Design and Realisation of Miniaturisation of SIW-Based BPF

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Design and Realisation of Miniaturisation of Substrate Integrated Waveguide (SIW)-Based Bandpass Filter (BPF) at L-band Frequency

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Abstract:-This paper studies a miniaturization of a substrate-integrated waveguide (SIW) bandpass filter (BPF) using only a square cavity. This cavity is to be loaded with a multi-sector circular patch, where each sectored patch is connected to the bottom surface of the cavity through a shorting blind via. Each of the shorting-via loaded sectored- patches and the cavity's top and bottom surfaces form a resonator. Hence, multiple resonators can be housed in a single square cavity and then are fed properly to construct a multi-pole BPF. For easy integration with surrounding circuit components, itis to be considered by only the case where the cavity is fed with the coplanar waveguide (CPW) rather than the coaxial cable. The downshifts in the resonance frequency of the proposed resonator structure for the different number of sectors obtained from a complete cicuit patch are studied. BPFs constructed using one, two, and threesectored patches are designed and compared. A sample BPF using three sectored patches is fabricated and measured. As compared with the third-order BPF using three empty SIW cavities, the size reduction rate of the fabricated one is up to 98%. A good agreement is obtained between simulated results and those measured.

Keywords:- *Miniaturization; trisection bandpass filter (BPF), SIW*

I. INTRODUCTION

Nowadays, a various of emerging wireless communication systems is developing rapidly. One of the most important devices for wireless communication systems is a filter to minimize interference by passing a frequency band of interest. It is a device which serves to select and/or reject specific frequency channels. High-performance filtering is critical since spectral crowding increases the need for interference mitigation. Interference mitigation will necessitate out-of-band attenuation. The such out-of-band attenuation is able to be provided by bandpass filters (BPFs). In general, a waveguide is used for designing a BPF with respect to a high selectivity and Q-factor. Disadvantages of the waveguide BPF are the size of the filter which is bulky and costly as well.

More than a decade ago, a laminated waveguide (also termed the substrate integrated waveguide, SIW), which is composed of a substrate with via-hole rows emulating the waveguide's side walls, was proposed [1]. Since then, it become an issue which has been attractive largely many researchers. It offers a new structure in BPF design. The smaller size make bandpass SIW filter very suitable for some applications such as satellite communication and radar systems. In addition, it possess highly integration capability with other planar structures. The SIW-based BPF can be fabricated in a single layer or multilayer employingprinted circuit board (PCB) or low temperature cofired ceramic (LTCC) technology [2]. For the time being, trisection BPFs on the basis of SIW have been reported in [3] using the circular cavity and in [4–5] using the rectangular cavity. These filters employ three cavities as resonators to construct a trisection BPF, while input and output resonators are cross-coupled. Nevertheless, the SIW's working frequency is in respect to the physical size of the component. Therefore, one of the SIW's shortcomings is still its larger dimension than that of the planar counterparts (e.g., microstrip line).

For circuit-size reduction purpose in BPF design, there have been many efforts to miniaturize SIW-based BPFs. Miniaturization techniques can be conducted using a half of conventional SIW, so-called HMSIW, while still maintaining its characteristics [6–7]. The further reduction of HMSIW results in a quarter of conventional SIW named quarter-mode SIW (QMSIW) whereas reserving its original characteristics as well [8]. Both of these techniques are to reduce physical dimensions of SIW resonators. The sense of miniaturization is not only the size reduction but also the resonant frequency decrease. In the latter case, miniaturization is able to be achieved by loading the SIW by means of capacitive and inductive loading in order to make it to work below its cutoff frequency as exhibited in [9–10], respectively. In both of these cases, the SIW's size is still same as the conventional one, however, its resonant frequency is shifted downward from the fundamental mode frequency. In [11], another miniaturization process is proposed for which the SIW cavity (SIWC) consists of three metal layers and two substrate layers. The circular patch is located in a sandwiched middle metal layer so that results in a large loading capacitance between the circular patch and the top/bottom SIWC walls.

