Networks

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This article describes the design and development of the microwave and intermediate-frequency filters, which provide the frequency selectivity required in the TD-3 microwave transmitter and receiver, and of the equalizers, which compensate for amplitude and delay distortions introduced by the filters.

I. INTRODUCTION

As described in the first article of this issue, all selectivity required by the microwave transmitter and microwave receiver of the TD-3 radio system is provided by passive networks.[1]

The general plan followed in the design and development of these networks was to:

(*i*) Determine, from the system requirements, the preliminary in band and out of band loss requirements for the microwave networks.

(*ii*) Design microwave networks meeting the loss requirements with a minimum of amplitude distortion and with only broad-shape types of amplitude variation over the passband. The return loss requirement of the networks was generally specified as 30 dB minimum.

(*iii*) Calculate and measure the in-band loss distortions of the microwave networks to set preliminary requirements for the IF equalization networks.

(*iv*) Develop a preliminary design of IF equalizer meeting these preliminary requirements in order to perform system tests to establish that the network characteristics were satisfactory.

(*v*) Construct and measure characteristics of a substantial number of models of the microwave networks which introduce distortions. These results were used to specify the actual requirements of the IF equalizing networks.

(*vi*) Redesign the IF equalizing networks to compensate for the nominal distortions introduced by the microwave networks.

(*vii*) Establish the IF selective network requirements and design the network to meet the systems requirements. Equalization sections were included to compensate for the in-band distortions of the filter sections.

II. MICROWAVE TRANSMISSION NETWORKS

Frequency selective microwave transmission* networks are required in the TD-3 microwave transmitter to select the desired sideband output of the transmitting modulator, to suppress unwanted modulation products which might fall within the passbands of microwave receivers operating on other channels, to provide a means for efficiently coupling as many as six microwave transmitters to a common waveguide output while providing sufficient isolation to prevent interaction, and to prevent the emission of harmonics and other modulation products which might interfere with other radio services.

In the microwave receiver, selective microwave networks are required to efficiently couple as many as six receivers to a common waveguide, to prevent adjacent channel interference[2] and to suppress interfering frequencies which would degrade the system's performance.

2.1 *Transmitter Microwave Transmission Path*

The location and nature of each selective microwave component in the transmission path of the TD-3 microwave transmitter is shown in Fig. 1a.† In Fig. 1a, the 1336 type bandpass filter transmits the desired sideband from the transmitting modulator to the TWT amplifier, but suppresses the carrier leak, undesired sidebands, and other out-of-band modulation products and noise which would otherwise degrade the performance of the system. The 1336 type bandpass filter following the TWT provides most of the adjacent- and adjoining-channel attenuation,‡ the 1326A low-pass filter prevents transmission of harmonics of the modulated carrier, which might otherwise interfere with services operating on harmonically related frequencies or, by modulation with corresponding harmonics from other bays, produce interfering tones within the TD-3 system itself. Finally, the 1418 type channel combining network§ provides additional adjacent- and ad-

* We define transmission networks as those which pass the modulated signal.

† Fig. 1 includes only those components essential to this discussion. For complete block diagrams, see Ref. 3.

‡ Adjoining channels are considered to be those within the 3.7 to 4.2 GHz common-carrier band. At frequencies appreciably higher than those within this band, the performance of narrow-band waveguide filters is unpredictable because of the possibility of multimode propagation.

§ For reasons discussed in Section 2.3, the channel-combining and channel-separation networks are identical.

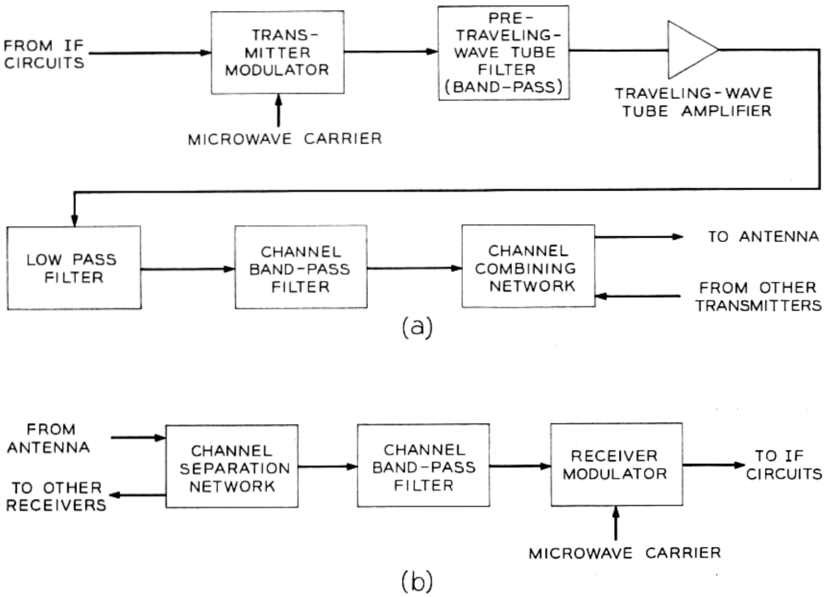

Fig. 1 — Location of selective microwave transmission components in TD-3 microwave (a) transmitter and (b) receiver.

joining-channel selectivity and couples the transmitter to the common antenna-feed waveguide without substantial loss.

2.2 *Receiver Microwave Transmission Path*

In the microwave receiver, as show in Fig. 1b, the 1418 type channel-separation network selects the desired channel, while passing the remaining ones on to the appropriate receivers via the common waveguide. The 1322 type bandpass filter provides supplementary adjacent- and adjoining-channel selectivity.

2.3 *Requirements, Constraints and Practical Considerations*

Identical in-band transmission objectives were established for each of the selective transmission networks, namely, a ripple-free insertion loss characteristic with a value no greater than 0.5 dB at midband, flat to ± 0.01 dB within the range f_0 (the midband frequency) ± 6 MHz, and to ± 0.10 dB within the range f_0 ± 10 MHz.

Similarly, common objectives governing permissible variations of network performance because of changes in ambient conditions were specified for all microwave networks as follows in response to any temperature change within $+40°$ to $+140°$F:

(*i*) The change in envelope delay distortion of any filter characteristic shall be less than 0.1 nanosecond within the range $f_0 \pm 6$ MHz.

(*ii*) The in-band insertion loss versus frequency characteristic of any selective network shall shift by no more than ± 1.0 MHz.

(*iii*) At frequencies of $f_0 \pm 70$ and ± 80 MHz, the insertion losses of the selective networks shall decrease by no more than 10 dB.

The first two of these requirements are necessary to ensure that the amplitude and delay characteristics of the microwave networks are at all times complementary to those of the correcting amplitude and delay equalizers, which operate at the intermediate frequency, and so exhibit relatively insignificant variation of characteristics with changes in ambient conditions. The third ensures that adjoining-channel interferences will not become intolerable. The first two requirements proved to be governing, and about equally restrictive, in all cases.

The out-of-band selectivity requirements for both the microwave transmitter and receiver were specified in terms of the total discrimination of the channel separation network and channel bandpass filters together; the distribution of these losses among the networks was left to the discretion of the networks' designer. In addition to those stated in the preceding paragraph these requirements were, for the transmitter, discrimination of at least 20 dB at frequencies $f_0 \pm 20$ MHz, and at least 80 dB at frequencies $f_0 \pm 70$ MHz; and for the receiver, discrimination of at least 20 dB at frequencies $f_0 \pm 20$ MHz, and of at least 100 dB at frequencies $f_0 \pm 80$ MHz.

