Microwave Transmitter and Receiver

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The basic blocks of the TD-3 radio system are the microwave transmitter and receiver units. Since up to 140 of these units may be connected in tandem, each must meet rather stringent requirements in order to satisfy the over-all system objectives described in the previous paper. The transmitter and receiver units use solid state semiconductor devices (except for the traveling wave electron tube amplifier in the transmitter), and operate from a single −24 volt battery source. The detailed design of these units is presented here with particular attention to thermal noise, cross modulation noise, selectivity, and equalization.

I. INTRODUCTION

The microwave transmitter and receiver units constitute the basic building blocks for the TD-3 radio system. They provide the gain necessary to compensate for the transmission loss of the station-to-station air path, and means for changing that gain during periods of radio signal fading. These units also include the selectivity required to keep the radio channels of a route properly separated.

As previously pointed out,[1] the Schottky barrier diode modulator is used in microwave receivers now being manufactured; therefore, most of our derivations pertain to this type of receiver. However, our analysis of receiver selectivity requirements pertains to a receiver equipped with a parametric amplifier, whose use was expected when the selectivity requirements were originally established.

II. DESIGN FEATURES AND PERFORMANCE OBJECTIVES

The distinction between a design feature and a performance objective is not always clear because they may be related; therefore, somewhat arbitrary distinctions are made.

2.1 *Design Features*

The electronic circuits in the microwave transmitter and receiver, with one exception, use solid-state semiconductor devices. The exception is the traveling wave electron tube amplifier in the transmitter used to deliver the high power required at microwave frequencies. Modern, solid-state low voltage battery plants were in use in the telephone plant at the beginning of the TD-3 development. Since these low voltage sources were compatible with transistors, the microwave transmitter and receiver have been designed to operate from a single −24 volt battery plant. A regulator[2] closely controls the voltages applied to solid-state devices and thereby eliminates degradation of performance with changes in battery voltage. An inverter-regulator,[2] powered from the −24 volt source, provides the several voltages required by the traveling wave tube.

At intermediate frequencies, compatibility with the TD-2 system and with the 100A protection switching system[3] is achieved by using a 70 MHz frequency, by providing 75 ohm impedance circuits, and by using signal powers consistent with the switching system.

2.2 *Performance Objectives*

2.2.1 *Thermal Noise*

The first paper in this issue established the system thermal noise objective and allocated 34.9 dBrnC0 to the radio line.[1] The system model for the radio line assumed 140 repeaters; therefore, the per-repeater (or transmitter-receiver pair) thermal noise objective is: $34.9 - 10 \log_{10} 140 = 13.4$ dBrnC0.

As the input signal to a microwave receiver decreases because of fading, the thermal noise contribution from that receiver increases and eventually controls total system noise. Noise of 55 dBrnC0 in a message channel has been established as the maximum permissible in a commercially acceptable circuit. Based on the best information available, a fade range of 40 dB for each receiver should insure that yearly outage time because of fading will provide acceptable circuit reliability if a 2 for 10* protection switching arrangement is used. For this fade range, the allowable nonfaded thermal noise requirement for the receiver becomes $55 - 40 = 15$ dBrnC0, which is slightly more lenient than the 13.4 dBrnC0 above.

* Two protection channels for ten working channels.

2.2.2 *Cross Modulation Noise*

The first paper in this issue allocated 36.3 dBrnC0 of cross modulation noise to the radio line.[1] Early in the development, this allocation was divided into two equal parts, one assigned to equalization misalignment resulting from changes in ambient temperature or humidity (discussed in Section 3.4.3), and one assigned to noise caused by residual amplitude and delay distortion in the signal path after equalization. Assuming that the noise adds on a power basis, the per-repeater allocation for the latter part is: $33.3 - 10 \log_{10} 140 = 11.8$ dBrnC0.

With 11.8 dBrnC0 allotted to a transmitter-receiver pair, the related magnitudes of amplitude and delay distortions were needed. To simplify finding these magnitudes, we assumed that:

(*i*) The transmission characteristics would need to be well controlled only over the frequency region $f_c \pm 6$ MHz, that is, the region occupied by the carrier and the first order sidebands of the signal.
(*ii*) The noise resulting from residual amplitude distortion would be negligible.
(*iii*) The residual delay distortion would include only slope and ripple components.

Using methods described in Refs. 4 and 5, the delay slope or ripple amplitudes which would produce 11.8 dBrnC0 of cross modulation noise were determined to be:

Delay slope, 0.15 nanoseconds per MHz.
Delay ripple, 0.2 nanoseconds peak amplitude for ripple periods of 10 MHz and greater; 0.6 to 0.2 nanoseconds peak amplitude for ripple periods from 2 to 10 MHz.

It is obvious that, if both delay slope and delay ripple distortions are present, power addition of the noise must not exceed 11.8 dBrnC0.

In addition, parabolic amplitude distortion must be controlled to achieve satisfactory system baseband amplitude response. An objective of ± 0.05 dB at $f_c \pm 6$ MHz, relative to the response at f_c, was specified.

Late in the development period, methods became available which permitted predictions to be made of the cross modulation noise produced by simple types of amplitude and delay distortions for the type of pre-emphasis used.[6] It was determined that either the original delay slope or the ripple allowances would consume the 11.8 dBrnC0 repeater allocation. Also, the noise from these two sources generally

TABLE I—MAGNITUDE OF DISTORTION*

Distortion type	Magnitude of distortion at ±6 MHz
	(nanoseconds)
Linear delay	0.9 (slope of 0.15 nanoseconds/MHz)
Parabolic delay	4.6
Cubic delay	1.1
	(decibels)
Parabolic gain	5.0
Cubic gain	0.12
Quartic gain	0.3

* Distortion causing 11.8 dBrnC0 of cross-modulation noise in the top message channel.

could be expected to add on a power basis. The magnitudes of specific types of distortion, each of which would account for 11.8 dBrnC0 of cross modulation noise, are given in Table I.

Notice that no allocation was made to cross modulation noise caused by AM to PM converting devices. For that reason, noise from this source had to be kept small relative to that from the other sources if system objectives were to be met.

2.2.3 *Interferences*

Spurious tones can be generated in a voice channel by interferences from inside and outside the microwave transmitter and receiver. To reduce these tones to acceptable levels, the interferences must be controlled by selective filters, by directional devices, and by reduction of leaks between circuits.

The objective for a single frequency tone interference in a voice channel is a maximum of −70 dBm0 under "no fade" conditions.[1] For interferences dependent on the received carrier power, the objective is relaxed to −42 dBm0 at the fade depth corresponding to a radio channel noise of 55 dBrnC0, the protection switch request point. In addition to the −70 dBm0 objective, spurious tones at intermediate frequencies must be kept sufficiently low so as not to materially affect the automatic gain control range of the microwave receiver, or the switch-on point of the carrier-resupply circuit used to replace the carrier in a failed or very deeply faded radio channel.

With the TD-3 frequency plan, two types of adjacent channel interference occur: spillover of higher order sidebands from the disturbing channel, and direct transfer of modulation from the disturbing to the disturbed channel. Direct transfer is usually called direct adjacent channel interference. The magnitudes of these interferences are determined primarily by the antenna side-to-side coupling loss, the degree of cross polarization discrimination achieved, and by the repeater selectivity. The design objective was to keep noise contributions from these sources negligible compared with the total noise objective.

The microwave transmitter and receiver must meet requirements for spurious radiation imposed by FCC regulations, and spurious radiation must not interfere with systems operating in the other common-carrier bands at 6 and 11 GHz.

2.2.4 *Baseband Amplitude Response*

The baseband signals applied to the FM terminal transmitter at the input to the radio line extend approximately from 0.5 to 6 MHz. The first paper specifies an objective of ± 0.25 dB for baseband flatness for each radio channel in an IF switching section.[1] Experience has indicated that this objective can be met if the variation in amplitude response of the microwave transmitter-receiver over the range of $f_c \pm 6$ MHz is held to ± 0.1 dB.

III. CIRCUIT DESIGN

3.1 *Circuit Description*

3.1.1 *Microwave Transmitter*

Figure 1 shows the microwave transmitter with typical signal powers. It accepts a -7 dBm signal at 70 MHz IF from the microwave receiver at a repeater station or from the FM terminal transmitter at a main station. It produces an output of $+37$ dBm at one of the 24 channels in the 3710–4170 MHz frequency range.

The IF limiter[7] removes unwanted amplitude modulation from the input IF signal and furnishes an IF control signal to the IF carrier resupply circuit. The IF carrier resupply[7] furnishes a carrier should the normal IF signal be lost.

