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2.9 - 8.4 GHz Filter Bank Using Stub-Loaded Multi-Mode Resonators

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Abstract

This paper consolidates the activity of design and fabrication of 2.9 - 4.32 GHz, 4.3 - 6.42 GHz, and 6.4 - 8.4 GHz filter bank. Novel compact microstrip bandpass filters with stub-loaded multi-mode resonators are proposed. Simulated results indicate that all the filters exhibit insertion losses less than 1.5 dB with passband ripples of 1 dB and sharp attenuations of above 40 dB in their stopbands. The maximums of input and output *VSWRs* are 1.742 and 1.734, respectively. Due to fabrication error, the initial measured passbands show frequency shifts and insertion losses in upper passbands deteriorate seriously. After tuning of the filter bank, measured results imply that the input and output *VSWRs* are found lower than 2.135 and 2.187, and the insertion loss in 1 dB bandwidth is less than 2.52 dB. Filter bank has a sharp skirt and out-of-band rejection levels approaching to 40 dB in all desired stopbands except at the frequencies near $2f_0$.

Keywords

Multi-Mode Resonator, 1 dB Bandwidth, Filter Bank

1. Introduction

A SFB (Switch Filter Bank) is commonly used to handle multi-band communication systems and frequency hopping radar systems, where better selectivity, sensitivity and high isolation are required between the individual channels [1]. When a compact SFB is used in modern ESM (electronic support measures) receiver, the system capacity over the different kinds of signal environment is improved. Filter bank is composed of several bandpass filters (BPFs), which allow signals inside a specific bandwidth at a certain center frequency to pass with minimum loss and reject unwanted signals at both higher and lower frequencies.

Compared with lumped filter bank, the microstrip filter bank configuration has the advantages of a more

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flexible design with very good reproducibility, better power handling capability, and better electrical performance at high frequencies [1]. Moreover, microstrip filter with light weight and low cost can be easily mounted on a dielectric substrate [2]. Due to inhomogeneous nature of the microstrip lines, the odd mode phase velocity is faster than the even mode phase velocity. Therefore, the spurious pass band of a conventional microstrip parallel coupled filter at $2f_0$ and $3f_0$ occurs [3] [4]. Besides, this kind of filter configuration is also found theoretically difficult to be directly employed for designing a wideband filter with a high rejection [5]. In contrast, a filter with a multi-mode resonator (MMR) has outstanding advantages including wide passband, miniature size, and high performance. The original MMR categorized as stepped-impedance resonator (SIR) is proposed in [6] for 3.1 - 10.6 GHz BPF, which consists of one $\lambda/2$ low-impedance line section in the center and two identical $\lambda/4$ high-impedance line sections at the two sides. After that, diverse structures of MMR based on the SIR are reported but all applied to 3.1 - 10.6 GHz BPF [7]-[9]. A cross-junction in the SIR always causes the first transmission pole to unexpectedly shift to the lower transmission zero, thus bringing about poor in-band flatness. Stub-loaded MMR-based filter has been confirmed to have a sharp out-of band rejection, but usually suffers from a narrow upper stopband due to the resonant mode in the MMR filter with SIR [10].

In this paper, to solve these issues, open-end stubs loaded MMR with modified SIR are developed for 2.9 - 4.32 GHz, 4.3 - 6.42 GHz, and 6.4 - 8.4 GHz microstrip BPFs with 20 MHz overlapping frequencies. The proposed SIR has two distinct high-impedance lines loaded at two sides of diverse parallel coupled lines. Two or three MMRs are cascaded to obtain wide upper stopbands with a high rejection. The initial physical dimensions of filters are calculated and optimized using ADS, and then the optimized sizes are tuned slightly using the fullwave electromagnetic simulator HFSS. AutoCad is utilized for preparing PCB layouts and mechanical housings. Three filters, integrated in a single mechanical housing, form an integral part of 2.9 - 8.4 GHz filter bank.

2. Proposed Filter with Multi-Mode Resonator

2.1. Modified Multi-Mode Resonator

Figure 1 shows the proposed multi-mode resonator formed by loading two open-ended stubs at the two opposite sides of a SIR in center [6], respectively. Different from conventional SIR in other papers, the two high-impedance lines with a similar length but not $\lambda/4$ (namely, $L1 = L2$) in the SIR are permitted to have a part of distinct lengths for different coupled lines. When the length of loaded stub is longer or shorter than $\lambda/2 - L1$ and $\lambda/2 - L2$, which can create a transmission pole at a frequency lower or higher than the desired passband, thus sharpening the roll-off skirt in either lower or higher cut-off frequencies. Here, the length of stub is set as a value close to $\lambda/2 - L1$, so that the resonant frequency f can locate at center frequency, which enhances the reflection loss and improves |S21| further.

$$f = \frac{c}{2(L1 + L3)\sqrt{\varepsilon_r}}. \tag{1}$$

where c is the speed of light and ε_r is the effective permittivity of substrate.

