DUAL MODE COUPLING BY SQUARE CORNER CUT IN RESONATORS AND FILTERS

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ABSTRACT

A new method for realization of dual mode coupling in rectangular waveguide cavities is described and analyzed. The new method completely replaces the coupling screw, and therefore can be used to eliminate the need for tuning in dual mode waveguide cavity filters. It also offers a wide range of coupling values and can achieve higher power handling capability than coupling screws. Mode matching method is used to calculate the mode chart of the infinite corner cut rectangular waveguide, the field distributions of each mode, and the resonant frequencies of the cavity. Dual mode coupling parameter is calculated from resonant frequencies and is verified by measurements. A C-band 4-pole dual mode elliptic function rectangular waveguide cavity filter using the new coupling method was constructed and tested. The experimental results showed excellent agreement with theory.

I. INTRODUCTION

Dual mode empty or dielectric loaded resonator filters are widely used in satellite communications, due to their sharper performance, smaller size and less mass than conventional single mode direct coupled filters [1]-[7]. The usual way to couple dual modes is by adding a coupling screw at 45° angle with the direction of the electric fields of the two dual modes. By changing the coupling screw penetration in the cavity, the coupling between the two dual modes can be adjusted. Due to the sharpness of the coupling screws, the filter's power handling capability is reduced. Although the tuning and the coupling screws provide flexibility for adjusting the filter's response, the tuning process itself is time consuming and makes the cost of dual mode filters production high. Recently, a method has been introduced to realize dual mode coupling in planar waveguide structures [7]. The realization is through a perturbation of the single mode microstrip line resonator. On waveguide structures, a similar method can be used to replace the coupling screws in dual mode filters. This paper introduces such a method which has the ability to provide a wide range of coupling values with minor reduction of cavity power handling capability. The new method has the potential of reducing production cost of dual mode waveguide cavity filters by eliminating the need for tuning altogether. The key is the ability to accurately analyze the coupling configuration. The method developed and presented in this paper uses mode matching technique and yields accurate results which were verified by measurements.

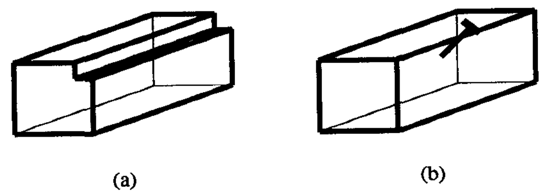

Fig. 1 Coupling configuration (a) new method,and (b) conventional coupling screw

To illustrate the application of the new coupling mechanism, an experimental C-band 4-pole dual mode elliptic function rectangular waveguide cavity filter was designed. In reference [6] a length of evanescent mode rectangular waveguide was used to provide the required coupling values between physically adjacent cavities and is accurately modeled by mode matching method. These couplings can be separately controlled by the two dimensions in the cross section. This method is used in the present paper to design the 4-pole dual mode filter. The experimental filter was tested, showing excellent agreement with theory, with no tuning.

II. ANALYSIS

Fig. 1(a) shows the proposed new coupling configuration under study. A square (or rectangular) waveguide cavity is perturbed by a square cut on one of its corners. With no cut, the square waveguide cavity has two dominant degenerate modes with the same resonant frequencies and perpendicular electric field distributions. Those two modes can be referred to as TE_{101} and TE_{011} modes and are decoupled from each other. With the square cut, the two dominant modes will be coupled to each other, and the resonant frequencies will split. As a comparison, the conventional method of coupling the two dual modes in a cavity by a coupling screw is shown in Fig. 1(b).

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To analyze the square cut waveguide cavity, first, the infinitely long perfectly conducting square cut waveguide, with the cross section shown in Fig. 2, is analyzed. Mode matching method is employed to solve this problem. The cross section is divided into two regions. Region I is the area $0 \le y \le b_1$, and region II is the area $b_1 \le y \le b_2$. The modes existing in the square cut waveguide are still *TE* & *TM* modes [8].

