The Microwave Bandpass Filter

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Microwave 帶域通過 濾波器

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Summary

The rectangular waveguide direct-coupled cavity filter and the interdigital filter be designed and constructed to meet virtually any frequency selective amplitude specification with, if required, possible prescribed type of phase characteristics. The appropriate design techniques are described for the two most common classes of filter by giving closed form design equations enabling the filters to be designed using only a hand calculator and appropriate tables and graphs.

1. Introduction

Microwave filters may be conveniently divided into four groups classed in accordance with the type of propagation used and bandwidth of operation. We shall consider devices supporting a TEM mode for both narrow and broadband operation and then those using waveguide modes (Rhodes, 1974, 1975; Levy, 1975; Atia et al, 1974) belonging to similar categories. Of the TEM components, the low-pass stepped impedance filter (Levy, 1965; Rhodes, 1976) and the bandpass interdigital and com-line filters (Rhodes, 1970; Levy et al, 1971) are probably the most important and readily designed for relatively narrow band operation using the explicit formulas for prototype maximally flat and Chebyshev filters (Matthaei et al, 1980). The waveguide direct coupled cavity bandpass filter (Williams, 1970) using shunt inductive irises located in a rectangular waveguide occupies a unique position of all microwave filters which are constructed as these devices have represented a very large proportion. Apart from the ease of designing from explicit formulas, accepting the waveguide environment for low loss or high power applications, the simplicity and consequent low cost of manufacture makes these devices attractive and hence are widely used in the commercial and military areas.

The purpose of this paper is to describe the appropriate design techniques for the two most common type of microwave bandpass filters and techniques used in multiplexing bandpass channel filters.

2. Waveguide Direct-Coupled Cavity Filters

In the rectangular waveguide case, the filter comprises of shunt inductive posts located in a uniform waveguide as illustrated in Fig. 1. For narrow bandwidth the design equations for a bandpass Chebyshev amplitude characteristic are (Matthaei et al, 1980)

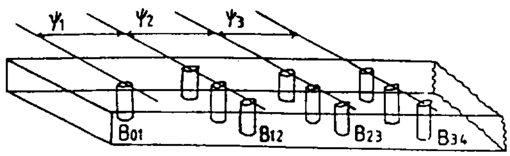

Fig. 1. The rectangular waveguide direct coupled cavity filter.

$$B_{r,r+1} = \frac{\alpha/\sqrt{g_r g_{r+1}}}{K_{r,r+1}} - \frac{K_{r,r+1}}{\alpha/\sqrt{g_r g_{r+1}}} \quad r=0\rightarrow n \qquad (1)$$

$$\Psi r = \pi - \frac{1}{2}\cot^{-1}\left(\frac{B_{r-1,r}}{2}\right) - \frac{1}{2}\cot^{-1}\left(\frac{B_{r,r+1}}{2}\right)$$

$$r=1\rightarrow n \quad \text{with } \alpha = \pi\left(\frac{\lambda g_1 + \lambda g_2}{\lambda g_1 - \lambda g_2}\right) \qquad (2)$$

$$K_{r,r+1} = \frac{\sqrt{\eta^2 + \sin^2\left(\frac{r\pi}{n}\right)}}{\eta} \quad r=0\rightarrow n \qquad (3)$$

$$g_r = \frac{2\sin\left(\frac{(2r-1)\pi}{2n}\right)}{\eta} \quad r=1\rightarrow \eta \qquad (4)$$

$$g_o = g_{n+1} = \frac{1}{\alpha} \qquad (5)$$

$$\eta = \sinh\left(\frac{1}{n}\sinh^{-1}\left(\frac{1}{\varepsilon}\right)\right) \qquad (6)$$

λg_1 and λg_2 are the guide wavelengths at the lower and upper bandedge frequencies respecitively and the overall insertion loss characteristic is approximately given by:

$$L = 10\log\left(1+\varepsilon^2 \, Tn^2\left(\alpha\pi(\lambda_g - \lambda_{go})\right)\right) \qquad (7)$$

and $Tn(x)=\cosh(n\cosh^{-1}x)$ is the Chebyshev polynomial of the first kind and

$$\lambda_{go} = \frac{\lambda g_1 + \lambda g_2}{2}$$

is the guide wavelength corresponding to the electrical length 2π radians enabling the Ψr to be related to physical lengths.

