# Inline Cavity Stepped Window Bandpass Filter With Two Transmission Zeros

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***Abstract*—This paper proposed the design of bandpass filter (BPF) based on stepped window resonators (SWRs) which allows inline resonators to produce transmission zero (TZ). Compared with traditional cavity resonator using iris-coupled structure, the proposed SWR consists of two metal cavities, the larger cavity is regarded as a bandstop resonator to provide TZ near the passband. Therefore, each SWR pair can generate a transmission pole (TP) and a TZ. Two cavities are directly cascaded without any coupling iris, which increases the interface window between these two metal cavities and reduces the fabrication tolerance. By properly dimensioning the bandstop cavity, TZs can be generated in the lower and upper stopbands to improve the frequency-selective performance. An equivalent circuit model of SWR is proposed to explain the generation of TZ and synthesize the BPF. Finally, a 4$^{th}$-order inline BPF with a fractional bandwidth of 1% at 10 GHz is designed using this SWR approach. The synthesis response and simulation results match well. The proposed BPF is fabricated and measured, the good agreement between the measured and simulated results verifies the proposed design methodology.**

***Index Terms*—Stepped window resonator (SWR), transmission zeros (TZs), inline topology, equivalent circuit model.**

## I. Introduction

MICROWAVE bandpass filters (BPFs) are widely used in microwave communication systems. Microstrip BPFs are very popular due to its planar structure [1]-[2]. However, compared with waveguide resonators, insertion loss (IL) and power capacity of microstrip BPFs are not competitive. Due to the high quality-factor and low insertion loss, waveguide cavity filters are commonly designed and used in base station and satellite communication systems. Traditional cavity filters are designed by cavity resonators [3]-[18], coaxial resonators [19]-[20], and dielectric resonators [21]-[22]. Iris/window resonators are rarely investigated to design microwave cavity filters [23]-[35]. Our previous works [23]-[24] use iris resonators to improve the frequency selectivity and get wider bandwidth for both filter and antenna. A slot filtenna using iris resonators is proposed in [23], it increases bandwidth by 20% without significantly increasing the circuit size. In [24], a cavity filtering crossover is presented, the structure excites the resonance modes in cavities and the iris resonators simultaneously. The iris resonator contributes a wider fractional bandwidth (24%) and a compact size as the iris resonator provides additional resonant mode without adding extra circuit size. In [25]-[29], some waveguide BPFs based on resonant irises have been presented, but the frequency selective performance is not very good since there is no TZ in the frequency response. In addition, the design formula in [25] is an empirical equation, which is only 50% accuracy. A highpass filter with three TZs at lower-stopband is presented in [30], the filter uses both cavities and irises as resonators simultaneously to make the passband wider. Two third-order BPFs with a TZ are presented in [31] and [32], three TPs consist of two cavity resonators and an iris resonator. Additionally, the TZ in [31] is attributed to higher-order modes which are excited in two cavities while the TZ in [32] is provided by the crossing coupling through the iris. Recently, some inline quasi-elliptic BPFs [33], which use the iris resonators are designed. It is noted that there are some TZs in the filters with inline topology, which improves the frequency selective performance. Moreover, two of these filters are fabricated by metal 3-D printing technology, which is easy to manufacture and reduces the cost of manufacturing. Three dual-band filters are presented in [34], cavity and iris resonators are used to get two passbands, respectively. Moreover, the capacitive stubs are utilized to obtain TZs between the two passbands. Finally, the filter is cascaded with staircase configuration to suppress the first spurious passband. In [35], a wideband BPF with an inline structure is proposed by utilizing both cavity and iris resonators. The bandwidth is from 6.75 GHz to 8.25 GHz because of the iris resonators, but the first spurious passband is from 9.5 GHz. Therefore, the filter is cascaded with the staircase configuration to suppress the first spurious passband, which increases the volume of the filter. This filter is designed by WR-90, but it does not work in this frequency range, it needs a waveguide transformer to convert the input to the desired frequency range, which increases the complexity and volume of the filter circuit.

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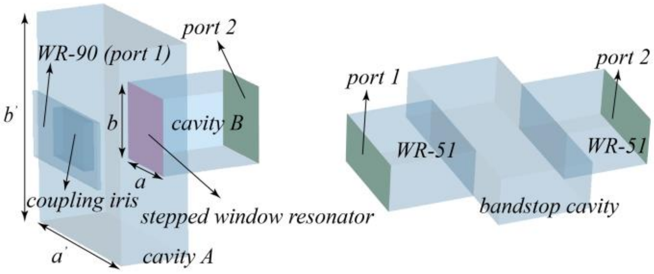


Fig. 1. Three-dimensional (3-D) view: (a) Proposed stepped window resonator. (b) Bandstop singlet in [13].

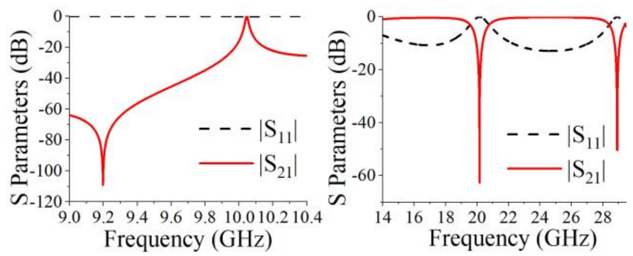


Fig. 2. Comparison of frequency responses: (a) Proposed SWR. (b) Bandstop singlet in [13].

The cavity filters with TZs are required in various communication systems. Traditionally, it is difficult to introduce a TZ in the inline topology, novel coupling paths need to be introduced into the inline topology for generating TZs. Some papers [10]-[14] introduce some novel resonant elements in the inline topology, or some novel coupling methods to create TZs. These works greatly contribute to the generation of TZs for inline topology structures. In [10], a novel coupling structure which is a wire inside the cavity to realize the bypass coupling is presented, the structure enables bypass coupling without introducing additional cavities and enhances the frequency selectivity. In [11], a non-resonating mode waveguide filter is presented, the filter consists of a cavity and two irises for coupling. $TE_{201}$ mode is excited at the cavity and $TE_{10}$ mode generates the source-load coupling to get a TZ. The filter in [11] is used to design two filters in [12] by cascading some filters in [11]. A stopband singlet is presented in [13], the singlet consists of three cavities and two irises. There are two poles and a TZ, and the $TE_{301}$ mode is excited at the coupled structure to generate source-load coupling. The structure is easy to fabricate. Therefore, the singlet can be used for designing waveguide filters. In [14], the singlet proposed in [13] is used to design a filter that has six poles and two TZs. It is worth noting that the paper [14] focuses on the synthesis design approach for waveguide cavity filters. The filter consists of two singlets in [14] and two cavities are an example of the synthesis design approach.

