A designing solution for bandpass filter using TE dual-mode resonator in waveguides

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ABSTRACT

The article presents a design for a bandpass filter using TE dual-mode resonator. The design is based on the general scatting matrix of each discontinuity in the structure. Different propagation modes in the same waveguide section are used to represent the matrices to facilitate the design. On the basis of ultra-high frequency theory and simulation software tools, the authors have calculated the design of a 4th order bandpass filter using the resonances, TE_{01} and TE_{10} with two symmetrical transmission zeros.

Keywords: Bandpass filter; TE dual-mode.

1. INTRODUCTION

Filters play a critical role in the transceiver system of radio communication systems in general and radar systems in particular. Bandpass filters are one of the essential components of today's radars. Many different types of filters are used for radar applications, such as microstrip filters, coaxial resonant cavity sets, waveguide filters, and bandpass filters using lumped elements. The microstrip filter has the advantages of small size, low cost, and easy fabrication. The main disadvantage of this filter is its high insertion loss because its Q-factor is significantly lower than that of other types [6]. The coaxial resonant cavity filter has many advantages, such as low loss, compact size. However, it has the disadvantage of medium power and is difficult to fabricate at high frequencies [5]. The waveguide filter has the advantages of low loss, high power, but the size of the waveguide filter is much larger than that of other filters. It is suitable for use at the output of transmitters. To reduce the waveguide filter size, there are two main methods, and the first is to use multimode theory to create multimode coupling filters. Another approach is to use a dielectric resonator to create a waveguide filter. The application of single-cavity multimode theory to design filters was proposed in 1951. After that, Atia and Williams proposed a relatively complete dual-mode waveguide filter design theory in 1971 [8]. Dual-mode filters are created using perturbations placed in the cavity to produce degenerate modes [1-3]. The TM dual-mode waveguide filter can achieve a compact structure by using unique modes to propagate electromagnetic waves. Compared with the TE dual-mode waveguide filter, the thickness of the waveguide filter is further reduced. In addition, the TM N-cavity dual-mode filter can realize 2N transmission zeros. TE dual-mode waveguide filters and coupling between resonant modes in a single cavity are also cross-coupling with resonant modes of adjacent cavities. Therefore, they can achieve the characteristic of the elliptic function filter, and TE dual-mode N-cavity can realize 2N-2 transmission zeros.

This paper studies a filter design solution using TE dual-mode because it has the advantages of low loss, high-frequency selectivity, and smaller size than conventional waveguide filters. Based on ultra-high frequency and simulation software theories, the paper will present a solution to design a bandpass filter using TE dual-mode for the frequency band between 11.9 GHz and 12.1 GHz.

2. PROBLEM

2.1. Coupling matrix of a TE dual-mode bandpass filter

Filters that use TE dual-mode resonant cavities, such as those shown in fig. 1, are often represented by a set of cross-coupled resonances, as shown in figure 2 [1-3]. Resonance is understood as the singular solutions of cavity structures, vertically and horizontally polarized resonances for such as the filters in fig.1.

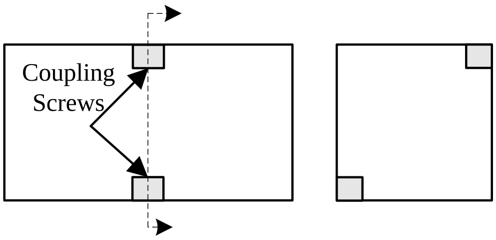

Figure 1. *The geometric shape of a dual-mode resonant cavity.*

Although the circuit in figure 2 can accurately depict the filter principle using TE dual-mode, it does not accurately describe the electromagnetic field inside the cavities. It is based on the cavity's modes. That is, each mode is considered a cavity. In addition, the symmetry of the perturbation cavity is different from that of the empty cavity. In short, the solutions of Maxwell's equations for a cavity and a cavity with perturbation are different.

