3A FM Terminal Transmitter and Receiver

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FM terminals form an important subsystem of long-haul microwave radio systems as the link between the baseband signal and the 70 MHz FM signal. Designed primarily for use on the TD-3 system, 3A FM terminals are also compatible with the TD-2 system on which they are finding wide application. Development objectives included the use of solid-state circuitry throughout, emphasis on reliability, and performance consistent with TD-3 objectives with up to 16 terminal pairs in tandem in 4000 miles.

The FM transmitter uses two voltage controlled oscillators at 186 and 256 MHz which are frequency modulated by the baseband signal and heterodyned down to the 70 MHz IF. Transmitter frequency stability is ensured by controlling the oscillator's environmental temperature and stringent pre-aging of critical circuit elements. The FM receiver uses a balanced, parallel resonant type discriminator preceded by two limiters which ensure good AM suppression and wide dynamic range. The gain of a terminal pair is 16 dB with 124 ohm balanced baseband input and output impedances.

I. INTRODUCTION

As indicated in a companion paper, FM terminals perform the initial and final modulation steps in the TD-3 microwave radio system.[1] The FM transmitter converts the baseband signal to a frequency modulated signal centered at 70 MHz; the FM receiver performs the reverse function of recovering the baseband modulation from the FM signal. Although primarily intended for use in the TD-3 system, the 3A FM terminal equipment* also is used for improved performance in the TD-2 system.[2]

Design emphasis was placed on: (*i*) reliability and minimum maintenance through the use of solid-state circuitry throughout, (*ii*) im-

* Manufactured by the Western Electric Company for Bell System use only.

proved performance consistent with the more stringent TD-3 objectives, and (*iii*) reduced cost and size.

The types of baseband signals to be transmitted over the terminals are discussed by S. D. Hathaway and others.[1] In general terms they consist of 1200 message circuits comprising two multiplex mastergroups or a single NTSC* color television signal. FM terminals are required at each end of a radio route and at intermediate points where the baseband signal or some portion of it must be added or dropped. The 3A design is based on a maximum of 16 terminal pairs in tandem in 4000 miles.

II. DESIGN OBJECTIVES

With appropriate assumptions about the law of impairment addition, the system allocation to terminals can be converted to the performance objectives for a single terminal pair given in Table I.

These objectives apply with the appropriate pre-emphasis shapes which are chosen to optimize over-all system performance. Since terminal noise and distortion are not controlling, preemphasis shapes were not available as a design parameter for optimizing terminal performance. The pre-emphasis shapes chosen (see Fig. 1), were based on an extension of the shapes that were found to be optimum for the TD-2 system.

The message and television signals carried by TD-3 have an upper frequency limit of approximately 6 MHz. However, to minimize the influence of terminals on high-end frequency response and to allow for possible future applications, the upper frequency limit was set at 10 MHz. The lower frequency limit is set by the 60 Hz component of a television signal. To minimize phase distortion of the low frequency components of the TV signal, it is necessary to keep all low frequency cutoffs well below 60 Hz. This is satisfactorily controlled by maintaining the transmission response essentially flat down to 6 Hz.

Cross modulation objectives can be stated in terms of nonlinearity or harmonic performance.[3] These characterizations, however, are less appropriate than a dBrnC0 objective with simulated message loading since situations can arise where optimum noise loading and optimum linearity are not coincident. Nevertheless, since good linearity is in a broad sense a necessary condition for satisfactory performance with

* Based on standards of the National Television System Committee.

TABLE I — DESIGN OBJECTIVES FOR ONE FM TERMINAL PAIR

Baseband transmission Bandwidth (±0.1 dB point) Gain stability	 6 Hz to 10 MHz ±0.25 dB per six-month maintenance interval
Total noise First order nonlinearity (±4 MHz) Second order nonlinearity (±4 MHz)	21 dBrnc0 <1.3% <2.0%
Differential gain	Satisfactory differential gain is ensured by the telephone cross-modulation objectives
Differential phase	±0.3°
Television weighted high frequency signal-to-noise ratio	>73 dB
Center frequency of FM transmitter Peak frequency deviation Deviation sense	70 MHz ±100 kHz 4 MHz Positive going signal on the tip of the input plug produces a decrease in transmitter output frequency
Transmitter longitudinal suppression	>53 dB up to 1 MHz
Change in receiver demodulation sensitivity for a 10 dB reduction in IF input	<0.25 dB
Microphonics	Negligible
Change in transmitter carrier frequency between zero and full modulation	<20 kHz
Operating temperature range	0 to 50°C

a live message load, it is useful to have a linearity objective for design purposes.

The requirement relating to carrier frequency stability as a function of modulation level is associated with the desire to use the carrier null or Crosby technique to set transmitter deviation accurately.[4] Since the detection of a carrier null in the presence of adjacent sidebands requires a narrow band receiver, it is essential that the application or removal of modulation does not shift the carrier outside the receiver passband.

III. FM TRANSMITTER

The FM transmitter provides a +10 dBm output signal centered at 70 MHz with frequency modulation linearly related to the baseband

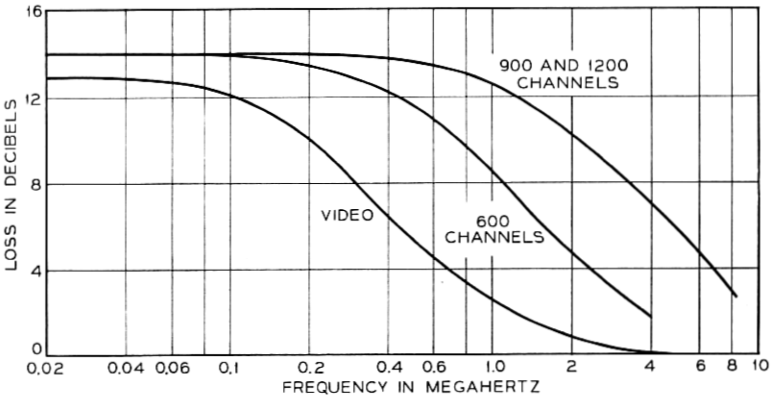

Fig. 1 — Pre-emphasis characteristics.

(input) signal. A −12 dBV input signal produces 8 MHz peak-to-peak deviation.

The development of the transmitter involved the examination of various alternative solid-state modulators. These fall into two broad categories: (*i*) direct modulators which operate and are deviated at the 70 MHz intermediate frequency; (*ii*) heterodyne type modulators in which the outputs of two higher frequency voltage-controlled oscillators are combined in a mixer to generate a difference frequency of 70 MHz. For reasons related to the status of device development, reproducibility, and high frequency baseband response, the heterodyne approach was used in the 3A development.

3.1 *General Description*

The broad features of the FM terminal transmitter are shown in the block diagram of Fig. 2.

The modulating signal is applied to the 124 ohm balanced input of the baseband amplifier. Two outputs from this amplifier, in antiphase, drive two voltage controlled oscillators with frequencies centered at 186.67 and 256.67 MHz. The deviated outputs from these oscillators are applied through buffer amplifiers to a mixer circuit where the 70 MHz difference frequency is generated. A low pass filter in the mixer output rejects unwanted products and passes the desired intermediate frequency to the succeeding filter-equalizer unit. The filter portion of this unit provides additional suppression to out-of-band products generated by the mixer. The equalizer provides amplitude equalization for the low pass filter as well as delay equali-

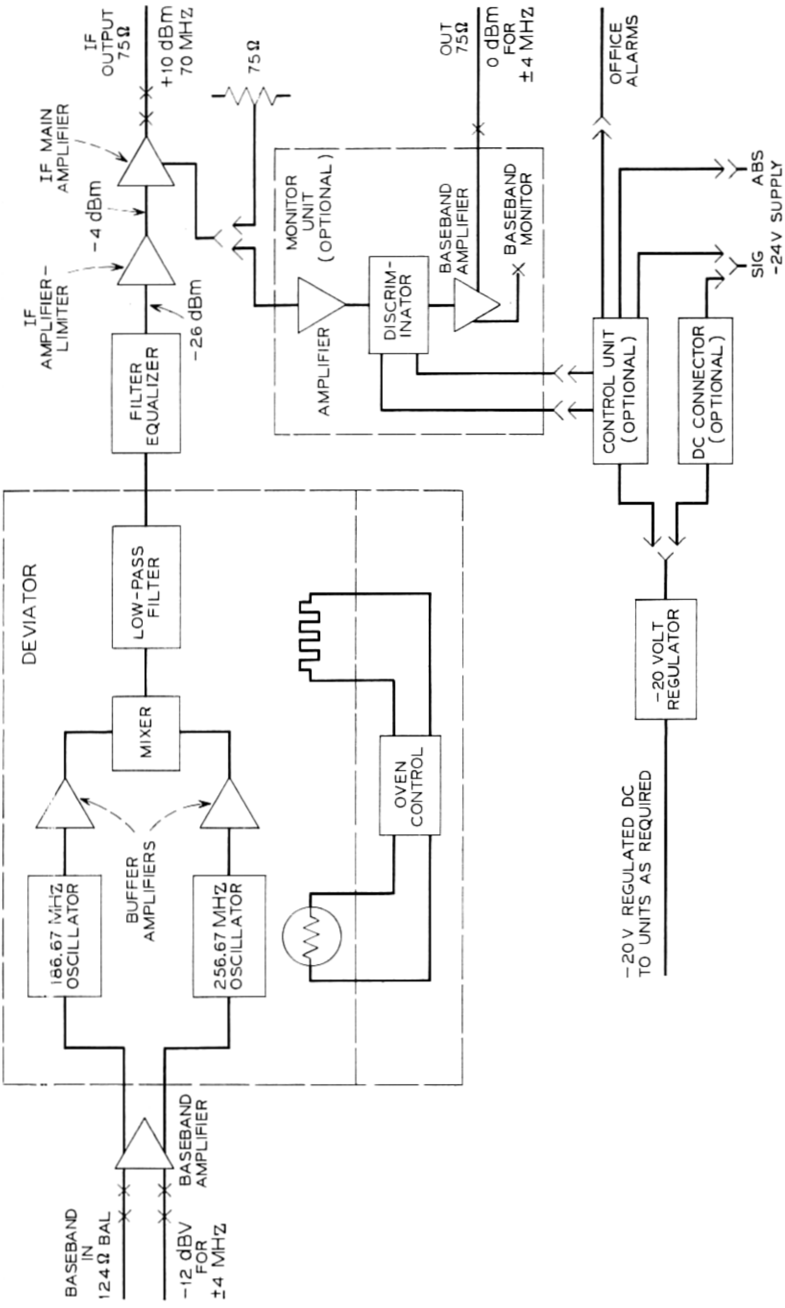

Fig. 2 — Block diagram of FM transmitter.

zation for the complete transmitter. The signal then passes to an amplifier-limiter and finally to an output IF amplifier which delivers +10 dBm.

