Microstrip Bandpass Filter with a High Selective Performance Using Triple-mode Stub-loaded Resonator

Yunxiu Wang, Wenhui Zou*

School of Electronic and Information Engineering
China West Normal University, Nanchong, Sichuan, China.
* Corresponding author: zouwench@163.com

***Abstract* — A high-selective microstrip bandpass filter (BPF) using triple-mode stub-loaded resonator is proposed in this letter. Firstly, a triple-mode stub-loaded resonator is formed by tap-connecting a λ/2 resonator with two open-ended stubs at its center. Thus, three resonant modes are excited to make up a three-pole passband, and meanwhile two transmission zeros at both side of the passband are produced as well. Secondly, the electrical paths of the two open-ended stubs are investigated so as to adjust the two transmission zeros to the desired frequency locations and herein the desired bandwidth can be obtained. Finally, a prototype filter is designed and fabricated, and its measured results are provided to verify the predicted frequency response.**

Keyword – Bandpass filter; roll-off skirt frequency response; stub-loaded resonator; triple-mode

I. INTRODUCTION

Bandpass filter (BPF) is one of essential RF and microwave components for various wireless systems. Planar wideband BPFs have been recently arousing much attention because they can be fabricated using printed circuit technology and easy integration with other active and passive circuit blocks. The conventional parallel coupled BPF based on λ/2 resonators exhibits some advantages such as simple synthesis procedure, good repetition, and wide range of realizable bandwidth, etc. But it suffers from poor out-of-band selectivity. This problem can be resolved by out-of-phase cross coupling [1]. Another simple stub-tapped scheme was recently proposed to design BPFs [2]-[6]. A two-pole BPF using a single stub-loaded λ/2 resonator is initially reported in [2], where the tapped open stub acts as an equivalent K-inverter between two λ/4 resonators. Moreover, the stub with a length longer or shorter than λ/4 can produce a transmission zero at a frequency lower or higher than the desired passband, and thus by coupling these two resonators and introducing parallel coupling structure at I/O ports, a filter with sharpened roll-off skirts in lower and higher cut-off frequencies can be obtained. To minimize the filter size, a pair of unsymmetrical open-ended stubs with unequal stub lengths and widths, tap-connected to the central point of a λ/2 line resonator, is proposed in [3]. However, due to the harmful impact of a cross-junction in this resonator, the first transmission pole is unexpectedly shifted to the lower frequency, herein causing poor in-band performance. In [4], a hairpin resonator with a quasi-900 open-ended stub and a very short short-ended stub loaded at the midpoint of the hairpin resonator is applied to design an alternative planar BPF, in which the resonator can produce three transmission poles but it can only generate one transmission zero at the higher side of the passband as the cross-coupling is inductive. Similarly, with the use of open-ended and short-ended stub-loaded multi-mode resonator, wideband BPFs with a transmission zero at the higher frequency of the passband were built up in [5] and [6].

In this letter, a three-pole BPF with a high selective performance using the proposed stub-loaded resonator in Figure 1 is presented. The proposed resonator is firstly theoretical analyzed, and then applied for BPF design. In final, a prototype filter is designed and fabricated to provide experimental verification on the predicted frequency responses.

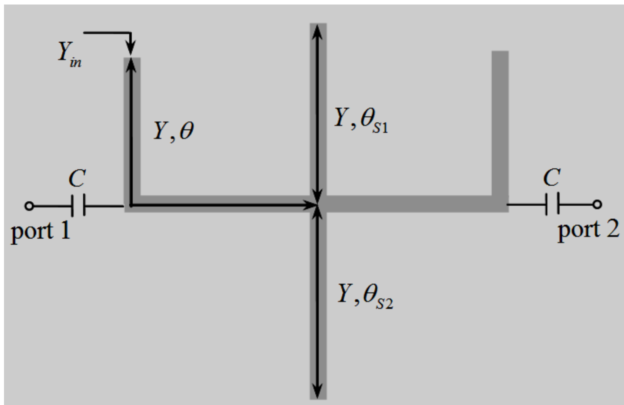


Fig.1 Proposed two unsymmetrical open-ended stubs loaded resonator.

