COMPACT MICROSTRIP BANDPASS FILTER USING MINIATURIZED HAIRPIN RESONATOR

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Abstract—A compact microstrip bandpass filter using miniaturized hairpin resonators is presented in this letter. Two modified stepped impedance hairpin resonators connected by parallel coupling are designed for the bandpass filter. The proposed miniaturized hairpin resonator is composed of a microstrip line and rectangular ring structures between parallel high impedance lines. A big capacitance in the hairpin resonator is provided by the gaps of rectangular ring structures in the parallel high impedance lines. Therefore, the proposed bandpass filter using the hairpin resonators has a low insertion loss, low return loss and compact size. The proposed bandpass filter with a center frequency of 4.96 GHz is designed with EM full wave simulator IE3D and verified with experiment.

1. INTRODUCTION

A planar compact, light weight and high performance microwave bandpass filter is generally desired in microwave communication systems. The conventional bandpass filter generally uses a large size for sharp rejection, low insertion loss and wide stopband [1].

The hairpin resonator using the modified stepped impedance is introduced in [2]. However, the undesirable attenuation pole in the stopband causes the narrow bandwidth for the out of band rejection. Specially to remove the undesired attenuation pole, the filter is added with a defected pattern on the ground plane as presented in [3]. However, the filter with aperture is difficult to fabricate and have no diverse characteristics. As an alternative method, the hairpin resonator using the interdigital capacitors is presented in [4, 5]. The internal coupling effect of the filter is depending on the interdigital

capacitance, which is related to the finger number. Increasing the number of the finger provides bigger capacitance for the low insertion loss, wide stopband, and sharp rejection. However, the filter needs to have a large size for the large number of the fingers with bigger capacitance [4, 5].

In this letter, a new compact hairpin resonator using rectangular loop structures in high impedance parallel lines is proposed for bandpass filter. The couplings in the high impedance parallel lines of the hairpin resonator are operated as the capacitance in the frequency response. Therefore, the proposed bandpass filter is easy to adjust the coupling coefficient inside the resonator and the coupling coefficient between hairpin resonators to provide an improved insertion loss, a wider and a deeper stopband without the change of the filter size. The proposed bandpass filter is also easy for fabrication because there is no pattern on ground plane. The dimensions of the bandpass filter are optimized by the EM simulation for the desirable frequency response. The simulated result and measured result agree with each other.

2. MODIFIED HAIRPIN RESONATOR

Figure 1(a) shows the schematic of the modified hairpin resonator using rectangular loop structure to obtain a strong coupling in the parallel high impedance lines. The proposed filter is connected with the feed lines. Figure 1(b) represents the equivalent circuit of the modified hairpin resonator [2]. The different microstrip line length L_t and the varied coupling capacitance C provide the different range of the passband and the location of the attenuation pole in the stopband [4]. C_j denotes the capacitance between the microstrip line, the parallel
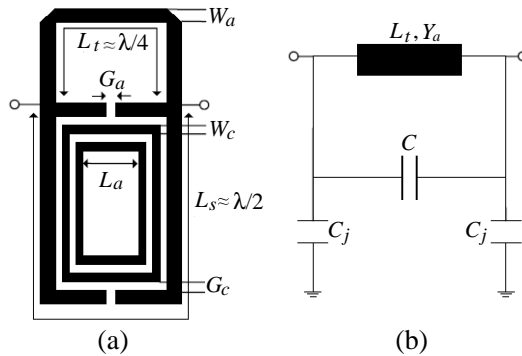
Figure 1. Modified hairpin resonator, (a) schematic, and (b) equivalent circuit model.

high impedance length, the rectangular loop structures and the ground.

The internal coupling coefficient in the modified hairpin resonator is affected by the gaps G_a, G_c between the parallel high impedance lines L_s. Therefore, the gaps between the coupled high impedance lines L_s are related to the capacitance C. In the equivalent circuit, the capacitance C is dependent on the coefficients of the gaps in the coupled high impedance lines L_s. Neglecting the junction discontinuity C_j for easy analysis, the S-parameter related to attenuation poles is given by [4]

$$S_{21} = j\frac{2Y_0(\omega C - Y_a \csc(\beta L_t))}{Y}$$
$$Y = Y_0^2 + Y_a^2 + j2Y_0(\omega C - Y_a^2 \cot(\beta L_t)) + \omega C Y_a(\cot(\beta L_t) - \csc(\beta L_t)) \tag{1}$$

where Y_0 is the admittance of the input and output, and Y_a and L_t are the admittance and the electrical length of the microstrip line. C is the capacitance of the gap in the parallel coupled high impedance lines, β the propagation constant, and ω the angular frequency.