If the IF signal at the limiter input is lost, the carrier resupply furnishes a tone-modulated FM signal to the transmitter modulator. In addition, an operated carrier resupply provides a dc output to bias the limiter to a high insertion loss state and thus attenuates noise at the limiter input during the lost carrier period. The reinserted carrier prevents AGC circuits in subsequent repeaters from driving

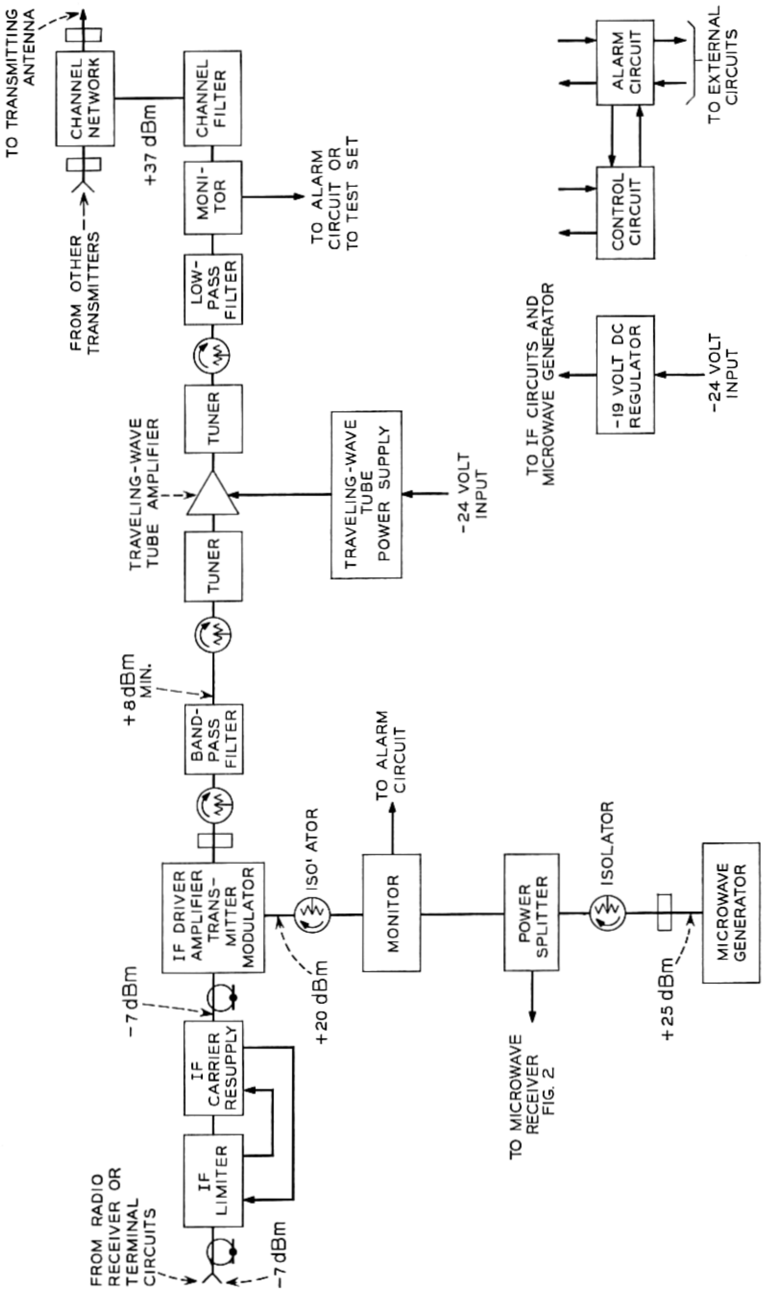


Fig. 1 — Block diagram of TD-3 microwave transmitter.

IF amplifiers to full gain and limiters from spreading high-level noise. Noise spreading would cause adjacent working channels to become unsatisfactory because of excessive noise spillover. The tone modulation on the resupplied carrier provides information to the automatic switching system to indicate that the channel is defective.

The IF driver amplifier-transmitter modulator[8] converts the 70 MHz IF signal to a microwave signal at the transmitter output frequency. The driver amplifier provides approximately 7 dB of gain. The modulator uses a waveguide hybrid junction and a pair of varactor diodes to achieve an upconversion gain of about 8 dB. The beat oscillator power, +20 dBm, is supplied by the microwave generator. The upper and lower sidebands and the beat oscillator leakage signal appear at the output port.

For some channels the modulator is used as an upper sideband upconverter; for the others, as a lower sideband upconverter. The sideband to be transmitted is selected by a band-pass filter. The undesired sideband, which serves as an idler frequency for the upconverter, is reflected by the filter and is dissipated in the reverse loss of an isolator between the upconverter and the filter. Allowing idler frequency currents to flow in this manner results in stable operation, as discussed in Ref. 8. The sideband selecting filter provides about 15 dB of loss to the beat oscillator frequency, and about 30 dB to the unwanted sideband. These losses assure that the power of the 2A-B product generated in the TWT by the beat oscillator and the unwanted sideband is acceptably low.

The desired input to the TWT is a frequency-modulated signal consisting of a carrier and FM sidebands. However, if the signal is transmitted through a circuit that has amplitude or delay distortion, AM components are introduced into the signal. These components are transformed by the relatively high AM-PM conversion coefficient of the TWT into PM components, and they appear as cross-modulation noise. By using low selectivity in the sideband selecting filter,[9] the "in-band" amplitude and delay distortion is small and AM-PM cross-modulation noise from this source is negligible.

The traveling wave tube amplifier uses permanent magnet periodic focusing.[10] It provides approximately 34 dB of gain with an output power of +37 dBm and an amplitude flatness of 0.02 dB over the 12 MHz radio channel. The low pass filter following the amplifier provides 50 dB of loss to second and third harmonics of the carrier frequency.

A monitor, consisting of a directional coupler and a diode detector, is used during transmitter alignment and for monitoring the output power.

The channel filter provides selectivity for the transmitter to assure that energy spillover into adjacent channels is within tolerable levels.[9]

The channel network provides additional selectivity and permits the tandem connection of up to six bays for combining transmitter signals for application to the common transmitting waveguide and antenna.[9]

Isolators, tuners, and attenuators are used in the signal path as as required to improve return loss and transmission, and to adjust signal powers.

The microwave generator provides an output of +25 dBm at the appropriate 4 GHz frequency for use as a beat oscillator signal for the radio transmitter and receiver.[11] It uses a crystal oscillator-amplifier unit producing about +37 dBm at $\frac{1}{32}$ of the output frequency. Varactor diodes in tandem connected quadrupler, doubler, and quadrupler circuits provide the multiplication of 32.

The power splitter, used in repeater bays, is a directional coupler which divides the microwave generator power between the transmitter and receiver loads. Resonance isolators are used to control the reverse transmission characteristics in the generator power distribution circuit. This insures that signals in the transmitter do not couple into the receiver (or vice versa) and produce interfering tones.

A directional coupler and diode detector at the output of the microwave generator are used for testing and monitoring the output power.

An inverter regulator unit supplies electrode voltages to the TWT amplifier. Other circuits are powered from a −19 volt regulator. Control and alarm circuits for alignment and surveillance are provided.

3.1.2 *Microwave Receiver*

Figure 2 shows the microwave receiver with typical signal powers. The receiver accepts an input signal on one of 24 radio channels located in the 3710–4170 MHz frequency range. The carrier input power depends on path length and has a nominal value of −28 dBm. The output power is −7.0 dBm at 70 MHz IF.

A channel separation network selects the proper channel and applies it to the channel band-pass filter which provides additional selectivity to reduce adjacent channel interference. An isolator

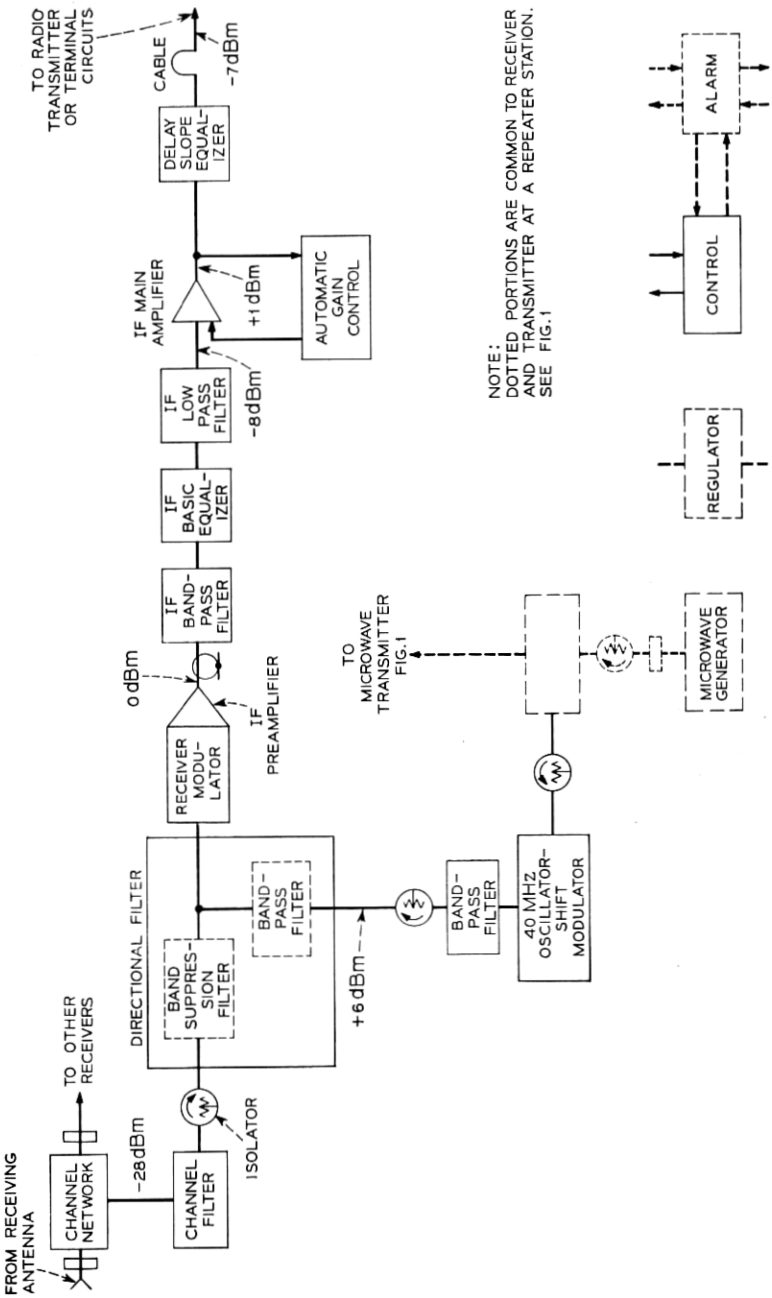


Fig. 2 — Block diagram of TD-3 microwave receiver.

terminates the image frequency generated in the receiver modulator.

The modulator preamplifier consists of an unbalanced Schottky barrier diode modulator and a low noise IF preamplifier.[12] The incoming microwave and beat oscillator signals are combined and applied to the input port of the modulator by a directional filter. The modulator translates the microwave input signal to the 70 MHz IF. The IF output of the modulator is applied to a five-stage preamplifier, which provides about 33 dB of gain. The gain can be adjusted to provide an output of 0 dBm for input signals ranging from −22 to −30 dBm. The noise figure of the modulator-preamplifier is about 6.8 dB.

