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PACKAGED ELECTRIC TUNED 35-200 MC PANORAMIC RECEIVER

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ABSTRACT

A method of utilizing the voltage tuning characteristics of ferroelectric capacitors in a packaged, wide-range, superheterodyne, electric-tuned, panoramic receiver is described. This receiver employs titanate ceramic capacitors as tuning elements in the tank circuits of the RF, mixer, and local oscillator stages. The capacity of the tuning elements is varied by changing the electric field applied to the capacitors. The receiver is designed for use in monitoring communication signals in the frequency range 35-200 Mc.

This Packaged Electric Tuned Receiver PANoramic (Petr Pan) is the result of a development program designed to indicate the direction engineers can take in developing and packaging a dielectric tuned receiver suitable for operational use.

ACKNOWLEDGEMENT

The author wishes to acknowledge the contributions of Dr. H. W. Welch, Jr. for his encouragement and direction during the course of this work; of Mr. H. Diamond for his development of the ferroelectric capacitors; and of Dr. L. W. Orr for his assistance in the circuit design work.

PACKAGED ELECTRIC TUNED 35-200 MC PANORAMIC RECEIVER

I. INTRODUCTION

The receiver described in this report is a packaged, wide range, superheterodyne, electric tuned, panoramic receiver employing titanate ceramic capacitors as tuning elements in the tank circuits of the RF, mixer and local oscillator stages. This receiver, designated Petr Pan, is designed to monitor communication signals in the frequency range 35 to 200 MC. It is the result of a development program designed to indicate the direction engineers can take in developing and packaging a dielectric-tuned receiver suitable for operational use.

In general, dielectric tuning techniques utilize the non-linear electrical characteristics of certain types of ferroelectric materials. Barium-strontium titanate materials constitute the major class of dielectrics that are presently being applied to ferroelectric tuning devices. The capacity of the tuning element is varied by changing the electric field applied to the capacitor.

A photograph of the Petr Pan receiver is shown in Figure 1 and a block diagram is shown in Figure 2.

The receiver front-end assemblies are plug-in units containing the electrically tunable stages; i.e., the RF, mixer, and local oscillator stages. A low noise, cascode input stage provides maximum gain with minimum noise. The characteristics of the FE assembly are given in Table 1.

The receiver makes use of three 20 MC, synchonously-tuned IF stages followed by a crystal diode detector. The IF strip provides 130 KC resolution throughout the operating range of the receiver. The IF strip is a plug-in unit

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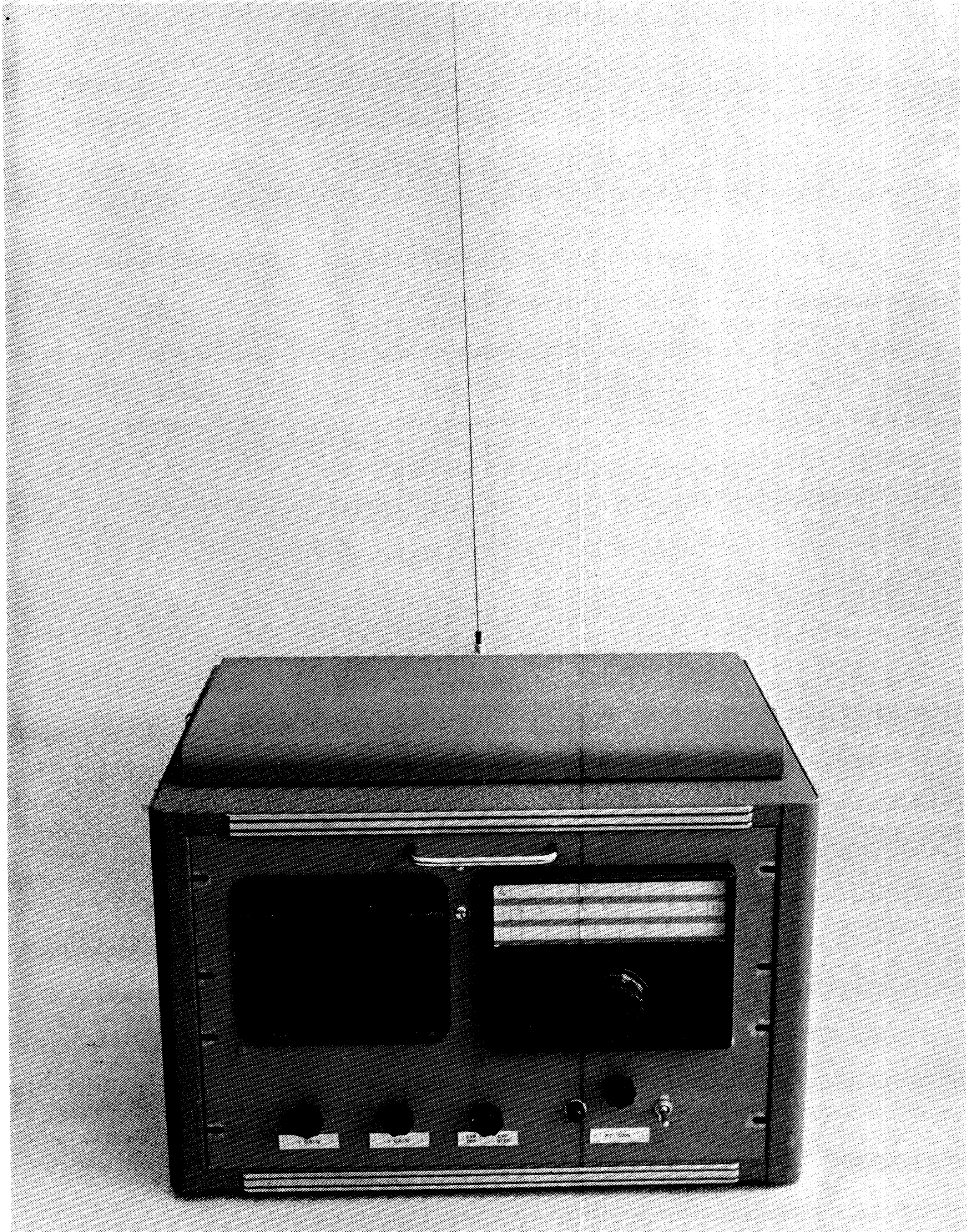

FIG. 1. PETR PAN RECEIVER.

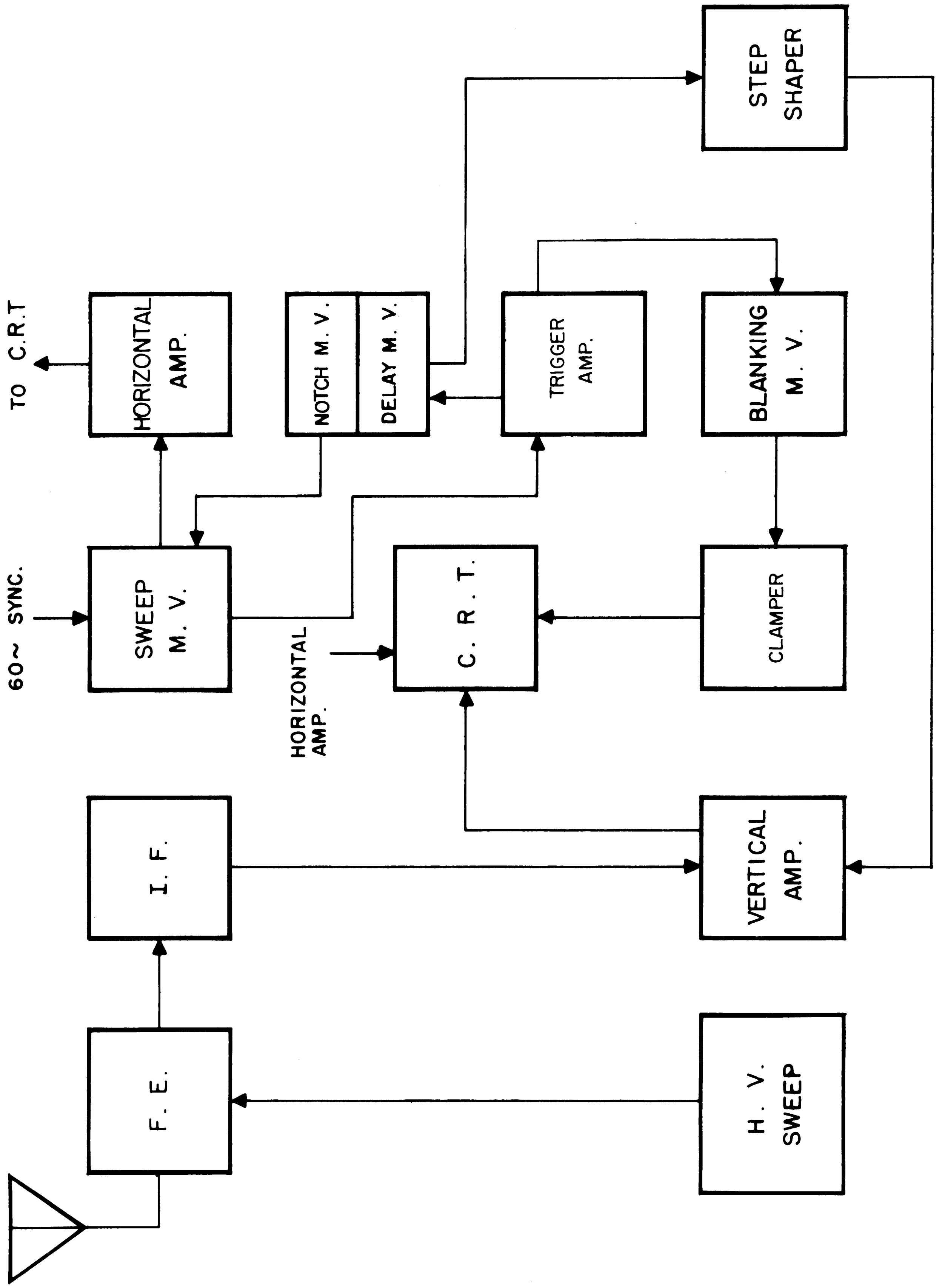

FIG. 2. BLOCK DIAGRAM
PETR-PAN RECEIVER

and may be removed from the main chassis and operated remotely by means of an auxiliary test cable.

Since it is desirable to have a linear frequency change with time, thus obtaining a uniform response rate, the voltage used to tune the capacitors must be shaped. The shape of the required tuning voltage to give a linear frequency change with time is approximated quite closely by a simple RC circuit.

Table 1
Characteristics of Receiver

FE Assy.	Freq. (MC)	Sensitivity (μv)	N.F. (db)	Scan Rate
A	35-70	2	7-10	60 cps
B	70-130	4-8	10-13	60 cps
C	130-200	4-8	12-15	60 cps

NOTE:
1. Step-Marker Circuit is provided to permit quick, accurate frequency measurement of received signals.
2. Expanded sweep is provided to permit separation of received signals that are close in frequency.
3. Display is 5" CRT.
4. All electrically-tuned circuits are swept by a single voltage source.
5. Input is 50 Ω.