Notice that the requirements for the transmitter and receiver can be fulfilled by very nearly identical sets of filters; the economics resulting from using actually identical sets proved to outweigh the disadvantages.

To provide an easily equalizable attenuation and delay characteristic, the maximally-flat-amplitude shape was chosen as the design basis for all selective microwave networks. The allocation of out-of-band insertion loss between the channel-separation and combining networks and the bandpass filters was dictated largely by the space available for the former, which, to permit side-by-side disposition and connection of transmitter-receiver bays, are best located horizontally and transversely in the 19-inch wide bays.

2.4 *Channel-Combining and Separation Networks*

The basic configuration selected for the channel-combining and separation networks is the constant-resistance one originally described by Lewis and Tillotson,[4] and used successfully in the TD-2, TH and TJ radio systems. However, instead of the rod-and-disc resonators used in Lewis and Tillotson's filters, the resonators of the band rejection filters which form the selective elements of the networks for the TD-3 system consist of rectangular waveguide cavities, spaced along the main waveguide at three-quarter wavelength intervals, and coupled to it by circular irises.* The cavities form filters with lower dissipative loss, improved stability, and higher passband return loss.

Asymmetry of the filter characteristic, caused by the series inductance inherent in the coupling irises, is corrected by capacitive compensating studs, centered on the broad wall of the main waveguide opposite each iris. Each resonator cavity is designed to resonate at a frequency slightly above the design center freqency of the filter and has a capacitive tuning screw, which is adjusted to resonate the cavity at exactly the design frequency during factory tuning.

To ensure minimum shifting of the transmission characteristics of these networks with temperature changes, the component filters are (except for the connecting flanges) completely fabricated from drawn WR229 waveguide tubing of copper-clad low-temperature coefficient nickel alloy (Invar). After fabrication, the filter is copper plated inside and out to 0.0002 inches thick to prevent corrosion on portions of the nickel alloy which have been exposed during machining, and to ensure low surface resistivity. A band rejection filter is shown in Fig. 2.

Available space limited the number of resonators per band-reject filter to three. This, in turn, limited the obtainable insertion loss to the separating port to about 18 dB at $f_0 \pm 80$ MHz, while satisfying the in-band requirements. These two parameters (the number of resonators, and the nondissipative insertion loss at one frequency) are sufficient to determine the electrical design of a maximally-flat amplitude filter.

Typical insertion-loss and delay characteristics are shown in Fig. 3. A detailed theoretical and experimental investigation has established that the asymmetries largely result from errors in the lengths of the waveguide sections connecting the resonators. A more refined design procedure which greatly improves the symmetry has resulted.[6] The ohmic losses cause the in-band transmission to not conform with the stated requirements. Intermediate frequency amplitude equalizers compensate for this deviation (see Section II).

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* Rectangular resonators are also used in the analogous band-rejection filters for the TJ, TH, and TM radio systems. See Ref. 5.

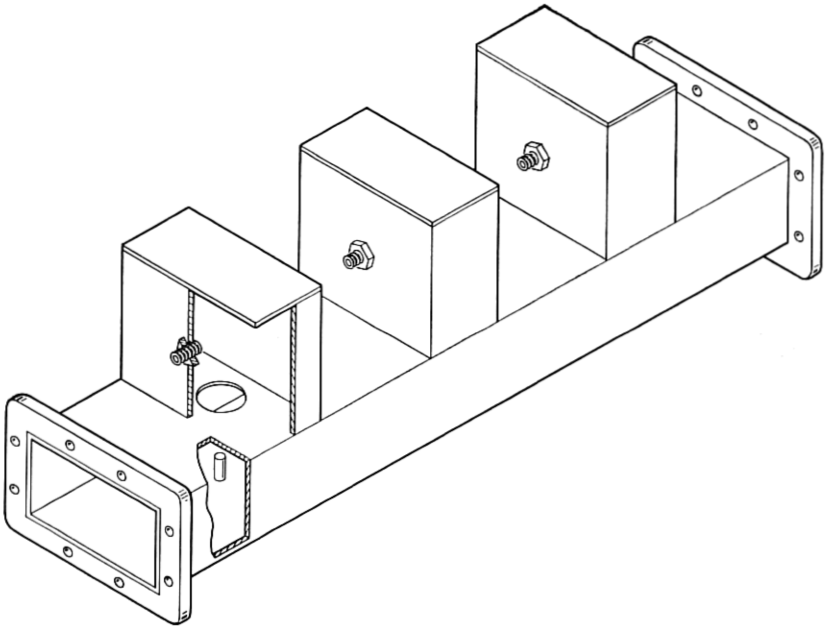

Fig. 2— Band reject filter used in channel combining and separating network.

A total of 24 codes of networks is required, one for each microwave channel. The input (4B) and output (5B) hybrid junctions are common to all codes; the band-reject filters, however, differ among codes dimensionally, as well as in center frequency, to provide the most uniform possible performance from channel to channel. Figure 4 shows a 1418 type channel separating or combining network.

2.5 *Channel Bandpass Filters*

The allocation of 18 dB of discrimination at $f_0 \pm 80$ MHz to the channel separating and combining networks leaves a residue of 82 dB at this frequency to be provided by the channel bandpass filters. This can be provided by a five-resonator maximally flat amplitude filter, which will also satisfy the in-band transmission requirements. Such a filter becomes somewhat unwieldy when composed of the usual assembly of approximately half-wavelength resonators, bounded by single cylindrical-post susceptances and coupled by three-quarter wavelength lines (necessary to avoid spurious coupling from evanescent modes created by the susceptive posts).

Effort was therefore directed toward reducing its size without sacrificing performance. A possible alternative, the direct-coupled filter, was rejected because of the extreme mechanical precision which would be required in forming and positioning the high-susceptance coupling elements needed in such narrow bandwidth filters. A second alternative, to select a configuration of susceptive elements which would reduce coupling caused by evanescent modes and so permit closer spacing of the resonators, appeared to be more desirable.

An array of three cylindrical posts, equal in diameter and uniformly spaced across the waveguide (see Refs. 7 through 10), was found to permit reducing the coupling lines to one-quarter wavelength long. The posts have reasonable dimensions for manufacturing. Cylindrical posts were favored because they are the simplest to form within close tolerances and to position accurately within the waveguide. Data were obtained empirically relating post array dimensions to susceptance. This information, gathered with the most accurate measuring methods available, was used in the filter designs, following Mumford's[11] method.

After several prototype filters performed satisfactorily, an IBM 7094

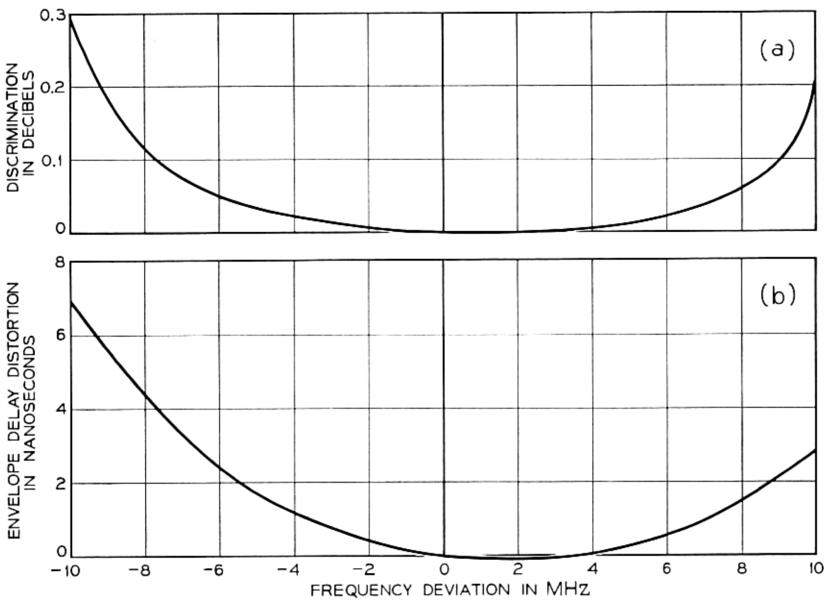

Fig. 3 — Discrimination and delay characteristics of channel combining and separating networks.