The IF band-pass filter provides additional receiver selectivity.[9] Its amplitude and delay are internally equalized over the 64–76 MHz range. The basic equalizer corrects the amplitude and delay distortion of the transmitter and receiver microwave filters and networks over the 64–76 MHz range. The IF low-pass filter attenuates harmonics of the 70 MHz carrier frequency. It also is internally equalized over the 64–76 MHz range.

The IF main amplifier has a gain of 9 dB under normal received signal conditions and produces an output of +1 dBm.[7] During receiver input signal fading, the amplifier variolosser circuits, controlled by the AGC circuit, maintain the output power constant for "down" fades as great as 35 dB and "up" fades as large as 10 dB. A linear (in dB) meter indication of received signal power is provided by a linearizing circuit which monitors the total variolosser control current.

Delay slope equalizers are used as required following the main amplifier to compensate for residual delay slope distortion. A length of IF coaxial cable, as required, is used to build all channels in a switching section to the same electrical length. This insures that large baseband signal phase changes will not occur during a switch to a protection channel. An attenuator is used to build out the loss between the IF main amplifier and the receiver output to maintain an output of −7 dBm.

In a repeater station bay, the beat oscillator signal for the receiver is supplied by a 40 MHz oscillator-shift modulator which shifts the microwave generator frequency by 40 MHz.[8] The shift modulator uses a waveguide hybrid junction and a pair of silicon diodes. The 40 MHz oscillator delivers +17 dBm to the diodes. Approximately +17 dBm of power at the microwave generator frequency also is applied to the diodes. Either the sum or difference frequency output signal, at a power of about +6 dBm, is selected by a band-pass filter for application to the receiver modulator. The filter reflects the undesired sideband into the modulator, thus improving modulator efficiency.

In a main station radio bay, the shift oscillator-modulator is not used because there are separate microwave generators for the microwave transmitter and receiver. Power, control, and alarm circuits are similar to those used in the transmitter.

3.2 *Thermal Noise*

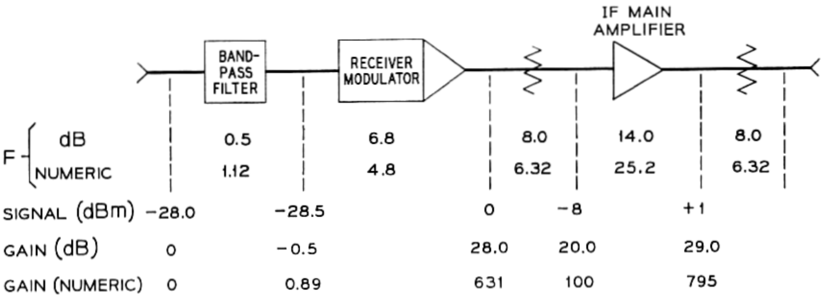
3.2.1 *Noise Figure of a Radio Receiver*

In the TD-3 microwave receiver, the input to the channel band pass filter is the reference point at which the received signal and the receiver noise figure are measured. Typical values indicating the noise contributions for the several parts of the receiver are shown in Fig. 3a. The nominal over-all noise figure of the receiver is 7.5 dB.

With the normal received carrier power and receiver noise figure known, the rms frequency deviation caused by receiver noise as a function of baseband frequency may be calculated. If the system signal deviation is also known, the thermal noise contributed by the receiver may be determined. The rms signal deviation per message band as a function of baseband frequency is shown in Fig. 4. Using the value for a message circuit at 5.77 MHz, and a 7.5 dB receiver noise figure, a top channel circuit noise of 13.5 dBrnC0 is obtained. This meets the receiver thermal noise objective imposed by fade margin, but is slightly larger (by 0.1 dB) than the noise allocated to the complete repeater (see Section 2.2.1). Also, so far only receiver signal path noise contributions have been considered. Thermal noise contributed by the microwave generator and the microwave transmitter also must be considered.

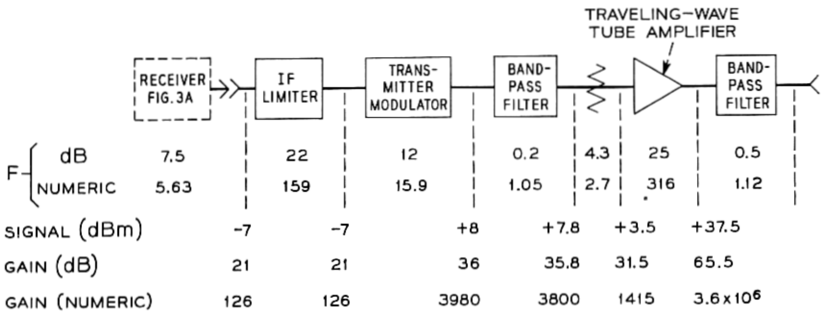
3.2.2 *Radio Repeater Noise*

Under normal received signal conditions, the microwave transmitter (excluding the microwave generator) makes a small but significant contribution to repeater noise. Figure 3b shows that the noise figure of the complete repeater is about 8.5 dB. Curve A of Figure 5 presents the computed thermal noise for a 4000 mile radio line with 140 repeaters, each with 8.5 dB noise figure and a −28 dBm received signal.



$$F_{RECEIVER} = 1.12 + \frac{4.8-1}{0.89} + \frac{6.32-1}{631} + \frac{25.2-1}{100} + \frac{6.32-1}{795}$$
$$= 1.12 + 4.25 + 0.008 + 0.242 + 0.007$$
$$= 5.63 \text{ OR } \sim 7.5\,\text{dB}$$

(a)



$$F_{REPEATER} = 5.63 + \frac{159-1}{126} + \frac{15.9-1}{126} + \frac{1.05-1}{3980} + \frac{2.7-1}{3800} + \frac{316-1}{1415} + \frac{1.12-1}{3.6\times10^6}$$
$$= 5.63 + 1.25 + 0.12 + 0 + 0 + 0.22 + 0$$
$$= 7.22 \text{ OR } \sim 8.5\,\text{dB}$$

(b)

Fig. 3 — Receiver (a) and repeater (b) noise figure.

3.2.3 *System Thermal Noise from Microwave Generators*

Noise sidebands surrounding the beat oscillator signal contribute to system noise. The sidebands appear as a result of noise within the local oscillator producing phase modulation of the output carrier. The magnitude of the modulation is directly proportional to the carrier-to-noise ratio. Since the output signal of a modulator used as a frequency shifter contains the phase modulation of both input signals, a given carrier-to-noise ratio at the beat oscillator input

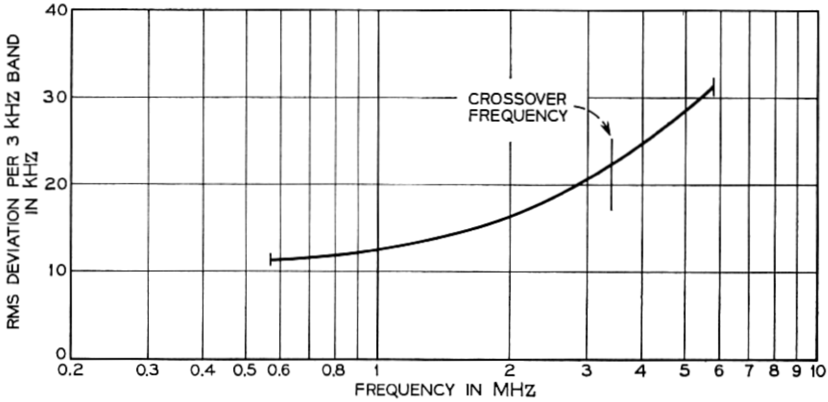


Fig. 4 — Frequency deviation characteristic for 1200 telephone channels. Total RMS deviation: 780 KHz, baseband range: 0.564 to 5.772 MHz.

will have the same effect as an equal carrier-to-noise ratio at the signal input to the receiver modulator.

Figure 6 shows the carrier-to-noise ratio (for a 1 Hz noise band) for a typical TD-3 microwave generator. If separate generators were used for receivers and for transmitters at all stations, the noise

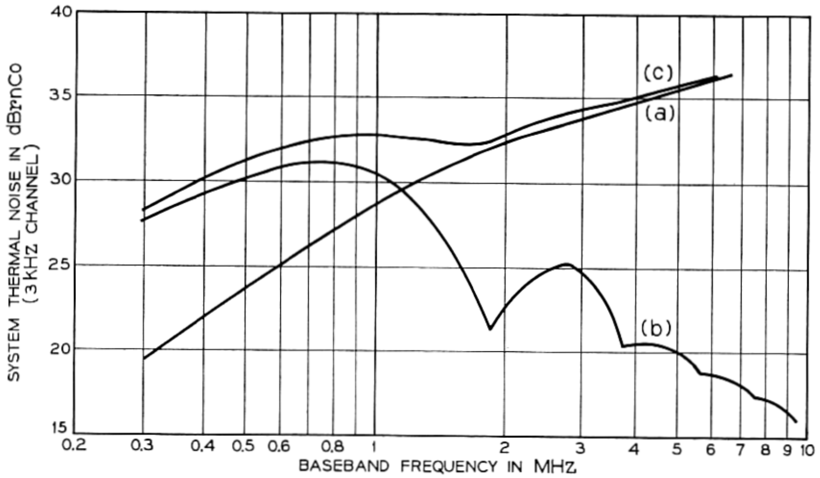


Fig. 5 — System thermal noise. (a) 140 repeaters, each with 8.5 dB noise figure and −28.0 dBm received signal level. (b) 182 generators, 98 at repeater stations and 84 at main stations. (c) Total of a and b.

contributed by the several generators could be added on a power basis to obtain the total system noise from this source. At repeater stations, however, a common generator is used for the receiver and transmitter modulators, and the generator noise introduced at the two modulators will add for some baseband frequencies, and will cancel for others.* The result is a periodic scalloping of generator noise. The scallop period is determined by the difference between the electrical lengths of two paths within the radio repeater. The first path is the direct one from the microwave generator to the transmitter modulator, and the second is from the generator to the transmitter modulator via the receiver modulator. The expression for calculating the noise addition at a repeater station is given in Appendix A.