The panoramic display utilizes a 5" CRT, and features (1) a variable-position notch sweep which is very useful when it is desired to separate and examine two signals that are very close in frequency, and (2) a passive marker circuit which provides a vertical step in the base-line at the frequency indicated on the calibrated slide-rule dial.

Self-contained power supplies furnish all necessary voltages at the currents required for satisfactory operation of the video and sweep portions of the receiver, and a Hewlett-Packard Model 712-A power supply provides the proper voltages and currents to the remainder of the receiver. All controls that are necessary for the operation of the receiver are located on the front panel.

II. CIRCUIT DESCRIPTION

2.1 Front-End Assemblies

The heart of this receiver, and the point at which it differs most widely from other receivers, is in the three front end assemblies. The assemblies are similar in design and construction, differing only in the values of the inductances used in the tank circuits. The schematic of a typical assembly is shown in Figure 3.

The principle points of difference in circuitry between this receiver and conventional receivers occur in the tuned circuits and in the method of connecting the sweep voltage. Capacitances to ground are kept as low as possible to reduce the effect of shunt circuit capacitance on the tuning range. The Q of the tank circuits used is lower than ordinarily encountered in receiver design, due to the relatively low Q of the tuning capacitors. This necessitates the use of high g_m tubes in order to obtain satisfactory results. The bias and sweep voltages are decoupled from the RF circuits to avoid increases in RF loading and stray capacitance. Pairs of series-connected capacitors are used in all tank circuits. This affords a lower minimum capacitance and at the same time supplies a convenient DC block for the bias voltage.

In considering the practical application of these principles to an electric-tuned receiver, the following features were considered to be of primary importance: (1) low noise figure; (2) gain; (3) single voltage tracking; and (4) selectivity.

Each FE assembly is a plug-in unit with 50-Ω output to the IF strip. The general parts layout and constructional details are shown in Figures 4 and 5.

2.1.1 RF Section. The maximum sensitivity of any receiver is determined largely by the characteristics of the first stage. The input circuit establishes the signal-to-noise ratio of the receiver and sets the upper limit of detectibility of weak signals.

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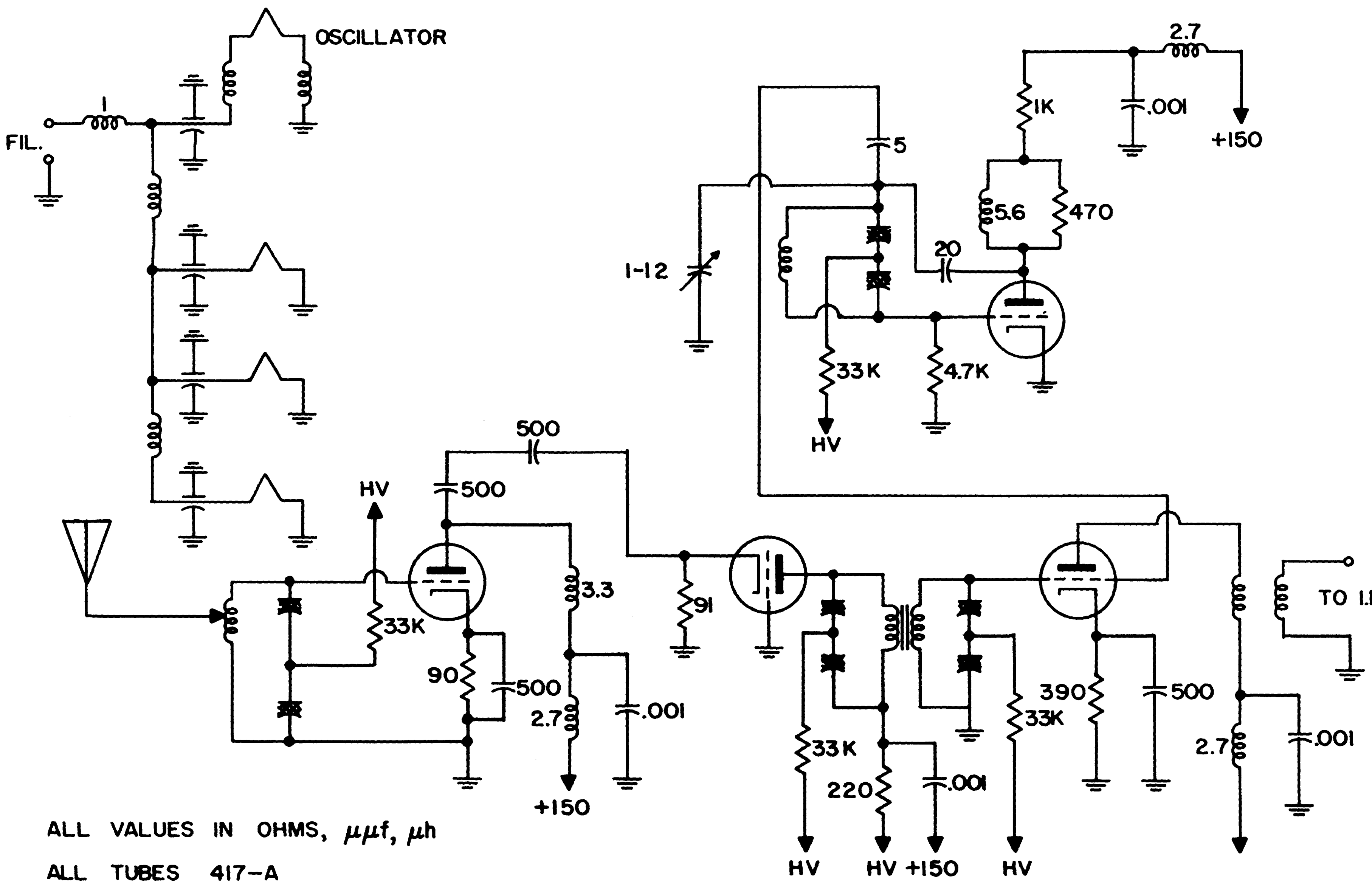

ALL VALUES IN OHMS, $\mu\mu f$, μh

ALL TUBES 417-A

1-P 4, 5, 7, 8 - g
2-NC 6 - K
3,9-FIL.
XXX-100$\mu\mu$f VOLTAGE

Fig. 3.

FRONT END ASSEMBLY "C" 130-200 MC

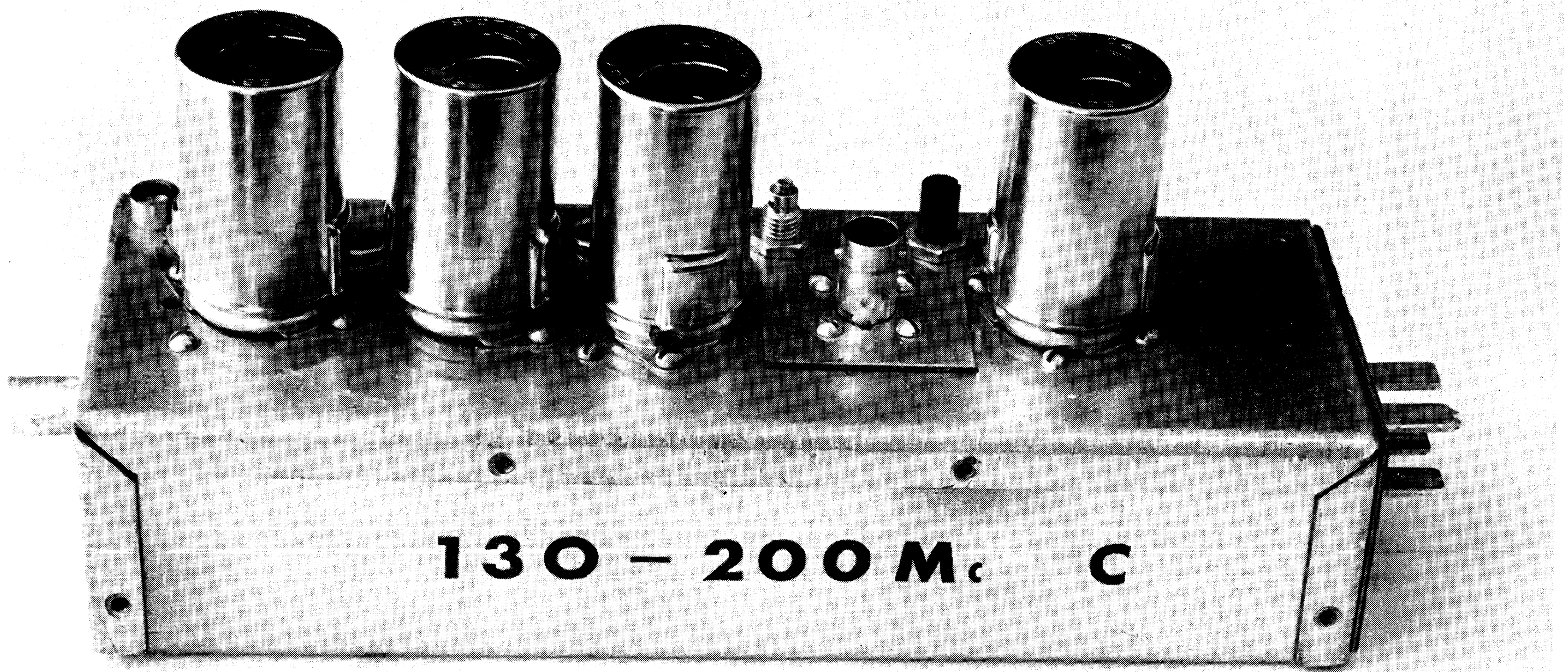

FIG. 4. EXTERIOR VIEW OF A
TYPICAL FRONT END ASSEMBLY.

