Compact Bandpass Filter with Sharp Out-of-band Rejection and its Application

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Abstract **─** A novel compact bandpass filter considered as the harmonic suppression circuit is designed in this paper. Because of the application of a T-shaped structure, the filter is improved in performance and reduced in size. Two transmission zeros at passband edge can be conveniently adjusted by changing the length of the open stubs located at the center of the T-shaped structure. Two filters with different open-stub structures are designed. Good agreement between the simulation and the measurement is acquired, which verifies the theoretical predictions. Benefiting from this feature, an active frequency multiplier with the proposed filter as the output matching network is designed. When input signal is set to be 6 dBm, output power of the second harmonic varies from 6 to 8 dBm with 20 dBc suppression for the first, third and fourth harmonics.

Index Terms **─** Harmonic suppression, multiplier, output power, passband, T-shaped structure, transmission zero.

I. INTRODUCTION

As one of the most important microwave component, filters with high performance and compact size are highly desirable in wideband microwave circuit. Structures of ring resonators, short/open stubs, multiple-mode resonators and so on [1-9] have been utilized to design the wideband bandpass filter in the past few years. In [1-3], traditional coupled lines are considered as the key substitute for the wideband filter of compact size and simple structure. However, due to the limit of process technique, it is difficult to realize small size of gap and line width. To tackle the problem of process, low-pass and high-pass filters are connected serially to achieve the wideband system [4-5]. Unfortunately, this kind of structure will lead to the increase of volume. In [6-7], kinds of patterns are etched on the ground plane of substrate to reach the wideband performance. Unfortunately, disadvantages of package, integration and electromagnetic leakage are inevitable. In [8-9], a novel concept of signal interaction is adopted to design the wideband filters by introducing two parallel transmission paths. In order to realize sharp-rejection bandpass filters, it is the most effective to create two transmission zeros at either side of the passband. In [10-12], quarter/half wavelength open stubs connected to the center of the resonator are proposed to realize transmission zeros located at lower or upper stopband. In summary, the performance of bandpass filter has been improved. However, the application of the proposed filter in microwave circuit is rarely involved.

In this paper, a novel structure with its series quarter-wavelength line replaced with an equivalent T-shaped structure is presented, as shown in Fig. 1. The new transmission zero can be controlled exactly by adjusting the length of the open-circuited stub. Two different wideband filters are designed and fabricated for demonstration. In addition, the filter size is reduced for use of the T-shaped structure. Two transmission zeros located at each side of the passband can be observed in simulation and measurement results. Besides, to demonstrate its advantage in engineering application, an active frequency multiplier based on the proposed filter is designed. And tunable transmission zeros produced by the filter can be employed to suppress the first, third and fourth harmonics

II. ANALYSIS OF THE WIDEBAND BANDPASS FILTER

Figure 1 (a) shows the conventional bandstop filter with two open stubs (Z_0) connected by the quarter-wavelength (Z) line. A passband can be realized between f_0 (the central frequency of the bandstop filter) and $3f_0$. But, the filter has spurious passband at $4f_0$ and several cascaded open stubs are needed to realize good impendence match. By replacing the quarter-wave length line with a T-structure, a novel bandpass filter is proposed

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in this paper, as shown in Fig. 1 (b).

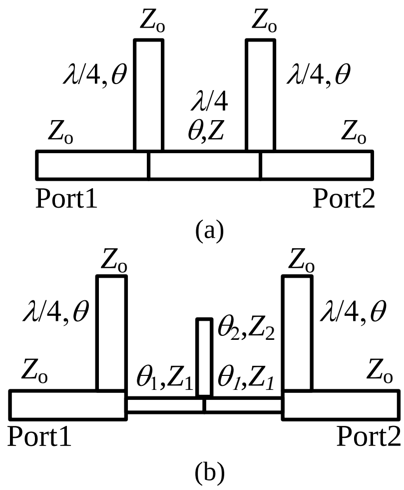


Fig. 1. (a) Conventional filter circuit and (b) novel bandpass filter circuit.