This paper is to study a miniaturization design of the substrate-integrated waveguide (SIW) multi-resonator bandpass filter (BPF) using only a square cavity on the basis of the proposed structure in [11].

II. RESONATOR DESIGN

Fig. 1(a) shows the proposed basic resonators to obtain a miniaturized SIW cavity bandpass filter (BPF) which consists of three 0.035-mmthick metal layers, two substrate layers, and a thick prepreg (PP) as shown in Fig. 1(b). The prepreg material is to be introduced for the binding purpose between the middle metal layer and the top substrate. Basically, the middle metal layer is a circular patch, which can be divided into some parts identically in order to obtain the desired filter order and each part is further connected to the ground by means of a shorting blind via as depicted in Figure 1(a). All metal layers are considered as copper, meanwhile substrate layers are made of Rogers RT/Duroid 5880 ($\epsilon_r = 2.2, \tan\delta = 0.0009$) with different thickness of whichthe top substrate has h_{top} = 0.254 mm and the bottom substrate has h_{bottom} = 1.58 mm. The SIW square-shaped cavity resonator occupies an area of 25 x 25 mm^2 with which a fundamental mode, TE101, exist around 5.67 GHz. These parameters and dimensions are to be implemented to study and design BPFs with the different order.

The resonance frequency of the resonator is primarily determined by the SIWC's loading capacitance, while the loading capacitance is mainly contributed by the region bounded by the middle patch and the top SIWC wall due to small thickness [11]. On the other hand, the inductance section is established by the shorting blind via. Fig exhibits that the proposed resonator structure results in the electric field uniformly on the surface of the middle metal. As a result, the resonance frequency of the proposed resonator can be greatly lowered from that of a standard patch-free SIWC resonator, and among these basic resonators, full circular patch resonator has the lowest resonance frequencyas can be seen in Fig. 3. However, lowering the dimension of the middle metal as a single resonator by sectioning the full circular patch into some sections will increase the resonance frequency of the resonator. It makes sense that lowering the dimension of the middle metal is to decrease loading capacitance, in turn, it will increase the resonance frequency of the resonator as described in [11].

Meanwhile, by modifying the location of the shorting blind via from the cavity center, the resonance frequency of the resonator will be increased so that for the fine tuning purpose, one

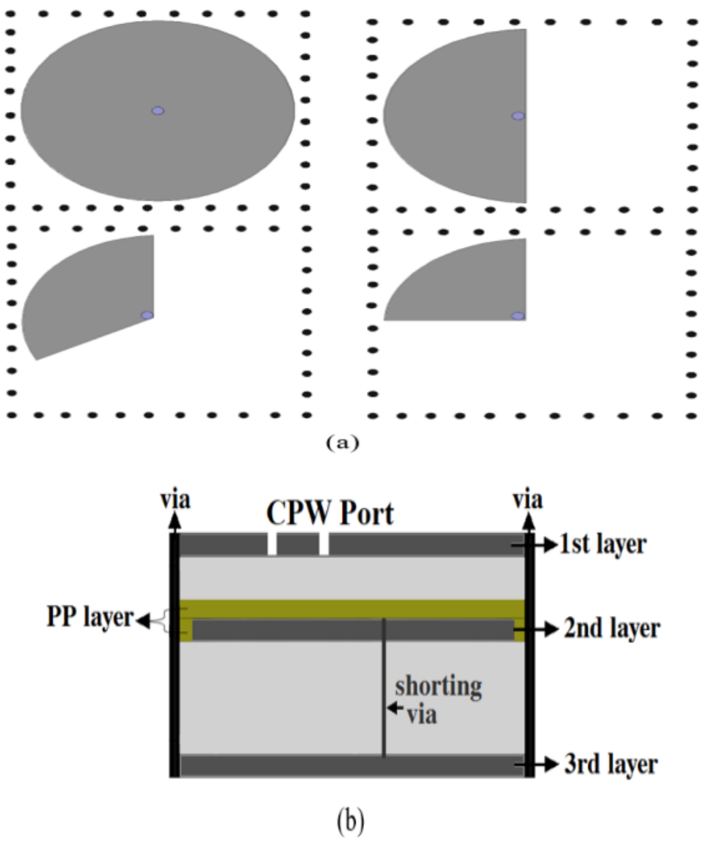

Fig. 1: Shape of middle metal layer dan resonator structure; (a) various shapes of middle metal layer; (b) cross-section of resonator structure