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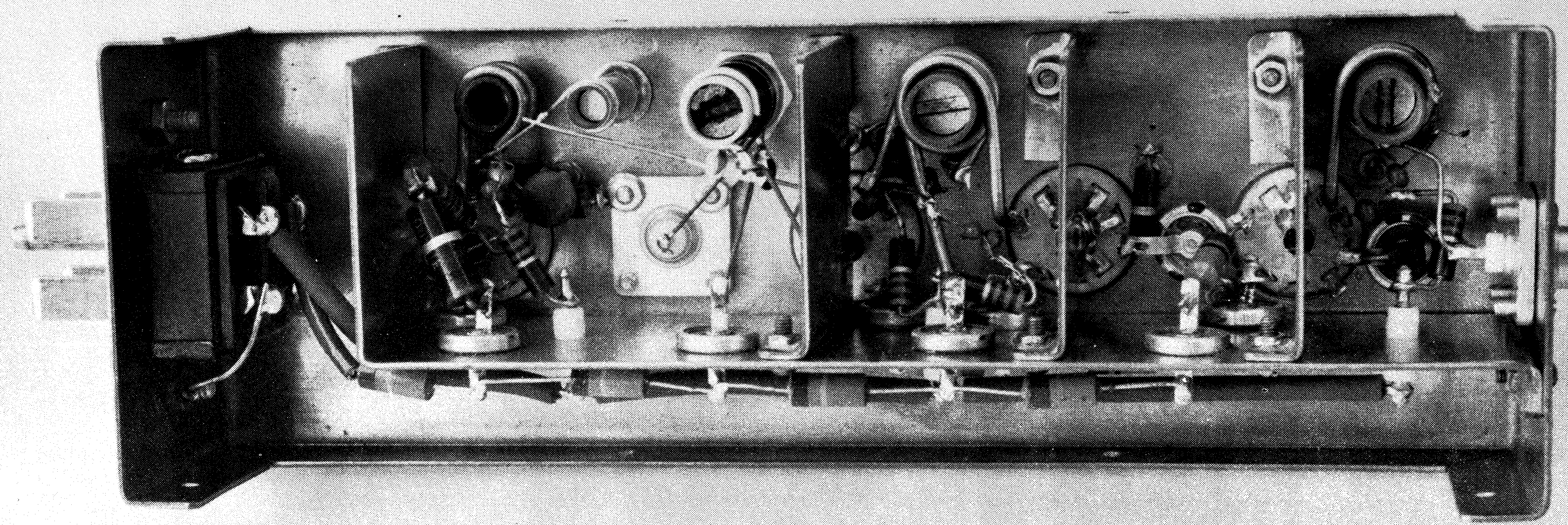

FIG. 5. INTERIOR VIEW OF A TYPICAL FRONT END ASSEMBLY.

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The minimum usable signal is determined by the noise voltages generated in the input stage. The antenna, power supply, passive circuit components, and tubes contribute noise. The most desirable combination for low noise and good gain is the cascode[1] circuit in which a grounded-cathode triode feeds a grounded-grid triode. Two triodes so connected provide an overall gain somewhat better than the gain from a single pentode in a grounded-cathode type RF amplifier. The noise figure improvement, however, is considerable.

As shown in Figure 3, a typical front end assembly is designed around an RF section consisting of a cascode input stage. The cascode stage consists of two 417-A triodes connected in a parallel arrangement. The parallel connection has an advantage over the series-type circuit in that lower plate voltage is required and the heater-cathode voltage is less. The parallel connection also permits the use of a common supply for the entire FE assembly.

2.1.2 Mixer Section. In designing the mixer section the conversion transconductance[2], the excitation requirements, and the noise figure were of primary interest. A triode was selected as a mixer since it provides a reasonable amount of gain with a good noise figure.[3]

2.1.3 Oscillator Section. A survey of various lumped-constant oscillator circuits suitable for dielectric tuning in the HF and VHF ranges was conducted and it was found that successful operation in all three ranges (see Table 1) could be obtained using a simple triode-connected 417-A in a Colpitts circuit.

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1 H. Wallman, A. B. Macnee, and C. P. Gadsden, "A Low Noise Amplifier", Proceedings of the IRE, page 700-708, June 6, 1948.

2 This is defined as the quotient of the IF output current and the signal input voltage.

3 S. N. Van Voorhis, Microwave Receivers, McGraw-Hill, page 145, 1948.

The major problem in the design of the oscillator section, especially in the two higher frequency ranges, was to obtain a wide tuning range with sufficient drive to develop the required mixer conversion transconductance over the entire frequency range.

2.1.4 Tracking and Alignment. Single voltage tracking was accomplished by feeding the same voltage to all tuned stages. If the capacitance of each tuned stage follows the law $C = C_o f(E)$, where $f(E)$, the function of applied voltage, is the same for all voltage sensitive capacitors, tracking can be accomplished by using conventional tracking theory.[1]

The tuned stages of the FE assemblies were aligned with a grid-dip meter. Proper adjustment for three-point tracking was obtained by the application of DC only to the capacitors. Figure 6 is a plot of the oscillator tracking error curve versus tuning frequency. In plotting this curve an average value was taken for the frequency of the signal circuits. However, since the circuits are quite broad-band by virtue of the relatively low Q of the ferroelectric capacitors, no loss in receiver sensitivity due to this effect could be detected. In all three assemblies the oscillator frequency is 20 MC above the signal frequencies.

2.1.5 Antenna Section. It is assumed that a quarter wave whip antenna will be used which can be adjusted to the center of the desired band. The loss due to mismatch in the antenna system across each band is not excessive. The use of voltage step-up and a tuned-input circuit makes possible a lower noise figure than is obtainable with untuned inputs. If the receiver is to be operated in a low-signal area the antenna could be tuned across the band by means of ferroelectric condensers.

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[1] R. R. Sturley, Radio Receiver Design, Part I, 2nd edition revised, Chapman and Hall, 1953.

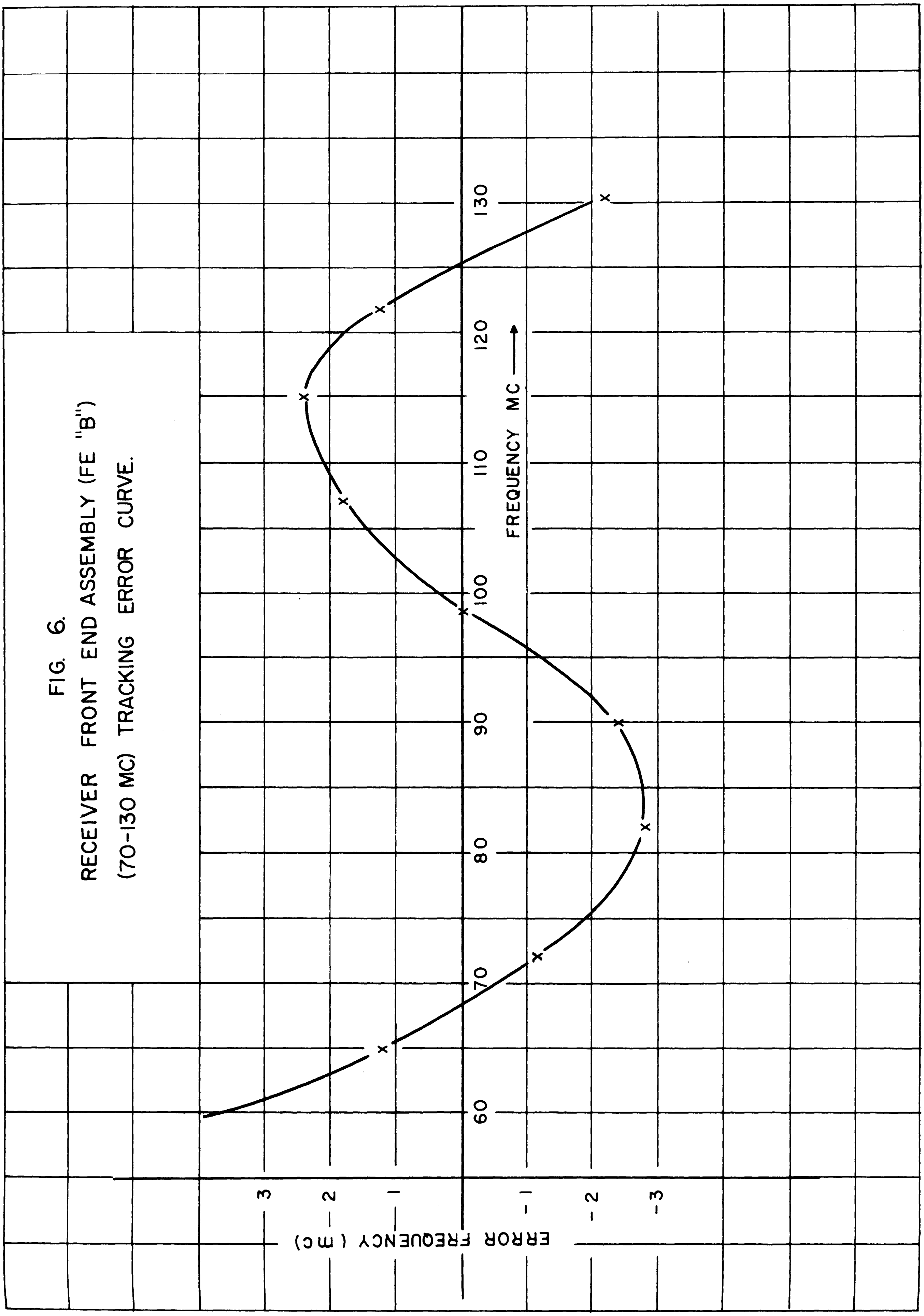


FIG. 6.
RECEIVER FRONT END ASSEMBLY (FE "B")
(70-130 MC) TRACKING ERROR CURVE.

2.2 IF Amplifier

2.2.1 Choice of Intermediate Frequency. To obtain single-signal reception it is necessary to remove all image and harmonic responses, or at least to reduce them to a large extent. Any reasonable amount of pre-selection will be adequate to take care of the harmonic responses. Image rejection is much more difficult to obtain. With ordinary RF amplifiers, rejection increases with increased spacing between the true response and the image. This calls for as high an intermediate frequency as is reasonably attainable. However, the noise figure of the amplifier deteriorates with increasing frequency. Furthermore, the ganging problem tends to increase in complexity when the local oscillator and preselector circuits are widely different in frequency, especially at the lower frequencies. Therefore, the final choice of the intermediate frequency is one of compromise between the desired image rejection ratio, and the desired sensitivity and complexity that may be tolerated in the ganging of the RF circuits.

The actual choice of intermediate frequency from the standpoint of image rejection must be made on the basis of the selectivity of the RF circuits preceeding the IF amplifier. From the RF selectivity curve the desired image spacing for a given image rejection may be chosen, and the intermediate frequency required will be equal to one-half of this spacing times the oscillator harmonic utilized. Since the lowest received frequency is 35 MC, an intermediate frequency of 20 MC was chosen. This particular choice of IF frequency will give the desired image rejection and still keep the IF out of the receiver's passband.

2.2.2 Bandwidth and Selectivity Characteristics. The bandwidth and selectivity characteristics of the IF amplifier are determined by several considerations. First, with regard to bandwidth, the study of the response of a linear resonant system to a sinusoidal driving function having a linear variation of frequency with time (as in a panoramic superheterodyne receiver)[1], dictates that $S/B^2 < 1$, where S = sweep rate of signal (radians/second^2) and B = bandwidth of filter (radians/second). The minimum bandwidth for this receiver is at least 32 KC.

Band	Frequency Range	No. of MC Swept Out
Band A	(35-70 MC)	35 MC
Band B	(70-130 MC)	60 MC
Band C	(130-200 MC)	70 MC

Second, the band must be sufficiently wide so that change of the local oscillator frequency with time under operating conditions will not cause the IF output of the mixer to drift out of the acceptance band of the IF amplifier. The smallest bandwidth consistent with these considerations should be used, since the noise figure of the amplifier deteriorates with increasing bandwidth.

The selectivity desired is influenced by three factors. These are: (1) bandwidth; (2) rejection of frequencies outside the passband; and (3) transient response.

