Transmit–Receive Multiplexer for the 12–14-GHz Band

HERBERT L. THAL, JR., SENIOR MEMBER, IEEE

***Abstract*—A compact low-loss transmit–receive multiplexer for direct satellite broadcast of two television channels is described. The response synthesis for the two contiguous transmit channels is outlined. A circuit model for the filter-to-manifold coupling, methods for its control, and the optimization of the manifold spacings are discussed. A technique for improving the transmitter-to-receiver isolation is shown. Experimental responses are given.**

I. Introduction

A TRANSMIT–RECEIVE multiplexer designed for direct TV broadcast operation in the 12–14-GHz band is shown in the photograph of Fig. 1 and the cross-section diagram of Fig. 2. The frequency plan is given in Table 1. The received signal (two TV channels plus command) from the antenna port at the right travels past the three transmit filters and through the six-pole rectangular waveguide filter to the receive port at the left. The TV transmit channels A and B are coupled through the two four-pole elliptic function filters in the center of the manifold [1]–[3], which form a contiguous diplexer. One end of each TV filter is connected to the sidewall of the WR62 waveguide; located in the broadwall at the same plane is an associated tuning aperture backed by a nonresonant cavity. At the other end of each filter there is a vestigial shorted manifold with a step change in width designed to yield a short-circuit in parallel with the filter terminals in the 14-GHz receive band; as a result, the net isolation between the transmitters and the receiver in the receive band is in excess of 100 dB. Two annulling cavities are located between these filters and the rectangular waveguide filter, which at the transmit frequencies has 100-dB rejection and appears as a virtual short-circuit. A two-cavity three-pole telemetry filter is coupled to the antenna end of the manifold. The entire assembly is fabricated from Invar for good thermal stability and silver plated for good RF conductivity. The cavity seams are laser welded. This single unit allows the receiver and transmitters to be connected to the same antenna port without the addition of a circulator or broadband diplexer to separate the transmit and receive signals.

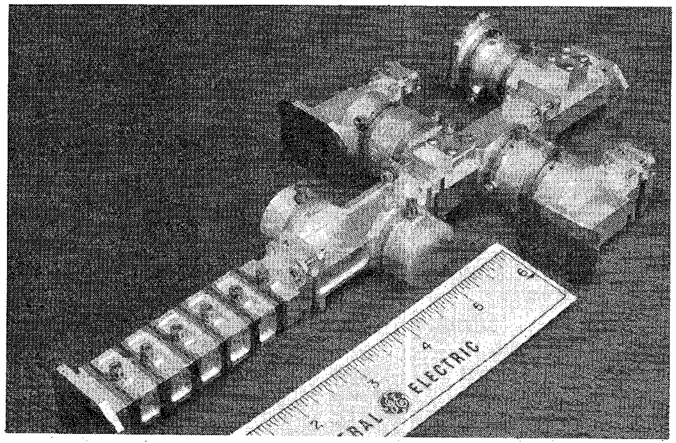


Fig. 1. Transmit receive multiplexer.

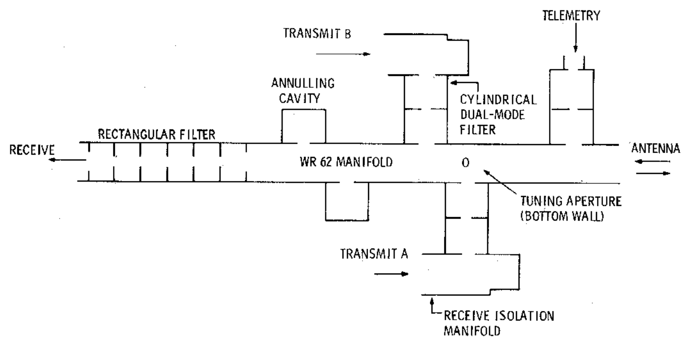


Fig. 2. Simplified cross-section view of bottom half of multiplexer.

TABLE I
FREQUENCY PLAN

	Signal Band (MHz)	Approximate Filter Bands (MHz)
Transmit:		
Telemetry	11710	11700-11720
Channel A	11905.5-11932.5	11880-11957.5
Channel B	11982.5-12009.5	11957.5-12035
Receive:		
Command	14005	
Channel A	14205.5-14232.5	13750-14550
Channel B	14282.5-14309.5	

II. Response Synthesis

The response of the two TV filters was optimized to meet the return loss and rejection requirements with minimum insertion loss and delay by assuming a large array of identical filters in parallel [4] and adjusting the locations of the S_{11} and S_{12} zeros. An interactive graphic program was used for this. Although the width of an individual channel is 27 MHz, the optimum results were

Manuscript received November 25, 1981; revised March 26, 1982.
The author is with the Valley Forge Space Center, General Electric Co., Philadelphia, PA 19101.

