

Improved microstrip hairpinline bandpass filter using via ground holes and capacitive gap

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Abstract A novel design methodology for an improved hairpinline microstrip narrowband bandpass filter is presented in this paper. The new proposed methodology incorporates use of via ground holes and a capacitive gap to improve the performance and dimensions of a conventional hairpinline bandpass filter. This design approach incorporates use of $\lambda/4$ resonators thus reducing the overall size of the filter by 15.5 % as compared to the conventional design. A capacitive gap is also introduced in the center of the center resonator. The proposed design achieves 3 dB fractional bandwidth less than 4 %, insertion loss less than 0.8 dB and return loss better than 40 dB at the center frequency of 1 GHz. The effects of tap point height, characteristic impedance and gap width are analyzed in detail and subsequent relationships are developed with the fundamental design parameters. Based on the proposed design, a bandpass filter is designed and fabricated on FR4 substrate, and good agreement is observed between measured and simulated results.

Keywords Bandpass filter · Narrowband · Microstrip · Hairpinline · Via ground

1 Introduction

Emerging applications of wireless communications in aerospace industry continue to challenge the RF/microwave signals with ever more stringent requirements of

higher performance, smaller size, lighter weight and lower cost. For narrow band applications, higher performance means fractional bandwidth (FBW) less than 5 % along with insertion loss (IL) less than 3 dB, and return loss (RL) greater than 15 dB. The electronic warfare (EW) applications in aerospace confine the assigned spectral limits to even more stern tolerances. The conventional filter design approaches, such as a coupled resonator filter and a parallel-coupled resonator filter, provide the basic guidelines, however they do not offer a ready solution for such challenging performance parameters. Using a dielectric substrate with high dielectric constant (ϵ_r) would result in a narrower microstrip line width that would cause reduction in size, however, a narrower line results in stronger input/output coupling or a smaller external quality factor (Q_e) [1]. Narrower bandwidth can be achieved through larger gaps between coupled resonators, but this in turn increases the filter size. Microstrip pseudo-interdigital bandpass filters using resonator slightly shorter than $\lambda/4$, have IL of 1.5 dB with 2.07 % FBW [2]. Use of an expensive substrate (LaAlO_3) with a lumped capacitor increases the complexity of design and synthesis cost. Filters designed using cross coupled hairpin resonators have even smaller sizes with FBW of 2.07 %, but IL of 3.8 dB does not fulfill the requirements of receiver side front-end bandpass filters [3, 4]. Compact split ring stepped impedance resonators have an FBW close to 4 % at 2.4 GHz, however the RL is limited to 20 dB with out of band rejection of 25 dB [5]. FBW of 1.5 % can be achieved using square loop bandpass filters, but they provide RL of 10 dB in pass band [6].

In this paper, a novel microstrip hairpinline narrowband bandpass filter using via ground holes and a capacitive gap is presented. The proposed design approach further improves the methodology presented by the authors for a microstrip narrowband bandpass filter using via ground

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holes [7]. The via hole diameter and metallization area are kept to a minimum as it constitutes the significant part of the total inductance of the via structure [8]. The resonators are kept $\lambda/4$ long with a capacitive gap in the center resonator. Both arms of the center resonator and the inner arms of third and fourth segment are grounded using via ground holes. An increased number of via holes reduces the overall high impedance effect of via grounds and provides greater flexibility in choosing the microstrip resonator characteristic impedance [9]. In the conventional design the resonator acts as an open circuited transmission line which resonates if its length is an integer multiple of the half wavelength i.e. $\lambda/2$, $3\lambda/2$ etc. By introducing via ground holes on arms of the resonator it acts as a transmission line with one end open circuited and the other end short circuited. Such transmission lines resonate at the frequency corresponding to the length being odd multiples of quarter wavelength i.e. $\lambda/4$, $3\lambda/4$ etc. The use of via ground holes reduces the overall size of the filter as compared to the conventional design. The capacitive gap in the center resonator caused an improvement in the IL. Filter design is explained using Rogers TMM 06 substrate with a dielectric constant 6.08, and the overall filter size is reduced by 15.5 % of the conventional hairpinline filter. FBW of 0.9–4.5 % is achieved at 1 GHz while maintaining IL less than 2 dB and RL better than 30 dB. The effect of tap point height, characteristic impedance and microstrip gap are studied in detail in this paper. The proposed filter may be designed for midband frequencies up to 3 GHz, beyond which folding the resonator becomes difficult. Design robustness is verified on two more substrates with dielectric constants 2.17 and 9.2. The frequency selectable aspect of proposed passive filter design on a commonly available substrate renders it viable for a wide range of applications.

