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A NOVEL AND ACCURATE METHOD FOR DESIGNING DIELECTRIC RESONATOR FILTER

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Abstract—The Development of numerical techniques enables us to analyze a large number of complex structures such as dielectric resonator (DR) filters and planar passive elements for coplanar monolithic microwave integrated circuits.

As for DR filters, numerical analysis of these structure is highly intricate mostly because of their non-homogenous composition (dielectric constant of DR is greater than 80, dielectric constant of the maintainer is less than 2 and dielectric constant of the atmosphere is 1). Hence, numerical analysis of such a structure, either in time domain (TLM, FDTD and ...) or frequency domain (FE, moment, mode matching, boundary element, FD and ...) is both complex and time-consuming. From one hand, the non-homogenous structure and from the other hand, the high frequency of applications, demand high density meshing in order to achieve accurate response [1–3].

The explained method in this paper enables us to design a Chebyshev band passes filter by coaxially placing high-Q $TM_{01\delta}$ dielectric resonators in a cutoff circular waveguide. In the presented work, discussions are made regarding high-Q resonators and inter-resonator coupling. It can be used for TE and TM modes of dielectric resonators. The advantages of such a method are simplicity and accuracy. Furthermore, this method is very quick in calculating resonant frequencies of a single dielectric resonator and coupling factor between two dielectric resonators. Compared with HFSS software, the total required time in this method is less than 1 percent.

Based on the presented method, a DR filter is designed, implemented and fabricated and the results are provided. The fabricated filter has an exclusive feature, i.e., it contains no screw for frequency tuning and it gives out the desired result without a need to modification.

1. INTRODUCTION

Dielectric Resonator filters have found widespread applications in microwave systems due to their exclusive features such as extremely high Q, low bandwidth (less than 2%), inexpensiveness, high power tolerance and high temperature stability. High frequency radio digitals and satellite telecommunications systems can be mentioned as instances of DR filters applications [4–6].

Designing such filters includes appropriately choosing resonators and coupling factors. This, by itself, demands a complicated process and resource. In this paper, a novel and accurate method is presented for calculation of coupling factor between two dielectric resonators which leads to rapid computations. Using such a method, a filter has been designed and fabricated in S band and the measurements are given, as well.

Many elaborator numerical methods can be applied in DR filter design such as mode-matching [7], finite-element [8], boundary element [9], transmission line [10], finite-difference time domain and finite difference frequency domain [11–15].

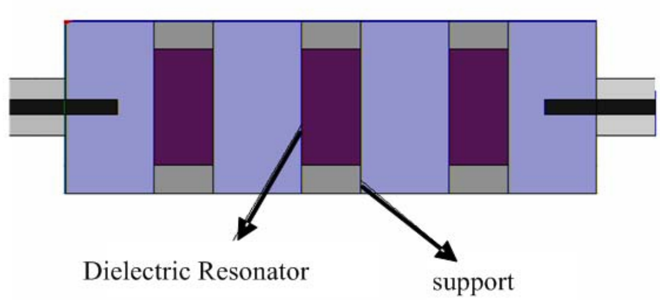

Figure 1. Configuration of dielectric resonator filter.

Figure 1 shows the structure of dielectric resonator filter. Since the configuration is three-dimensional and non homogeneous, mode matching and boundary element methods are not efficient approaches. In contrast, finite-difference time-domain (FDTD), finite-difference frequency-domain (FDFD), transmission line and finite element methods may provide accurate results.

A three dimensional non orthogonal FDTD method has been proposed in [17]. Time-domain processing and fast Fourier transform are needed to reduce the number of iterations. Frequency-domain finite-difference method has been used by Guillon et al. [4, 20]. However, only the TE and TM modes are treated [4]. For the structure in question, the full wave approach (FEM, FDTD, TLM, FDFD, MM, etc) are very intricate and time-consuming.

The TE and TM modes of general configuration have been analyzed in [18, 19], using the finite-element method. It should be noted that the analysis of hybrid modes is only involved in [16] for open dielectric resonators.