0018-9480/82/0900-1324$00.75 ©1982 IEEE

obtained by widening the filter bandwidths to the channel separation of 77 MHz so that a contiguous response was formed. The pole and zero locations of the low-pass prototype are given in Table II. Only the frequencies of the zeros plus a single real scale factor (e.g., the midband return loss) are chosen independently. These values uniquely specify the pole locations which are then computed by the program. Because of the symmetry requirements on the placement of the zeros, the only degrees of design freedom for the four-pole case are two complex S_{11} zeros, an imaginary (or real) S_{12} zero, and a scale factor. The complex reflection coefficient at the common port of this idealized parallel configuration was computed across the desired contiguous passband and stored as a reference for subsequent adjustment of the actual manifold configuration of Figs. 1 and 2.

The half-power output coupling bandwidths at ports 1 and 2, B_1 and B_2, are simply related to the real parts of the poles P and S_{11} zeros Z by

$$B_1 = -\sum_{n=1}^{N} P_m + \sum_{n=1}^{N} Z_n \tag{1}$$

$$B_2 = -\sum_{n=1}^{N} P_n - \sum_{n=1}^{N} Z_n. \tag{2}$$

The external Q's equal the resonant frequency divided by the respective bandwidth. For the multiplexer filter of Table II, the manifold end (port 1) has a narrower bandwidth or higher external Q since the sum of the real parts of the S_{11} zeros is negative; for a symmetrical filter the zeros lie on the $j\omega$ axis and this sum is zero. The internal coupling values are readily obtained numerically from the computed pole-zero response [5]. The coupling bandwidths between the m and nth resonances are given in Table III; zero refers to the manifold, and five, to the waveguide at the opposite end. The negative sense of the 1-4 coupling yields imaginary axis S_{12} zeros, whereas a positive sense places them on the real axis.

The filters were assumed to be connected in parallel with the manifold through short lengths of transmission line as described in the next section. Both annulling cavities were placed between the filters and the virtual short-circuit provided by the rectangular waveguide filter. In this location, their design is relatively insensitive and many configurations are possible including a long waveguide stub without cavities, but tuning of the chosen design is still critical. They form a reactive stub whose only degrees of freedom (neglecting losses) are the frequencies at which the phase crosses zero and 180° at some reference plane. In the selected design, the annulling cavities are tuned quite close to the operating band to reduce the size of their coupling apertures and thus their reflections at 14 GHz. Each annuller introduces an additional S_{12} null in each channel approximately at its resonant frequency.

The nominal manifold separations are one-quarter guide wavelength from the rectangular waveguide filter input (virtual short-circuit) to the first annulling cavity, one-half wavelength to the next annuller, one wavelength to the first filter, and a half between filters. Since a half guide wavelength at the transmit frequencies is approximately three-quarters at the receive, the residual reflections from the two filters tend to cancel in the receive band as do those of the annulling cavities. The one wavelength separation between the filter and annulling cavity allows the waveguide to be parted to facilitate fabrication and alignment. (For a transmit-only multiplexer, the virtual short-circuit of the rectangular filter would be replaced by a physical short-circuit).

TABLE II
POLES AND ZEROS OF LOW PASS PROTOTYPE (MHz)

Poles	S_{11} Zeros	S_{12} Zeros
-26.6 + j20.4	-4.0 + j 8.7	0. + j80.0
-11.3 + j44.2	-11.2 + j40.0	(2)

(Port 1 is the manifold end. S_{22} zeros are negative conjugates of S_{11} zeros.)

TABLE III
COUPLING BANDWIDTH VALUES (MHz)

	m	n	Coupling	
Port 1 (manifold)	0	1	45.4	(B_1)
	1	2	46.0	
	2	3	60.1	
	3	4	83.6	
Port 2	4	5	106.2	(B_2)
	1	4	-10.5	

The final values of the manifold separations and annulling cavity couplings were found by optimizing to match the stored complex reflection coefficient of the idealized multiplexer across a 120-MHz band. The tuning of the manifold-end resonance in each filter was adjusted during this procedure. An essentially perfect match was achieved. (This same basic approach has been used for computer designs up to twelve contiguous channels with twelve percent bandwidth; in these cases, however, a more complex annulling arrangement was used, and the manifold (0-1) and internal (1-N) couplings, plus the end resonance in each filter, were involved in the optimization.)