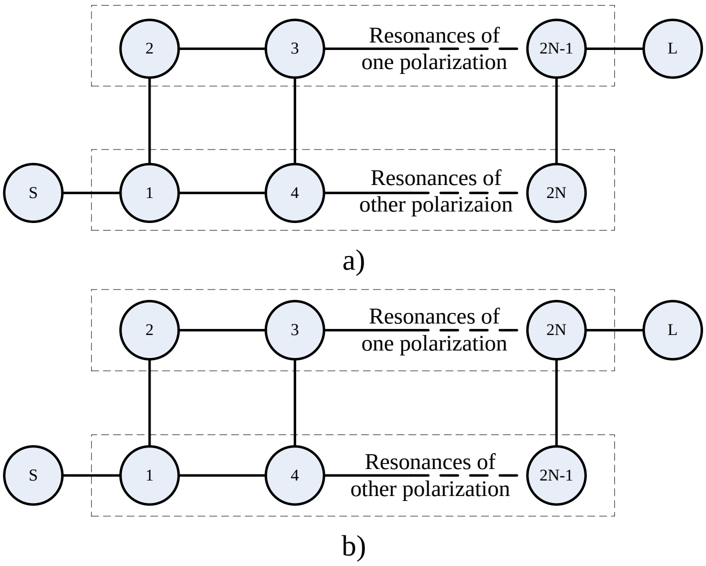

Figure 2. *Coupling scheme of a dual-mode filter for odd (a) and even (b) number of cavities.*

Assume the coupling matrix M is based on the coupling scheme in fig. 2, where M_{ij} is the coupling coefficient between the i^{th} and j^{th} resonant cavities, the diagonal elements on the coupling matrix represent the frequency displacement. T is the (2N+2)x(2N+2) transformation matrix.

$$[T] = \begin{bmatrix} 1 & 0 & 0 & \cdots & 0 \\ 0 & & & & 0 \\ 0 & & P & & 0 \\ \vdots & & & & \vdots \\ 0 & 0 & 0 & \cdots & 1 \end{bmatrix} \quad (1)$$

where P is a 2Nx2N real orthogonal matrix.

$$P = \begin{bmatrix} [A_1]_{2x2} & 0 & \cdots & 0 \\ \cdots & [A_2]_{2x2} & \cdots & 0 \\ \cdots & \cdots & \cdots & \\ 0 & 0 & \cdots & [A_N]_{2x2} \end{bmatrix} \quad (2)$$

Here, A_i is the eigenvector of the submatrix representing the modes in the i^{th} cavity, and N is the number of resonant cavities. I.e., that it's a 2N-order filter, as shown in fig. 2, when the resonator is tuned synchronously, the coupling matrix of the i^{th} cavity can be expressed in the following form:

$$M_{si} = \begin{bmatrix} 0 & M_{2i-1,2i} \\ M_{2i-1,2i} & 0 \end{bmatrix} \quad (3)$$

The eigenvalues of this matrix are:

$$\lambda_i = \pm M_{2i-1,2i} \quad (4)$$

Normalized eigenvectors, given by the matrix A, are as follows:

$$A_i = \begin{bmatrix} \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & \frac{-1}{\sqrt{2}} \end{bmatrix} \quad (5)$$

From (5), we can see that the partial diagonalization, corresponding to a rotation, is equal to 45^0. Suppose this rotation is applied to the original coupling matrix, with the coupling structure given in fig. 2. In that case, the two new resonances are polarized along the symmetry plane of the cross-section. For a square cavity with a perturbation, the eigenmode resonances have a wall of electric and magnetic fields along the plane of symmetry. The plane of symmetry is the diagonal containing the coupling perturbation when only one angular step is used. When two identical angular steps are used, as shown in fig. 3(a), both diagonals are planes of symmetry. Using diagonalization, we get the new coupling matrix as follows:

$$[M]' = [T][M][T]^T \quad (6)$$

We can see that the new coupling matrix has the following characteristics:

1) The diagonal elements are λ_i,
2) Between two modes in the same physical cavity, there is no coupling.

It should be noted that due to the discontinuity, i.e., when the coupling elements interfere with the symmetry of the characteristic modes p and q, the modes in the consecutive cavities are cross-coupled. Fig. 3(a) and fig. 3(b) show the topology for the coupling matrix, transformed by (6).