II. ANALYSIS ON THE PROPOSED BPF

As depicted in Figure 1, the proposed stub-loaded resonator is formed up by tap-connecting two different length open-ended stubs to the center of a λ/2 uniform impedance resonator (UIR), and the electrical lengths of the UIR and these two stubs are 2θ, θs1 and θs2, respectively. Then the input admittance of the resonator can be expressed as

$$Y_{in} = jY\frac{\tan\theta_{s1} + \tan\theta_{s2} + 2\tan\theta}{1-\tan\theta(\tan\theta_{s1} + \tan\theta_{s2} + \tan\theta)} \quad (1)$$

where Y is the characteristic admittance of the UIR and the stubs. It is known that the resonance condition of the resonator can be obtained from Yin=0.

Now, let's investigate the resonant-mode characteristics of the proposed resonator in Figure 1. Firstly, for a special case of θ=900, we can get

$$Y_{in} = \lim_{\theta\to 90^0} jY\frac{(\tan\theta_{s1} + \tan\theta_{s2})/\tan\theta + 2}{1/\tan\theta - (\tan\theta_{s1} + \tan\theta_{s2} + \tan\theta)} = 0 \quad (2)$$

Actually, this special frequency is corresponded to the second resonant frequency (f2), and the principle has been discussed in [7]. The other resonant frequencies can be attained from

$$Y_{in} = 0 \quad (3)$$

In this design, we set the first three resonant frequencies of the resonator as f1, f2, and f3 ($f_1 < f_2 < f_3$). Herein, the first and the third resonant frequencies can be deduced as

$$\begin{cases}\tan(k_1\cdot\theta_{s1}) + \tan(k_1\cdot\theta_{s2}) + 2\tan(k_1\cdot 90^0) = 0\\ \tan(k_2\cdot\theta_{s1}) + \tan(k_2\cdot\theta_{s2}) + 2\tan(k_2\cdot 90^0) = 0\end{cases} \quad (4)$$

where k_1 is the ratio of the first resonant frequency to the second resonant frequencies (f_1/f_2), while k_2 denotes the ratio of the third resonant frequency to the second resonant frequency (f_3/f_2). Keeping θ=90^0, Figure 2 plots the variation of k_1 and k_2 when θ_{s2} is taken a series of fixed values (10^0-120^0) while θ_{s1} varies from 10^0-120^0.

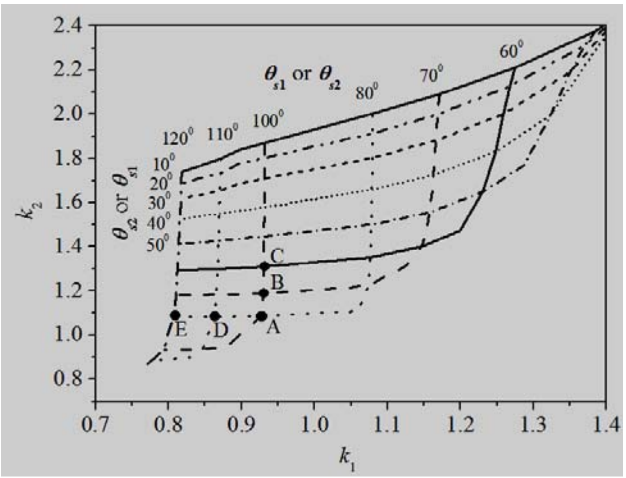
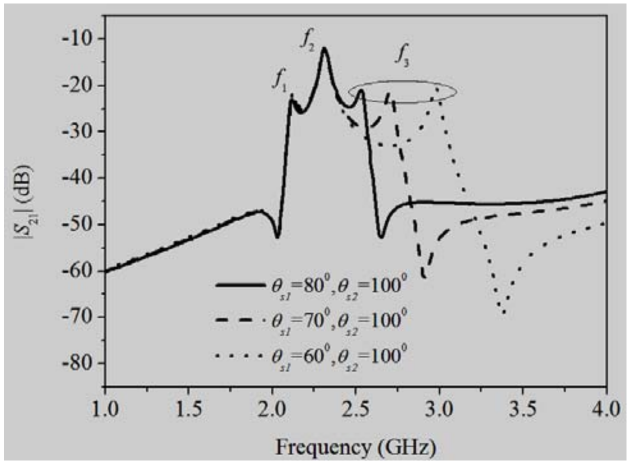
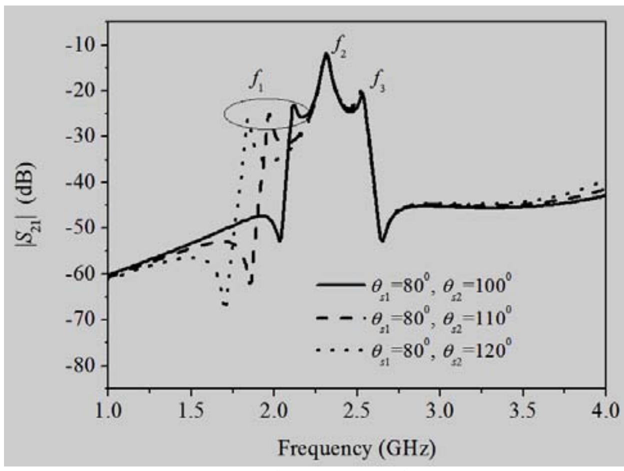