The narrow telemetry filter which is approximately 240 MHz below the TV channels was designed using the measured reactance of the other components. It was adjusted to have very low reflections in the TV transmit and receive bands as will be discussed.

III. Single Filter on Manifold

The filters in this multiplexer are coupled to the sidewall of the waveguide. This technique avoids excitation of the cylindrical TM_{01x} filter cavity modes which would be excited with broad-wall coupling and almost certainly degrade the receive band performance. Furthermore, the (transverse) sidewall currents decrease with increasing

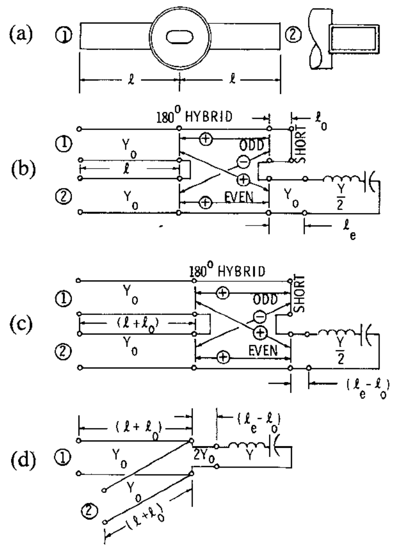


Fig. 3. Diagrams for sidewall manifold coupling.

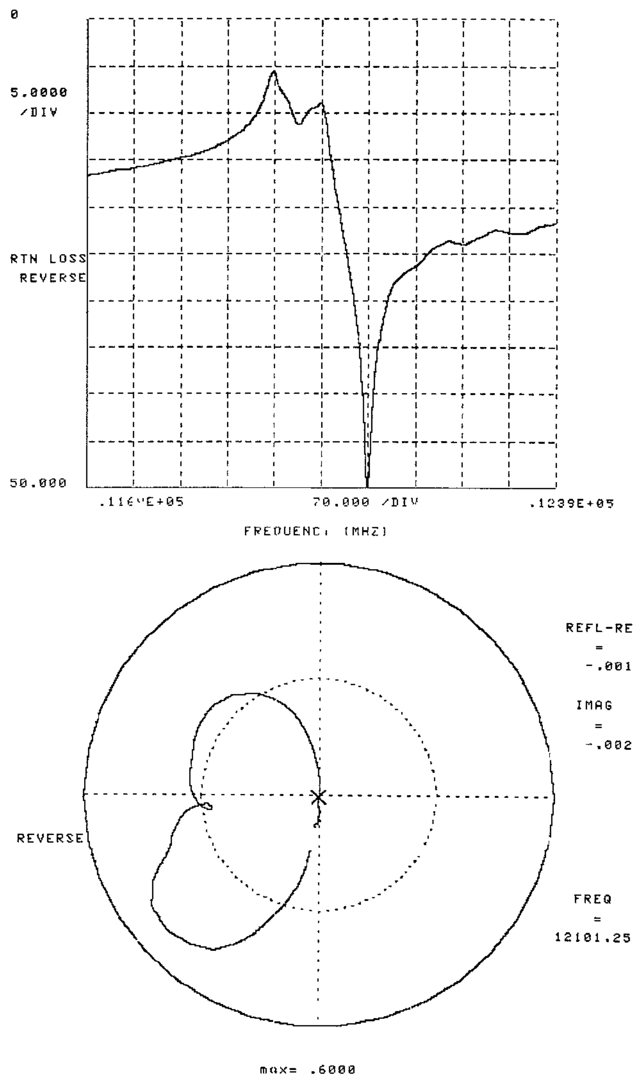


Fig. 4. Return loss and reflection coefficient of a single filter on a matched manifold.

frequency thereby offsetting some of the response skewing caused by increasing aperture coupling.