As shown in Fig. 1 (b), one open stub of Z_0, θ and another stub of Z_2, θ_2 are adopted to produce two transmission zeros which can be utilized not only to suppress unwanted harmonics, but also to adjust the bandwidth of the filter. And the filter is much more compact due to the small size of T-shaped structure over the quarter-wavelength line. In Fig. 1 (b), Z_1, Z_2, θ_1 and θ_2 represent the characteristic impedances, the electrical lengths of the series and shunt sections of the T-shaped structure respectively. ABCD matrix is used to obtain design equations and prove equivalence between T-shaped structure and a quarter-wavelength line. The ABCD matrix of a microstrip line with electrical length θ_0 is:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cos\theta_0 & jZ\sin\theta_0 \\ jY\sin\theta_0 & \cos\theta_0 \end{bmatrix}. \tag{1}$$

The ABCD matrix of the T-shaped structure can be written as $M_1 \times M_2 \times M_1$:

$$M_1 = \begin{bmatrix} \cos\theta_1/2 & jZ_1\sin\theta_1/2 \\ jY_1\sin\theta_1/2 & \cos\theta_1/2 \end{bmatrix}, \tag{2}$$

$$M_2 = \begin{bmatrix} 1 & 0 \\ jY_2\tan\theta_2 & 1 \end{bmatrix}. \tag{3}$$

In the work here, the T-shaped model is equivalent to the quarter-wavelength line (θ_0=90°at f_0,), and we thus have:

$$\begin{bmatrix} A_T & B_T \\ C_T & D_T \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 0 & jZ \\ jY & 0 \end{bmatrix}. \tag{4}$$

From (1)-(3), Z_1, Z_2 can be found based on Eqs. (5) and (6):

$$Z_1 = Z/\tan\theta_1, \tag{5}$$

$$Z_2 = \frac{Z\times\tan\theta_2}{1-\tan^2\theta_1}. \tag{6}$$

When θ_2=90°, the open stub looks like an impedance inverter, and if it is terminated in an impedance Z_U on one port, the impedance Z_L seen at the other port can be calculated by Eq. (7). As is known, Z_U here is infinite, so its impedance at another port is zero which makes the symmetrical part of the filter shorted. In this case, the transmission zero $f_{\theta 2}$ that makes θ_2=90° appears and can be acquired by Eq. (8):

$$Z_L = K^2/Z_U, \tag{7}$$

$$f_{\theta 2}/f_0 = 90^o/\theta_2. \tag{8}$$

In order to achieve a compact equivalent T-shape model, θ_1 should be less than 45°. And, the transmission zero $f_{\theta 2}$ created by the stub θ_2 can be adjusted easily, with Z_1, Z_2, f_0 and θ_1 fixed. Relation curve of $f_{\theta 2}/f_0$ and θ_2 (0° < θ_2 < 360°) is shown in Fig. 2. With the increase of θ_2, $f_{\theta 2}/f_0$ becomes less and less. When θ_2=90°, the novel created transmission zero of $f_{\theta 2}$ and f_0 are coincident.

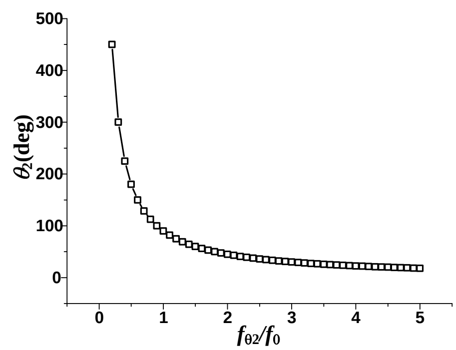


Fig. 2. Relationship between transmission zero $f_{\theta 2}$ and θ_2.

