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Extracted-Pole Filter Manifold Multiplexing

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***Abstract*—A transformation method is introduced for enabling filters of the extracted-pole variety to be match-multiplexed onto a manifold using standard waveguide multiplexer computer programs. Thus the advantages that accrue from the extracted-pole realization for filters may now be extended to multiplexers, which will be particularly useful in narrow-band high-power low-loss multiplexing applications. The measured performance of a 12-GHz contiguous-channel quadraplexer comprising TE_{011} cavity extracted-pole elliptic filters is presented, demonstrating the very low insertion losses attainable with this form of realization. Since the majority of applications envisaged for this type of multiplexer is in high-power output circuitry, a discussion on thermal aspects is included.**

I. Introduction

In a previous paper [1], the synthesis procedure for extracted-pole filters[1] was introduced and the measured results of a laboratory model presented. This model demonstrated the advantages of building bandpass filters in this way, which stem chiefly from the single-sign coupling elements throughout the filter that the extracted-pole procedure ensures. For when the coupling elements are uniform in sign, advanced self-equalized pseudo-elliptic characteristics may be realized with a single-layer arrangement of TE_{01n}-mode resonance cylindrical cavities. This in turn yields the advantages of a high unloaded Q for optimally low in-band insertion losses, relatively large dimensions for immunity from multipactor effect in space, for high-power handling capability, and insensitivity to manufacturing tolerances. Also, the device has a flat bottom for easy transfer of dissipated heat to a flat cooling plate, if necessary. One of the most important applications foreseen for this type of filter is in high-power low-loss contiguous or near contiguous channel multiplexing. When elliptic function characteristics can be used in such multiplexers, performance may be enhanced even further since the design bandwidths may be made greater and sometimes the degree of filter necessary may be reduced, both of which tend to reduce loss and in-band signal distortion. Particularly, application was envisaged for direct-broadcasting TV satellites using recently developed high-power TWTA's and Klystrons, whose output powers may be as high as 600 W. The aim of this paper is to introduce a

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[1]US Patents pending.

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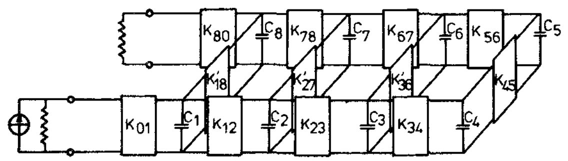


Fig. 1. Generalized cross-coupled double-array network—8th degree example.

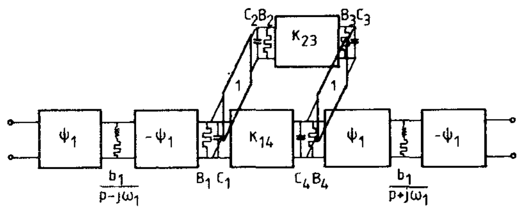


Fig. 2. Example of a pseudo-elliptic extracted-pole network.

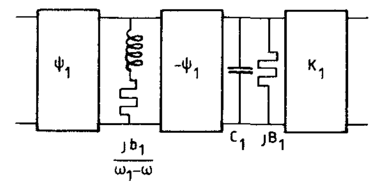


Fig. 3. Section of an extracted-pole network (network 1)

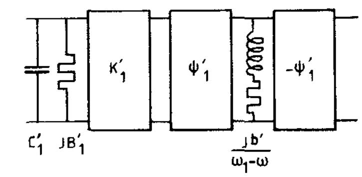


Fig. 4. Transformed section (network 2).

method for multiplexing extracted-pole filters which is an adaptation of previously established manifold multiplexer technology, and to present the measured performance results of a quadraplexer designed using the theory about to be described. The configuration chosen for this quadraplexer is that for a direct-broadcast TV satellite working for the Scandinavian countries which have been allocated four "contiguous" TV broadcast channels, numbers 22, 24, 26, and 28 (1979 World Administrative Radio Conference). These 27-MHz channels, center frequencies 38.4 MHz separated at about 12 GHz, are the closest that any TV broadcast satellite will be required to transmit from one antenna, and as such represents an appropriate challenge for this new technology.

II. Multiplexing Procedure

When computer-aided methods for manifold multiplexing of extracted-pole filters were being considered it was decided early on not to develop a specialized extracted-pole manifold multiplexer computer program. Apart from being specialized, it would require an in-depth discussion of manifold multiplexing methods which tend to be lengthy and complicated. Since there are a variety of methods and computer programs available for the manifold multiplexing of waveguide filters of conventional configuration (Fig. 1), it was thought that a simpler course to take would be to transform the extracted-pole filter configuration (Fig. 2) to resemble this conventional configuration, especially at the end nearest the manifold. The idea behind this transformation method is to run any unmodified waveguide manifold multiplexer program with the filter response embodied in the conventional topology network, which the program will deal with. It will then be noticed that due to the influence of neighboring filters, the element values of the network will have changed from those of the original (double-ended) network, these changes being greatest in the elements nearest the manifold, and will diminish rapidly in significance the farther they are from the manifold. The next step is to take the same filtering characteristic but now synthesize it in an extracted-pole configuration. Then exact transformations (later referred to as "forward" transformations) are applied to this network in order to shift all the "extracted-pole elements" to one side of the network. Having done this, the configuration of the other side of the network, which will eventually become the side nearest to the manifold, will then be the same as the "conventional configuration" referred to earlier. The percentage changes that were noted in the elements of the conventional configuration network after multiplexing may now be applied as far as is possible to the corresponding elements of the transformed extracted-pole network. Finally, the extracted poles of this modified transformed network may be restored to their original positions ("reverse" transformation). Thus, the original extracted-pole configuration has been restored, with the elements modified by the necessary amounts for match-multiplexion of the network onto a manifold.

