Novel dual-mode filters implemented by quarter mode folded substrate integrated waveguide

Ling Yang,1 Feng Xu,1,✉ Qing Liu,2 Jingxia Qiang,1 and Junlin Zhan1

1School of Electronic and Optical Engineering, Nanjing University of Posts and Telecommunications, Nanjing, People’s Republic of China
2PLA Strategic Support Information Engineering University, Zhengzhou, People’s Republic of China

✉Email: feng.xu@njupt.edu.cn

Two dual-mode filters implemented by folded substrate integrated waveguide (FSIW) are proposed here. The high order modes (TE103 and TE301) in the FSIW cavity are employed to realize the function of filters. The non-resonating node TE202 mode is employed in the first filter to realize the transmission zero below the passband. A metal pin and an L-shaped slot are used in another dual-mode filter to restrict the parasitic mode, TE202 mode. The proposed FSIW dual-mode filters with simple structures exhibit compact size and good frequency selectivity. As per knowledge, it is the first time that the quarter mode FSIW (QMFSIW) cavities are utilized to design the multimode filters. The measured results agree well with the simulated ones, which validates the feasibility of the compact dual-mode filters based on the technology of double-layered FSIW.

Introduction: Filters which can be utilized to combine or separate different frequencies are of great concern in wireless communication systems. Unlike traditional bulky metallic waveguide, substrate integrated waveguide (SIW) [1–4] has advantages of compact size, low cost and low profile. Besides that, the planar structure makes it possible to integrate with other planar circuits. On the basis of SIW, the folded substrate integrated waveguide (FSIW) is proposed to realize miniaturizing in the lateral direction while preserving all the merits of SIW [5]. For the same cut-off frequency, the size of the FSIW is nearly half of SIW. As a consequence, the FSIW is a promising transmission line to design many microwave/RF components. The geometry of the asymmetrical FSIW as well as the cross-section view of the FSIW is illustrated in Figure 1. As shown in Figure 1, the electric field and the magnetic field of the upper layer and the bottom layer have the same magnitude but reverse direction. The propagation constant of the FSIW has been investigated in [6]. An accurate analysis of the FSIW which shows precise phase constant without an empirical factor for the solution of the equivalent folded rectangular waveguides (FRWGs) is demonstrated in [7]. Dual-mode FSIW filters in the K-band are realized by the technology of low-temperature co-fired ceramics (LTCC) in [8] in which the single resonant cavity with degenerate modes folded TE102 and TE201 is used to design the filters. However, to excite the dual modes, the feed lines are realized by open microstrip coupling through the rectangular slots on the ground plane which needs an extra dielectric layer and which causes unnecessary losses. In addition, the quarter mode FSIW (QMFSIW) cavities have never been used to realize dual-mode high-performance miniaturized filters. Hence, here, QMFSIW cavity is chosen to design the dual-mode filters. The configuration of the QMFSIW cavity which consists of two substrate layers and three metal layers is depicted in Figure 2.

Two different kinds of dual-mode filters implemented by QMFSIW cavities, are designed and fabricated here. In consideration of the modes distributed in the QMFSIW cavities, to realize TZs on the basis of the even mode and the odd mode, TE103 and TE301 are used to realize the passband of the filters here. Compared with the dual-mode SIW filters [8–12], the proposed dual-mode filters implemented by the FSIW cavities have better insertion loss, simple structure, compact size and an enclosed structure. The dual-mode FSIW filters are simulated by the high frequency structure simulator (HFSS). The filters designed here are fabricated on the Rogers 5880 substrate with a nominal dielectric constant of 2.2 and a thickness of 0.508 mm.

Figure 3. As shown in Figure 3 TE103 mode has a symmetric field with respect to the diagonal line, while TE101 mode has an antisymmetric pattern. Considering the field distributions of the degenerate modes, the feed lines of the dual-mode filters are symmetric with respect to the diagonal line and perpendicular to each other.

Type A: Dual-mode filter with two transmission zeros: A dual-mode FSIW filter with two transmission zeros is designed and fabricated first. The prototype and the topology of the dual-mode FSIW filter is shown in Figure 4. The operating frequency of the passband is determined by the resonant frequency of the even mode TE103 and odd mode TE101...
while the mode TE_{202} acts as a non-resonating node (NRN) [13]. The entry of the coupling matrix is denoted by $M_{ij}$, then the loop currents, which are grouped in a vector are given by a matrix equation of the form [14].