In the above equation, the attenuation poles are provided by $S_{21} = 0$. As shown in Figure 2(a), the number of the attenuation pole is determined by the intersection between ωC and $Y_a \csc(\beta L_t)$. The value of the lumped capacitor C provides the number of the attenuation pole. Therefore, different dimensions of the square loop structure in the coupled high impedance lines can change the gap G_c which provides the total capacitance C. The unloaded quality factor of the modified resonator is calculated by the definition [7]. Figure 2(b) shows the number of the attenuation poles and unloaded quality factor according to the capacitance affected by different gaps. As shown in Figure 2(b),
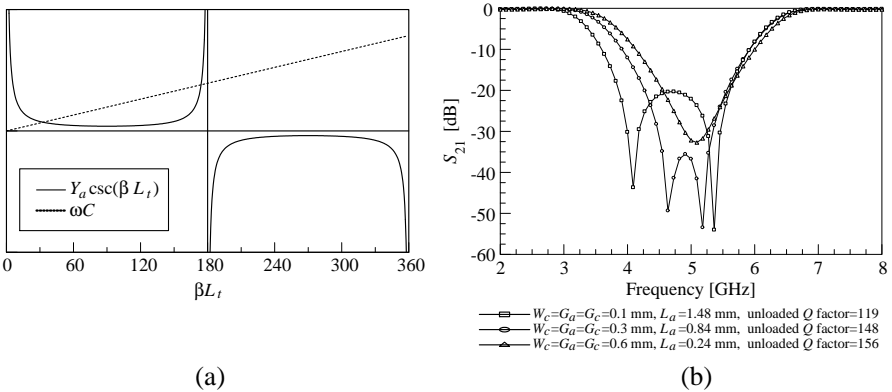
Figure 2. (a) Pictorial description of the capacitor. (b) S_{21} and unloaded quality factor with different capacitances ($L_t = 5.4$ mm, $L_s = 11.2$ mm, $W_a = 0.3$ mm for $64.5\,\Omega$, $W_c = 0.1$ mm).
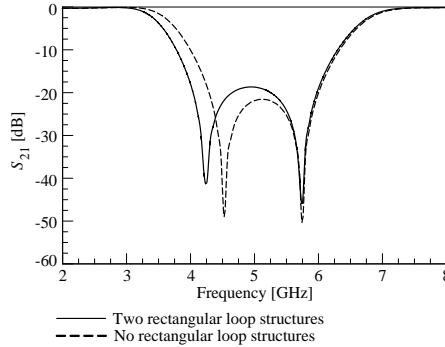
Figure 3. S_{21} differences between the inserted rectangular loop structures in the parallel high impedance lines without them ($L_t = 5.34 \, \text{mm}$, $L_s = 10.62 \, \text{mm}$, $W_a = 0.3 \, \text{mm}$ for $64.5 \, \Omega$, $G_a = G_c = W_c = 0.1 \, \text{mm}$, $L_a = 1.45 \, \text{mm}$).

the loose coupling of the structure provides only one attenuation pole. The more the coupling coefficient increases, the more the number of the new locations the one attenuation pole can be split into two locations. The bigger capacitance creates the large gap between the location of the first attenuation pole and the location of the second attenuation pole.

In this paper, the attenuation pole is always provided more than one because of the structure. In the paper [4], the smaller length L_t and bigger capacitance C change the locations of the attenuation pole and the cutoff frequency for the sharp rejection and wide stopband. In this paper, the microstrip line L_t and parallel high impedance line L_s in the modified hairpin resonator are approximately set with the electrical length of $\lambda/4$ and $\lambda/2$ optimized by a full wave electromagnetic simulator IE3D. Figure 3 shows the main effect of the inserted rectangular loop structures in the parallel high impedance lines is the coupling coefficient. The inserted rectangular loop structures in the parallel high impedance lines increase the coupling coefficient. Also to compare the size of the conventional hairpin resonator of the paper [4], the modified hairpin resonator has the same center frequency of 4.9 GHz is designed on the same substrate of RT/Duroid 6010.8 has a thickness of 25 mil and the relative dielectric constant $\varepsilon_r = 10.8$. The overall size of the conventional hairpin resonator with the five-finger capacitors is $2.78 \times 6.57 \, \text{mm}^2$ in the paper [4]. As shown in Figure 3, the overall size of the new compact hairpin resonator is $2.82 \times 6 \, \text{mm}^2$. The new compact hairpin resonator has a 7.65% size reduction compared to the conventional hairpin resonator.

This paper suggests a simple design of the hairpin resonator

for the desirable resonant frequency. After adding the rectangular loop structures in the parallel high impedance lines for the bigger capacitance, the modified hairpin resonator provides the locations of the suitable attenuation poles to optimize the stopband and sharp rejection response with same size.