III. Forward and Reverse Transformation of the Extracted-Pole Configured Network

Referred from the original extracted-pole filter paper [1], a 6th degree pseudoelliptic network in extracted-pole configuration is shown in Fig. 2. This represents a typical extracted-pole network with a cross-coupling (K_{14}) for phase self-equalization. In general, this cross-coupling will be light or nonexistant in output multiplexer applications, but in any case is ignored initially.

Assuming that the left-hand port of the network is to be nearest the manifold, a section may be identified containing a resonant-pole section separated by a unity impedance ideal phase shifter prior to the cross-coupled array in the center of the filter. This is illustrated in Fig. 3. Here, the phase shifters are of unity impedance and phase length $\pm\psi_1$ and the impedance inverter has a characteristic admittance of K_1. ($K_1 = 1$ in the example of Fig. 3.)

The first step is to transform this section into the configuration shown in Fig. 4 (network 2). After this first (forward) transformation, the new section will have exactly the same electrical characteristics as the original section, but the shunt capacitor, frequency-invariant reactance, and inverter have moved to the left (or manifold) side of the network, and the pole-producing elements to the right. With the pole-producing elements in their new position, another section similar to network 1 may now be identified

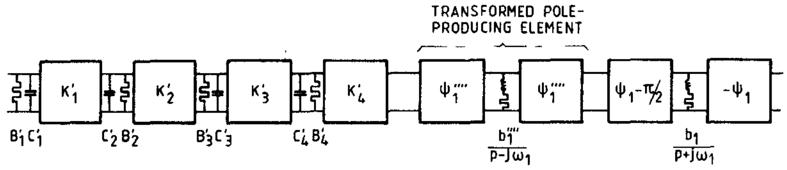
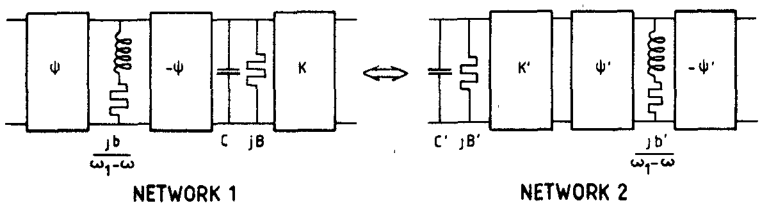


Fig. 5. Extracted-pole network after 4 forward transformations.

TABLE I
FORWARD AND REVERSE TRANSFORMATION FORMULAS.



Reverse transformations	Forward transformations
$C = C'\left[1+\frac{b'C'\cos^2\psi'}{K'^2}\right]$	$C' = \frac{C}{1+bC\sin^2\psi}$
$\psi = -\cot^{-1}\left[C'\omega_1 + B' + K'^2\tan\psi'\right]$	$\psi' = -\tan^{-1}\left[\frac{C\omega_1 + B + \cot\psi}{K^2}\right]$
$K = \frac{C}{C'}K'$	$K' = \frac{C'}{C}K$
$b = \frac{b'\cos^2\psi'}{K K'\sin^2\psi}$	$b' = \frac{b K K'\sin^2\psi}{\cos^2\psi'}$
$B = -[K^2\tan\psi' + \cot\psi + C\omega_1]$	$B' = -[\cot\psi + K'^2\tan\psi' + C'\omega_1]$

and the transformation applied again, moving the pole-producing section farther to the right and the next shunt capacitor, frequency-invariant reactance, and inverter (C_2, jB_2, K_2) towards the input. The transformation may be repeated until all the shunt capacitors are on the left of the network and all the pole-producing sections are on the right. If there is more than one pole-producing element on the left of the network, the transformation may be applied to these others until they too appear on the right of the network. For the case of the extracted-pole network in Fig. 2, the resultant network will be as shown in Fig. 5. It can now be seen that the left hand (manifold) side of the network has conventional configuration.

Now the element triplets $C_1', B_1', K_1', C_2', B_2', K_2', \cdots$ may be adjusted in value by the same proportion that the corresponding elements in the double-folded array realization of the same characteristic have been changed by, after application of a standard manifold multiplexer program [3]–[6]. Usually it is found that the first two or three element triplets nearest to the manifold suffer the greatest changes, after which the effect of the multiplexing process on the element values diminishes progressively the farther the element is from the manifold.

Having modified the values of C_1', B_1', K_1', etc., the reverse transformations may now be applied to restore the transformed pole-producing element to its original position at the left of the network, at which stage the network will have its original topology but with the element values altered in such a way that the network may be match-multiplexed onto a manifold. These transformed element values will be referred to as $\psi_{1t} b_{1t}$, $C_{1t} B_{1t} K_{1t}$, $C_{2t} \cdots$, etc. The formulas for the forward and reverse transformations are given in Table I. At each stage the network is electrically identical to the original. For some applications it may be advantageous to leave the poles at the right of the network, i.e., not to use the reverse transformation, since this would result in a shorter length of waveguide between the first iris and the manifold, a feature which tends to improve the performance over wide bandwidths. However, experience has shown that it tends to leave a very high susceptance value for the iris nearest to the pole-producing elements, which may be undesirable from a power handling point of view.