2. RESONANT FREQUENCY

The analyzed structure of the dielectric resonator cavity is illustrated in Figure 2. The perfect conductor is a metallic cylindrical cavity of radius c_2 and height h_2, loaded axiomatically with a pillbox dielectric rod of radius c_1, height h_1 and dielectric constant ε_{rd}. The dielectric is supported with a maintainer of radius c_2, height h_1 and dielectric constant ε_b. Because of the low ε_r of the maintainer, it can be neglected in analysis. However, in this work, all the computations have been performed in the presence of the characteristics of the maintainer with the purpose of increasing the accuracy of the method.

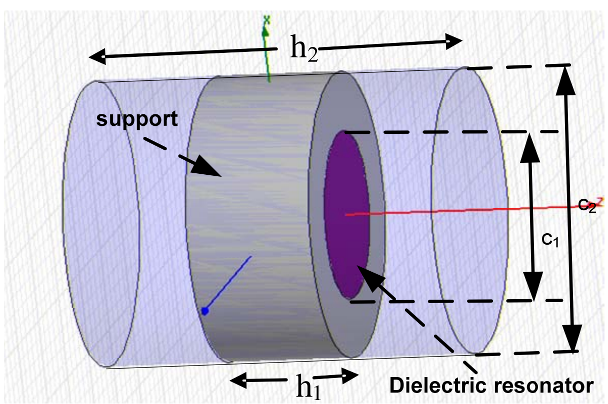

Figure 2. Configuration of dielectric resonator cavity.

It can be readily seen that for the azimuth invariant modes, the azimuth field H_p is decoupled from the other two fields, H_r and H_z.

Similar situation holds for the electric fields. Thus, there exist TM and TE modes, of which the non vanishing fields are $(H_\varphi \cdot E_r \cdot E_z)$ and $(E_\varphi \cdot H_r \cdot H_z)$ respectively [21, 22].

To find the resonant frequency, the eigenvalue of the Helmholtz equation in this structure should be calculated [23].

$$\nabla^2 H_Z + K^2 H_Z = 0 \qquad \text{TE} \tag{1a}$$

$$\nabla^2 E_Z + K^2 E_Z = 0 \qquad \text{TM} \tag{1b}$$

The solution is complex and that why Maxwell equation should be used [24]:

$$\nabla \times E = -j\omega\mu H \tag{2a}$$

$$\nabla \times H = j\omega\varepsilon E \tag{2b}$$

In TE and TM modes, vanishing fields are $(H_\varphi \cdot E_r \cdot E_z)$ and $(E_\varphi \cdot H_r \cdot H_z)$ respectively. So Equation (2a) can be rewritten as

$$\frac{1}{r}\frac{\partial E_Z}{\partial \varphi} - \frac{\partial E_\varphi}{\partial z} = -j\omega\mu H_r \tag{3a}$$

$$\frac{\partial E_r}{\partial z} - \frac{\partial E_z}{\partial r} = -j\omega\mu H_\varphi \tag{3b}$$

$$\frac{\partial E_r}{\partial z} - \frac{\partial E_z}{\partial r} = -j\omega\mu H_\varphi \tag{3c}$$

And the Equation (2b) would be

$$\frac{1}{r}\frac{\partial H_z}{\partial \varphi} - \frac{\partial H_\varphi}{\partial z} = j\omega\varepsilon E_r \tag{4a}$$

$$\frac{\partial H_r}{\partial z} - \frac{\partial H_z}{\partial r} = j\omega\varepsilon E_\varphi \tag{4b}$$

$$\frac{1}{r}\frac{\partial (rH_\varphi)}{\partial r} - \frac{1}{r}\frac{\partial H_r}{\partial \varphi} = j\omega\varepsilon E_z \tag{4c}$$

For the TM mode, the governing equation in homogeneous regions is

$$\frac{\partial^2 (rH_\varphi)}{\partial Z^2} - \frac{1}{r}\frac{\partial (rH_\varphi)}{\partial r} + \frac{\partial^2 (rH_\varphi)}{\partial r^2} = -K^2((rH_\varphi)) \tag{5}$$

And for the TE mode

$$\frac{\partial^2 (rE_\varphi)}{\partial Z^2} - \frac{1}{r}\frac{\partial (rE_\varphi)}{\partial r} + \frac{\partial^2 (rE_\varphi)}{\partial r^2} = -K^2 (rE_\varphi) \tag{6}$$

The resonant frequency of TE or TM mode and the coupling factor between two dielectrics can be generated by solving Equation (5) for TM mode and Equation (6) for TE mode and finding the eigen values of the matrix.