Since wide ranges in the input signal are often encountered, it is important that the off-band rejection, commonly called skirt selectivity, be as high as possible in order to prevent strong signals close to the passband from producing a response in the receiver. From a transient consideration, the rounded

[1] H. W. Batten, "The Response of a Panoramic Receiver to CW and Pulse Signals", Technical Report No. 3, Task EDG-3, Engineering Research Institute, University of Michigan, June 1952.

selectivity curve of synchronous single-tuned circuits[1] is desirable. Thus a compromise must be made between the desired transient response, skirt selectivity required, and gain-bandwidth products available with various types of coupling. From the above considerations it was decided that synchronous, single-tuned amplifier stages would be used with an overall bandwidth of 130 KC.

2.2.3 IF Amplifier Gain. It was desired to provide enough gain in the IF strip to raise the IF output signal to a level sufficient to produce linear operation of the second detector. The minimum level required for operation is approximately 0.5 to 1.0 volt. This also is a convenient input level for various types of presentation circuits.

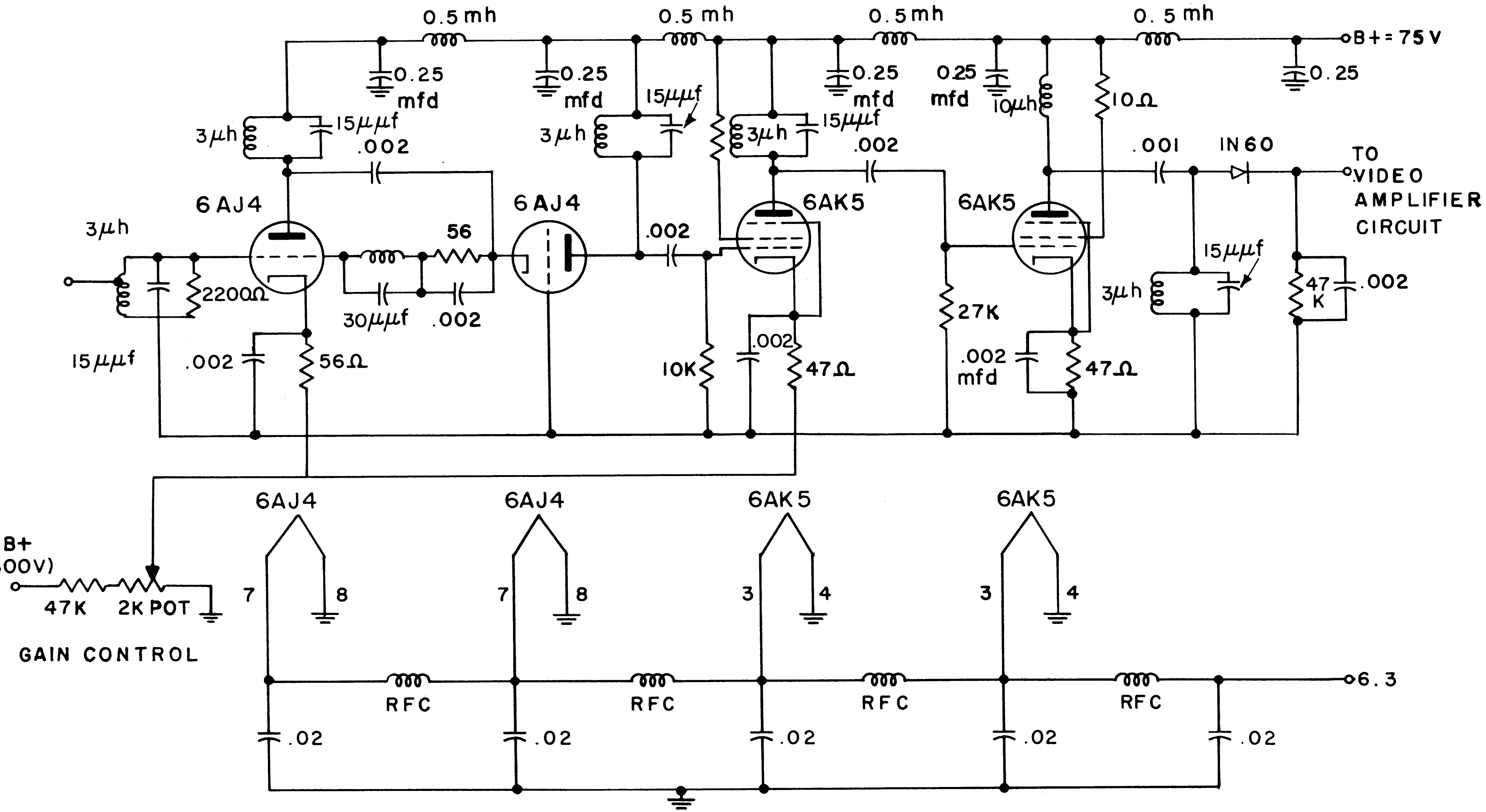
2.3 Second Detector

A crystal diode was selected in preference to a vacuum tube as the second detector for the following reasons. (1) The diode conductance in the forward direction is considerably higher than that of any tube. (2) Interelectrode capacitance and capacity to ground are very low (around .5 to 1 μμf) in the diode, and this property is helpful in both the IF and video circuit. (3) The diode requires no heater power and contributes no hum. (4) The diode is very efficient at small input voltages since it is easy to reach the linear region at a low voltage.

The IF amplifier and second detector consists of a cascode input stage followed by two pentode amplifiers and a germanium crystal diode (Figure 7). The 50-Ω input stage was designed around the cascode circuit to insure that the contribution of noise from the IF amplifier to the overall receiver system was negligible.

To facilitate maintenance and repair the amplifier-detector was built as a plug-in unit. Photographs of a completed unit are shown in Figures 8 and 9. Considerable care was taken, both in design and construction, to build a stable amplifier operating from a single 75 volt source.

[1] The term synchronous tuning is used to designate a series of circuits tuned to the same frequency.



f = 20 mc
B + = 75 VOLTS
Δf = 133 KC (3 db DOWN)
NOISE FIGURE = 3.5 db

FIG. 7. I. F. STRIP AND 2ND DETECTOR

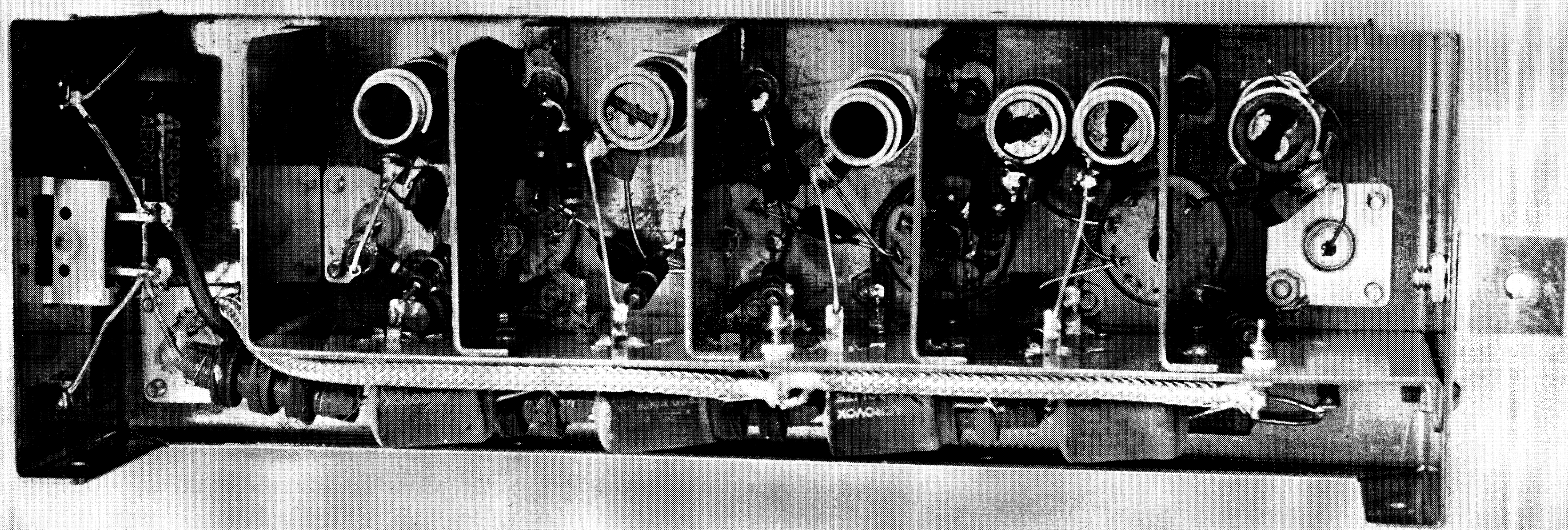


FIG. 8. INTERIOR VIEW OF 20 MC I.F. STRIP.

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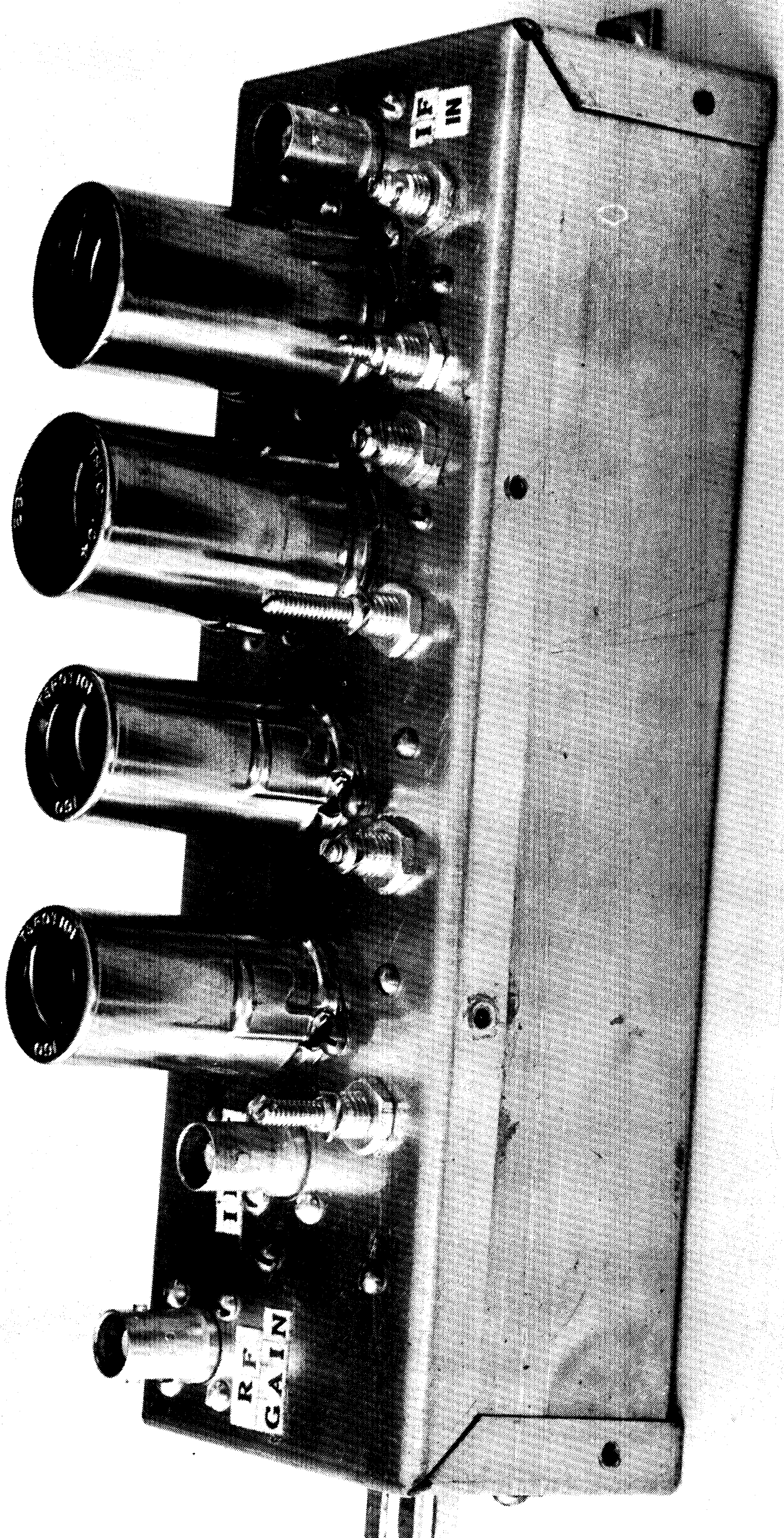

FIG. 9. EXTERIOR VIEW OF 20 MC I.F STRIP.