A cavity coupled to the waveguide sidewall is illustrated diagramatically in Fig. 3(a). Since the structure has a symmetry plane, its response may be analyzed in terms of odd and even excitation modes which may be exactly

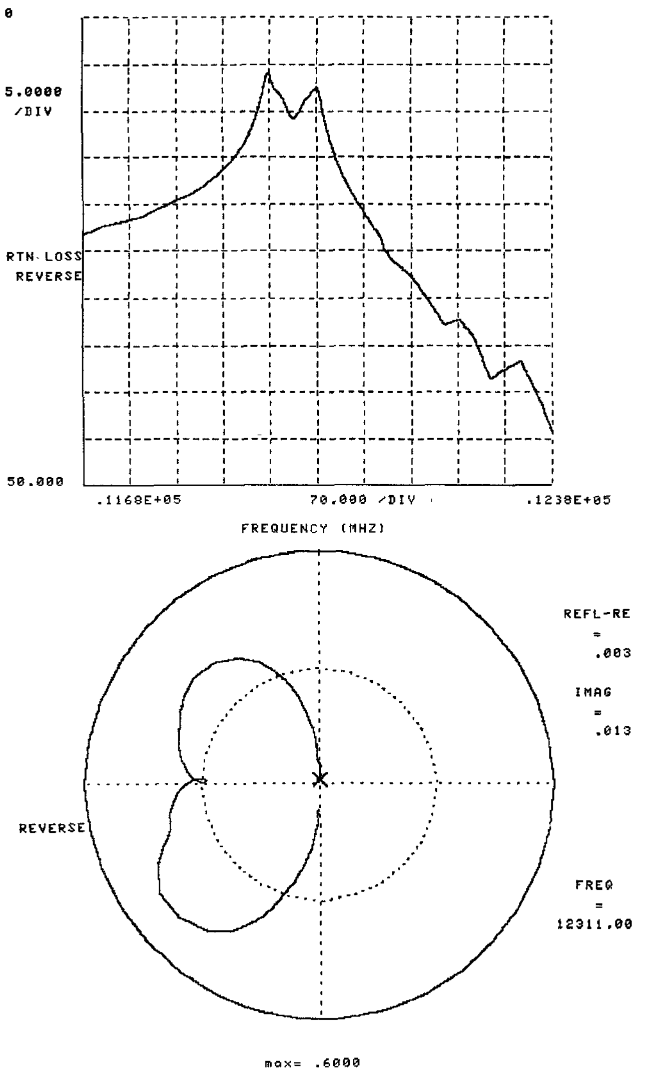


Fig. 5. Return loss and reflection coefficient of a single filter on a matched manifold with a tuning aperture.

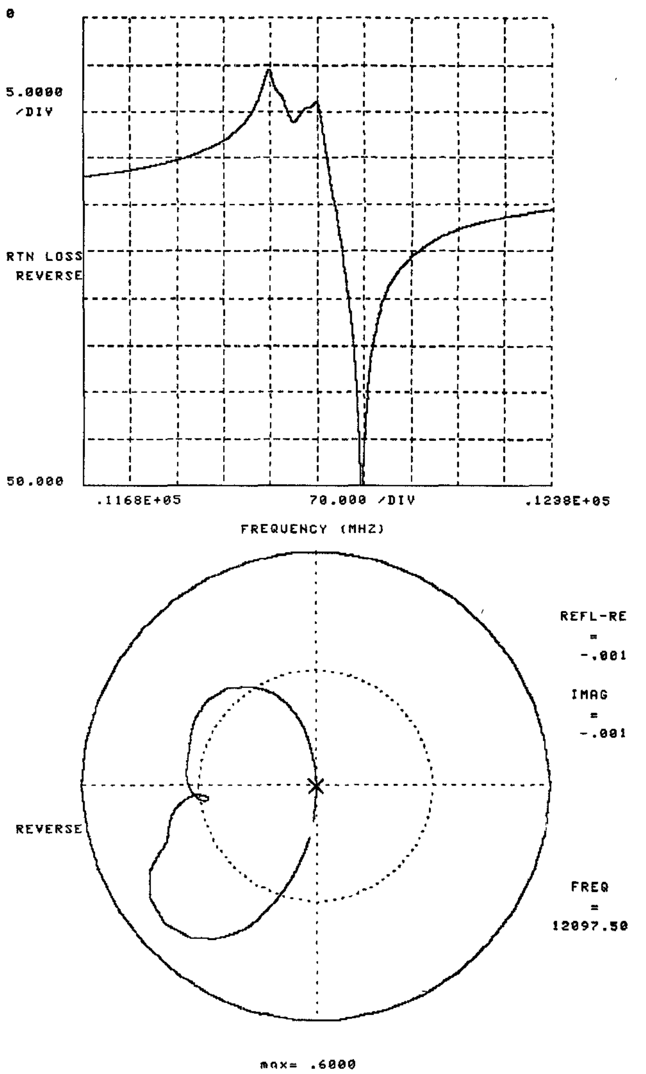


Fig. 6. Computed response of a single filter on a matched manifold.