In order to validate above design ideas, a bandpass circuit with Z_0 = Z = 50Ω, f_0 = 3 GHz and θ_0 = 90° are simulated with Advanced Design System (ADS). Here we choose θ_1 = 26.5°, two different θ_2, saying, 23° and 35° are chosen to indicate the transmission zero produced by open-ended stub can be adjusted. From (4), it is easy to get Z_1 = 100Ω. From (5), the corresponding Z_2 in two cases of θ_2 are found to be 28Ω and 46.6Ω, respectively. In addition, transmission zeros for two different θ_2 can be calculated by (8). $f_{\theta 2}$ = 11.7 GHz and 7.7 GHz respectively. Figure 3 shows the simulated results for two cases with different impedance values. Comparing the simulated and the calculated results, we may see that when θ_2 < 30°, a wideband bandpass filter can be implemented between f_0 and 3f_0, while the created transmission zero $f_{\theta 2}$ can be used to suppress the fourth harmonic, thus resulting in wider upper stopband for the bandpass filter; when 30° < θ_2 < 90°, a narrowband bandpass filter between f_0 and $f_{\theta 2}$ is achieved, and the transmission zero $f_{\theta 2}$ can be used to improve the rejection performance. It is true that a transmission $f_{\theta2}$ zero can be also located below f_0, when θ_2 > 90°. However, the passband performance is dissatisfactory. In this way, two different bandpass filters with an adjustable bandwidth can be realized by controlling the location of the transmission zero $f_{\theta2}$ created by the open stub (θ_2). In addition, the filter size is much reduced by the introduced T-shaped lines.

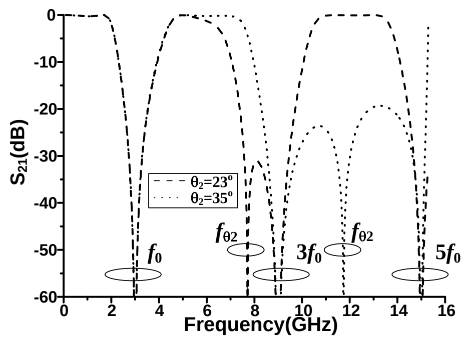


Fig. 3. Simulated results for two different $f_{\theta2}$, Z_0 = 50 ohm, f_0 = 3 GHz, θ_0 = 90°, θ_1 = 26.5°, Z_1 = 100Ω.

III. TWO WIDEBAND BANDPASS FILTERS

In this work, the quarter-wavelength line of the conventional bandstop filter is replaced with proposed T-shaped structure. For demonstrating the design strategies discussed in Section 2, two different wideband bandpass filters are designed. Here we choose θ_2 < 30° for filter A and θ_2 > 30° for filter B to realize wide passband. The two filters are all simulated with Ansoft HFSS and constructed into the Rogers4350B substrate with ε_r = 3.66 and h = 0.508 mm. Figure 4 illustrates the simulation model of bandpass filters. For filter A, Z_0 = 50Ω, f_0 = 3.2 GHz, θ_0 = 90°, Z = 50Ω, θ_1 = 27°, θ_2 = 24°, Z_1 = 80Ω, and $f_{\theta2}$ = 4f_0 = 12.8 GHz. To obtain better passband characteristics, the optimized impedance Z_2 = 80Ω. For filter B, Z_0 = 50Ω, f_0 = 3.2 GHz, θ_0 = 90°, Z = 50Ω, θ_1 = 15°, θ_2 = 40°, Z_1 = 100Ω, $f_{\theta2}$ = 7.9 GHz, and Z_2 = 120Ω.

The simulated and measurement performances of the two filters are shown in Fig. 5, good agreement can be observed between the results. For filter A, the central frequency is 6.07 GHz with two transmission zeros at f_0 and 3f_0 at either side of the passband. The transmission zero $f_{\theta2}$ created by the open stub (θ_2) is located at around 12.8 GHz to suppress the fourth harmonic 4f_0 and a wider upper stopband is thus realized by this simple and effective method. In addition, in the whole passband there are two poles with return loss below 20 dB, as can be seen in Fig. 5 (a). From Fig. 5 (b), the bandpass filter B covers the band of 4.6-7.1 GHz, the transmission zero $f_{\theta2}$ is located at around 7.9 GHz, three transmission poles are observed with return loss below 20 dB in the whole passband. The comparisons of measured results for several wideband filters [16, 17] are shown in Table 1.