1. *TE* modes

For region I,

$$j\omega\mu H_{z1} = \sum_n A_{1n} \cos(k_{x1n}x)\cosh(k_{y1n}y) \quad (1a)$$

$$E_{x1} = -\sum_n \frac{A_{1n}}{k_c^2} k_{y1n} \cos(k_{x1n}x)\sinh(k_{y1n}y) = j\omega\mu H_{y1} Z_{TE} \quad (1b)$$

$$E_{y1} = -\sum_n \frac{A_{1n}}{k_c^2} k_{x1n} \sin(k_{x1n}x)\cosh(k_{y1n}y) = j\omega\mu H_{x1} Z_{TE} \quad (1c)$$

where

$$k_{x1n}^2 - k_{y1n}^2 = \gamma^2 + k^2 = k_c^2 \quad (1d)$$

$$k_{x1n} = \frac{n\pi}{a_2}, \quad k^2 = \omega^2\mu\epsilon, \quad Z_{TE} = \frac{1}{\gamma} \quad (1e)$$

γ, k_c, and Z_{TE} are the propagation constant, the cut-off wave number, and the wave impedance, respectively.

For region II,

$$j\omega\mu H_{z2} = \sum_m A_{2m} \cos(k_{x2m}x)\cosh(k_{y2m}y) \quad (2a)$$

$$E_{x2} = -\sum_m \frac{A_{2m}}{k_c^2} k_{y2m} \cos(k_{x2m}x)\sinh(k_{y2m}y) = j\omega\mu H_{y2} Z_{TE} \quad (2b)$$

$$E_{y2} = -\sum_m \frac{A_{2m}}{k_c^2} k_{x2m} \sin(k_{x2m}x)\cosh(k_{y2m}y) = j\omega\mu H_{x2} Z_{TE} \quad (2c)$$

where

$$k_{x2m}^2 - k_{y2m}^2 = \gamma^2 + k^2 = k_c^2 \quad (2d)$$

$$k_{x2m} = \frac{m\pi}{a_1} \quad (2e)$$

By applying the boundary conditions at $y = b_1$, taking the inner products, and using the orthogonality relations on the eigen-fields in each region, the following characteristic equation is obtained:

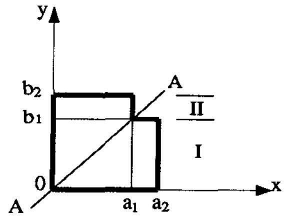

Fig.2 Coordinate system and the cross section of the square cut waveguide

$$\det[X] = 0 \quad (3)$$

where X is an $(N \times N)$ matrix and N is the number of eigen-modes used in region I. The elements of the matrix X are

$$x_{ij} = \cosh(k_{y1j}b_1) \sum_{m=0}^{M} \left[T_m k_{y2m} \tanh(k_{y2m}(b_1 - b_2)) < \hat{e}_{2m}, \hat{h}_{1j} >< \hat{e}_{2m}, \hat{h}_{1i} >\right] - \delta_{ij} \frac{a_1 a_2}{4T_i} k_{y1i} \sinh(k_{y1i}b_1) \quad (4a)$$

$$T_m = \begin{cases} 1, & m \neq 0 \\ \frac{1}{2}, & m = 0 \end{cases}, \quad \delta_{ij} = \begin{cases} 0, & i \neq j \\ 1, & i = j \end{cases} \quad (4b)$$

where $(M + 1)$ is the number of modes used in region II.

Solving equation (3), a group of eigen-values of γ's can be obtained. Each γ corresponds to a mode in the square cut waveguide.

2. *TM* modes

Similar procedures can be used to describe *TM* waves in the square cut waveguide. The same characteristic equation as equation (3) is obtained with the matrix elements given by

$$x_{ij} = \cosh(k_{y1j}b_1) \sum_{m=1}^{M} \left[\frac{tanh(k_{y2m}(b_1 - b_2))}{k_{y2m}} < \hat{e}_{2m}, \hat{h}_{1j} >< \hat{e}_{2m}, \hat{h}_{1i} >\right] - \delta_{ij} \frac{a_1 a_2}{4} \frac{sinh(k_{y1i}b_1)}{k_{y1i}} \quad (5)$$

And the wave impedance is

$$Z_{TM} = -\frac{\gamma}{k^2} \quad (6)$$

A computer program using the mode matching technique was developed to calculate the propagation constants for the modes that can exist in the waveguide. Fig. 3 shows the variation of the normalized propagation constants with the normalized size of the square cut for several *TE* and