The boundary conditions at permittivity discontinuity are [25]

$$(rH_\varphi)_1 = (rH_\varphi)_2 \tag{7a}$$

$$\frac{1}{\varepsilon_1}\frac{\partial\,(rH_\varphi)}{\partial n} = \frac{1}{\varepsilon_2}\frac{\partial\,(rH_\varphi)}{\partial n} \tag{7b}$$

and

$$(rE_\varphi)_1 = (rE_\varphi)_2 \tag{8a}$$

$$\varepsilon_1\frac{\partial\,(rE_\varphi)}{\partial n} = \varepsilon_2\frac{\partial\,(rE_\varphi)}{\partial n} \tag{8b}$$

And the associated boundary conditions at perfect electric are

$$E_\varphi = 0 \tag{9}$$

$$\frac{\partial(rH_\varphi)}{\partial n} = 0 \tag{10}$$

The associated boundary conditions at the perfect magnetic walls are

$$H_\varphi = 0 \tag{11}$$

$$\frac{\partial rE_\varphi}{\partial n} = 0 \tag{12}$$

Using the above equations and the mentioned boundary conditions, the equations of the field can be solved.

3. COUPLING FACTOR

As it was previously discussed, in order to design a DR filter, the coupling factor needed for the filter must be computed. In this job, the required coupling factor is calculated based on 3rd order Chebyshev filter relations. It is also needed that the distance between two resonators is being calculated in a way that the desired coupling factor is achieved. For this reason, the graph of the coupling factor versus the distance two resonators has to be found.

The configuration of coupled resonators to be analyzed is exhibited in Figure 3. Two dielectric rod resonators having relative permittivity ε_r, diameter c_1 and length h_1 are arranged coaxially in a cutoff circular waveguide of diameter c_2 with dielectric rings of relative permittivity

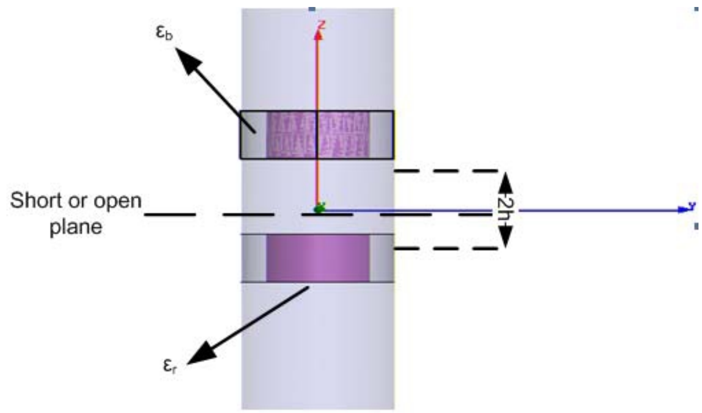

Figure 3. Coupled dielectric rod resonator.

ε_b ($\varepsilon_b \ll \varepsilon_r$). The space between rods is $2h$. It is assumed that the TM mode in the circular waveguide is evanescent, that is

$$\frac{\pi f_0 c_2}{C} < 2.405$$

where C is the light velocity in vacuum. The coupling coefficient of coupled TM dielectric resonators, K, is given by [26–32].

$$K = \frac{f_{op}^2 - f_{sh}^2}{f_{op}^2 + f_{sh}^2} \qquad \text{TM} \tag{13}$$

And the coupling coefficient of coupled TM dielectric resonators, K, is given by

$$K = \frac{f_{sh}^2 - f_{op}^2}{f_{sh}^2 + f_{op}^2} \qquad \text{TE} \tag{14}$$

where f_{op} and f_{sh} correspond to the resonance frequencies, when the structurally symmetric T plane shown in Figure 3 is open and short circuit respectively. Figure 4 shows the structure used to calculate the resonance frequency between two dielectric resonators with $2h$ space between dielectrics. In this figure, theoretically, the length of s should move toward infinity. However, in practice, it is sufficient that the length is chosen to be 4 times larger than dielectric width.