The midband susceptances of the irises Br, r+1 may be converted into physical discontinuities using appropriate graphical information (Saad, 1980).

The dissipation loss in dependent upon the manufacturing and plating techniques but due to the uniform dissipation and normally low passband ripple level, is approximately proportional to the group delay characteristic.

Fine tuning is accomplished by tuning screws located at the maximum electric field points in the center of the cavities and is normally achieved using a waveguide reflectometer arrangement (Rhodes, 1976).

For broader bandwidths the insertion loss becomes (Levy, 1967)

$$L = 10\log\left(1+\varepsilon^2 Tn^2\left(\frac{\alpha\lambda g}{\lambda go}\sin\left(\frac{\pi\lambda go}{\lambda g}\right)\right)\right) \qquad (8)$$

and approximate design equations are as before apart from α and λ_{go} being determined from eq. (8) and eq. (4) becomes

$$g_r = \frac{2\sin\frac{(2r-1)\pi}{2n}}{\eta} - \frac{1}{4\eta\alpha^2}\left(\frac{\left(\eta^2+\sin^2\left(\frac{r\pi}{n}\right)\right)}{\sin\left(\frac{(2r+1)\pi}{2n}\right)} + \frac{\left(\eta^2+\sin^2\left(\frac{(r-1)\pi}{n}\right)\right)}{\sin\left(\frac{2r-3)}{2n}\right)}\right) \quad r=1\rightarrow n \qquad (9)$$

and the selectivity on the high frequency side of the passband deteriorates.

3. Interdigital Filters

An interdigital filter consists of rectangular bars or rods alternatively short circuited to ground located between ground planes as depicted in Fig. 2.

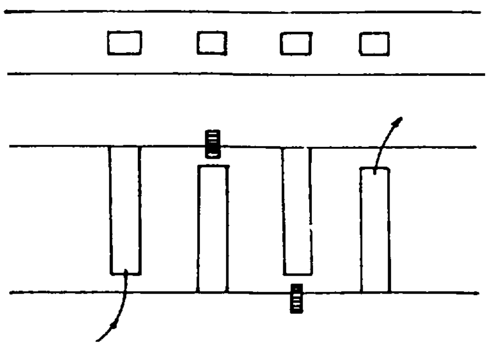

Fig. 2. The interdigital filter.

The tri-diagnoal characteristic admittance matrix defines the structure and from its entries, using appropriate graphs (Getsinger, 1962) the physical dimensions may be obtained. For arrow band-widths the insertion loss characteristic for an equiripple response is given approximately by

$$L = 10\log\left(1+\varepsilon^2\, Tn^2\left(\frac{\cos\omega}{\cos\omega_o}\right)\right) \qquad (10)$$

the passband extending from $\frac{\pi}{2}-\omega_o$ to $\frac{\pi}{2}+\omega_o$. For a physically realizable narrow band filter, redundant transformer elements are necessary at the input and output leading to an n+2 wire line defined by characteristic admittance matrix.

$$[Y] = \begin{bmatrix} Y_{00} & -Y_{01} & 0 & 0 & \cdots\cdots & 0 \\ -Y_{01} & Y_{11} & -Y_{12} & 0 & & \\ 0 & -Y_{12} & Y_{22} & -Y_{23} & & \\ 0 & 0 & -Y_{23} & \ddots & & \\ & & & & \ddots & -Y_{n,n+1} \\ 0 & & & & & Y_{n+1,n+1} \end{bmatrix} \qquad (11)$$