related to the actual waveguide ports 1 and 2 by the use of an ideal 180-degree hybrid junction as shown in 3(b). For example, in an unperturbed piece of waveguide, the odd mode appears to be shorted in the center, and the even mode open. If these terminations are placed on the hybrid, it is seen that signal is coupled from 1 to 2 without reflection as expected. Placing an aperture backed by a nonresonant cavity in the sidewall shifts the phase of the even-mode response. (The even- and odd-mode reflections always have unit amplitude in the lossless case.) This can, at least for small perturbations, be approximated by a length of waveguide l_e on the even-mode port. The odd mode is affected similarly but to a much smaller extent. If the cavity is made resonant, the even-mode open-circuit is replaced by a series resonant circuit (which appears as an open off resonance), but the odd mode is unaffected, as depicted in 3(b).

This junction could alternatively be represented by a lattice of these odd- and even-mode reactances or by pi or T networks which would contain negative elements. However, the hybrid junction is convenient for changing reference planes as shown in 3(c), where it is moved to the right by l_o so that the odd-mode termination becomes simply a short-circuit, the same as for the unperturbed waveguide. In this case, the lattice, pi, and T networks all reduce to simply a shunt element, i.e., the circuit representation of the side-wall coupled cavity becomes a series-tuned circuit shunt coupled through a line with admittance $2Y_0$ and length $(l_e - l_o)$ to the midpoint of an elongated waveguide as depicted in 3(d).

Now consider how a broad-wall aperture backed by a nonresonant cavity may be used to modify the values in this circuit. Since this aperture is located in a region of magnetic field for the odd mode and electric for the even, their effective lengths, l_o and l_e, are increased and decreased, respectively. Thus, adjustment of its dimensions allows $(l_e - l_o)$ to be decreased or set equal to zero, in which case the resonant circuit appears directly in shunt, but the effective length of the waveguide is longer than its physical length.

The validity of this formulation over a frequency range greater than that of the multiplexer was verified experimentally. Fig. 4 shows the reflected signal due to a single filter on a matched manifold without a broad-wall tuning

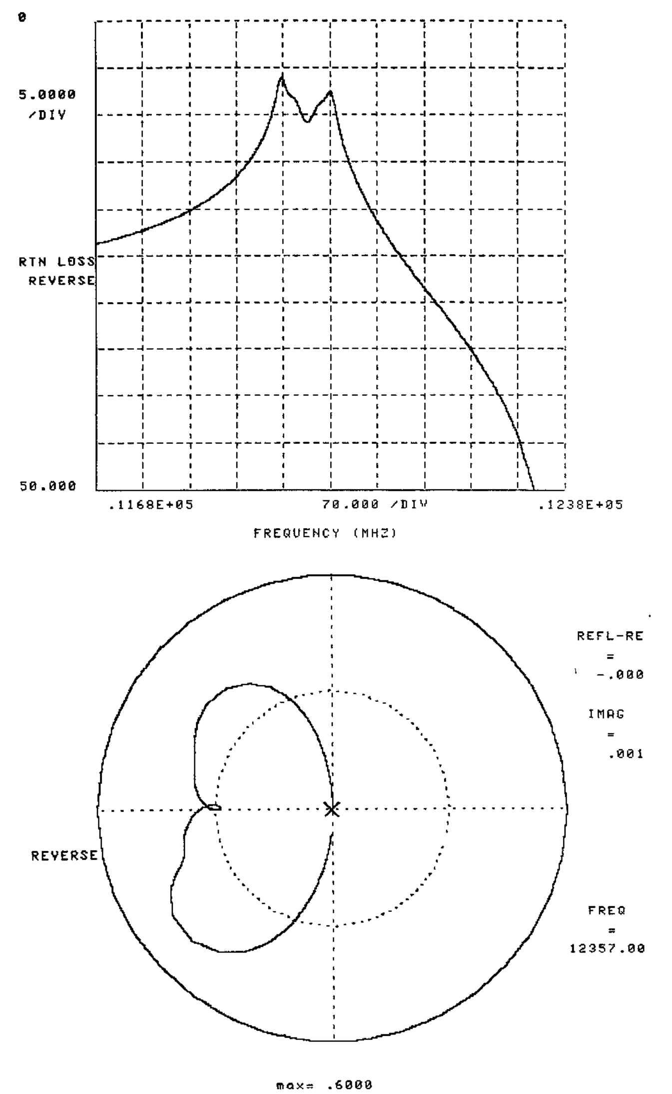


Fig. 7. Computed response of a single filter on a matched manifold with a tuning aperture.

