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High Isolation Substrate Integrated Waveguide Diplexer With Flexible Transmission Zeros

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Abstract—This letter presents a high isolation substrate integrated waveguide (SIW) diplexer using dual-mode resonators. Each dual-mode SIW resonator can generate one transmission pole (TP) and one transmission zero (TZ) for the channel filter. The frequencies of the TZs can be easily controlled by the perturbation vias. One channel filter's two TZs' frequencies are designed to cover the other channel filter's passband. Thus, high isolation performance is achieved. The proposed diplexer is designed and fabricated with lower and upper channels centered at 8.04 and 9.07 GHz, respectively. Simulated and measured results are in good agreement for the two channels with better than 49/53 dB channel isolation.

Index Terms—Substrate integrated waveguide, high isolation, diplexer, dual-mode.

I. INTRODUCTION

SUBSTRATE integrated waveguide technology has attracted more and more research interests because of its merits of high Q , low-cost, light-weight, and easy to connect to other planner circuits [1]. With the rapid expansion of microwave communication systems, SIW has been widely applied to the design of diplexer which is one of the major limiting components in a transceiver. High isolation between the two channel filters is required to design the diplexer with excellent performance due to the requirement of good interference suppression between the two channels.

To improve the isolation characteristics of the diplexer, one method is to put one channel's TZs to the other channel's passband. Several techniques are proposed to generate the TZs, including dual-mode SIW resonator [2]-[4], cross-coupling technology [5], and mixed electric and magnetic coupling technology [6]. However, the TZ generated by the dual-mode SIW resonator cannot be independently controlled which is also associated with the bandwidth of the channel filter [2]-[4]. Furthermore, the mixed electric and magnetic coupling method is complicated because electric and magnetic coupling coefficients were deliberately tuned to satisfy both the bandwidth and TZ requirement [6]. In [3] and [5], the suppression level is limited with only one TZ at the other

channel filter's passband. If more TZs are introduced, the isolation level can be improved.

In our former work [7], a circular dual-mode SIW filter with two flexible TZs above the passband was proposed. The merit is that the TZs can be controlled by the location of the perturbation vias. However, the TZs only can be above the passband. This kind of resonator can be used as the lower channel filter. In this letter, a more general circular dual-mode SIW resonator is proposed which can be used to design the filter with TZs above or below the passband. Unlike the planar structure in [7], the resonators in this letter are vertically stacked to reduce the size. Moreover, one channel's two TZs are designed to cover the other channel filter's passband. Thus, the isolation level between the two channels is improved. Extra TZs near the passband of each channel are also generated which could further improve the selectivity of the channels. The design concept is proved by the measured results of the diplexer which centers at 8.04 and 9.07 GHz. The measured isolation is greater than 49/53 dB which is much better than that in the previous works.

II. DUAL-MODE RESONATOR DESIGN

Fig. 1. Structure of the proposed dual-mode resonator.

Fig. 2. Magnetic field distributions of the dual-mode resonators.

The proposed circular dual-mode cavity resonator is shown in Fig. 1. Two arrays of orthogonal metallized vias are placed in a circular cavity. The two arrays have $2m$ and $2n$ ($m, n = 0, 1, 2 \dots$) vias, respectively. Fig. 2 shows the magnetic field distributions of the original dual-mode resonator ($m = 0, n = 0$) and the proposed resonators, namely Type-A ($m = 1, n = 0$), Type-B ($m = 2, n = 0$), Type-C ($m = 2, n = 1$). The mode whose magnetic line is not cut by the vertical line is labeled mode I and

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the other one is labeled mode II. The resonant frequencies of the two degenerate modes in the original dual-mode resonator equals to each other which can be calculated by [8]