Part of the IF output from the main amplifier can be connected to a monitor circuit which is essentially an FM receiver without limiting. The discriminator output from this circuit is used to provide alarm indications when the transmitter IF output power or average frequency departs by a prescribed amount from nominal. The alarm outputs are not provided directly from the monitor but through alarm relays in the control unit. Both the monitor unit and control unit are optional and are generally not used when the terminal is included within an automatic terminal protection switching system.

The frequency of the transmitter is maintained within 100 kHz of nominal by closely regulating the temperature of the entire deviator assembly. This is accompanied by special pre-aging of critical components within the deviator to ensure that the stability objectives are met for extended periods. The use of this form of frequency stabilization as opposed to automatic frequency control has the merit of placing no limitation on the low-end frequency response of the transmitter.

Figure 3 is a photograph of the FM transmitter.

3.2 *Baseband Amplifier*

The baseband amplifier provides a nominal 19.5 dB of voltage gain flat to ±0.05 dB in the frequency band 6 Hz to 10 MHz between the balanced 124 ohm input and the balanced varactor diode loads presented by the deviator. It must also suppress low frequency common mode tones by more than 53 dB.

Figure 4 is a simplified schematic diagram of the amplifier. The amplifier is balanced in all stages and operates in push-pull class A manner with identical stages for "tip" and "ring" sides of the signal.

The input stages Q_1 and Q_{101} provide no signal gain, but together with transistor Q_7 provide more than 65 dB of voltage loss to common mode tones applied to the input. In addition, transistor Q_7 provides a canceling voltage for low frequency power supply noise suppression. This is accomplished by applying the power supply noise to the emitter of Q_7 through resistor R_{16} which is adjusted to provide a noise voltage at the bases of Q_2 and Q_{102} equal to the power supply noise present at their emitters. In this manner the noise volt-

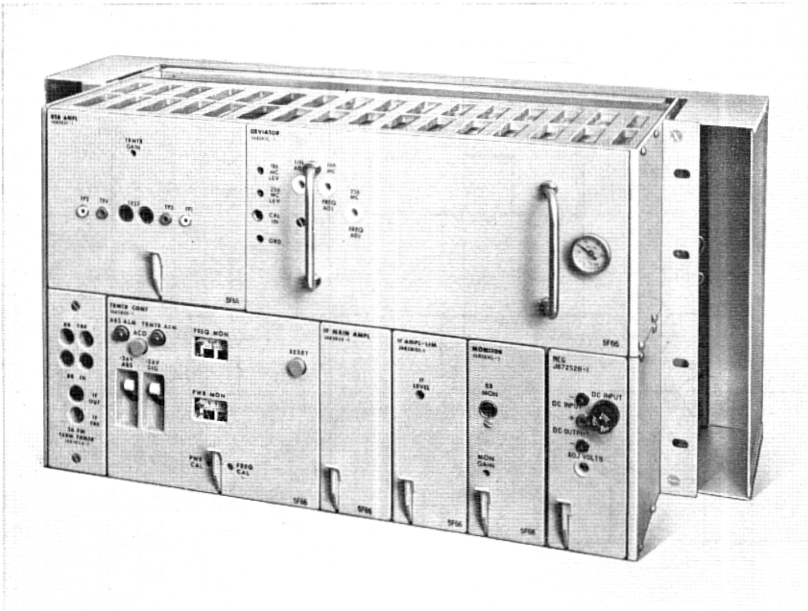

Fig. 3 — 3A FM transmitter.

age at either side of the amplifier output is reduced by more than 20 dB.

The amplifier gain is provided by two 2-stage series input, shunt output balanced feedback amplifiers. Bias stability is obtained by dc feedback. The high input impedance of the series input feedback circuit allows frequency response down to 0.6 Hz with coupling capacitors of reasonable size (500 μf) between stages and without the use of low frequency loop gain compensation.

Gain adjustment is accomplished in the first balanced feedback amplifier by a single potentiometer common to the "tip" and "ring" sides. The gain potentiometer is connected to equal dc voltage points in the two amplifiers which eliminates bothersome transients during gain adjustment.

The voltage gain of the first balanced feedback amplifier is adjustable from 4 to 12 dB and the gain of the second amplifier is fixed at 11 dB. As shown in Fig. 5, the output amplifier has 38° of phase margin at 50 MHz with no phase crossover below 500 MHz. The use of ferrite beads in place of resistors in the phase shaping network C_5 and CM_3 provides smooth loop transmission where the inductance

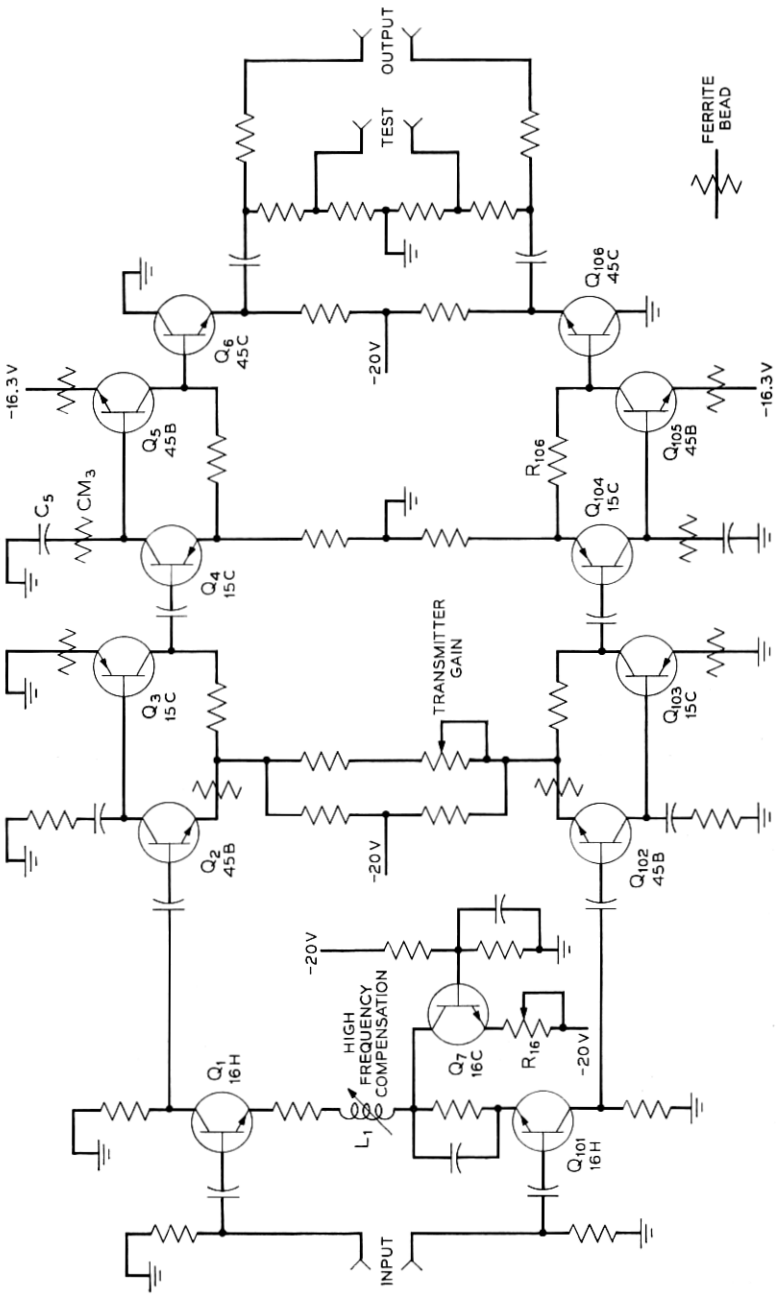

Fig. 4 — Simplified schematic diagram of transmitter baseband amplifier.

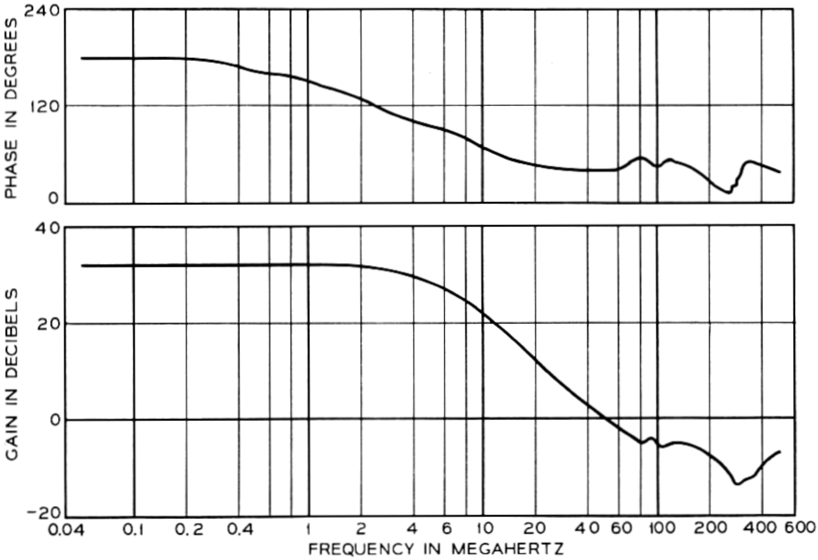

Fig. 5 — Output amplifier open-loop gain and phase.

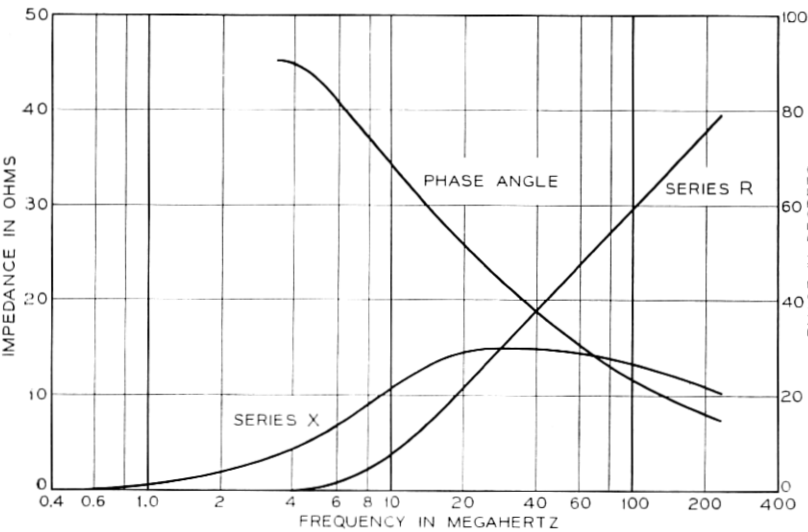

Fig. 6 — Ferrite bead impedance characteristics.

of an equivalent resistor would have produced a resonant ripple at 50 MHz.

