Dual Mode Square SIW Cavity Filter Using Corner Cut Perturbation

Morteza Rezaee and Amir Reza Attari

Abstract. This paper presents a square SIW cavity that is perturbed by displacement of corner vias as a corner cut perturbation. This perturbation causes mode splitting of degenerate modes, TE_{102} and TE_{201}, of the square SIW cavity. A dual mode filter using this cavity is realized with 2.8% Fractional Bandwidth (FBW) at 6.1GHz and two transmission zeros at 5.5GHz and 6.8GHz. This filter is modeled by a coupling matrix representation based on the global eigen modes. Also, the simulation and measurement results of this filter are presented.

Keywords: SIW cavity, dual mode filter, coupling matrix.

1. Introduction

Dual mode filters are used widely in the satellite communications and cellular base stations, due to their high selectivity and small size with respect to the single mode filters [1-4]. Generally, dual mode filters are realized by introducing a perturbation in a microwave cavity that supports two degenerate modes, in which the perturbation causes the mode splitting [5,6] or mode shifting [7]. This perturbation may be realized by introducing a screw at 45º with respect to the electric field direction of two degenerate modes [8], a corner cut in a rectangular waveguide cavity [3] and also, introducing an additional patch [5] or corner cut [1] in a microstrip cavity. Equivalent lumped element circuit (coupling matrix) [9], equivalent distributed model [10], and equivalent circuit based on propagation [11] are used to model and design of dual mode filters.

Recently, substrate integrated waveguide (SIW), which consists of two arrays of metallic vias in the substrate between top and bottom metal layers, has been used as a planar structure for microwave and millimeter applications [12]. SIW components have high Q and low loss similar to conventional waveguide components and can be fabricated easily by standard printed circuit board (PCB) process or low temperature co-fired ceramic (LTCC) technology [13]. In [14], the first SIW filter has been proposed based on the classical post-wall waveguide technique. So far, dual mode SIW filter using circular and elliptical SIW cavity [15,16], near square SIW cavity [2], and square SIW cavity with slot perturbation [17,18], have been proposed. Dual mode SIW filter is also loaded by Defected Ground Structure (DGS) [19] and Complementary Split Ring Resonator (CSRR) [20] to improve the filter performance.

Generally, slot perturbation, DGS, or CSSR structures cause higher insertion loss and malfunction in packaging due to the radiation via slots. In this paper, a square SIW cavity with corner cut perturbation is introduced and by using this resonator, a dual mode SIW filter is realized. Also, coupling matrix representation of this filter based on the global eigen modes, is presented and finally, the simulation and measurement results are investigated.

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2. Resonator Structure

Figure 1a demonstrates the proposed square SIW cavity which is perturbed by a square corner cut. In this figure, a is side length of the cavity, d is diameter of metallic vias and p is the via pitch size. For values of $p \le 2d$ and $d \le 0.1\lambda_0$, the leakage signal from the gap between adjacent vias can be neglected [2]. A square SIW cavity can be considered as a classical waveguide cavity with the effective side length of [21]

$$a_{eff} = a - \frac{d^2}{0.95p}. \tag{1}$$

It can be shown that TM^z modes cannot be excited in the SIW structure due to radiation from sidewall gaps and also, no variation exists in EM-fields of cavity in y direction because of small height of the substrate, h. Thus, only TE^z_{m0n} modes are excited in the SIW cavity with the following resonant frequencies