All the filters that use iris resonators are designed to make the passband wider, none of them are narrowband filters with iris resonators. In this paper, a stepped window BPF with two TZs is first presented. The fractional bandwidth is 1% at 10 GHz. The proposed BPF is fabricated by full metal cavities, and the stepped cavities produce four stepped window resonators (SWRs). Besides, the equivalent circuit models are created to explain the proposed SWR and BPF. Finally, compared with the inline cavity filters with TZs in [14] and [35], the volume of this proposed stepped window BPF is smaller.

## II. Stepped Window Resonator With TZ

This section demonstrates the study of the proposed stepped window resonator and its equivalent circuit model of the proposed resonator structure. Non-resonating discontinuity is a very common filter design method that has been applied to both microstrip filters and cavity filters. Therefore, the stepped window resonator filter proposed in this article may be misunderstood and considered to be the same structure as in [13]. What we would like to emphasize is that the stepped window filter proposed in this paper resonates at the window region in between the large and small cavity, which uses the length of the rectangular waveguides that creates the SWR to adjust the coupling between two window resonators. In contrast, the traditional cavity filters use the cavity itself as a resonator and adjust the coupling strength through the size of iris, which is the most essential difference between the structure proposed in this paper and the traditional filters.

Fig. 1(a) illustrates the three-dimensional (3-D) view of the proposed SWR. Obviously, it is shown in Fig. 1(a) that the stepped window resonator consists of two different cavities (Cavity A and B), the window (purple part) created by the discontinuity is a resonator. A standard waveguide (WR-90) is used as the input waveguide feeding port and an iris controls the coupling strength between the input and cavity A. Fig. 1(b) gives the three-dimensional (3-D) view of the stopband singlet in [13], it consists of two standard waveguides (WR-51) and a bandstop cavity. The structure in [13] uses a waveguide cavity as a resonator, and the input port is coupled to waveguide cavity through an iris. In order to prove that it is not resonating in the window and is different from our proposed SWR, the size of the coupling irises is enlarged to be same as the waveguide, and the input and output ports are directly added to the WR-51. The designed SWR and the singlet in [13] are simulated in full-wave commercial software CST. Fig. 2(a) depicts the frequency response of the proposed SWR. Obviously, there is a transmission pole (TP) at 10.05 GHz and a TZ at 9.20 GHz in Fig. 2(a), the TP is created by the SWR. Besides, Fig. 2(b) gives frequency response of the stopband singlet in Fig. 1(b). There is only a bandstop response at 20.16 GHz. Therefore, we can conclude that the singlet resonates at two cavities instead of the two windows in [13].

The two structures are simulated by eigenmode solver in CST. Fig. 3(a)-(b) shows the magnetic and electric field distributions of proposed SWR in two cavities. By looking at the field distribution, different modes in different cavities are excited. It can be seen from Fig. 3(a)-(b) that $TM_{120}$ mode is excited in cavity A, while $TE_{101}$ mode is excited in the adjacent cavity B. Fig. 3(c)-(d) depicts the surface current distribution of the proposed SWR and the bandstop singlet in [13]. It is obvious that the surface current distributions of the two structures are different, the SWR resonates at the stepped window, while the bandstop singlet does not resonate at the window.

The high current density is caused by the presence of discontinuities in both E-plane and H-plane inside the cavity, which stores the electric and magnetic energy and produces parasitic inductance and parasitic capacitance around the stepped window structure [37]. With reference to [37], when an electromagnetic wave propagates in a rectangular waveguide and the magnetic field is compressed. When the electric and magnetic fields pass through the stepped window, they are simultaneously compressed in both directions. Thus, a parallel resonance circuit is equivalent to this SWR, as shown in Fig. 4(e)-(f). Therefore, when the condition of stored electric field energy equal to the stored magnetic field energy is satisfied, the stepped window can be regarded as a resonator [37].

The cavity A and cavity B in Fig. 1(a) can be regarded as transmission lines in the equivalent circuit. Therefore, the characteristic impedances of the two transmission lines are matched when the stepped window resonates. According to [25] and [37], the characteristic impedances of TE and TM modes satisfy following equations:

$$Z_{cTE_{101}} = \frac{\eta}{\sqrt{1-(\frac{v}{2af_0})^2}} \quad (1),$$

$$Z_{cTM_{120}} = \eta\sqrt{1-\left(\frac{v}{2a'f_0}\right)^2-\left(\frac{v}{b'f_0}\right)^2} \quad (2),$$

where $a$, $b$, $a'$, and $b'$ are the dimensions marked in Fig. 4(e). $\eta$ is the wave impedance of the cavity in the free space and $\lambda$ is the wavelength. $f_0$ and $v$ are the central frequency and propagation speed of electromagnetic waves in the free space. Since the impedance of $TE_{101}$ mode and $TM_{120}$ mode is matched, we get:

$$Z_{cTE_{101}} = Z_{cTM_{120}} \quad (3).$$

From (1)-(3), we solve the expression of central resonance frequency:

$$f_0 = \frac{v}{2}\sqrt{\frac{b'^2+4a'^2}{a^2b'^2+4a^2a'^2+a'^2b'^2}} \quad (4).$$

However, this formula is derived from impedance matching condition, which does not consider the parasitic effect caused by the discontinuity of the stepped window. A correction of (4) is needed, we multiply $a'$ by a correction factor $p$:

$$f_0 = \frac{v}{2}\sqrt{\frac{b'^2+4(pa')^2}{a^2b'^2+4(pa^2a')^2+(pa')^2b'^2}} \quad (5).$$

