

High Selectivity Wideband Bandpass Filter Based on Transversal Signal-Interaction Concepts Loaded with Open and Shorted Stubs

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Abstract—A new design of high selectivity wideband bandpass filter based on transversal signal-interaction concepts loaded with open and shorted stubs is proposed in this paper. Two transmission paths are used to realize signal transmission. Path 1 is composed of a T-shaped structure with shorted stub, and Path 2 consists of two open coupled lines loaded with open stubs. A wide five-order passband and high selectivity stopband with four transmission zeros can be achieved in the proposed filter. Finally, a wideband bandpass filter operating at 3 GHz with 3-dB fractional bandwidth of 83.3% (1.55 to 4.05 GHz) is designed, fabricated, and measured. Good agreement between the simulation and experiment is obtained.

1. INTRODUCTION

In recent years, some filter structures based on transversal signal-interaction concepts have been proposed [1–6]. Two ultra-wideband bandpass filters using shorted coupled lines based on the ideal 180° phase inversion were proposed in [1, 2]. But one transmission zero in the upper stopband made the 3-dB bandwidth of the filters not easy to adjust. In [3, 4], dual-bandstop and dual-bandpass filters based on transversal signal-interaction concepts were realized. And high selectivity wideband bandpass filters with multiple transmission zeros were achieved due to the superposition of signals of the two transmission paths in [5, 6]. In addition, coupled lines were vital elements to create bandpass filters. Some mentioned researches were realized by using coupled lines [1, 2] and some by using coupled lines and T-shaped structures [3, 5, 6]. The wideband bandpass filters in [7–9] and dual-bandpass filter in [10] were also realized by using coupled lines. Besides, other design methods of filter were also investigated in [11, 12].

In this paper, a high selectivity wideband bandpass filter based on transversal signal-interaction concepts loaded with open and shorted stubs is proposed. A T-shaped structure consisting of two open coupled lines and a shorted stub is utilized to realize a wide passband, and the open coupled lines loaded with open stubs are used to improve passband return loss and increase stopband attenuation. Wide five-order passband and high selectivity upper stopband can be realized. The two transmission zeros close to passband can be adjusted by changing the characteristic impedances of the open stubs. Detailed theoretical design, simulation and measurement results for the wideband bandpass filter are demonstrated and discussed.

2. ANALYSIS AND DESIGN OF THE PROPOSED WIDEBAND BANDPASS FILTER

Figure 1(a) shows the T-shaped structure bandpass filter consisting of two open coupled lines (even/odd-mode characteristic impedance Z_{e1} and Z_{o1} , electrical length θ) and a shorted stub (characteristic

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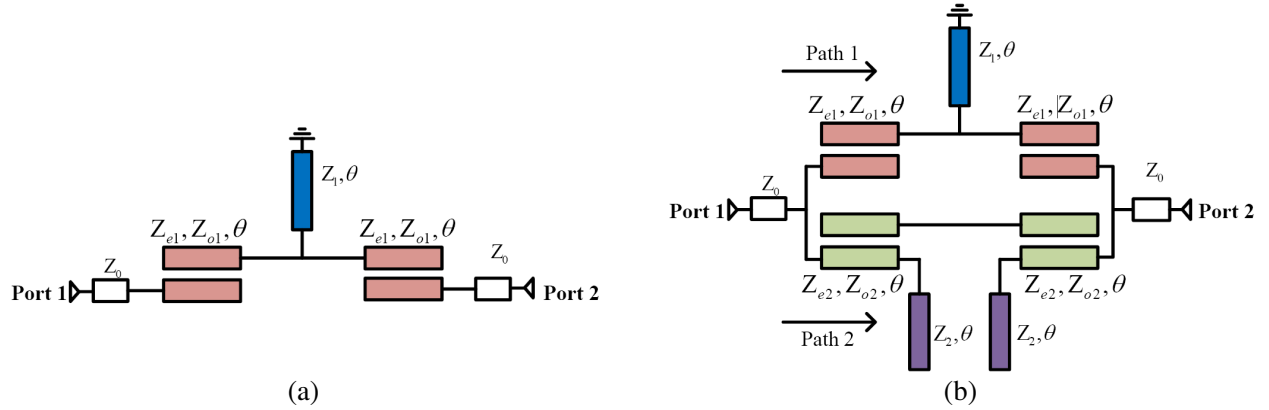


Figure 1. Filter circuits. (a) Wideband bandpass filter in [7]. (b) The proposed wideband bandpass filter.