$$Y_{r,r+1} = \frac{\sqrt{\eta^2+\sin^2\left(\frac{r\pi}{n}\right)}}{2\alpha\sqrt{\sin\left(\frac{(2r-1)\pi}{2n}\right)\sin\left(\frac{(2r+1)\pi}{2n}\right)}} \quad r=1\rightarrow n \qquad (12)$$

$$Y_{01} = Y_{n,n+1} = \sqrt{\frac{\eta}{2\sin\left(\frac{\pi}{2n}\right)}} \qquad (13)$$

$$Y_{r,r+1} = 1 \quad r=2\rightarrow n-1 \qquad (14)$$

$$Y_n = Y_{n,n} = 1+\frac{\eta}{2\sin\left(\frac{\pi}{2n}\right)} \qquad (15)$$

$$Y_{0,0} = Y_{n+1,n+1} = 1 \qquad (16)$$

and

$$\alpha = \cos\omega_o \qquad (7)$$

with η defined in eq. (6).

The line is one quarter of a wavelength long at the center frequency and fine tuning is obtained by locating tuning screws at the open circuited ends of the lines as illustrated in Fig. 2 and adjusted using a reflectometer arrangement. Dissipation loss is again proportional to the group delay and inversely proportional to the ground plate spacing. The latter however, is limited by the onset of higher ordered modes occurring when the spacing is approximately one half of a wavelength.

For broad bandwidths, transformer elements are not normally necessary and the insertion loss characteristic is modified to:

$$L = 10\log\left(1+\varepsilon^2\, Cn^2\right) \qquad (18)$$

where

$$Cn = \cosh\left(n-1\right)\cosh^{-1}\left(\frac{\cos\omega}{\cos\omega_o}\right) + \cosh^{-1}\left(\frac{\cot\omega}{\cot\omega_o}\right) \qquad (19)$$

and the elements of the characteristic admittance matrix

$$[Y] = \begin{bmatrix} Y_{11} & -Y_{12} & 0 & 0 & \cdots\cdots & 0 \\ -Y_{12} & Y_{22} & -Y_{23} & 0 & & \\ 0 & -Y_{23} & -Y_{33} & \ddots & & \\ 0 & & & \ddots & \ddots & \\ & & & & & Y_{n,n} \end{bmatrix} \qquad (20)$$

are approximately given by Rhodes (1976).

$$Y_{r,r+1} = \frac{\sqrt{\eta^2+\sin^2(\frac{r\pi}{n})}}{\eta} \qquad r = 1 \to n-1 \qquad (21)$$

$$Y_{r,r} = Y_{n-r+1,\, n-r+1}$$

$$= \frac{2\sin(\frac{(2r-1)\pi}{2n})}{\eta}\left(\frac{1}{\alpha} - \frac{2\alpha}{2n}\right)$$

$$-\frac{\alpha}{4\eta}\left(\frac{(\eta^2+\sin^2(\frac{r\pi}{n}))}{\sin(\frac{(2r-1)\pi}{2n})} - \frac{(\eta^2+\sin^2(\frac{(r-1)\pi}{n}))}{|\sin(\frac{(2r-3)\pi}{2n})|}\right)$$

$$r = 1 \to [\tfrac{n}{2}] \qquad (22)$$

with $\alpha = \cos\omega_0$

For alternative physical realizations of the same electrical characteristic, internal rows and columns of the characteristic admittance matrix may be scaled, i.e.

$$[Y^1] = \begin{bmatrix} Y_{11} & -Y_{12}n_2 & 0 & 0 & & \\ -Y_{12}n_2 & Y_{22}n_2^2 & -Y_{23}n_2n_3 & & & \\ 0 & -Y_{23}n_2n_3 & Y_{33}n_3^2 & -Y_{34}n_3n_4 & & \\ & & \ddots & & -Y_{n-1,n}n_{n-1} & \\ & & & \ddots & & Y_{n,n} \end{bmatrix} \qquad (23)$$

to produce the most desirable physical arrangement.

