Design of Nonradiative Dielectric Waveguide Filters

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Abstract —**An efficient design technique of nonradiative dielectric waveguide filters for use at millimeter wavelengths is developed. Filter structures considered here are a gap-coupled type and an alternating-width type. According to present theory, 3-pole, 0.1-dB Chebyshev ripple bandpass filters with a 2-percent bandwidth at a center frequency of 49.5 GHz were designed and fabricated with Teflon dielectric. Calculated and measured filter responses agree quite well, and excess insertion losses are found to be as small as 0.3 dB for both types of the fabricated filter circuits.**

I. Introduction

VARIOUS FILTER circuits have been proposed for microwave and millimeter-wave applications [1]–[3]. Filter structures based on the nonradiative dielectric waveguide (NRD-guide) [4] are also promising because of their applicability to millimeter-wave integrated circuits. Actually, an NRD-guide filter has been previously fabricated and tested at 50 GHz [5]. But, since there was no reliable design theory at that time, the result was not necessarily satisfactory, and the excess insertion loss was unexpectedly as large as 1 dB due to the lack of optimization in design parameters. An efficient design technique is needed to deduce the full potentiality of the NRD-guide filter.

One of the filter structures considered here consists of longitudinally coupled dielectric chip resonators inserted between the input and output ports, as shown in Fig. 1. The synthesis technique developed for gap-coupled coplanar waveguide filters [3] can also be applied to the present case if the coupling property between adjacent dielectric sections is known. Analysis of the coupling can be elucidated by means of a variational technique. The air gaps in this filter structure can be replaced with below-cutoff narrow dielectric strips joining adjacent resonators (see Fig. 7). Such a modification is of practical interest, since the whole dielectric structure can be built in one rigid body. The latter type of NRD-guide filter will be called an alternating-width filter in this paper.

In order to verify the theory and to check filter performance as well, bandpass filters were designed, fabricated, and tested at a center frequency of 49.5 GHz. It was found that the specified filter response was realized with an excess insertion loss of 0.3 dB for both types of filter circuits investigated here.

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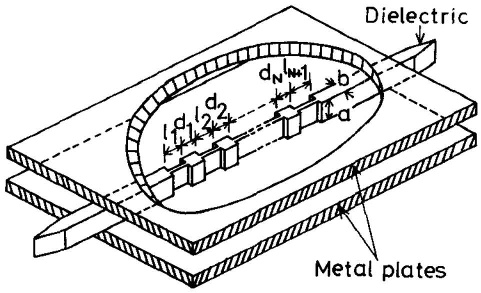

Fig. 1. Structure of gap-coupled NRD-guide filter.

II. Design Considerations of Gap-Coupled NRD-Guide Filters

A. Equivalent Circuit Representation of Gap-Coupled Dielectric Strips

The key elements in the gap-coupled NRD-guide filter structure are dielectric sections coupled by way of an air gap which can be idealized by a pair of semi-infinite dielectric strips longitudinally arranged with an air gap, as shown in Fig. 2(a). Such simplification is admissible because all modes except for the dominant one decay away from the gap region very quickly and can hardly reach the far end of the chips due to the inherent cutoff nature of the metal plates separated by a distance of less than half a wavelength. The gap region can be represented by an impedance inverter network with transmission-line sections connected on both sides, as shown in Fig. 2(b) [3]. The inverter parameter K and the electrical line length ϕ are given by

$$K = \left|\tan \tfrac{1}{2}\left(\tan^{-1} b_s - \tan^{-1} b_0\right)\right| \tag{1a}$$

$$\phi = -\pi - \tan^{-1} b_s - \tan^{-1} b_0 \tag{1b}$$

where b_s and b_0 are normalized susceptances at the truncated end of a dielectric strip looking toward an electric and a magnetic wall boundary, respectively, assumed in the transverse midplane of the air gap. Therefore, the discontinuity problem in the NRD-guide has to be solved in order to evaluate the normalized susceptances. Fortunately, some of the mathematical techniques developed for metal

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waveguide analysis can also be applied to the NRD-guide, since field configurations in both the waveguides are quite similar. The variational method is certainly one such technique and is chosen here because of its popularity.