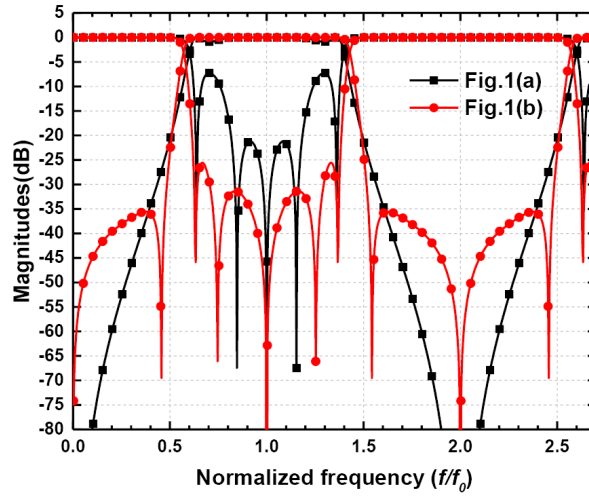


Figure 2. Simulated frequency responses of Fig. 1.

impedance Z_1 , electrical length θ , $\theta = 90^\circ$ at the center frequency f_0) in [7]. The proposed wideband bandpass filter in this paper consisting of two transmission paths based on transversal signal-interaction concepts is shown in Fig. 1(b). The circuit in Fig. 1(a) is used as transmission Path 1. For Path 2, two open coupled lines (even/odd-mode characteristic impedance Z_{e2} and Z_{o2} , electrical length θ) loaded with open stubs (characteristic impedance Z_2 , electrical length θ) are connected. Two microstrip lines with characteristic impedance $Z_0 = 50 \Omega$ are connected to Port 1 and Port 2. The simulated results of Fig. 1(a) and Fig. 1(b) are shown in Fig. 2.

Due to the superposition of signals for Paths 1 and 2, the in-band reflection coefficient is improved, and four out-of-band transmission zeros can be achieved. The passband order will not be changed by adding transmission Path 2; however, the in-band $|S_{11}|$ is greater than 25 dB. In addition, the two transmission zeros close to the passband, which enhances the proposed filter performance, are created by transmission Path 2.

The theoretical analysis of the circuit structure of the proposed filter is given. The $ABCD$ matrices of open coupled lines, shorted stub and open coupled lines loaded with open stub are shown below. The $ABCD$ matrices M_c (open coupled lines) and M_s (shorted stub) can be obtained from [13]. M_{cs1} and

M_{cs2} (open coupled lines loaded with open stub) can be derived from [14].

$$M_c = \begin{bmatrix} \frac{Z_{e1} + Z_{o1}}{Z_{e1} - Z_{o1}} \cos \theta & j \frac{(Z_{e1} - Z_{o1})^2 - (Z_{e1} + Z_{o1})^2 \cos^2 \theta}{2(Z_{e1} - Z_{o1}) \sin \theta} \\ j \frac{2 \sin \theta}{Z_{e1} - Z_{o1}} & \frac{Z_{e1} + Z_{o1}}{Z_{e1} - Z_{o1}} \cos \theta \end{bmatrix}, \quad (1)$$

$$M_s = \begin{bmatrix} 1 & 0 \\ -j \cot \theta / Z_1 & 1 \end{bmatrix}, \quad (2)$$