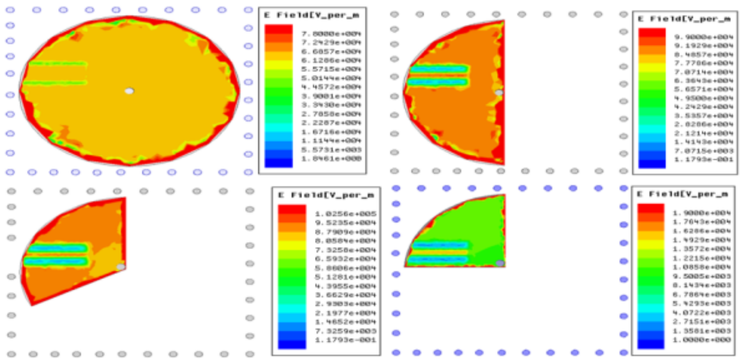

Fig. 2: E-field distribution inside the middle metal layer shown in Fig. 1 at the corresponding resonance frequency

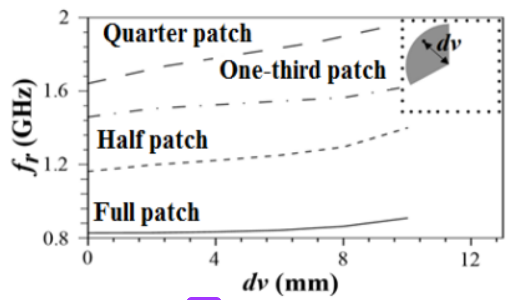

Fig. 3: The resulted resonance frequencies of the basic resonators at various positions of the shorting blind via from the cavity center can adjust shorting blind-via position. In addition, as depicted in Fig. 4, increasing the thickness of the bottomsubstrate will decrease the resonance frequency of the resonator. In contrast, increasing the thickness of the top substrate will also increase the resonance frequency. Shortly, lowering the thickness of the top substrate and increasing the thickness of the bottom substrate will increase the miniaturization factor. However, these means will be restricted by available materials and allowable fabrication limits.

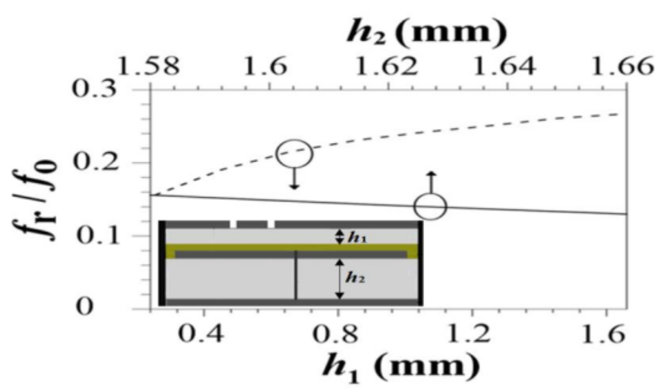

Fig. 4: Effect of the substrate thickness on the resonance frequency

III. FILTER DESIGN

A. *One-Pole Bandpass Filter*

The proposed one-pole filter along with its dimensions is obtained by employing the full patch shape of the middle metal as exhibited in Fig. 5(a). In order to design a such filter, the dimensions of the square SIW cavity can be obtained as described in [11]. The unloaded Q factor, Q_u, of the proposed structure is 221. The structure is excited by using coplanar waveguide (CPW) structure. Therefore, the required external quality factor (Q_e) can be controlled by varying the inner-strip length of the CPW and maintaining the width of the CPW and other dimensions constantly, as depicted in Fig. 5(b). The Q_e value can be extracted by the singly-loaded expression [12]

$$Q_e = f_0/\Delta f_{\pm 90^0} \quad (1)$$

where f_0 denotes the simulated resonance frequency, while $\Delta f_{\pm 90^0}$ represents the frequency difference between phase-shift $+90^0$ and phase-shift -90^0 occuring in the S_{11} phase response.