Having obtained the modified element values, it remains to adjust the cross-couplings. (K_{14} in the example of Fig. 2.) During the multiplexing design process the cross-coupling inverters are modified in a manner which tends to retain the location of the transmission zeros. For the extracted-pole case, the locations of the transmission zeros on the $j\omega$-axis are automatically retained, while the re-

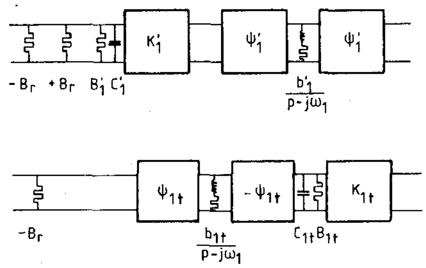


Fig. 6. Section of network before and after final reverse transformation showing inclusion of the additional reactance.

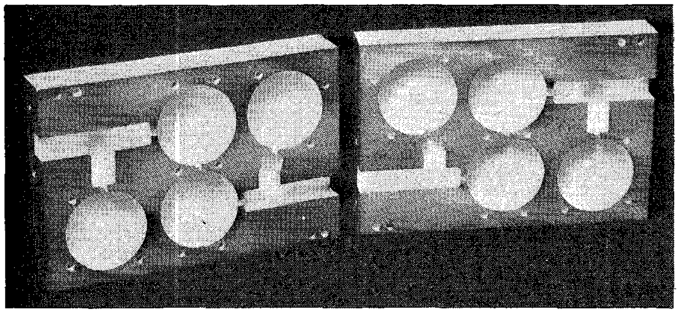


Fig. 7. Disassambled extracted-pole filter.

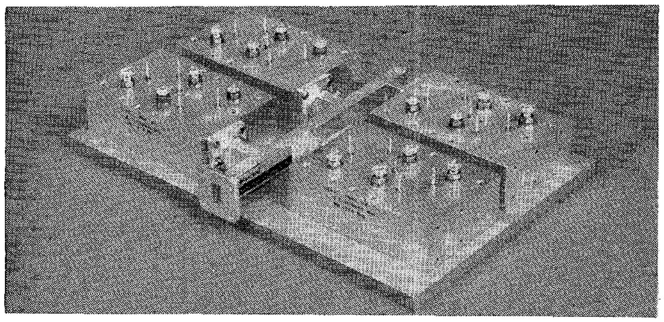


Fig. 8. General view of assembled quadraplexer.

mainder are contained in the central cross-coupled array and are adjusted by the same proportion as the corresponding cross-couplings of the multiplexed double array [6].

During the reverse transformation process a small modification at the stage of the final transformation will greatly aid the later construction and tuning stages. The modification involves the inclusion of an additional shunt frequency-independent reactance before the final reverse transformation, after which the phase length ψ_{1t} will be the same as the original phase length ψ_1. Physically, the multiplexed filter will then resemble the double-ended filter and may be independently tuned up as such initially. Then, when the filter is to be tuned with the manifold, the necessary changes in the iris susceptance values and cavity center frequencies may be achieved by screw adjustments and the additional reactance realized with a capacitive or inductive screw.

To calculate the value of this reactance, consider the state just before the final reverse transformation. From Table I

$$\psi_{1t} = -\tan^{-1}\left\{\frac{1}{C_1'\omega_1 + B_1' + K_1'^2 \tan\psi_1'}\right\}. \tag{1}$$

If it is required that $\psi_{1t} = \psi_1$, two shunt susceptances $\pm B_r$ must be added to the left of the network just before the final reverse transformation (Fig. 6). Then, if one of these additional reactances is included with B_1'

$$\psi_1 = -\tan^{-1}\left\{\frac{1}{C_1'\omega_1 + (B_1' + B_r) + K_1'^2 \tan\psi_1'}\right\}$$

or

$$-\cot\psi_1 = C_1'\omega_1 + (B_1' + B_r) + K_1'^2 \tan\psi_1')$$

$$\rightarrow B_r = -\left(K_1'^2 \tan\psi_1' + \cot\psi_1 + B_1' + C_1'\omega_1\right). \tag{2}$$

The result is that the phase lengths after the final transformation are the same as those before the transformation process started, and an additional reactance has appeared on the manifold side of the filter at an electrical distance of an odd number of quarter-wavelengths from the effective reference plane of the impedance inverter produced by the iris feeding the first main cavity. This reactance may be positive or negative and is usually low in value. The best realization is a screw in the broadwall of the rectangular waveguide which may be withdrawn while the filter is being tuned in isolation, and adjusted for best common-port return loss during the final stages of multiplexer tuning.

IV. Breadboard Model of a High-Power Quadraplexer

To demonstrate the feasibility of using TE_{01}-mode elliptic function filters in high-power multiplexer applications, a quadraplexer was designed and built for four semi-contiguous TV channels, WARC numbers 22, 24, 26, and 28, whose center frequencies are spaced at 38.4 MHz. The useable bandwidth specified for each channel is 27 MHz.

To meet the rejection specifications, four 4th degree TE_{011}-mode extracted-pole filters were designed, built, and tuned to the center frequencies. The design equiripple bandwidths were 33 MHz, in-band return loss levels 22 dB, and the poles were at $\pm j1.818$. A photograph of the two halves of one filter is shown in Fig. 7 and of the assembled