$$M_{cs1} = \begin{bmatrix} a & b \\ c & d \end{bmatrix}, \quad (3)$$

$$M_{cs2} = \begin{bmatrix} d & b \\ c & a \end{bmatrix}, \quad (4)$$

Where

$$a = \frac{(Z_{e2} + Z_{o2})(2Z_2 \cos^2 \theta - (Z_{e2} + Z_{o2}) + (Z_{e2} + Z_{o2}) \cos^2 \theta)}{(Z_{e2} - Z_{o2})2Z_2 \cos \theta}, \quad (5)$$

$$b = j \frac{Z_{e2} Z_{o2} (Z_{e2} + Z_{o2} + Z_2) + Z_2 (Z_{e2}^2 + Z_{o2}^2) / 2}{2Z_2 (Z_{e2} - Z_{o2})} \sin \theta - j \frac{2Z_{e2} Z_{o2}}{(Z_{e2} - Z_{o2}) \sin \theta}, \quad (6)$$

$$c = j \frac{(2Z_2 + Z_{e2} + Z_{o2})}{Z_2 (Z_{e2} - Z_{o2})} \sin \theta, \quad (7)$$

$$d = \frac{Z_2 Z_{e2} + Z_2 Z_{o2} + 2Z_{e2} Z_{o2}}{Z_2 (Z_{e2} - Z_{o2})} \cos \theta. \quad (8)$$

The $ABCD$ matrices of the Path 1 circuit can be defined as $M_{c1} \times M_s \times M_{c1}$. And the $ABCD$ matrices of the Path 2 circuit can be defined as $M_{cs1} \times M_{cs2}$. After $ABCD$ - and Y -parameter conversions, the Y matrices (Y_1 -Path 1 and Y_2 -Path 2) can be obtained. Then the Y matrices of the proposed filter circuit structure can be illustrated as

$$Y = Y_1 + Y_2. \quad (9)$$

According to Y matrices, the S -parameters of the proposed BPF can be expressed as [15]

$$S_{11} = \frac{1 - Y_{11}^2 + Y_{21}^2}{(1 + Y_{11})^2 - Y_{21}^2}, \quad (10)$$

$$S_{21} = \frac{-2Y_{21}}{(1 + Y_{11})^2 - Y_{21}^2}. \quad (11)$$

When S_{21} is equal to zero, transmission zeroes can be obtained. After theoretical analysis according to [16], the circuit structure of the proposed filter is a whole stopband structure when the electrical length θ is equal to 0° ($f = 0$) and 180° ($f = 2f_0$). So there are two inherent transmission zeros located at $f = 0$ and $f = 2f_0$. The other two transmission zeros are not obtained by equation because the equation is too complex and cumbersome. Then the influence of characteristic impedances of the transmission lines is shown in Fig. 3. It can be seen that the coupling coefficient $K = (Z_e - Z_o) / (Z_e + Z_o)$ of the coupled lines and the characteristic impedances of the shorted stub have little influence on the position of the other two transmission zeros. However, the location of the other two transmission zeros can be adjusted by changing the characteristic impedances of the open stubs. In addition, when S_{11} is equal to zero, transmission poles can be obtained. The solutions of $S_{11} = 0$ are not obtained because the equation is too complex. But seen from Fig. 3, the solutions have relationship with K_1 , Z_1 , K_2 and Z_2 . When K_1 , Z_1 , K_2 and Z_2 are properly selected, five passband transmission poles reflect the fact that $S_{11} = 0$ has five real solutions, and three passband transmission poles reflect the fact that $S_{11} = 0$ has three real solutions. Then five passband transmission poles are selected in this paper.