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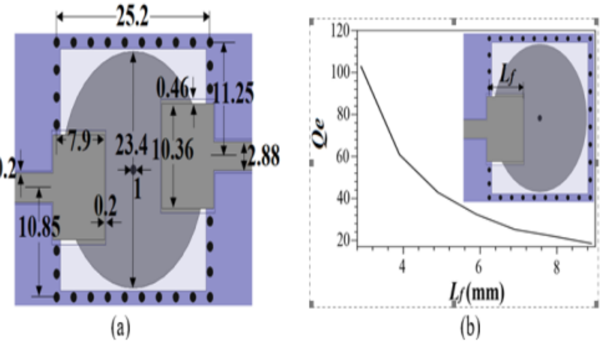

Fig. 5: One-pole bandpass filter; (a) the proposed structure, (b) External quality factor (Q_e) of the one-pole BPF vs. inner-strip length

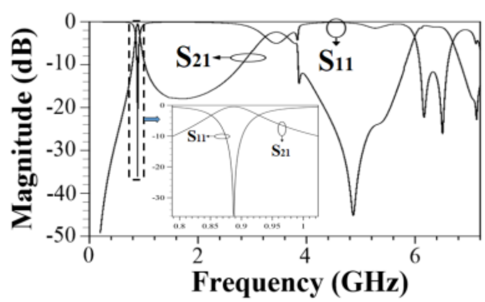

Fig. 6: Simulated S-parameter response of the one-pole bandpass filter

By using Q_e value of 22 which is realized by the inner-strip length of 7.9 mm, Fig. 6 exhibits the S-parameter response of the one-pole BPF. As can be seen that the center frequency is 0.888 GHz. The return loss is better than 30 dB, whereas the insertion loss is 0.45 dB. The 3-dB fractional bandwidth is 9.08%. The entire size of the proposed one-pole filter is 0.01 λ_d^2 where λ_d is the wavelength in the medium of the dielectric material at the operating frequency.

B. Two-pole Bandpass Filter

Fig.7 shows the proposed structure of the two-pole BPF using the proposed half patch resonator with its associated dimensions. The structure implements magnetic coupling among resonators. This coupling is caused by the induced currents on the shorting blind vias. Actually, electric couplingexists between adjacent edges of the half patches. However, electric coupling between them is very weak. This is because the top substrate is so thin that the electric field is strongly bound in the region between each half patch and the top SIWCwall. As a result, the coupling among them is the dominated magnetic coupling, and is established by the shorting blind via.

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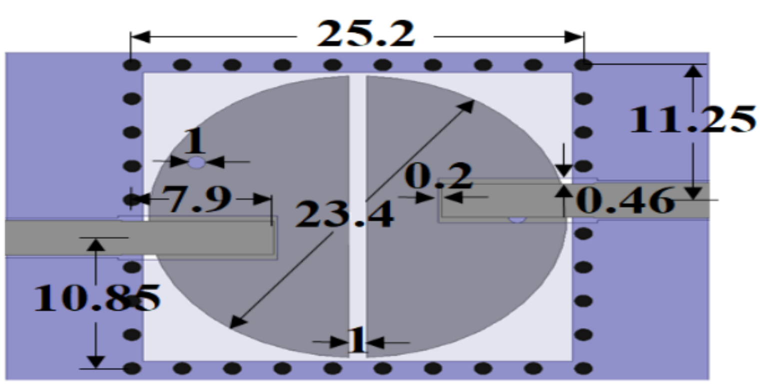