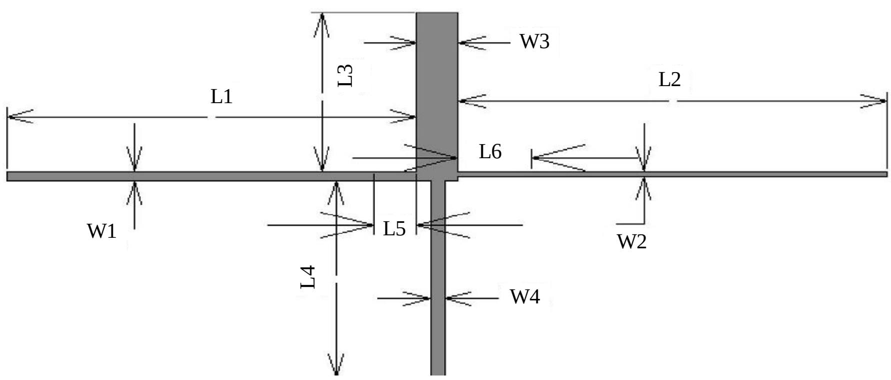

Figure 1. Proposed multi-mode resonator.

2.2. Proposed Band Pass Filters

Operation frequency and substrate material affects the size of a filter, and substrate thickness the bandwidth of a filter. To realize the size-miniaturization and obtain a wide band, all the filters are designed and implemented on the Rogers RT Duroid dielectric substrate having permittivity of 10.2, tangent loss of 0.0022, and thickness of 0.625 mm. For the convenience of discussion later, these three filters are denoted as Filter A, Filter B, and Filter C.

Figure 2 illustrates the geometrical sketches of designed filters using MMR proposed in **Figure 1**. The Filters A and B contain two cascaded MMRs with similar two stubs. If the MMRs do not have these stubs, due to the coexistence of two pairs of different coupled lines, the filter will show two transmission poles (TPs) near f_0 and $2f_0$, respectively. When four stubs are added in the MMR, the TPs near $2f_0$ move toward the lower frequency significantly and locate near the upper passband, while the first two resonant frequencies almost remain unchanged. Therefore, the open-circuited stubs can be applied to adjust the high resonant modes from the MMR into desired passband, which deepens the upper stopband [11] and improves the rejection near $2f_0$ simultaneously. The introduction of stubs in the MMR also lessens the passband ripple and increases the reflection loss. The loaded open-circuit stubs at RF ports generate TPs at lower and upper passbands or reallocate the resonant modes from MMR. Similarly, among Filter C with three cascaded MMRs, the two stubs added in the central MMR move the TPs near $2f_0$ into passband, while two mirror images MMRs in geometrically symmetrical location flatten the fluctuation of insertion loss. In the proposed filters, two identical parallel coupled lines at input/output ports are stretched longitudinally so as to raise the frequency-dispersive coupling degree with a coupling peak near the lower passband. By adjusting the length of the coupled lines, the frequencies of pass and stop bands can be varied. In addition, we can also increase the magnitude of |S21|, and when the length is equal to the center frequency of passband, |S21| at center frequency approximates to the desired 0 dB.