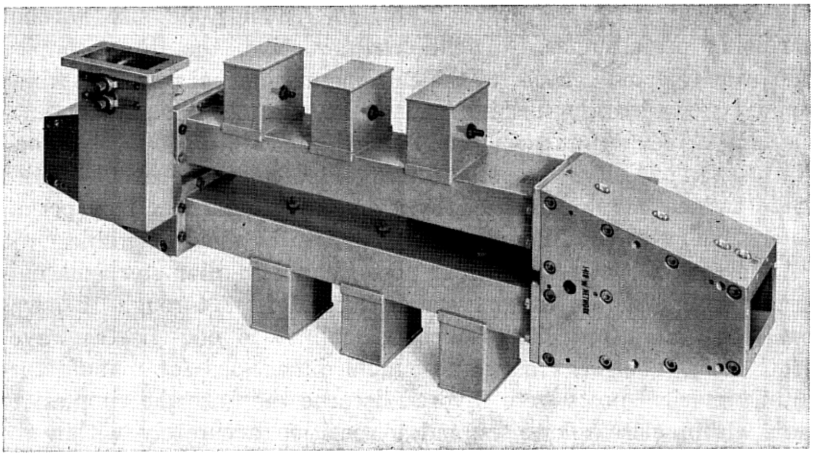

Fig. 4 — Channel combining and separating networks.

computer was programmed to mechanically and electrically design filters for the remaining channels. As with the band rejection filters, each of the 24 filter codes entails a distinct physical arrangement of elements for the greatest possible uniformity of performance from channel to channel.

Each filter consists of a section of drawn WR229 waveguide of copper-clad low-temperature coefficient nickel alloy.The posts, composed of the same alloy, are cylindrically ground to a tolerance of ± 0.0002 inch, then copper plated and inserted in holes formed in the broad walls of the waveguide by electrical discharge machining (to avoid burrs), and soft-soldered in place. After fabrication, the entire filter is copper-plated to a thickness of 0.0002 inch to prevent corrosion and to ensure a low-resistivity surface. Again, each resonator is designed to resonate slightly above the nominal center frequency of the filter and has a capacitive tuning screw to permit exact adjustment to the specified frequency.

In Fig. 5, a 1322 channel bandpass filter is cut to show the internal arrangement of susceptive posts. Figure 6 shows typical measured amplitude and delay distortion characteristics, as well as the asymmetry which is the subject of investigation. Ohmic losses also contribute to the departure from the ideal "maximally flat" characteristic, which is compensated for in the IF amplitude equalizer. The channel bandpass filters, as well as the channel separating and combining networks, have a frequency-temperature coefficient of -5 kHz per degree Fahrenheit.

2.6 *Filter Before TWT*

The first article in this issue mentioned that excessive delay distortion in the transmitting channel bandpass filter (which preceeded the TWT), and TWT amplitude to phase conversion were producing excessive modulation noise in the initial TD-3 installation.[1] To alleviate this the bandpass filter was moved to follow the TWT. To prevent overloading the TWT by the carrier leak and the unused sideband produced by the transmitting modulator, a very broad bandpass filter was designed and put between the transmitting modulator and the TWT. It was to be only selective enough to acceptably reduce signals while introducing negligible amplitude and delay distortion in the passband. Because its amplitude and delay distortions are very small, this filter can be made of all copper construction in a WR229 waveguide. This filter, number 1336, consists of three triple-post resonators. It provides 15 dB of discrimination at f_m ± 70 MHz, less than 0.1 ns of delay distortion, and immeasurably small amplitude distortion in the band.

2.7 *Low-pass Filter*

The attenuation characteristics of the bandpass filters are unpredictable at frequencies for which the waveguide containing them can propagate modes besides the dominant mode. Furthermore, even without high-order modes, the distributed element resonators of the filters will exhibit resonances at frequencies which are approximately integral multiples of the operating frequencies, and so introduce spurious passbands. Therefore, to ensure adequate attenuation of harmonic frequencies of the microwave carrier, it is necessary to provide supplementary high-frequency attenuation. A low-pass filter

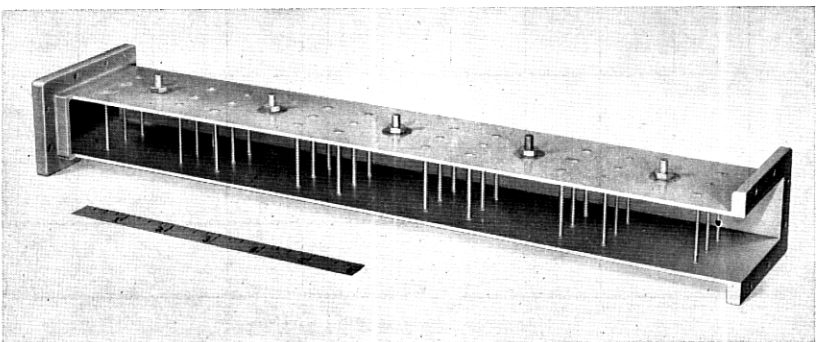

Fig. 5 — Channel bandpass filter cut to show internal construction.

which follows the highly nonlinear TWT of the microwave transmitter does this.

The requirements placed on the low-pass filter were: insertion loss less than 0.2 dB at 3.7 to 4.2 GHz, and greater than 50 dB at 7.4 to 12.64 GHz; and return loss greater than 30 dB from 3.7 to 4.2 GHz.

A coaxial structure of alternating sections of low- and high-impedance coaxial line provides the required stop-band loss. In-line coaxial-to-waveguide transducers couple the coaxial element to a WR229 waveguide, as shown in Fig. 7.

The coaxial attenuating element has seven sections, mid-shunt termination, constant K, and a nominal image impedance of 50 ohms. The inside diameter of the outer conductor is small (0.288 inches) to minimize spurious passbands from propagation in modes other than the TEM in the normally attenuating frequency range. To ensure minimum loss in the passband, only two thin discs were used as dielectric supports, one at either end of the coaxial structure. Shock and vibration tests have proven this design mechanically adequate.