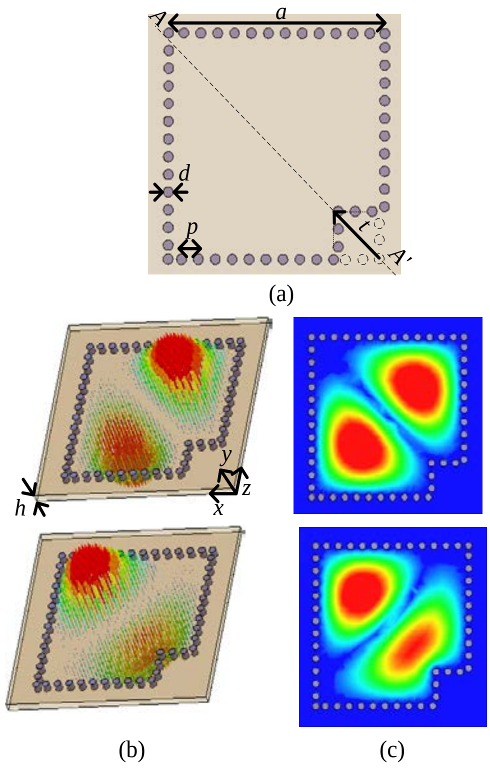


Fig. 1. (a) SIW cavity with square corner cut perturbation, (b) Simulated E-field of the odd mode, and (c) Simulated E-field of the even mode.

$$f_{m0n} = \frac{c}{2\sqrt{\varepsilon_r}} \sqrt{\left(\frac{m}{a_{eff}}\right)^2 + \left(\frac{n}{a_{eff}}\right)^2} \qquad (2)$$

where c is the velocity of light in free space [2]. The first degenerate modes of the square SIW cavity are TE_{102} and TE_{201}. By introducing the corner cut perturbation, based on the symmetry of structure with respect to A-A' axis, field distributions of eigen modes is symmetric. Figures 1b and 1c demonstrate the simulated *E*-field distribution of these modes that have odd and even symmetry, respectively. As can be seen, these modes are corresponding to the sum and difference of TE_{102} and TE_{201} modes, except around the perturbation. To study the mode splitting due to perturbation, the dual mode SIW resonator has been simulated using HFSS. As shown in Fig. 2, with increasing the perturbation size t, resonant frequency of odd mode remains almost fixed due to locating the corner cut at null position of the electric field distribution. However, resonant frequency of the even mode increases because the effective volume of cavity for this mode is decreased.

3. Dual Mode SIW Filter

3.1. Design of Dual Mode SIW Filter

Figures 3a-c show three square SIW cavities fed by 50Ω microstrip lines in the center of side length of cavities. The substrate of these cavities is Taconic RF-35 with ε_r=3.5, tanδ = 0.0018, and h =1.524mm. Figures 3d and 3e depict reflection and transmission responses of these structures, respectively. In Structure I, shown in Fig. 3a, the side length of the square SIW cavity, a, is equal to 30.7mm in order to set the f_{102} at 6GHz. This structure has no perturbation and as can be seen from figure 3e, it has a transmission zero around 6GHz. In this structure, because input ports are located at the center of cavity side length, one port excites only TE_{102} mode and the other port excites only TE_{201} mode. Therefore, because of locating each port at the null position of the field distribution excited by the other port, energy cannot be carried from one port to the other port at 6GHz and the transmission zero is generated.

The perturbed square SIW cavity using a corner cut perturbation, Structure II, is shown in Fig. 3b. The corner

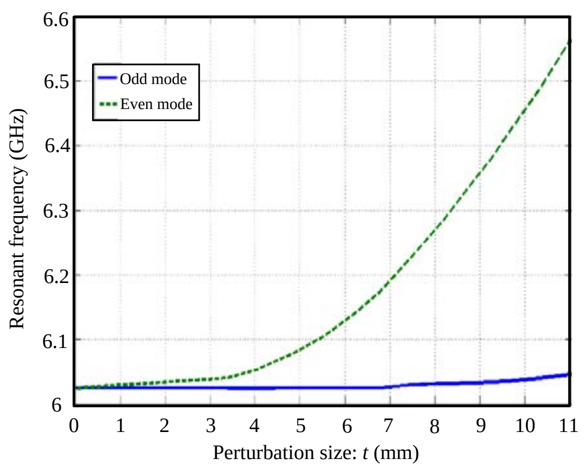


Fig. 2. Simulated resonant frequencies of the even and odd modes shown in figures 1b and 1c, with ε_r = 3.5, h=1.524mm, a= 30.7mm, d= 1.4mm, and p=2.4mm.

cut perturbs the TE_{102} and TE_{201} modes and generates two odd and even modes, as depicted in figures 1(b) and (c). In this case, each port excites both of these two modes and the two modes participate in transferring the energy from input to output port. Therefore dual mode operation is realized and as it can be seen from Fig. 3d, two poles (reflection zeros) are excited around the center frequency at 6GHz. However, from figure 3e, the selectivity of this filter is insufficient.