2 Improved bandpass filter design

Based on the conventional hairpin line filter design, the proposed filter is a significant improvement in terms of size reduction, narrower FBW and higher Q_e . To demonstrate the design approach, a three-pole filter was designed and the design parameters: external quality factor and coupling coefficients; were calculated using design equations reported in [10]. The proposed filter incorporates a capacitive gap in addition to the via ground holes. The 1st and the 5th hairpin-line sections do not resonate around the passband center frequency but only act as part of the input and output feeding structures. Both arms of the center resonator and the inner arms of the 1st and 3rd resonator are grounded using via ground holes. The diameters and top pad area of via holes are kept within minimum

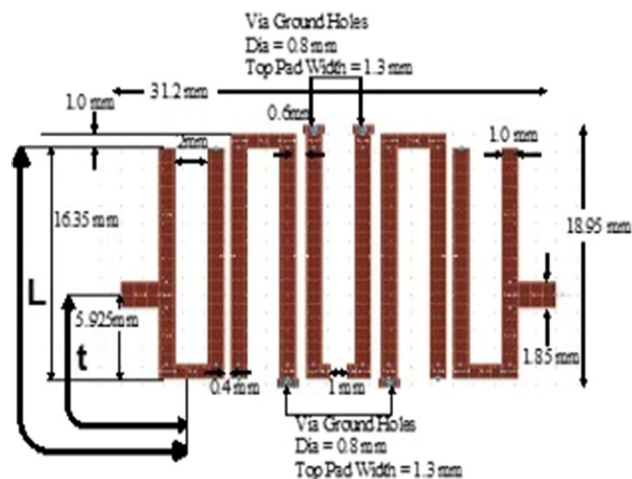


Fig. 1 Layout of the proposed filter designed using Rogers TMM06 ($\epsilon_r = 6.08$)

practicable limits of available prototyping facility. Furthermore, a microstrip gap is introduced in the center resonator. A commercially available substrate of Rogers TMM 06 with relative dielectric constant 6.06 and thickness 1.27 mm is used for this design. The characteristic impedance of the hairpin line (Z_r) is kept greater than the terminating impedance (Z_o) with 68Ω typically set as the starting value for Z_r (1.0 mm wide). Spacing between the resonator arms is set to the typical value of 2 mm and spacing between the adjacent resonators is kept at 0.4 and 0.6 mm. After selecting the tap point height and Z_r corresponding to the target specifications, the gap width is also optimized for the best performance results. Gap width is set equal to the microstrip resonator width as the starting value. Filter layout is shown in Fig. 1 and S-parameter results are shown in Fig. 2. It can be observed that a quite narrow 3 dB FBW of 3.5 % is achieved with IL of 0.734 dB. The RL is greater than 42, and 20 dB FBW of just 11 % indicates good out of band rejection. The independent parameters for this design include tap point height t , characteristic impedance Z_r and the capacitive gap width w_g . Dependant parameters are FBW, Q_e , RL and IL. Through sensitivity analysis, RL and IL are related to the independent variables using design graphs. Relationships are developed to relate Q_e and FBW to the independent variables.