The cutaway in Fig. 7 shows details of the coaxial-to-waveguide transducer. The tuneable stub and axial probe form the inductive

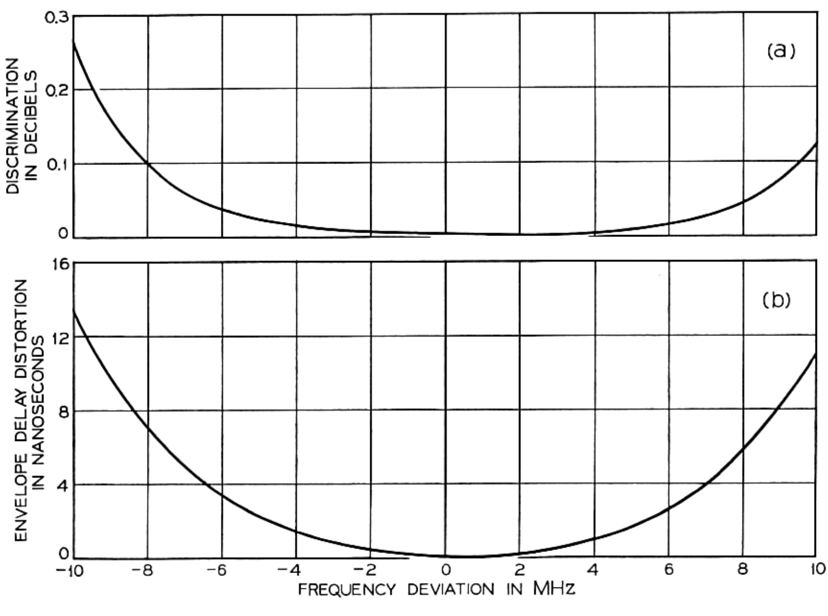

Fig. 6 — Discrimination and delay characteristics of channel bandpass filter.

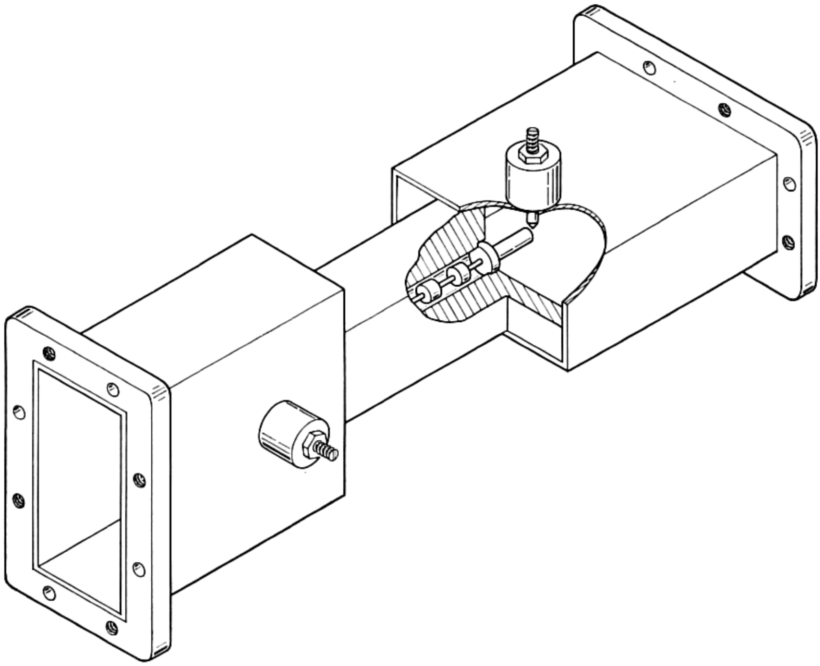

Fig. 7 — Microwave low pass filter.

elements and the air gap between them forms the capacitive element of a series-resonant coupling loop with reactances small enough to fulfill the inband return loss requirement. Besides being a convenient in-line arrangement of waveguide parts, these transducers permit the waveguide sections to rotate 90 degrees as shown in Fig. 7, thus eliminating the need for a waveguide twist, which would otherwise be required.

The insertion loss and return loss characteristics of the low-pass 1326A filter, are shown in Fig. 8.

2.8 *Miscellaneous Microwave Networks*

The TD-3 transmitters and receivers contain conventional attenuators, directional couplers, terminators, and a single-resonator filter. There is no need to describe them here.

III. IF BANDPASS FILTER

An equalized IF bandpass filter, with extremely good passband performance, was required to increase the selectivity of the TD-3

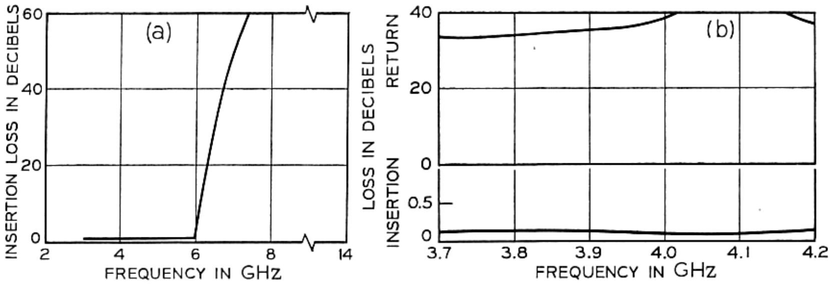

Fig. 8. (a) Insertion loss (3.0 to 12.6 GHz) and (b) insertion and return loss characteristics of low pass filter in TD-3 frequency band.

receiver. This filter, with a passband from 62-78 MHz, selects against out-of-band interference which could cause improper operation of the AGC or carrier resupply circuits. The filter allowable passband distortion has to be kept to a minimum. Therefore, the following stringent objectives were specified: a maximum peak delay distortion of ±0.3 nanoseconds, a maximum peak amplitude distortion of ±.05 dB, and a minimum return loss of 30 dB. This filter was called 745A; its stopband objectives are shown in Fig. 9.

3.1 *Electrical Design*

Insertion loss and image parameter methods of designing the filter were explored. The insertion loss technique is a more exact method of design in that the insertion loss and return loss characteristics can be specified, and a circuit configuration realizing these characteristics is then synthesized. By the image parameter method a circuit configuration is first chosen, then the circuit parameters are adjusted until the required characteristics are obtained. The image parameter design was chosen for this filter because a circuit that better compensates for parasitic elements was more readily obtained.

The filter consists of two double M-derived, asymmetrical, transformed half-sections. The transformation was performed to obtain capacitors to ground that can be adjusted to compensate for the parasitic capacitances. Figure 10 is a schematic diagram of the filter. Figure 9 shows typical measured characteristics of the unequalized filter. The delay distortion is 12 nanoseconds, and the loss distortion is 0.3 dB over the 62-78 MHz frequency range.

A slight change in the frequency of the filter attenuation peaks by temperature and aging will cause a large change in the delay at the band edge. Therefore greater emphasis was placed on equalizing the