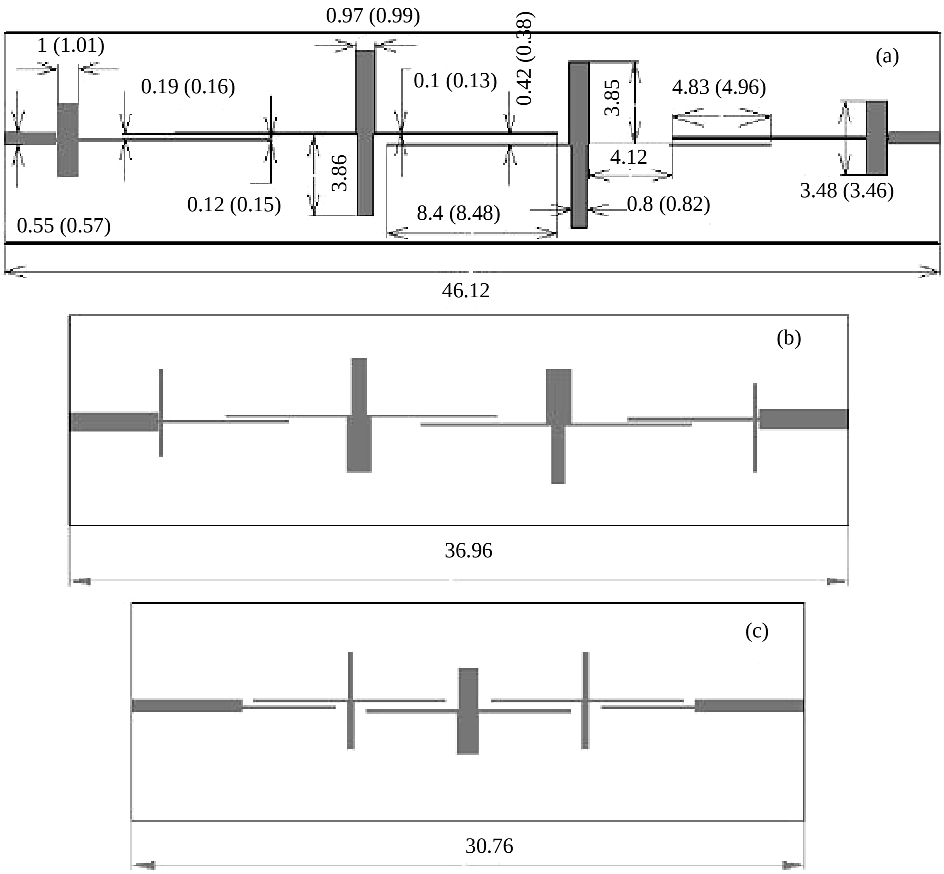

Figure 2. Topologies of the designed filters (a) 2.9 - 4.32 GHz; (b) 4.3 - 6.42 GHz; (c) 6.4 - 8.4 GHz BPFs. The designed and fabricated sizes in brackets are labeled in mm.

Comparing the three filters shown in **Figure 2**, the resonators with a longer electrical length are used to design the lower passband, while the resonators with a shorter electrical length are used to design the upper passband [12]. The actual bandwidth of the filter appears to be always less than the value assumed in the design, and therefore it would be desirable in calculating the parameters of a given filter to use a somewhat larger bandwidth than actually is required. In order to have the wave impedance of 50 ohms, the width of microstrip lines in the input and output ports are kept as 0.55 mm. However, for achieving better *VSWRs*, the width of the two lines are set as 0.88 mm in Filter B. The sizes of other lines are optimized to have a good performance. To have a degree of freedom in tightening the coupling degree and achieve a coupling peak in the lower passband, the strip and slot widths in the coupled lines should be chosen carefully. An important part of understanding any particular design problem is a consideration of the trade-offs. Increasing the bandwidth means less loss in the passband but reduced selectivity. Increasing the ripple will increase the selectivity but at the cost of decreased return loss [13]. Therefore, the tradeoff among these targets should be balanced

3. Simulated and Measured Results

Figures 3-5 display the |S21| in dB and *VSWRs* against the frequency. The dash lines represent the simulated frequency responses and the solid and symbolic lines the final measured frequency response. For simplicity, the original measured data, in which all passbands present a shift of about 300 - 500 MHz to low frequencies, are not included in this paper. Lalitha *et al*. [1] thought the frequency shift in their filters could be attributed to poor packaging and mechanical housings. In this paper, in our opinion, it is mainly dominated by the discrepancy between the designed and fabrication sizes. From the designed and fabricated sizes marked in **Figure 2**, the fabrication tolerances of width and length are 0.02 mm at least, while that of the spacing between coupled lines is below −0.03 mm, which significantly influences the properties of filters. Zhu *et al*. [14] also reported the shifted bandwidth due to fabrication tolerance, even if very small. As stated before, resonator with longer electrical length causes a lower passband. Based on these opinions, the lengths of high-impedance lines in SIRs are shortened slightly in tunning. However, the enlarged widths and the shrinked spacing in the MMR, which also importantly dictate the electrical performance, are difficult to optimize to obtain better actual characteristics.