The unit was aligned in the usual manner. The voltage gain was measured by applying an unmodulated signal at the input and measuring the deflection at the output of the detector with an oscilloscope. Noise figure measurements were made as described in Appendix B. Figure 10 is a plot of the passband of the IF amplifier. The 3 db bandwidth was measured and found to be 133 KC. As shown in Figure 11, which is a plot of the response curve of the unit, the voltage gain is about 95 db, but it can be boosted over 100 db by simply raising the supply voltage. The unit has a noise figure of approximately 3.5 db.

2.4 Tuning-Voltage Shaping Circuit

To obtain a linear frequency scale on the display scope the shape of either the tuning voltage or of the voltage applied to the horizontal sweep may be altered. Since it is desirable to have a linear frequency change with time, thus obtaining a uniform response rate, a linear sawtooth voltage is used to drive the horizontal sweep of the display scope and the tuning voltage is shaped.

2.4.1 Approximating the Driving Function f(E). Although the function f(E) in the equation $C = C_o \cdot f(E)$ may be approximated by a mathematical function, it is of no particular advantage in a practical design. A graphical method is therefore used and the steps of this are briefly outlined below.

First, the capacity variation of the voltage sensitive capacitors is plotted as a function of applied voltage, E. A typical plot is shown in Figure 12a. In the particular circuit to be tuned, the capacitance of the coil, associated tube, and wiring is lumped together as C_s. This value is added to C and the capacity variation is then replotted.

The second step in the graphical method is shown in Figure 12b. This is a plot of the frequency variation, and is obtained by plotting $K(C + C_s)^{-1/2}$ versus E, where the constant K is equal to $1/2\pi \, L^{-1/2}$.

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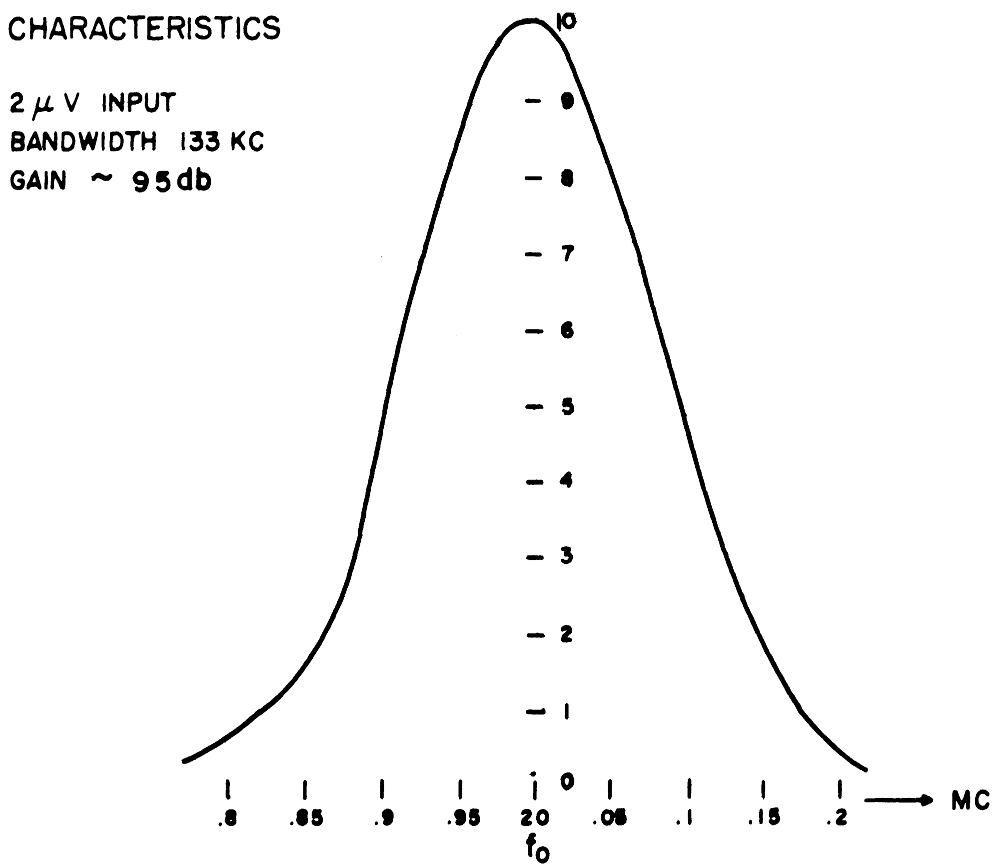

FIG. 10. PASS BAND OF 20 MC I.F. AMPLIFIER.