$$f_{TM110} = (0.610c) / (R\sqrt{\mu_r\epsilon_r}), \quad (1)$$

where, c is velocity of light in the vacuum, R is the radius of the circular cavity, μ_r and ϵ_r are relative permeability and permittivity of the material, respectively. For other dual-mode resonators ($m \neq 0, n \neq 0$), the vertically distributed perturbation vias are located in the node of the standing wave of mode I. So those vias would not influence mode I's resonant frequency f_I . On the other hand, the vias would force mode II's electrical field to be zero. The field of mode II is squeezed. The resonant frequency f_{II} would be pushed to a higher frequency. For the horizontal distributed perturbation vias, similar effect can be achieved. Thus, we have the relationship of the resonant frequencies shown in Table I. There are two methods to adjust the resonant frequencies. One is to add more vias. With more vias, the resonant frequency can be push higher, vice versa. The other one is to move the via along the diameter line. If the via is placed closer to the center, a higher resonant frequency can be generated, vice versa. It should be noted that when one mode's resonant frequency is adjusted, the other one's resonant frequency keeps the same. It is the key to the design of flexible TZ. In table I, the lower frequency of the dual-mode resonator is defined by mode I which equals or is larger than f_{TM110} . If the vertical perturbation vias are connected together, there would be a metal wall in the middle ($m = \infty, n = 0$). In this situation, the magnetic field distribution is similar to TM_{210} mode whose resonant frequency can be calculated by [8]

$$f_{TM210} = (0.817c) / (R\sqrt{\mu_r\epsilon_r}). \quad (2)$$

This frequency defines the higher frequency of mode II. So the maximum frequency ratio of the dual-mode is

$$R_f = f_{TM210} / f_{TM110} = 1.34. \quad (3)$$

TABLE I
RELATIONSHIP OF THE RESONANT FREQUENCIES

Type-A	$f_I = f_{TM110} < f_{II}$
Type-B	$f_I = f_{TM110} < f_{II}$
Type-C	$f_{TM110} < f_I < f_{II}$

III. DIPLEXER DESIGN

Fig. 3 shows the structure and coupling scheme of the lower and upper channel filters. As shown, the dual-mode resonators are vertically stacked and coupled by the rectangular slot on the middle metal layer. The slot is parallel to the magnetic line of mode I and perpendicular to that of mode II in the lower channel filter. Thus, only mode I in the two resonators is magnetically coupled to each other. In the upper channel filter, mode II is used to generate the passband that the coupling slot is parallel to the magnetic line of mode II. So the lower channel has two TZs above the passband while the upper channel's two TZs are below the passband. The feeding structure has an angle of 45° to the lines of the vias. It can excite both mode I and mode II in the resonators.

The two channel filters are designed to have center frequencies at 8 and 9 GHz with $BW_{3dB} = 360$ MHz ($BW_{0.0432dB} = 148$ MHz). The 8 GHz filter has two TZs $f_{z1} = 8.89$ GHz, $f_{z2} = 9.06$ GHz while the 9 GHz has two TZs $f_{z3} = 8.17$ GHz, $f_{z4} = 7.88$ GHz. The synthesis method is described our previous work [7]. The self-coupling coefficients can be calculated by

$$M_{ii} = f_0 / f_i - f_i / f_0, \quad (4)$$

where f_0 is the center frequency of the filter, and f_i is the resonant frequency of the i th resonator. The coupling coefficients for the two filters are given as $\{M_{S1}, M_{S3}, M_{L2}, M_{L4}, M_{12}, M_{11}, M_{22}, M_{33}, M_{44}\} = \{0.0233, 0.0297, 0.0233, 0.0297, 0.0309, 0.0063, 0.0063, -0.2114, -0.2495\}$ and $\{0.0224, 0.0398, 0.0224, 0.0398, 0.0276, -0.0102, -0.0102, 0.1938, 0.2666\}$, respectively. Other elements in the coupling matrix are all zero.

Fig. 3. Configuration and coupling scheme of each channel filter. (a) Lower channel filter. (b) Upper channel filter. ($R_1 = R_2 = 14.17$, $R_3 = R_4 = 15.36$, $d_1 = 10.9$, $d_2 = 9.2$, $d_3 = 10.9$, $d_4 = 7.74$, $d_5 = 12.9$, $L_1 = 4.24$, $L_2 = 4.42$, $s = 1$, unit: mm, $\theta_1 = \theta_2 = 30.4^\circ$, $\theta_3 = \theta_4 = 24^\circ$).

After the coupling matrices are obtained, the following steps are taken to get the filters' physical dimensions. Firstly, the F4B substrate (dielectric constant = 2.6, loss tangent = 0.001, substrate thickness = 0.8 mm and copper thickness = 0.018mm) is chosen to design the channel filters. Then, the dimensions of each resonators can be derived by the center frequencies and TZ frequencies. The dimensions of the external and internal coupling structures are defined by the extracted coupling coefficients in Fig. 4. Finally, all the physical dimenmsions are

optimized to satisfy the specifications. The final dimensions of the filters are listed in the caption of Fig. 3.