Fig. 7: The proposed structure of the two-pole BPF

In order to design a such BPF, once the resonator is dimensioned, one can follow the conventional procedure by determining the design values of the low-pass Butterworth/Chebyshev prototype response on the basis of the specification target. On the basis of the coupled-resonator filters design, the general coupling matrix is implemented to represent the employed coupling topology. Therefore, the relationship between coupling coefficients or external quality factors and physical dimensions of the coupled resonators must be established. In this case, spacing of the two shorting blind vias represents the physical dimensions, and is employed to derive the proper coupling. The simulation is carried out by providing weak coupling for the input/output. The values of the simulated coupling coefficient are extracted by the well-known expression [12]

$$k = \pm \{[(f_1)^2 - (f_2)^2] / [(f_1)^2 + (f_2)^2]\} \quad (2)$$

where f_1 and f_2 represent the high and low resonant frequencies, respectively, and k denotes the value of coupling coefficient between two resonators in terms of the distance between two shorting blind vias. In addition, $k > 0$ and $k < 0$ indicate magnetic and electric couplings, respectively.

As with the one-pole BPF, the required external quality factor (Q_e) can also be controlled by the inner-strip length of the CPW at the input/output, and can be extracted by using (1). In order to turn out the relationship between the physical dimensions and the required theoretical design values, the coupling coefficient as a function of the spacing between the shorting blind vias, and the external quality factor value as a function of the inner-strip length is plotted in Fig.8(a) and (b), respectively.

As shown in Fig.8, the strong coupling can be achieved when the distance between two shorting blind vias are in the close proximity. In other hand, the weak coupling can be obtained by making larger distance between them. Roughly speaking, it is possible to control coupling coefficient by adjusting distance between two blind vias. The horizontal distance between the two shorting blind vias can be employed for coarse adjustment of the coupling coefficient, meanwhile the vertical distance can be used for fine tuning purpose.

With the coupling coefficient of 0.0193 between two resonators and external quality factor of 21.47, the proposed two-pole BPF has S-parameter response as shown in Fig.9. The center frequency is 1.258 GHz. The return loss is better than 25 dB, whereas in-band insertion loss is 2.1 dB. The 3-dB fractional bandwidth is 2.6%. The proposed two-pole BPF provides an out-of-band rejection better than 20 dB and 30 dB up to 4.7721 and 4.7698 GHz, respectively. The unloaded Q factor,Q_u,of the structure is 214. The overall size of the proposed two-pole BPF, excluding CPW feed lines, is 0.024 $(\lambda_d)^2$.

C. Three-pole Bandpass Filter

Fig. 10 exhibits the structure and its associated dimensions of the proposed SIWC three-pole BPF. Such filter is well-known as a trisection BPF using one-third patch as the basic resonator. The proposed trisection BPF is a central reflection-type resonator as reported in [13].

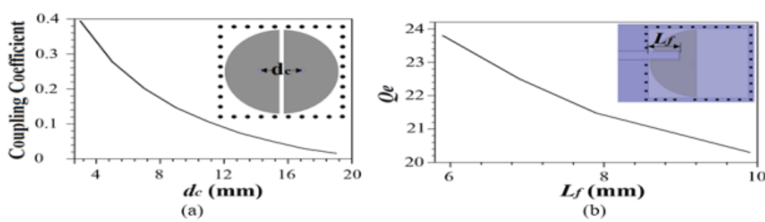

Fig. 8. (a) Coupling coefficient vs. distance between two blind vias in allignment, (b) External quality factor vs. inner-strip length

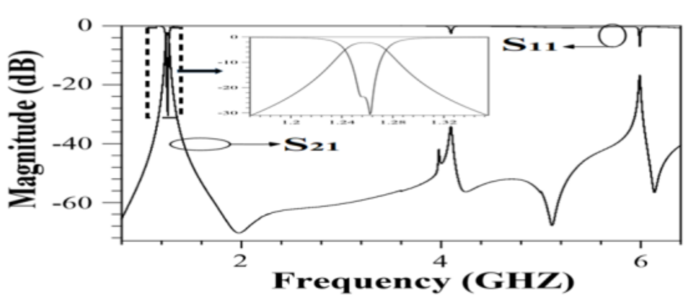

Fig. 9: Simulated S-parameter response of the two-pole BPF

A such structure introduces a cross coupling between the first resonator in terms of the input resonator and the third resonator in terms of the output resonator. Therefore, there are two possible paths from input to output. As a result, the structure introduces one TZ in its response. The associated TZ occurs near the upper edge of the passband. This TZ is to improve the selectivity performance of the filter. Its asymmetrical response benefits some applications requiring higher selectivity only on one side of the passband. The derived TZ is able to be related to coupling coefficients as follows [13]

$$f_z = f_0 + (f_0 / 2)(k_{12})^2/k_{13} \quad (3)$$

where f_z and f_0 represent the frequency of transmission zero and the center frequency, respectively, whereas k_{12} and k_{13} indicate coupling coefficient of resonator 1 and 2 and cross coupling, respectively.