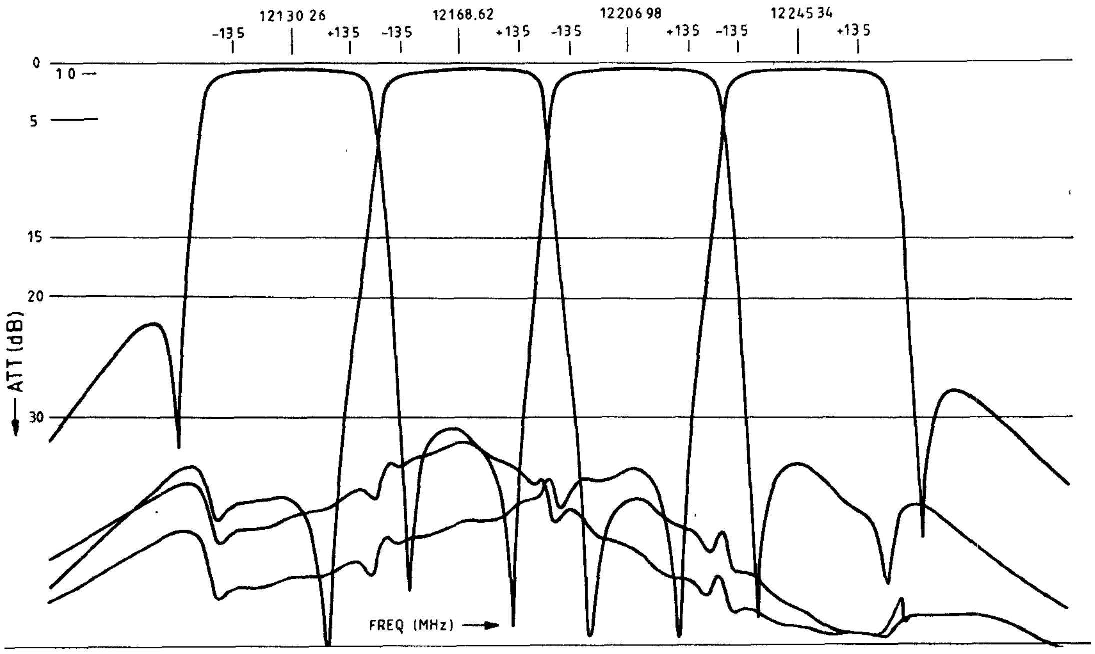


Fig. 9. Superimposed channel rejection characteristics.

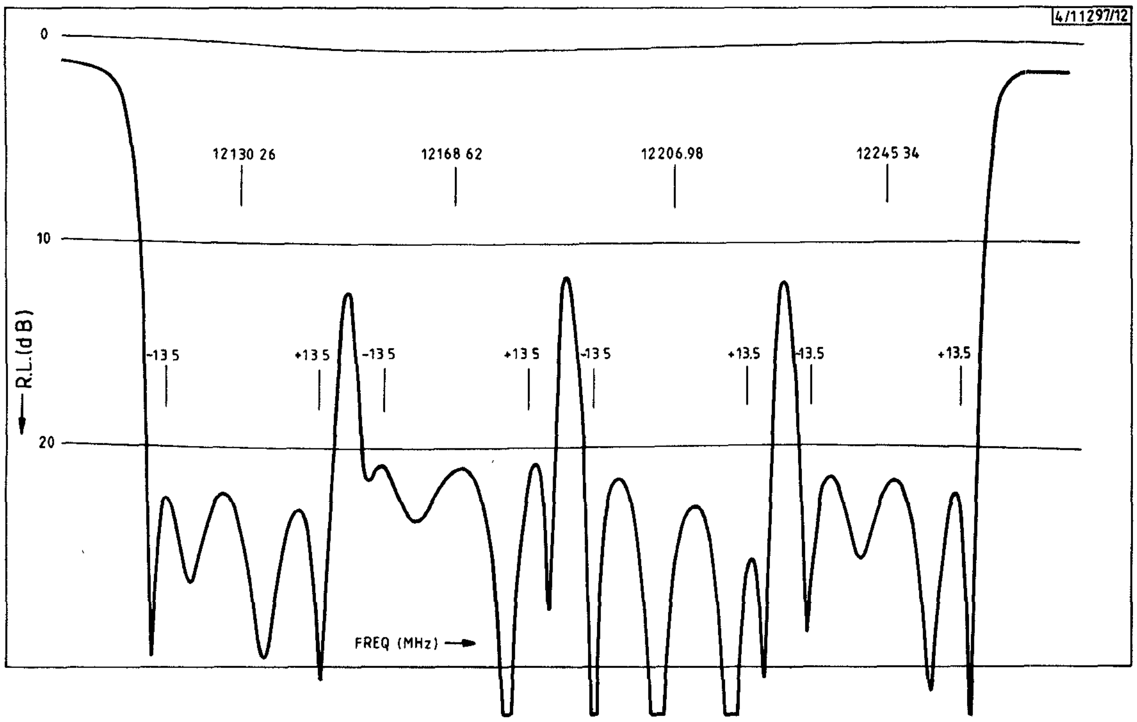


Fig. 10. Common port return loss.

quadraplexer in Fig. 8. After tuning the individual filters, they were assembled onto the manifold and the susceptance and tuning screws nearest the manifold readjusted until the common-port return loss was restored to >22 dB over the channel bandwidths.