B. Variational Expressions for Normalized Susceptances

A below-cutoff narrow dielectric strip instead of the air gap is assumed in the analysis for purposes of generalization. The dielectric strip is a in height, b and c in width in the normal and thinned portions, respectively, and the length of the thinned portion is l. According to image theory, a conducting plate may be placed longitudinally along the vertical midplane of a dielectric structure. Another conducting plate may also be placed to cover the side opening of the NRD-guide at a distance of $D/2$ far from the image plane without disturbing field distribution appreciably. Thus, the solution of the original structure is reduced to that for step discontinuities in the partially filled rectangular waveguides, as shown in Fig. 3. Similar problems in which a partially filled and an empty rectangular guide of semi-infinite length are connected with each other have been treated elsewhere [6], [7]. These theories, however, have limited consideration to either LSM-type modes [6] or LSE-type modes [7]. The waveguide geometry of Fig. 3 allows LSE modes to be generated at the junction between the normal and thinned dielectric strips for an incident LSM mode. In addition, the finite length of the narrow dielectric strip has to be taken into account in the present analysis. Considering these facts, the variational expression for the normalized susceptance b_s for the electric wall boundary placed at the transverse midplane of the below-cutoff region can be expressed as follows:

$$b_s = \frac{\iint \bar{z}\times E(\rho)\cdot \tilde{G}(\rho|\rho')\cdot \bar{z}\times E(\rho')\,d\rho\,d\rho'}{jY_1^{(1)'}\left[\int \bar{z}\times E(\rho)\cdot h_1^{(1)'}(\rho)\,d\rho\right]^2} \tag{2}$$

where

$$\begin{aligned}\tilde{G}(\rho|\rho') = &\sum_{n=3,5,\cdots} Y_n^{(1)'}h_n^{(1)'}(\rho)h_n^{(1)'}(\rho') \\ &+\sum_{n=1,3,\cdots}\Bigg[Y_n^{(1)''}h_n^{(1)''}(\rho)h_n^{(1)''}(\rho') \\ &+\coth\frac{\gamma_n^{(2)'}l}{2}\,Y_n^{(2)'}h_n^{(2)'}(\rho)h_n^{(2)'}(\rho') \\ &+\coth\frac{\gamma_n^{(2)''}l}{2}\,Y_n^{(2)''}h_n^{(2)''}(\rho)h_n^{(2)''}(\rho')\Bigg].\end{aligned} \tag{3}$$

In the above equations, $\bar{z}$ is the unit vector in the longitudinal direction, $E(\rho)$ the electric field assumed at the junction between dielectric strips of normal and narrow width, $h(\rho)$ the nth vector mode function of magnetic type, Y_n the mode admittance, γ_n the mode propagation constant, and integration is carried out over the entire cross-sectional area at the junction. Furthermore, superscripts (1) and (2) are associated with the waveguide portions filled with normal and thinned dielectric strips,

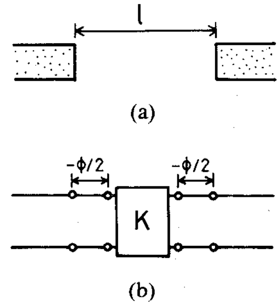

Fig. 2. (a) A pair of gap-coupled semi-infinite dielectric strips and (b) its equivalent inverter circuit with additional transmission-line sections connected on both sides.

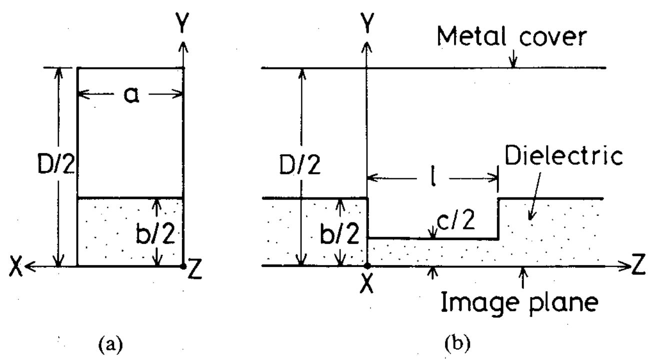

Fig. 3. Analytical model for gap-coupled semi-infinite dielectric strips in NRD-guide. (a) Front view and (b) side view.

respectively, and the single and double primes denote quantities related to the LSM and LSE modes, respectively. With this scheme of notation, the mode admittance and the mode propagation constant are related by

$$Y_n' = j\frac{\omega\epsilon_0}{\gamma_n'} \qquad Y_n'' = \frac{\gamma_n''}{j\omega\mu_0} \tag{4}$$

where ω is angular frequency, ϵ_0 the free-space permittivity, and μ_0 the free-space permeability. Explicit expressions for the mode functions and the corresponding characteristic equation of LSM modes are given in [6], while those of LSE modes are given in [7]. The variational expression of the other normalized susceptance b_0 can be obtained by replacing "coth" function in (3) with "tanh" function.