2.1 Effects of tap point height

For a conventional design, the tap point height (t) is linked with Q_e for conventional hairpinline filter by the expression shown in Eq. (1) in which L is the resonator length [2]. The conventional design employs $L = \lambda/4$ while in the proposed design, $L = \lambda/8$ as depicted in Fig. 1.

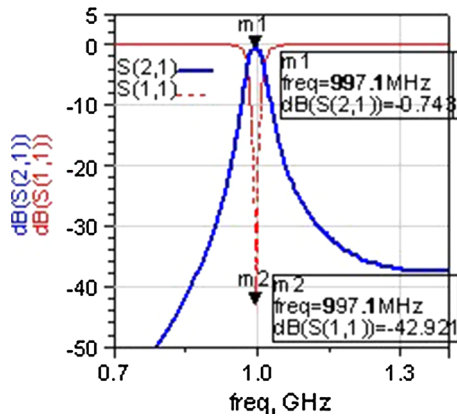


Fig. 2 S-parameter results of proposed filter designed using Rogers TMM06 ($\epsilon_r = 6.08$)

$$t = \frac{2L}{\pi} \sin^{-1} \left[\sqrt{\frac{\pi Z_o/Z_r}{2 Q_e}} \right] \tag{1}$$

Target FBW of proposed design is less than 5 % which corresponds to Q_e higher than 30. The lowest practicable value of tap point (t) can be calculated simply by adding microstrip resonator width, half of the microstrip tap width and one half of the separation between two arms of first/last resonator. The lowest practicable value of t thus calculated for proposed design is 3.43 mm, while the maximum value of t calculated using Eq. (1) is 2.934 mm. Therefore, Eq. (1) does not give practicable values for t for the design presented in this paper. Tap point locations vary over a much larger range for the proposed design in terms of Q_e . As t is increased from 4.93 to 13.93 mm, the Q_e correspondingly decreases from 32.64 (FBW 3.513 %) to 30.34 (FBW 3.779 %). Using the least square fit through points, a second order polynomial was developed to explain the relationship between Q_e and t/L for the proposed design as shown in Eq. (2).

$$Q_e = -10.268 \left(\frac{t}{L}\right)^2 + 5.8803 \left(\frac{t}{L}\right) + 31.803 \tag{2}$$

Q_e can be calculated by using Eq. (1). The expression for Q_e extracted from Eq. (1) is shown in Eq. (3)

$$Q_e = \left(\frac{\pi}{2}\right) \left(\frac{Z_o}{Z_r}\right) \frac{1}{\sin^2\left(\frac{\pi t}{2L}\right)} \tag{3}$$

Values thus calculated from Eq. (1) or (3) are not same as the results of Eq. (2) and are thus does not hold valid for the proposed design. In the proposed design Q_e improves by lowering the tap point, however this improvement cannot be calculated using Eq. (1). Sensitivity analysis for the tap point height in terms of S_{11} was carried out, as shown in Fig. 3. There is an observable effect of t on RL.

It can be observed that RL improves as t is lowered from 13.93 to 7.93 mm, however from 7.93 till 5.93 mm