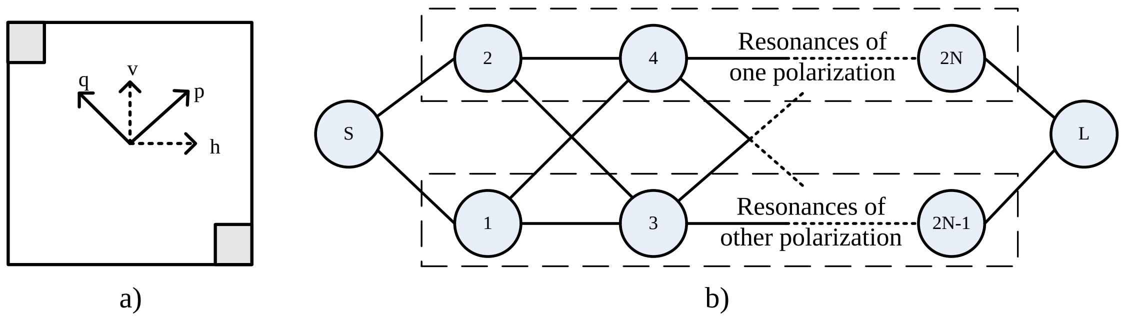

Figure 3. *a) The polarization modes in the cavities that have perturbation; b) Topology of the coupling matrix.*

The scheme in fig. 3b contains information about the load of the diagonal modes due to perturbation. This is reflected in the resonant frequencies of the diagonal elements of the coupling matrix (6). However, using this scheme to describe the resonance between the coupling elements between the cavities is imprecise. Therefore, an equivalent circuit is required to accurately model and design a dual-mode filter, which allows accurate load calculation. It has been done in a propagation-based model [4]. Fig. 4 depicts the structure diagram of a bandpass filter using dual-mode resonance.

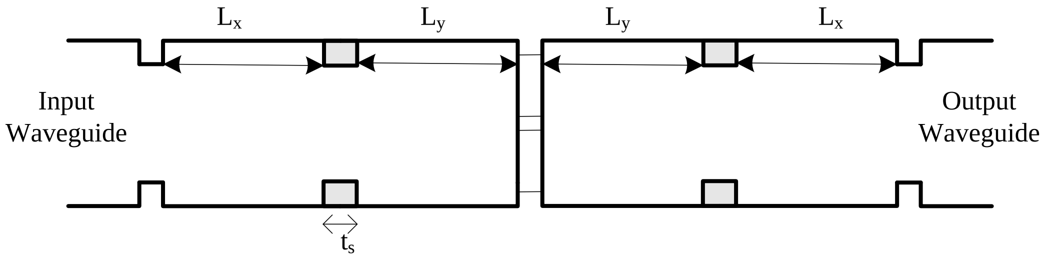

Figure 4. *Structure diagram of the 4th order dual-mode bandpass filter.*

2.2. Calculations and design a bandpass filter using TE dual-mode resonator

Based on the theory in Section 2.1, the research team then proposed the steps to calculate and design a bandpass filter using TE dual-mode resonator. These are explained in the example of a 4th order bandpass filter with a square waveguide and corner perturbations. The same steps can be used to design a higher-order filter.

The ideal coupling matrix for a filter with the required specifications can be obtained by analysis or optimization. Through either of these two methods, it is possible to get the coupling matrix. In general, the ideal coupling matrix has the following form:

$$M = \begin{bmatrix} 0 & M_{01} & 0 & 0 & 0 & 0 \\ M_{01} & 0 & M_{12} & 0 & M_{14} & 0 \\ 0 & M_{12} & 0 & M_{23} & 0 & 0 \\ 0 & 0 & M_{23} & 0 & M_{12} & 0 \\ 0 & M_{14} & 0 & M_{12} & 0 & M_{01} \\ 0 & 0 & 0 & 0 & M_{01} & 0 \end{bmatrix} \quad (7)$$

where, M_{ij} is the coupling factor.