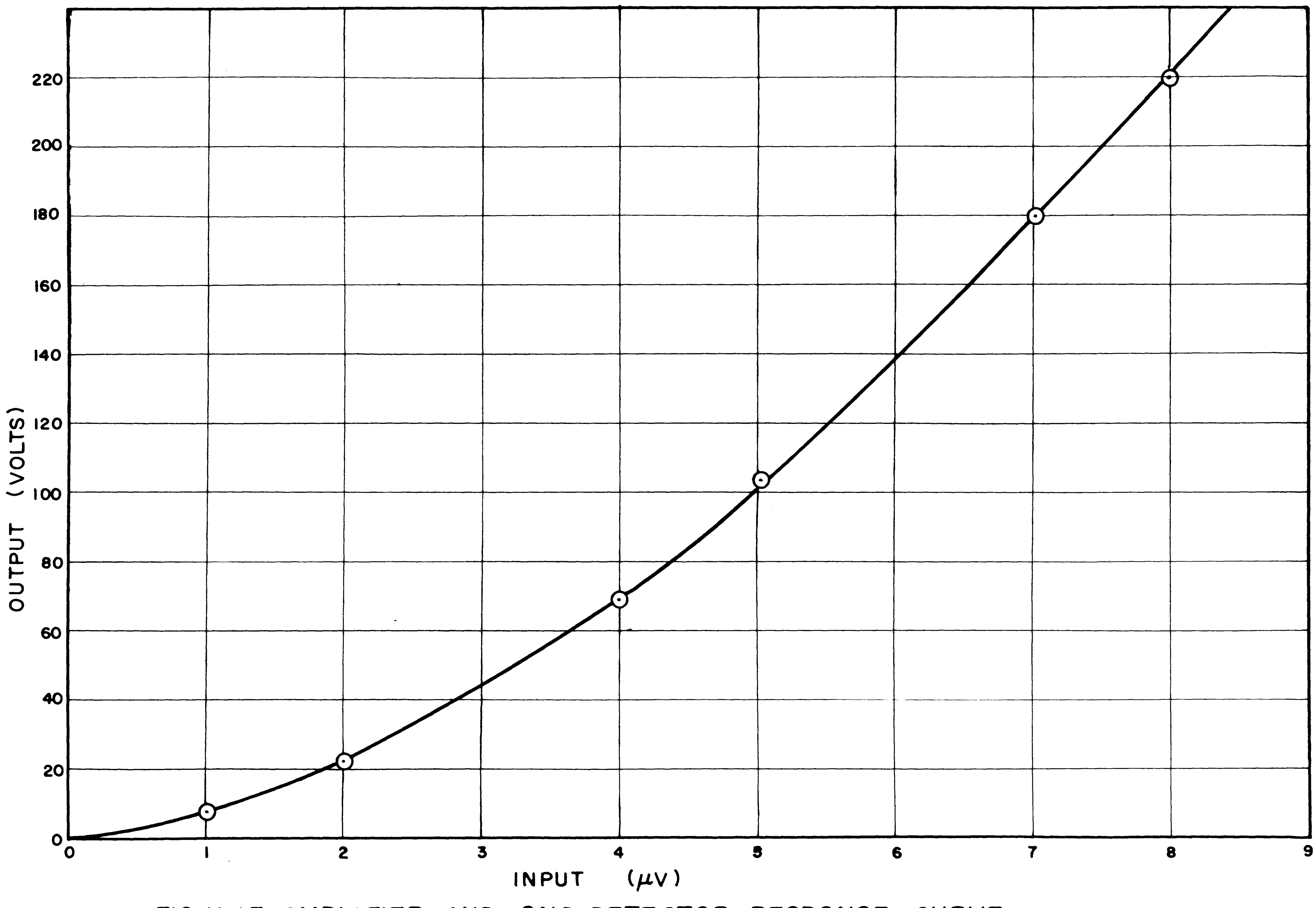

FIG. 11. I.F. AMPLIFIER AND 2ND DETECTOR RESPONSE CURVE

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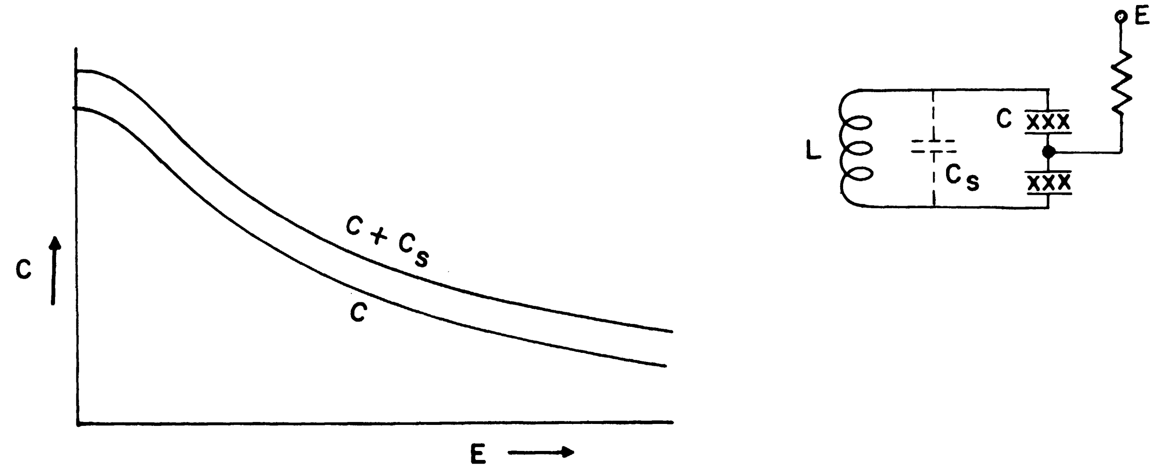
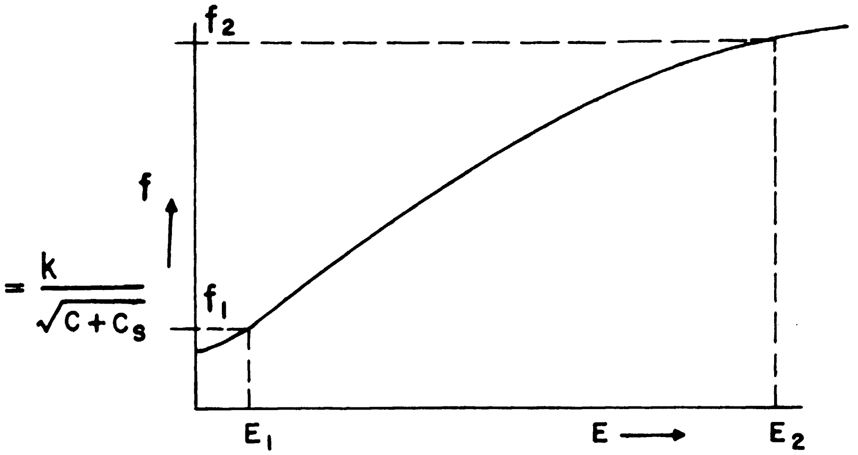
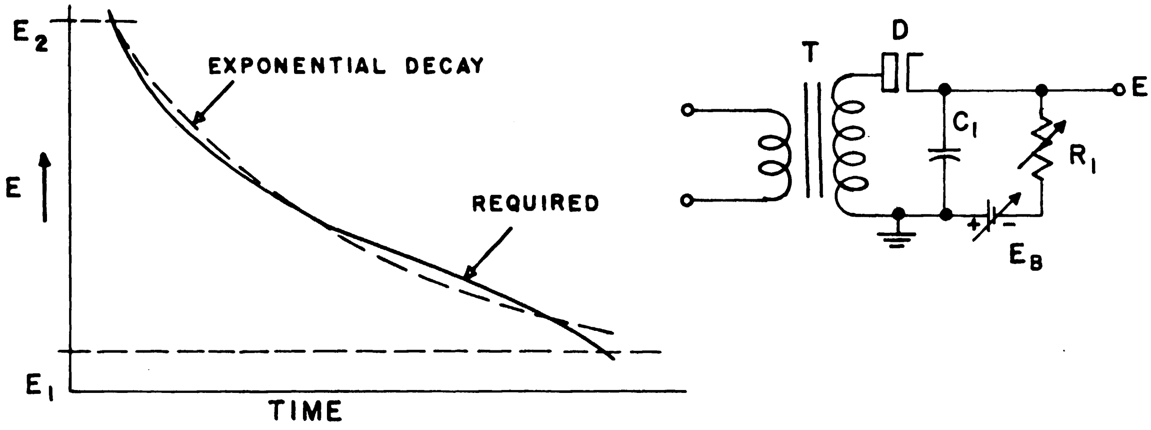


A. CAPACITY VARIATION



B. FREQUENCY VARIATION



C. SWEEP VOLTAGE FOR LINEAR FREQUENCY

FIG. 12.

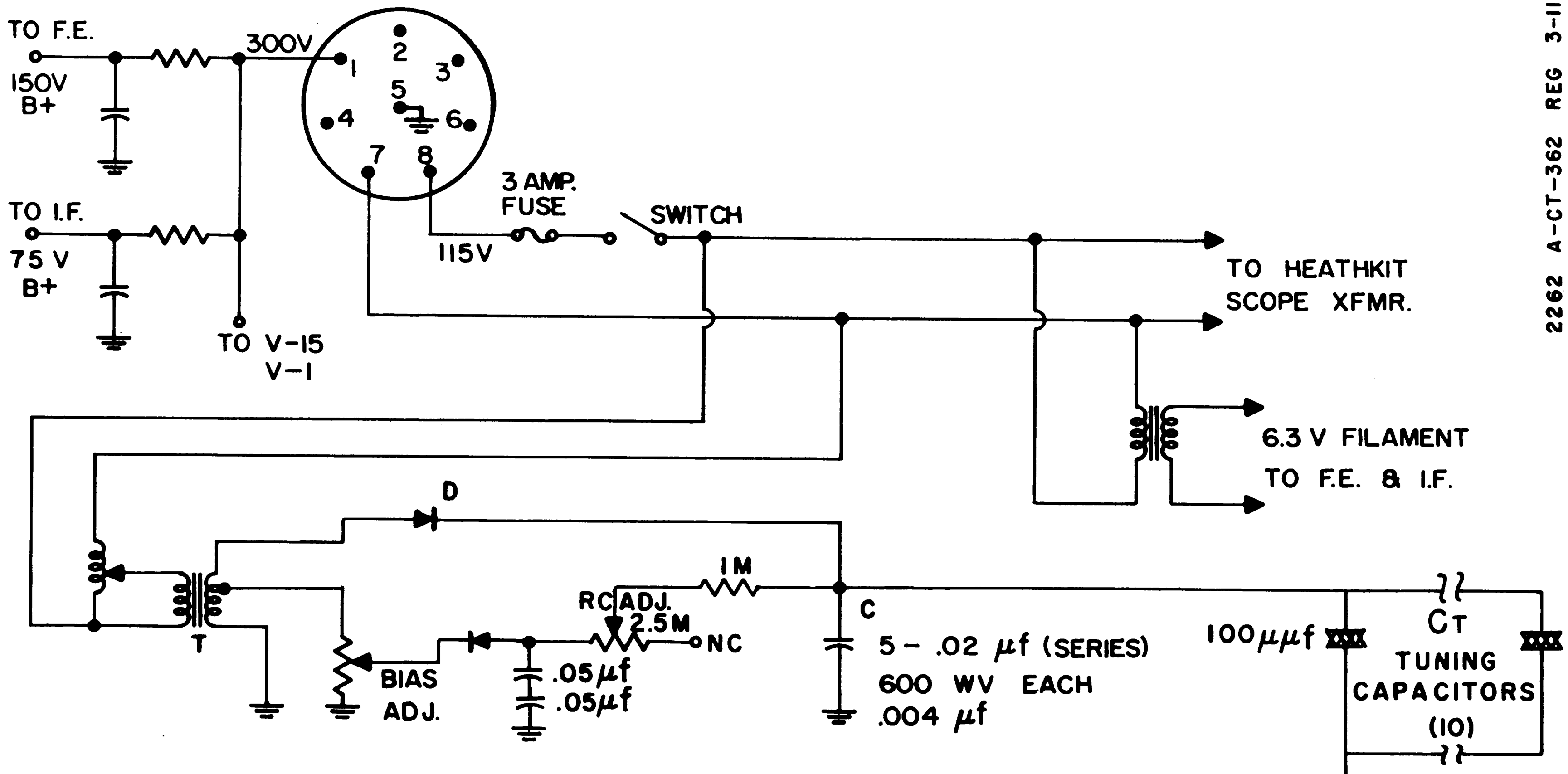
LINEARIZING THE FREQUENCY SWEEP

It is now necessary to decide whether to sweep from high to low frequency or from low to high frequency. It is obvious that in sweeping from high to low the voltage must change rapidly at first and slowly later. This suggests an electrical circuit with a negative exponent, which is readily realizable. There is an additional advantage in sweeping from the high frequency end to the low frequency; that is, the flyback, or return sweep, is generally much faster than the main sweep, which allows considerably more time for discharge of the capacitors than for charging. Previous investigation[1] shows that the polarization lag, or response of the capacitor to changes in applied voltage, is more rapid on charge than on discharge.

The frequency limits f_2 and f_1 are noted on Figure 12b, and from these the required limits of voltage E_2 and E_1 are obtained. To obtain the required shape of voltage wave for linear frequency sweep this curve is replotted in Figure 12c. Hence, equal frequency intervals are replaced by equal time intervals along the baseline, and the solid curve shows the required voltage variation. It is generally possible to approximate this curve fairly closely by a single RC decay circuit as suggested by the simplified circuit to the right of the curve. The time constant, R_1C_1, and the bias battery, E_b, may be adjusted to give the approximation indicated by the dashed curve.

2.4.2 Circuit Design. The circuit of Figure 13 is used to shape the tuning voltage to give a linear frequency change with time. The transformer "T" and selenium rectifier stack "D", connected as a half wave rectifier, serve to charge capacitors $C + C_T$ (all in parallel) to a peak voltage of 2000 volts once each cycle. This voltage in turn is allowed to discharge across the tuning capacitors with a time constant as shown in Figure 14. The operation of this

1. "Interim Report on Ferroelectric Materials and Their Application," L. W. Orr and H. Diamond, Electronic Defense Group Technical Report No. 31, The University of Michigan, Engineering Research Institute, Ann Arbor, Michigan.



1	+300 RED	5	GRND. BLACK
2	NC GREEN	6	-50 WHITE
3	-300 YELLOW	7	115 V PURPLE
4	+50 ORANGE	8	115 V BROWN

FIG. 13. H. V. SWEEP CIRCUIT

circuit is best illustrated by the sketch of Figure 14. E_1 is a rectified 60 cps sine wave voltage, and E_c is the shaped tuning voltage. Both ends of the shaped tuning voltage curve, E_c, may be adjusted to give the required approximation. The "cut out" point can be adjusted by means of the variac while the "cut in" point can be adjusted by means of the bias adjust potentiometer. It is interesting to note that sweeping during 244° with the remaining 116° used as fly-back time represents a gain of 36% in receiver on-time when compared with 60 cycle sweep.

2.4.3 Comparison of Sine Wave vs Shaped Wave Sweep. Figure 15 is a comparison of the frequency response with time when sweeping the range 30-50 MC with (1) a sine wave, and (2) with shaped wave. In both cases the marker pips are spaced at 5 MC intervals from 30 MC. Note that in Figure 15a the lower frequencies are squeezed together and the higher frequencies are spread out, while in Figure 15b the marker pips are evenly distributed in time across the display scope.

2.4.4 Power Required for Sweep Circuit. The voltage across the tuning capacitors is approximately 2000 volts peak. Although this may be considered high, the actual power consumed in the charging circuit is almost completely reactive and is practically negligible. The high voltage transformer (Figure 13) is rated at 50 VA which is much larger than required. The capacitor (Figure 13) should be rated at nearly twice the value of the peak charging voltage to avoid being damaged under conditions of full electronic tuning.