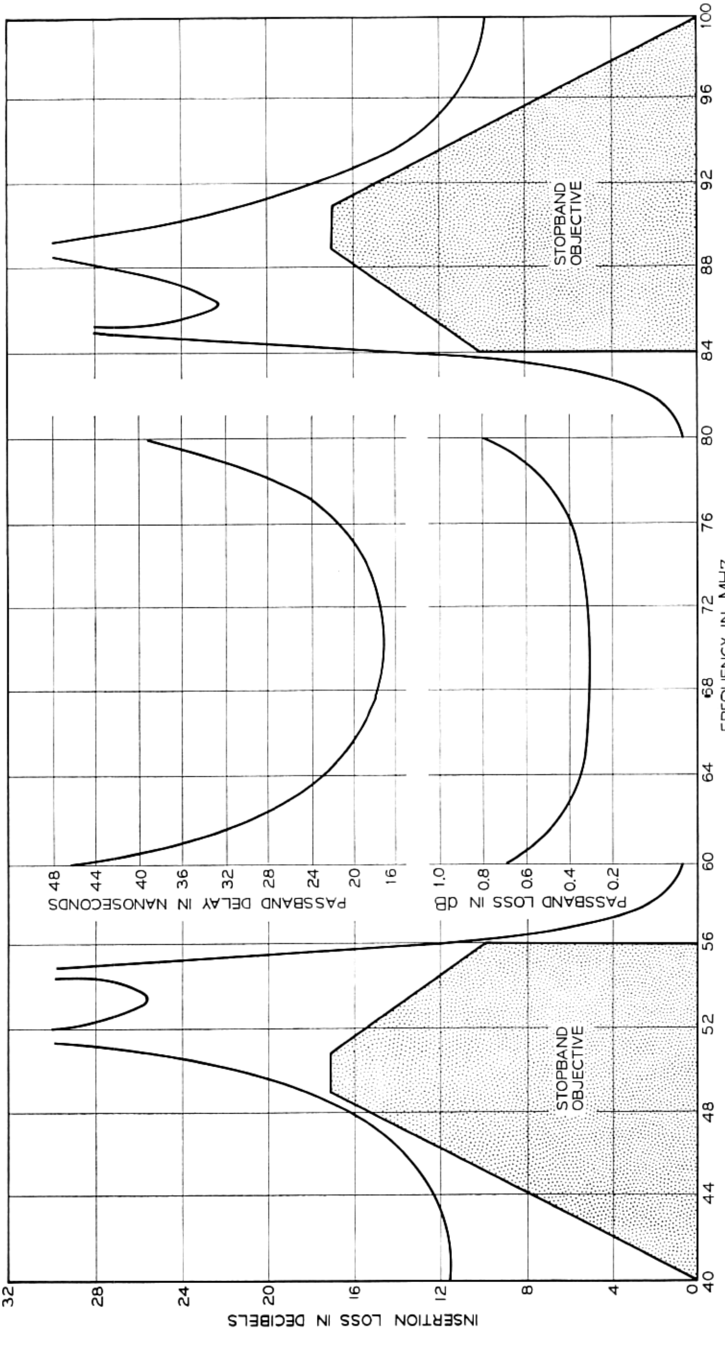

Fig. 9 — Characteristics of the unequalized 745A filter.

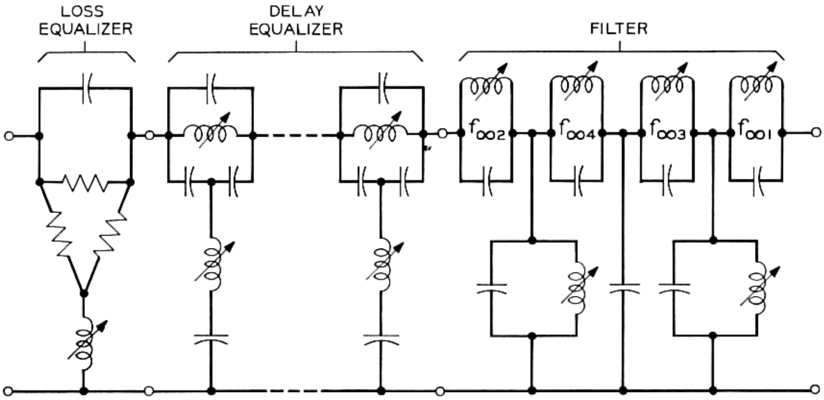

Fig. 10 — Schematic diagram of the 745A filter.

filter over the 64-76 MHz frequency range than over the 62-78 MHz frequency range. Four bridged-T 360° all-pass sections are used to equalize the delay distortion of the filter to within the specified objectives. One bridged-T amplitude equalizer section is used to compensate the combined amplitude distortions of the filter and the four delay equalizing sections. The filter and its equalizers are connected in tandem as shown in Fig. 10.

The typical measured characteristics of the equalized filter are shown in Fig. 11. The delay distortion is ± 0.25 nanoseconds over the 62-78 MHz frequency range. The peak-to-peak loss distortion is 0.05 dB over the same frequency range. The return loss (compared with 75 ohms) is greater than 30 dB over the 62-78 MHz frequency range.

3.2 *Mechanical Design*

The 745A filter is housed in a sectionalized, cast-aluminum base with a folded sheet aluminum cover (see Fig. 12). Each inductor is enclosed in an individual compartment in the channel to minimize the magnetic coupling between inductors. The closely controlled mechanical tolerances achieved with this type of construction help to keep constant any parasitics caused by proximity. Printed wiring paths are used where possible to eliminate varying lead lengths of hand wiring.

3.3 *Summary*

A filter with extremely good performance has been developed by using a digital computer program to get the best electrical design and

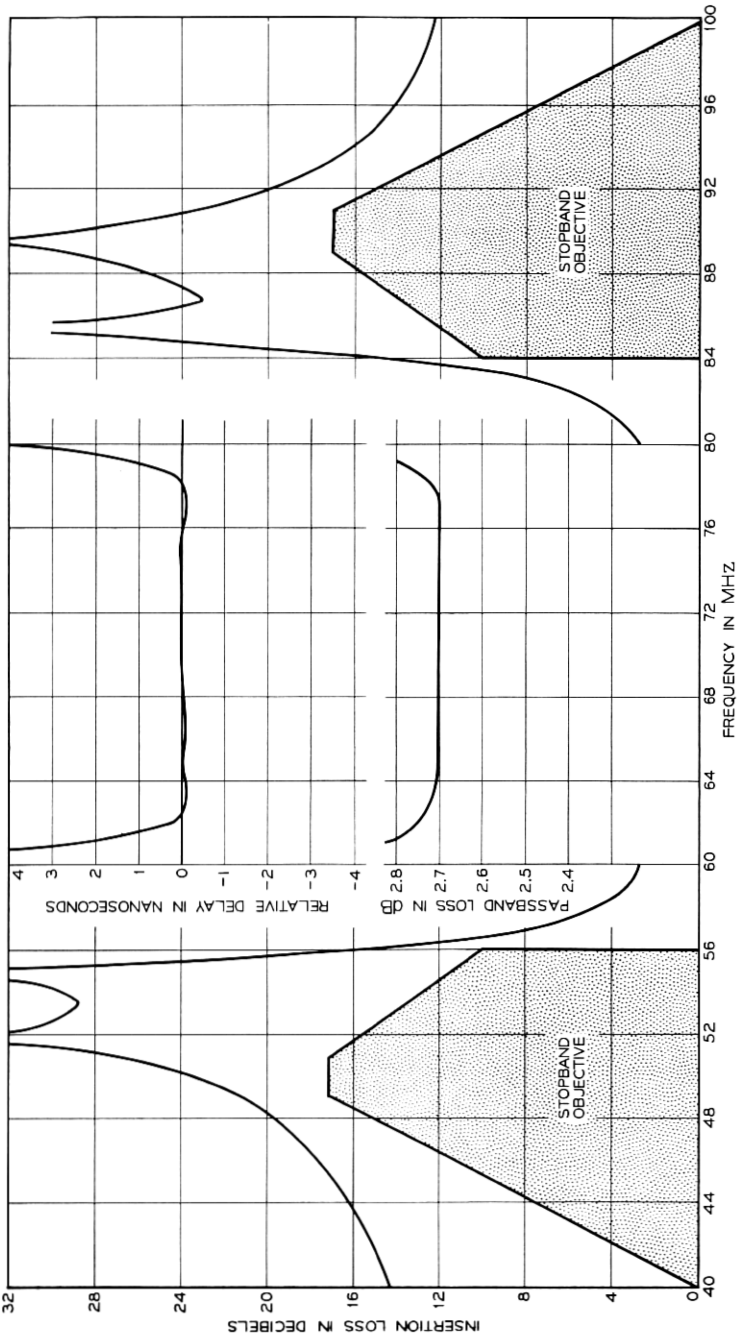

Fig. 11 — Characteristics of the equalized **745A** filter.