By changing the value of $p$ and comparing the calculation results with the simulation results, we find that the two results are in good agreement when $p$ is 1/3, which is shown in Fig. 5(a). Therefore, the final expression is as follows:

$$f_0 = \frac{v}{2}\sqrt{\frac{9b'^2+4a'^2}{9a^2b'^2+4a^2a'^2+a'^2b'^2}} \quad (6).$$

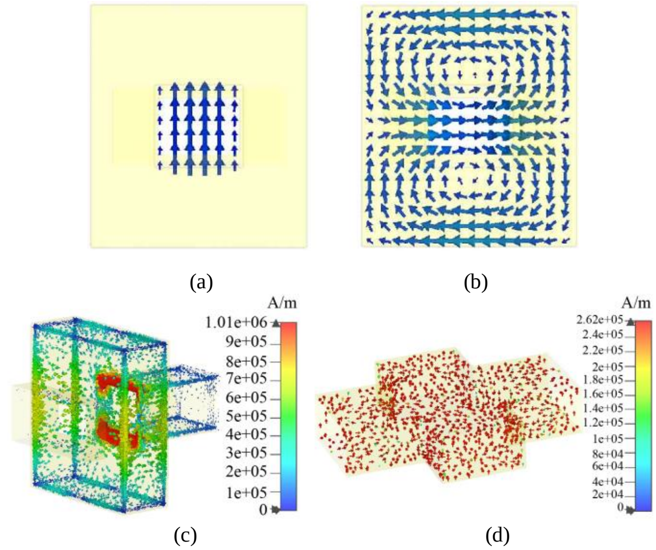


Fig. 3. (a) $TE_{101}$ mode in cavity B. (b) $TM_{120}$ mode in cavity A. (c) Surface current distribution on the stepped window. (d) Surface current distribution on the bandstop singlet in [13].

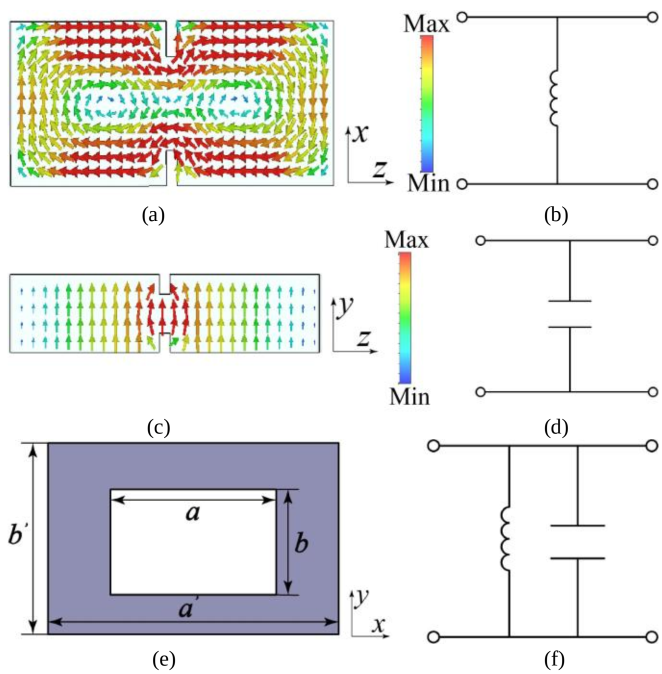


Fig. 4. (a) Symmetrical inductive diaphragm, (b) Equivalent circuit, (c) Symmetrical capacitive diaphragm, (d) Equivalent circuit, (e) Rectangular resonant iris, (f) Equivalent circuit.

For verification of (6), some results are shown in Fig. 5(a), variable $m$ on the x-axis represents the multiplication factor by which all filter dimensions are scaled up or down. $f_{0sim}$ and $f_{0cal}$ are the central resonance frequencies of the proposed SWR that are simulated in CST and calculated by equation (6), respectively. It can be found that the simulation and calculation results are in good agreement, which shows that the derived equation (6) can accurately predict the resonator frequency within a wide frequency band ranging from 3-40 GHz. It is shown in Fig. 5(b) that changing the size ratio in the structure keeps the difference between $f_{0sim}$ and $f_{0cal}$ within a small range.

In order to explain the generation of the TZ and the factors

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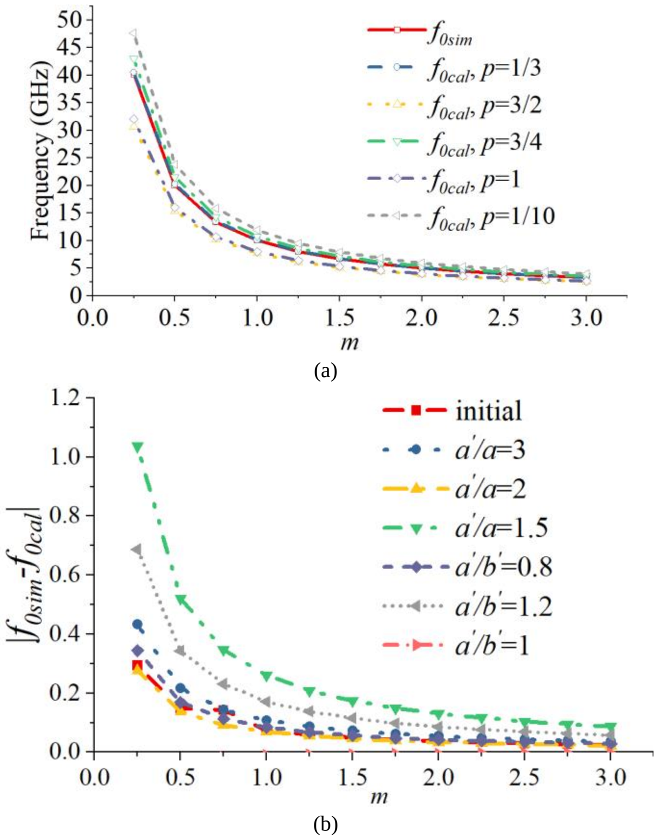



Fig. 5. (a) Resonant frequency comparison of SWR between EM simulated and calculated results in (5). (b) The effect of changing the size ratio in the structure on the difference between $f_{0sim}$ and $f_{0cal}$.