2.5 Display Circuits

The panoramic display section of the receiver utilizes a 5" CRT and features a variable-position notch sweep and a moveable step-type marker. The notch, or expanded sweep, is useful in separating signals which are close in frequency, while the step-type marker is useful in determining the frequency of displayed signals. The display control switch on the front panel allows selection

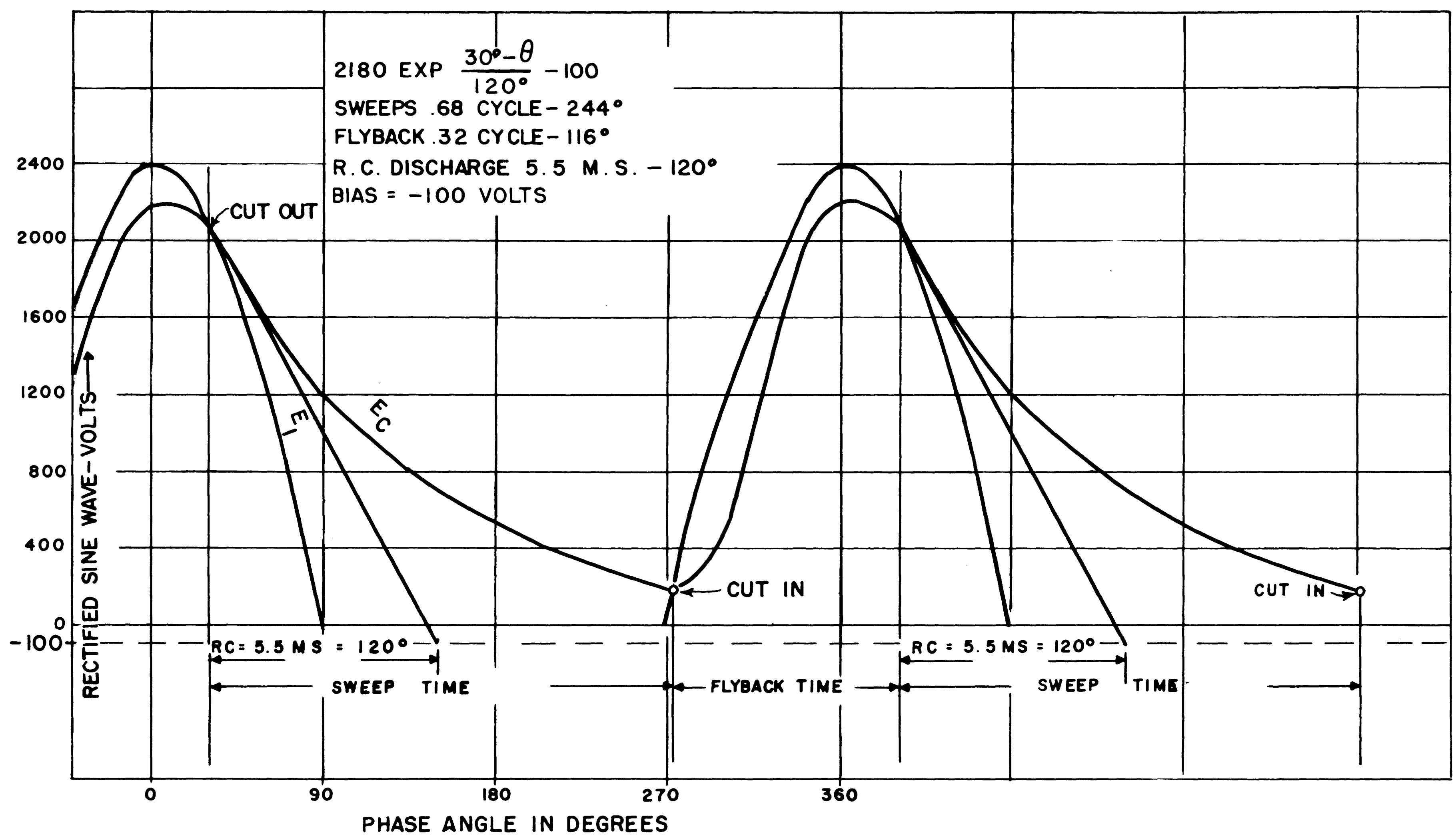


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FIG. 14. SHAPED TUNING VOLTAGE FOR FE UNITS OF PETR PAN

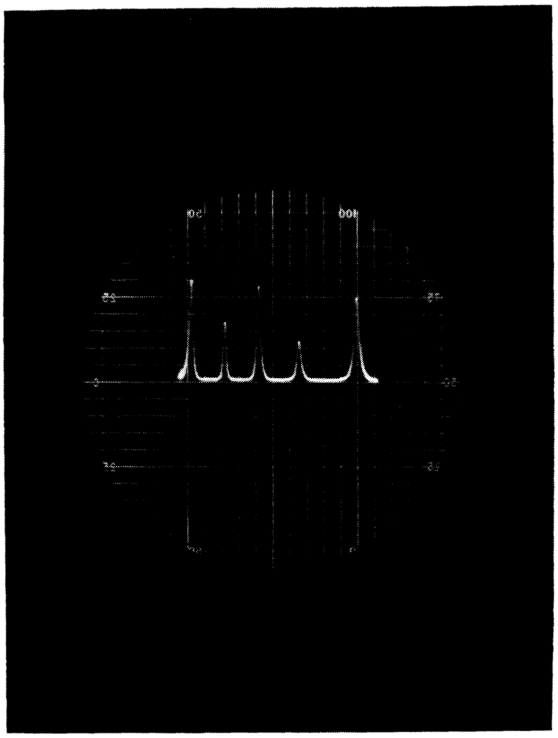


FIG. 15. (a) -SINE WAVE SWEEP-MARKER PIPS SPACED AT 5 MC INTERVALS IN THE FREQUENCY RANGE 30-50 MC

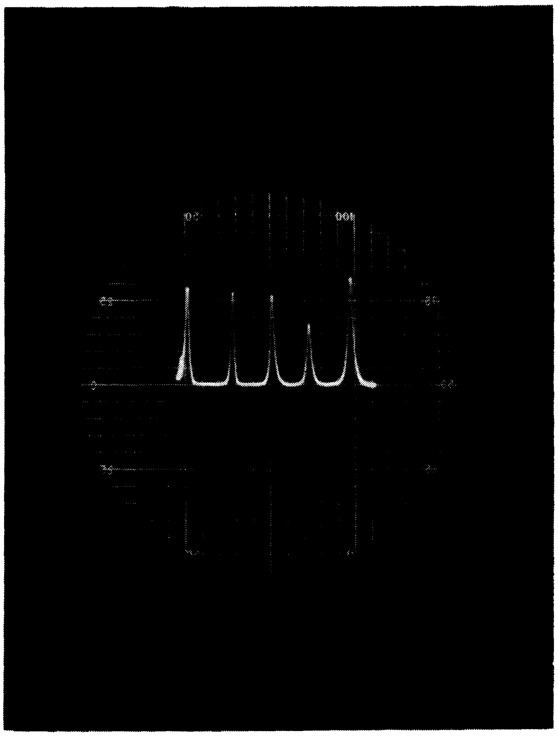


FIG. 15. (b) -SHAPED SWEEP- MARKER PIPS SPACED AT 5 MC INTERVALS IN THE FREQUENCY RANGE 30 -50 MC

of any one of three displays: (1) linear sweep; (2) linear sweep with notch (variable in position); or (3) notch which begins on a step in the sweep (variable in position). Typical operation of the display circuit is illustrated by Figure 16.

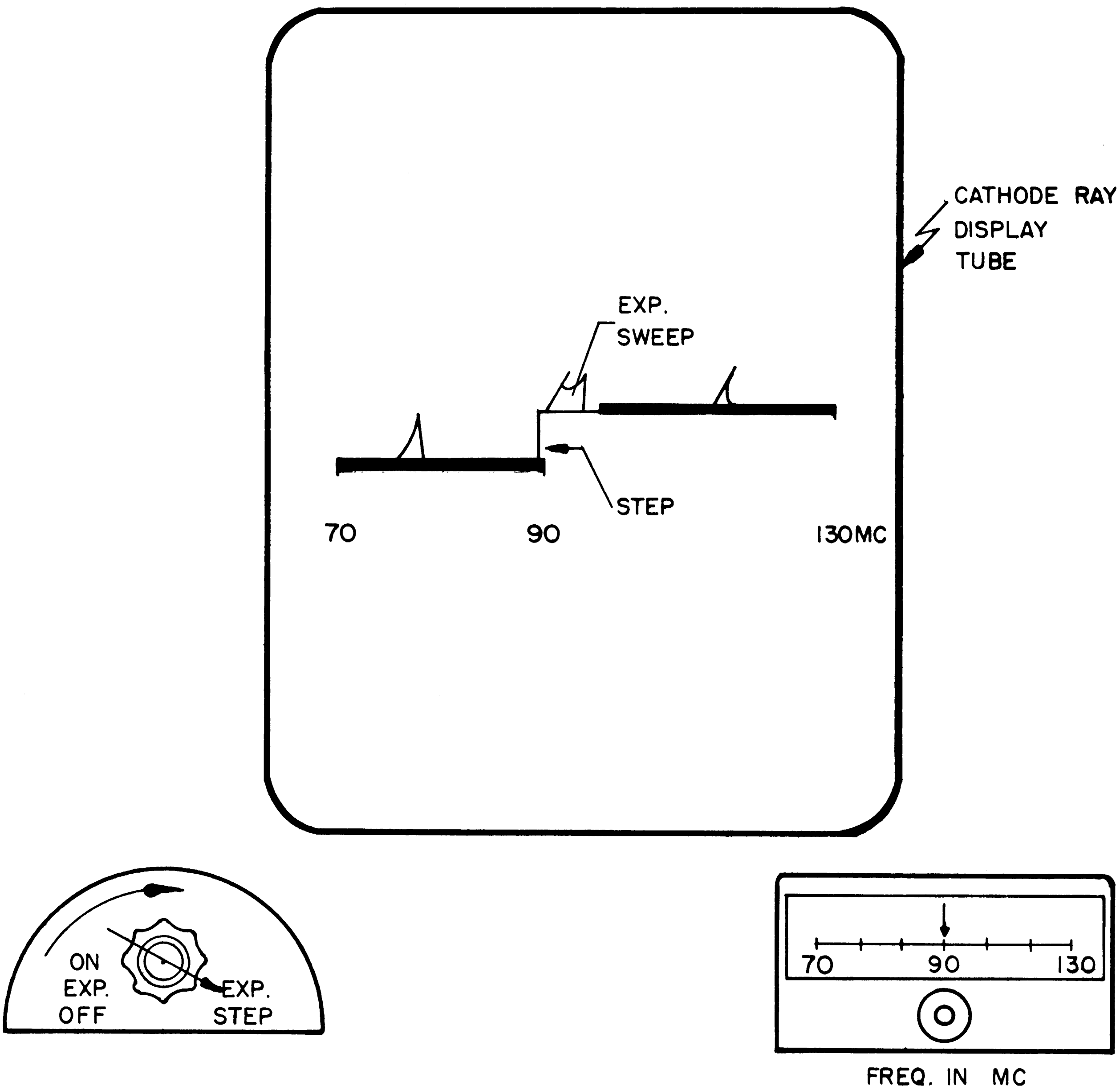
The block diagram of Figure 17 indicates in a general way the overall functioning of the panoramic display.