The Structure 3, shown in Fig. 3c, has two transmission zeros at the upper and lower sides of the pass band, and hence the selectivity of filter is improved. Indeed, by changing the perturbation position with respect to the

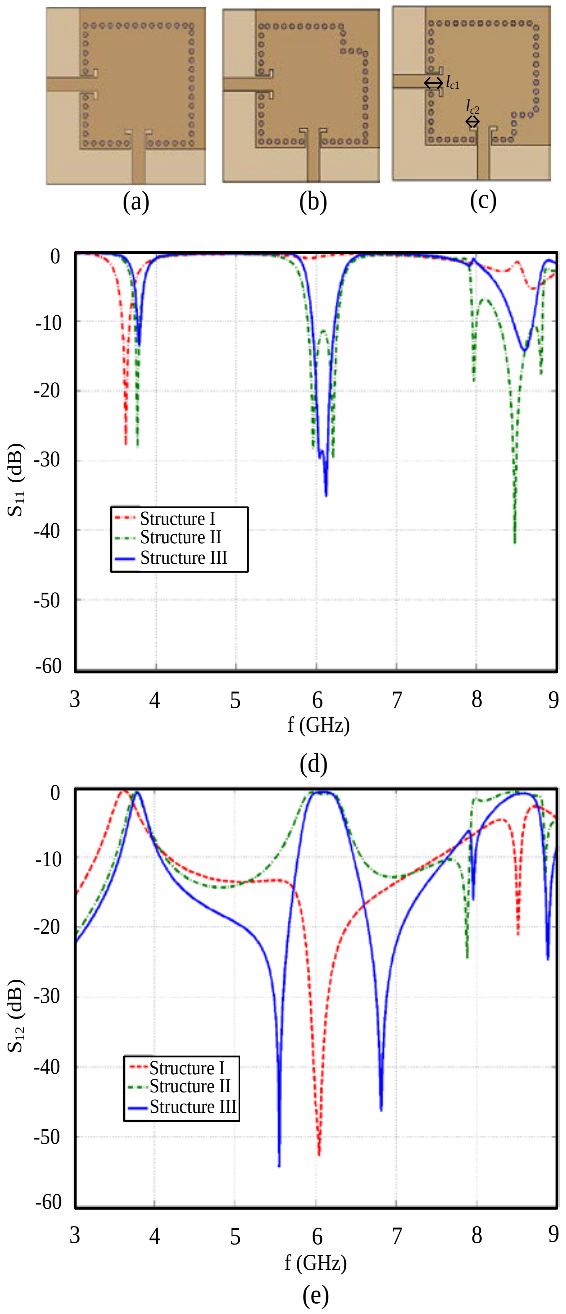


Fig. 3. (a) Square SIW cavity (Structure I), (b) perturbed square SIW cavity (Structure II) , (c) perturbed square SIW cavity with perturbation in the other corner of cavity (Structure III), (d) simulated reflection response of these structures, and (e) simulated transmission response of these structures.

position of ports, relative phase of fields at the ports changes and two transmission zeros are generated due to interference of resonant modes. Also, a transmission zero around 8GHz is generated because the ports are located at the null position of TE_{202} mode.