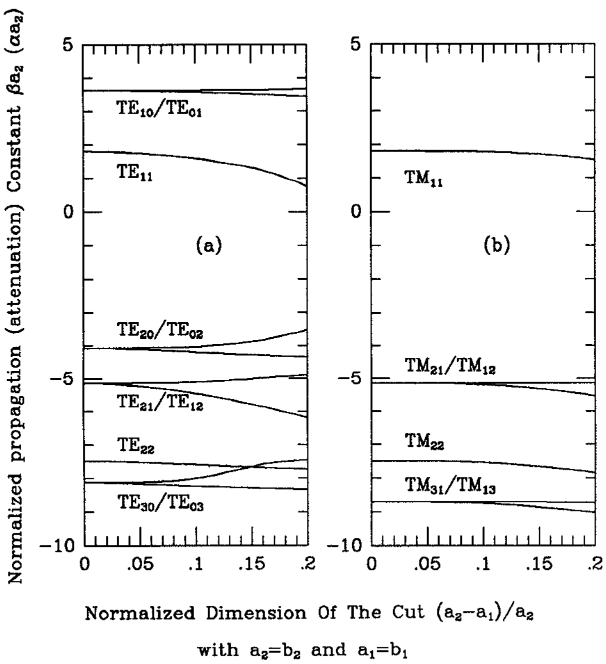

Fig. 3 Mode chart for TE and TM modes

TM modes at normalized frequency $fa_2 = 9GHz \cdot inch$. The normalized propagation constants βa_2 for propagating modes are shown as positive, and the attenuation constants αa_2 of cut-off modes are shown as negative. The split of the dual modes (i. e. $TE_{10}/TE_{01}, TE_{20}/TE_{02}, TE_{21}/TE_{12}$ and TE_{30}/TE_{03} in (a)) is clearly apparent when the dimension of the square cut increases. Due to the different propagation constants of all the dual modes, once the two conducting short circuit planes are added to form a waveguide cavity, the corresponding resonant frequencies for the dual modes will also be different.

III. DUAL MODE COUPLING

The two equivalent circuits shown in Fig. 4 can be used to model the coupling structures between two cavities [4]-[6]. If the coupling coefficient between the cavities is M, then from the symmetry of the structure, it is possible to calculate M from a knowledge of the resonant frequencies f_e and f_m, where f_e and f_m are the resonant frequencies assuming the symmetric plane $A - A$ as a perfect electric conductor and a perfect magnetic conductor, respectively. The coupling coefficient is simply:

$$c = \frac{M}{L} = \frac{f_e^2 - f_m^2}{f_e^2 + f_m^2} \tag{7}$$

The field distribution in the cross section for the first dual modes in the square cut waveguide are calculated and plotted in Fig. 5(a) and (b). From these two plots, it is seen that the plane perpendicular to the cross section, containing the z-axis and the line parallel to the z-axis

