$$Q = \left(\frac{\lambda_0}{4a}\right)\frac{1}{\left(1-\frac{\lambda_0^2}{4a^2}\right)^{3/2}}\left\{\cot^2\left(\frac{\pi D}{2a}\right) - \frac{1}{D^2}\left[\frac{\pi(a^2-D^2)}{4aD\cos\left(\frac{\pi D}{2a}\right)}\right]^2 \times\left[\left(\frac{b^2}{3}+\frac{d^2}{2}-\frac{8bd}{\pi^2}\right) - 2\left(\frac{b}{\pi}\right)^2\sum_{N=1}^{\infty} J_0^2\left(\frac{N\pi d}{b}\right)K\left(\frac{2N\pi(a-D)}{b}\right)\Big/ N^2\right]\right\} + \left(\frac{b}{\lambda_0}\right)\frac{\left(1+\frac{\lambda_0^2}{4D^2}-\frac{\lambda_0^2}{2a^2}\right)}{\left(1-\frac{\lambda_0^2}{4a^2}\right)^{3/2}}\left[\frac{\pi(a^2-D^2)}{4aD\cos\left(\frac{\pi D}{2a}\right)}\right]^2 \log\csc\left(\frac{\pi d}{2b}\right). \quad (8)$$

Contours of equal Q and contours of equal resonant frequency have been mapped out in the D-d plane for irises in waveguides for different bands. These maps permit the quick determination of resonant frequency and loaded-Q for an iris of any dimension $D \times d$.

The bandwidth of the iris filter over which the VSWR is below an assigned value S is given by [11]

Iris Number	$D\times d$ inches	Resonant Frequency		Loaded-Q Measured	Total Q Q_T	Loaded Q Equation (8)
		Equation (4)	Measured			
1 7	0.558 ×0.108	11 500	11 405 11 370	1.56	6.99	1.703
2 6	0.542 ×0.080	11 500	11 420 11 500	2.11	3.36	2.233
3 5	0.533 ×0.060	11 500	11 445 11 500	2.86	3.18	2.879
4	0.528 ×0.045	11 500	11 415	3.35	3.35	3.339

$$w = f_0\left(\frac{S-1}{2\sqrt{S}}\right)^{1/n}\left(\frac{1}{Q_T}\right) \quad (9)$$

in which f_0 is the midband frequency of the filter. The overall frequency response can be evaluated from

$$\text{Loss in dB} = 10\log_{10}\left[1+Q_T^{2n}\left(\frac{f}{f_0}-\frac{f_0}{f}\right)^{2n}\right]. \quad (10)$$

The on-band insertion loss decreases slowly with n, because the factor $Q_T(f/f_0 - f_0/f)$ is less than unity in the passband. The off-band attenuation rises rapidly with n, because the same factor exceeds unity in the upper and lower stopbands.

The table in Fig. 1 includes data for 4 iris filters constructed to suppress the spurious outputs in X-band varactor frequency multipliers. Because of the interaction between the closely spaced irises, the mid-frequency of the filter becomes higher than the resonant frequency of the irises composing the filter. The loaded-Q for the outer-most irises are larger than those of the others to obtain sharp cutoff at the band edges. Thus, higher Q_T than the maximally flat requirement obtains when Q_T is calculated on the basis of the loaded-Q of the outer-most irises. The attenuation close to the band edges of the filter shown in Fig. 3 agrees with the result calculated from (10) with $Q_T = 6.99$. In the far off-band ranges, the attenuation in Fig. 3 conforms to the value found from (10) but with $Q_T = 3.35$. The insertion loss amounts only to 0.2 or 0.3 dB over a major portion of the passband. The matching property of the iris filter is illustrated by Fig. 4 which indicates that the band over which S is below 2.0 extends from 10 300 to 13 500 MHz. This checks with $w = 3120$ found from (9) with $f_0 = 11\,900$, $S = 2.0$, and $Q_T = 3.35$.