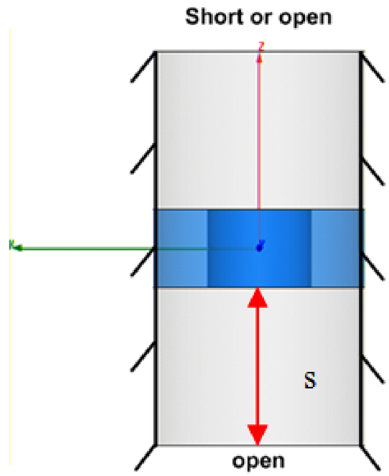

Figure 4. Structure for calculating coupling factor.

4. CHEBYSHEV FILTER THEORY

For the new filter design, design equations for dielectric resonator filter have been used. The first step is choosing an appropriate filter type compatible with the application and calculating its parameters. In this approach, we have chosen 3rd order Chebyshev filter with 2% bandwidth.

The filter can be designed using the following equations [33, 34].

$$
\begin{aligned}
g_0 &= 1, \qquad\qquad g_1 = \frac{2a_1}{\gamma} \\
g_k &= \frac{4a_{k-1}a_k}{b_{k-1}g_{k-1}}, \qquad k = 2, 3, \ldots, n \\
g_{n+1} &= 1 \qquad\qquad\quad \text{for } n \text{ odd} \\
g_{n+1} &= \coth^2\left(\frac{\beta}{4}\right)
\end{aligned}
$$

where

$$
\begin{aligned}
a_k &= \sin\left[\frac{(2k-1)\pi}{2n}\right], \qquad k = 1, 2, \ldots, n \\
b_k &= {}^2 + \sin^2\left[\frac{k\pi}{n}\right], \qquad k = 1, 2, \ldots, n \\
\beta &= \ln\left[\coth\left(\frac{A}{2 \times 8.686}\right)\right], \qquad \gamma = \sinh\left(\frac{\beta}{2n}\right)
\end{aligned}
$$

where A is the pass band ripple in decibels. The relation between A and ε is given by

$$A = 10\log(1 + \varepsilon^2)$$

Depending on the order of the filter, the termination resistance could be different from the source resistance. The bandpass design parameters Q_e is calculated using [35].

$$Q_{e1} = \frac{g_0 g_1}{FBW}$$
$$Q_{en} = \frac{g_n g_{n+1}}{FBW}$$

where Q_{e1} and Q_{en} are external quality factors of the resonator at the input and output.

5. FILTER DESIGN

The $\mathrm{TM}_{01\delta}$ resonators used in this filter structure were fabricated from low loss ceramics ($\varepsilon_r = 82$, QF = 9500) and Plexiglas supports ($\varepsilon_b = 1.2$ tg$\delta = 0.001$). High Q design of these resonators was performed as described below.

Solving (5) in the structure illustrated in Figure 4 and computing the eigen values of that equation, the resonance frequency can be calculated in TM mode in two states of perfect electric and perfect magnetic for different distances.

Figure 5 shows the resonance frequency graph for perfect electric (f_{sh}) and perfect magnetic (f_{op}) versus distance. Figure 6 shows the coupling factor between two dielectrics with $2h$ spacing.

Derived from graph of Figure 5 and Equation (5), coupling factor graph versus the distance between two resonators can be sketched as shown in Figure 6. Based on the above equations and graphs, a 3rd order band pass Chebyshev filter is constructed without tuning screw.

Figure 7 demonstrates the designed Chebyshev filter using the graph in Figure 6. This filter has been also simulated with HFSS and the result is depicted in Figure 8.

Figure 8 shows the transitions and reflection factor of the 3rd order Chebyshev band pass filter. It is seen that the insertion loss is lesser than 0.5 dB and the reverse loss is lesser than 20 dB. The bandwidth of the filter is 1.25% in center frequency of 3.2 GHz. It is understood that the designed filter using graph in Figure 6 has given desirable results in simulation.

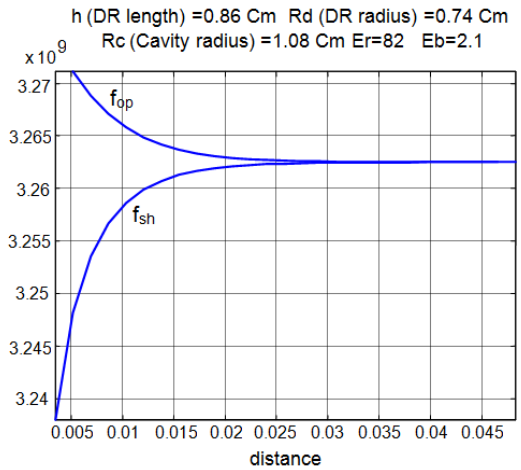

Figure 5. Calculated results of f_{sh}, f_{op}.