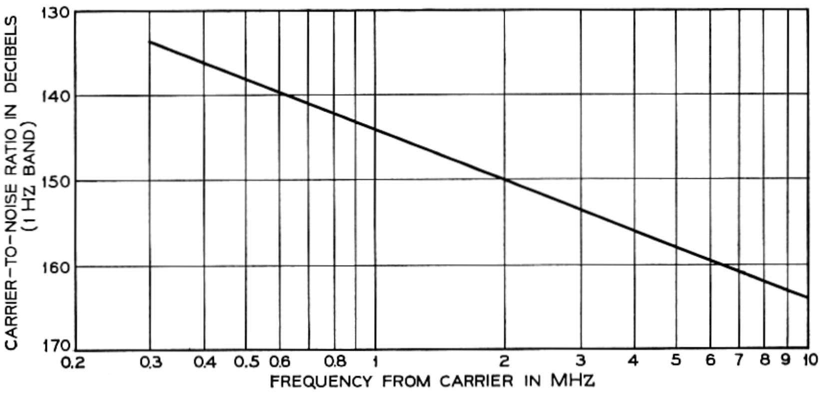


Fig. 6 — Typical carrier-to-noise performance of microwave generator.

Using the data of Fig. 6, and a TD-3 typical delay difference, D_3–D_1–D_2 (see Appendix A), of 555 nanoseconds, curve b of Fig. 5 shows the total noise contributed by the two generators at each of 42 main stations, and the single generators at the 98 repeater stations for a 4000-mile system. The gradual decrease of the envelope amplitude from low to high baseband frequencies results from the attenuation characteristic of the band pass filter at the output of the shift modulator used at repeater stations.

* The noise contribution of the shift modulator-oscillator is small and may be neglected.

3.2.4 *Total Thermal Noise of the Radio Line*

Curve c of Fig. 5 results from power addition of curves a and b, and represents the total thermal noise of the radio line model. Notice that generator noise is controlling at low baseband frequencies, and that the total thermal noise is less than 33 dBrnC0 for frequencies below 2 MHz. Above this frequency, noise from sources other than the generators generally dominates, and the total thermal noise gradually increases to 36 dBrnC0 in the top message channel.

The top message circuit thermal noise objective was established at 34.9 dBrnC0, as discussed in Ref. 1. Figure 5 shows that this objective will be missed by about 1 dB. To meet total noise objectives, therefore, the cross modulation noise must be kept below the 36.3 dBrnC0 assigned to it in Section 2.2.2.

3.3 *Selectivity Considerations*

3.3.1 *Introduction*

The TD-3 microwave transmitter and receiver provide required selectivity in passive networks. The active circuits, designed for broadband response, contribute a negligible amount to the total amplitude and delay distortion. The equalization compensates for the amplitude and delay distortion introduced by the passive selective networks.

The selectivity chosen was influenced greatly by the need to use the frequency plan and antenna arrangement in current use for the TD-2 radio system.[13, 14] This plan uses interspersed transmitter and receiver frequencies, with as little as 20 MHz spacing between carrier frequencies. The antenna side-to-side and back-to-back coupling losses, and the cross-polarization discrimination provide some isolation between the interfering and the disturbed radio channels. These losses, however, are usually insufficient to meet performance objectives, and selective networks must be provided in the transmitter and receiver. In addition to control of intrasystem interferences, selectivity must be provided to meet FCC regulations.

3.3.2 *Selectivity in the Microwave Transmitter*

Selectivity is required in the transmitter to suppress the unwanted sideband and beat oscillator signals at the output of the transmitting modulator, and to suppress carrier harmonics generated in the modulator and the TWT amplifier.

The selectivity required falls into two categories: that which must be placed between the transmitting modulator and the TWT amplifier, and that which must be placed after the TWT amplifier. If the TWT amplifier were perfectly linear, all of the selectivity could be located either ahead of or after it, or it could be arbitrarily divided. Since the amplifier is nonlinear, two considerations apply. First, enough selectivity must be placed ahead of the amplifier to keep sufficiently low the third order modulation products generated in the amplifier by the transmitter modulator products (beat oscillator and upper and lower sideband signals). Second, low selectivity must be used ahead of the amplifier to minimize in-band distortion. This reduces the amount of amplitude modulation added to the FM signal and thus restricts the AM to PM conversion occurring in the TWT amplifier to a tolerable amount.

The $2A-B$ product generated in the TWT amplifier by the beat oscillator and undesired sideband signals is an exact replica of the desired signal, both in spectrum (no frequency inversion) and index of modulation. However, this product generated in the tube will be delayed relative to the desired sideband; this will be a source of echo-type cross modulation noise. To keep this noise within tolerable limits, it is necessary to attenuate the beat oscillator and undesired sideband signals before they are applied to the traveling wave tube amplifier.

A derivation of the requirements for selectivity at this location is given in Appendix B, which is an example of the analytical methods used to derive selectivity requirements. The requirements in Appendix B were based on measured intermodulation coefficients of typical traveling wave tube amplifier models, and 25 dB suppression of the beat oscillator signal at the transmitter modulator output. In practice, the 25 dB suppression may not always be maintained, and some TWTs may have higher modulation coefficients. Therefore, the sideband selecting filter ahead of the traveling wave tube amplifier has losses (about 15 and 30 dB at 70 and 140 MHz, respectively, from the desired sideband) somewhat greater than those derived in Appendix B. The in-band distortion of this filter is sufficiently low so that cross modulation noise from the AM-PM conversion in the amplifier is negligible.

This filter does not attenuate sufficiently the beat oscillator and unwanted sideband signals to meet requirements at the transmitter output. Therefore, additional selectivity is provided by the channel band-pass filter and by the channel separation network. Actually,

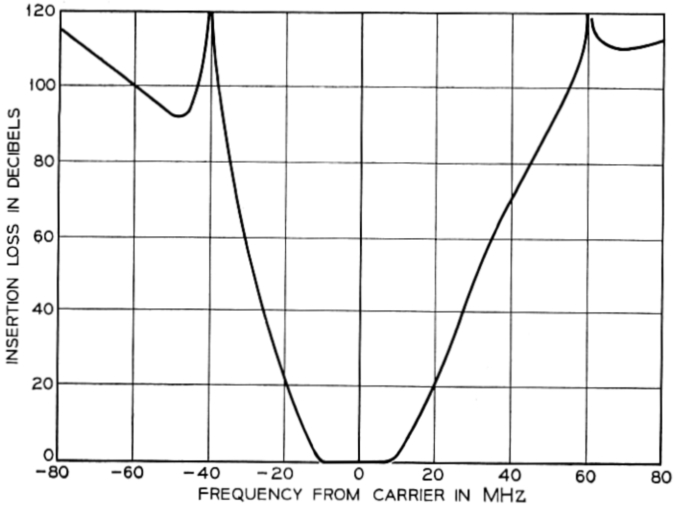


Fig. 7 — Microwave selectivity characteristic for microwave receiver. (This characteristic also applies to the microwave transmitter because the same network-filter combination is used.)

the channel band pass filter has somewhat more selectivity than needed, but to reduce the number of filter designs, a filter identical to the one used in the microwave receiver was adopted. The combined selectivity characteristic for the channel network and the channel filter is shown in Fig. 7.

The channel separation network and the channel band pass filter cannot be relied on to provide sufficient suppression of carrier harmonics generated in the TWT amplifier since these waveguide designs can propagate modes other than the dominant one at harmonic frequencies. Therefore, a coaxial low-pass filter is used. It provides more than 50 dB of loss at the second and third harmonic frequencies of the output signal.

3.3.3 *Microwave Receiver Selectivity Requirements*

Selectivity is needed in the receiver to:

(*i*) Sufficiently reduce the levels of interfering tones at the receiver output to permit the message circuit tone objectives to be met.

(*ii*) Prevent interference from signals at the receiver image frequency.

(*iii*) Prevent the receiver automatic gain control circuit operation from being affected by intrasystem interference.

(*iv*) Prevent the "turn-on" threshold of the carrier resupply circuit of the transmitter from being affected by interferences.

(*v*) Reduce the noise from direct adjacent channel interference to acceptable values.

As with the transmitter, selectivity can be placed at several points in the circuit. Considerable latitude is possible in choosing between microwave and IF selectivity. The plan followed was this: to provide only sufficient selectivity at microwave frequencies (ahead of the receiver modulator) to control third order modulation products generated in the receiver modulator, and to provide the remainder of the selectivity at the intermediate frequency.

In determining the selectivity required, it was assumed that system antenna side-to-side coupling losses of 80 dB or more, and antenna cross-polarization discriminations of 25 dB or more, would be realized. As we already explained, a parametric amplifier was originally used ahead of the receiver modulator. The microwave selectivity provided was based on the modulation performance of this combination. This resulted in more selectivity than that required without the parametric amplifier, but is of little importance because the total selectivity requirements remain about the same. If the microwave selectivity were to be reduced, additional IF selectivity would be needed.

3.3.3.1 *Interfering Tones.* The considerations governing interfering tones are treated in Ref. 1. The required carrier-to-interfering tone ratio, at the output of the receiver, is given by Fig. 6 of that article. For example, a single tone at a frequency corresponding to a baseband frequency of 5.77 MHz (75.77 or 64.23 MHz at the receiver output) must be 94 dB below the 70 MHz IF carrier at the receiver output. An IF tone of this power will produce a tone of −70 dBm0 in a message circuit located at a baseband frequency of 5.77 MHz.

Tones will result from third order modulation products formed in the receiver modulator by unwanted carriers spaced at 20 MHz intervals and transmitted to a receiver through antenna coupling losses, cross-polarization discrimination, and leaks. Products of the $A+B-C$ and $2A-B$ type produced by unwanted carriers at ± 20 and ± 40 MHz from the desired carrier are the most serious. If all carriers were precisely on assigned frequency, the modulation products would fall at the desired carrier frequency, and unless the products were unreasonably large, no system degradation would result. However, because of frequency errors in the system, the product may appear as a tone in the message portion of the baseband of the disturbed channel.