Fig.2 The values of θ s1 and θ s2 under the given frequency ratios k1 and k2.

From the graph, we can get the lengths of the two loaded stubs once the frequency ratios k_1 and k_2 are given. It can be observed that k_1 and k_2 can be controlled independently when θ_{s1} and θ_{s2} are taken some special values. For example, k_2 is decreased whereas k_1 remains approximately unchanged when θ_{s2} is taken as 100^0 while θ_{s1} varying from 10^0 to 80^0. It means in this case that the third resonant frequency can be adjusted only by changing the length of θ_{s1}. Whereas when we keep θ_{s1}=80^0 to be unchanged and increase θ_{s2} from 100^0 to 110^0 and 120^0, here k_2 almost remains unchanged while k_1 is pushed down, which means the first resonant mode can be flexibly controlled by the θ_{s2} under this situation. To further clearly show the characteristic, *Agilent Momentum* software is used to investigate the frequency response. All the filter structures discussed as below are formed on the RT/Duroid 6010 substrate with a thickness of 1.27 mm and a dielectric constant of 10.8. In our design, firstly the second resonant frequency is set to be at f_2=2.3 GHz and θ is set to be 90^0. Meanwhile, characteristic admittance Z=1/Y is chosen as 90 Ω in order to minimize the unwanted discontinuity effect around several T-junctions involved in the proposed resonator shown in Figure 1. Figure 3 illustrates the simulated |S_{21}| of the proposed resonator with different lengths θ_{s1} and θ_{s2} under loose-coupled with I/O ports by C=0.075 pF.



(a) Different θ s1 with fixed θ s2=1000.



(b) Different θ s2 with fixed θ s1=800.

Fig.3 Simulated |S21| of the proposed resonator with different lengths θ s1 and θ s2.

As previously described, the third mode location, shown in Figure 3(a), is basically pushed up as θ_{s1} decreases from 80^0 to 70^0 and 60^0 when keeping θ_{s2}=100^0 to be unchanged, which correspond A, B, and C in Figure 2. It can also be observed from the Figure 3 that there appears two transmission zeros, which are caused by virtual grounding of the two loaded open-ended stubs. The first transmission zero remains unmoved as θ_{s1} decreases from 80^0 to 70^0 and 60^0 as described in Figure 3(a). Similarly, as depicted in Figure 3(b), the first mode now tends to be shifted to the lower frequencies as θ_{s2} increases from 100^0 to 110^0 and 120^0 under the fixed $\theta_{s1}=80^0$, and obviously the values of k_1 and k_2 here are corresponded to A, D, and E points in Figure 2. Meanwhile, the results of Figure 3(b) herein show us that the first transmission zero is now moved to the lower frequencies as θ_{s2} increases from 100^0 to 110^0 and 120^0 while the location of the second transmission zero doesn't change.

III.FILTER DESIGN AND RESULTS

To verify the above design method, a BPF is designed with the center frequency at 2.3 GHz and the 3-dB fractional bandwidth of 16.5 % by using the proposed resonator, herein we can get θ=900. Following the method described in Section II, k1 and k2 are chosen to be 0.92 and 1.07 respectively in order to guarantee the first three resonant frequencies of the resonator are properly excited and placed within the desired passband. Then θs1 and θs2 can be determined to be 800 and 1000 according to Figure 2. So, the dimension parameters of the proposed BPF shown in Figure 4(a) are obtained and optimized as follows: L1=4.0 mm, W=0.2 mm, L2=7.5 mm, S=1.0 mm, Ls1=11.4 mm, Ls2=14.3 mm, and g=0.2 mm. Figure 4(b) plots its simulated and measured frequency responses.