Figure 6 shows the resistance and inductive reactance of these beads versus frequency. The phase angle rapidly decreases from about 90° at 5 MHz to 34° at 50 MHz. The increase in resistance with frequency and the decreasing phase angle give these devices ideal characteristics for reducing loop gain at high frequencies with minimum phase shifts. These beads are also used for emitter feedback in the second stage of the feedback amplifiers.

Emitter followers are used as output stages to isolate the feedback loop from the load impedance. The emitter followers work into the high impedance of varactor diodes through an equalizing network for flat baseband transmission.

The high frequency response of the over-all amplifier is adjusted by the common feedback inductor L_1 to achieve a response flat to ± 0.05 dB from 6 Hz to 10 MHz. Figure 7 shows the frequency response of a baseband amplifier. The 3 dB down points are 0.6 Hz and 45 MHz.

The nonlinearity of this amplifier is less than 0.15 percent, which adds virtually no degradation to the terminal.

3.3 *Deviator*

Frequency modulation in the FM transmitter is performed by the deviator. A heterodyne deviator was chosen since it requires higher oscillator frequencies which: (*i*) simplify the separation of oscillator

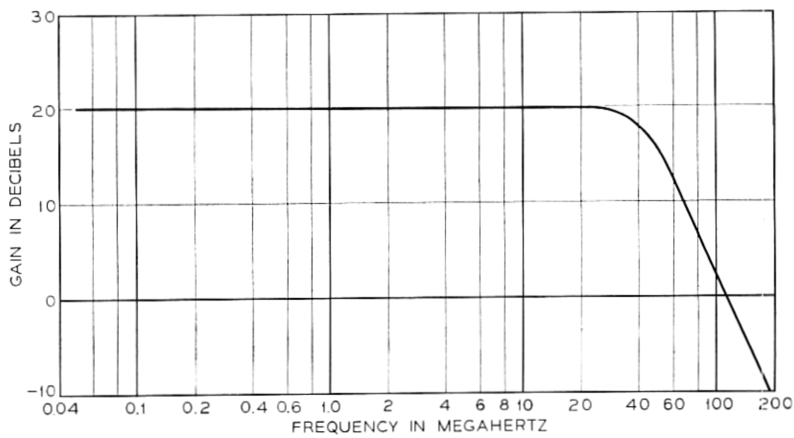

Fig. 7 — Baseband amplifier frequency response.

and baseband signals, (*ii*) have an inherently higher modulating frequency capability, and (*iii*) have lower nonlinearity because of their smaller percentage deviation. The principal disadvantage of the heterodyne method is the generation of undesired products in the mixer which in practice necessitates lower deviator output powers.

The choice of oscillator frequencies for the heterodyne deviator should, on the one hand, be as high as possible to minimize percentage deviation and hence improve linearity, and on other hand, should be kept as low as possible to avoid device limitations and reduce the importance of strays. In the 3A deviator, the percentage deviation on each oscillator is halved by deviating both oscillators in antiphase. As discussed in Appendix A, this push-pull type situation leads to the cancellation of first-order nonlinearities so that the first significant nonlinearity is the smaller second-order term.

The results of Appendix A indicate that to keep this second-order term below two per cent, using graded junction varactor diodes as the voltage sensitive tuning elements, the lowest heterodyne oscillator frequency should lie above 150 MHz. An examination of spurious mixer products as a function of oscillator frequencies shows that the nearest satisfactory location above 150 MHz is 186 MHz with the high frequency oscillator 70 MHz above this at 256 MHz. Designating these frequencies as V and C, respectively, the lowest order mixer product falling within the 60 to 80 MHz band is the 10V-7C term. This product will be least interfering if it falls exactly at 70 MHz leading to the final choice of rest frequencies at 186.667 MHz and 256.667 MHz. However, in arriving at requirements for the 10V-7C product level, no advantage was taken of its optimum location at 70 MHz on the grounds that it would not be accurately enough controlled in frequency to assure its continued location at the chosen design frequency.

To insure IF frequency stability, temperature control was chosen in preference to automatic frequency control. This was based on economic advantage, the elimination of low frequency response limitations because of an automatic frequency control loop, and the general advantages which accrue from temperature stabilization of the semiconductor circuit environment. However, the lack of an automatic frequency control circuit imposes stringent requirements on oscillator frequency stability.

The voltage-controlled oscillators are Colpitts-type circuits such as in Fig. 8. The varactor diodes which have the voltage capacitance

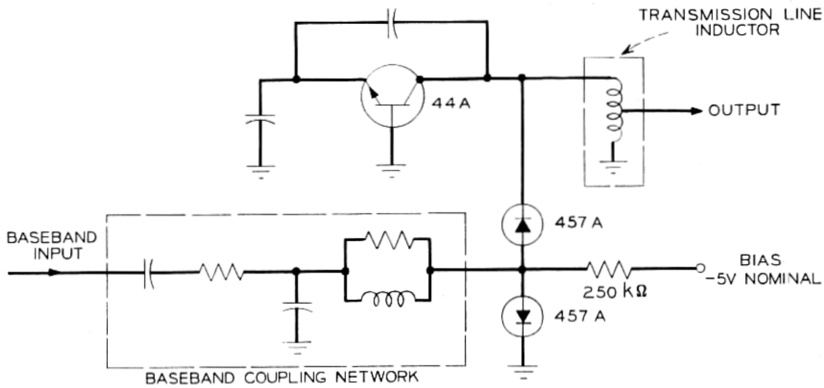

Fig. 8 — Simplified voltage controlled oscillator circuit.

law illustrated in Fig. 9 are connected in opposite directions across the oscillator tank circuit with the modulating signal applied at the center point. This reduces modulation of the tank circuit capacitance by the oscillator output signal. An R-L-C coupling network, terminated by the two varactor diodes in parallel, ensures that the modulating voltage applied to the diodes is virtually independent of frequency over the entire baseband range.

The capacitance of the two varactor diodes in series at the nominal −5 volts bias is 8 pF. This value was chosen as a compromise between providing a reasonably high ratio of diode capacitance to fixed capacitance and at the same time ensuring that the tank circuit inductance did not become unduly small in relation to circuit strays.

The tank circuit inductance consists of a 100-ohm short-circuited coaxial transmission line of approximately 27 degrees electrical length.

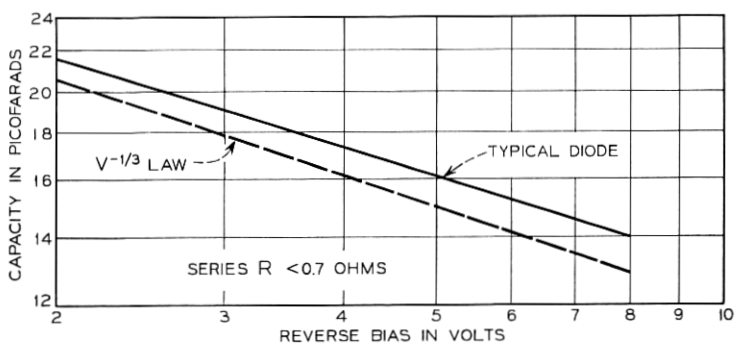

Fig. 9 — Typical capacitance characteristics of 457A varactor diode.

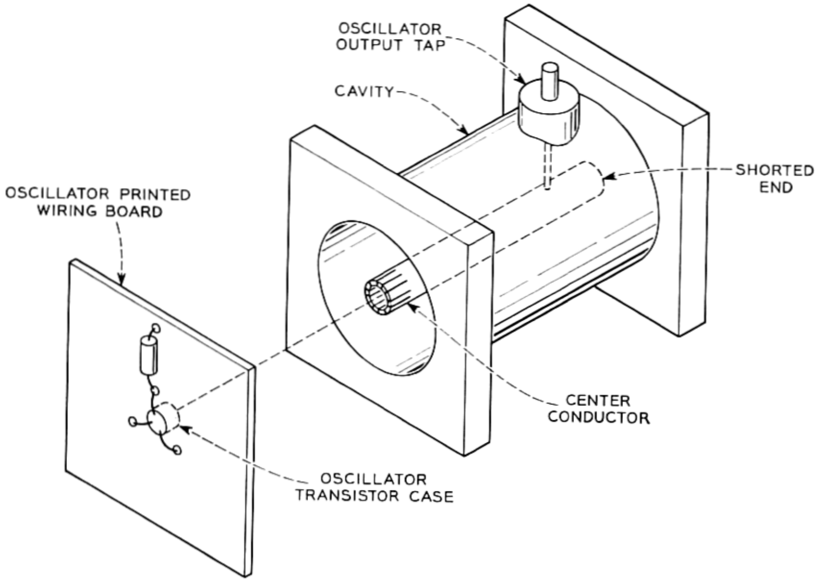

Fig. 10 — Transmission line inductor assembly.

As illustrated in Fig. 10, a chuck on the open end of the transmission line center conductor accepts the transistor case which is connected internally to the collector. This gives an intimate and well-controlled connection between the collector and tank circuit inductance.

The transmission line inductor is very stable, highly reproducible through control of the mechanical dimensions, and quite insensitive to mechanical shock. The center conductor also provides a good heat sink for thermal stabilization of the oscillator transistor. Output power is derived at a convenient 50-ohm impedance level by tapping the center conductor near the short-circuited end of the transmission line. To eliminate hysteresis effects and frequency discontinuities during warmup, the same material is used on inner and outer conductors. One disadvantage of the transmission line inductor is that its first-order equivalent circuit is an inductor in parallel with a small capacitor which increases the total effective stray capacitance.

The approximate effective equivalent circuit of the 186 MHz oscillator tank circuit is shown in Fig. 11. The ratio of stray to total tank circuit capacitance is 0.5 and, using the results of the Appendix, the computed second-order deviator nonlinearity is 1.8 percent for ± 4 MHz deviation. This is in good agreement with typical measured performance.