Figure 3 shows the simulated results of the circuit structure in Fig. 1(b). The original parameters of the circuit structure are defined as follows: $Z_{e1} = 170 \Omega$, $Z_{o1} = 75 \Omega$ ($K_1 = 0.388$), $Z_{e2} = 99 \Omega$,

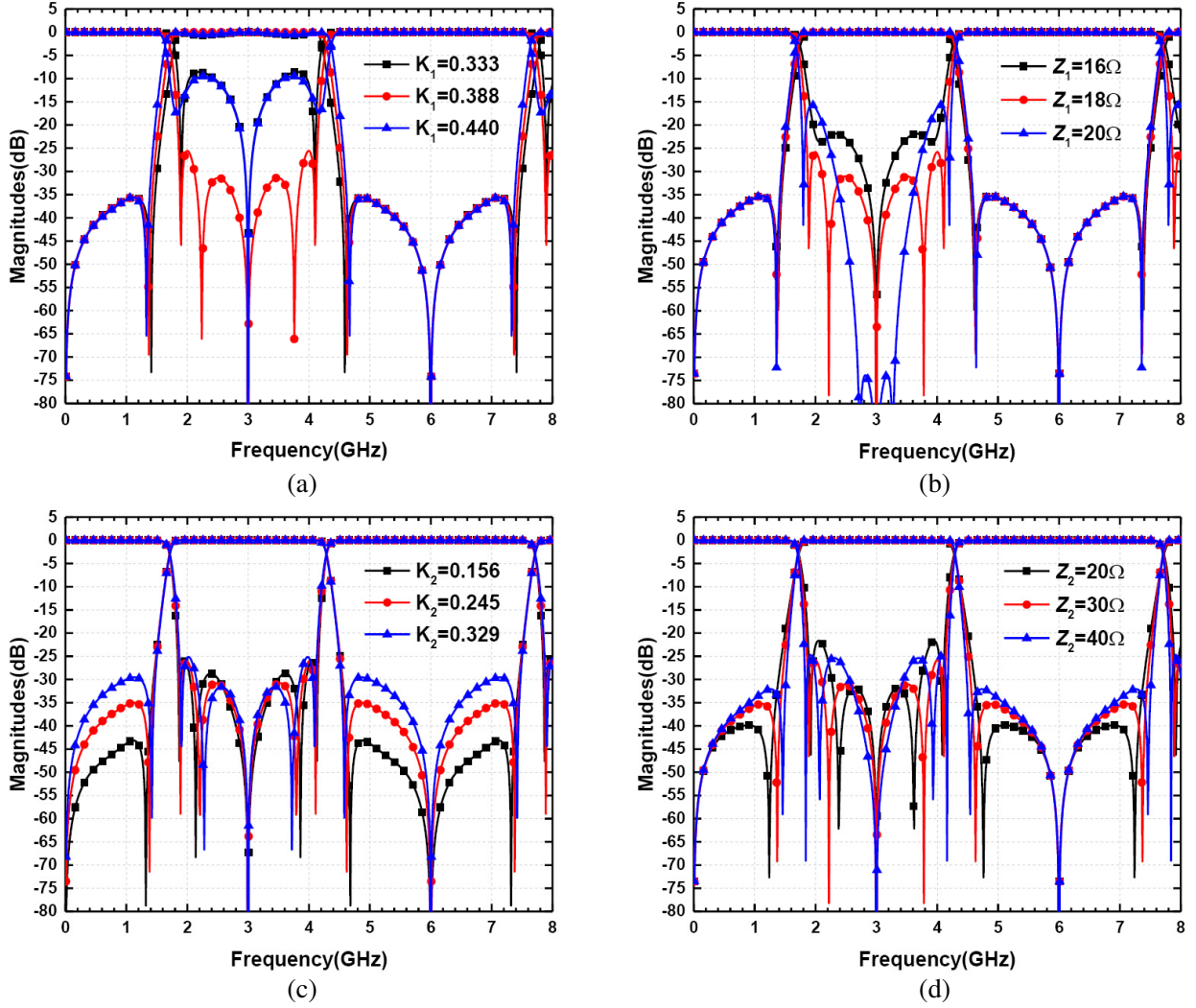


Figure 3. The influence of characteristic impedances of transmission lines. (a) $K_1 = \frac{Z_{e1} - Z_{o1}}{Z_{e1} + Z_{o1}}$. (b) Z_1 . (c) $K_2 = \frac{Z_{e2} - Z_{o2}}{Z_{e2} + Z_{o2}}$. (d) Z_2 .