Some of the measured results appear in Fig. 9 (rejection), Fig. 10 (common-port return loss), Fig. 11 (group delays), and Fig. 12 (wideband sweep to check for out-of-band spurious responses). The band-center insertion loss of each channel was between 0.52 and 0.55 dB, composed of a filter loss of about 0.42 dB (corresponding to a Q_u of about 16 000) and a manifold loss of between 0.1 and 0.13 dB. The measured results corresponded closely to the predicted, demonstrating clearly the rejection enhancement that occurs in the regions around the crossovers of two adjacent filters. The group delay curves show the slight asymmetric distortions that each channel suffers as a result of this enhancement, especially the outer channels. It will be possible to correct these distortions by appropriate asymmetric placement of the poles and/or (exceptionally)

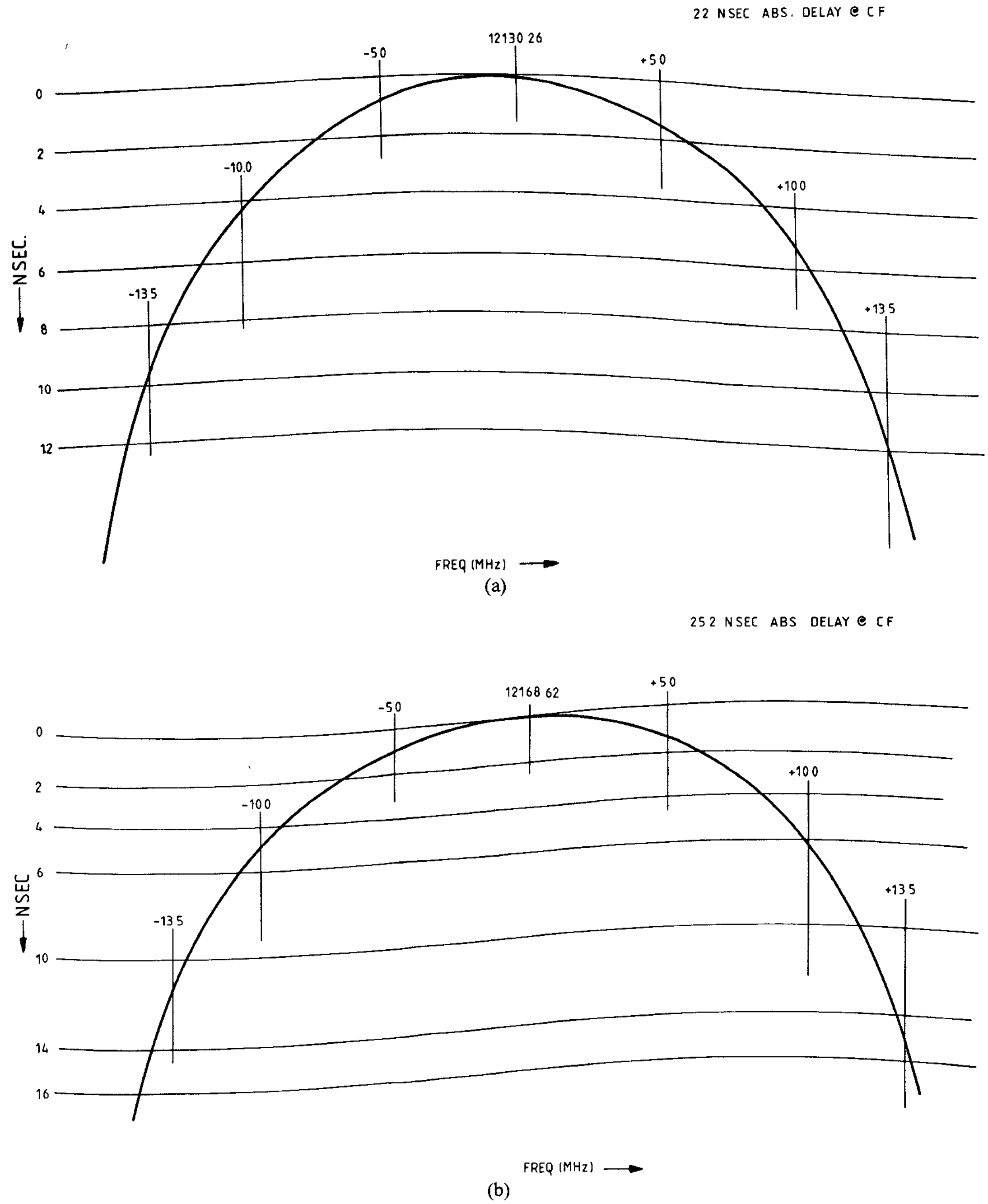


Fig. 11. (a) Channel 22 group delay. (b) Channel 24 group delay.

the inclusion of complex-plane transmission zero pairs asymmetrically arranged about the complex-plane real axis. The exact procedures for the combination of such prescribed asymmetric transmission zero patterns with a prescribed in-band equiripple level have recently been solved, and the generating and network synthesis programs written. Using such programs, asymmetric group delay distortions may be made symmetric, or completely flattened.

V. Thermal Considerations

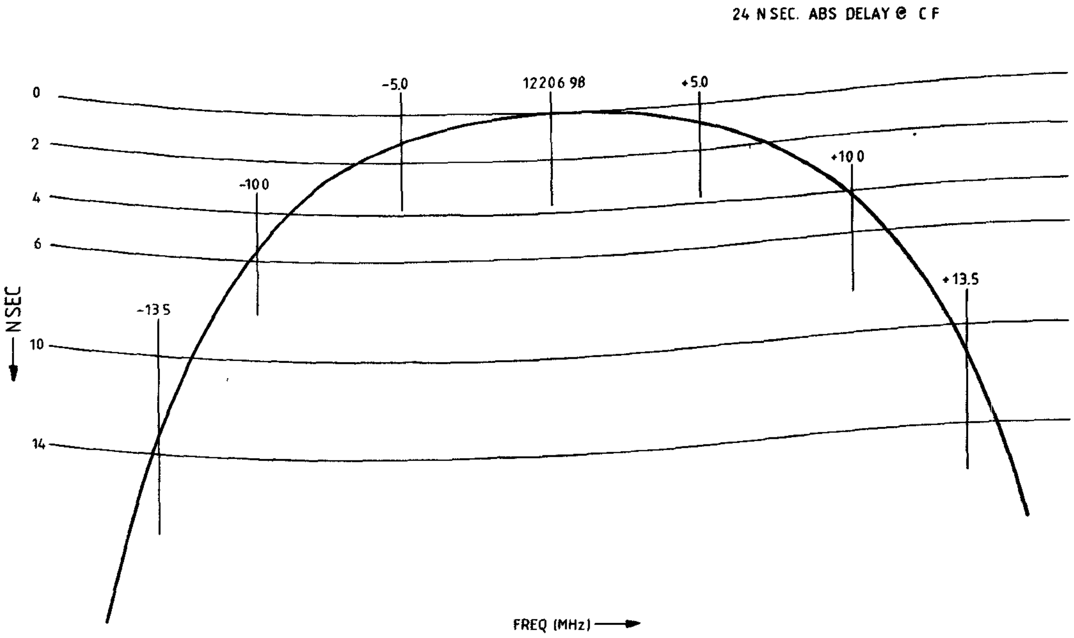
Since this multiplexer is primarily intended for high-power space applications, it is necessary now to pay some attention to the thermal aspects. The points to be considered are listed below.

