EXPANDABLE BANDWIDTH AND IMPROVED GROUP DELAY FILTERS USING THE PERTURBED SIW DUAL-MODE RESONATORS

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Abstract

An alternative method for realizing the expandable bandwidth microwave and millimetre-wave bandpass filters is presented in this paper. A substrate integrated waveguide (SIW) dual-mode circular resonator with a small perturbation is proposed to adjust a transmission zero (TZ) for widening the bandwidth and improving the selectivity by properly changing the location of an additional via-hole placed inside the cavity. X-band and Q-band bandpass filters have been designed, fabricated and measured. The experimental results show good agreement with simulation results. The proposed planar filter has the merits of simple design process, expanded bandwidth, improved group delay and compact size.

Keywords: Dual-mode, Filter, Group delay, Perturbation, SIW, Transmission zero, Wideband.
1. Introduction
The modern microwave and mm-wave communication systems have an increasing requirement of developing wideband filters. The dual-mode circular cavity (DMCC), in particular, is gaining popularity for high performance bandpass filters (BPFs) due to its higher selectivity, wider bandwidth and less mass than conventional single-mode direct coupled filters [1]. Usually, the conventional DMCCs require tuning-screws to achieve the desired coupling, which leads to a generally time-consuming tuning process. In order to overcome this drawback, a new ridged circular waveguide section has been proposed in [2]. In [3], the authors presented a flexible method for designing waveguide BPFs. It offers wider bandwidths by changing the angle of a perturbing notch etched on the sidewall of the DMCC as shown in Fig. 1(a). However, these designs have high production cost and are directly incompatible with planar integrated circuits.

Moreover, SIW technique has been widely implemented for designing planar BPFs [4, 5]. As shown in Fig. 1(b), millimetre-wave dual-mode SIW filters based on circular or elliptic SIW cavities with complementary features have been investigated to design a high selectivity diplexer [6]. The advantages of these filters are low cost, and high integration capability. However, their bandwidth, insertion loss and flexibility need to be enhanced due to the limitation of the geometry as well as the unloaded quality factor of SIW cavity [7-11].

Recently, the Q-band planar SIW BPFs with high selectivity have been proposed by T.V. Duong et al. [7]. Therein, the authors introduced a technique which involves a small additional via-hole inside the DMCC with a perturbing angle which can flexibly adjust TZs to improve the selectivity of the filter. In [12-15], not only the filter is designed with a compact structure, but it also allows for the adjustment of the passband in the first band, the second band, or both bands.

Firstly, this paper focuses on detailed results of the electromagnetic analysis simulation of DMCC filter with expandable bandwidth and the designs of SIW BPFs operating at X-band, and Q-band. Secondly, experimental results are presented in Section III, along with the comparison to previous works given in the discussion. Finally, conclusions are presented in Section IV.
2. Filter Design with Expandable Bandwidth

2.1. The proposed structure of the perturbed SIW filters

The layout of a planar single-DMCC filter (BPF1) is shown in Fig. 2(a). It consists of transmission lines as input/output ports, inductive windows, and a perturbed SIW cavity. The CPW (coplanar waveguide)-variant transitions at the input/output port has the characteristic impedance of 50Ω. It works as transmission lines which excites quasi-TEM wave modes. The inductive windows with different sizes works as admittance inverters that can control the couplings between the input/output ports and DMCC. A perturbed SIW DMCC is designed to support two orthogonal resonant modes corresponding to the given microwave/millimeter wave range. In order to analyze the proposed method in filter design, the first investigated frequency range is selected as Q-band (33-50 GHz). The widths of induction windows are denoted by w1, w2, w3, and w4. DMCC holes are made from a series of holes with diameter d, and the distance between adjacent holes should be calculated to ensure the protection from radiation leakage [10]. The dv diameter is located in position Lν relative to the center of the cavity and created with output port one angle β while the angle created by two input/output ports is α. The equivalent radius r of DMCC is evaluated by [10]

\[
f_{\text{TM}_{nm,l}} = \frac{c}{2\pi \sqrt{\varepsilon h}} \sqrt{\left(\frac{p_{nm}}{r}\right)^2 + \left(\frac{l \pi}{h}\right)^2}
\]