Using the matrix transformation formula (6), we have

$$M' = \begin{bmatrix} 0 & M'_{01} & M'_{02} & 0 & 0 & 0 \\ M'_{01} & -\lambda_1 & 0 & M'_{13} & M'_{14} & 0 \\ M'_{02} & 0 & \lambda_1 & M'_{23} & M'_{24} & 0 \\ 0 & M'_{13} & M'_{23} & -\lambda_2 & 0 & M'_{01} \\ 0 & M'_{14} & M'_{24} & 0 & \lambda_2 & M'_{02} \\ 0 & 0 & 0 & M'_{01} & M'_{02} & 0 \end{bmatrix} \quad (8)$$

The calculation and design steps are as follows:

Step 1: Determine the size of the perturbation and the waveguide part

In this step, we need to determine the depth d of the perturbation and the length L of the square waveguide on either side of the perturbation.

First, the length t_s of the perturbation is determined through the practical value, specifically based on the width of the tuning screw. Then use electromagnetic simulation software such as CST to determine the dimensions in fig. 5. The depth, *d* determines the distance between the two resonant frequencies, while the length *L* will change both frequencies.

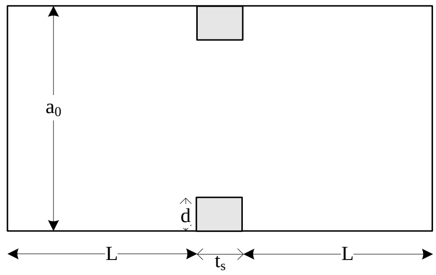

Figure 5. *Setting up the electromagnetic field simulation for the eigenmode of the dual-mode resonance cavities.*

Step 2: Determine the size of the coupling window between the dual-mode resonant cavities

Fig. 6 shows the geometry of the coupling window between two dual-mode resonant cavities. It consists of two identically symmetrical rectangular windows positioned with a distance o from the longitudinal centerline of the resonator. This type of coupling is the most suitable for resonant cavities with a square cross-section. It is more convenient to use vertically and horizontally polarized propagation modes in the design due to their cross-sectional symmetry. The distance o mainly controls the combination of vertical polarization modes on both sides, while the horizontal polarization mode is controlled primarily by b_2. It is due to the magnetic field distribution of the two modes. As the distance o increases, the magnetic field of the longitudinally polarized mode decreases while that of the transversely polarized mode remains almost constant in that direction.

On the other hand, increasing the b_2 dimension leads to stronger magnetic coupling between horizontally polarized propagation modes while having virtually no effect on the coupling between vertically polarized propagation modes. It means that the coupling for both modes can be controlled almost independently.

To design a coupling window between two dual-mode resonant cavities, the process is similar to that in the design of a conventional cross-coupling window. The general scattering matrix of the scheme was extracted by EM simulation. When placed in a uniform waveguide, the vertically and horizontally polarized propagation modes on both sides of the coupling window, TE_{10} and TE_{01}, remain separated.

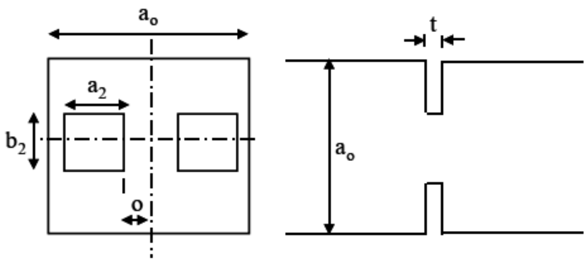

Figure 6. *Coupling scheme between two dual-mode resonant cavities.*

Step 3: Define the input coupling window

In this step, we define the input coupling window model and size. Unlike the coupling window of a dual-mode resonant cavity, the input coupling window represents a geometrically and electrically asymmetrical discontinuity. Fig. 7a shows the geometry of the input coupling window and the propagation modes on both sides. The modes on the right can be implemented as horizontal and vertical polarization modes or cross-polarized modes, TE_{10} and TE_{01}, while the input rectangular waveguide on the left only supports a propagation mode TE_{10} in the frequency band of interest.