The iris filter by itself can be connected to the output of an X-band traveling-wave tube; spurious output lying outside of the passband in Fig. 3 will be eliminated. When two iris filters having overlapping passbands are cascaded, a filter of intermediate bandwidth is obtained for use in tunable solid-state generators. For narrowband applications, the iris filter is connected in tandem with a Mumford inductive-post filter, Fig. 5; the former compensates for the deficiency in attenuation found in the latter. Figure 6 portrays the frequency response of the composite filter which, when used in the output of a varactor multiplier, reduces practically all the extraneous signals to a level 80 dB below the desired signal.

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Additional Considerations in Comb-Line Bandpass Filter Interstage Couplings

In a prior correspondence,[1] various aspects of narrowband comb-line bandpass filters were discussed. Experimental data were presented showing that *nonzero* coupling is obtained between adjacent quarter-wave combline resonators when a *nonuniform* (i.e., compound) center conductor is employed. In this correspondence, additional techniques for controlling this interstage coupling will be discussed.

The comb-line filter structure previously employed has been retained herein for additional experiments with center frequencies still at 2000 MHz. From Fig. 4 of a previous correspondence,[1] a coupling bandwidth $\Delta f_{12} = 88$ MHz is obtained as the coupling with no partitions between adjacent resonators. *Nonzero* coupling exists because the compound center conductor provides *unequal* magnetic and electric couplings.

The comb-line filter structure (see Fig. 1) has been modified by use of a 0.250 inch diameter metallic coupling post located midway between the adjacent resonators and connecting the top and bottom walls of the filter. The post centerline is located 1.125 inch from the plane of the short (68.7 electrical degrees at 2000 MHz). Upon inclusion of this post into the filter structure, the measured coupling bandwidth was increased to 254 MHz.

An additional modification of the coupling mechanism entailed use of a metallic coupling screw passing through the coupling post, as shown in Fig. 1. With the end of the coupling screw flush with the exterior surface of the coupling post (i.e., gap = 1.000 inch), the mea-

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[1] R. M. Kurzrok, "Design of comb-line band-pass filters," *IEEE Transactions on Microwave Theory and Techniques* (*Correspondence*), vol. MTT-14, pp. 351–353, July 1966.

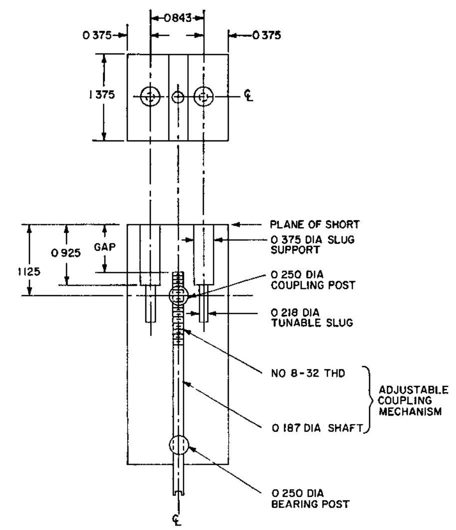

Fig. 1. Comb-line filter structure with modified interstage coupling mechanism.