Generally, an *N*th-order cross coupled filter can be modeled by an equivalent circuit and represented by a coupling matrix [*M*]. In this model, the voltage-current relationship is given by [22]

$$\begin{aligned}&[E]=[A][I]\\&[A]=[R]+j\Omega[W]+j[M]\end{aligned} \tag{3}$$

where [*R*] is $(N+2)\times(N+2)$ diagonal matrix with $R_{ii}=1$ for i=1 and N+2, and otherwise $R_{ii}=0$ and [*W*] is $(N+2)\times(N+2)$ diagonal matrix with $W_{ii}=1$ for resonating nodes; and otherwise $W_{ii}=0$. Also, the normalized low pass frequency is obtained as $\Omega=\frac{f_0}{BW}(\frac{f}{f_0}-\frac{f_0}{f})$. S-parameters of the cross coupled filter are calculated by equation (4) [22]

$$S_{11}=1-2[A^{-1}]_{11}\ ,\ \ S_{21}=2[A^{-1}]_{N+2,1} \tag{4}$$

Using optimization a convenient coupling matrix for a dual mode filter with FBW = 2.5% at the center frequency of f_0 = 6GHz and two transmission zeros at 5.5GHz and 6.5GHz is obtained as follows

$$[M]=\begin{bmatrix}0.00 & 0.90 & -0.90 & -0.09\\ 0.90 & -2.00 & 0.00 & 0.90\\ -0.90 & 0.00 & 2.00 & 0.90\\ -0.09 & 0.90 & 0.90 & 0.00\end{bmatrix} \tag{5}$$

At first the self coupling coefficient of *i*-th resonance, M_{ii} , i =1 and 2, is realized. Using the following relation [22]

$$M_{ii}=\frac{f_0^2-f_i^2}{BW\ f_i},\quad i=1,2 \tag{6}$$

the resonant frequency of *i*-th resonance, f_i, is obtained. The side length of the square SIW cavity, *a*, should be determined so that to set the TE_{102} resonance mode at f_1. Also, the perturbation size, *t*, is selected from figure 2 to achieve f_2. From the input and output coupling coefficients to the *i*-th resonance mode, that is M_{si} and M_{li}, respectively, the convenient external quality factor, Q_{ei}, should be obtained with the following relation [22]

$$Q_{ei}=FBW\ M_{s/l,i}^2\ ,\qquad i=1,2 \tag{7}$$

Two L-shaped slots located at the ports are used to achieve the required quality factors for the input and output. These slots act as a current probe [7].

Detailed dimensions of the designed filter shown in Fig. 3(c) which realize the coupling matrix in equation (6), are presented in Table 1. The simulated FBW of the proposed filter with return loss better than 20dB is 2.8% at the center frequency of f_0=6.1GHz. Also, two transmission zeros at 5.5GHz and 6.8GHz are generated in the transmission response. It should be noted that the reported value of *a* in Table 1 is the same value considered for Structure I and hence the center frequency obtained in simulation results is a little greater than 6GHz.

3.2. Modeling of Dual Mode SIW Filter with Transversal Coupling Matrix

A filter can be modeled using the transversal coupling matrix that is based on the global eigen modes of the entire cavity including the perturbation [23]. The transversal coupling scheme of the proposed dual mode SIW filter is depicted in Fig. 4. This scheme includes the resonant nodes as well as the source and load nodes. The resonant nodes are corresponding to the eigen modes of the perturbed cavity. Nodes 2 and 3 (hachured circles) represent the odd and even modes shown in figures 1b and 1c that realize the dual mode operation of the filter around 6GHz. Higher and lower order resonant modes should be considered to provide a wideband representation. Node 1 is corresponding to the quasi TE_{101} mode and nodes 4 and 5 are corresponding to two modes that are similar to sum and difference of TE_{103} and TE_{301} modes. These nodes provide additional paths from input to output. It should be noted that TE_{202} mode is not excited because the feed lines are located at the null position of *E*-field distribution of this mode. Finally, the coupling matrix of filter can be determined by the following equations [22]

$$\begin{aligned}&M_{Si}=\frac{1}{\sqrt{FBW\ Q_{ei}}}\quad ,\quad Q_{ei}=\frac{\omega_i\ GD_{S11}(f_i)}{2}\\&M_{Li}=\frac{1}{\sqrt{FBW\ Q_{ei}}}\quad ,\quad Q_{ei}=\frac{\omega_i\ GD_{S22}(f_i)}{2}\\&M_{ii}=\frac{f_0^2-f_i^2}{BW\ f_i}\qquad ,\quad i=1,2,\ldots,5\end{aligned} \tag{8}$$