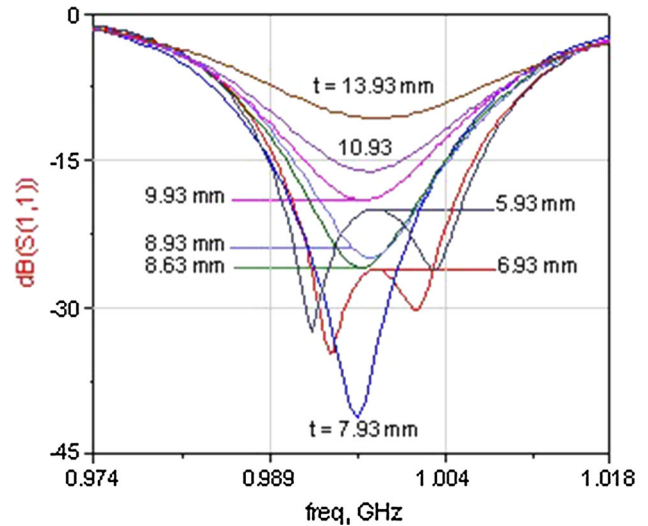


Fig. 3 $S(1,1)$ for different tap point heights t

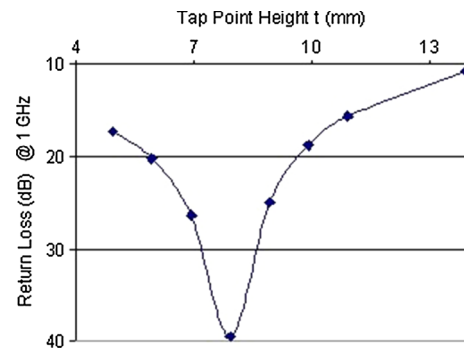


Fig. 4 Return loss versus t

reflections start to increase and RL decreases. Values of RL and IL were recorded for different tap point heights and the design graphs relating RL to t and IL to t were developed as shown in Figs. 4 and 5, respectively.

Although a lower tap point guarantees a better Q_e , however design optimization is required to find the best combination of IL and RL. Best values of RL correspond to the value of t that gives best impedance matching and minimum reflections. Impedance matching can be successfully accomplished by adjusting the taping point. It was observed that the input impedance reduces with considerably larger increase in the argument as the tap point is lowered and vice versa.

2.2 Effects of characteristic impedance of hairpinline (Z_r)

Sensitivity analysis was carried out to study the effect of microstrip width and the results in terms of S_{11} and S_{21} are shown in Figs. 6 and 7, respectively.

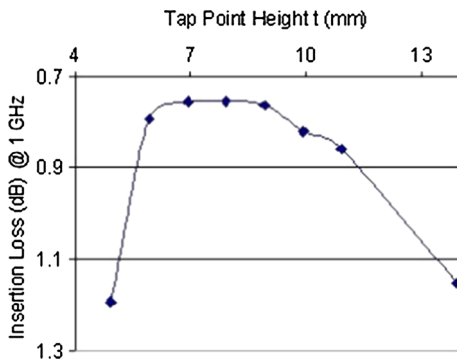


Fig. 5 Insertion loss versus t

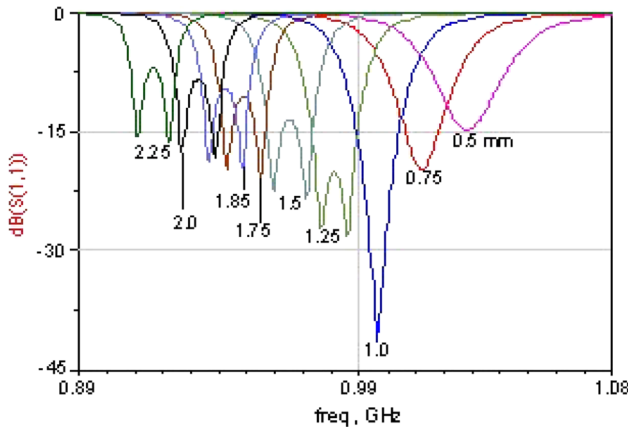


Fig. 6 S_{11} plotted for microstrip width 0.5, 0.75, 1.0, 1.25, 1.50, 1.75, 1.85, 2.0, and 2.25 mm