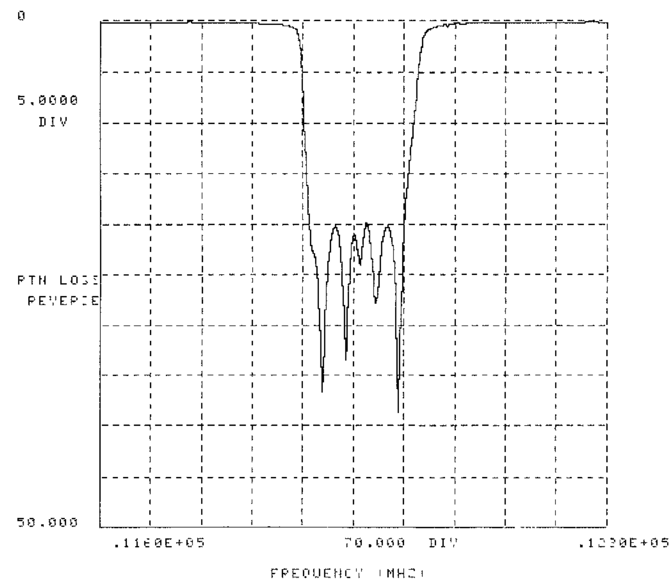


Fig. 8. Return loss of contiguous transmit diplexer.

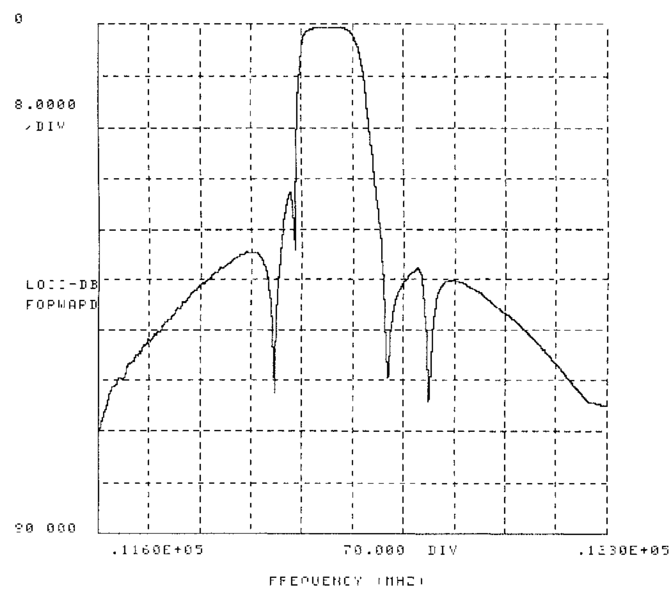


Fig. 9. Transmission loss from channel A to antenna (transmit band).

aperture in terms of return loss and complex reflection coefficient transformed to the effective shunt plane. The line length $(l_e - l_o)$ rotates the "Q circle" so that the phase becomes zero just above the resonance instead of at infinity, thereby introducing the deep null in the reflected signal. Fig. 5 shows the effect of a 0.237-in diameter tuning aperture; the reflection null has increased in frequency, and although not shown, the shoulder above the null is much lower.

These experimental data (complete S matrix) were computer processed [5] using the manifold transformations of Fig. 3 to determine the coupling values, resonant frequencies, and odd- and even-mode lengths. Fig. 6 shows the reconstructed response without the tuning aperture; $(l_e - l_o)$ is 0.029 in and l_o is virtually zero. Fig. 7 shows the tuning aperture case; the length of the coupling line $(l_e - l_o)$ has been reduced to 0.008-in, and l_o is 0.010 in; in addition, the coupling through the sidewall aperture was increased somewhat. These models provide adequate agreement through the receive band.

The use of the broad-wall tuning apertures to make the filters more nearly shunt elements allows the diplexer response to match the ideal case more accurately. It also provides a reduction in the reflections in the receive band. The TV filters were adjusted to have $(l_e - l_o) = 0.004$ in. In the case of the telemetry filter, the reflection null was tuned to the center of the TV transmit bands.