Table 1. Detailed Dimensions of the Proposed Dual Mode SIW Filter (units in Millimeter)

a	d_p	l_{p1}	l_{p2}	l_{c1}	l_{c2}	d_{via}	s_{via}
30.7	5	16.35	17.75	4.7	2.75	1.4	2.4

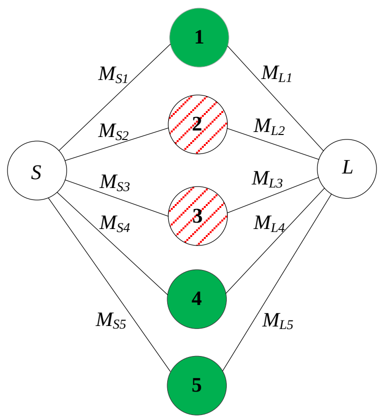


Fig. 4. Coupling scheme of the proposed dual mode filter (structure 3 shown in Fig. 3c).

where f_i and Q_i are the resonant frequency and quality factor of i-th resonator, respectively and GD is the group delay that are obtained from full wave simulation. Sign of the matrix elements is determined from relative phase of the *E*- (or *H-)* field at the input and output [24].

3.3. Results and Discussions

Coupling matrix of the proposed dual mode SIW filter is obtained by using equation (7) as follows

$$[M]=\begin{bmatrix} 0.0000 & 0.8266 & 0.8401 & -0.9724 & -1.2255 & 0.6672 & 0.0000 \\ 0.8266 & 32.6277 & 0.0000 & 0.0000 & 0.0000 & 0.0000 & 0.9121 \\ 0.8401 & 0.0000 & 2.3077 & 0.0000 & 0.0000 & 0.0000 & 0.8624 \\ -0.9724 & 0.0000 & 0.0000 & -1.8468 & 0.0000 & 0.0000 & 0.9322 \\ -1.2255 & 0.0000 & 0.0000 & 0.0000 & -23.8026 & 0.0000 & 1.2998 \\ 0.6672 & 0.0000 & 0.0000 & 0.0000 & 0.0000 & -27.3090 & 0.4678 \\ 0.0000 & 0.9121 & 0.8624 & 0.9322 & 1.2998 & 0.4678 & 0.0000 \end{bmatrix} \tag{9}$$

The frequency responses calculated from equations (4), (5), and (9) are depicted in figures 5a and 5b and compared with simulation results. These figures show a good agreement between the simulated and calculated responses. A filter prototype is fabricated in standard PCB technology and its photograph is shown in figure 5c. Measurement results of the manufactured filter are also presented in figures 5a and 5b and as it is seen these results are in good agreement with the simulation results. The measured FBW of the filter is 2.6% at the center frequency of 6.05GHz with return loss better than 20dB. Also, the minimum measured and simulated insertion loss of the filter are 1.1dB and 1.2dB, respectively. Also, from the simulation results, insertion loss due to dielectric loss, conductor loss, and other losses including leakage and return loss are about 0.4dB, 0.2dB, and 0.5dB, respectively. A comparison of the proposed filter with some prior SIW filters is summarized in Table 2 and as can be seen, the proposed filter has acceptable characteristics.