The 256 MHz oscillator is similar to its lower frequency counter-

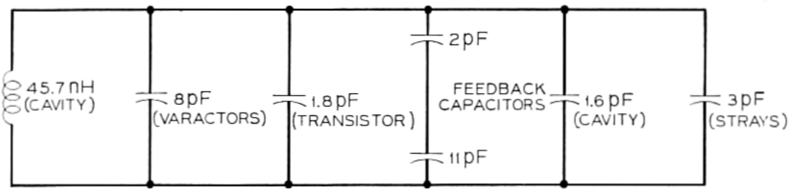

Fig. 11 — Equivalent tank circuit of 186 MHz oscillator.

part except for an additional trimmer capacitor across the tank circuit. Assuming equal deviation sensitivities, one would expect the 256 MHz oscillator to be more linear than the 186 MHz unit as a result of the smaller percentage deviation. Since it is intended that first-order nonlinearities in the two oscillators should cancel, the linearity of the 256 MHz oscillator must be diminished by providing additional "strays" in the form of the trimmer.

This trimmer is used to optimize linearity by adjusting for zero first-order deviator nonlinearity. The need for the trimmer can also be demonstrated from equation 15 of the Appendix. For $n = 1/3$, this equation requires that the high frequency oscillator stray capacitance must be greater than the stray capacitance of the low frequency oscillator for zero first-order nonlinearity.

To ensure a stable bias voltage for the oscillator varactor diodes, the −20 volt output from the dc regulator is first reduced to −15 volts by a voltage regulating diode circuit. A second voltage regulating circuit using a 446AC diode further reduces this to −8.2 volts which is applied to two adjustable voltage divider networks which supply the bias for the individual oscillators. Since the −8 volt bias supply is common to the two oscillators, frequency changes resulting from residual voltage variations tend to cancel.

Special care has been used in the detailed design of the oscillators to ensure good frequency stability. The principal factors influencing long-term frequency stability are breakdown-voltage aging on the 446AC diode, capacitance aging on the 457A varactor diodes, and leakage current changes on the varactor diodes. Device aging objectives were determined starting from an intermediate frequency stability objective of 100 kHz over a six-month maintenance interval. Of this total, 50 kHz was allocated to semiconductor aging and ambient temperature effects while the remaining 50 kHz was allocated to all other causes of frequency drift. The 50 kHz allowance for semiconductor and temperature effects was further divided in the manner shown by Fig. 12 to provide objectives for individual devices.

To achieve these aging objectives, the 446AC diode and 457A varactor diode are subjected to stringent pre-aging as described by S. M. Forst and others.[5]

The major means for controlling frequency is the proportionally controlled oven which maintains the environmental temperature of the oscillator, buffer-amplifier, and mixer assembly within 1° of 50°C. The control characteristic for the oven is shown in Fig. 13. The loss of control at 45°C results from power dissipation in the circuitry within the oven. Since the design operating temperature range for the terminal is 0 to 50°C, the loss of oven control at 45°C results in a degradation in frequency stability over the top 5° of the range. This degradation is small, however, since the individual oscillators have temperature coefficients from 15 to 18 kHz per degree Centigrade with a net temperature coefficient of approximately 3 kHz per degree Centigrade for the intermediate frequency output.

Field measurements indicate that the long-term frequency stability objectives are being met. Figure 14 shows a typical frequency-time record for a seven-month monitoring period.

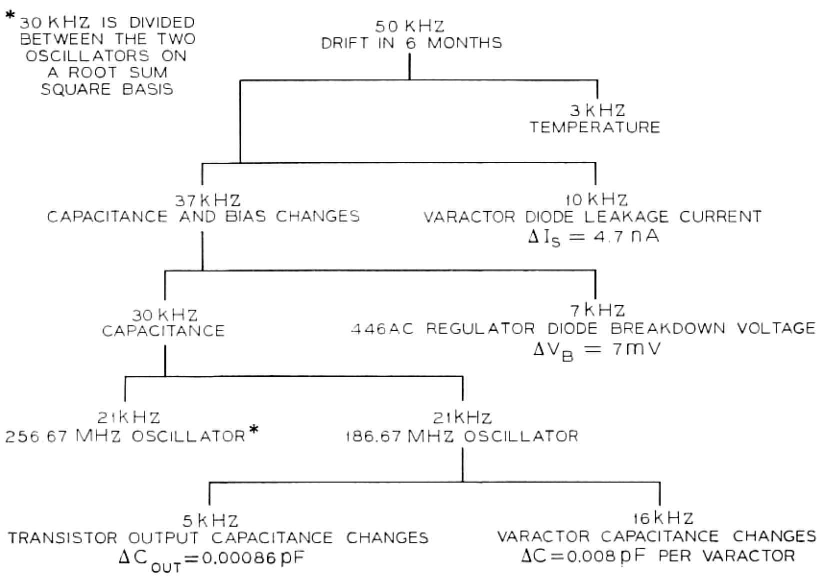

Fig. 12 — Frequency tolerance allocations.

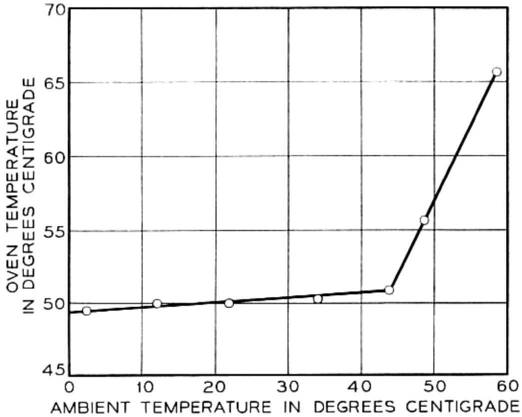

Fig. 13 — Oven temperature control characteristic.

Diode detector circuits are bridged across the oscillator outputs to provide a dc indication of oscillator output level. In the case of the 256 MHz oscillator, provision is also made for applying an external signal in parallel with the oscillator output and its detector circuit. For alignment, this external signal is the output of a crystal-controlled 256.667 MHz oscillator to which the frequency of the internal 256 MHz oscillator can be locked.

The lock-in is indicated by a dip in the dc output of the detector diode. Under lock-in conditions, the 186 MHz oscillator can be adjusted to obtain a precise 70 MHz output from the deviator. Thereafter, with the external reference removed, the internal 256 MHz oscillator can be accurately set by re-establishing a precise 70 MHz intermediate frequency. This procedure is used so that only one reference frequency oscillator need be provided.

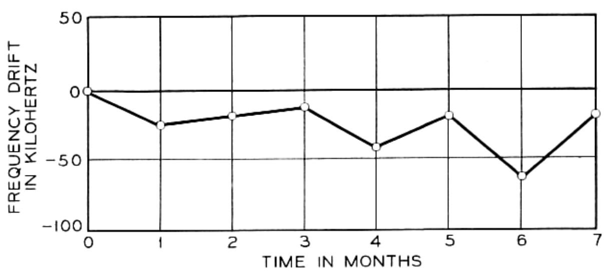

Fig. 14 — Typical frequency stability record for a production transmitter.

The outputs of the deviated oscillators at approximately +2 dBm are connected through single-stage, common-base amplifiers to the transistor mixer circuit as shown in Fig. 15.

The emitter circuit of the mixer transistor is biased to minimize undesired products and provides the desired difference output at −24 dBm. The transistor mixer has the desirable feature that the impedance presented by the output circuit has very little influence on the modulation performance of the emitter circuit. The relatively high 186 and 256 MHz oscillator signals in the mixer output are suppressed by the 697A low pass filter. The loss characteristic of the 697A filter is shown in Fig. 16.

Out-of-band mixer products must be limited in magnitude to minimize adjacent channel interference and in-band interference resulting from modulation sidebands. Undesired in-band products must be at least 96 dB below the carrier. The only significant in-band tone is the high order 10V-7C product and it is the 96 dB objective for this tone that limits the maximum output of the mixer to −24 dBm.

The modulation sensitivity of the deviator ranges from 2.9 to 4.2 MHz per volt, depending upon the particular 457A varactor diodes used. Since all transmitters must have the same sensitivity, this variation is compensated for by providing adequate gain range in the baseband amplifier. The sensitivity is virtually constant over the whole baseband range and in combination with the 3A receiver meets the ±0.1 dB baseband response objective.

3.4 *IF Amplifier-Limiter*

Residual amplitude modulation in the output of the deviator is removed by the amplifier-limiter which also provides 22 dB of gain. The design is an adaptation of the amplifier-limiter used in the 100A protection switching system for pilot stabilization.[6] The limiter is very similar to the TD-3 repeater limiter described in Ref. 7 except that the level-to-phase neutralizing network has been eliminated.

The amplifier-limiter provides 20 dB of AM suppression and has

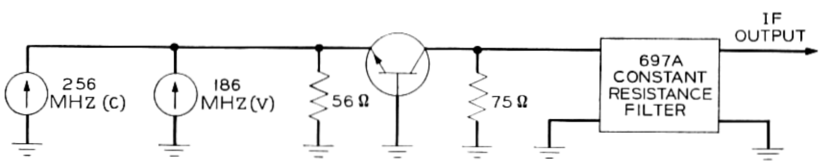

Fig. 15 — Simplified mixer circuit.

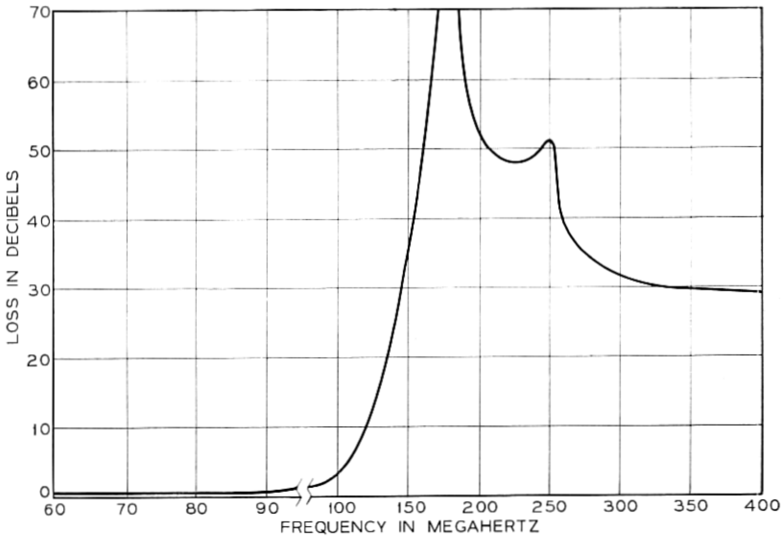

Fig. 16 — 697A filter loss characteristic.

input and output return losses in excess of 32 dB. The passband frequency response is flat within 0.03 dB from 60 to 80 MHz.