Again, in order to design a such filter, one can follow procedures which are described in the sub section B. Fig. 11(a) shows the relationship between the coupling coefficient and the physical dimensions of the filter in terms of the distance between two blind vias of the resonators. The larger coupling coefficient values can be obtained by shortening the distance between two blind vias with respect to the direction perpendicular to the edge of the patch. Fig. 11(b) describes the relationship between the external quality factor for resonator 1 and the inner-strip length of CPW, while the other dimensions of the CPW are maintained to be same as Fig. 10. Eventually, the tuning processes must be carried out to obtain an optimal response.

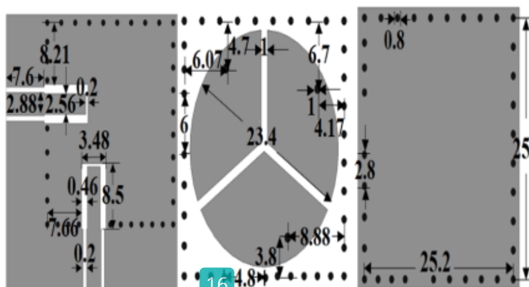

Fig. 10: The proposed structure of the trisection SIWC BPF; (a) top metal layer, (b) middle metal layer, (c) bottom metal layer

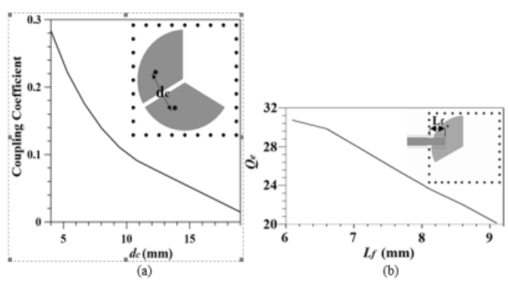

Fig. 11. (a)Coupling coefficient vs. distance between the adjacent shorting vias of the resonators, (b) External quality factor vs. inner-strip length

The simulated S-parameter response is shown in Fig. 13. It can be seen that the trisection BPF has the higher selectivity performance than one-pole and two-pole BPFs with which a transmission zero appears closely to the passband at upper side of stopband due to cross coupling between the first and the third resonators. The unloaded Q factor, Q_u, of the proposed structure is 216. Hence, for the fabrication purpose, our concern is only for the trisection BPF.

D. Fabrication Results

The designed trisection BPF is realized by using three 0.035-mmthick metal layers and two substrate layers. The top and bottom Rogers RT/Duroid 5880 substrate layers with dielectric constant ε_r = 2.2 and loss tangent tanδ = 0.0009 have the thicknesses of 0.254 and 1.58 mm, respectively. A 0.08-mmthick prepreg (PP) layer with dielectric constant 4 and loss tangent 0.013 is placed between the middle metal layer andthe top substrate for the binding purpose. The fabricated structure is shown in Fig. 10 and Fig. 12. The inset of Fig. 13 shows the measured and simulated narrowband S-parameter responses of the proposed SIWC trisection BPF, whose photos are given in Fig. 14. The measured results are in good agreement with simulated ones.

