008.
20. Kapii, K. B., "A vacartor-tunable filter with constant bandwidth and loss compensation," *Microwave Journal*, 105–110, 2007.
21. Chaudhary, G., Y. Jeong, and J. Lim, "Harmonic suppressed dual-band bandpass filter with tunable passbands," *IEEE Trans. Microwave Theory Tech.*, Vol. 60, No. 7, 2115–2123, 2012.
22. Chaudhary, G., Y. Jeong, and J. Lim, "Dual-band bandpass filter with independently tunable center frequencies and bandwidths," *IEEE Trans. Microwave Theory Tech.*, Vol. 61, No. 1, 107–116, 2013.
23. Makimoto, M. and S. Yamashita, *Microwave Resonators and Filters for Wireless Communication: Theory, Design and Application*, Springer-Verlag, Berlin Heidelberg, 2001.
24. Hong, J.-S., *Microstrip Filters for RF Microwave Applications*, 3rd Edition, J. Wiley & Sons, 2011.
25. Matthaei, G. L., E. Young, and E. M. T. Jones, *Microwave Filters, Impedance-matching Networks, Coupling Structures*, Artech House, Norwood, MA, USA, 1980.