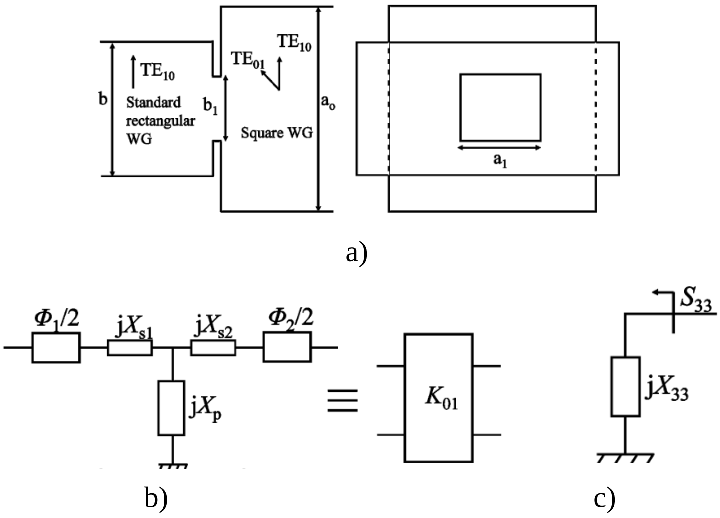

Figure 7. *a) EM simulation diagram for an input coupling window; b) Equivalent circuit for an input coupling window with vertical polarization mode; c) Equivalent circuit diagram for an input coupling window with horizontal polarization mode.*

Due to the symmetry, the vertical and horizontal polarization of the modes is not coupled across the coupling window. The following 3x3 scattering matrix relates the incident and reflected fields on both sides of the discontinuity:

$$\begin{bmatrix} b_{v1} \\ b_{v2} \\ b_{h2} \end{bmatrix} = \begin{bmatrix} S_{11v} & S_{12v} & 0 \\ S_{21v} & S_{22v} & 0 \\ 0 & 0 & e^{j(\varphi_{33}+\pi)} \end{bmatrix} \begin{bmatrix} a_{v1} \\ a_{v2} \\ a_{h2} \end{bmatrix} \tag{9}$$

where a_i, b_i are feature vectors of the incident and reflected waves at the i^{th} port. It is easy to show that the field of the cross-polarized modes is related to the input modes as follows:

$$\begin{bmatrix} b_{v1} \\ b_p \\ b_q \end{bmatrix} = R_1^{-1} \begin{bmatrix} S_{11v} & S_{12v} & 0 \\ S_{21v} & S_{22v} & 0 \\ 0 & 0 & e^{j(\varphi_{33}+\pi)} \end{bmatrix} R_1 \begin{bmatrix} a_{v1} \\ a_{v2} \\ a_{h2} \end{bmatrix} \tag{10}$$

where R_1 is the transformation matrix:

$$R_1 = \begin{bmatrix} 1 & 0 & 0 \\ 0 & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \\ 0 & \frac{1}{\sqrt{2}} & \frac{-1}{\sqrt{2}} \end{bmatrix}$$

The coupling window exhibits asymmetric discontinuity for the vertical mode, which the equivalent circuit can model in fig. 7b. On the other hand, the horizontally polarized mode propagates only in the square waveguide portion, while it does not propagate in the input rectangular waveguide. This means that the coupling window and input waveguide represent the load impedance for horizontal mode, as shown in fig. 7c. This reactance can be controlled by the vertical dimension of the input window.

The circuit in fig. 7b is an impedance inverter. The characteristic impedance parameter of the inverter in the diagram of fig. 7b is calculated according to the following formula:

$$K_{01} = \sqrt{\frac{\pi}{2}} B_\lambda M_{01} \tag{11}$$

where, $B_\lambda = \left|\frac{\lambda_{g1}-\lambda_{g2}}{\lambda_{g0}}\right|$

The coupling coefficient of the longitudinal mode to the input waveguide is determined by the a_1 dimension and it has little effect on b_1 dimension. On the other hand, the horizontal mode load, φ_{33}, depends mainly on b_1.