Carriers spaced at 80 MHz which are fed to the common waveguide connected to the microwave receivers can produce tones by a similar process. However, when sufficient microwave selectivity is provided to control the effects of the ± 20 and ± 40 MHz carriers, it will provide adequate suppression of the carriers at 80 MHz spacing. The selectivity provided is shown in Fig. 7.

Tones can be generated in the receiver modulator by the unwanted sideband signal of the shift modulator. A filter with 112 dB loss at the unwanted sideband frequency (80 MHz from desired output) is used at the shift modulator output to assure that such tones are within tolerable limits.

3.3.3.2 *Image Interference.* Signals at the receiver image frequency (140 MHz from the signal frequency and on the same side as the beat oscillator frequency) will produce a 70 MHz IF output signal. The selectivity provided to control third order modulation products is more than adequate to suppress interference at the image frequency.

Radiation of the beat oscillator frequency must be controlled to prevent interference within the system and with other systems. Such protection is provided by the networks just treated, and by the signal combining directional filter (Ref. 12) immediately ahead of the modulator.

3.3.3.3 *AGC Circuit Protection.* Both the IF main amplifier and the AGC detector circuits are broad band. Therefore, selectivity must be provided ahead of the AGC detector to prevent interfering tones from affecting its output. Tones at 50 and 90 MHz, caused by adjacent channel carriers, are the most troublesome. These tones may remain constant in amplitude while the desired signal is fading, and the AGC operation may be locked up for deep fades. The selectivity provided by the microwave networks is inadequate to prevent such lock-up. However, an IF band-pass filter must be used to control other effects of adjacent channel interference. By locating this filter ahead of the IF main amplifier, the AGC circuit is also protected. The filter used is equalized in amplitude and delay over a 16 MHz band centered at 70 MHz. Its attenuation characteristic is shown in Ref. 9.

3.3.3.4 *Carrier Resupply Operation.** Normally, the carrier resupply

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* This might have been included in Section 3.3.2, but because its performance is influenced primarily by the receiver selectivity, it is included here.

is set to turn on when the receiver input signal fades about 50 dB. A nonfading interfering carrier will therefore affect this unit before it affects the AGC circuit; hence, additional filtering is included in the carrier resupply itself.

3.3.3.5 *Adjacent Channel Interference.* The modulation on an adjacent channel can be a serious source of interference.[1] The usual worst cause of this interference is an adjacent channel (±20 MHz) transmitter coupling into the receiver of the disturbed channel via the antenna side-to-side coupling. The interference is most severe during fades of the disturbed channel, and it affects both the message circuit portion of the baseband and the 9 MHz noise slot used to determine the need for a switch to a protection channel. The amount and distribution of selectivity required to provide a specified degree of protection from this type of interference cannot be easily determined. It is fairly certain that the total receiver selectivity, and probably the transmitter selectivity, is important in controlling this type of interference. More work is planned to obtain understanding of the mechanisms involved. In any event, the field tests on the TD-3 system indicate that adequate protection is provided against adjacent channel interference.

3.4 *Equalization*

3.4.1 *Amplitude and Delay*

All active circuits are designed with broadband response, and IF selective networks are self-equalized. The remaining equalization needed is that necessary to compensate for the amplitude and delay distortion of the microwave networks in the transmitter and receiver. At the time TD-3 was developed, suitable microwave equalizers could not be designed. Therefore an IF equalizer, located in the microwave receiver, is used to compensate for the distortion of the microwave networks of a transmitter at one station, and for the distortion of the microwave networks of the associated receiver at the following station. Thus, the equalization is applied on each hop and a corrected signal is made available for the next hop.

To determine the shape needed for the basic equalizer, a large number of simulated radio hops were measured in the laboratory. The amplitude response of the hop was measured to an accuracy of about 0.01 dB, and the delay response to an accuracy of about 0.1 nanosecond. Measurements were made from the IF input of the microwave transmitter to the IF output of the associated microwave receiver. For the delay measurements, the accuracy available using swept frequency measurements was inadequate, and it was necessary to design and construct a precise point-by-point IF delay measuring set. With this set and a broadband upconverter and downconverter it was possible to measure individual microwave networks.

The measurements showed that one group of distortion patterns involve receivers and transmitters that use beat oscillator frequencies below the signal frequencies, and another group with frequencies above the signal frequencies. Another finding, and a very welcome one, was that within a group, the distortion characteristics to be equalized were virtually independent of channel frequency. This meant that just two types of IF basic equalizer were needed. Figure 8 shows the delay distortion characteristic for a hop.

Notice that the curve is not symmetrical about the carrier frequency. Therefore, an IF equalizer designed to compensate for this distortion will need to have this same shape for a channel with the beat oscillator frequency located below the signal frequency. It will need to have a reversed frequency shape for a channel that has the beat oscillator frequency located above the signal frequency.

Using the data collected in the measurement program, a sixth order least squares approximation fit was developed for each type of shape to be equalized. This same procedure was followed for both delay and amplitude shapes. The resulting delay shape to be equalized is shown in Fig. 8, the amplitude shape in Fig. 9. As described

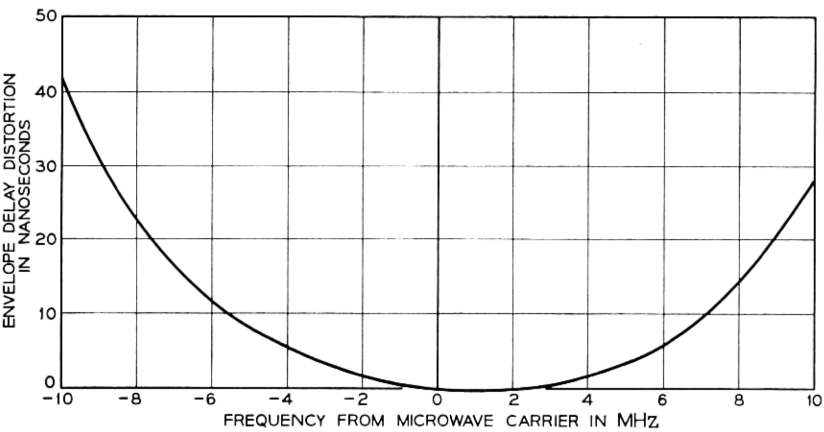

Fig. 8 — Envelope delay characteristic of unequalized radio hop. Tandem connection of two channel networks and two channel filters.

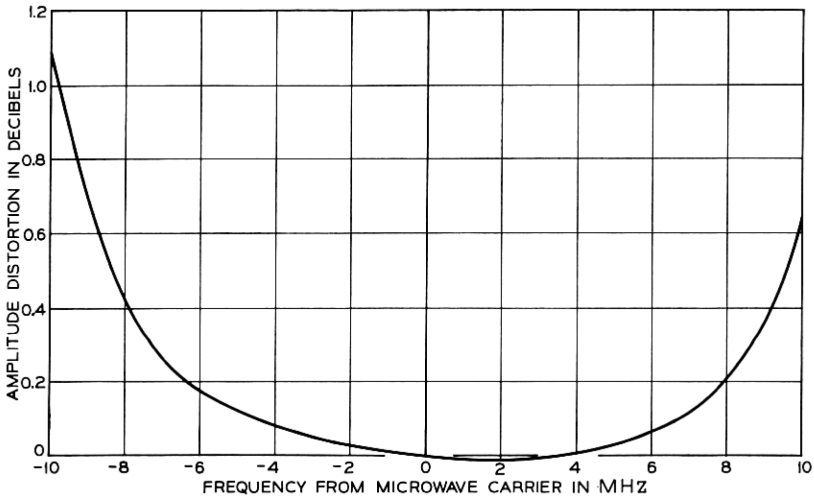

Fig. 9 — Amplitude characteristic of unequalized radio hop. Tandem connection of two channel networks and two channel filters.

in Ref. 9, IF equalizers were designed to compensate for these shapes. Adding this equalizer to the hop results in the delay characteristic of Fig. 10, and the amplitude characteristic of Fig. 11.

Over the signal first order sideband spectrum of 70 $\pm$ 6 MHz, the residual delay slope is approximately 0.025 nanoseconds per MHz compared with an objective of 0.1 nanoseconds per MHz. The residual

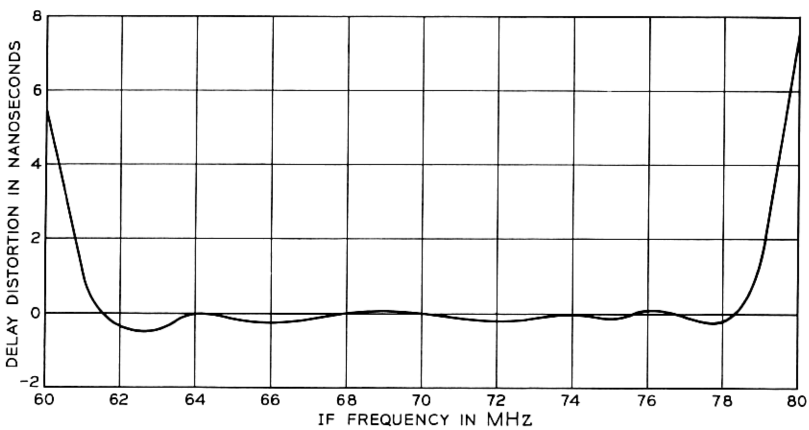

Fig. 10 — Envelope delay characteristic of equalized radio hop.

delay ripple is 0.2 nanoseconds peak with a period of about 4 MHz; the objective for a delay ripple of this period is 0.9 nanoseconds peak. Over the 70 ± 6 MHz band, the amplitude distortion is less than 0.03 dB. This meets the objective of 0.05 dB. These data illustrate the close fit between the distortion introduced by the microwave networks and the equalization provided.