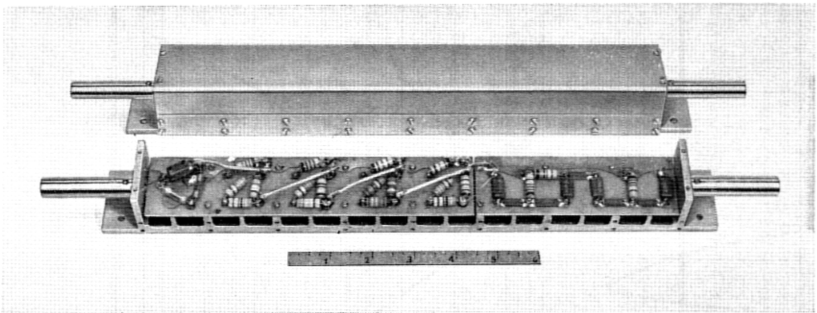

Fig. 12 — The 745A filter.

parasitic-corrected element values, and by using a mechanical configuration with closely controlled tolerances to keep constant parasitics caused by proximity.

The 745A filter represents a great improvement in performance over previous IF filters. For example, the manufacturing limit on the delay distortion is reduced to one third that of the TD-2 intermediate frequency filter (574A), the limit on loss distortion has been cut in half, and the minimum return loss has been increased by three decibels.

IV. BASIC IF DELAY AND AMPLITUDE EQUALIZERS

4.1 *Function*

Most circuits in the TD-3 repeater contain active devices which are relatively broadband and therefore introduce little transmission distortion. The IF filters used in the TD-3 system are self equalized. Therefore almost all of the delay and amplitude distortion is introduced by the No. 1418 microwave channel separation networks and No. 1322 channel bandpass filters.

There are 24 No. 1418 microwave channel separation networks and 24 No. 1322 channel bandpass filters which operate in the 24 microwave channels. Many of these networks and filters were measured to determine similarities in passband characteristics about their respective center frequencies.

These passband characteristics are in two groups, those where the beat oscillator is above the signal frequency and those where the beat oscillator is below the signal frequency. Consequently, only two basic IF equalizers are needed. The 794A equalizer is used with the first group of channels, and the 793A equalizer is used with the second. The delay and amplitude distortions of the first group is the mirror image (about 70 MHz) of the second group.

4.2 *Requirements*

The average characteristics of two No. 1322 and two No. 1418 networks in tandem was determined from the measurements. To reduce the influence of measurement inaccuracies, the average delay and amplitude characteristics were approximated by a sixth order polynomial (using the least squares method). These smooth delay and amplitude characteristics are shown in Figs. 13 and 14.

TD-3 was designed for 1200 message circuits, which requires the transmission characteristic of the basic equalizers to be carefully controlled over a 12 MHz band. However, in order to provide some transmission performance margin for possible future, but presently unknown, system loads, the transmission characteristic is controlled over a 16 MHz band. Therefore the basic equalizers were designed to meet stringent inband tolerances between 62 and 78 MHz.

The objective of the basic equalizers is to equalize the one-hop

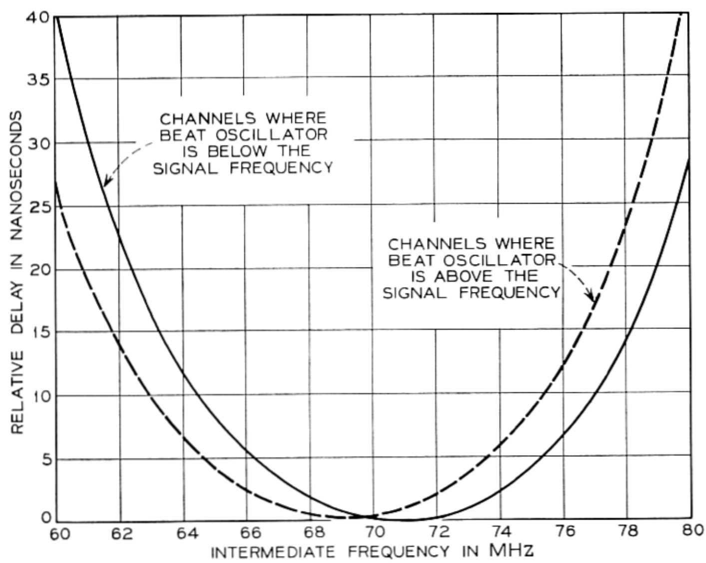

Fig. 13 — Average one hop delay characteristic.

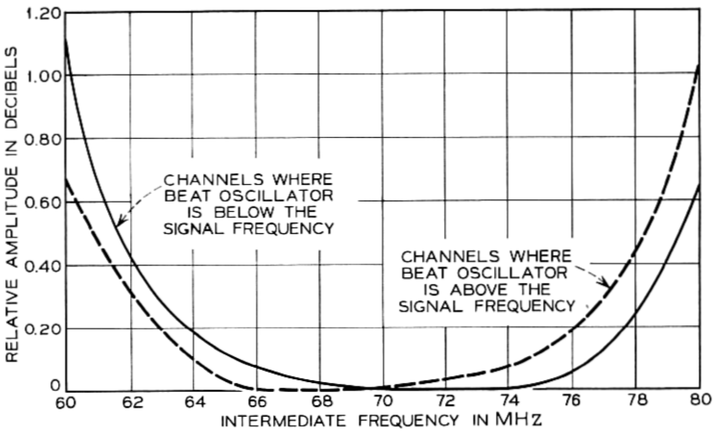

Fig. 14 — Average one hop amplitude characteristic.

characteristics, shown in Figs. 13 and 14, to within the following tolerances:

Frequency (MHz)	Delay		Insertion loss (dB)		Minimum return loss (dB)
	Slope	Ripple	Slope	Parabolic	
62–78	—	±0.2 ns	—	±0.05	30
60–80	≤0.1 ns/MHz	±2.0 ns	0.20 max.	±0.50	25

The input-output impedance level of the basic equalizers is 75 ohm-unbalanced. The constant resistance, unbalanced bridged-T type of delay and insertion loss sections is used to achieve a high return loss.

4.2.1 *Theoretical Design*

Since the design procedure for both of the basic equalizers is the same, the design of only one equalizer, the 794A, is presented, along with the results of both equalizers.

The requirements on the basic equalizers are very stringent, and a great deal of work was needed to achieve a balance between performance and cost. Hand calculations indicated that the required delay match could be achieved using six 360 degree all-pass delay sections. In order to optimize these results, the b (stiffness parameter) and fc (resonant frequency parameter) for each delay section were used as input to a computer program. This program matched the sixth order polynomial requirements with equalizers of six, seven, and eight delay sections. A final design using seven delay sections was chosen as a compromise between the number of delay sections and practical considerations such as element values and the values of b. Experience shows values of b between 2.5 and 10.0 to be the most desirable to build in the IF range.

The computer delay match over the 62-78 and 60-80 MHz bands was ± 0.02 and ± 0.17 ns, respectively.

Since there is a finite Q associated with the inductors used in the delay sections, the delay sections are lossy. The total delay of the delay equalizer is concave down, and it is a characteristic of this type of delay equalizer to have a similar loss shape. Therefore, the loss of the delay sections tend to self-equalize the loss shape of the microwave networks.