1) Any heat generated inside the cavities by RF losses will have to be conducted down to a flat cooling plate.

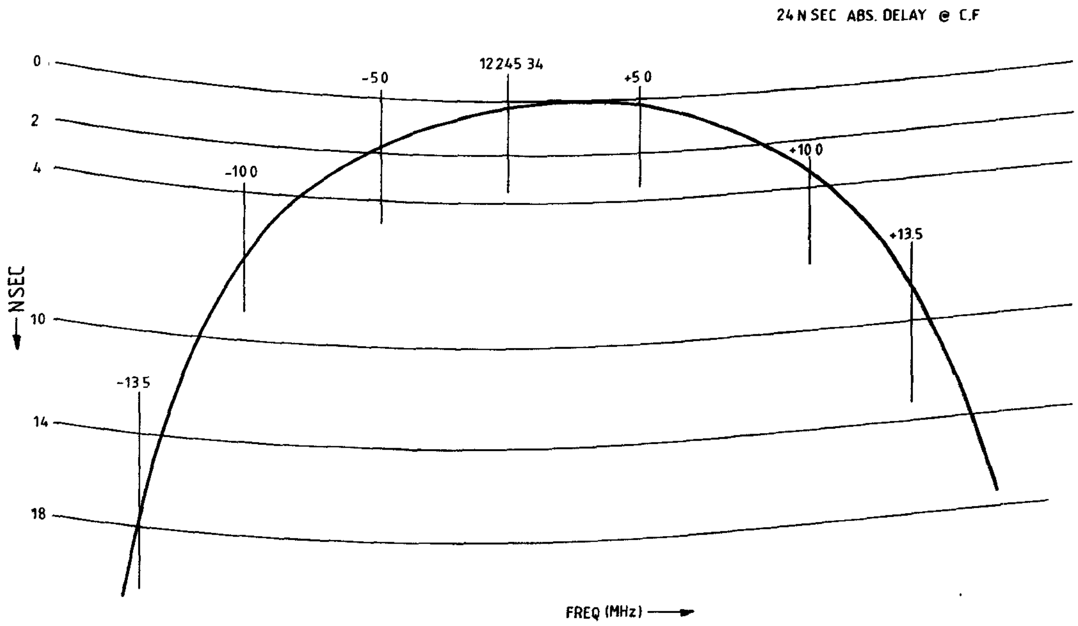
2) The hotter an uncompensated device runs, the more it will drift from its center frequency, which in this narrow-band application could be serious.

3) If the device is constructed from thick-wall aluminum, which has a fairly good thermal conductivity, gains are made in terms of lightness and machinability and the device will run relatively cool, minimizing frequency drift.

4) If thin-wall Invar is used the device will run hotter because of the poor thermal conductivity of the metal, and then the frequency drift and weight will probably be in the same order as the thick-wall aluminum solution in spite of the thermal stability of Invar.



(c)



(d)

Fig. 11. (c) Channel 26 group delay. (d) Channel 28 group delay.

5) A thick-wall Invar solution will be too weighty.

To aid in the trade-offs, some curves were generated to illustrate the normalized cumulative distribution of dissipated RF energy within a cavity supporting the TE_{01n}-mode resonance, using the diameter/height ratio as a parameter.

The cumulative power dissipation distribution contained in a concentric circular area of fractional radius r on the top or bottom surface of a TE_{01n} cavity of diameter/height ratio R, is

$$P_{ce} = \frac{(\pi nRr)^2}{4K}\left(J_1^2(y_0 r) - J_0(y_0 r)J_2(y_0 r)\right)$$

where $y_0 = 3.83171 =$ 1st solution of $J_1(y) = 0$

$$K = y_1^2\left\{\frac{(\pi nR)^2}{2} + \frac{2y_0^2}{R}\right\}$$

$$y_1 = J_0(y_0) = -0.40276 \tag{3}$$

and J_0, J_1, and J_2 are Bessel functions of the first kind of order zero, 1, and 2, respectively.