3.5 *IF Main Amplifier*

The IF main amplifier raises the -4 dBm output of the amplifier-limiter to the FM transmitter output level of $+10$ dBm. The individual stages use the same basic design as the repeater IF amplifier.[7]

A second output, 12.2 dB below the main output, is available for driving the optional monitor circuit. Input and output return losses on the main amplifier exceed 35 dB. The frequency response is flat within 0.03 dB in the 60 to 80 MHz band.

3.6 *Frequency and Level Monitor*

The frequency and level monitoring circuit is essentially an FM receiver without limiting. It provides output indications proportional to the error in average intermediate frequency and the error in IF output power relative to its nominal $+10$ dBm. The error signals are applied to meter relays in the transmitter control unit and the relay-operate points are set to provide alarm outputs when the intermediate frequency error exceeds ± 200 kHz or the change in IF output power exceeds ± 2 dB. The monitor unit also provides two independent 75-

ohm baseband outputs that can supply a 0 dBm signal for 8 MHz peak-to-peak sine wave deviation.

The frequency alarm point at an error of ±200 kHz is beyond the ±100 kHz objective for transmitter frequency stability. Since a properly functioning transmitter will stay within the 100 kHz objective, the primary function of the 200 kHz alarm indication is to call attention to a trouble condition when it reaches a point where it will begin to have an adverse effect on system performance.

The discriminator-baseband amplifier portion of the monitor circuit is very similar to the corresponding circuitry in the 3A FM receiver which is described in Section 4.3. Power and frequency alarm indications are derived from the sum and difference outputs, respectively, of the two balanced discriminator detector diodes.

3.7 *Transmitter Control Unit*

The transmitter control unit provides individual alarm outputs, an alarm cutoff feature, and alarm circuit breaker protection, self-contained within the terminal. This unit is optional and is not provided when the monitor unit is omitted.

Off-frequency and IF level error signals from the monitor unit operate individual meter relays with high and low latching type contacts. The meter relays provide external alarm indications and a local visual indication of trouble conditions.

IV. FM RECEIVER

The FM receiver accepts the 70 MHz FM signal and delivers a balanced baseband output of +4 dBV in a 124 ohm circuit for 8 MHz peak-to-peak deviation.

Performance requirements with respect to linearity, baseband response, and stability are similar to (and must be compatible with) the over-all terminal pair objectives.

The design is fairly conventional and similar to the approaches used on corresponding equipment in the TH and TD-2 systems.[2, 3]

4.1 *General Description*

Figure 17 is a block diagram of the FM receiver. The IF input to the receiver is applied to the 75-ohm input of the limiter-amplifier at a power of −7 dBm (+3 dBm on an earlier version of the receiver). Here, the first of two limiters is the major source of AM suppression in the receiver and has low amplitude-to-phase conversion. Follow-

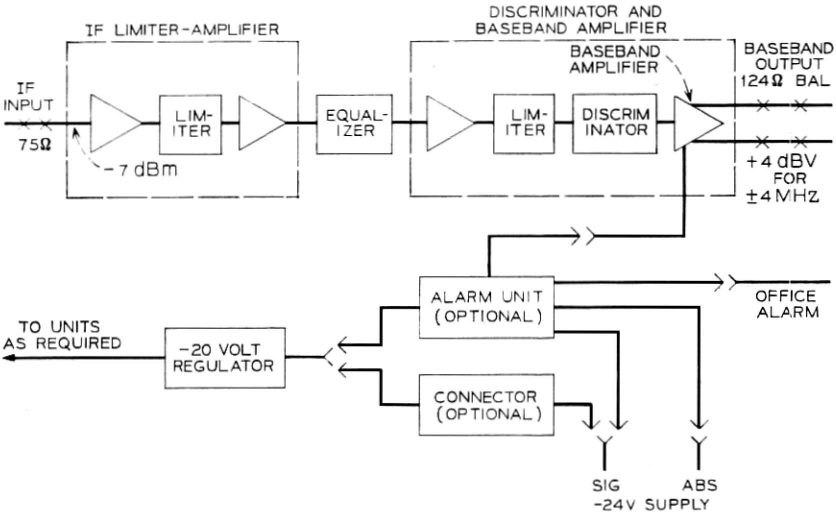

Fig. 17 — Block diagram of FM receiver.

ing the limiter are several stages of gain and an equalizer which corrects for delay distortion in the over-all receiver. The second limiter extends the dynamic range of the receiver such that performance is essentially unaffected by reductions up to 10 dB in IF input power.

The output of the second limiter is applied to the discriminator where it is demodulated. The recovered signal is then amplified and an output of +4 dBV is provided from a balanced 124-ohm impedance for a peak-to-peak deviation of 8 MHz.

An optional alarm unit in the receiver provides an indication of low or high IF level at the discriminator and an indication of changes in the dc conditions on the baseband amplifier transistors. Special care was taken in the design of the receiver to ensure reproducible adjustment of the discriminator networks which are the major source of envelope delay distortion in the terminal. Field adjustment controls were also carefully selected to minimize the possibility of degrading the delay equalization.

Figure 18 is a photograph of the receiver.

4.2 *IF Limiter-Amplifier*

Early models of the limiter-amplifier operated at an IF input power of +3 dBm. This was later changed to −7 dBm to compensate for office cabling losses.

A common emitter configuration is used for the input stage while all other gain stages have the common base configuration used in the repeater IF amplifier described by G. L. Fenderson and others.[7] The series diode limiter is of the compensated type described in the same paper.

The limiter-amplifier suppresses amplitude modulation by at least 40 dB at nominal drive. It provides an output of +8 dBm, flat within 0.03 dB between 60 and 80 MHz.

4.3 *Discriminator-Baseband Amplifier*

The discriminator-baseband amplifier performs the demodulation function in the FM receiver. The first stages of the unit comprise a limiter and a low pass filter. This limiter reduces amplitude modulation by 25 dB and stabilizes the IF input to the discriminator. Because of the combined action of the two receiver limiters, a 10 dB drop in IF input (from nominal) produces less than a 0.25 dB reduction in baseband output. This is illustrated in Fig. 19 where receiver sensitivity is plotted as a function of IF input power.

The limiter output is temperature stabilized by thermistor control of the limiter diode bias current. The limiter circuit is very similar to corresponding stages of the amplifier-limiter described in Section 3.4. A low pass filter following the limiter rejects harmonics of the 70 MHz IF to prevent distortion products being generated in the discriminator detectors.

To make the baseband amplifier noise contribution negligible, it

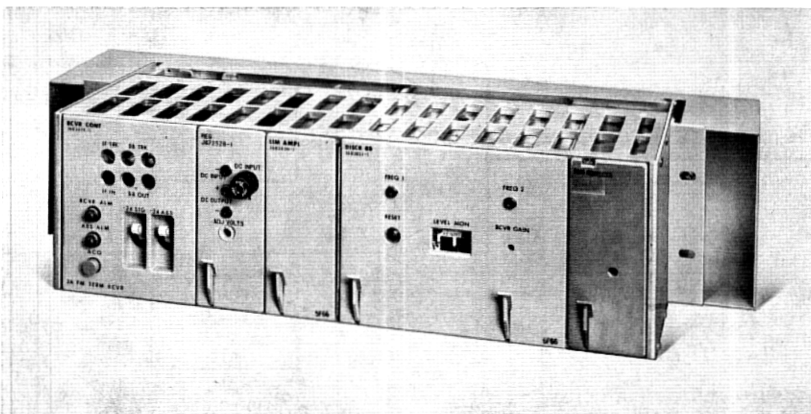

Fig. 18 — 3A FM receiver.

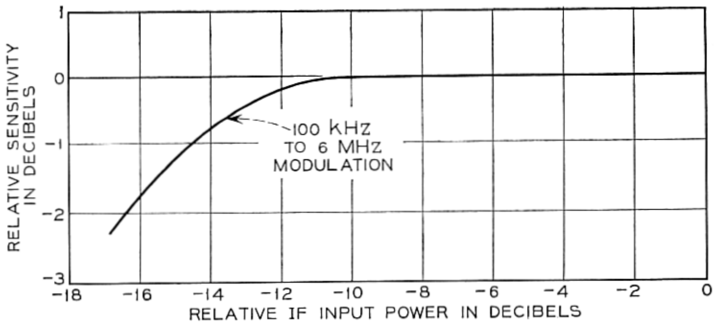

Fig. 19 — Relative receiver demodulation sensitivity as a function of IF input power.

was necessary to provide a discriminator sensitivity of at least 0.25 volt per MHz. For a given input power and degree of complexity, there is in general a trade-off between discriminator sensitivity and linearity. Using the ratio of sensitivity to nonlinearity as a figure of merit, several circuit configurations were compared before selecting the modified balanced resonant discriminator similar to that used in the TH radio system.[3]

To determine the relationship between sensitivity and nonlinearity in the balanced resonant discriminator, the idealized circuit of Fig. 20 was optimized on the computer. The computer was programmed to adjust the circuit elements to obtain the best match between $|E_2| - |E_3|$ as a function of frequency and a straight line of specified slope

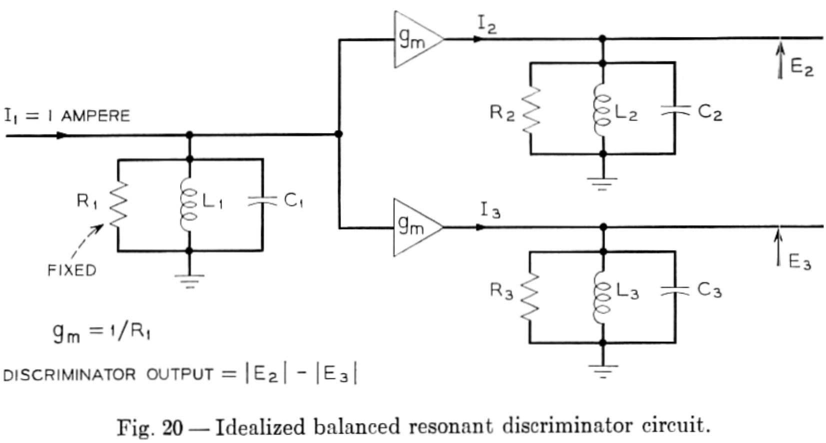

Fig. 20 — Idealized balanced resonant discriminator circuit.

passing through zero at 70 MHz. By repeating this process for several different slope objectives corresponding to different sensitivities, the relationship between sensitivity, linearity, and element values was determined.