where \(f_{\text{res}}\) is the resonant frequency; \(m, n,\) and \(l\) are indexes of \(\text{TM}_{nm,l}\) modes; \(c\) is the speed of light in free space; and \(\varepsilon\) and \(h\) are the dielectric constant and thickness of the substrate, respectively. The value of \(p_{nm}\) is 3.832 for the TM110 mode. In order to analyse the designed filter, an initial value of \(r=3.25\) mm is used. This value can be adjusted to analyse the bandwidth of the BPF. The angle α produces a weak coupling between the input and the output. Thus, a TZ is generated at upper stopband [6]. In our design, the angle α is selected as 1050. As the result, TZ located at \(f_{2a}\) in Fig. 3 can enhance the rejection level at the stopband of the proposed SIW filter. In the case of BPF1, all indicators are poor. The cavity has two orthogonal wave modes TM110, thus, it will create an initial transmission zero \(f_{2a}\) at the right side of the passband (see Fig. 3). If there is a perturbing via-hole, we will use it to move the transmission zero \(f_{2a}\) to the left of the filter passband.

Figure 2(b) presents the layout of BPF2 which is formed by merging two DMCCs. A segment of SIW is adopted to connect two DMCCs in BPF2. DMCC-II creates one transmission zero at the right of the passband. However, DMCC-I uses the perturbing via-hole to move it to the left of the passband. The inductive coupling window has the widths of \(w_0, w_5,\) and \(w_6,\) the distance between the two cavities is \(L_0.\) The radius of the first cavity \(r_1\) is larger than that of the second cavity \(r_2.\) Thus, the resonant frequency of the first cavity is smaller than that of the second cavity.

In practical applications, cavity resonators are often modified by changing their shapes, or by applying small pieces of dielectric or metallic materials. In certain instances, the effect of perturbations on the cavity performance can be precisely calculated, but approximations generally have to be made. This is due to that fact that the precise calculation of cavity disturbance can only be performed in a few specific cases, with precise boundary conditions. For simplicity, if the angle of β is set at 90 degrees, then sin β = 1 and cos β = 0. In other
circumstances, exact calculations are extremely difficult, and electromagnetic field simulation software tools (such as HFSS, CST) are required for the analysis. A useful technique is the perturbation method [11]. It assumes that the actual fields within DMCC with a small perturbation are not greatly different from those of the unperturbed one. In particular, placing an additional via-hole within DMCC as shown in Fig. 2 can be considered as a change in the shape of the cavity. These small changes can also be treated by the perturbation technique. In this section, we will derive expressions for the approximate change in resonant frequency when a DMCC is perturbed by small changes in its shape.

![Fig. 2. The proposed Q-band DMCC SIW filters.](image)

![Fig. 3. Simulated response of the proposed filter with/without the coupling via-hole at the angle β varies from 15° to 75°.](image)

The performance of these filters is analysed by using EM Simulator HFSS.

Let $\mathbf{E}_0, \mathbf{H}_0, \omega_0$ be the fields and resonant frequency of the original DMCC that have volume $V_0$ and surface area $S_0$, and let $\mathbf{E}, \mathbf{H}, \omega$ be the fields and resonant frequency of the perturbed DMCC of the volume $V$ and surface area $S$. As a result, the perturbing via-hole has a volume of $\Delta V=V_0-V=\pi d^2 h/4$. We assume $\Delta S$ is small, and approximate $\mathbf{E}, \mathbf{H}$ to be equivalent with the unperturbed values $\mathbf{E}_0, \mathbf{H}_0$. After several modifications from Maxwell's curl equations, the fractional change in resonant frequency is extracted as [11]:

\[
\frac{\omega-\omega_0}{\omega_0} \approx \frac{\int_{V_0} (\mu |\mathbf{H}_0|^2 - \epsilon |\mathbf{E}_0|^2) \, dV}{\int_{V_0} (\mu |\mathbf{H}_0|^2 + \epsilon |\mathbf{E}_0|^2) \, dV}
\]

(2)

Equation (2) can be written in terms of stored energies as:
where $\omega_m + \omega_e$ is the total stored energy in the DMCC, $\Delta W_m$, $\Delta W_e$ are the changes in the stored magnetic energy and electric energy, respectively. This result shows that the resonant frequency may increase or decrease; depending on the location of the perturbation and whether the cavity volume is increased or decreased.