The responses of the synthesized and simulated channel filters are shown in Fig. 5. Those results agree well with each other except some extra spurs in the simulated $|S_{21}|$ which are introduced by the weakly coupling of mode II in Type-A resonators and the coupling of mode I in Type-B and Type-C resonators. However, those spurs would not influence the isolation performance or channel filter's passband performance.

The two channel filters are combined by a microstrip line T-junction to form the diplexer in Fig. 6. The width of each port is 2.3 mm with 50 Ohm characteristic impedance. The lengths of the arms are tuned to make sure that each channel filter in the diplexer is an open circuit to the other filter at its center frequency. In other words, the length of the arms should make sure that $Z_{in1}'@9$ GHz and $Z_{in2}'@8$ GHz in Fig. 6 are infinity.

Fig. 4. Extracted coupling coefficient versus the slot angle/slot length. (a) Lower channel filter. (b) Upper channel filter.

Fig. 5. Synthesized and simulated frequency response of the channel filters. (Solid line: synthesized, dash line: simulated).

Fig. 6. Explode 3-D view of the proposed diplexer. ($L_3 = 16.91$ mm, $L_4 = 15.6$ mm).

IV. MEASUREMENT

The photograph, simulated, and measured results of the fabricated diplexer are shown in Fig. 7. The measured results show that two channels have center frequencies of 8.04 and 9.07 GHz, with 3-dB fractional bandwidth of 4.23% and 4.19%, respectively. The measured insertion losses (2.35/2.33 dB at 8.04/9.07 GHz) are a little larger than the simulated ones

(1.38/1.45 dB at 8.04/9.07 GHz). It is conjectured that the actual loss tangent of the substrate is larger than the specification. The isolation between two channels is greater than 49 or 53 dB. The measured responses have a small frequency shift than the simulated ones which may be caused by the substrate's relative dielectric constant errors. Extra TZs at 8.3 and 8.7 GHz are introduced by mode I of Type-C resonator and mode II of Type-A resonator, respectively. Those TZs further improve the selectivity of the channels. The measured stopband for the lower/upper channel is better than 23/16 dB up to 13.87/12.95 GHz. Table II compares our diplexer with other designs in the literature. Although the diplexer in this work seems a little larger due to the T-junction, it exhibits a better isolation performance. To further minimize the diplexer size, many methods can be used such as folded T-junction [9] and common resonator method [10]-[12].

Fig. 7. Photograph, simulated and measured results of the diplexer (Dash line: simulated results, solid line: measured results).

TABLE II

COMPARISON BETWEEN THE PROPOSED DIPLEXER AND THE REFERENCES

Ref.	Order	f_0 (GHz)	FBW (%)	IL (dB)	Isolation (dB)	Size (λ_g^2)	Layer
[2]	4, 4	24.98, 26.65	5.41, 5.25	2, 2.5	40	NA	single
[3]	2, 2	5, 5.25	1.95, 2.08	2.2, 2.4	45	2.12	single
	2, 2	5.5, 5.25	1.94, 2.05	1.8, 1.5	40	2.12	single
[4]	4, 4	24.93, 26.59	5.42, 5.15	2.09, 1.95	20	NA	single
[5]	3, 3	9.5, 10.5	4.21, 3.81	1.6, 2.1	35	3.89	single
[6]	2, 2	9.55, 10.61	2.95, 3.21	2.62, 2.34	45	3.56	double
[11]	3, 3	12, 14	4.9, 5.65	1.34, 1.41	27	2.77	single
[12]	2	10, 12.25	2.58, 3.51	2.05, 1.65	20	2.34	single
	2	8, 12	2.94, 1.75	1.84, 2.88	20	1.54	single
[13]	2	2.35, 29.87	19.15, 6.29	0.94, 1.32	22	0.16	double
This work	2, 2	8.04, 9.07	4.23, 4.19	2.35, 2.33	49, 53	4.6	double

V. CONCLUSION

A SIW diplexer with high isolation is proposed by employing dual-mode resonators in this letter. The dual-mode resonators can generate flexibly controllable TZs. With those TZs, high isolation is achieved. The simulated and measured results show good agreement which validates the design concept. The proposed high isolation diplexer is low-cost and can be applied to improve the performance of RF front end.

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