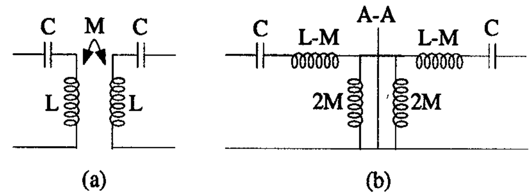

Fig. 4 Equivalent circuits of the coupling in resonators

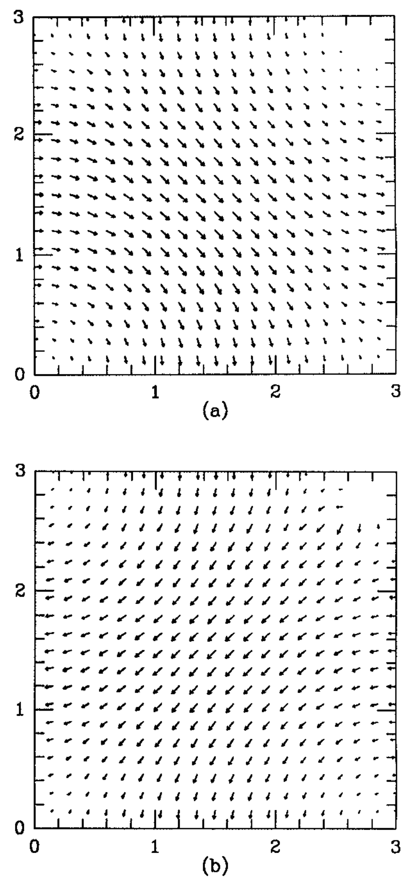

Fig. 5 Field distribution plot of (a) TE_{1e}, and (b) TE_{1m}

through the vertex of the square cut (a_1, b_1) in the cross section is the symmetric plane $A - A$ as marked in Fig. 2. The electric fields at this plane are maximum and are parallel to the plane for the odd modes (magnetic wall) and perpendicular to the plane for the even modes (electric wall). These two modes are just the split dual modes due to square cut perturbation. Equation (7) can be used to calculate the coupling coefficient, where f_e and f_m are the two split resonant frequencies with the higher one as f_e (smaller propagation constant, electric wall) and lower one as f_m (larger propagation constant, magnetic wall).

IV. RESULTS

The developed computer program, which computes the propagation constants for each mode in the square cut waveguide and its field distribution, also calculates f_e, f_m, and c. Fig. 6 shows the calculated results of TE_{10}/TE_{01} modes on a C-band cavity resonator and the measured data. It is noticed that both f_e and f_m are affected by the cut and f_e increases while f_m decreases with the cut size.

In order to verify the usefulness and validity of the new coupling method, a C-band 4-pole dual mode filter using the new technique was designed with the configuration shown in Fig. 7(a), where the couplings between the two cavities are through an evanescent mode rectangular hole [6]. The coupling between dual modes in the cavities are $M_{12} = M_{34} = c = 0.01628$, which are provided by the cut in each cavity. The coupling through the evanescent mode rectangular hole are $M_{14} = -0.00603$, and $M_{23} = 0.01622$, which are controlled by the dimensions of the hole. The measured filter responses are shown in Fig. 7(b). These results agree very closely with the theoretical filter responses.

V. CONCLUSIONS

The new dual mode coupling mechanism presented in this paper is simple, reliable, and flexible. It allows a wide range of variation of the coupling with reasonable cut. The same strategy can be used to couple dielectric loaded resonator dual mode filters. Mode matching method was successfully applied to accurately compute the propagation constants of the waveguide and then the resonant frequencies and the couplings of the dual mode cavity. Experimental verification of the calculated results were done. Finally, a C-band 4-pole dual mode filter was designed using the new coupling mechanism. Measurements on the filter show the validity of the method. This configuration has the potential for relatively inexpensive production of dual mode filters and millimeter wave filters, since the cavities can be easily machined (e.g. using an Numerical Controlled Mill), then assembled and yield the required response with no tuning.

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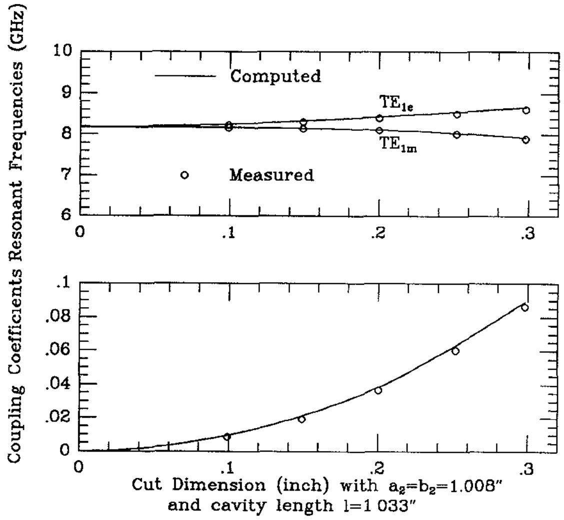

Fig. 6 Comparison of computed and measured f_e, f_m, and k for a C–band cavity resonator

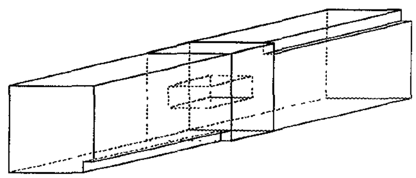

Fig. 7(a) 4-pole dual mode filter configuration

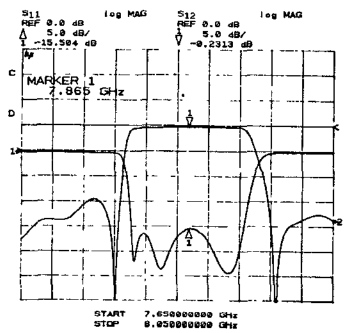

Fig. 7(b) Measured responses of the filter

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