To determine those, we start with a square coupling window and resize it until the required value of the coupling factor is reached. The thickness of the coupling window we choose to be fixed at 1 mm.

After fixing the input coupling window and phase shift (φ_2 and φ_{vv}) for the longitudinal mode, the waveguide section's dimensions, L_x and L_y, in fig. 4 are determined as follows:

$$L_x = L + \frac{\lambda_{g0}}{2\pi}\frac{\varphi_2}{2} \tag{12}$$

$$L_y = L + \frac{\lambda_{g0}}{2\pi}\frac{\varphi_{vv}}{2} \tag{13}$$

When all design parameters have been determined, except horizontal mode's load, φ_{33}. Using the condition of phase equilibrium of the two modes, the required value can be obtained as:

$$\varphi_{33} = \varphi_2 + \varphi_{vv} - \varphi_{hh} \tag{14}$$

3. RESULT AND DISCUSSION

3.1. Filter's requirements

The design of the *4th* order bandpass filter using TE dual-mode is done with technical parameters according to table 1.

Table 1. *Technical requirements of a bandpass filter using TE dual-mode resonator.*

No	Specification	Unit	Value
1	Central frequency	GHz	12
2	Bandwidth	MHz	200
3	Reflection loss	dB	≥ 20
4	Transmission zeros	MHz	11840/12160

3.2. Calculation and design for the bandpass filter

Based on the required parameters in table 1, using the optimization method of calculating the coupling matrix on Matlab software, we get the ideal coupling matrix and the parameter S response of the filter as shown in fig. 8.

Using (6), we transform the ideal coupling matrix to:

$$M' = \begin{bmatrix} 0 & 0.7193 & 0.7193 & 0 & 0 & 0 \\ 0.7193 & -0.8325 & 0 & 0.2608 & -0.5519 & 0 \\ 0.7193 & 0 & 0.8325 & -0.5519 & 0.2608 & 0 \\ 0 & 0.2608 & -0.5519 & -0.8126 & 0 & 0.7193 \\ 0 & -0.5519 & 0.2608 & 0 & 0.8126 & 0.7193 \\ 0 & 0 & 0 & 0.7193 & 0.7193 & 0 \end{bmatrix}$$

Based on the values in the new coupling matrix obtained after the calculation, we calculate the physical dimensions of the filter. First, we choose the cross-sectional size of the rectangular waveguide as $a_0 = 16.93\ mm$, the length of the perturbation $t_s = 3mm$. Using CST software [7] to simulate and optimize the parameters of the perturbation, such as the depth d and cavity length L. Then determine the coupling window dimensions. The results after the optimization simulation and calculation is shown in table 2. The 3D model of the filter is shown in fig. 9. The results of the optimization simulation of the filter are shown in fig. 10.

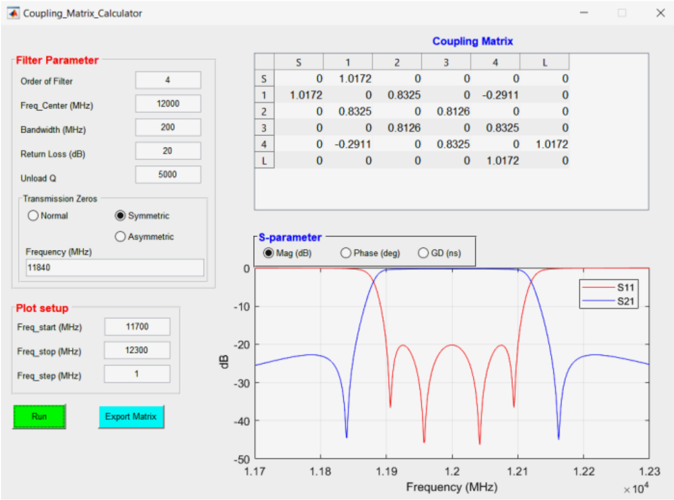

Figure 8. *Calculation results of the coupling matrix and parameter S response of the filter.*

Table 2. *The results after optimization simulation for a bandpass filter using TE dual-mode.*