The amplifiers in Fig. 20 have infinite input and output impedances whereas in the actual circuit, common-base transistor stages with low input impedances are used. If the damping resistor R_1 is in series with the emitters, the emitter currents are proportional to the voltage across R_1 so that the idealized circuit becomes a satisfactory representation of the actual circuit. Good agreement was obtained between measured and computed sensitivity and linearity. In Fig. 21, the high and low frequency tank circuits in the collectors of Q_9 and Q_4 provide impedance peaks at 89 and 51 MHz, respectively. The highly damped tank in the collector circuit of Q_3 provides shaping of the common IF signal around 70 MHz to attain an over-all discriminator linearity of better than 0.5 per cent for 8 MHz peak-to-peak deviation.

Printed wiring gives close control of most of the stray inductance and, to a considerable extent, the stray capacitance. By using close tolerance, tension wound inductors, and by adjusting the circuit capacitance with trimmers to correct for differences in transistor and diode capacitances, excellent reproducibility of the discriminator delay characteristics was obtained. The high and low frequency tank circuits of the discriminator need no field adjustments. In the common or center frequency tank circuit, both the inductor and capacitor are variable to facilitate alignment.

An important and useful technique to reduce and control stray inductance and couplings in high frequency ground paths was the use of a ground plane as illustrated in Fig. 22. This ground plane is placed about one half inch below the printed wiring board with threaded studs connecting wherever ground reinforcement is needed on the printed wiring board.

A high-speed epitaxial silicon Schottky-barrier diode, coded 479B, with low forward voltage and high reverse breakdown voltage was developed for the 70 MHz detector. High reverse breakdown voltage allows relatively high discriminator drive with correspondingly improved discriminator sensitivity. A combination of high detector load impedance and high detection efficiency results in the detector circuits contributing less than 10 per cent to discriminator tank circuit loading.

The high sensitivity of the discriminator provides such high signal

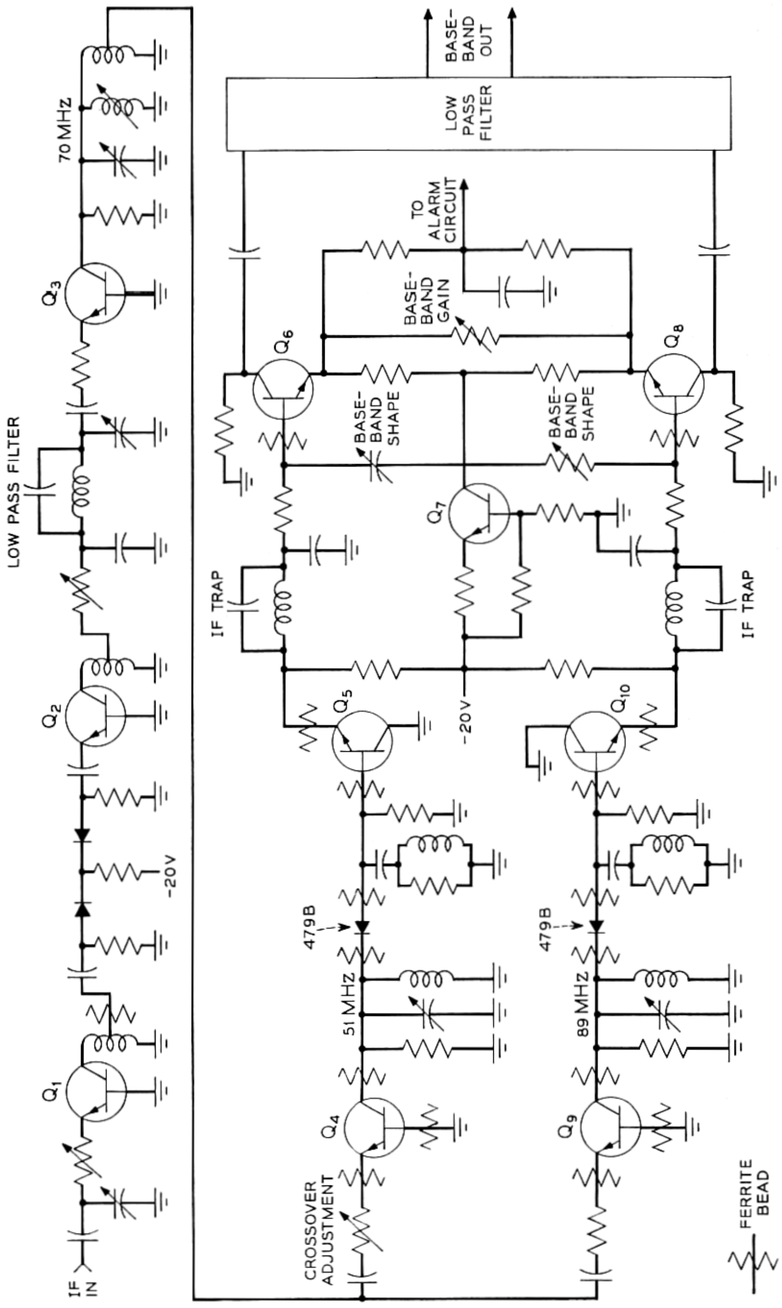

Fig. 21 — Simplified schematic diagram of discriminator-baseband amplifier unit.

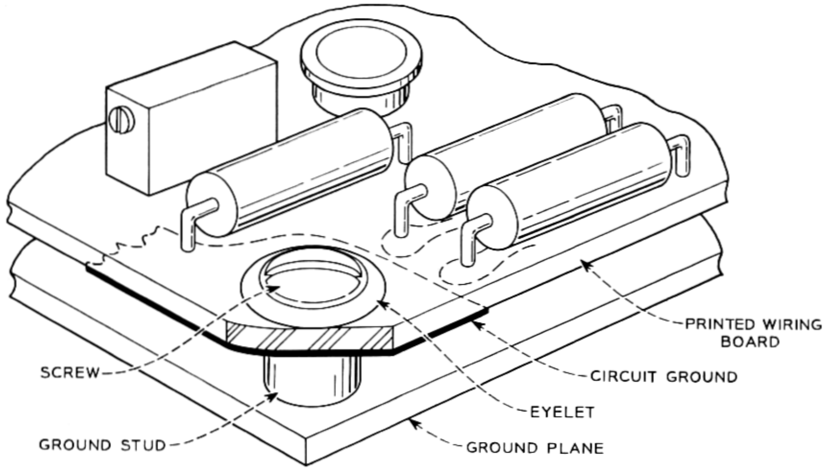

Fig. 22 — Grounding technique.

input to the baseband amplifier that little voltage gain is needed and the amplifier noise is relatively unimportant. The controlling design requirements for this amplifier were impedance transformation to the 124-ohm output, differential balance, and frequency response. The low-end frequency response of 3 dB down at 0.6 Hz led to the choice of dc coupling between stages and a circuit configuration permitting the use of low-voltage output coupling capacitors to minimize space requirements.

Emitter followers were chosen for the input stages of the amplifier to provide a high impedance load for the discriminator. Resonant 70 MHz traps in the emitter follower outputs attenuate the IF signal and its harmonics. These IF signals are further attenuated by the low pass filter in the amplifier output to be less than 3 millivolts from tip or ring to ground. The high end frequency response is controlled by an adjustable RC filter which is connected between the inputs of Q_6 and Q_8 to adjust the over-all transmitter and receiver frequency response flat to ± 0.1 dB from 6 Hz to 10 MHz.

The baseband signals from the two halves of the discriminator must be subtracted to cancel their first-order nonlinearities. The differential amplifier, Q_6, Q_7, and Q_8, performs this combining function.

An optional IF level alarm is used to detect the loss of IF signal in the receiver. The discriminator diode detector outputs are dc coupled through the baseband amplifier to the emitters of Q_6 and Q_8.

The IF level on both halves of the discriminator as well as baseband amplifier bias conditions are therefore monitored by summing these emitter voltages. The monitored voltage is stored in a capacitor and compared to a reference voltage by a zero center, latching-type, meter relay. The meter relay will operate with either an increase or decrease in monitored voltages equivalent to 3 dB changes in IF level at the discriminator. The capacitor smooths rapid voltage changes to prevent alarms on hit type failures. It also stores sufficient energy to latch the meter relay when circuit power is removed. The meter relay contacts energize alarm relays in the control unit to provide external alarm indications, and a local visual indication of trouble.

4.4 *Receiver control Unit*

The receiver control unit serves the same purpose and is very similar to the corresponding transmitter unit. In this case, however, the meter relay which responds to loss of IF signal at the discriminator is located on the discriminator-baseband amplifier unit.

V. POWER REQUIREMENTS

The FM terminals operate from a −24 volt office battery. The transmitter requires a maximum of 3.0 amperes from the −24 volt supply (signal battery) with a typical requirement of 1.5 amperes at ambient temperatures around 20°C. The dependence on temperature results primarily from variations in power requirements on the deviator oven. The −24 volt receiver drain is 0.75 ampere. When equipped with alarm and control units, the transmitter and receiver each require a maximum of 0.2 ampere from the −24 volt alarm battery supply.

In both the transmitter and receiver, the signal battery supply is fed through a voltage regulator providing an output of −20 volts ±1 percent.

VI. TANDEM PERFORMANCE

Based on laboratory and production experience, the tandem performance of the FM transmitter and receiver has shown a very satisfactory degree of reproducibility so that the quoted performance can be regarded as typical. Compared with similar vacuum tube equipment, performance variations with time are relatively small and in most instances there is not yet enough experience to quantitatively characterize the stability.

6.1 *Noise Load Performance*

The most satisfactory measure of terminal performance from the standpoint of message applications is based on the use of band-limited fluctuation noise to simulate the multi-channel speech load. Noise and cross-modulation performance measured in this manner is presented in Figs. 23, 24, and 25 corresponding to 600, 900, and 1200 channel message loading, respectively. The performance at reference drive with 1200 channel loading is well within the TD-3 objective of 21 dBrnC0.