For the walls, the cumulative power dissipated at a fractional height z of the cylinder is given by

$$P_{cw} = \frac{2y_0^2y_1^2}{RK}\left(z - \sin(2\pi nz)/2\pi n\right). \tag{4}$$

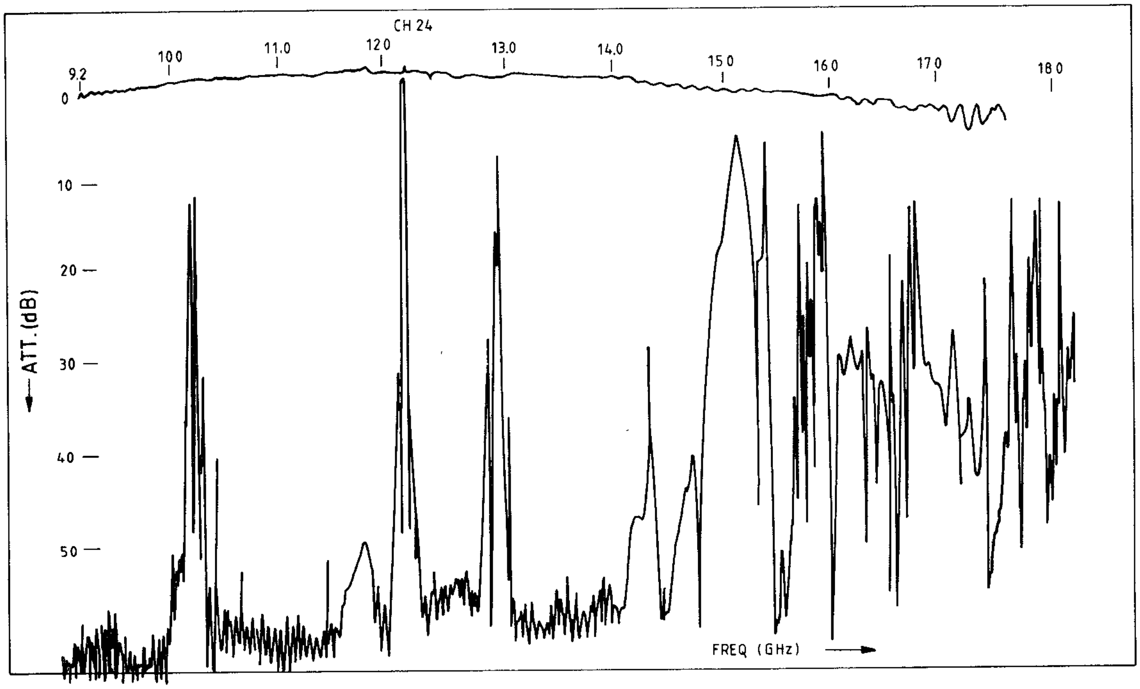


Fig. 12. Wideband sweep 9–18 GHz on channel 24.

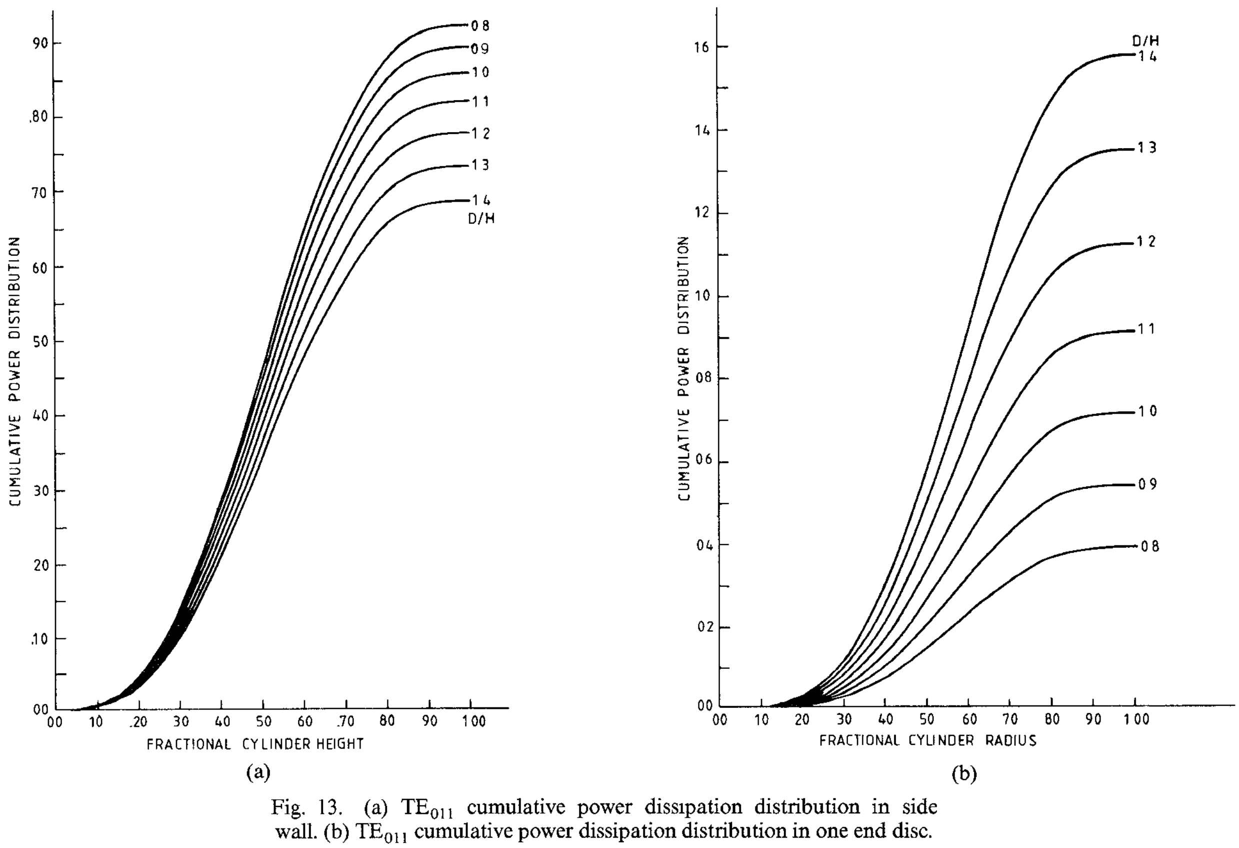


Fig. 13. (a) TE_{011} cumulative power dissipation distribution in side wall. (b) TE_{011} cumulative power dissipation distribution in one end disc.