4. Multiplexing Techniques

A large percentage of microwave filters which are manufactured are used in multichannel systems. Thus, the problem of multiplexing becomes important and the techniques vary according to whether the multiplexing system is required to transmit high or low power and whether or not group delay distortion is important.

We shall concentrate upon the problem of combining several bandpass channel filters since the diplexer case possesses many solutions which are not applicable in the multichannel case. In all the cases considered, either circulators, 3 dB hybrids or a waveguide manifold are used. Circulators are normally 3-port, non-reciprocal, symmetrical matched devices which use a common junction containing magnetized ferrite designed to produce a scattering matrix over the communication band of interest of the form:

$$\begin{bmatrix} 0 & e^{j\psi} & 0 \\ 0 & 0 & e^{j\psi} \\ e^{j\psi} & 0 & 0 \end{bmatrix} \qquad (24)$$

where ψ is approximately a linear function of frequency and the losses have been neglected. 3 dB hybrids are 4-port reciprocal, symmetrical, matched devices which may be phsyically realized in several ways but posses a scattering matrix

$$\begin{bmatrix} 0 & 0 & \frac{e^{j\psi}}{\sqrt{2}} & j\frac{e^{j\psi}}{\sqrt{2}} \\ 0 & 0 & j\frac{e^{j\psi}}{\sqrt{2}} & \frac{e^{j\psi}}{\sqrt{2}} \\ \frac{e^{j\psi}}{\sqrt{2}} & j\frac{e^{j\psi}}{\sqrt{2}} & 0 & 0 \\ j\frac{e^{j\psi}}{\sqrt{2}} & \frac{e^{j\psi}}{\sqrt{2}} & 0 & 0 \end{bmatrix} \qquad (25)$$

where ψ is approximatelly a linear functon of frequency over the communication band of interest. These components have negligible loss and will operate under high power conditions.

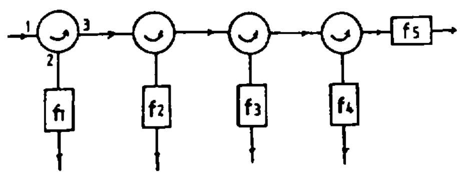

Fig. 3. Multiplexing conventional filters using circulators.

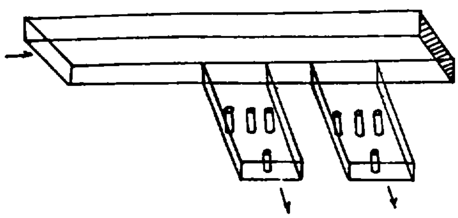

Fig. 4. Direct waveguide multiplexer using manifold.

The most common arrangement in medium capacity systems is to use a circulator and filter for each channel as shown in Fig. 3 for 5 channels. On the receiving side the signal enters port 1 of the first circulator and emerges as port 2 without distortion. Signals oter than those contained in the passband of the first filter(f_1) are reflected and emerge as port 3 distorted by the reflection group delay of the filter. However, if the amplitude is the primary concern, then the first channel has been sucessfully extracted leaving the remaining channels without amplitude distortion and the process may be repeated. One great advantage of this arrangement is that channels may be extracted in any order or even omitted altogether without changing the system. This may be contrasted to the case where filters are directly connected to a waveguide manifold or equivalent as illustrated in Fig. 4 where every filter interacts and the whole has to be designed as a single unit.

5. Conclusion

Most of the specifications for microwave systems may be a known design technique. Whether only amplitude or combined amplitude and phase specifications are required at least one of the designs will be capable of exhibiting the desired response, the only difficulty being able to choose the most appropriate configuration taking into account size, loss, economic and environmental constraints for the filters and possible multiplexing.

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國 文 抄 錄

Microwave 帶域通過 濾波器

任意의 必要한 特性을 나타내는 micro波 導波管 直結空胴 濾波器를 設計하였다. 이 設計 技術은 가장 一般的인 두가지 濾波器를 表 및 graph들만 使用하여 設計式에 매우 近似한 濾波器를 設計할 수 있다.