For the perturbed DMCC, in term of the radius $r$ and the height $h$ investigated in this section, we can assume that a small perturbing via-hole with volume $\Delta V$ etches at an initial position $(l_i, \beta)$ as shown in Figure 2(a). According to Reference [10], the fields with the reference amplitude of $E_0$ for each of the $\text{TM}_{mn}$ modes in the circular cavity; the propagation constant $k$ can be simplified in a cylindrical coordinate system $(\phi, \rho, z)$ as:

$$
\begin{align*}
E_z &= E_0 \frac{\text{p}_{mn}}{r} \cos(n\phi) \\
H_\rho &= -j\omega \mu \frac{r}{\text{p}_{mn}} \frac{1}{\rho} \frac{\text{p}_{mn}}{r} \sin(n\phi) \\
H_\phi &= -j\omega \mu \frac{r}{\text{p}_{mn}} \frac{1}{\rho} \frac{\text{p}_{mn}}{r} \cos(n\phi) \\
H_z &= E_\phi = E_\rho = 0
\end{align*}
$$

These formulas show that the fields will be affected by both the different angles $\phi$ and the radii $\rho$. In other words, the electric and magnetic fields of the perturbed DMCC will also be affected by the location of the perturbing via-hole as illustrated in Fig. 4.

![Fig. 4. Electric field distribution of two orthogonal modes TM110-like in DMCC with $\beta=30^\circ$ at (a) fp_1b and (b) fp_2b.](image)

If the via-hole is small, we can assume that the fields of $\text{TM}_{110}$ mode are constant over the cross section of the via-hole and Eq. (4) can be expressed at the $\phi=90^\circ$ as:

$$
\begin{align*}
H_\rho &= -j\omega \mu \frac{r}{\text{p}_{11}} \frac{1}{\rho} \frac{\text{p}_{11}}{r} \sin(90^\circ) = -j\omega \mu \frac{E_0 \rho}{4} \\
H_z &= H_\phi = E_z = E_\phi = E_\rho = 0
\end{align*}
$$

Consequently, the numerator of Eq. (2) can be evaluated as

$$
\int_\Delta r (\mu \mu_0 |H_\rho|^2 - \epsilon_0 |E_\rho|^2) dV = \mu \int_\Delta |E_\rho|^2 dV = -\mu(\omega\mu)^2 \frac{k_0^2}{16} \Delta V
$$

and the denominator of Eq. (2) is the total energy stored in the DMCC and extracted as:
\[
\int_{V_0} (\mu |\bar{H}_0|^2 + \varepsilon |\bar{E}_0|^2) dV = \frac{\mu (\omega \mu)^2 r_1^2 E_0^2 \pi h}{8 (p_{11})^2} \left[ 1 - \frac{1}{(p_{11})^2} \right] f_f(p_{11}) = \frac{\mu (\omega \mu)^2 r_1^2 E_0^2}{32 (p_{11})^2} \left[ (p_{11})^2 - 1 \right] V_0 \]

where \( V_0 = \pi r_1^2 h \) is the volume of the unperturbed DMCC. From Eq. (2), we can extract the fractional change in resonant frequency from the unperturbed DMCC with \( \rho = \frac{d}{p_{11}} = 0.8 \) and \( p_{11} = 3.832 \) as:

\[
\frac{\omega - \omega_0}{\omega_0} = \frac{2 \rho^2 (p_{11})^2}{r^2 [ (p_{11})^2 - 1]} \frac{\Delta V}{V_0} = -1.37 \frac{\Delta V}{V_0} \quad (8)
\]

It can be seen that the negative sign indicates a downward trend of the fractional change in resonant frequency of the unperturbed DMCC and this trend will be proportional to the via-hole volume, while the electric fields are affected by the angle change. Those effects will be utilized to improve the bandwidth performance of DMCC BPFs with the support from EM simulator HFSS.