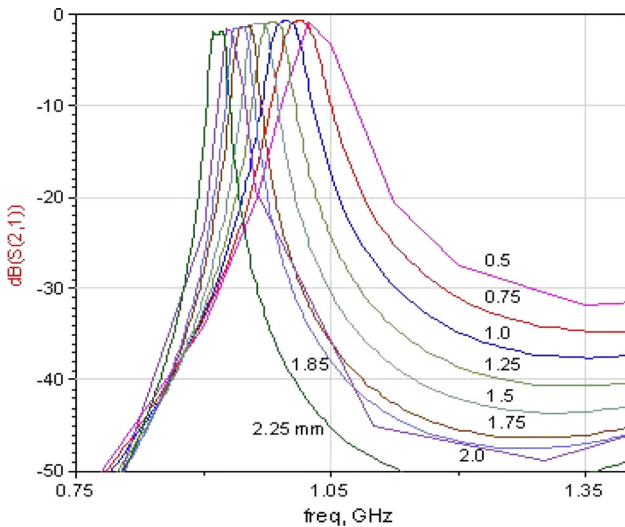


Fig. 7 S_{21} plotted for microstrip width 0.5, 0.75, 1.0, 1.25, 1.50, 1.75, 1.85, 2.0, and 2.25 mm

The design graphs relating the effect of varying microstrip resonator width on IL and RL are depicted in Figs. 8 and 9 respectively.

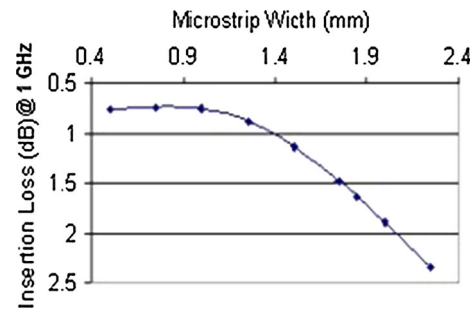


Fig. 8 Microstrip width versus IL

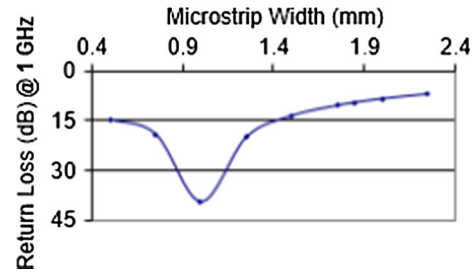


Fig. 9 Microstrip width versus RL

Generally, IL improves with reduction in microstrip width [1], however it is observed that for microstrip width smaller than 1mm, IL starts to degrade because of the higher reflection losses. It is also observed that FBW decreases noticeably with an increase in the resonator width. As the value of t/L is already fixed, Q_e becomes dependent only on Z_r as shown in Eqs. (4) and (5).

$$Q_e = \left[\frac{\pi}{2 \sin^2\left(\frac{\pi t}{2L}\right)} \right] \frac{Z_o}{Z_r} \tag{4}$$

$$\Rightarrow Q_e \propto \frac{Z_o}{Z_r} \tag{5}$$

Q_e calculated using Eq. (3) against Z_o/Z_r shows a linear relationship as verified from Eq. (6) (established through curve fitting on data generated using Eq. (4)).

$$Q_e = 0.5302 \left(\frac{Z_o}{Z_r} \right) + 0.538 \tag{6}$$

However, the relation is practically non-linear. Least square fit through points was used to establish a relationship between Q_e and Z_o/Z_r using the simulated data as shown in Eq. (7).

$$Q_e = 42.442 \left(\frac{Z_o}{Z_r} \right)^2 - 20.33 \left(\frac{Z_o}{Z_r} \right) + 24.47 \tag{7}$$

In this design, higher Q_e can be achieved with lower Z_r , resulting in narrower bandwidth. However, increasing the resonator width would increase the IL along with the

overall size of filter. To operate on the sweet spot of this design tradeoff, microstrip resonator width is chosen to be 1–1.25 mm which results in the best input/output impedance matches.

2.3 Effects of gap width

A parametric study was carried out to understand the relationship between gap width (w_g) and filter performance parameters in terms of S_{21} and S_{11} . Gap width has a significant effect on S_{11} and distinct patterns can be observed as the gap width is changed from 0.01 to 0.8 mm and from 0.8 to 1.9 mm as shown in Fig. 10. RL improves as the gap is increased from 0.1 to 0.8 mm. After reaching a maximum value it starts to decrease as the gap width is increased further from 0.8 to 1.9 mm.