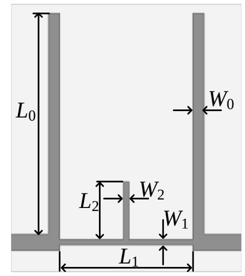
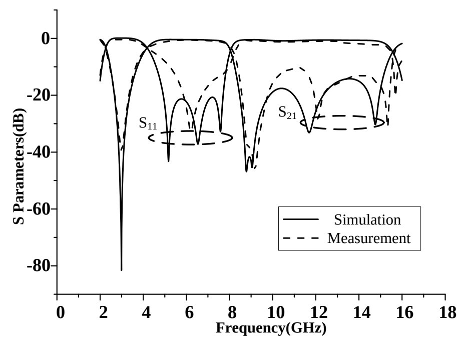
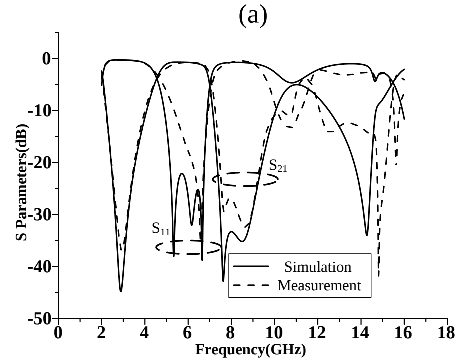


Fig. 4. The configuration of proposed filters: filter A (L_0 = 14.8 mm, L_1 = 9.0 mm, L_2 = 3.88 mm, W_0 = 0.77 mm, W_1 = 0.37 mm, W_2 = 0.36 mm), and filters B (L_0 = 14.8 mm, L_1 = 4.99 mm, L_2 = 6.08 mm, W_0 = 1.94 mm, W_1 = 0.26 mm, W_2 = 0.11 mm).



(a)



(b)

Fig. 5. (a) The simulated and measured results of filters A, and (b) The simulated and measured results of filters B.

Table 1: Comparisons with wideband filters

Different Structures	Filter A in This Paper	Ref. [16]	Ref. [17]
f_0	6.07 GHz	1.54 GHz	6.8 GHz
Circuit size (λ_o)	0.54×0.37	0.05×0.21	0.17×0.32
(f_{40dB}-f_{3dB}) / f_{3dB}	9%	23%	13%
Transmission poles	3	3	3
Second harmony suppression	20 dB	30 dB	20 dB

IV. THE APPLICATION OF THE NOVEL FILTER

As analyzed in the Section 3, the novel filter can be used to suppress the undesired harmonics of f_0, $3f_0$ and $4f_0$, but keeping the harmonic of $2f_0$. And according to the given center frequency f_0, one is able to calculate the physical size of the filter based on (4) and (5). Besides, to make the filter more compact, a T-shaped structure with an open-circuited stub is employed to replace the conventional quarter-wavelength line. Thus, due to the advantages presented above, it is extremely useful in RF circuit design [13-15]. In this paper, an active frequency doubler is designed by incorporating the proposed filter in the output matching network, as shown in Fig. 6. A quarter-wavelength line at the center frequency f_0 is adopted to bias the gate of FET, while a quarter-wavelength line at the second harmonic $2f_0$ is employed to bias the drain of FET. Properly adjusting the location of three transmission zeros at f_0, $3f_0$ and $4f_0$ as shown in Fig. 5, the output harmonics except $2f_0$ will be suppressed greatly. And, it behaves as a 50Ω line at the second harmonic $2f_0$ and also plays an important role in matching output circuit of the multiplier. So, when determining the size of the filter, harmonic suppression and port reflection coefficient should be considered simultaneously.

In fact, the designed filter can be considered as a 50Ω transmission line at the second harmonic $2f_0$ for its excellent matching that has been proved in Section 3. Thus, as shown in Fig. 6, a novel output matching network in which the proposed filter is introduced is designed. And, due to the compact feature of the filter, the size of multiplier is reduced further.