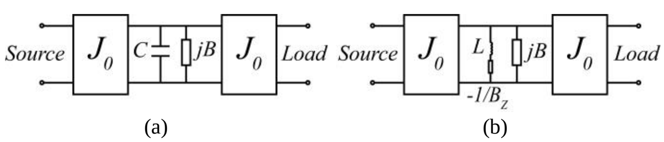



Fig. 6. Lowpass prototype networks: (a) Bandpass resonator, (b) Bandstop resonator.

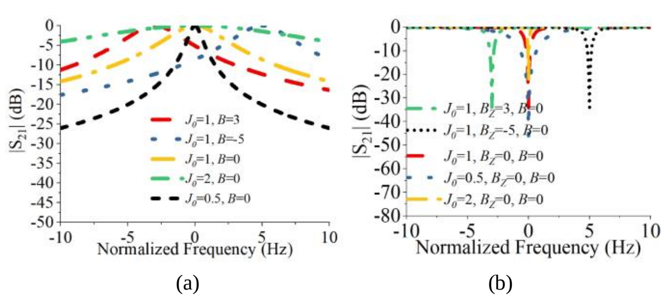



Fig. 7. Different S-parameters versus the changes of circuit component values: (a) Normalized frequency response of bandpass resonator, (b) Normalized frequency response of bandstop resonator.

which have an influence on the TZ and TP, an equivalent circuit model will be established. In general, the first step in modeling an equivalent circuit model is to design a low-pass circuit model at normalized frequency. The frequency transformation formula (7) [36] can be used to convert the desired frequency to the normalized frequency:

$$F = \frac{1}{FBW}\left(\frac{f}{f_c} - \frac{f_c}{f}\right) \quad (7),$$

where $f_c$ is central frequency, $FBW$ is the fractional bandwidth of the bandpass filter, and $F$ is the normalized frequency. Each component also has a corresponding transformation process to complete the modeling of the equivalent circuit. Therefore, we can first design the model of the normalized frequency domain, and then convert it to the desired frequency through the transformation formula.

The number of lumped capacitance and lumped inductance which are frequency-dependent reactive elements determines the order of the lowpass filter prototype network. In a ladder network, capacitance and inductance are interchanged by the dual-network theorem. Frequency-invariant reactance (FIR) elements can be used to design resonators, where FIR represents the offset between the resonant frequency of the resonator and the nominal frequency. Before modeling the entire circuit, there is some study about two kinds of basic resonators: bandpass resonator and bandstop resonator. Fig. 6 illustrates the lowpass prototype networks of the bandpass resonator and bandstop resonator, respectively. $J_0$ represents the $J$ inverter, $jB$ and $jB_z$ are FIRs, which control offset of the resonant frequency, $C$ and $L$ are capacitance and inductance, respectively.

The process of establishing the equivalent circuit model starts from the transmission (ABCD) matrixes of each unit. The transmission (ABCD) matrixes of units are as follows:

$$[A_{\mathrm{J}}] = \begin{bmatrix} 0 & j/J \\ jJ & 0 \end{bmatrix} \quad (8),$$

$$[A_{\mathrm{C}}] = \begin{bmatrix} 1 & 0 \\ jfC & 1 \end{bmatrix} \quad (9),$$

$$[A_{\mathrm{jB}}] = \begin{bmatrix} 1 & 0 \\ jB & 1 \end{bmatrix} \quad (10),$$

$$[A_{\mathrm{LB_Z}}] = \begin{bmatrix} 1 & 0 \\ (jfL + jB_Z)^{-1} & 1 \end{bmatrix} \quad (11).$$

From (8)-(11), we get the transmission (ABCD) matrixes of two basic resonators:

$$[A_{\mathrm{BPR}}] = [A_{\mathrm{J_0}}][A_{\mathrm{C}}][A_{\mathrm{jB}}][A_{\mathrm{J_0}}] \quad (12),$$

$$[A_{\mathrm{BSR}}] = [A_{\mathrm{J_0}}][A_{\mathrm{LB_Z}}][A_{\mathrm{jB}}][A_{\mathrm{J_0}}] \quad (13).$$

With reference to [37], we have the transformation formula for transmission (ABCD) matrix to S-parameters:

$$S_{11} = \frac{A + B/Z_0 - CZ_0 - D}{A + B/Z_0 + CZ_0 + D} \quad (14),$$

$$S_{21} = \frac{2}{A + B/Z_0 + CZ_0 + D} \quad (15),$$

where $Z_0$ is the port impedance of the equivalent circuit. From (8)-(15), we solve the $S_{11}$ and $S_{21}$. Next, the numerators of $S_{11}$ and $S_{21}$ are extracted and solved, and the resulting roots are the frequency of the TP:

$$f_{0BPR} = -B \quad (16),$$

$$f_{0BSR} = -B_z \quad (17),$$

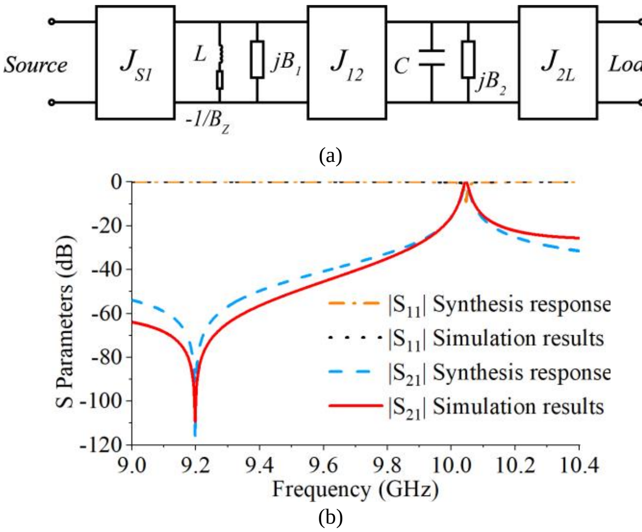


Fig. 8. (a) Equivalent circuit model of the SWR (b) Comparison of $|S_{21}|$ between simulation response and synthesis response.

where the $f_{0BPR}$ and $f_{0BSR}$ are the resonance frequency of the bandpass and bandstop resonators, respectively. Fig. 7(a) and 7(b) shows the relationship between the values of the circuit components and the S-parameters of the entire circuit in Fig.6(a) and 6(b) respectively.