The trial function $E(\rho)$ is represented in terms of a linear combination of the LSM- and LSE-mode functions appropriate to the above-cutoff waveguide portion. The first six to twelve LSM-mode functions and the first LSE-mode function are found to be sufficient to obtain accuracy of four digits depending on the cross-sectional dimensions of the dielectric strip.

C. Design of Gap-Coupled NRD-Guide Filter

The gap-coupled filters will be considered first ($c=0$ in Fig. 3). Substituting the computed normalized susceptances into (1) yields the inverter parameter and the electrical line length as a function of the gap length, as shown in Fig. 4. In this calculation, Teflon strips ($\epsilon_r = 2.04$) 2.7 mm in

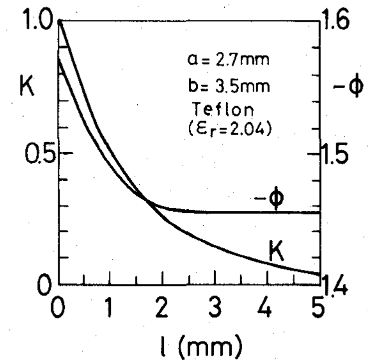

Fig. 4. Computed inverter parameter and electrical line length of the equivalent circuit for gap-coupled semi-infinite dielectric strips as a function of the gap length.

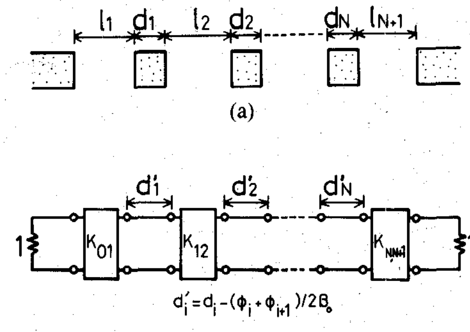

Fig. 5. (a) Gap-coupled filter and (b) its equivalent circuit representation.

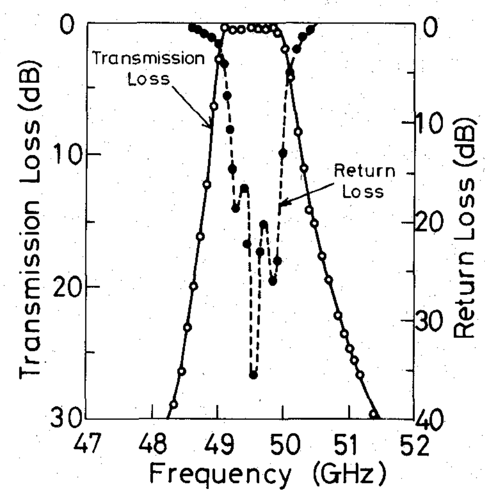

Fig. 6. Measured transmission loss and return loss of gap-coupled filter. Theory (———). Measured (○, --●--).

height and 3.5 mm in width and a frequency of 49.5 GHz are assumed for later reference.

By replacing each gap in the original filter structure with the equivalent impedance inverter, a canonical bandpass filter circuit can be derived, as shown in Fig. 5. Based on this circuit representation, design of the filter can be carried out as described below. First, calculate the required inverter parameters using the tabulated low-pass prototype filter parameters, the specified fractional bandwidth, and the reactance slope parameter [8] given by

$$x=\frac{\pi}{2}\left(\frac{\sqrt{\epsilon_r}k_0}{\beta_0}\right)^2\frac{2(q_0^2+\epsilon_r p_0^2)+p_0 b(q_0^2+\epsilon_r^2 p_0^2)}{2\epsilon_r(q_0^2+p_0^2)+p_0 b(q_0^2+\epsilon_r^2 p_0^2)} \quad (5)$$

where k_0 is the free-space wavenumber, β_0 is the phase constant of the dominant NRD-guide mode, and p_0 and q_0 are the lowest eigenvalues of the characteristic equations

$$p_0=\frac{1}{\epsilon_r}q_0\tan\frac{q_0 b}{2} \quad (6a)$$

$$p_0^2+q_0^2=(\epsilon_r-1)k_0^2. \quad (6b)$$

Then, determine the gap lengths by equating (1a) with the inverter parameters computed above, and finally find dielectric chip lengths by requiring that the electrical line lengths between the adjacent impedance inverters be π, that is,

$$d_i=\frac{1}{\beta_0}\left\{\pi+\frac{1}{2}(\phi_i+\phi_{i+1})\right\}. \quad (7)$$

Thus, all the design parameters needed to realize the specified shape of response are completely determined.