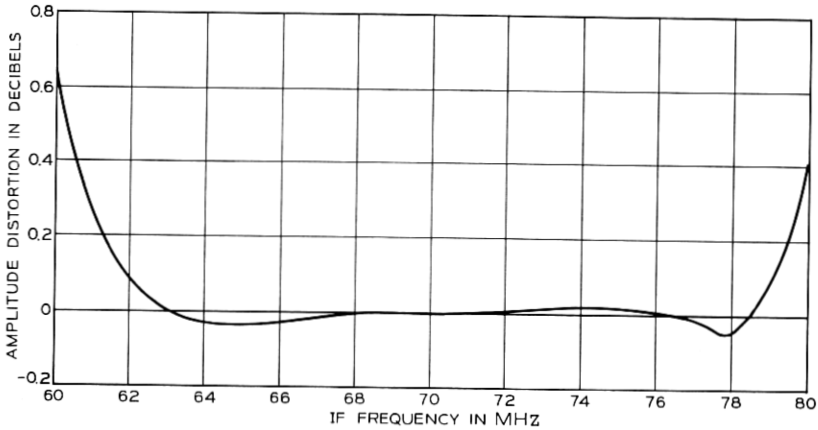

Fig. 11 — Amplitude characteristic of equalized radio hop.

3.4.2 *Mop-up*

Although the match between the characteristics of the microwave networks and the IF equalizer is excellent, some residual distortion will accumulate as a number of hops are connected; therefore, mop-up equalization is provided.

Laboratory and field measurements have shown that the residual amplitude distortion is negligible and that the residual delay distortion is predominantly slope contributed mainly by the adjacent channel network.* Therefore, a series of fixed delay slope equalizers are available to provide delay slope values of +0.5, +0.25, 0, −0.25, and −0.5 nanoseconds per MHz.

* In a TD-3 radio station, up to six bays may be connected in tandem, with the channel networks connected in ascending order by frequency, and the lowest frequency network closest to the antenna. For a full bay lineup the networks are spaced in frequency by 80 MHz. Thus the incoming signal to the first bay in the line-up passes through no previous network, the signal for the second bay passes through one previous network, and so on. Only the previous network (80 MHz lower in frequency) adds appreciable delay slope to the desired signal.

The type and number of slope equalizers needed is determined by measuring the delay distortion of a switching section. The slope correction then is distributed among the receivers of the switching section, with not more than one slope equalizer per receiver. This plan is less detrimental to baseband response than equalizing at one location.

3.4.3 *Effects of Temperature and Humidity*

It was recognized early in TD-3 development that changes in ambient temperature or humidity would shift the resonant frequencies of the cavities in the microwave networks and filters, but the IF basic equalizer, operating at a much lower frequency, would remain relatively stable. Therefore, as ambient conditions changed, mismatch would occur. To minimize this mismatch, the microwave networks considered in Section 3.3 are constructed of invar, and the radio stations are air conditioned.

As a result of incomplete control, however, some mismatch still will occur. The residual delay distortion from this mismatch will be essentially delay slope whose effect on system cross modulation noise performance can be estimated if some assumptions are made.

To help make such an estimate, a model of a 4000-mile route was hypothesized. The route extended from Boston to San Francisco via Atlanta and Dallas, and was comprised of 140 repeaters. It was assumed that the route was divided into 16 sections with FM terminals at the ends of each section, and with complete baseband circuit frogging at each terminal. Study of three years' weather records for four months in summer and early fall revealed that extremes of temperature and humidity were quite likely to occur simultaneously at all stations in a section of the model. It was logical to assume that cross modulation noise caused by misequalization from weather would add systematically for stations within a section. Because of circuit frogging, noise of the 16 sections might be expected to add randomly. Therefore, using the 33.3 dBrnC0 allocation from Section 2.2.2, the per-repeater noise allocation becomes:

$$N_{REP} = 33.3 - 10 \log_{10} 16 - 20 \log_{10} \frac{140}{16}$$

$$= +2.5 \text{ dBrnC0}$$

Figure 12 shows the center frequency shift versus temperature for invar networks. Figure 13 shows the shift versus relative humidity.

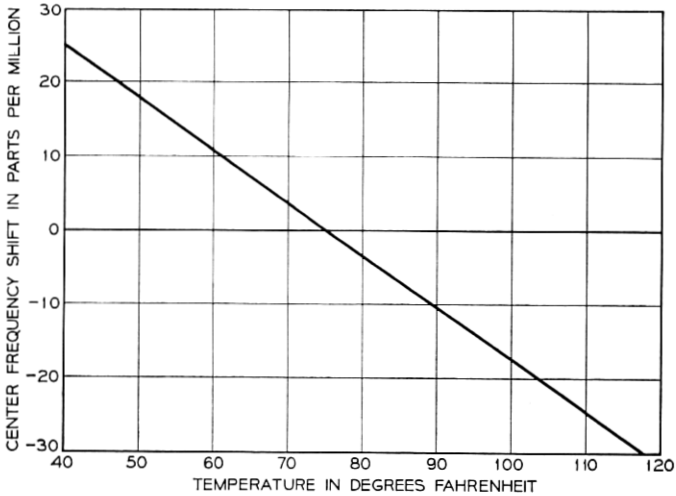

Fig. 12 — Frequency shift of tuned cavities versus ambient temperature with relative humidity at 0 per cent.

Values of frequency shift, determined from Figs. 12 or 13, may be used with Fig. 14 to determine the repeater cross-modulation noise resulting from temperature or humidity. For instance, if relative humidity is held at 0 per cent, the repeater cross modulation noise caused by the temperature changing from 75 to 120°F will be +3.5 dBrnC0.* Similarly, if a fixed temperature of 75°F is assumed, the repeater cross modulation noise caused by the relative humidity changing from 40 to 100 per cent will be in excess of 10 dBrnC0. Thus control of temperature and relative humidity is required to meet the cross modulation noise objective of +2.5 dBrnC0.

To reduce the cost of air-conditioning equipment, a decision was made to control the temperature within stations only to 75 ± 20°F. This temperature range requires the application of dry air (less than 5 per cent relative humidity) to the waveguide networks. The cross modulation noise caused by misalignment between the equalizer and the repeater then will meet the +2.5 dBrnC0 objective.

3.4.4 *Cross Modulation Noise*

No allowance was made for noise caused by transmission distortion followed by an AM-PM converting device when cross modulation

* Had the microwave networks been constructed of copper rather than invar, this would have been +23.5 dBrnC0.

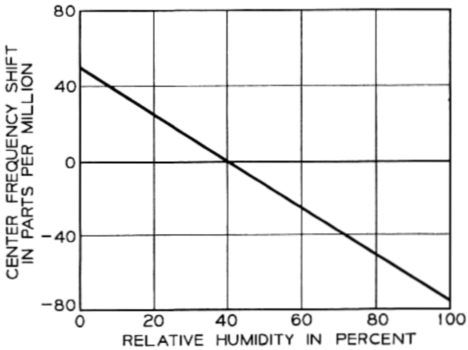

Fig. 13 — Frequency shift of tuned cavities versus relative humidity with temperature at 75°F. (Adjusted at 40 per cent relative humidity.)

noise was allocated. It was recognized that circuits would have to be designed to minimize the AM-PM conversion factor. An objective of 0.25 degree per dB maximum was established for this factor for each of the IF active circuits. The objective for the traveling wave tube amplifier was established at 4 degrees per dB maximum.

Tests on the initial installation indicated that excessive cross modu-

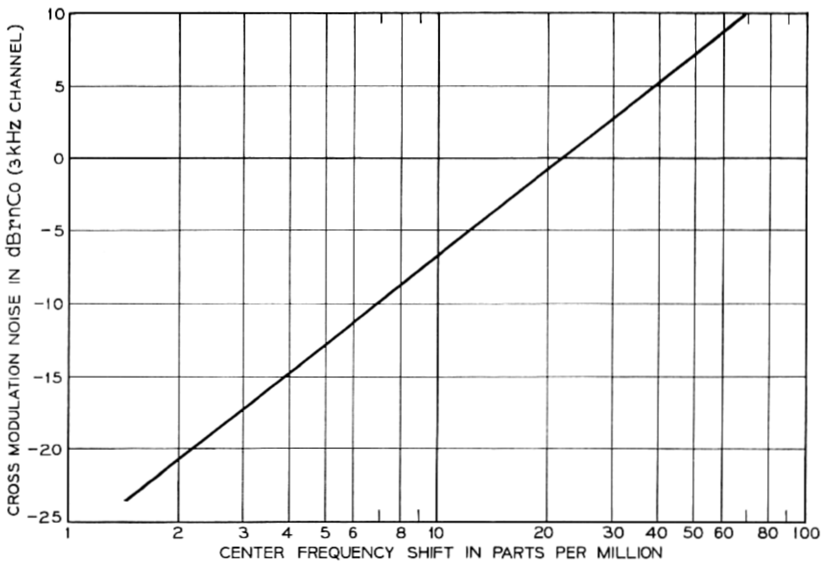

Fig. 14 — Cross modulation noise in one repeater caused by frequency shift of tuned cavities.

lation noise was occurring in the IF main amplifier and in the traveling wave tube amplifier. This noise was made negligible by relocating the basic equalizer ahead of the IF main amplifier, by moving the highly selective microwave channel band pass filter to the output side of the TWT amplifier, and by providing a relatively broad (low distortion) sideband-selecting filter ahead of the TWT.

Analytical techniques serve to confirm the validity of these choices.[6] For instance, using typical values of 0.4 and 3.5 degrees per dB for the IF amplifier* and the traveling wave tube amplifier, respectively, Table II gives maximum distortions which may be permitted ahead

TABLE II — DISTORTION AHEAD OF AN AM–PM CONVERTER†

Distortion type	Magnitude of distortion at ±6 MHz	
	K = 0.4°/dB	K = 3.5°/dB
	(nanoseconds)	
Linear delay	19.2	6.0
Parabolic delay	4.9	0.6
Cubic delay	16.2	1.9
	(decibels)	
Linear gain	3.8	1.2
Parabolic gain	0.8	0.1
Cubic gain	5.1	0.4
Quartic gain	0.5	0.04

† Distortion which will produce +1.8 dBrnC0 of cross modulation noise in the top message channel.

of either of these converting devices to produce cross modulation noise of 1.8 dBrnC0.