It is illustrated in Fig. 7(a) that various normalized frequency responses can be obtained by setting different $B$ while $C$=$L$=1. When the value of $B$ is the same, the selectivity of the curve can be changed by adjusting the value of $J$. Moreover, the selectivity of the curve will reduce by increasing the value of $J$. In contrast, as shown in Fig. 7(b), the selectivity of the curve will increase when $J$ increases. And the location of the bandstop resonance point is as described in formula (17).

The TZ could be analyzed by an equivalent circuit of the proposed SWR. Cavity A in Fig. 1(a) is a bandstop resonator in the proposed SWR, and the equivalent circuit model of the SWR is depicted in Fig. 8(a). Besides, Fig. 8(b) gives the comparison of $|S_{21}|$ between simulation results and synthesis response where the $J_{S1}$=0.08, $J_{12}$=0.08, $J_{2L}$=0.30, $B_Z$=103.80, $B_1$=0, $B_2$=-0.50, $L$=$C$=1.00. It is obvious that the synthesis response and simulation results match well.

In the equivalent circuit, the position of the TPs and the TZ can be adjusted by changing the values of the FIR elements. Correspondingly, the sizes of the structure can control the resonance frequency and TZ. Moreover, Fig. 9(a)-(b) illustrates the relationships among central frequency, TZ and the size of cavity A. It is obvious that the central resonance frequency and TZ are moving to the lower frequencies with the rises of $a^{'}$ and $b^{'}$. This provides two good design parameters for fine-tuning the resonance frequency and TZ to the desired frequency. Meanwhile, the TZ and resonance frequency can also be adjusted by the size of cavity B. It is shown in Fig. 9(b) that the resonance frequency is moving to the lower frequencies with the increase of $a$. The position of the TZ is not affected by $a$, but will move to the higher frequencies by increasing $b$. These physical parameters correspond to those of the equivalent circuit in Fig. 8(a).

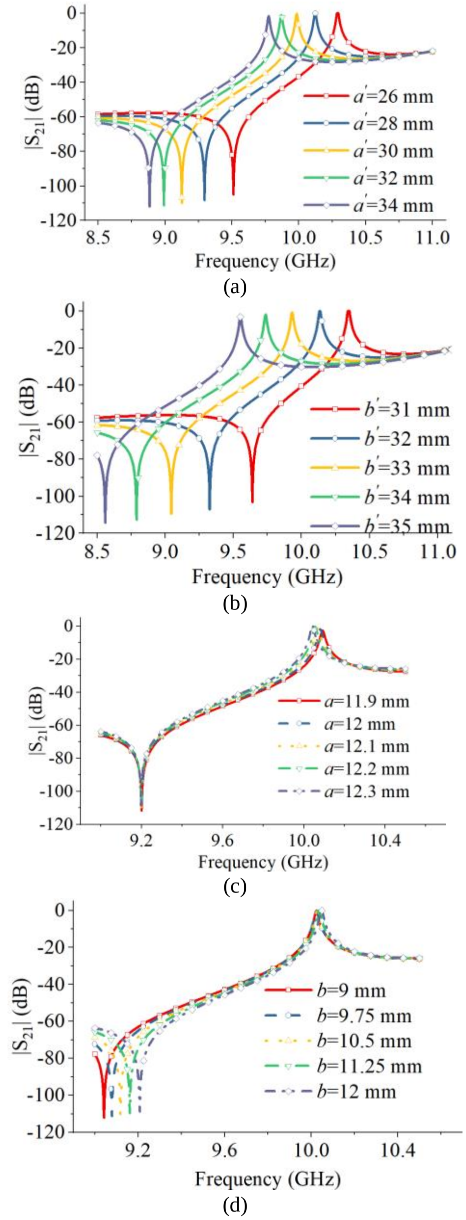


Fig. 9. Different resonant frequencies and TZs versus the changes of the sizes.

## III. Inline Stepped Window Cavity Filter With Two TZs

Based on the circuit analysis of the resonator above, a 4th-order inline stepped window cavity filter with three TZs is proposed, with its equivalent circuit and side view shown in Fig. 10. The four stepped window resonators are marked in dotted lines. Cavity A', B' and C' control the coupling between the adjacent stepped window resonators and generate TZs at the same time. Fig. 11(a) shows the 3-D view of the proposed cavity filter, which is a symmetrical structure consisting of four SWRs. Since the structure is symmetric along three-axis of the structure, the dimensions of cavities A' and B' are the same. These two cavities correspond to $B_{Z1}$ and $B_{Z3}$ in Fig. 10, so there are two TZs at the same frequency. Fig. 11 illustrates the comparison of S-parameters between simulation results and

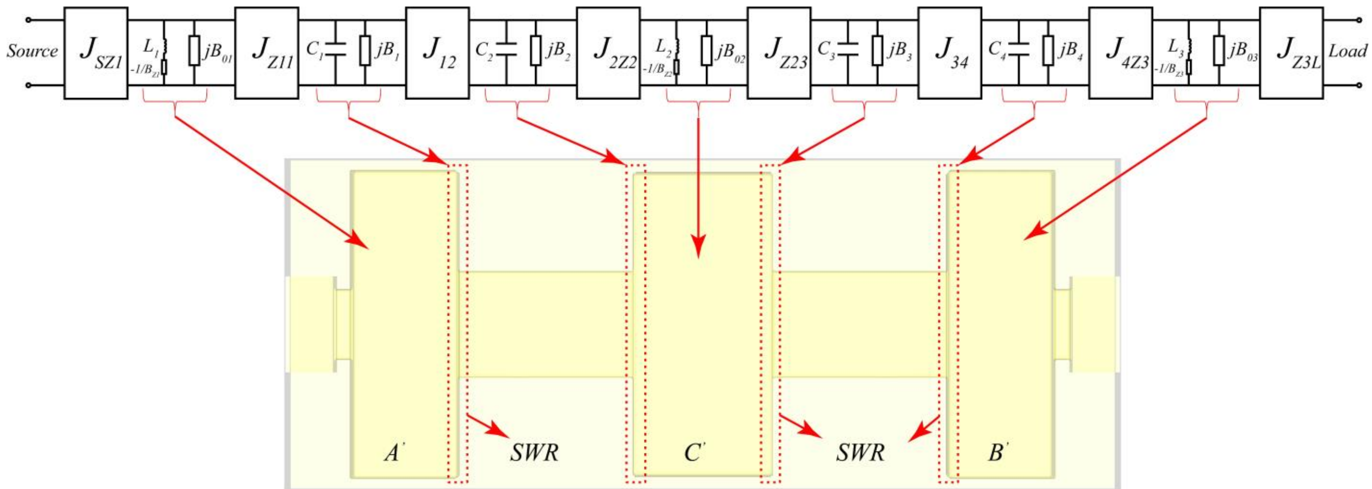



Fig. 10. Equivalent circuit model of the inline stepped window cavity filter.