There is also some delay shaping associated with amplitude equalizers. One amplitude section meets the design objectives for this equalizer; however, the single section is complex and would be difficult to tune in the IF range. Therefore, two simpler sections in tandem were designed which are simpler to tune and can be constructed in the standard mechanical assembly. The computed amplitude match with the two sections was ± 0.03 and ± 0.08 dB over the 62-78 and 60-80 MHz ranges, respectively. Figure 15 shows the complete schematic for the 794A equalizer. The schematic for the 793A equalizer is the same except that it has one additional delay section.

4.2.2 *Laboratory Measured Performance*

The measured performance of the first laboratory models of 793A and 794A equalizers is:

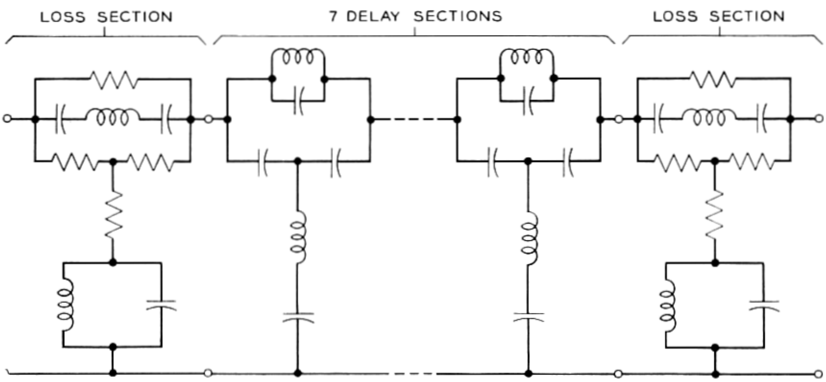

Fig. 15 — Schematic for the TD-3 basic delay and amplitude equalizer, 794A.

Frequency (MHz)	Delay match (ns)	Loss (dB)	
		Insertion match	Return
62–78	±0.15	±0.04	35 min
60–80	±0.50	±0.20	33 min

The insertion loss at 70 MHz is 5.3 dB, and the absolute delay at 70 MHz is 180 ns.

Additional models of both equalizers were constructed, using capacitors and resistors that were within ±1 percent of those used in the first laboratory models. Using the ±1 percent variations in element values, which is the closest tolerance available, the same degree of delay match and high return loss was not achieved.

4.2.3 *Manufacturing Procedure and Requirements*

The manufacturing requirements for the basic equalizers are:

Frequency (MHz)	Delay match (ns)	Loss (dB)	
		Insertion match	Return
62–78	±0.3	±0.05	28 min
60–80	±2.0	±0.50	25 min

The delay and amplitude tolerances of ±0.3 ns and ±0.05 dB is the best accuracy that can be achieved on the equipment available for testing the basic equalizers. In order to obtain this accuracy, the equalizers are measured by comparison; that is, an unknown equalizer is measured against a standard 793A and 794A equalizer. This adjusting and testing is done on a test set which visually displays the input and output return loss of the equalizer being tested as well as the delay and insertion loss differences between the equalizer being tested and standard equalizer.

Each delay and loss section is tuned individually before it is assembled into a complete equalizer. Once the equalizer is completely assembled, it is returned by the comparison method.

4.2.4 *Mechanical Design*

The basic equalizers are stacked in assembly. Each section, delay and loss, is assembled on a printed wiring card and inserted in cans mounted on a frame. The cards are then wired together and each can is sealed with a separate cover, as the bottom of Fig. 16 shows. After the equalizer has been tested, the completed assembly is enclosed by

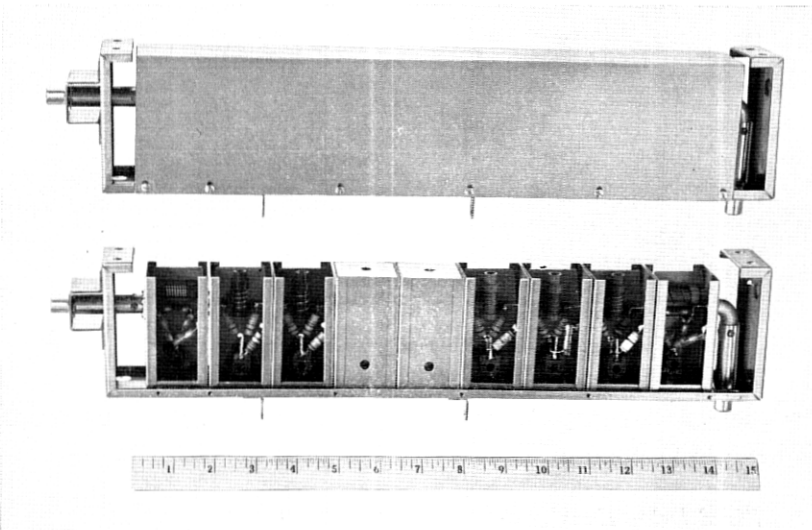

Fig. 16 — 794A equalizer.

a single cover, as shown at the top of Fig. 16, to increase mechanical strength.

Each section is in a separate can to reduce the interaction between sections; Printed wiring boards reduce the differences that could occur with hand wiring. The two inductors in the 360-degree all-pass bridged-T delay sections are mounted perpendicular to each other to reduce the magnetic coupling between them, and to enable the inductors to be adjusted with the covers over the individual cans. The bridging capacitors are hand wired between the inductors; the distance between inductors is about the same as the bridging capacitor length, so there is little room for extra lead. Figure 17 shows the printed wiring card and its assembly. This assembly has the good mechanical symmetry needed for a high return loss.

V. "MOP-UP" EQUALIZATION

5.1 *Function*

Since each basic equalizer is used with different sets of RF networks, it was not possible with only one fixed IF equalizer in each repeater to achieve the nearly perfect equalization required by the TD-3 radio system. Additional "mop-up" equalization is required on

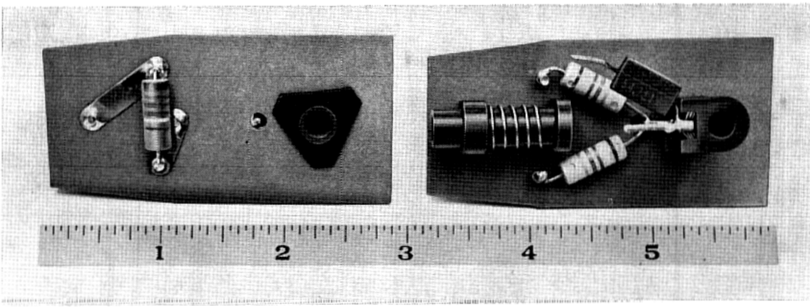

Fig. 17 — Assembly of printed wiring cards.

each hop. Delay slope is the most prevalent and potentially serious component of the residual characteristic of each hop.

5.2 *Fixed Delay Equalizers*

5.2.1 *Requirements*

The amount of delay slope required to mop-up a radio channel is subject to change. Thus, slopers must be added, removed or replaced from time to time. In order that the absolute delay at 70 MHz can be retained with minimum effort, it is required that all mop-up equalizers have the same absolute delay at 70 MHz.

Based on the TD-3 field trial experience and the per-repeater linear envelope delay distortion requirements, it was concluded that four codes of slope equalizers were needed.

Nominal slopes of +0.25, −0.25, +0.50 and −0.50 ns per MHz are required, along with an equalizer which has the same absolute delay at 70 MHz as the slopers, but which is nominally without envelope delay distortion (zero delay slope) and amplitude distortion.