IV. Experimental Multiplexer Response

The return loss at the antenna port measured across the TV transmit band by an automatic network analyzer is shown in Fig. 8 (without the telemetry filter); it is better than 20 dB across the contiguous passbands of channels A and B. The return loss across the receive band also is better than 20 dB. The rejection response between channel A and the antenna port is shown in Fig. 9. There are two trans-

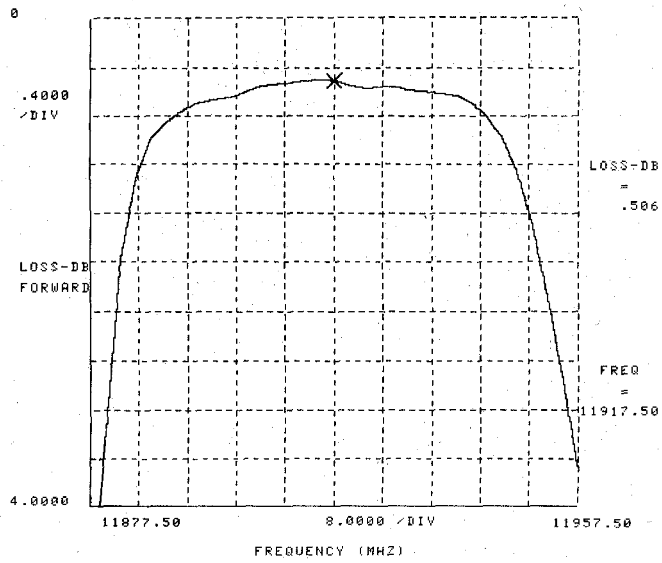


Fig. 10. Transmission loss from channel A to antenna (passband).

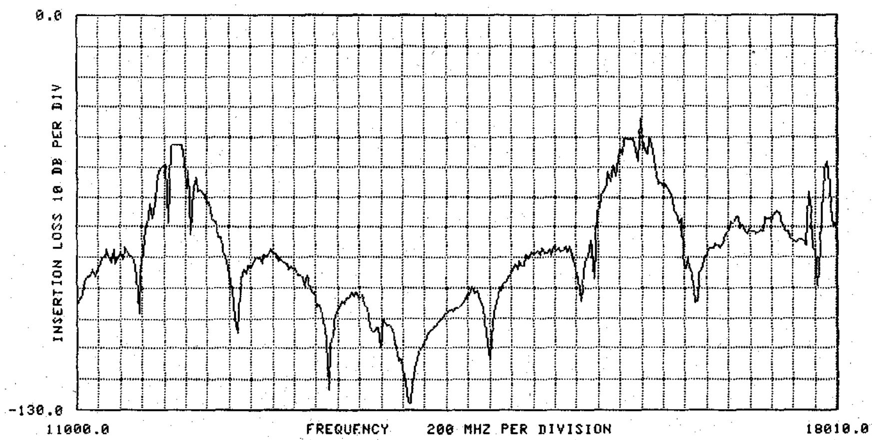


Fig. 11. Transmission loss from channel A to antenna (transmit/receive bands). The detector is saturated for levels greater than approximately −40 dB.

mission nulls on each side of the passband; the lower frequency member of each pair is due to the elliptic filter response; the upper one is due to an annulling cavity. The isolation at the lower edge of the channel B signal band is 26.5 dB. The channel B response is essentially a mirror image, and both channel responses are in good agreement with the computed responses.

Fig. 10 shows the passband insertion loss which is 0.51 dB at midrange and 3.68 dB at the crossover frequency of 11957.5 MHz at the right edge of the graph; the usable bandwidth is appreciably greater than the 27-MHz signal bandwidth. The effective unloaded Q of the entire multiplexer including losses in the manifold and receive isolation stub is approximately 7200 which is 0.61 times the theoretical value of 11 800 for a cylindrical silver TE_{111}-mode cavity of the shape used in the filters. The wideband rejection response measured on an automatic spectrum analyzer to a level of 130 dB is shown in Fig. 11; the response is saturated for losses less than approximately 40 dB. The isolation across the range of the received command and TV channels (14 000–14 310 MHz) is in excess of 100 dB. The isolation between either TV transmit channel and the receive port is greater than 100 dB in both the transmit and receive bands due to the combined effect of the transmit filters and the rectangular receive filter.

Acknowledgment

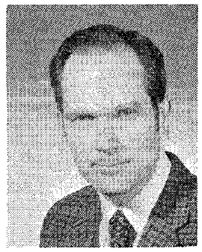
The author wishes to acknowledge the contributions of W. McLaughlin to the analysis and testing, L. Bickel to computer techniques, R. Dietz to the mechanical design, and M. Marcelli and M. Curtis to the testing.