synthesis response, which shows the well-matched between results. The parameters of optimized synthesis response are as follows: $J_{SZ1}$=0.3656, $J_{Z11}$=0.3986, $J_{12}$=0.9864, $J_{2Z2}$=0.8340, $J_{Z23}$=0.7659, $J_{34}$=0.9565, $J_{4Z3}$=0.6913, $J_{Z3L}$=0.6779, $B_{Z1}$=18.4, $B_{Z2}$=-1.79, $B_{Z3}$=18.4, $B_1$=-0.4483, $B_2$=0.4858, $B_3$=0.3461, $B_4$=-0.1504, $C_1$=1, $C_2$=1, $C_3$=1, $C_4$=1, $L_1$=1, $L_2$=1, $L_3$=1, $B_{01}$=0, $B_{02}$=0, $B_{03}$=0.

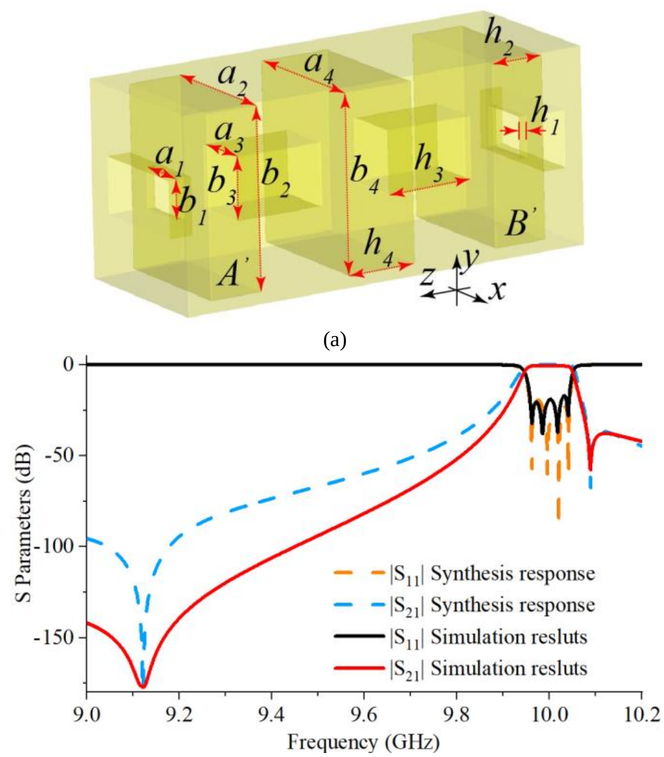
The external quality factors $Q_e$ and coupling coefficient $K$ between two resonators could be extracted through (18)-(20) [38], as follows:

$$Q_e = \frac{f_m}{\Delta f_{m\pm 90^\circ}} \quad (18),$$

$$K = \pm\frac{1}{2}\left(\frac{f_{02}}{f_{01}} + \frac{f_{01}}{f_{02}}\right)\sqrt{\left(\frac{f_{p2}^2 - f_{p1}^2}{f_{p2}^2 + f_{p1}^2}\right)^2 - \left(\frac{f_{02}^2 - f_{01}^2}{f_{02}^2 + f_{01}^2}\right)^2} \quad (19),$$

$$K = \pm\frac{f_{p2}^2 - f_{p1}^2}{f_{p2}^2 + f_{p1}^2} \quad (20),$$

where $f_m$ is the resonance frequency of the resonator, $\Delta f_{m\pm 90^\circ}$ can be extracted from the absolute phase at $f_m$ where a phase shift $\pm 90°$. When the structures of the two coupled resonators are different, equation (19) can be used to extract coupling coefficient $K$ where the $f_{01}$ and $f_{02}$ are the resonant frequencies of the two resonators, respectively. Besides, $f_{p1}$ and $f_{p2}$ are the respective resonant frequencies when two resonators are cascaded. Similarly, equation (20) can be used to extract coupling coefficient $K$ when the structures of the two resonators are identical. In order to extract the TPs more accurately, a weak input coupling is adopted to show two resonance peaks clearly. Fig. 12(a)-(b) illustrates the structures for extracting coupling coefficients $K_{12}$ and $K_{23}$, and Fig. 12(c)-(d) shows the simulation results of two structures, respectively. Since the dimensions of the two resonators are different, equation (19) is used to extract the coupling coefficient $K_{12}$. There are two TPs: $f_{p1}$ as well as $f_{p2}$, and two TZs in Fig. 12(c). Obviously, this is consistent with the previous circuit analysis. Since the structure is symmetrical, the sizes of the two resonators are the same. Therefore, the equation (20) can be used to extract the $K_{23}$. There are two TPs and one TZ. Compared with Fig. 1(b) and Fig. 2(b), it can be seen that the proposed filter in this paper is completely different from the singlet in [13].




(a)

(b)

Fig. 11. (a) 3-D view of the proposed cavity filter (b) Comparison of S-parameters between simulation results and synthesis response.