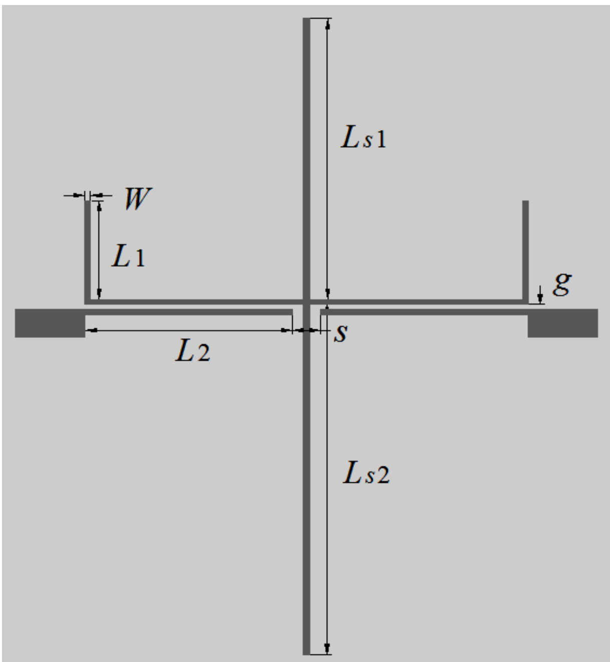


Fig.4 (a) Layout of the proposed filter.

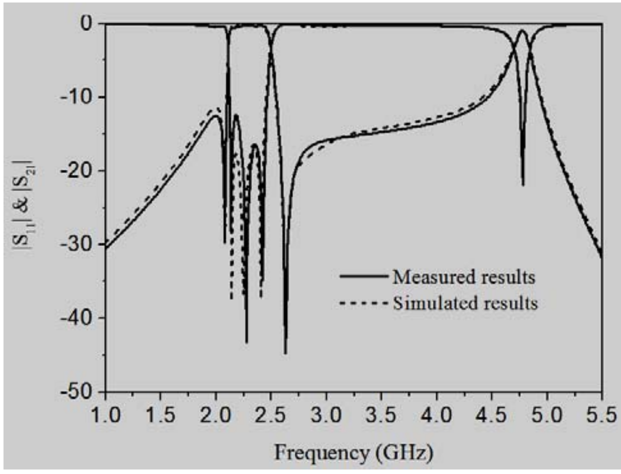


Fig.4 (b) Simulated and measured results.

The two sets of results show good agreement with each other. The measured passband is exactly centred at 2.3 GHz with the minimum insertion loss achieves 0.7 dB, and the return loss is lower than 12.5 dB within the passband, respectively. Furthermore, two transmission zeros, appearing at 2.1 and 2.6 GHz, so as to guarantee the constituted filter to achieve very sharp roll-off skirts out of the desired passband. In final, Table I is provided to compare our proposed filter with those previous works [8-9], where λg is the guided wavelength at 2.3 GHz and roll-off rate is defined as |αmax-αmin|/|fs- fc| (αmax and αmin stand for the 30-dB and 3-dB attenuation points, respectively, fs is the 30-dB stopband frequency and fc is the 3-dB cutoff frequency). It is apparent that the proposed filter has sharp-ended roll-off skirts near cutoff frequencies and thus possesses a high selectivity.

IV. CONCLUSION

This letter presented a simple and efficient method for designing a microstrip BPF with a sharp roll-off skirt frequency response. By proposing the two stubs loaded resonator, the resultant bandwidth can be flexibly controlled and adjusted. Measured results of the fabricated filter have exhibited excellent performances as predicted in analysis.

TABLE I. COMPARISON WITH OTHER REPORTED BANDPASS FILTERS

	3-dB bandwidth	Roll-off rate (dB/GHz)	Size (λg ×λg)	Transmission zeros near cutoff frequencies
Ref. [8] Triple-mode	16.7%	79	0.22×0.55	one
Ref. [9] Without U-shaped I/O coupled lines	15.0%	150	0.59×0.31	one
This work	16.5%	225	0.41×0.65	two

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