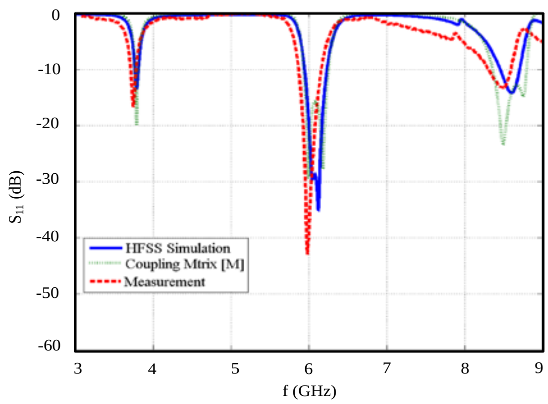
Transmission response of the filter for different values of perturbation size is shown in Fig. 6 and as can be seen with increasing t, the bandwidth of filter increases because of more splitting the modes, but achieving a good matching will be more difficult.

Table 2. Comparison with Some Prior SIW Filters

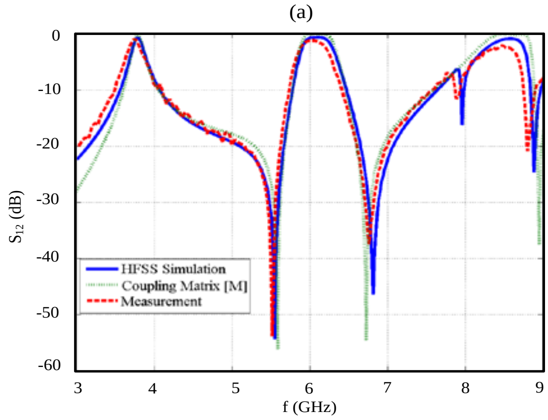
Ref.	Order, Cavity Type	Dimension $\lambda_0 \times \lambda_0$	Insertion Loss (db)	FBW, f_0 (GHz)	Number of TZ below and above pass band
This paper	2, dual mode	0.61×0.61	1.1	2.8%, 6.0	1 and 1
[18]	2, dual mode	0.72 × .72	2.6	2.6%, 14.4	1 and 1
[25]	2, dual mode	0.65 × .62	1.6	3 %, 5.0	1 and 1
[20]	3, dual mode	0.76 × .76	1.3	5%, 5.0	1 and 2
[2]	4, dual mode	1.82 × .73	2.6	1 %, 10	1 and 1
[26]	4, single mode	0.93 × .91	0.9	3.4%, 20.5	1 and 1
[7]	3, triple mode	0.61 × .61	0.7	9.9 %, 6.0	0 and 3

4. Conclusion

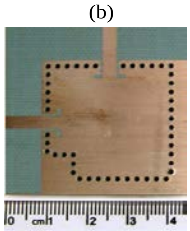
In this paper, a dual mode square SIW cavity with a corner cut perturbation is presented. Using two orthogonal micro strip feed lines, the even and odd modes corresponding to the sum and difference of TE_{102} and TE_{201} modes are excited and a dual mode filter with two transmission zeros is realized. The coupling matrix representation of the proposed filter based on the global eigens modes of cavity is presented. Frequency responses of the filter from full wave simulation, coupling matrix, and measurement are obtained and compared with each other.



(a)



(b)



(c)

Fig. 5. (a) Reflection coefficient and (b) transmission coefficient of the proposed filter obtained from full-wave simulation, coupling matrix, and measurement. (c) Photograph of fabricated filter.

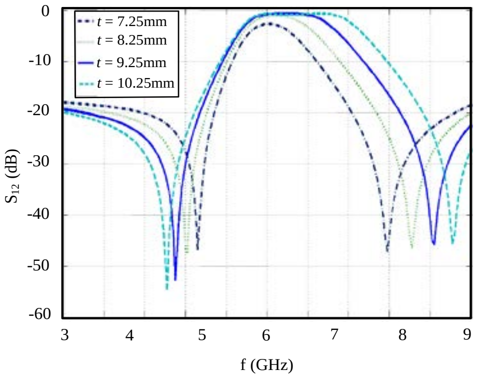


Fig. 6. Transmission coefficient of filter shown in figure 3(c) for different values of perturbation size *t*.

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