Fig. 13 depicts the value of $Q_e$ and coupling coefficient $K$ as a function of various dimensions. Fig. 13(a)-(d) shows the variation of external quality factors with the sizes of the feeding coupling structure: $a_1$, $b_1$, $h_1$, and the dimensions of the first SWR: $a_2$, $b_2$, $h_2$, $a_3$, $b_3$. The $Q_e$ decreases with the rise of $a_1$, $b_1$. In contrast, $Q_e$ increases with the rise of $a_2$, $a_3$, $b_3$, $h_1$, $h_2$. As for the $b_2$, $Q_e$ will first shift down and then increase. It is illustrated in Fig. 13(b) that the size of the first and second SWRs can have an influence on the $K_{12}$ of the proposed BPF. As $a_3$ increases, $K_{12}$ will first shift down and then rise dramatically, while when $b_3$ becomes larger, $K_{12}$ will gently rise first and then fall. Finally, as $h_3$ increases, $K_{12}$ will decrease sharply. Besides, it is shown

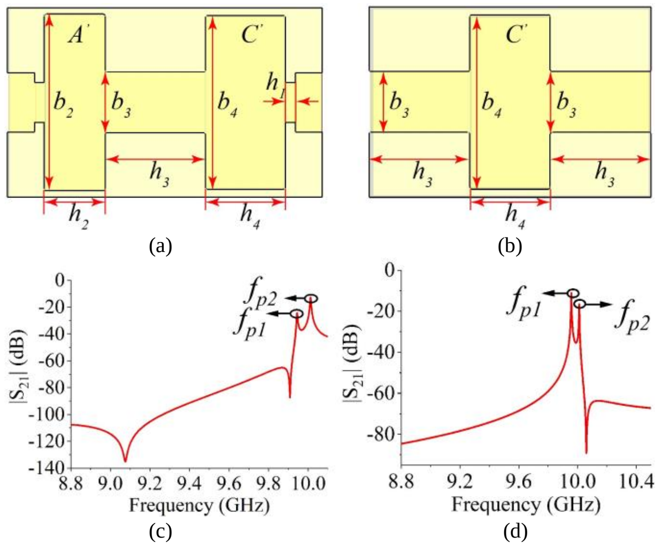



Fig. 12. (a) Side view of the extracting $K_{12}$ structure. (b) Side view of the extracting $K_{23}$ structure. (c) Simulation results for extracting $K_{12}$. (d) $|S_{21}|$ for extracting $K_{23}$.

in Fig. 13(c-d) that the $a_4$, $b_4$ and $h_4$ can have various influences on the $K_{23}$. $K_{23}$ will decrease as $a_4$ increases. Moreover, $K_{23}$ will first shift down and then increase with the rise of $b_4$. And $K_{23}$ will decrease and then increase with the rise of $h_4$.

## IV. Experimental Results

A fourth-order stepped window BPF is proposed and fabricated using brass. Fig. 14(a)-(b) shows the photographs of the external view and internal view of the fabricated filter, respectively. The physical size parameters are as follows: $a_1$ = 10.80, $b_1$ = 7.38, $h_1$ = 1.81, $a_2$ = 29.12, $b_2$ = 32.56, $h_2$ = 11.23, $a_3$ = 12.04, $b_3$ = 11.21, $h_3$ = 18.56, $a_4$ = 31.04, $b_4$ = 32.02, $h_4$ = 14.79 (all in mm). The simulated and measured results are illustrated in Fig. 15(a) with wideband frequency response. It is shown that two results match well, the measured center frequency is at 10 GHz, and the measured 3-dB fractional bandwidth of the filter is from 9.94 to 10.06 GHz (1%). The measured return loss of the filter is below 10 dB within the passband and the IL is around 0.9 dB in the passband. The IL is around 0.5 dB in simulation, the discrepancy between simulation and measurement may be caused by the roughness and gaps between metal parts which is not considered in simulation. It is noteworthy that, the proposed SWR filter compares with the conventional inline cavity resonator filter with filter same specifications as shown in Fig. 15(b). The selectivity is much better at roll-off region in higher frequency.

From 10.08 to 12.97 GHz, there is a good upper stopband rejection has more than 20 dB. Two standard coaxial-to-waveguide transitions (WR-90) are used as input/output waveguide feeding ports for measurement. Therefore, the differences between simulation results and measurement results are mainly caused by fabrication and the losses from the parasitic ohmic loss between the installation gap of each metal structure. Two TZs are produced by three shunt loading cavities, one TZ is located at 10.1 GHz, and the other TZ is below the noise floor of the Vector Network