$Z_{o2} = 60 \Omega$ ($K_2 = 0.245$), $Z_1 = 18 \Omega$ and $Z_2 = 30 \Omega$. Then one of the characteristic impedances is varied while others remain constant. As we can see from Figs. 3(a) and 3(b), the coupling coefficient K_1 of the open coupled lines in Path 1 and the characteristic impedance Z_1 of the shorted stub have great impact on the passband performance of the proposed filter. Fig. 3(c) shows that the coupling coefficient K_2 of the open coupled lines in Path 2 increases while the harmonic suppression decreases. Seen from Fig. 3(d), besides adjusting the position of the two transmission zeros close to the passband, the variation of the characteristic impedance Z_2 of the open stubs has influence on both passband and stopband performances.

3. HIGH-SELECTIVITY WIDEBAND BANDPASS FILTER IMPLEMENTATION

Based on the above theory analysis, the circuit parameters of wideband bandpass filter in Fig. 1(b) are $Z_0 = 50 \Omega$, $Z_1 = 18 \Omega$, $Z_2 = 30.7 \Omega$, $Z_{e1} = 170 \Omega$, $Z_{o1} = 75 \Omega$, $Z_{e2} = 99.8 \Omega$, $Z_{o2} = 61.4 \Omega$ and $f_0 = 3 \text{ GHz}$. A prototype fabricated on a Rogers4350B substrate with $\epsilon_r = 3.66$ and $h = 0.762 \text{ mm}$ is shown in Fig. 4. The dimensions are selected as follows (all in millimeters): $w_0 = 1.72$, $w_1 = 7.25$, $l_1 = 14.4$, $w_2 = 0.26$, $s_2 = 0.2$, $l_2 = 16$, $w_3 = 0.76$, $s_3 = 1$, $l_3 = 12.89$, $w_4 = 4.2$, $l_4 = 17.52$, $s = 1.14$ and $d = 0.5$.

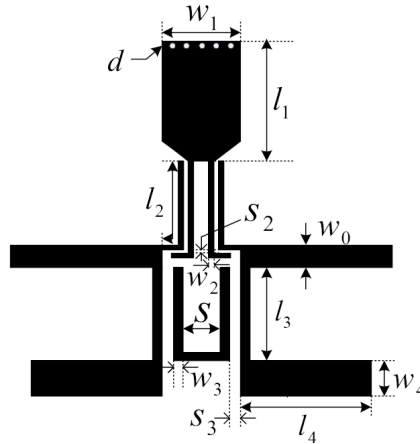


Figure 4. Layout of proposed wideband bandpass filter.