When the DMCC is operated without a perturbing via-hole, two orthogonal modes are very close. This space is slightly expanded using asymmetric inductive coupling windows, hence an additional cross coupling between two modes can be obtained and another TZ at \( f_2 \) is generated in the upper side of passband to improve the steep slope of the BPF [10]. However, the BPF is still limited by bandwidth. In order to expand the BW, a small additional perturbing via-hole is placed inside the DMCC at an initial angle \( \beta \) of 30\(^\circ\). The perturbing via-hole can increase the frequency distance of two orthogonal modes, thus achieve a BW wider angle of 50%.

Figure 4 shows the simulated electrical field distributions at the locations of two separated resonant peaks: \( f_{p1b} \) and \( f_{p2b} \) (as can be seen from Fig. 3), where two orthogonal resonating modes TM\(_{110}\) can be clearly observed and the coupling is improved by adjusting the location of the additional via-hole. In Fig. 3, the location of the second TZ has been changed when the angle \( \beta \) varies from 75° to 15°. As angle \( \beta \) decreases, it shifts to lower frequency. When \( \beta \) is chosen at 15° or 30°, the second TZ is placed at a lower stopband. When \( \beta \) is larger than 75°, the second TZ is placed at the upper stopband.

It is worth noting that the first TZ, which is controlled by the angle \( \alpha \), is always located at the upper stopband when the perturbation angle varies from 15° to 75°. In other words, two TZs can be individually adjusted. Thus, it offers more flexibility while reducing the complexity of the design. As a result, the filter bandwidth is expanded and the selectivity is improved as well.

### 2.2. Improving group delay in perturbed SIW filters

In this section, we will improve the group delay peak and the unloaded \( Q_u \) factor of the proposed SIW filters while operating at X-band frequency range. The layout of the proposed X-band DMCC SIW filters is shown in Fig. 5.
Figure 6 shows the dependencies of the frequency locations of the 2nd-TZ and group delay peak on the angle $\beta$ which varies from 20° to 90°. It can be seen that as the angle $\beta$ increases and $L_v$ decreases, the 2nd-TZ shifts to a higher frequency and the bandwidth is also subsequently changed. Particularly, at $L_v$=5 mm, as the angle $\beta$ moves out of the range (50°÷60 degree), the 2nd-TZ makes the bandwidth expand to a further range (9.88 - 10.65 GHz) or the FBW is larger than 8%. Moreover, by properly choosing the perturbing via-hole location and the value of $\beta$=20°, 30°, 40°, 70° or 80°, the change of the group delay is also achieved both in term of frequency and amplitude.

Figure 7 shows the apparent dependence of the 2nd-TZ and group delay peak value on the value of the perturbing angle on the left-side of the passband, while the frequency location of the group delay peak is nearly unchanged. Moreover, in the range of (50°, 55°), the group delay peak quickly diminishes as the 2nd-TZs reach the filter’s slopes.

In general, by cascading $n$ DMCCs, we can obtain $2n$-order filters and a maximum of $2n$ controllable TZs. Figures 2(b) and 5(b) show the 4th-order double-cavity filter (i.e., BPF2 and BPF4) operating at Q-band and X-band, respectively. Moreover, Fig. 3 also shows that it has no significant effect of the additional via-hole position on the second TZ at the upper slope of the passband when the angle $\beta$ is larger than 75°. Therefore, the double-cavity proposed filter needs only one via-hole placed at an initial angle $\beta$=30° inside the cavity $I$. 

**Fig. 5.** The proposed X-band DMCC SIW filters.

**Fig. 6.** The influence of perturbing via-hole location.

**Fig. 7.** The influence group delay peak magnitude at $L_v$= 4 mm.
Figure 8 shows the dependence of the group delay peak on the angles $\beta$ in BPF3 and BPF4, respectively. We can see that, the larger the perturbing angle, the higher the group delay will peak, especially in BPF4 because the corresponding TZ is close to the transitional slope. Meanwhile, the unloaded Q factors of the first resonance mode, whose pole frequencies are close to the peaks of the group delay, still remained at high levels (~ 500) in both BPF3 and BPF4.