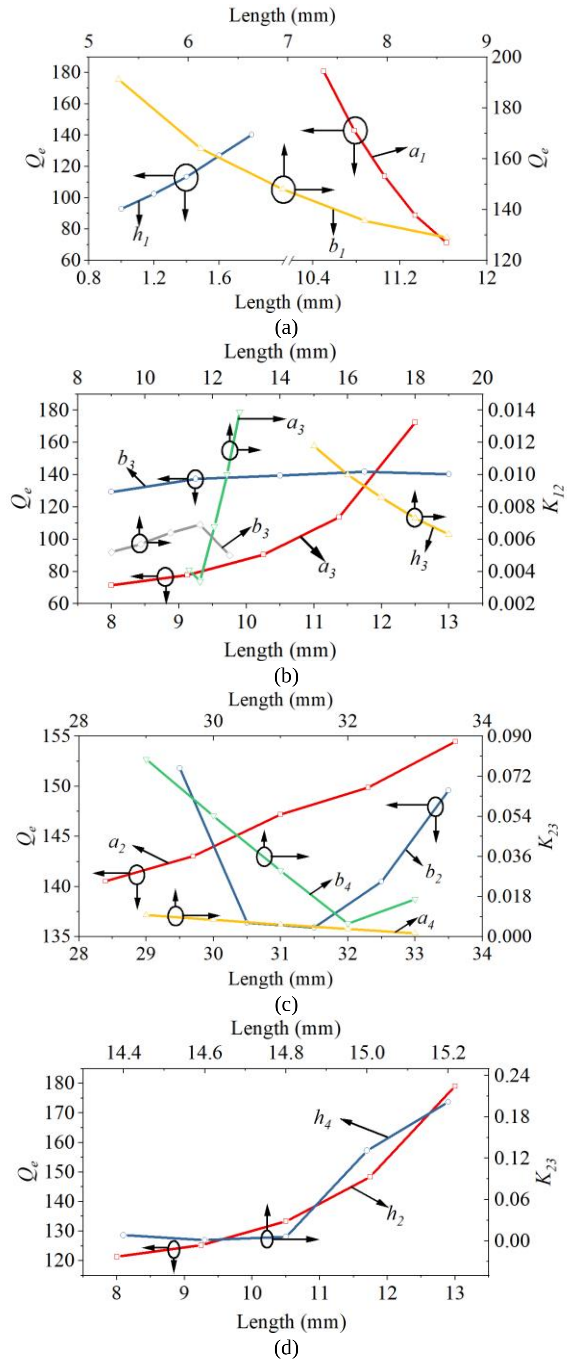


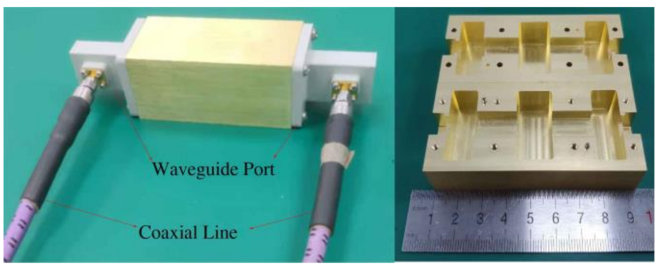

Fig. 13. (a) Relationships between $Q_e$ and value of $a_1$, $b_1$ and $h_1$. (b) Effect to $Q_e$ and $K_{12}$ against with the changes of $a_3$, $b_3$ and $h_3$. (c) Values of $Q_e$ and $K_{23}$ against various $a_2$, $b_2$, $a_4$, $b_4$. (d) Different $Q_e$ and $K_{23}$ against with the changes of $h_2$ and $h_4$.

Analyzer. It can be viewed from the narrow band frequency response in Fig. 15(b) that the selectivity of the proposed stepped window BPF is better than that of the conventional cavity resonator BPF using iris coupling structure. Moreover, Table I illustrates the comparison between this work and the state-of-the-art cavity BPFs. As can be seen from Table I, the size of the proposed filter is much smaller than the reported filter which only use rectangular waveguide resonators. Finally, the $Q_u$ and the first spurious passband also have good performance.

TABLE I Comparisons with previously reported cavity BPFs.

| Ref. | Resonator | $f_0$ (GHz) | FBW (%) | IL (dB) | $Q_u$ | TP | TZ | Manufacturing techniques | Size ($\lambda_c$ is the wavelength of the center frequency) |
|---|---|---|---|---|---|---|---|---|---|
| [14] | rectangular cavity | 19.82 GHz | 1.21 % | N/A | N/A | 6 | 2 | N/A | 2.14*0.56*7.38 $\lambda_c^3$ |
| [27] | Iris resonator | 14.25 GHz | 4.56 % | N/A | N/A | 7 | 0 | N/A | 0.75*0.38*3.01 $\lambda_c^3$ |
| [32] | Iris resonator and rectangular cavity resonator | 9.08 GHz | 9.50 % | 0.15 dB | N/A | 3 | 1 | 3-D printing | 1.64*1.00*0.61 $\lambda_c^3$ |
| [35] Filter I | Iris resonator and rectangular cavity resonator | 7.50 GHz and 10.00GHz | 13.33 % and 5.00 % | 0.17 dB | 1415 | 5 and 4 | 2 and 2 | CNC | 0.57*1.78*2.96 $\lambda_c^3$ |
| [35] Filter II | Iris resonator and rectangular cavity resonator | 7.55 GHz | 17.54 % | 0.19 dB | 1986 | 9 | 3 | CNC | 0.58*1.95*3.80 $\lambda_c^3$ |
| This work | Stepped window resonator | 10.00 GHz | 1.00 % | 0.90 dB | 3883 | 4 | 2 | CNC | 2.67*1.03*1.09 $\lambda_c^3$ |

N/A: Not available; CNC: Computer Numerical Control; FBW: Fractional Bandwidth




(a) (b)

Fig. 14. Photograph of the fabricated stepped window BPF: (a) External view, (b) Inside view.

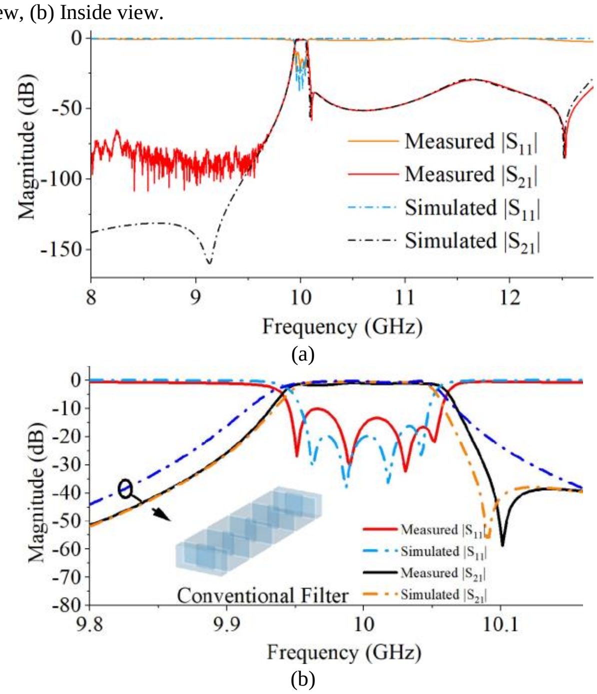



Fig. 15. Comparison of simulated and measured S-parameter frequency responses: (a) Wideband frequency response, (b) Narrow band frequency response with conventional cavity filter comparison.

## V. Conclusion

An inline stepped window BPF with two TZs is proposed in this paper. The structure of this BPF is very simple and easy to fabricate. It is noted that the volume of the proposed filter is much smaller than the cavity filter using cavity resonators. Compared with the reported iris BPFs, the $Q_u$ of the proposed filter is large, which is a very good candidate to design narrow band bandpass filter. Moreover, the equivalent circuit is created to explain the working mechanism of the proposed filter. The synthesis frequency response, simulation results and fabricated measurement match well to experimental verify the synthesis theory and the proposed SWR filter. Finally, the proposed BPF is useful in the X-band for the radar communication system application because of its easy fabrication, low cost, and simple design.

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