From **Figure 3**, six transmission poles are formed in the passband of the simulated *VSWRs* curves. The first and last poles is contributed by the stubs loaded at RF ports, while the second and third resonant frequency are due to the coupled lines in MMR. As described before, the fourth and fifth poles, almost merging together, are adjusted into the passband from the frequencies near $2f_0$ by the stubs loaded in MMRs. However, the sixth pole disappears in the measurements, as a result of which, the insertion loss and *VSWRs* in upper passband increase fast. In contrast, the loaded stubs at two sides of RF ports do not produce new resonant modes in Filter B (see **Figure 4**). The values at its second and third poles keep very small. The same number of poles exists in the final

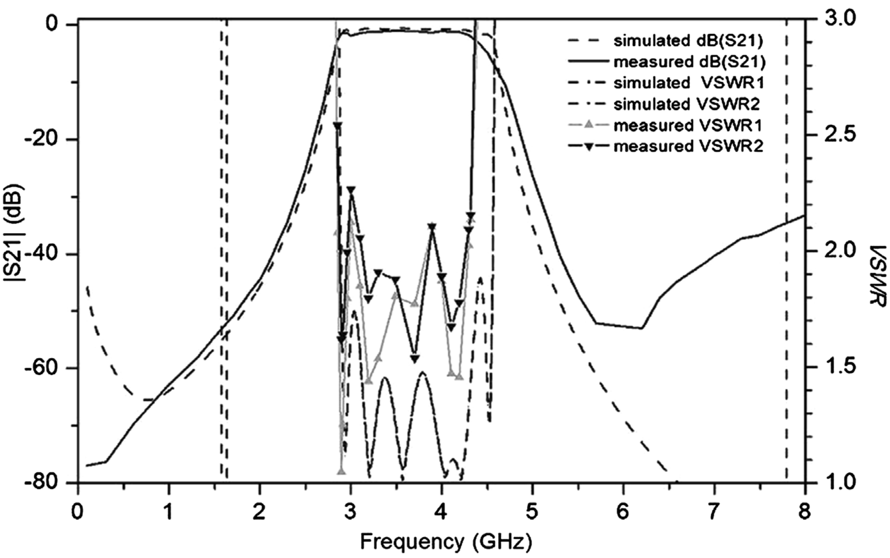

Figure 3. Simulated and measured results of 2.9 - 4.32 GHz BPF.

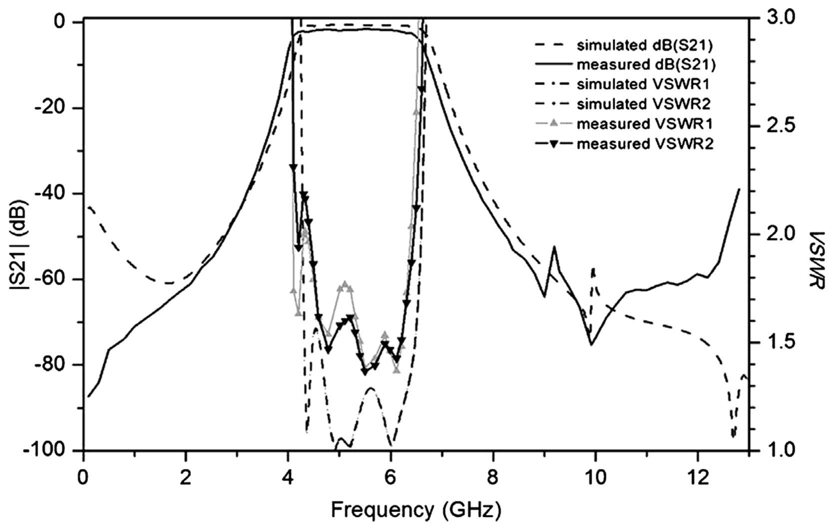

Figure 4. Simulated and measured results of 4.3 - 6.42 GHz BPF.

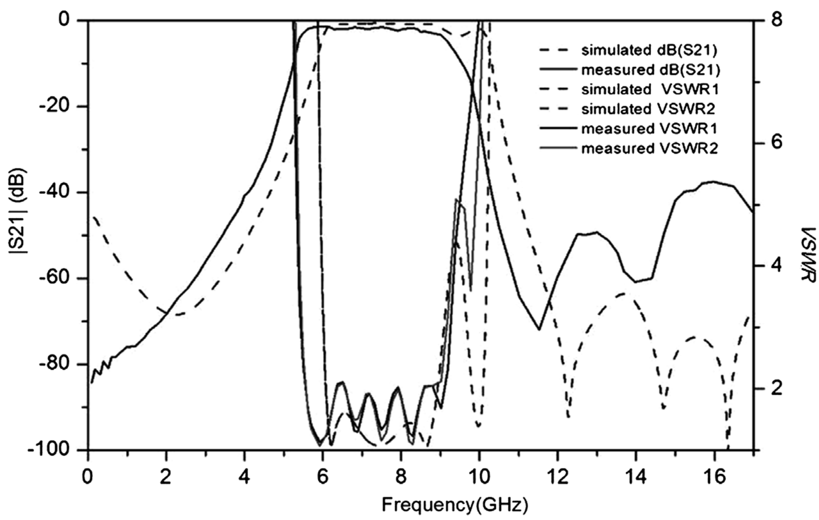

Figure 5. Simulated and measured results of 6.4 - 8.4 GHz BPF.