As mentioned in the sub section C, there is a transmission zero nearby upper edge of the passband, and it can be seen clearly that the proposed structure reflects its type as reported in [13]. The such transmission zero will sharpen one side of the passband skirt. The measured and the simulated is and minimum insertion loss of the proposed BPF are 2.9 dB and 1.95 dB,respectively. Meanwhile the measured and the simulated 3-dB fractional bandwidth (FBW) are

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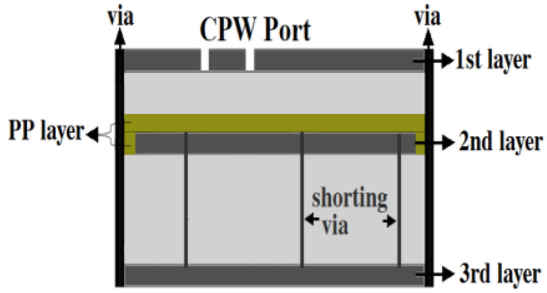

Fig. 12: Cross-section of the fabricated trisection BPF

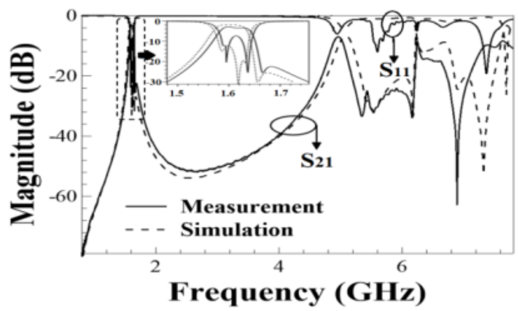

Fig. 13: The measured and simulated S-parameter responses of trisection BPF

3.97% and 4% with the center frequencies of 1.612 and 1.6 GHz, respectively. The measured and the simulated transmission zero are 1.66 and 1.65 GHz. These values approach the value obtained by using (3) as large as 1.648 GHz.

From Fig. 13, the measured high-end stopband BW is 2.69 and 3.04 GHz under the criterion of insertion loss greater than 30 and 20 dB, respectively, corresponding to an upper stopband fractional bandwidth (FBW) of 166.873% and 190%. In order to exhibit our structure superiorities, Table 1 provides comparison between our work and related SIWC trisection BPFs. In this table, the datum with a tilde sign in front denotes that such a datum is not given in the reference paper, but is estimated by us using curves or other relevant data available. Clearly, our circuit design has a much better area efficiency than those of the others. In particular, the occupied circuit area of our proposed BPF is 0.04 λ_d^2, which is much smaller than the area of 2 λ_d^2 required by a regular patch-free three-cavity SIWC trisection BPF, that is, miniaturization factor of 98% can be achieved. In addition, our circuit design yields the largest upper stopband fractional bandwidth (FBW) with the criterion of $S_{21} \leq -20$dB.

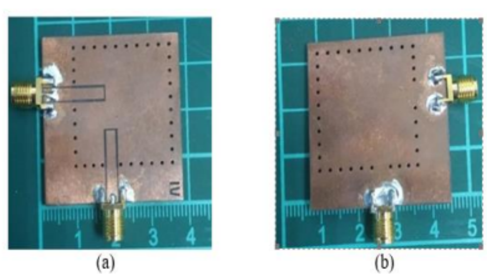
Fig. 14: Fabricated trisection BPF; (a) top view, (b) bottom view

	Circuit size (λ_d^2)	Insertion loss (dB)	FBW (%)	Upper-stopband FBW (%) ($S_{21} \leqq -20$ dB)
[2]	0.126	2.4	~6	NA
[3]	~2.6	2.9	1.5	NA
[4]	0.511	1.15	5.8	~20
[5]	~0.98	2.46	1.95	>43
This work	0.04	2.9	3.97	190

IV. CONCLUSION

This paper has already studied and shown a miniaturization design of the bandpass filters (BPFs) on the basis of substrate integrated waveguide technology. The proposed structure provides a miniaturization factor of 98% in trisection BPF which is the highest record as long as the author knowledge. The excitation structure using CPW provides the advantage of being easy for integration with other planar circuit. The circuit design and performance are verified by measurement, and it shows a good agreement with the simulation. The fabricated BPF can provide the center frequency of 1,6 GHz which is at L-band frequency. Hence, this device is capable of being utilized for radar application at L-band .The slight discrepancy between the measurement insertion loss and the simulation insertion loss comes from the effect of practical errors due to fabrication as well as SMA connector loss.

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