Finally, the four broadband multi-mode SIW BPFs with different expandable bandwidth can be optimized by using software HFSS to satisfy the following specifications:

- Wider FBW of passband: > 5%
- Insertion loss: > -1.2 dB
- Return loss: < -20 dB
- Rejection: > 25 dB in both the stopbands
- Asymmetrical frequency responses with the individually controlled TZs located on both sides of the passband.

The geometrical parameters of four optimized SIW BPF1, BPF2, BPF3, and BPF4 are listed in Table 1, respectively.

<table>
<thead>
<tr>
<th>Sym.</th>
<th>Unit</th>
<th>Q-band BPF1</th>
<th>Q-band BPF2</th>
<th>X-band BPF3</th>
<th>X-band BPF4</th>
</tr>
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<tbody>
<tr>
<td>$w_{30}$</td>
<td>mm</td>
<td>0.775</td>
<td>0.775</td>
<td>1.55</td>
<td>1.55</td>
</tr>
<tr>
<td>$r_1$</td>
<td>mm</td>
<td>3.28</td>
<td>3.27</td>
<td>12.45</td>
<td>12.40</td>
</tr>
<tr>
<td>$r_2$</td>
<td>mm</td>
<td>--</td>
<td>3.03</td>
<td>--</td>
<td>11.50</td>
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<tr>
<td>$s$</td>
<td>mm</td>
<td>0.70</td>
<td>0.70</td>
<td>0.80</td>
<td>0.80</td>
</tr>
<tr>
<td>$d$</td>
<td>mm</td>
<td>0.40</td>
<td>0.40</td>
<td>0.40</td>
<td>0.40</td>
</tr>
<tr>
<td>$h$</td>
<td>mm</td>
<td>0.254</td>
<td>0.254</td>
<td>0.508</td>
<td>0.508</td>
</tr>
<tr>
<td>$w_1$</td>
<td>mm</td>
<td>2.00</td>
<td>2.30</td>
<td>6.00</td>
<td>8.00</td>
</tr>
</tbody>
</table>
3. Fabrication and Measurement

The single-cavity and the dual-cavity bandpass SIW filters have been fabricated using the standard single layer PCB process on the Rogers 5880 (\(\varepsilon=2.2\), tan\(\delta=0.001\)). BPF3 and BPF4 have the thickness of 0.254 mm while BPF1 and BPF2 have a thickness of 0.508 mm. Figure 9 shows the photos of four fabricated filters whose parameters are specified in Table 1. Fabricated BPFs are measured by using the vector network analyser (VNA) Agilent N5245A and Anritsu device with the inherent insertion loss of (0.4 \(\div\) 0.7) dB within the range of (10 \(\div\) 50) GHz, respectively.

<table>
<thead>
<tr>
<th>w2</th>
<th>mm</th>
<th>2.60</th>
<th>2.89</th>
<th>8.25</th>
<th>9.00</th>
</tr>
</thead>
<tbody>
<tr>
<td>w3</td>
<td>mm</td>
<td>2.00</td>
<td>1.80</td>
<td>5.00</td>
<td>6.00</td>
</tr>
<tr>
<td>w4</td>
<td>mm</td>
<td>3.16</td>
<td>2.82</td>
<td>7.85</td>
<td>8.17</td>
</tr>
<tr>
<td>w5</td>
<td>mm</td>
<td>--</td>
<td>2.66</td>
<td>--</td>
<td>10.63</td>
</tr>
<tr>
<td>w6</td>
<td>mm</td>
<td>--</td>
<td>2.68</td>
<td>--</td>
<td>--</td>
</tr>
<tr>
<td>w7</td>
<td>mm</td>
<td>--</td>
<td>4.00</td>
<td>--</td>
<td>--</td>
</tr>
<tr>
<td>L0</td>
<td>mm</td>
<td>--</td>
<td>7.00</td>
<td>--</td>
<td>28.00</td>
</tr>
<tr>
<td>d1</td>
<td>mm</td>
<td>0.80</td>
<td>0.50</td>
<td>0.40</td>
<td>0.65</td>
</tr>
<tr>
<td>L1</td>
<td>mm</td>
<td>2.34</td>
<td>2.00</td>
<td>3.80</td>
<td>6.80</td>
</tr>
<tr>
<td>(\alpha)</td>
<td>degree</td>
<td>106</td>
<td>110</td>
<td>114</td>
<td>109</td>
</tr>
<tr>
<td>(\beta)</td>
<td>degree</td>
<td>30</td>
<td>32</td>
<td>30</td>
<td>40</td>
</tr>
</tbody>
</table>