Design graphs relating RL and IL to the gap width are shown in Figs. 11 and 12. As the gap varies across the complete range of 0.1–1.9 mm, IL has its variation only in the range of 0.74–0.77 dB. Although the effect is negligible but interestingly it gives the best value for w_g that corresponds to maximum RL as well.

It was observed that the gap width has more impact on the parameters calculated from S_{11} (RL and input/output impedance) as compared to the parameters calculated from S_{21} (FBW, Q_e and IL). For front-end applications, which deal with the noisiest stage with a very weak input signal, these small gains cannot be ignored. Moreover, the value of gap width which yields the lowest IL also corresponds to the best value of RL; however FBW at that point is the widest. This creates a typical design tradeoff between IL and FBW, however as best IL is coupled with maximum RL, the situation is comparatively favored for choosing the gap width corresponding to the best IL.

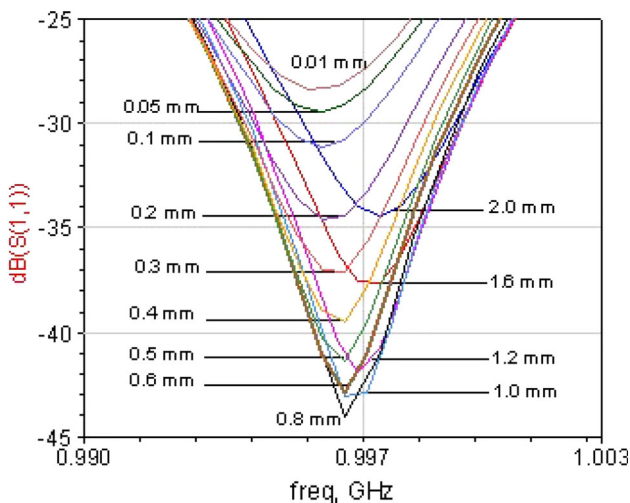


Fig. 10 Effect of gap width on S_{11}

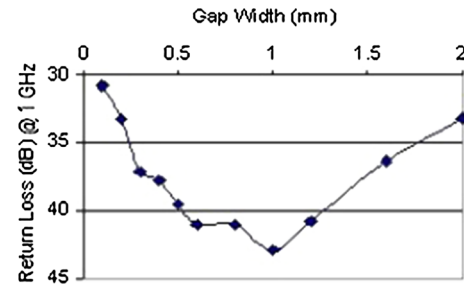


Fig. 11 Effect of gap width on RL

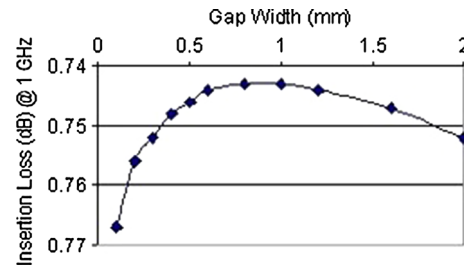


Fig. 12 Effect of gap width on IL

2.4 Frequency range of design

A filter was designed using proposed methodology with center frequency of 1 GHz. To determine the frequency range of the proposed design approach, filters were designed at center frequencies of 2 and 3 GHz as well. Desired center frequency could be easily tuned by adjusting the resonator length L corresponding to $\lambda/8$. Other geometrical dimensions including spacing between resonator arms, spacing between adjacent resonators, diameter of via ground hole and width of top pad were kept the same as for 1 GHz design. The sequential procedure of optimizing the response through adjusting the tap point height, the microstrip resonator width and the gap width was followed. For the center frequency of 2 GHz, 3 dB FBW of 2.99 % was achieved with good out of band rejection. Sharp skirting was observed as 20 dB FBW was only 9.9 %. RL was greater than 38 dB with an IL of 0.927 dB at pass band. For the filter designed at midband center frequency 3 GHz, the IL was maintained at less than 1.3 dB with 3 dB FBW of 2.464 %. Performance parameters of filters designed at center frequencies of 2 and 3 GHz are given in Table 1.