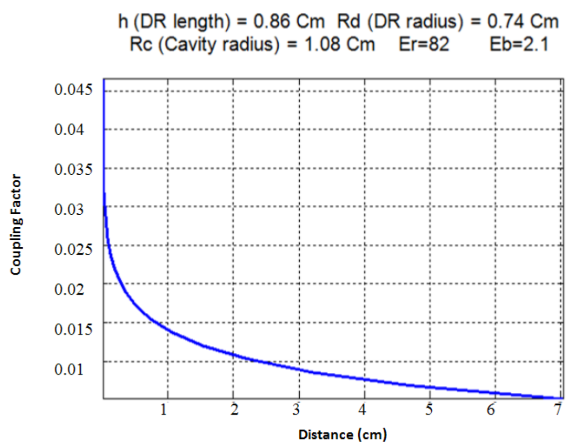

Figure 6. Calculation of coupling factor.

6. MEASUREMENT RESULTS

The designed filter was fabricated and its parameters were measured using hp 8756 network analyzer. The return loss and insertion loss are shown in Figure 9.

The depicted graph of Figure 9 reveals that the insertion and return losses are lesser than 0.5 dB and 25 dB, respectively. It is worth mentioning that a 60 MHz drift in the center frequency has occurred

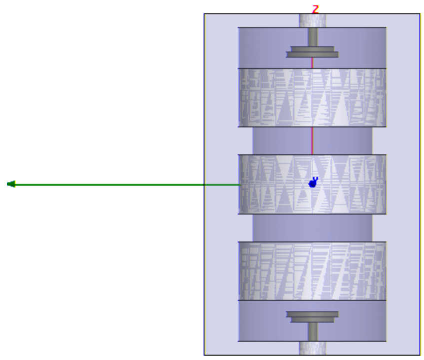

Figure 7. Filter implementation.

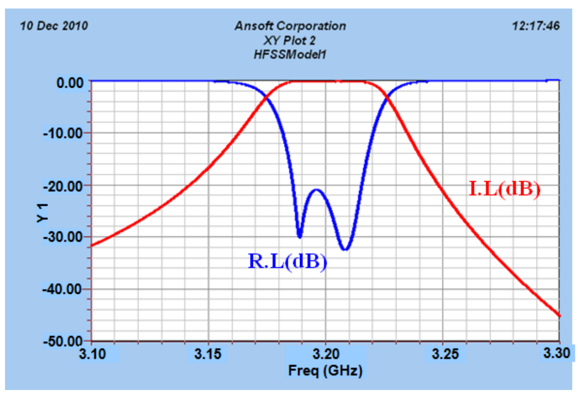

Figure 8. Transitions and reflection factor of the filter.

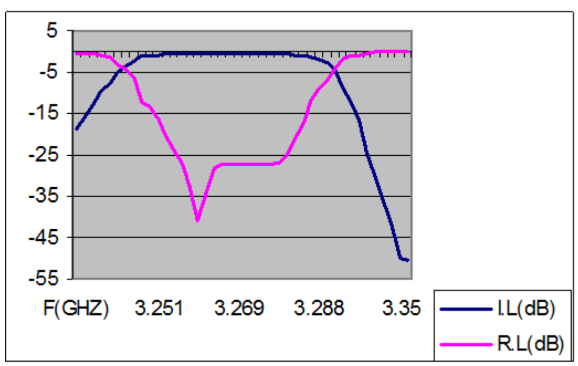

Figure 9. Measurement result.

due to the inaccuracy of the dielectric coefficient of the plaxy glass which has been used as a maintainer.

7. CONCLUSION

We used the analytical and numerical (finite difference) techniques as efficient numerical procedures. The total time consumed for this calculation is less than 20 minutes compared to the time needed for a simulation by means of HFSS software which takes 2880 minutes. The filter which has been designed with these methods has low loss characteristics (less than 0.5 dB).

This filter has been fabricated without any tuning screw and the filter response demonstrates desirable matching with simulations.

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