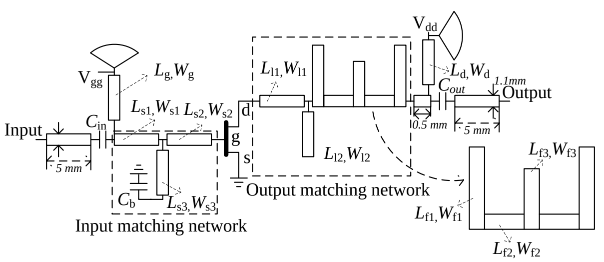


Fig. 6. The circuit of the frequency doubler ($L_g = 6.9$ mm; $W_g = 0.2$ mm; $L_d = 3.3$ mm; $W_d = 0.2$ mm; $L_{s1} = 15.6$ mm; $L_{s2} = 1.69$ mm; $L_{s3} = 9.9$ mm; $W_{s1} = 2.0$ mm; $W_{s2} = 0.2$ mm; $W_{s3} = 0.2$ mm; $L_{11} = 9.0$ mm; $L_{12} = 1.5$ mm; $W_{11} = 0.2$ mm; $W_{12} = 0.25$ mm; $L_{f1} = 9.25$ mm; $L_{f2} = 1.81$ mm; $L_{f3} = 2.85$ mm; $W_{f1} = 0.83$ mm; $W_{f2} = 0.15$ mm; $W_{f3} = 0.46$ mm).

A Hetero-Junction FET NE3210S01 whose model can be found on official website is adopted for the design of the multiplier simulated by the Agilent's Advanced Design System (ADS). The gate voltage is set to -0.25 V to reach the nonlinear field of the FET and drain voltage is fixed to be 2 V. The PCB with 14 internal ports inserted is optimized in ADS software, and then the external power, DC blocking capacitor, FET and so on is connected to be ports. Finally, a 14 ports block diagram based on S parameters of PCB is simulated to test the performance of multiplier.

For demonstrating the introduced technique and validating the simulation result, an active multiplier of 4.5~5.5 GHz implemented on Rogers 4350B substrate with ε_r =3.66 and h = 0.508 mm is designed. It can be seen in Fig. 7 that input return loss is below 20 dB for simulation and 10 dB for measurement, while the output return loss below 10 dB both for simulation and measurement in the whole band. Figure 8 tells that under 6 dBm input power, the output power of second harmonic variates from 6 to 8 dBm, and the elimination for first harmonic, third harmonic and forth harmonic is lower than 20 dBc. To indicate the relationship between the input power and output power of second harmonic more clearly, Fig. 9 is introduced. Besides, the total length of output matching network is 15 mm, while the circuit without novel filter reaches 21 mm. And, nearly no suppression for the harmonic is produced. Finally, picture of the fabrication circuit is shown in Fig. 10.

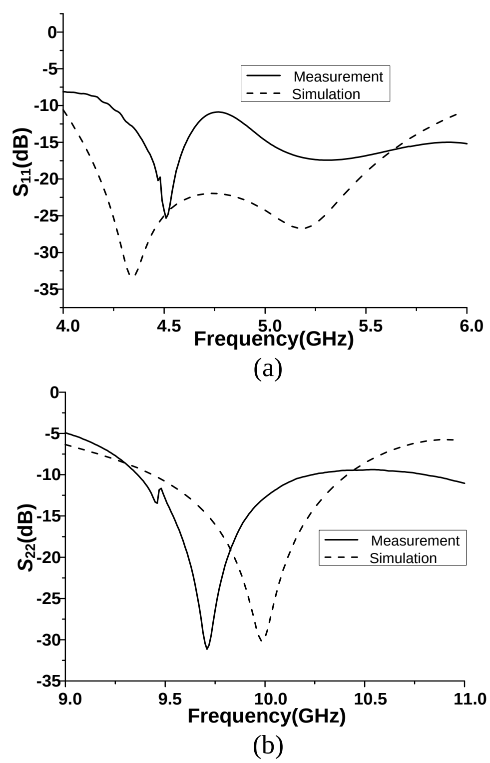
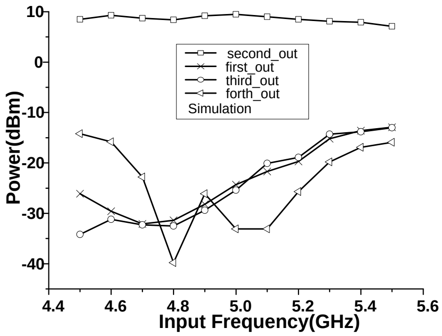


Fig. 7. (a) The simulated and measured results of S_{11}, and (b) the simulated and measured results of S_{22}.