If the cross modulation noise from either of these sources is assumed to add on a power basis for 140 repeaters, the system noise would be 1.8 + 21.5 or 23.3 dBrnC0. This value is 13 dB lower than the total cross modulation noise allocation of 36.3 dBrnC0, and therefore would increase the system cross modulation noise by about 0.2 dB.

Figure 8 shows that the delay distortion of the microwave networks and filters of a hop is approximately 12 and 6 nanoseconds at 6 MHz removed from the carrier frequency. If the basic equalizer is located after the IF amplifier, this distortion will appear in the IF signal applied

* The objective of 0.25 degree per dB maximum for the IF amplifier was not met; a typical number is 0.4 degree per dB. However, this was a tradeoff with the noise figure requirement. Since the basic equalizer was repositioned ahead of the IF amplifier thereby reducing the noise caused by AM-PM conversion, it was judged more important to keep the noise figure low.

to the amplifier. The amplitude distortion of the microwave networks and filters is less than 0.2 dB at ± 6 MHz: this appears to be no problem.

If the highly selective channel filter was located ahead of the traveling wave tube, it would produce delay distortion of about 11 and 13 nanoseconds at ± 6 MHz removed from the carrier frequency. By moving the channel filter to the output of the traveling wave tube, and using a broad sideband selecting filter, the delay distortion preceding the tube is reduced to 0.3 nanoseconds at ± 6 MHz. This will contribute substantially less than +1.8 dBrnC0 of cross modulation noise. Similarly, the 0.2 dB amplitude distortion in the channel filter has been reduced to 0.03 dB by using the broad side-band-selecting filter ahead of the tube. As in the case of the IF amplifier, the delay distortion ahead of the AM-PM converter is more serious than the amplitude distortion.

IV. EQUIPMENT DESCRIPTION

At a repeater station, the microwave receiver and transmitter in a bay serve one direction of transmission only, and the IF output of the receiver is connected directly to the IF input of the associated transmitter. This combination is called a repeater station bay, and is illustrated in Fig. 15. The bay uses a single microwave generator and a single −19 volt regulator. The equipment described here is manufactured by the Western Electric Company for Bell System use only.

At a main station, the microwave transmitter and receiver serve opposite directions of transmission. The IF output of the receiver, and the IF input to the transmitter, are connected to IF switching, patching, and distribution circuits in the station. This main station bay uses one microwave generator and one regulator for the transmitter, and a second one of each for the receiver to provide independent operation for the two directions of transmission. This improves reliability and facilitates maintenance.

The transmitter, receiver, and associated control units are packaged on a 9-foot, 19-inch, unequal flange, duct-type bay framework about 22½ inches wide and 15½ inches deep.

A clear plastic cover (top of bay in Fig. 15) protects the precisely tuned channel networks from mechanical damage which would change electrical characteristics.

The bays can be mounted back-to-back or against a wall; all mounted equipment is accessible and removable from the front.

The lower portion of the bay contains one regulator and one micro-

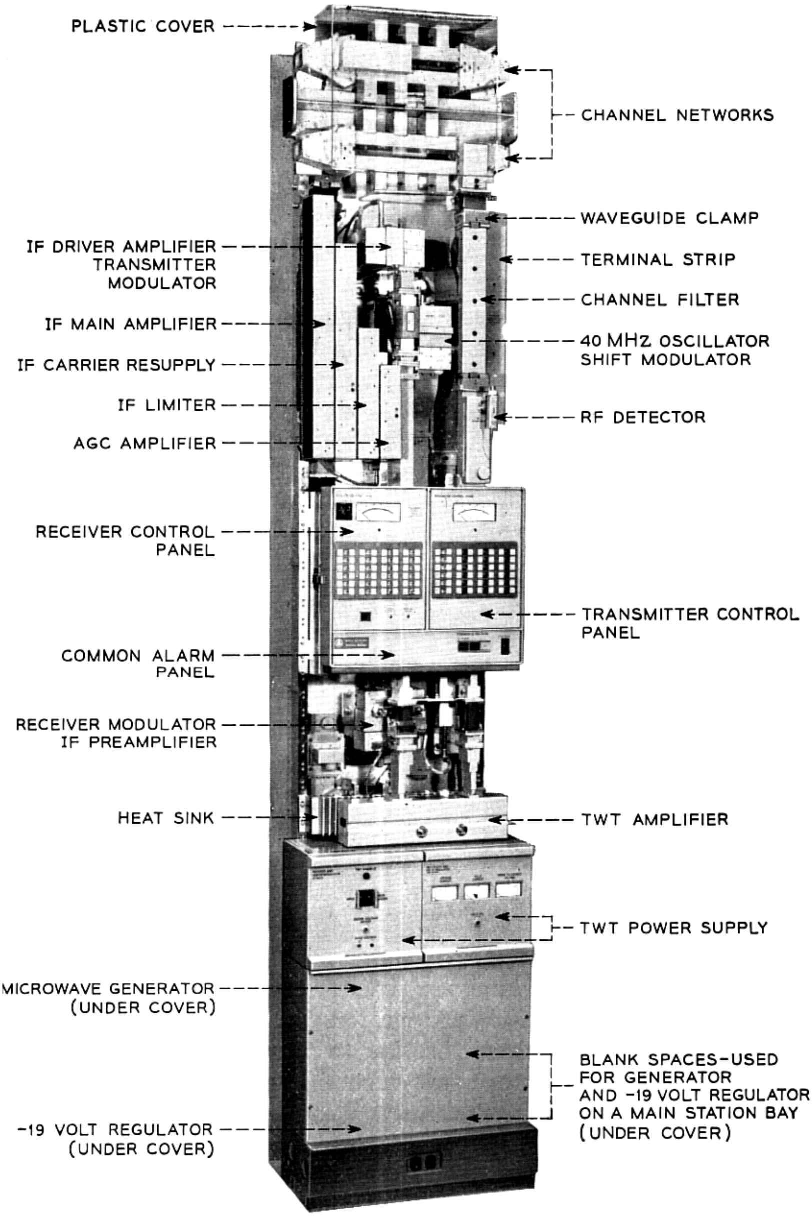

Fig. 15 — TD-3 transmitter-receiver bay for repeater station.

wave generator for a repeater station bay, and two regulators and two generators for a main station bay. These components plug in to permit easy removal. They are supported by nylon slides and protected by a front cover with quick release fasteners.

The traveling wave tube amplifier power supply mounts just above the upper microwave generator. It consists of two plug-in units: the low-voltage inverter unit on the left must be removed before the high-voltage rectifier unit on the right can be removed. Removing the low-voltage unit de-energizes the high-voltage unit making it safe for personnel. The power supply voltages are applied to the TWT amplifier by a cable that passes through an opening in the top of the power supply framework. The cable terminates in a plug that connects to a receptacle on the tube.

The TWT amplifier is on top of the power supply, with its power receptacle facing the supply. This arrangement encloses the power cord and its connector for greatest personnel safety. A dummy test load may be mounted in two slots on top of the supply framework without removing the TWT, and the power connection may be made to it. The dummy load is used to determine if the TWT amplifier or the power supply is faulty in case of transmitter output power failure.

The collector electrode of the TWT amplifier is connected to a cooling block on the TWT package. To achieve long life for the tube, it is necessary to restrict the cooling block temperature to 150°F maximum. To accomplish this, the TWT cooling block is connected to a heat sink made of cooling fins bolted to the bay uprights.

The control and alarm circuits are housed in a doorlike type framework mounted near the middle of the bay. The receiver control panel is at the left, the transmitter panel is at the right, and the common alarm panel is at the bottom. The door latches on the left side and swings open for access to apparatus behind the door, to read meter-relays, or to remove control units easily.

The receiver control units for repeater and main station bays are identical, and include a circuit breaker to prevent overloading of the −24 volt source, pushbuttons for connecting metered functions to a panel meter, a switch to select manual or automatic gain control, and a meter-relay to monitor the microwave generator output power. When the microwave generator output drops by a predetermined amount, the meter-relay provides an alarm. A pushbutton switch permits checking the voltage regulator output.

The transmitter control unit for a repeater station bay includes pushbuttons for connecting metered functions to a panel meter, and a meter-relay to monitor the transmitter output power and to provide a low output alarm. The transmitter control circuit for a main station is similar except that a circuit breaker and a pushbutton switch for checking the voltage regulator output is provided.

The common alarm unit receives transmitter and receiver alarms as grounds on the input leads and translates them into closed contacts on the output leads for operation of external audible and visible alarms. It also has an alarm cutoff key for disabling an external audible alarm, and an alarm reset switch for resetting the receiver and transmitter meter-relays.

The receiver, shift, and transmitter modulators are physically and electrically associated with the IF preamplifier, 40 MHz shift oscillator, and IF driver amplifier, respectively.

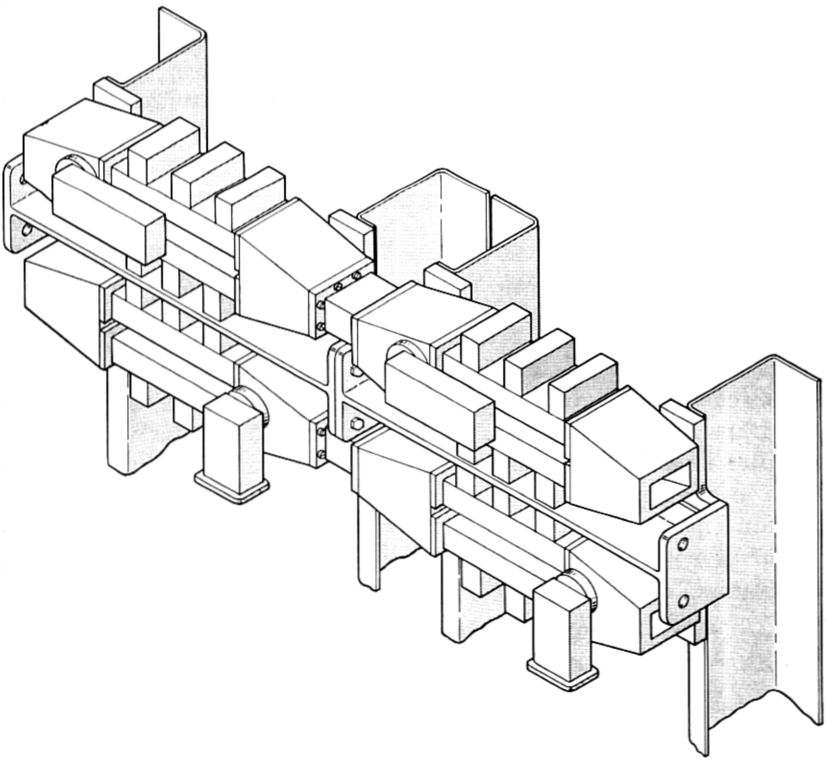

Fig. 16 — Channel separation network casting assembly.