6.2 *Fluctuation Noise*

Fluctuation noise, generated by a tandem transmitter and receiver, and measured at the receiver output, is shown in Fig. 26. This total noise is made up of several distinguishable contributors as indicated in the figure. Fluctuation noise at high baseband frequencies is controlled by the noise figure of the transmitter amplifier-limiter which has a relatively low input signal power of -26 dBm. In the baseband region below about 1 MHz, the main fluctuation noise sources in

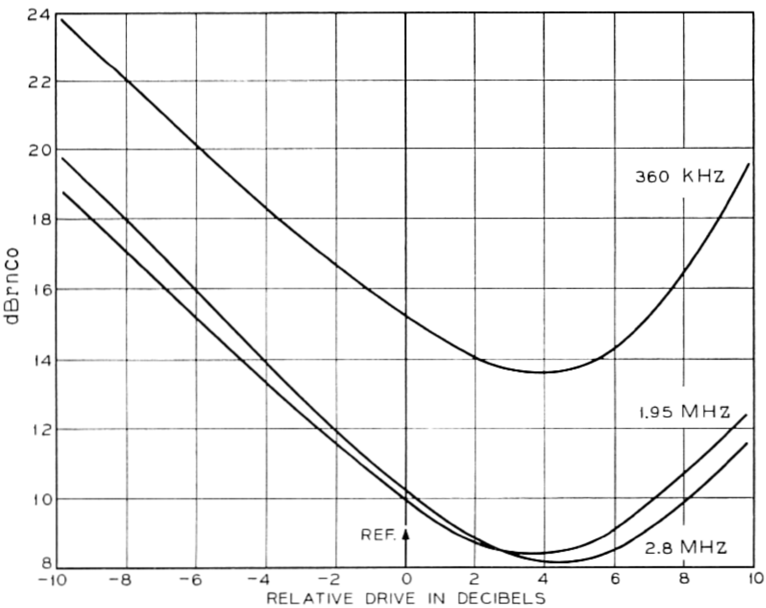

Fig. 23 — 600 channel noise load performance of a terminal pair; reference drive at -23.8 dBm.

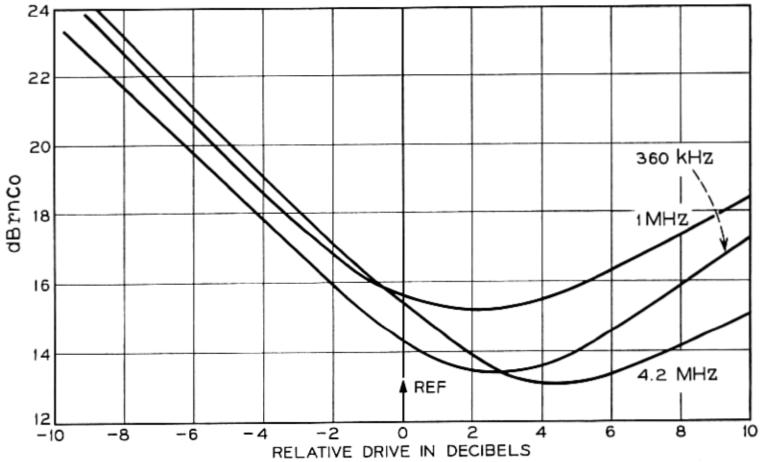

Fig. 24 — 900 channel noise load performance of a terminal pair; reference drive at —23.5 dBm.

order of importance are the transmitter baseband amplifier, the deviator, and the receiver.

For television, in addition to the spectral nature of the noise, the ratio of signal to total noise is of interest. This is usually expressed as the ratio of peak-to-peak video signal to weighted (rms) noise

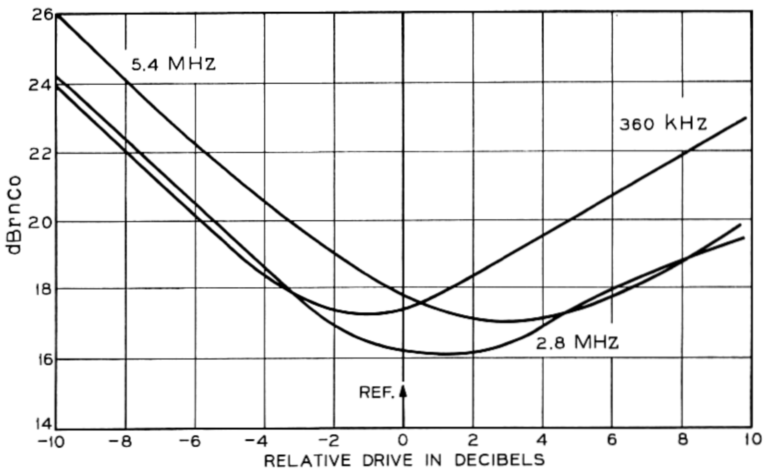

Fig. 25 — 1200 channel noise load performance of a terminal pair; reference drive at —23.2 dBm.

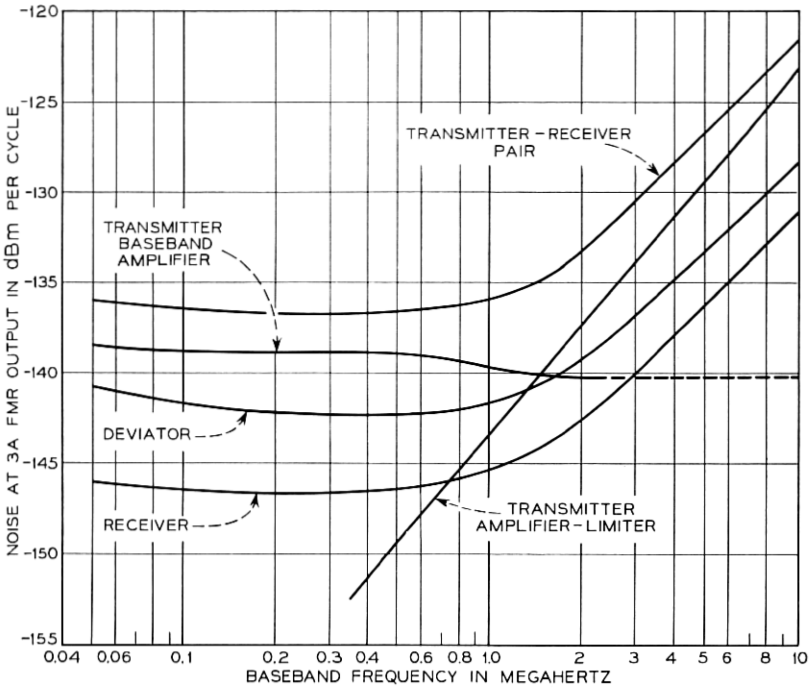

Fig. 26 — Composition of FM terminal fluctuation noise.

where the color weighting curve is as shown in Fig. 27. For the noise measurement, the weighted baseband noise is generally separated into two regions, one encompassing the low frequency end of the band up to 4 kHz and the other encompassing the range from 4 kHz up to approximately 4.5 MHz. The noise is measured at the FM receiver output following the video de-emphasis network. Typical values for signal-to-weighted low frequency noise and signal-to-weighted (color) high frequency noise are 91 and 84 dB, respectively, meeting the corresponding 40 and 73 dB objectives with a comfortable margin.

6.3 *Baseband Amplitude Response*

Figure 28 shows a typical baseband response characteristic.

For television purposes the low frequency response is frequently characterized by the response to a 60 Hz square wave. Slope on the square wave output, expressed as a percentage of the peak-to-peak

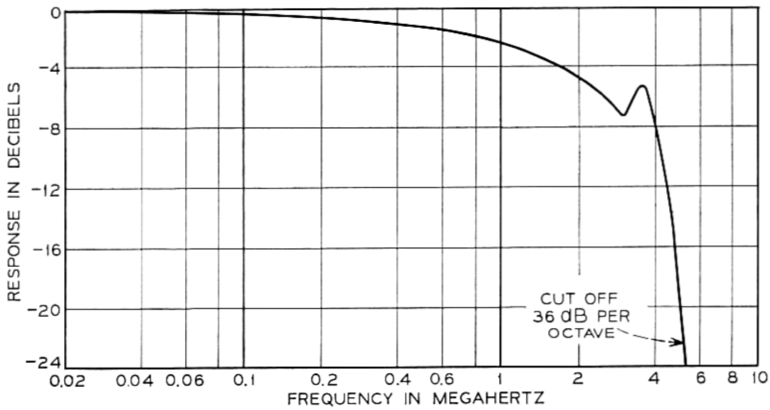

Fig. 27 — Color TV noise weighting curve.

signal is a measure of the phase fidelity at the fundamental frequency. Typical slope, measured at the output of a 3A terminal pair, is 1.5 per cent, which is consistent with a lower 3 dB cutoff frequency of approximately 0.5 Hz.

6.4 *Differential Gain, Differential Phase and Linearity*

With 8 MHz peak-to-peak deviation, the differential gain is 0.1 dB and the differential phase is 0.15 degree; both are predominantly linear in shape. Since pre-emphasis reduces the low frequency video components by 12 dB, the differential gain and phase experienced by the TV signal will be a factor of four less and is therefore well within the design objectives. Over-all nonlinearity is primarily determined by the transmitter which has a residual parabolic component of typically 1.5 per cent for 8 MHz peak-to-peak deviation.

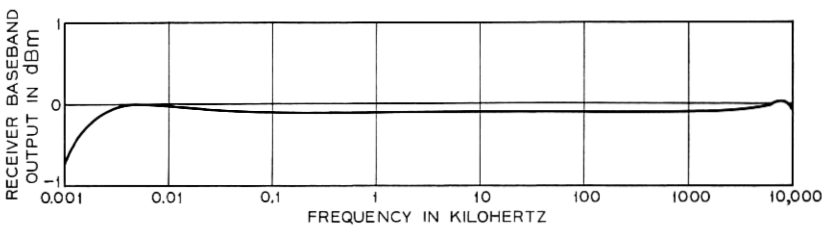

Fig. 28 — Baseband frequency response of a terminal pair.

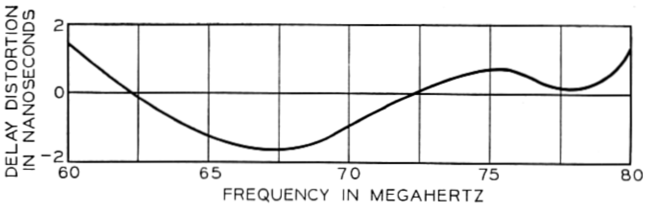

Fig. 29 — Envelope delay distortion of a terminal pair.

6.5 *Envelope Delay Distortion*

For all practical purposes the FM transmitter and receiver are separately delay equalized. In the transmitter there is residual envelope delay distortion of approximately 2 nanoseconds parabolic at ± 10 MHz. The residual delay distortion of the over-all terminal pair is shown in Fig. 29.

6.6 *Spurious Transmitter Output Tones*

Spurious tones present in the deviator output are band-limited and amplitude-limited in succeeding portions of the transmitter. The resulting relative amplitudes of these spurious tones are listed in Table II.