References

[1] A. E. Atia, "Computer-aided design of waveguide multiplexers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-22, pp. 332–336, Mar. 1974.
[2] M. H. Chen, F. Assal, and C. Mahle, "A contiguous band multiplexer," *COMSAT Tech. Rev.*, vol. 6, no. 2, pp. 285–307, Fall 1976.
[3] J. D. Rhodes and R. Levy, "Design of general manifold multiplexers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, pp. 111–123, Feb. 1979.
[4] G. L. Matthaei, L. Young, E. M. T. Jones, *Microwave Filters*,

Impedance Matching Networks and Coupling Structures. New York: McGraw-Hill, 1964, pp. 965–999.

[5] H. L. Thal, Jr., "Computer-aided filter alignment and diagnosis," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, pp. 958–963, Dec. 1978.



Herbert L. Thal, Jr. (M'58–SM'82) was born in Mount Vernon, NY, on February 15, 1932. He received the B.E.E., M.E.E., and Ph.D. degrees in electrical engineering from Rensselaer Polytechnic Institute, Troy, NY, in 1953, 1955, and 1962, respectively.

From 1953 to 1956 he was a Research Associate at R.P.I. In 1956 he joined the General Electric Power Tube Department, Schenectady, NY, where he performed research and development on circuits and beam interactions in fixed-frequency and voltage-tunable magnetrons, multiple-beam klystrons, and distributed amplifiers. In 1967 he transferred to the Re-Entry Systems Department, King of Prussia, PA, where he was involved with the prediction and control of radar cross section and the analysis of antennas. Since joining the Space Division in 1974, he has been concerned with the development and testing of microwave filters, multiplexers, and antenna feed systems, and with the analysis of reflector and lens antennas and radar scattering.

Dr. Thal is a registered professional engineer in Pennsylvania and is a member of Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

Microwave Filter Loss Mechanisms and Effects

HERBERT L. THAL, JR., SENIOR MEMBER, IEEE

Abstract —**The losses in coupled cavity filters due to coupling apertures, tuning screws, and surface roughness have been determined experimentally. The results show, for example, that the shapes of the coupling apertures have a significant effect on filter losses. They may be used to estimate how the effective unloaded Q of a filter, and thus its insertion loss, varies as a function of bandwidth, configuration (e.g., single- or dual-mode), and response type (e.g., Chebyshev or elliptic). Also, they may be used to predict other effects such as response skewing and spatial distribution of dissipated power.**

I. Aperture Losses

FIG. 1 SHOWS one of the solid copper test structures used for the investigation of aperture losses [1]. The aperture being tested is shown in the center; machined on each side is a 0.861-in diameter cylindrical waveguide section, which is one-half wavelength for the TE_{11} mode. The small excitation aperture shown at the right has a half-wavelength section on the near side and WR75 waveguide flange on the back side. A shorting plate is shown at the left. Thus, when the parts are stacked, a TE_{111}-mode cavity is formed between the test and excitation apertures, and a half-length, nonresonant cavity between the test aperture and the short. The irises are 0.020-in thick. The test procedure consisted of measuring the unloaded Q, Q_u, and the TE_{111}-mode resonant frequency of a cavity without a test aperture and then repeating these measurements on the same cavity for a series of progressively larger apertures. The accuracy required for this study was achieved by using an automatic network analyzer with direct computer reduction of the data to resonant frequency and Q values.

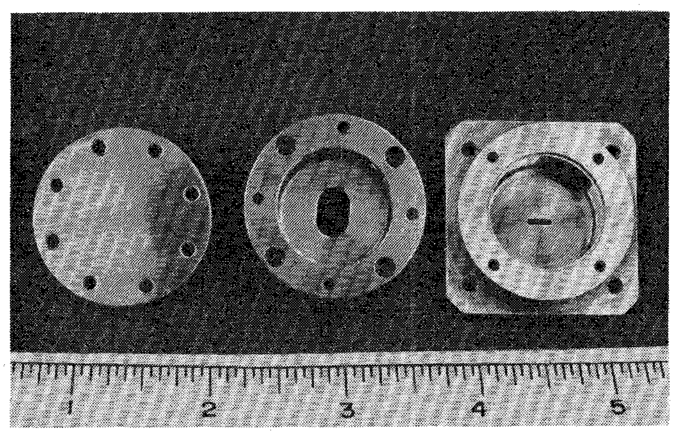


Fig. 1. Copper test structure, scale in inches.

Fig. 2 gives an equivalent circuit for the experimental configuration. The unperturbed cavity is represented by the series resonant circuit L, C, R. The test aperture and

Manuscript received November 27, 1982; revised March 26, 1982.

The author is with the Valley Forge Space Center, General Electric Co., Philadelphia, PA 19101.

0018-9480/82/0900-1330$00.75 ©1982 IEEE