With considerable reduction in resonator length at higher frequencies, the length to width ratio of coupled microstrip lines becomes smaller and folding the resonator becomes difficult due to impracticable dimensions. The proposed design gains its compactness and size reduction through use of $\lambda/4$ resonators as compared to a $\lambda/2$ resonator of the conventional design. Therefore, the limitation

Table 1 Performance measures of the proposed filter designed using Rogers TMM 06 ($\epsilon_r = 6.08$) with mid band centre frequency 2 & 3 GHz

Freq	Area (mm ²)	FBW 3 dB (%)	FBW 20 dB (%)	RL (dB)	IL (dB)	Q_L	Q_U	Q_e
2 GHz	315.12	2.99	9.94	38.59	0.93	33.41	296.72	38.32
3 GHz	218.4	2.46	8.49	27.36	1.28	40.58	256.63	46.54



Fig. 13 Snap shot of filter on FR4 ($\epsilon_r = 4.34$)

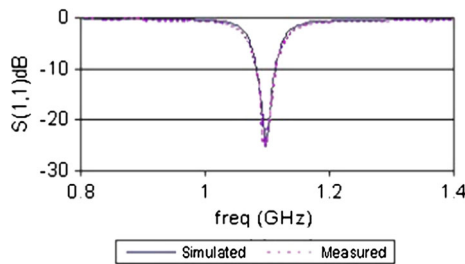


Fig. 14 Simulated & measured S_{11}

linked to folding the resonator at higher frequencies gets more pronounced in the proposed design. This design approach allows folding of resonators with practicable dimensions up to 3 GHz. A substrate with a higher dielectric constant and/or less height will further reduce the filter size and increase the design frequency range. Another limitation noticed at higher frequencies in the previous design was $IL \geq 3$ dB [7]. This limitation is addressed in this design through a capacitive gap and increased number of via holes. In order to demonstrate the robustness of this design methodology, filters were designed and tested on two more substrates: Rogers TMM 10 ($\epsilon_r = 9.2$), and Arlon CuClad 217 ($\epsilon_r = 2.17$). Equivalent design performance with similar design equations was observed across the complete frequency range of design methodology.

3 Fabrication and measurements

The biggest limitation of FR4 is its high loss tangent. High dissipation factor causes high insertion loss. The reason of using FR4 was to achieve a good agreement between measured and simulated results to validate the proposed design methodology. This would assure the desired performance when fabricated using suitable substrates. FR4