$$[-jR + \Omega W + M][I] = [A][I] = -j[e], \quad j^2 = -1.$$  

The transmission coefficient $S_{21}$ is given by (load and source resistors = 1)

$$S_{21} = -2j[A^{-1}]_{n+1,1}.$$  

According to [15], the resonate frequency of the even mode and odd mode can be calculated by the following equations:

$$f_{even} = f_0 \left(1 - \frac{M_{S1} \times \Delta f}{2f_0}\right),$$  

$$f_{odd} = f_0 \left(1 - \frac{M_{S2} \times \Delta f}{2f_0}\right).$$  

In which, $f_0$ is the centre frequency of the bandpass filter and $\Delta f$ is the bandwidth. The expression relating the coupling elements and the TZ caused by different paths from even modes and odd modes as well as the NRN which can be calculated from (2) is provided in a lowpass prototype as follows:

$$a = M_{S1},$$  

$$b = M_{S1}\Delta S_{11} + M_{S2}\Delta S_{22} - M_{S1}^2 + M_{S2}^2,$$  

$$c = M_{S1}\Delta S_{11} - M_{S2}\Delta S_{22} + M_{S1}M_{S2},$$  

$$\omega_b = \frac{-b \pm \sqrt{b^2 - 4ac}}{2M_{S2}}.$$  

For the proposed filter, $|M_{S1}| > |M_{S2}|$, $f_{even} < f_0$ and $f_{odd} > f_0$ ($M_{S1} > 0$ and $M_{S2} < 0$). $M_{S2} < 0$. It is clear that, $a \cdot c < 0$ and $\sqrt{b^2 - 4ac} > |b|$, therefore, two TZs always exist and are located in the upper stopband and lower stopband, respectively. For this case no extra path caused by the NRN exists between the source and load, only one TZ in the upper stopband is resulted by the different signal paths from the even mode and odd mode, and the TZ can be calculated by:

$$\omega_b = \frac{(M_{S1}M_{S2}^2 - M_{S2}M_{S1}^2)}{(M_{S1}^2 - M_{S2}^2)},$$  

Which coincides with that derived in [15], meanwhile the TZ is above the passband.

**Type B: Dual-mode filter with mode suppression:** The response of the filter like filter A but which is excited from the middle of a magnetic wall is illustrated in Figure 5. A spurious passband caused by the by-pass coupling between the TE_{202} and the TE_{103} mode is located at 7.8 GHz. To improve the out-of-band rejection, the structure of filter B with mode suppression is shown in Figure 6. A metal pin and an L-shaped slot which is located at the corner are utilized to suppress the TE_{202} mode in the QMFSIW cavity so as to eliminate the parasitic passband at lower frequency. Different from the electric field of the TE_{010} and TE_{011}, the electric field in the upper layer and the bottom layer of the TE_{202} mode has the same direction which results in the stripline that do not excite the mode adequately. According to (9), a TZ is formed above the passband. Owing to the existence of the fundamental mode, a spurious passband would appear near 6 GHz and a TZ is realized at 7 GHz as the resonant frequency of the TE_{010} mode will shift towards higher frequencies by introducing the L-shaped slot and the metal pin.

**Fabrication and measurement:** The proposed two compact QMFSIW filters consist two substrate layers and three metal layers which are realized by the standard PCB technology. The comparisons between simulated and synthesized results as well as the measured ones are depicted in Figure 7. It is clear that the measured return loss of type A is better than 15 dB while the minimum measured insertion loss of the passband is 1.1 dB. The measured return loss of the type B filter is better than 14 dB, and the measured minimum insertion loss is 1.2 dB. From the comparison between Figures 5 and 7, it can be concluded that the parasitic passband caused by the TE_{202} has disappeared which validates the feasibility of the proposed filter. Good
consistency can be seen between the simulated results and the measured ones. Owing to the mismatching tolerance and the loss caused by the SMA connector, there is a slight deviation between the simulated and measured results. Table 1 lists some related works presented in the references.

**Conclusion:** Two dual-mode filters with different feed positions (electric wall and magnetic wall, respectively) and implemented by the dual-layered FSIW technology are proposed and fabricated here. By introducing an NRN in the filter of type A, two TZs are located in the lower stopband and upper stopband, respectively. For the filter of type B, one metal pin and an L-shaped slot are introduced creatively to restrict the TE_{202} mode and improve the stopband characteristics. The filters realized here exhibit good passband performance, low insertion loss, good selectivity and compact size. Besides that, the enclosed structure makes it possible to integrate with other planar circuits.

**Acknowledgements:** This work was supported by the major project fund of natural science research in colleges and universities of Jiangsu province under grant No. 16KJA510003.

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**Table 1. Comparison with other works**

|       | \(f_0\) (GHz) | FBW (%) | IL (dB) | Size (\(\lambda^2\)) | TZ |
|-------|----------------|---------|---------|----------------------|----|
| Type A| 9.32           | 9       | 1.1     | 0.92 x 0.92          | 2  |
| Type B| 9.30           | 4.9     | 1.2     | 0.92 x 0.92          | 2  |
| [8]   | 20.5           | 5       | 2       | 1.2 x 1.2            | 2  |
| [11]  | 15             | 4.3     | 1.7     | 2.5 x 1.34           | 5  |

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