The four active IF units are mounted on a doorlike framework at the left of the bay directly above the control door assembly. The units are from left to right: main amplifier, carrier resupply, limiter and AGC amplifier. The door is hinged on the left side and swings open to expose the connectors at the rear and the waveguide apparatus behind the door.

The passive IF units are mounted to the bay framework and are located so as to keep the interconnecting cable to a minimum. The units are: band-pass filter, low-pass filter, basic equalizer, and delay slope equalizer.

The channel separating networks are attached to a rugged casting at the top of the bay. The casting is designed to connect to castings of adjacent bays as illustrated by Fig. 16. This arrangement aligns adjacent network ports within a fraction of an inch to permit them to be connected by short sections of flexible waveguide; it eliminates mounting and positioning the networks of the bays separately in a line since the connection of the castings provides sufficient alignment.

Several sections of waveguide within the bay are removable to permit access for routine testing. These sections are connected by waveguide clamps to permit removal.

APPENDIX A

Baseband Noise

It can be shown that the output signal of a modulator used as a frequency shifter contains the phase modulation of both the input signal and beat oscillator signal. In a repeater station, one microwave generator serves as the beat oscillator for both the receiver and transmitter modulators. Therefore, generator noise will appear as phase modulation at the output of both of these modulators. The addition of these two noise outputs is dependent upon their phase relationships. This appendix derives the equation for calculating the baseband noise due to the addition of noise at the outputs of the two modulators.

let $\phi_m(t)$ = message modulation on signal

$\phi_n(t)$ = noise modulation on beat oscillator signal

D_1 = absolute delay in the RF path between the microwave generator and the receiver modulator

D_2 = absolute delay in the IF path between receiver and transmitter modulators

D_3 = absolute delay in the RF path between the microwave generator and the transmitter modulator.

Receiver Modulator
Input Signal $= \cos [\omega_1(t) + \phi_m(t)]$

Receiver Modulator
Beat Oscillator Signal $= \cos [\omega_2(t + D_1) + \phi_n(t + D_1)]$

Transmitter Modulator
Beat Oscillator Signal $= \cos [\omega_3(t + D_3) + \phi_n(t + D_3)]$

Then the receiver modulator output signal is:

$$K_1 \cos [(\omega_1 - \omega_2)t - \omega_2 D_1 + \phi_m(t) - \phi_n(t + D_1)]$$

The transmitter modulator input signal is:

$$K_2 \cos [(\omega_1 - \omega_2)(t + D_2) - \omega_2 D_1 + \phi_m(t + D_2) - \phi_n(t + D_1 + D_2)]$$

The transmitter modulator output signal is:

$$K_3 \cos [(\omega_1 - \omega_2)(t + D_2) - \omega_2 D_1 + \omega_3(t + D_3) + \phi_n(t + D_3) + \phi_m(t + D_2) - \phi_n(t + D_1 + D_2)$$

Collecting the phase modulation terms and shifting the time axis by $D_1 + D_2$:

$$\phi_n(t + D_3 - D_1 - D_2) - \phi_n(t) + \phi_m(t - D_1)$$

The output of the discriminator in the FM receiver is proportional to the time derivative of phase modulation terms at its input. Therefore $\phi_n(t)$ becomes $\phi_n'(t)$ at the discriminator output. The transfer function which operates on $\phi_n'(t)$ may be found as follows:

$$\phi_n'(t + D_3 - D_1 - D_2) - \phi_n'(t) = \phi_n'(t) \exp [-j(D_3 - D_1 - D_2)\omega] - 1$$

$$= -\phi_n'(t) \exp \frac{-j(D_3 - D_1 - D_2)\omega}{2} \cdot \left[\exp \frac{j(D_3 - D_1 - D_2)\omega}{2} - \exp \frac{-j(D_3 - D_1 - D_2)\omega}{2} \right]$$

where

ω = baseband frequency

The magnitude of the above function

$$= 2\phi_n'(t) \sin \frac{(D_3 - D_1 - D_2)\omega}{2}$$

or in dB,

$$= 6 + 20 \log_{10} \phi_n'(t) + 20 \log_{10} \sin \frac{(D_3 - D_1 - D_2)\omega}{2}$$

This transfer function may be used to calculate the addition of noise at the output of a receiver modulator and transmitter modulator at a repeater station. It is seen that the transfer function has a maximum value (in-phase addition) of

$$6 + 20 \log_{10} \phi_n'(t) \quad \text{when} \quad \frac{(D_3 - D_1 - D_2)\omega}{2} = \frac{\pi}{2}, \frac{3\pi}{2}, \frac{5\pi}{2}, \text{ etc.}$$

The transfer function has a minimum value of $-\infty$ (cancellation) when $(D_3 - D_1 - D_2)\omega/2 = 0, \pi, 2\pi$, etc.

It should be noted that $D_3 - D_1 - D_2 = D_3 - (D_1 + D_2)$.

Therefore the difference in delay between the two paths from the generator to the transmitter modulator determines the period of this function.

APPENDIX B

Selectivity

This appendix shows the derivation of selectivity requirements at the traveling wave tube amplifier input in the microwave transmitter. The signal applied to the input from the transmitter modulator has three components which are assumed to have the following powers:

(*i*) The desired sideband, +8.5 dBm.

(*ii*) The beat oscillator frequency, displaced 70 MHz with respect to the desired sideband, −5 dBm. (The transmitter modulator is assumed to provide 25 dB of suppression to the +20 dBm beat oscillator input.)

(*iii*) The undesired sideband, displaced 140 MHz with respect to the desired sideband (in same direction as the best oscillator frequency), +8.5 dBm.

The beat oscillator and undesired sideband signals will produce a $2A-B$ product which is an exact replica of the desired signal. However, because of delay differences, this product will be delayed relative to the desired sideband and will be a source of echo-type cross modulation noise.

To establish filter selectivity requirements, a delay difference of 35 nanoseconds between the desired and undesired sidebands, and the beat oscillator frequency has been assumed. Available contour curves showed that the ratio between desired sideband power and the echo power can be 24 dB lower at RF than the required signal-to-noise ratio at baseband.[4]

Now if the noise resulting from the $2A-B$ product is permitted to increase system intermodulation noise by 0.1 dB, the echo noise should be $36.3 - 16 = 20.3$ dBrnC0. If systematic addition is assumed for 140 repeaters, the per-repeater requirement for echo noise is $20.3 - 20 \log_{10} 140 = -22.7$ dBrnC0.

The signal-to-noise requirement at baseband may be derived as follows:

The signal at baseband can be derived using the system constants listed in Ref. 1; that is, a -10 dBm0 power for an average talker and a 25 per cent activity factor. This yields a talker power of -16 dBm0 in each of the 1200 voice channels. Reference 5 shows that a power of 0 dBm0 corresponds to $+88$ dBrnC0. Therefore, the signal at baseband is $-16 + 88 = +72$ dBrnC0.

The noise at baseband is -22.7 dBrnC0 as derived above.

Therefore, the signal-to-noise ratio at baseband becomes $+72 - (-22.7) = 94.7$ dB.

When the 24 dB factor from above is included, the ratio between desired sideband power and echo power at the tube input must be $94.7 - 24 \cong 71$ dB.

Thus, for a desired sideband power of $+8.5$ dBm, the $2A-B$ product power referred to the input to the tube must be less than -62.5 dBm. The intermodulation performance of the traveling wave tube amplifier can be described by the following equation:

$$P_{2A-B} = 2P_A + P_B + K_{2A-B} \text{ dBm.} \tag{1}$$

where all powers are referred to the tube input and where

P_{2A-B} = Power of $2A - B$ product in dBm
P_A = Power of beat oscillator frequency in dBm
P_B = Power of undesired sideband in dBm
K_{2A-B} = -22 dB = $2A - B$ intermodulation coefficient measured on traveling wave tube amplifier.

From above,

$$-62.5 = 2P_A + P_B - 22$$

$$\therefore 2P_A + P_B = -40.5 \text{ dBm.} \tag{2}$$

For no filtering between the upconverter and the traveling-wave-tube amplifier, the values for P_A and P_B are equal to -5 and $+8.5$ dBm, respectively, and obviously will not satisfy equation (2).

The problem at this point is to select filter loss values at 70 MHz and 140 MHz removed from the desired carrier in order to reduce P_A and P_B at the amplifier input so that the conditions of equation (2) may be satisfied.

By substituting the following values into equation (2),

$$P_A = (-5 - FL_{70})\ \text{dBm}$$

$$P_B = (+8.5 - FL_{140})\ \text{dBm}$$

where FL_{70} and FL_{140} represent the filter loss at 70 and 140 MHz, respectively, from the desired sideband,

$$2(-5 - FL_{70}) + (+8.5 - FL_{140}) = -40.5\ \text{dBm}.$$

$$\therefore\ 2FL_{70} + FL_{140} = 39\ \text{dB}. \tag{3}$$

The following tabulation shows the range of losses that may be provided to satisfy equation (3):

Loss in dB		dBm		
FL_{70}	FL_{140}	P_A	P_B	$2P_A + P_B$
5	29	−10	−20.5	−40.5
10	19	−15	−10.5	−40.5

The loss of the filter, therefore, should be in the range of 5 to 10 dB at 70 MHz from the desired carrier, and in the range of 29 to 19 dB at 140 MHz from the desired carrier.

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