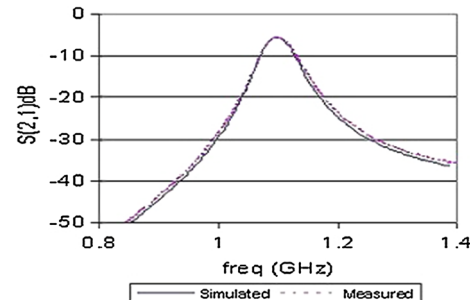


Fig. 15 Simulated & measured S_{21}

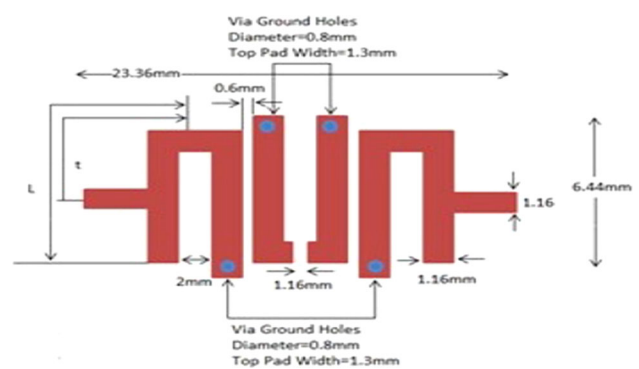


Fig. 16 Layout of the filter without 1st and 5th hairpin-line sections

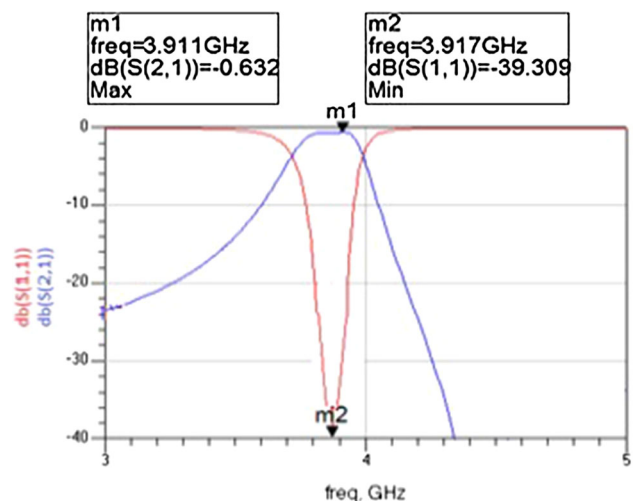


Fig. 17 S parameter result of the filter without 1st and 5th hairpin-line sections

Table 2 Comparison of proposed filter with conventional hairpinline filter using Rogers TMM 06 ($\epsilon_r = 6.08$)

Design Type	Area (mm ²)	FBW 3 dB (%)	FBW 20 dB (%)	RL (dB)	IL (dB)	Q_L	Q_U	Q_e
Conventional	699.56	25.03	33.95	26	0.31	3.99	108.59	4.58
Proposed	591.24	3.55	11.92	42.92	0.74	28.19	315.76	32.34

substrate with dielectric constant (ϵ_r) 4.34, thickness 1.5 mm and loss tangent ($\tan \delta$) of 0.02 was selected for the design. A fabricated filter is shown in Fig. 13.

Figures 14 and 15 shows comparison between measured/simulated RL and IL, respectively. Good agreement between measured/simulated results is observed with errors in IL, RL and center frequency of 1.6, 0.01 and 0.09 %.

It has been observed that first and fifth resonators in the proposed design do not resonate at the design frequency. By removing the first and fifth sections, an interesting result was observed. The frequency shifted from 2 to 4 GHz. The layout is shown in Fig. 16 and simulation results are shown in Fig. 17.

4 Conclusion

In this new, compact and improved filter design, a capacitive gap is used in addition to the via ground holes. This results in reduced resonator lengths corresponding to $\lambda/8$, causing a 15.5 % reduction in the overall size of the filter and better performance in terms of IL and RL. Mid-band frequency can be adjusted through resonator lengths. Less than 4 % FBW is achieved without increasing the gap between adjacent resonators, while the IL is maintained below 0.8 dB in the passband. Although 0.8 dB IL is not ideal however it is still less than 3 dB. Moreover the proposed design is simple and inexpensive which makes it a suitable option for many applications. A comparison between the conventional design and the proposed design is tabulated in Table 2. Higher Q_e is achieved with greater flexibility in the tap point location. This flexibility can be used to improve impedance matching, thereby improving the RL. Characteristic impedance of the hairpinline resonator has a more pronounced effect on the FBW, IL and RL as compared to the tap point height. Moreover, the FBW and IL can be improved by optimizing the gap width. Microstrip bandpass filters can be designed using the proposed methodology for frequencies up to 3GHz, beyond which folding the resonator becomes practically difficult. Design robustness is verified by designing the filter on three different substrates: Rogers TMM 06 ($\epsilon_r = 6.15$), Rogers TMM 10 ($\epsilon_r = 9.2$), and Arlon CuClad 217 ($\epsilon_r = 2.17$). This design approach is validated by designing and

fabricating a bandpass microstrip filter on FR4 substrate. Measured results show close agreement with the simulated results.

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