No	Parameter	Value (mm)	No	Parameter	Value (mm)	No	Parameter	Value (mm)
1	a_0	16.94	5	b_1	7.89	9	a	22.86
2	t_s	3	6	a_2	5.06	10	b	10.16
3	t	1	7	b_2	5.38	11	L_x	6.52
4	a_1	9.97	8	o	2.09	12	L_y	7.68
						13	d	1.88

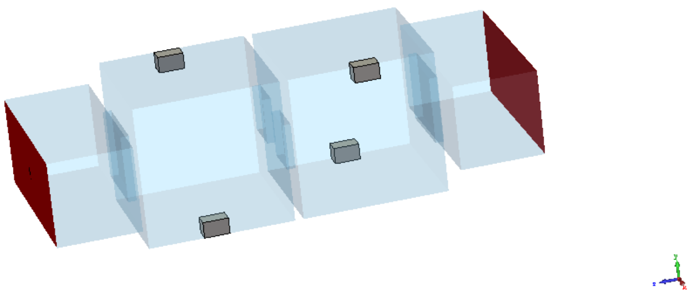

Figure 9. *3D model for a bandpass filter using TE dual-mode.*

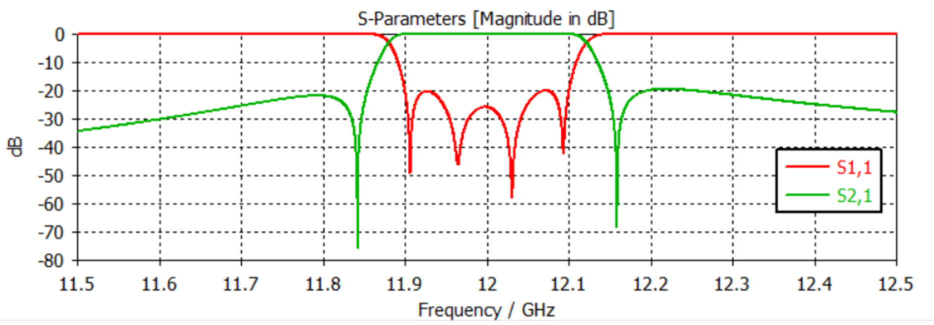

Figure 10. *Simulation result for the bandpass filter using TE dual-mode.*

From the simulation results, we can see that all parameters of in-band loss, out-of-band interception, and reflection loss of the waveguide bandpass filter meet the required specifications of the filter. The simulation design results and the required response are relatively consistent.

4. CONCLUSION

The paper presented the steps to calculate and design a bandpass filter using the resonance of TE dual-mode on waveguides based on theoretical calculations of the filter parameters combined with CST simulation to calculate the optimal physical values for the filter. A 4^{th} order bandpass filter using TE dual-mode resonance on the waveguide has been designed and simulated. The results meet the strict requirements of in-band loss, reflection loss, and especially the accuracy of the out-of-band cut-off. The results of this study are the basis for expanding the research, design, and manufacture of filters on waveguides with strict out-of-band cut-off requirements for applications in receiver systems and radar systems.

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TÓM TẮT

Về một phương pháp thiết kế bộ lọc thông dải sử dụng cộng hưởng của mode TE kép trên ống dẫn sóng

Bài báo trình bày kết quả nghiên cứu thiết kế bộ lọc thông dải sử dụng cộng hưởng của mode TE kép. Thiết kế dựa trên ma trận tán xạ tổng quát của từng điểm gián đoạn trong cấu trúc. Các mode lan truyền khác nhau trong cùng một phần ống dẫn sóng được sử dụng để biểu diễn các ma trận nhằm tạo điều kiện thuận lợi cho việc thiết kế. Trên cơ sở lý thuyết siêu cao tần và các công cụ phần mềm tính toán mô phỏng, nhóm tác giả đã tính toán thiết kế bộ lọc thông dải bậc 4 sử dụng cộng hưởng TE_{01} và TE_{10} với hai điểm không truyền dẫn đối xứng.

Từ khóa: Bộ lọc thông dải; Mode TE kép.