Requirements for these equalizers are:

Frequency (MHz)	Delay match (ns)	Loss (dB)	
		Amplitude	Return
62–78	±.3	±0.02	33.0 min
60–80	±.5	±0.05	30.0 min
70		1.0 max	

In addition, the absolute delay at 70 MHz for all five equalizers is within ±1.0 ns.

5.2.2 *Design and Performance*

The electrical design and adjustment procedures for all five mop-up equalizers are the same as those for the IF basic equalizers.

The codes and manufacturing requirements for each equalizer are:

Code	Delay slope (ns per MHz)	Insertion loss, 70 MHz (dB)
918A	−.50	0.35 ± 0.1
918B	−.25	0.35 ± 0.1
919A	+.25	0.60 ± 0.1
919B	0.00	0.50 ± 0.1
920A	+.50	0.80 ± 0.1

Additional manufacturing requirements common to all five equalizers are:

Frequency (MHz)	Delay match (ns)	Amplitude distortion (dB)	Return loss (dB min)
62–78	$\pm.3$	$\pm.03$	33
60–80	$\pm.5$	$\pm.05$	30

Absolute delay at 70 MHz $= 23.4 \pm 1.0$ ns.

VI. 747A LOW PASS FILTER

6.1 *Function*

The IF preamplifier used in the TD-3 repeater generates harmonics of the 60-80 MHz fundamental signal. This amplifier is followed by an IF main amplifier. When the IF basic equalizer is placed between these units, a substantial delay ripple occurs. This ripple occurs because the absolute delay of the basic IF equalizers is considerably greater at the fundamental frequencies than at the harmonics. Thus, the fundamental frequency is substantially delayed in relation to its harmonics.

Because of small nonlinearities in the main amplifier, the delayed fundamental and its harmonics produce modulation products at the fundamental frequency which have a different phase from the incident fundamental frequency. This causes the substantial delay distortion. When a low pass filter, which attenuates the second and third harmonics of the 60-80 MHz range, is inserted before the repeater equalizer, the ripple is greatly reduced. The low pass filter prevents the harmonics from reaching the IF main amplifier.

6.2 *Electrical Design*

6.2.1 *Requirements*

It is not practical to incorporate the low pass filter as part of the basic equalizer circuit. Further, separation of the low pass filter from the equalizer is consistent with the original philosophy of keeping the basic repeater equalization separate from other IF networks.[3]

For reasons described in Section IV, the equalized passband of the low pass filter is 62-78 MHz. However, the tightest inband requirements are placed on the 64-76 MHz band. The requirements for the low pass filter are:

Frequency (MHz)	EDD (ns)	Loss (dB) Insertion	Loss (dB) Return
64–76	±0.05	±0.01	30.0 min
60–80	±0.10	±0.05	27.0 min
70		1.0 max	
120–240		21.0 min	

The input and output impedances are 75 ohm, unbalanced.

6.2.2 *Theoretical Design*

The 747A is a fifth order Cauer elliptic low pass filter which is designed with an equal ripple passband and equal minima stopband. The computed passband loss distortion of this filter with a finite Q of 150 is 0.1 dB. The minimum return loss in the passband is 34 dB. The delay equalization is achieved with two 360-degree all-pass bridged-T sections, and the loss equalization with one bridged-T amplitude section. Figure 18 is a schematic diagram of the 747A filter.

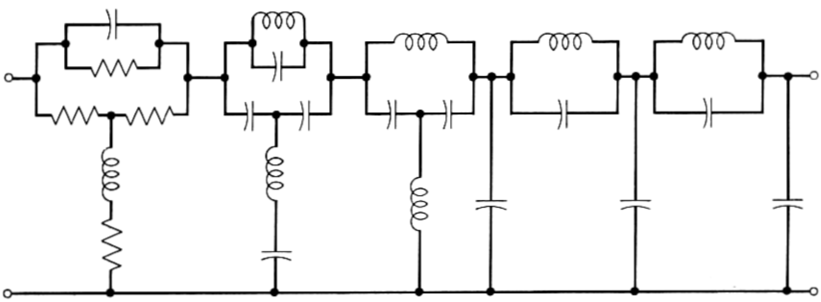

Fig. 18 — Schematic for the 747A filter.

6.2.3 *Measured Performance*

The measured performance and manufacturing requirements for the **747A** filter are:

Frequency (MHz)	EDD (ns)	Loss (dB) Insertion	Loss (dB) Return
64–76	±0.30	+0.02	30.0 min
60–80	±0.30	±0.05	27.0 min
70		0.60 max	
120–240		21.0 min	

Figure **19** shows its typical characteristics.

6.2.4 *Mechanical Design*

The **747A** filter is assembled on a printed wiring board which is mounted on an eight-section cast aluminum base. The inductors are mounted on the component side of the printed wiring board and shielded from each other by walls in the base. The capacitors and resistors are mounted on the conductor side of the board. Figure **20** shows the assembly.

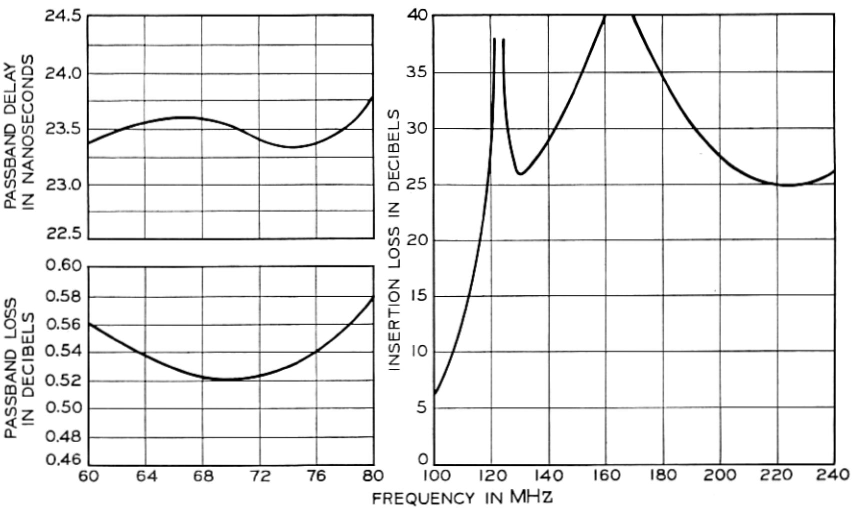

Fig. 19 — Characteristics of the **747A** filter.

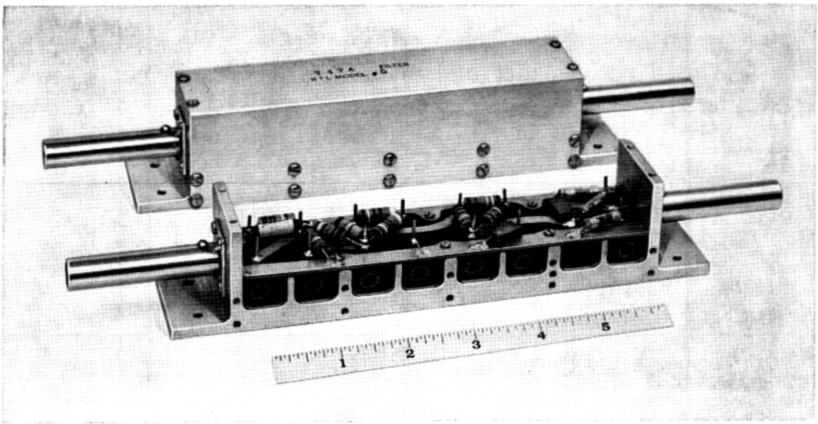

Fig. 20 — 747A filter.

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