measured results, but the frequency shift can still be seen clearly. What is worse, more notable frequency shift occurs in the final measurements as shown in **Figure 5**. There are three resonant frequencies staying in the passband of 6.4 - 8.4 GHz, which are attributed to the coupled lines and stubs in MMRs. However, the loaded stubs also generate the resonant mode centered at 10GHz and it is difficult to keep the four resonant frequencies at the desired frequencies simultaneously.

Stub-loaded MMR-based filter usually suffers from a high *IL* of about 2 dB in the upper passband and a narrow upper stopband. The former is mainly caused by parasitic radiation from the central part with wide strip conductor at high frequency, while the latter is due to the resonant mode in the MMR filter with SIR. However, the *ILs* in our simulated filters with a 1 dB ripple in passbands are smaller than 1.5 dB. On the other hand, the upper stopbands are extended up to high frequencies of 8.4 GHz, 12.8 GHz, and 16.8 GHz and the lower stopbands are stretched to DC, with sharp attenuations above 40 dB, even at $2f_0$.

Figures 3-5 imply that the final measured results are similar to the simulated ones in passbands, and have better rejections below lower stopbands but poorer rejections above upper stopbands. Even though the measured frequency response of the filter do not coincide with the designed ones well, their performances are still reasonably acceptable. For better understanding the properties of the fabricated filters, their main performances are summarized in **Table 1**. **Table 1** indicates that the developed filter bank has an *IL* less than 2.52 dB with input

Table 1. Performance of the fabricated filters.

Performance	Designed Filters		
	Filter A	Filter B	Filter C
Passband (GHz)	2.9 - 4.32	4.3 - 6.42	6.4 - 8.4
IL (dB)	1.053 - 2.057	1.568 - 2.518	1.565 - 2.339
*VSWR*1	1.048 - 2.135	1.383 - 2.037	1.229 - 2.11
*VSWR*2	1.539 - 2.157	1.427 - 2.187	1.101 - 2.087
Stopband (GHz)	DC - 2.15 & 5.8 - 8.4	DC - 3.2 & 8.4 - 12.8	DC - 4.2 & 12.8 - 16.8
Rejection (dB)	>40 & 31.11	>40 & 39.84	>40 & 37.48

and output *VSWRs* below 2.2. The rejection in each filter is better than 40 dB in their stopbands except at the frequencies near $2f_0$. This larger loss originates likely from the losses of the coaxial connectors and their poor contacts to the microstrip line. Shortened lengths of coupled lines by us compensate the shift in passband, but increase the *IL* to a large extent. The degraded *IL* can also be explained by the fabrication error, which deteriorates the *VSWRs* as well. For achieving good agreement between measurement and simulation, fabrication tolerance of facilities should be considered.

4. Conclusions

This paper presents the design, fabrication, and measurement of 2.9 - 8.4 GHz filter bank consisting of 2.9 - 4.32 GHz, 4.3 - 6.42 GHz, and 6.4 - 8.4 GHz microstrip filters with multi-mode resonators. This paper demonstrates the approach to overcoming the limitation of rejection as well as suppression of the harmonic, which is prominent in the microstrip coupled filter topology.

Simulated results of filter bank indicate that the insertion loss is no more than 1.5 dB in 1 dB bandwidth, input and output *VSWRs* are below 1.742 and 1.734, and higher than 40 dB rejection is achieved. Due to fabrication error, the final measurements give degraded but acceptable filter bank characteristics. The fabricated filter bank provides an insertion loss of less than 2.52 dB with 1 dB ripple in each channel, *VSWRs* below 2.2, and rejection of above 40 dB in their stopbands except at the frequencies near $2f_0$. Poor insertion loss and *VSWRs* result from the size change of filters in fabrication as well as in tuning due to frequency shift. For achieving good agreement between measurement and simulation, fabrication tolerance of facilities should be considered.

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