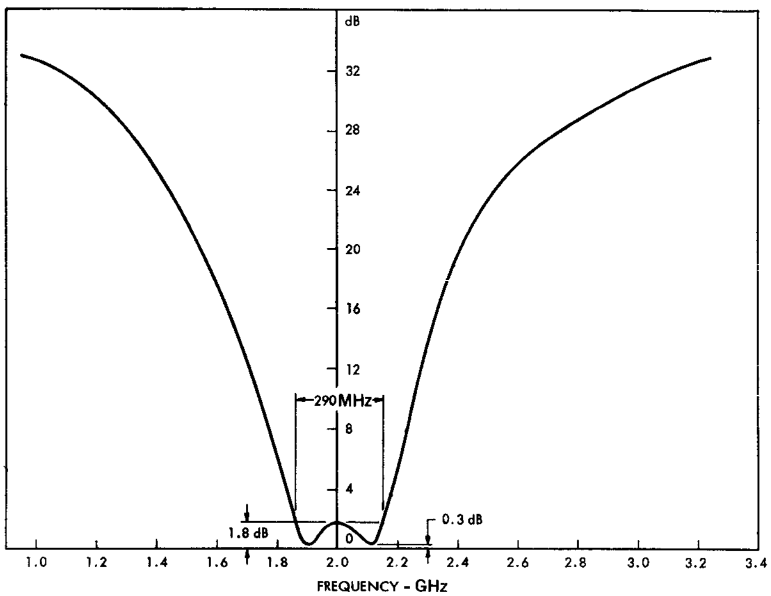

Fig. 2. Measured filter response (insertion loss versus frequency).

sured coupling bandwidth was 280 MHz. Upon further insertion of the coupling screw, reducing the gap to 0.730 inch, the measured coupling bandwidth was 353 MHz.

Both the coupling post and the coupling screw (as used herein) increased the net magnetic coupling and provided closer coupling between adjacent resonators. It is anticipated that both the coupling post and the coupling screw techniques can be used to obtain appreciable magnetic coupling between quarter-wave *uniform* comb-line resonators. Such uniform resonators are normally decoupled due to cancellation of magnetic and electric fields.[2]

The comb-line structure of Fig. 1, employing the 0.250 inch diameter coupling post, has been converted into a two-resonator bandpass filter by adding input/output couplings. Direct taps were implemented by using number 12 wire between the inner conductors of type UG-290A/U BNC fittings and tap points on the 0.375 inch diameter slug supports. With input/output tap points located 0.222 inch from the plane of the short, the measured filter response curve of Fig. 2 was obtained. Passband insertion loss comprised a 1.5 dB ripple plus 0.3 dB dissipation loss. The measured ripple bandwidth was about 290 MHz. The insertion loss on the high-frequency skirt drops down to 32 dB at 3500 MHz and 11 dB at 4000 MHz. This is due to the 1.375 inch dimension of the filter enclosure, which can propagate the TE_{10} rectangular waveguide mode at 4280 MHz.

Inductive posts have been used extensively as interstage couplings in waveguide bandpass filters. The brief experiments reported herein suggest the efficacy of using coupling posts in TEM filters employing comb-line resonators. Multiresonator bandpass filters can be fabricated quite economically using round-rod resonators parallel to the broad walls of nonpropagating rectangular waveguide tubing. Fixed resonator spacings could accommodate a range of filter bandwidths and/or response shapes by the use of coupling posts.

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High-Frequency Transistor Evaluation by Three-Port Scattering Parameters

Introduction

Semiconductor technology has developed sufficiently for transistors to retain useful properties up into the microwave region. One GHz is the approximate lower-frequency limit of this region, although a preferable concept might be to consider a transistor in microwave terms when the encapsulation design and associated jig or holder impedances have significant effect. One such effect is the reduction of measurement accuracy for admittance parameters, due to the increasing difficulty of specifying reference planes. Frequency characteristics of the jig and its interaction with the device must also be considered.

Transistors have so far been characterized as two-port networks whose admittance parameters are measured in two-port configurations. An alternative and more general method of three-terminal device characterization is by the three-port or three-terminal pair scattering parameters. The theory and technique for using scattering parameters are well known for multiport passive microwave junctions. The purpose of this correspondence is to apply the wave scattering concept to the evaluation of transistors.

Ideally, the three-terminal device is inserted at the center of the junction of three transmission lines. Reflection and transmission measurements made on each port then specify the device completely. An active device, in these circumstances, might exhibit circulation or gain, which could become appropriate quantities for characterization in the microwave region.

Because each port sees the characteristic waveguide impedance, power gain, circulation and dissipation in the matched arms of the junction can take place. Thus, algebraic expressions for the scattering parameters take on a more complicated form than the admittance parameters in terms of the transistor equivalent circuit elements. However, the practical consequences of measuring scatter-

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