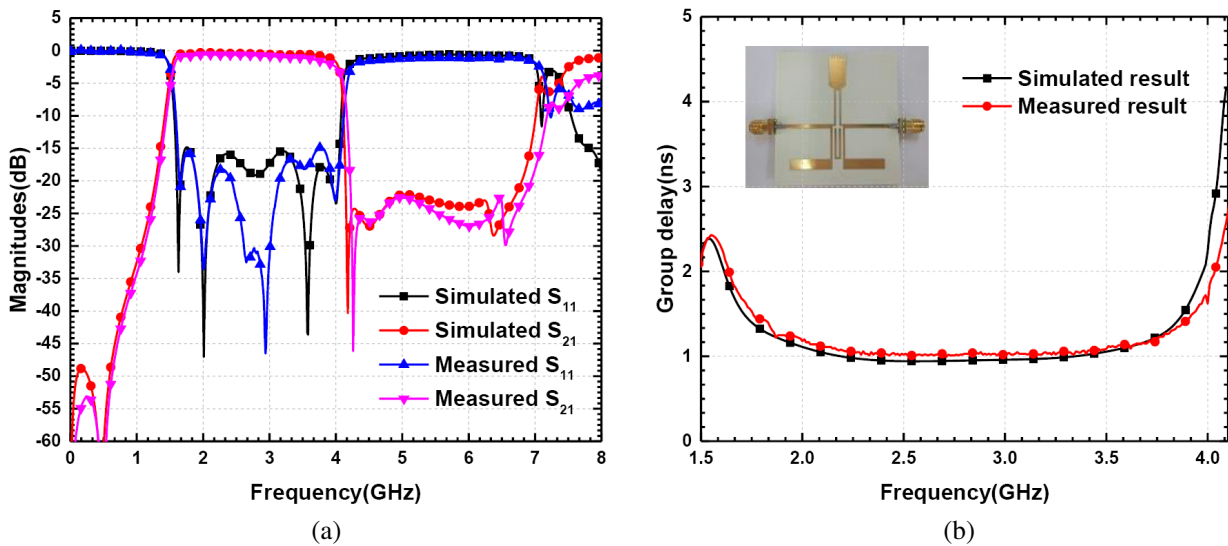


Figure 5. The simulated and measured results of the fabricated bandpass filter. (a) S -parameter. (b) Group delay.

The simulated and measured results of this fabricated filter are presented in Fig. 5. A good agreement between the simulated and measured results can be observed. Due to manufacturing errors, there are six transmission poles of the measured results in the passband in Fig. 5(a). As shown in Fig. 1(b), there are two pairs of coupled transmission lines (Z_{e1} and Z_{o1}) in Path 1 and two pairs of coupled transmission lines (Z_{e1} and Z_{o1}) in Path 2. The odd- and even-mode phases of the coupled lines are not equal. The phase difference of the odd- and even-modes of the coupled lines causes some frequency shift. So the transmission zero located at the low passband is not close to passband, and the transmission zero located at $2f_0$ is not at 6 GHz in Fig. 5(a). The measured 3-dB fractional bandwidth of this filter is almost 83.3% (from 1.55 to 4.05 GHz). Within the passband, the measured insertion loss for the filter is less than 3 dB while the return loss is over 15 dB. Furthermore, the stopband attenuation is greater than 20 dB from 4.2 GHz to 6.95 GHz and 15 dB from 4.2 GHz to 7.09 GHz. Seen from Fig. 5(b), the measured group delay varies from 1.01 to 2.43 ns in the whole passband with a maximum variation of 1.42 ns. Measurement and fabrication errors cause the slight frequency shift between the measured and simulated results. Finally, to further demonstrate the performance of the proposed filter, Table 1 compares some reported wideband bandpass filters with this work.

Table 1. Comparison with previous bandpass filters.

Filter structure	3-dB FBW (%)	Transmission zeros ($0-2f_0$)	Return loss passband (dB)	Insertion loss stopband (dB)	Group delay variation (ns)	Size ($\lambda_g \times \lambda_g$)
[1]	124.6%	2	> 15	> 15($2.5f_0$)	< 0.3	0.4×0.35
[5]	43.3%	6	> 11	> 20($2.67f_0$)	< 1	0.70×0.40
[6]	61.7%	6	> 20	> 15($2.7f_0$)	< 0.7	0.51×0.34
	48.2%	9	> 10.5	> 15($2.8f_0$)	< 0.8	0.51×0.40
[8]	88%	2	> 12.2	> 15($2.53f_0$)*	< 0.62	0.7×0.38
This work	83.3%	4	> 15	>20(2.36f₀)	<1.42	0.56 × 0.52

*: estimated value

4. CONCLUSIONS

A wideband bandpass filter based on transversal signal-interaction concepts is proposed. Path 1 utilizes a T-shaped structure consisting of two open coupled lines and a shorted stub to realize a wide passband, and Path 2 employs open coupled lines loaded with open stubs to achieve transversal signal-interaction. Due to the superposition of signals for the two paths, the in-band reflection coefficient is reduced to -15 dB, and the out-of-band harmonics are well suppressed to -20 dB. Compared with the previous works, the proposed wideband bandpass filter in this paper has the following advantages: 1) miniaturization; 2) high stopband rejection and frequency selectivity; 3) good return loss.

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