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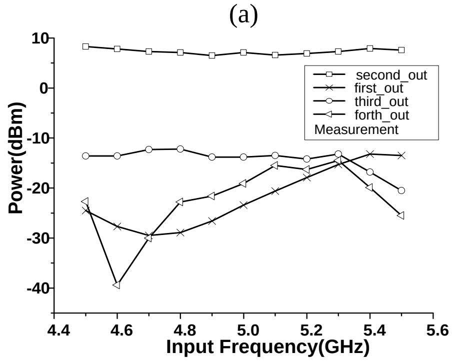
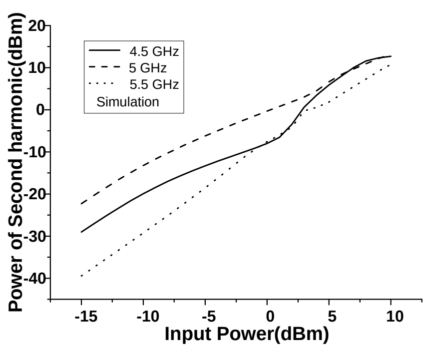
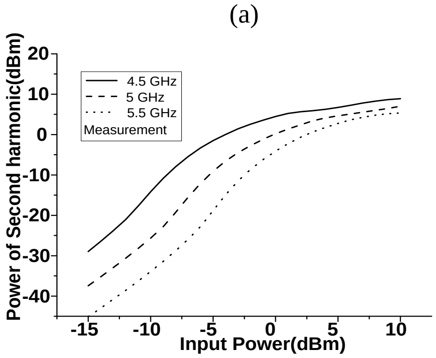


(b)

Fig. 8. (a) The output power of different harmonics for simulation, and (b) the output power of different harmonics for measurement.



(a)



(b)

Fig. 9. (a) The output power of second harmonic for simulation, and (b) the power of output second harmonic for measurement.

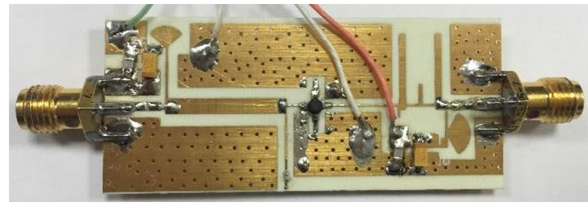


Fig. 10. The photograph of the designed multiplier.

V. CONCLUSION

In this work, two kinds of novel bandpass filters are proposed and demonstrated. What's more, the designed filter is applied to form an active multiplier. The highlight of the design is to replace the quarter-wavelength line of bandstop filter with a T-shaped open stub structure. One advantage is the size reduction of the structure; another is the flexible adjustability of transmission zero created by the open stub and thus easy bandwidth control of the passband. In addition, the transmission zero created by the central open stub of the T-shaped lines can be used to suppress the harmonic or improve the roll-off skirt selectivity; high out-of-band rejection can thus be realized. Two novel compact wideband bandpass filters are modeled and simulated for demonstration, good agreement can be observed between the simulation and theoretical analysis, indicating the validity of the proposed design strategies. With those good performances, a kind of active frequency doubler is designed based on the novel filter. In on hand, the size of output matching network is reduced; on the other hand, the output undesired harmonics is suppressed greatly.

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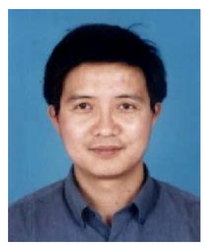

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