3. PROPOSED BANDPASS FILTER

The proposed bandpass filter with two modified resonators is shown in Figure 4. The proposed bandpass filter has a coupling structure separated by s and offset by d. The strength of the fringe field near the modified resonators is determined by spacing s and offset d. The bandpass filter with high performance requires a strong fringe field determined by the coupling coefficient. The coupling coefficient K between two resonators is given by [6]:

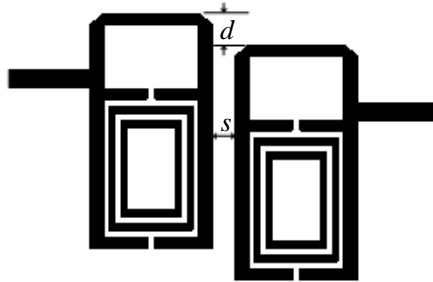$$K = \frac{f_2^2 - f_1^2}{f_2^2 + f_1^2} \tag{2}$$

Figure 4. The schematic of the proposed bandpass filter.
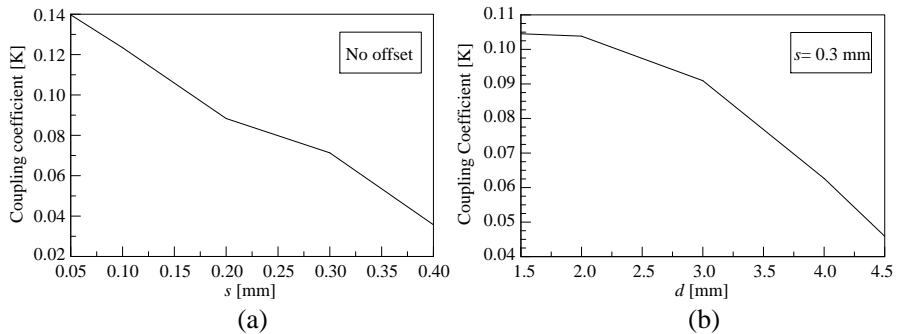

Figure 5. Coupling coefficient with (a) different offset s, (b) different offset d.

Here, f_1 and f_2 are the first and second resonant frequency. The coupling coefficient of the proposed bandpass filter calculated by a full wave electromagnetic simulator is shown Figure 5. If the space s and offset d between two resonators decreases, the coupling coefficient increases. The location of the transmission zero can be closed to the passband of the bandpass filter according to the coupling coefficient.

In this letter, the proposed bandpass filter is designed by the modified hairpin resonators with a center frequency of 4.8 GHz. The substrate RT/Duroid 6010.8 with a thickness of 25 mil and the relative dielectric constant $\varepsilon_r = 10.8$ is used to build the modified hairpin resonator with rectangular loop structures. The desirable frequency for the proposed bandpass filter is also controlled by the space s and offset d between two resonators in the proposed bandpass filter. The simulated result for the desirable frequency response for the filter is obtained by a full wave electromagnetic simulator IE3D.

Figure 6 shows the fabricated filter prototype and the results of the proposed bandpass filter using the modified resonators. The space s and offset d in the bandpass filter are set with 0.3 mm and 4 mm for the strong coupling coefficient. The measurement results of the bandpass filter give a return loss better than 10 dB from 4.8 to 5.3 GHz and an insertion loss of less than 2 dB from 4.8 to 5.4 GHz. Two attenuation poles of the bandpass filter are located at 3.5 GHz with 51 dB attenuation and 7.2 GHz with 43 dB next to the pass band. Therefore, the proposed bandpass filter achieves the high performances which are the wider and deeper rejection bandwidth in the stopband.

(a)   (b)
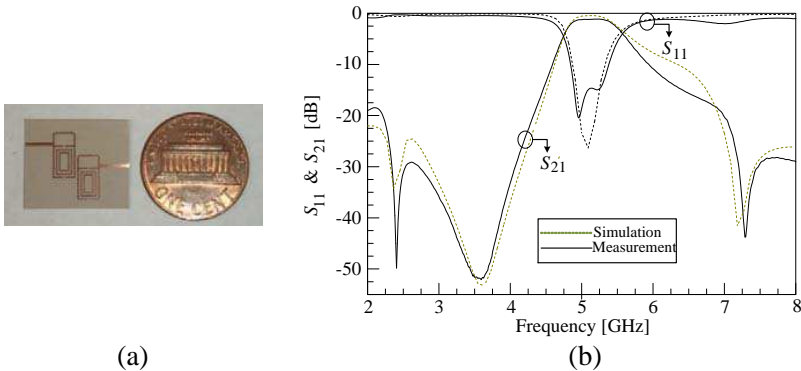
Figure 6. (a) Fabricated filter prototype. (b) Simulated and measurement results of the proposed bandpass filter ($L_t = 5.4$ mm, $L_s = 11.2$ mm, $W_a = 0.3$ mm for $64.5\,\Omega$, $G_a = G_c = W_c = 0.3$ mm, $L_a = 1.29$ mm).

4. CONCLUSION

A proposed bandpass filter using modified hairpin resonators is presented. A simple design and analysis method of the miniaturized hairpin resonator is given for the desired frequency performance. The gaps between the high parallel impedances in the hairpin resonator can be easily used to control a capacitance without changing the dimensions of the resonator. Therefore, the proposed miniaturized hairpin resonator provides a compact size, wider and deeper rejection. The proposed hairpin resonators are connected with the coupling method for a bandpass filter. The proposed bandpass filter provides a low insertion loss, good return loss, deeper rejection, and compact size. The proposed bandpass filter can be used for many applications.

ACKNOWLEDGMENT

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