A 3-pole, 0.1-dB Chebyshev ripple bandpass filter with a 2-percent bandwidth at a center frequency of 49.5 GHz was built using Teflon strips which had the same dimensions as those assumed in Fig. 3. The calculated filter dimensions are $d_1=d_2=d_3=2.72$ mm, $l_1=l_4=1.60$ mm, and $l_2=l_3=3.5$ mm (see Fig. 5 for definition). The excess insertion loss of this filter is estimated to be 0.28 dB from a theoretical unloaded Q-factor of the Teflon NRD-guide resonator, which has been calculated to be 3400 [9].

III. Characteristics of Fabricated Gap-Coupled NRD-Guide Filter

In the initial measurement, the 3-dB bandwidth of the fabricated filter was found to be wider by 70 MHz than the specified value of 1 GHz. This discrepancy could be eliminated by expanding the inner two gap lengths, l_2 and l_3, by 50 μm, while keeping other dimensions unchanged. The measured filter response is shown in Fig. 6, together with the return loss curve, which has the highest value of 38 dB at the center frequency. The computed response curve is also plotted for comparison. Agreement between theory and measurements is quite satisfactory. The excess insertion loss was measured and found to be 0.3 dB, which is very close to the theoretically predicted value of 0.28 dB. The fabricated NRD-guide filter becomes radiative above 55.6 GHz, as the plate separation of 2.7 mm implies. Though failed due to the limited frequency range of the klystron used, measurements should have been made in the radiative range of frequency as well to demonstrate the actual extent of the stopband.

Another filter structure, the alternating-width type, which can be built in one rigid body by joining chip resonators with below-cutoff narrow strips, as shown in Fig. 7, was also designed and tested. This filter is of practical interest

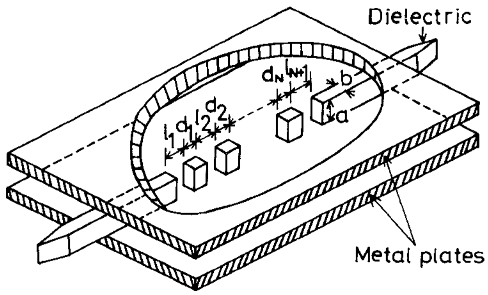

Fig. 7. Structure of alternating-width filter.

since the relative positions of the dielectric resonators are kept unchanged, once the dielectric structure is fabricated. It is obvious that the design procedure for the gap-coupled filter circuit holds equally for the alternating-width structure as well, so long as the inverter parameter and the associated electrical line length are computed for the nonvanishing thickness of the below-cutoff strip. By specifying the same filter response as that of the gap-coupled filter, the required dimensions of the dielectric structure are found to be $d_1 = d_3 = 2.04$ mm, $d_2 = 2.00$ mm, $l_1 = l_4 = 2.45$ mm, and $l_2 = l_3 = 4.95$ mm. Calculated and measured responses are shown in Fig. 8. The response shape is the same as that of the gap-coupled type shown in Fig. 6 in the near-midband frequency range, but the difference between the two cases becomes rather notable for other frequencies. The response curve for the alternating-width filter tends to bend toward the higher frequency side at the skirts compared to that of the gap-coupled filter. The excess insertion loss was found to be 0.3 dB in this case as well.

IV. Conclusions

Low-loss NRD-guide bandpass filters were designed and fabricated with Teflon strips at a center frequency of 49.5 GHz. Measurements showed that the desired filter performance could be achieved for both the gap-coupled and the alternating-width type filter. The excess insertion losses of these filters were experimentally found to be 0.3 dB at a center frequency of 49.5 GHz. This supports the assertion

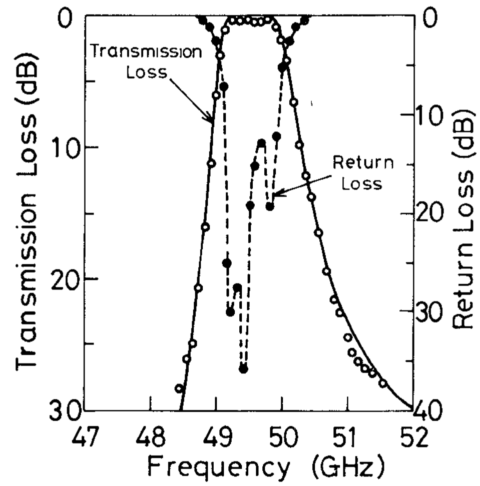

Fig. 8. Measured transmission loss and return loss of alternating-width filter. Theory (———). Measured (○, --●--).

that the NRD-guide filter is of practical use at millimeter wavelengths.

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