The ratio of power dissipated in the circular end plates and the cylindrical wall is simply given by

$$\frac{P_{ce}}{P_{cw}} = \left\{ \frac{n\pi}{2y_0} \right\}^2 R^3. \tag{5}$$

Fig. 13 illustrates the cumulative power distribution curves for TE_{011} cavities ranging in D/H ratio from 0.8 to 1.4, and Fig. 14 is the curve of the ratio of power dissipated in the two end walls to that dissipated in the side wall, as a function of D/H ratio.

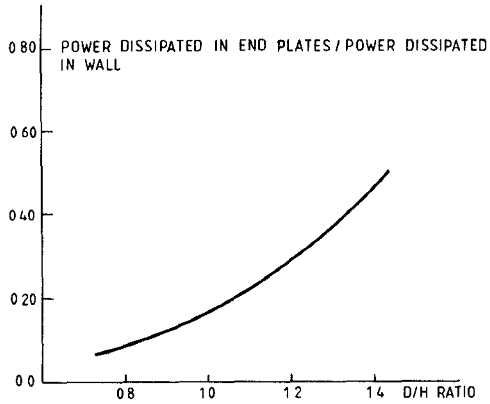


Fig. 14. Power dissipated in end-discs/power dissipated in side wall ratio.

The conclusion to be drawn from these curves is that when the D/H ratio is close to unity (best for maximum Q in the TE_{011} cavity) then only about 8 percent of the total dissipated power is absorbed by the top end disc of the cavity, i.e., that farthest from the cooling plate. The majority of the heat is dissipated in a band round the center of the cavity, and if a good conducting path could be provided between this band and the cooling plate, the temperature difference between the filter and ambient will be relatively small. (A similar exercise performed with TE_{111}- and TE_{113}-mode cavities showed that a much greater proportion of the total dissipated heat was lost in the end discs.)

Apportioning the dissipated heat to areas inside the cavity and estimating the distribution round two typical irises in the cavity sidewall, a thermal analysis program was run which showed that the excess temperature of an Invar cavity could be six times that of an aluminum cavity, for a given wall thickness. To increase the wall thickness of Invar to bring the temperature down to that of the aluminum cavity would render it exceedingly heavy, and cavity interconnection considerations would probably prevent it anyway. The conclusion that was drawn from this exercise was that although the temperature stability if Invar is $10\times$ that of aluminum, the weight of an equally-performing Invar cavity would be prohibitive. It seems to be worthwhile to build the filter of silver-plated aluminum and incorporate some simple device for partial or total compensation for center-frequency drift with temperature on each cavity. Some initial work has already been performed in this area on a single cavity, and results are encouraging.

VI. Conclusions

A procedure has been introduced to match-multiplex extracted-pole filters onto a common manifold using an adaptation of a method which has already been proved successful with cross-coupled double-array filters.

Some measured performance curves have been presented of a semi-contiguous narrowband-channel breadboard quadraplexer with TE_{011}-mode elliptic filters, intended for high-power direct-broadcast TV-satellite applications. The model represents a prototype for a new generation of high-power microwave output multiplexing devices, demonstrating the principal features which would be required of such devices, namely lowest insertion loss through the use of the high Q TE_{011}-mode resonance, realizability of advanced special-purpose symmetric or asymmetric transfer characteristics for low distortion and loss, and high-power handling capability. Mechanically, advantages are gained in lightness, relatively large dimensions for ease of manufacture from about 10 to 40 GHz, insensitivity to manufacturing tolerances, a flat-bottomed construction for efficient transfer of internally dissipated RF energy to a cooling plate of some kind, and ruggedness. Because the TE_{01}-mode is a higher order mode in a cylindrical cavity, care should be excercized in dimensioning the cavity to avoid close-to-band spurious responses, and it may be necessary in some applications to include a low-pass filter in the output manifold. However, the cutoff region of this low-pass filter may be over several gigahertz and so its contribution to the overall loss will be small. For narrow-band high-power applications where the multiplexer is constructed from aluminum, some type of temperature compensation will be needed and some promising methods are currently being studied [7].

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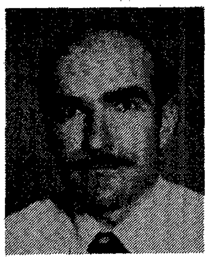

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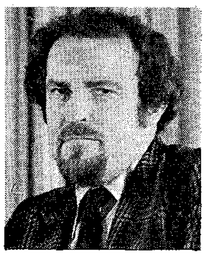

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A New Approach to the Computer-Aided Design of Nonlinear Networks and its Application to Microwave Parametric Frequency Dividers

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***Abstract*—A new cost-effective method allowing nonlinear microwave circuits to be designed by computer is demonstrated by application to parametric frequency dividers. The method is based on frequency-domain representations of both nonlinear circuit components and network voltages and currents. A special optimization strategy determines the unknown parameters of the linear part of the circuit while eliminating the need for a complete analysis of the nonlinear network at each step of the iterative process.**

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I. Introduction

IN THIS PAPER we demonstrate a new, very effective approach to the computer-aided design of nonlinear networks by working out in detail a specific microwave design problem and showing the experimental validity of the solutions obtained. The circuit to be dealt with—the