(a) BPF 1  (b) BPF 2  (c) BPF 3  (d) BPF 4

Fig. 9. Photographs of the manufactured filters.
The measured frequency response of BPF1 is shown in Fig. 10. A high level of agreement is obtained between the simulated and measured results. The first TZ is at 45.5 GHz, and the second TZ is at 35.6 GHz which is close to the lower slope of the passband. The expanded fractional BW is about 8.5%, the realistic insertion loss is 1.88 dB at the centre frequency (39.5 GHz). However, the achieved selectivity is poor.

![Fig. 10. Simulated and measured results of the BPF1.](image)

Figure 11 shows the measured frequency response of BPF2. Two TZs are at 36.5 GHz and 42.6 GHz for the improved selectivity. The measured filter offers an expanded fractional BW of 8% and a realistic insertion loss of 1.74 dB at the centre frequency (39.5 GHz). The measured S11 is approximately 20 dB, and the out-of-band rejection is about 35 dB. A spur exists around 47.5 GHz as a result of the higher order TM$^{210}$ mode.

![Fig. 11. Simulated and measured results of the BPF2.](image)

Figure 12 shows the measured properties of the fabricated X-band BPFs in terms of S-parameter and the group delay. These parameters are generally superior to those of Q-band filters. BPF3 offers an insertion loss of 1 dB and a peak group delay of 2.07 $\text{ns}$ around 10.00 GHz. BPF4 offers an insertion loss of 1.3 dB, and a peak group delay of 3.12 $\text{ns}$ around 10.15 GHz.
In order to compare these filters, we use the rectangle ratio $K_S$ to represent selectivity:

$$K_S = \frac{BW_{-20dB}}{BW_{-3dB}}$$

When $K_S \rightarrow 1$, the transition region is getting smaller, and the out-of-band attenuation is improved. Consequently, the in-band selectivity is enhanced. We can see that the rectangle ratio of BPF4 is the best among these filters.

Table 2 lists the performance comparisons among the proposed broadband perturbed SIW Filters.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Unit</th>
<th>Q-band</th>
<th>X-band</th>
</tr>
</thead>
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<tr>
<td>Centre Frequency</td>
<td>GHz</td>
<td>39.5</td>
<td>10.25</td>
</tr>
<tr>
<td>Fractional BW</td>
<td>%</td>
<td>8.5</td>
<td>8</td>
</tr>
<tr>
<td>Rectangle ratio</td>
<td>%</td>
<td>2.95</td>
<td>1.64</td>
</tr>
<tr>
<td>Insertion loss</td>
<td>dB</td>
<td>-1.88</td>
<td>-1.74</td>
</tr>
<tr>
<td>Return loss</td>
<td>dB</td>
<td>-18.6</td>
<td>-21.3</td>
</tr>
<tr>
<td>Rejection@left</td>
<td>dB</td>
<td>22±30</td>
<td>40±42</td>
</tr>
<tr>
<td>Rejection@right</td>
<td>dB</td>
<td>10±20</td>
<td>33±35</td>
</tr>
<tr>
<td>Group delay Peak</td>
<td>ns</td>
<td>1.75</td>
<td>2.62</td>
</tr>
<tr>
<td>Cavities radius</td>
<td>mm</td>
<td>3.28</td>
<td>3.27/3.03</td>
</tr>
</tbody>
</table>

