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MICROSTRIP FILTER DESIGN INCLUDING DISPERSION EFFECTS AND RADIATION LOSSES

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ABSTRACT

A numerical technique for the design of parallel-coupled bandpass microstrip filters is discussed. The method accounts accurately for the dispersion effects due to the presence of the substrate, the associated surface wave propagation and radiation losses. The presented technique does not have any frequency limitations and can be applied to various microstrip filters. Comparison with experimental results shows excellent agreement.

SUMMARY

Introduction

This work presents the analysis and synthesis of microstrip bandpass filters (Fig. 1). The implemented method of analysis is based on solving Pocklington's integral equation by employing the method of moments and accounts for conductor thickness, dispersion as well as radiation effects. The basic advantage of the method is the accurate evaluation of losses due to surface and radiation waves, a major problem in the design of microstrip filters in millimeter-wave frequencies.

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Microstrip filters find extensive applications in monolithic arrays [1]. The methods previously adopted in the design of microstrip filters fail in high frequencies where the substrate thickness becomes larger than a few hundreds of the wavelength. The reasons for the inaccuracies are the dispersion effects and radiation losses not considered by these techniques. To overcome the limitation to electrically thin substrates an integral equation approach has been suggested and has provided excellent results for microstrip antennas and microstrip discontinuities in high frequencies [5]-[6].

Formulation and Results

This paper applies the integral-equation method to the analysis and design of microstrip parallel-coupled bandpass filters. The approach presented is based on solving Pocklington's equation for the unknown current density on the strip conductors by employing the method of moments. The integral equation for the electric field is shown below

$$\bar{E}(\bar{r}) = \iint_S (k_0^2 \bar{\bar{I}} + \bar{\nabla}\bar{\nabla}) \cdot \bar{\bar{G}} \cdot \bar{J} \, ds \tag{1}$$

where $\bar{\bar{I}}$ = the unit dyadic,

$\bar{\bar{G}}$ = the dyadic Green's function for an infinitesimally small printed dipole,

$\bar{J}$ = the unknown current distribution and

s = the surface of the strip conductors.

The model developed here also accounts for conductor thickness and it assumes that the transmission-line and coupler widths are much smaller than the wavelength. The latter assumption insures that the currents are nearly unidirectional. For the analysis of a given filter the method of moments is applied to determine the current distribution in the longitudinal direction, while the longitudinal current dependence in the transverse direction is chosen to satisfy the edge condition at the effective width location [5]. Upon determining the current distribution, transmission line theory is invoked to determine the elements of the scattering matrix of the filter. For the evaluation of the scattering coefficients the two transmission lines are excited in-phase and out-of-phase resulting in two modes of excitation, even and odd. These modes are characterized by propagation constants β_e, β_o and normalized self impedances Z_e, Z_o, respectively [6]. The two impedances are evaluated with respect to reference planes at the same distance from the open ends of the transmission lines. The elements of the normalized impedance matrix are evaluated by the following expressions:

$$Z_{11} = \frac{Z_o + Z_e}{2} , \tag{2}$$

$$Z_{12} = \frac{Z_o - Z_e}{2} . \tag{3}$$

These values of Z_{11} and Z_{12} are then used for the evaluation of the scattering coefficients as shown below

$$S_{11} = \frac{Z_{11}^2 - Z_{12}^2 - 1}{(Z_{11} + 1)^2 - Z_{12}^2} , \tag{4}$$

$$S_{12} = \frac{2Z_{12}}{(Z_{11} + 1)^2 - Z_{12}^2} .$$

The computation of these two coefficients permits prediction of the filter response. Specifically, $|S_{11}|^2$ is the normalized reflected power while $|S_{12}|^2$ is the normalized transmitted power. For a filter without any dielectric or conductor losses the following equation is satisfied

$$|S_{11}|^2 + |S_{12}|^2 + P_r = 1 \quad , \tag{6}$$

where P_r is the power radiated in the form of surface and space waves.

Because of Eq. (6), the insertion loss defined as

$$L_{in} = 10 \log \frac{1}{1 - |S_{11}|^2} \tag{7}$$

cannot describe the response of the filter. For the complete characterization of the device two more parameters are defined; the radiation loss L_r,

$$L_r = 10 \log \frac{1}{1 - |S_{11}|^2 - |S_{12}|^2} \tag{8}$$

and the transmission loss L_t

$$L_t = 10 \log \frac{1}{|S_{12}|^2} \quad . \tag{9}$$

From Eqs. (7) and (9) one could see that in lossless filters $L_{in} = L_t$ and $L_r = 0$.

As an example, results for three different parallel-coupled, bandpass filter geometries are presented (Fig. 2). In Fig. 3, the transmission loss L_t is plotted for a 10.25 GHz two-section filter on a 0.635 mm alumina substrate with w/h = 1, $L_1 = L_2 = L_3$ = 2.54 mm, $d_1 = d_3$ = 0.762 mm and d_2 = 1.259 mm. This filter has a 3.0 dB bandwidth of 15.35 percent with a ripple of 1.2 dB. The theoretical results are in very good agreement with the experimental ones performed at Hughes Aircraft Company, Torrance (Fig. 4). The bandwidth of this device can be increased by adding more couplers of slightly different widths. A number of theoretical and experimental results will be presented which will reveal the effect of various parameters on the bandwidth. As a second example, Figs. 5 and 6 show the scattering parameters in amplitude and phase of a 41 GHz two-section filter on a 0.159 mm alumina substrate with w/h = 1, $L_1 = L_2 = L_3$ = 0.635 mm, $d_1 = d_3$ = 0.190 mm and d_2 = 0.317 mm. The performance of the filter is very good even in these high frequencies because of the small substrate thickness. Since in practice it is not easy to fabricate such thin dielectric substrates, it is very interesting to investigate the performance of these filters on thicker substrates. For this reason, the above structure was printed on a 0.635 mm alumina substrate and its performance in terms of insertion loss L_{in}, transmission L_t and radiated loss L_r is shown in Fig. 7. From this response one can hardly characterize the device under consideration as a filter. Actually, it can be concluded that the performance has been deteriorated badly because of the unacceptably high radiation losses. In Fig. 8, the normalized radiated power for the two high-frequency devices is plotted as a function of frequency.

A wealth of theoretical and experimental results will be presented for most of the structures of Fig. 1 with the emphasis placed on the development of a design scheme.

Conclusions

The application of an integral-equation approach has resulted in a very accurate analysis of parallel-coupled bandpass filters. The advantage of the technique is that it is quite general and can be used very efficiently to describe the effects of various parameters on the performance of filters. With this method as a basis a very accurate iterative procedure can be used for the design of millimeter-wave filters without neglecting dispersion effects and radiation losses.

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