VII. TEST FACILITIES

Because of the expected frequent application of 3A FM terminals in existing TD-2 stations, it was decided to make maximum use of existing terminal test facilities. For the maintenance of 3A terminals, the TD-2 FM terminal test console is supplemented by a new test panel which replaces a blank panel in the test console. The composi-

TABLE II — OUT-OF-BAND TONES AT THE FM TRANSMITTER OUTPUT

Tone frequency (MHz)	Product	Power relative to 70 MHz IF output (dB)
23.33	4V-3C	<-87
46.67	3V-2C	-76
93.33	5V-4C	-79
116.67	2V-C	-69
140.00	2C-2V	-23
186.67	V	-86
256.67	C	<-87

tion and functions of the test panel are described fully in Ref. 8.

The general approach to maintenance testing has been to confine routine tests to input-output measurements. Trouble on individual units is isolated by substituting a spare unit for the suspected defective unit.

VIII. EQUIPMENT FEATURES

All major components in the 3A transmitter and receiver are plug-in units for easy replacement and maintenance. They are located in an open box framework suitable for mounting on a 19-inch duct bay. The transmitter has two rows of plug-in units totaling 10½ inches high. The receiver has one row of units totaling 5¼ inches high. The transmitter and receiver weigh about 45 and 18 pounds, respectively.

The terminals also are available in portable carrying cases. A companion 117 volt, 60 Hz rectifier unit is available with an output capacity of 4 amperes at −24 volts. This unit can power a transmitter or receiver individually or both units simultaneously. Since the FM receiver may be used more frequently in portable form and because its power requirements are modest, an optional 117 volt rectifier unit is available, mounted inside the rear cover of the receiver case. In all instances the portable terminals are equipped for optional connection direct to the −24 volt office battery. Figure 30 shows the portable terminals with front covers removed and stacked for use.

APPENDIX

A.1 *Linearity of Varactor Diode Voltage Controlled Oscillator*

In the region where a quasistationary analysis is appropriate, the relationship between instantaneous frequency and modulating voltage may be expressed as a power series of the form

$$f(v) = f_0 + \sum_i a_i v^i \tag{1}$$

where $f(v)$ = instantaneous oscillator frequency
f_0 = oscillator rest frequency
v = applied signal voltage.

Using an approach similar to that discussed in Appendix A of Ref. 3, the following relationships can be defined in terms of the coefficients in equation (1).

$$\text{Deviation sensitivity} = a_1 \tag{2}$$

$$\text{First order nonlinearity} = \frac{4a_2\ \Delta f}{a_1^2} \tag{3}$$

$$\text{Second order nonlinearity} = \frac{3a_3\ \Delta f^2}{a_1^3} \tag{4}$$

where Δf = peak frequency deviation.

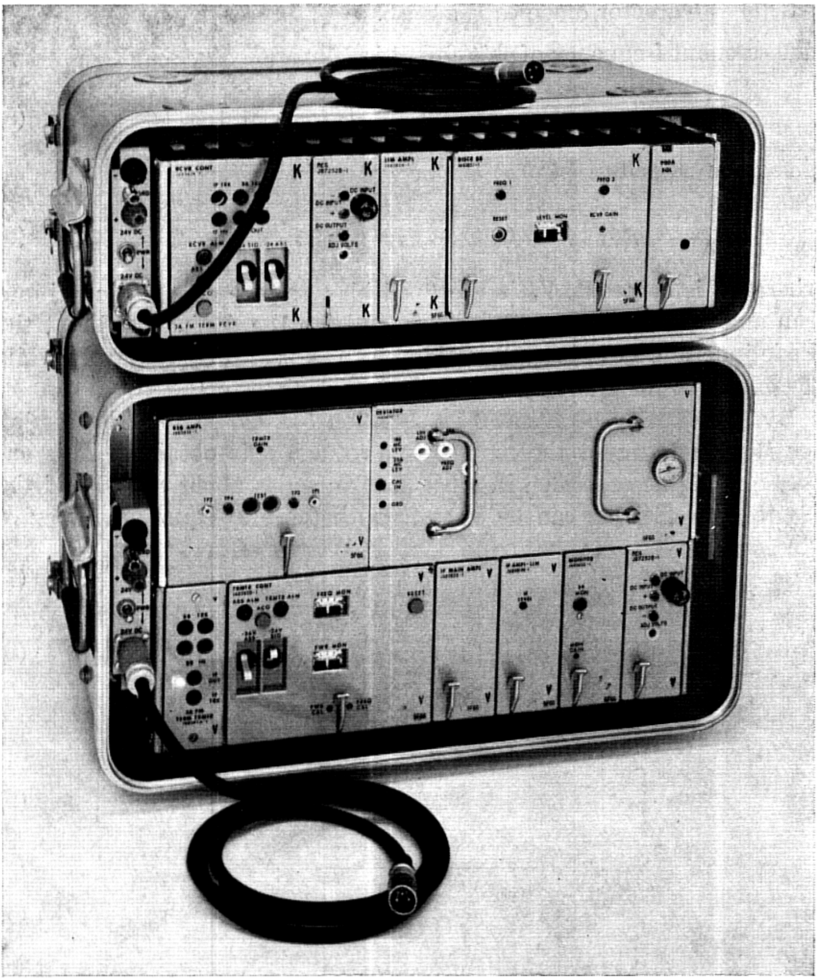

Fig. 30 — Portable transmitter and receiver equipped for dc operation.

Notice that equation 3 defines peak-to-peak nonlinearity in contrast with Ref. 3, which defines the peak value. The choice is somewhat arbitrary except that the peak-to-peak definition is more in keeping with values commonly measured on linearity test equipment.

Consider the parallel resonant tank circuit of a voltage controlled oscillator as shown in Fig. 31.

L = effective shunt inductance
C_s = total stray capacitance
$C_v(v)$ = varactor diode capacitance.

The resonant frequency of this circuit is

$$f(v) = \frac{1}{2\pi}\,[L(C_v(v) + C_s)]^{-1/2}. \tag{5}$$

The diode capacitance $C_v(v)$ can be written in the form

$$C_v(v) = \frac{C_0}{(V_0 + v)^n} \tag{6}$$

where C_0 is constant, V_0 is the fixed bias voltage on the diode junction and n is an exponent which depends on the doping profile at the diode junction. For abrupt junction diodes n is approximately ½ while the exponent for graded junction devices is close to ⅓.

By substituting equation 6 in equation 5 a closed form expression for $f(v)$ in terms of the modulating voltage v is obtained. This expression can be expanded in a MacLaurin series about $v = 0$ and the resulting coefficients can be equated with those in equation 1 to give

$$a_1 = \frac{nf_0}{2V_0}\left(1 - \frac{C_s}{C_B}\right) \tag{7}$$

$$a_2 = \frac{1}{2}\,\frac{nf_0}{(2V_0)^2}\left(1 - \frac{C_s}{C_B}\right)\left(n - 3n\,\frac{C_s}{C_B} - 2\right) \tag{8}$$

$$a_3 = \frac{1}{6}\,\frac{nf_0}{(2V_0)^3}\left(1 - \frac{C_s}{C_B}\right) \cdot\left[n^2 - 6n + 8 - 6n(2n - 3)\,\frac{C_s}{C_B} + 15n^2\left(\frac{C_s}{C_B}\right)^2\right]. \tag{9}$$

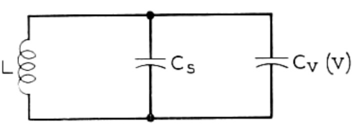

Fig. 31 — Voltage controlled oscillator tank circuit.

In the above expressions, C_B is the total tank circuit capacitance with $v = 0$ or

$$C_B = C_s + \frac{C_0}{V_0^n}. \tag{10}$$

Equations 7, 8 and 9 may now be substituted in equations 2, 3 and 4 to yield the following results.

$$\text{Deviation sensitivity} = \frac{nf_0}{2V_0}\left(1 - \frac{C_s}{C_B}\right) \tag{11}$$

$$\text{First order nonlinearity} = \frac{\Delta f}{f_0}\left[\frac{2\left(n - 3n\dfrac{C_s}{C_B} - 2\right)}{n\left(1 - \dfrac{C_s}{C_B}\right)}\right] \tag{12}$$

Second order non-linearity

$$= \left(\frac{\Delta f}{f_0}\right)^2\left[\frac{n^2 - 6n + 8 - 6n(2n - 3)\dfrac{C_s}{C_B} + 15n^2\left(\dfrac{C_s}{C_B}\right)^2}{2n^2\left(1 - \dfrac{C_s}{C_B}\right)^2}\right]. \tag{13}$$

Notice that the deviation sensitivity is inversely proportional to the bias voltage and that the nonlinearities depend on the percentage deviation.

A.2 *Linearity of the Heterodyne Deviator*

The heterodyne deviator in the 3A transmitter uses two voltage controlled oscillators whose rest frequencies differ by 70 MHz and which are deviated by baseband signals in antiphase. The desired difference frequency at the mixer output may therefore be expressed as

$$f(v) = f_h(v/2) - f_\ell(-v/2) = f_{0h} - f_{0\ell} + \sum_i (1/2)^i[a_{ih}v^i - a_{i\ell}(-v)^i] \tag{14}$$

where the subscripts h and ℓ refer to the high and low frequency oscillators, respectively. Also notice that v in equation 14 is the total balanced input voltage to the deviator which accounts for the $(\frac{1}{2})^i$ term in the power series.

With $V_{0\ell}$ and $C_{s\ell}/C_{B\ell}$ defined, there are two degrees of freedom in the high frequency oscillator, namely, V_{0h} and C_{sh}/C_{Bh}. This permits satisfying two arbitrary conditions by appropriate choice of these quantities. In the 3A deviator it was chosen to design the two oscillators with equal sensitivities and equal a_2 terms. To satisfy these requirements, C_{sh}/C_{Bh} must be related to $C_{s\ell}/C_{B\ell}$ in the following way.

$$\frac{C_{sh}}{C_{Bh}} = 1 - \frac{2(n+1)\left(1 - \dfrac{C_{s\ell}}{C_{B\ell}}\right)}{3n\left(1 - \dfrac{C_{s\ell}}{C_{B\ell}}\right) - \dfrac{f_{0h}}{f_{0\ell}} - \dfrac{f_{0h}}{f_{0\ell}}\left[n - 3n\dfrac{C_{s\ell}}{C_{B\ell}} - 2\right]}. \tag{15}$$

When the preceding conditions have been satisfied, the first order nonlinearity is zero and the dominant second order nonlinearity is given by

$$\frac{3(a_{3h} + a_{3\ell})\,\Delta f^2}{(a_{1h} + a_{1\ell})^3}. \tag{16}$$

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