4. Discussion

In recent works [12-15], the authors proposed a number of designs and fabricated filters with the advantages of compact structures and adjustable passband with the first, the second, or both bands. Table 3 shows the comparisons of our proposed filters with the previous ones in terms of Centre Frequency, Independent tunability, Insertion loss, and Resonator type. It can be seen that our filters offer a strong improvement in term of Insertion loss. Furthermore, we offer a technique to adjust a TZ for widening the bandwidth and improving the selectivity by...
properly changing the location of an additional via-hole placed inside the cavity. Filters in [12-15] are tunable bandpass filter (BPF) with the first band, second band, and both bands being independently controllable. However, it lacked flexibility for bandwidth expansion. The selectivity of our proposed filter is significantly improved through the $K_s$ factor given in the comparison Table 3. Our solution successfully improved group delays while maintaining a flexible bandwidth expansion. Moreover, the proposed filters work on the Q and X bands which means the alternative technique has a new unique feature.

<table>
<thead>
<tr>
<th>Works</th>
<th>Centre Frequency (GHz)</th>
<th>Insertion Loss (dB)</th>
<th>FBW (%)</th>
<th>$K_s$</th>
</tr>
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<tbody>
<tr>
<td>[12]</td>
<td>7.71/9.64</td>
<td>-1.9/1.65</td>
<td>5.45/8.1</td>
<td>1.42/1.85</td>
</tr>
<tr>
<td>[13]</td>
<td>7.89/8.89</td>
<td>-1.5/-1.9</td>
<td>3.42/3.93</td>
<td>1.48/1.95</td>
</tr>
<tr>
<td>[14]</td>
<td>9.32/11.32</td>
<td>-2.43/-2.35</td>
<td>4.3/4.2</td>
<td>1.24/2.67</td>
</tr>
<tr>
<td>[15]</td>
<td>3.6/6.4</td>
<td>-1.3/-1.8</td>
<td>3.3/2.4</td>
<td>3.36/2.34</td>
</tr>
<tr>
<td>This Work</td>
<td>10.42/39.5</td>
<td>-1.20/-1.32</td>
<td>7/8</td>
<td>1.18/1.64</td>
</tr>
</tbody>
</table>

**5. Conclusions**

This paper proposed an alternative technique to design the wideband BPFs with controllable TZs. A novel perturbation via-hole is provided for the resonators, which can easily adjust the TZ locations in order to improve selectivity of the BPF and allows for expandable bandwidth as well. Four dual-mode perturbed SIW BPF prototypes operating at Q-band and X-band have been fabricated and tested. A high level of agreement is achieved between the measured and simulated results. This shows that the proposed technique opens up new possibilities for designing compact, low cost, broadband microwave and mm-wave planar filters while rendering any additional mechanical tuning element in post-fabrication unnecessary. This new proposed result can be applied to the next generation communication systems.

**Nomenclatures**

- $c$: Speed of light in free space, m/s
- $d$: Diameter, mm
- $E_0, H_0$: Fields of the original DMCC
- $f_{mnl}$: Resonant frequency, GHz
- $\varepsilon_r, \varepsilon_t$: Dielectric constant
- $K_s$: Rectangle ratio
- $L$: Distance between the two cavities, mm
- $m, n, l$: Indexes of TM_{mnl} modes
- $Q_u$: Unloaded quality factor
- $r, s$: Radius and the height of the cavity, mm
- $S$ parameter: Insertion Loss and Return loss
- $s$: Distance between the two via-hole, mm
- $T$: Transmission zero
- $V$: Volume of DMCC, mm$^3$
- $W_e$: Electric stored energy in the DMCC
$W_m$ Magnetic stored energy in the DMCC
$w_0, w_5, w_6$ Widths of inductive coupling window, mm
$w_1, w_2, w_3, w_4$ Widths of induction windows, mm

**Greek Symbols**

$\alpha, \beta$ Coupling angles, degree
$\Delta W_m$ Changes in the stored magnetic energy
$\Delta W$ Changes in the stored electric energy
$\varepsilon$ Resonant frequency of the original DMCC

**Abbreviations**

BPF Bandpass filter
BW Bandwidth
CPW Coplanar Waveguide
DMCC Double mode Circular Cavity
PCB Printed Circuit Board
Quasi-TEM Quasi Transverse Magnetic Mode
SIW Substrate Integrated Waveguide

**References**