2.5.1 Vertical Amplifiers. As shown in Figure 17 the output of the 2nd detector is fed to the vertical amplifiers which consist of a video amplifier, a cathode follower stage, a single-ended voltage amplifier, and a balanced output stage whose load is the vertical deflection plates of a 5" CRT. Vertical gain and vertical centering controls are brought out to the front panel.

2.5.2 Horizontal Sweep and Amplifier Circuits. A linear time base is obtained in the following manner. As shown in Figure 18, the 60 cps line voltage is stepped up and clipped. The resulting semi-square wave is differentiated and the positive-going spike is discriminated against. The negative spike is fed to pin No. 2 of V-3, amplified and fed to pin No. 7 of V-3. This positive spike results in a surge of current which charges the cathode capacitor of V-3. The charge on the cathode capacitor is sufficient to cutoff this section of the tube through the long RC discharge of the cathode network. A sawtooth is taken from pin No. 8 of V-3, amplified and fed to the horizontal plates of the CRT in push-pull fashion.

2.5.3 Notch (Expanded Sweep) Circuit. Since the receiver is to be used to sweep over wide frequency ranges it was felt that a variable position notch sweep, which would spread the display in any given interval so that it would be easy to separate and examine two signals very close in frequency, would be useful (see Figure 19).

The notch is about .7 ms long, thus expanding the display so that 1/16 of the frequency band is displayed on 1/4 of the sweep width and 15/16 of the



THE ABOVE CONTROL OPERATES AS FOLLOWS:

(a) "OFF" STRAIGHT SWEEP

(b) "ON" EXP. STRAIGHT SWEEP WITH EXP. ANSION

(c) "EXP." STEP STEP AT BEGINNING OF EXPANDED SWEEP — VARIABLE IN AMPLITUDE

FIG. 16. PETR PAN DISPLAY

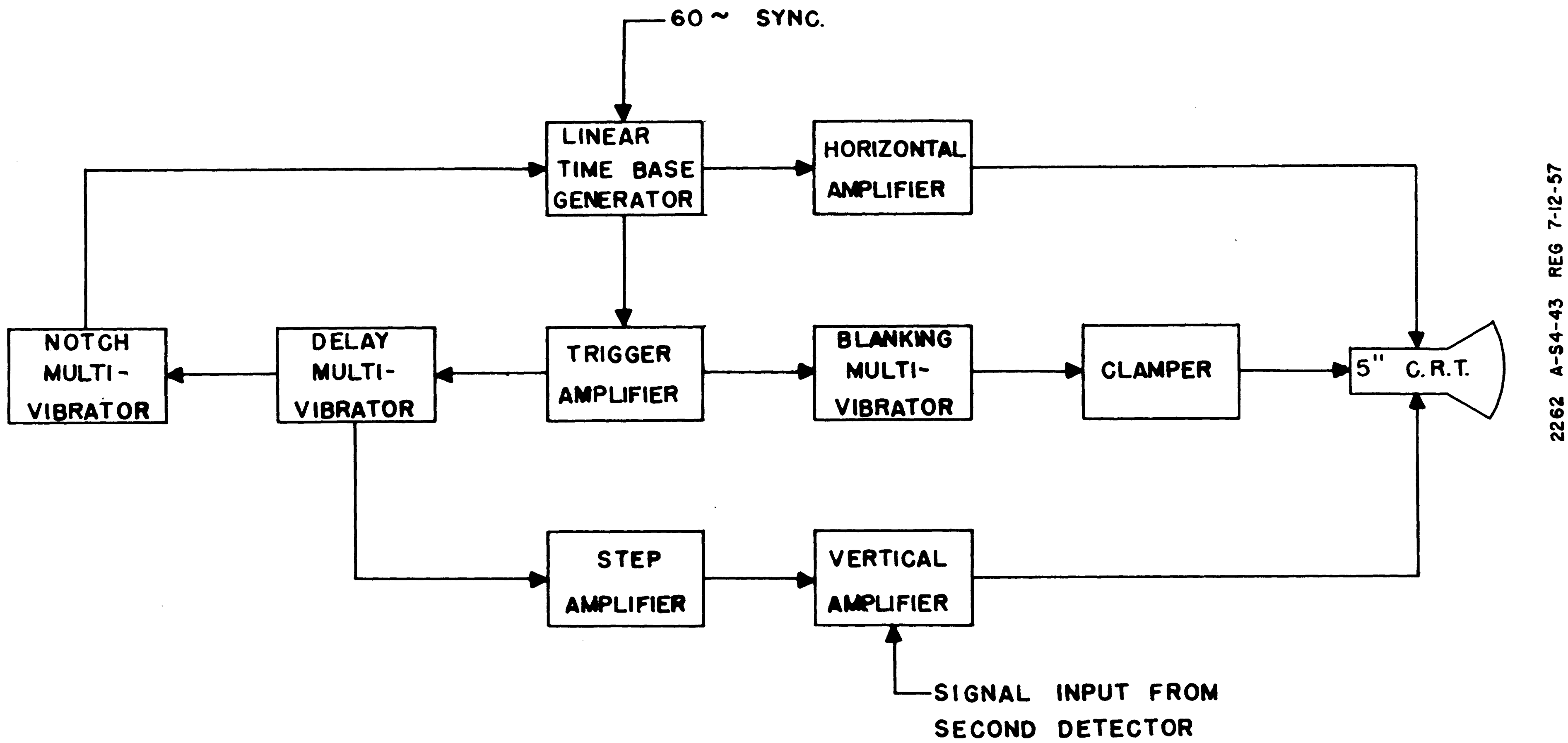


FIG. 17. BLOCK DIAGRAM
DISPLAY CIRCUITS
PETR PAN RECEIVER

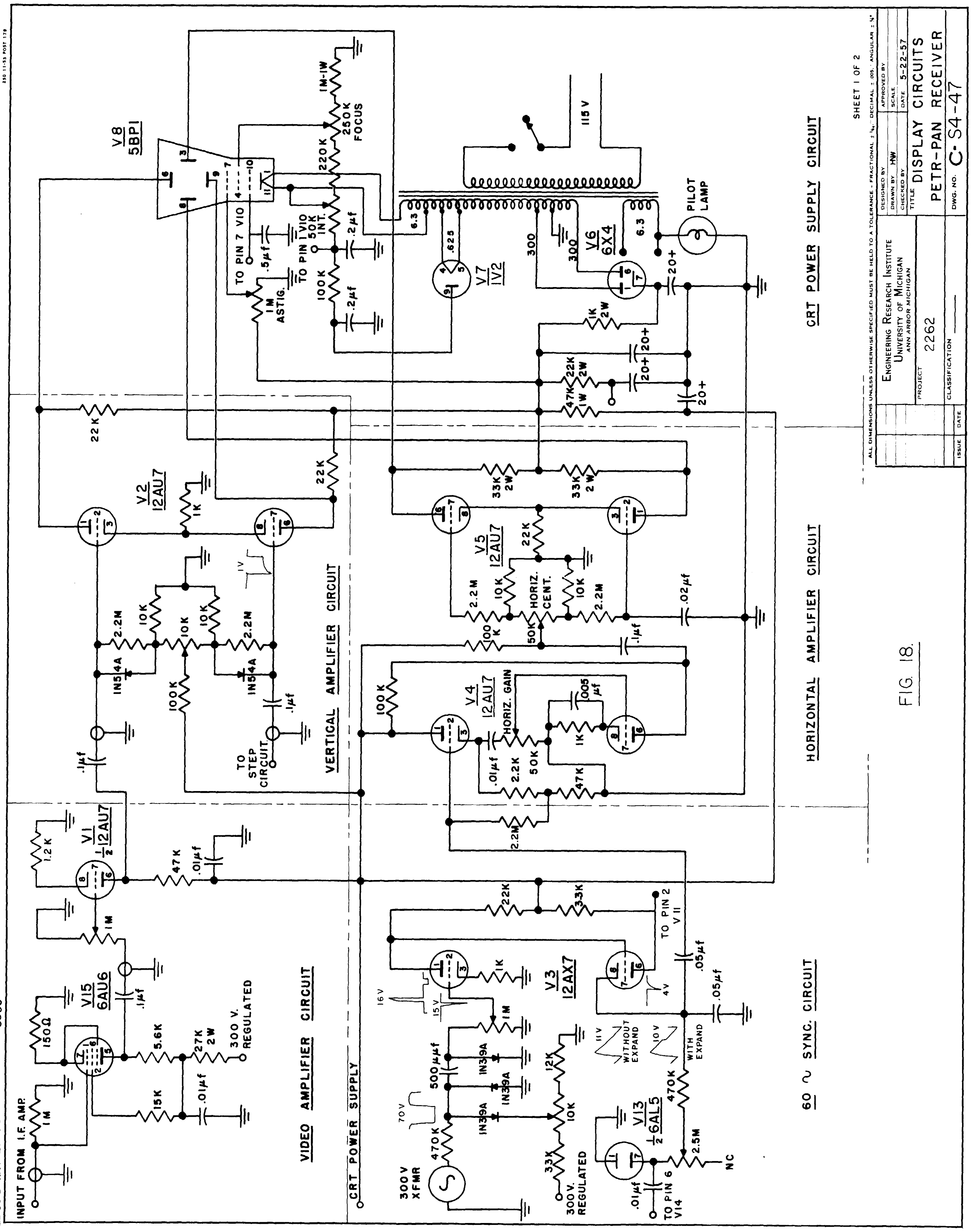
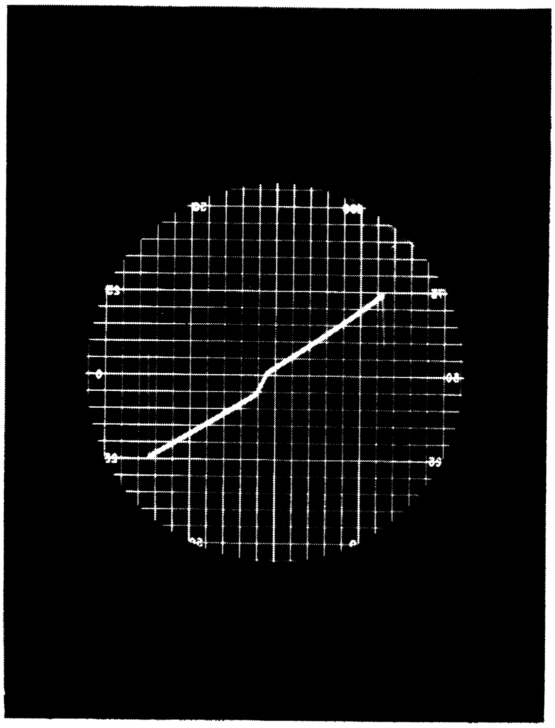
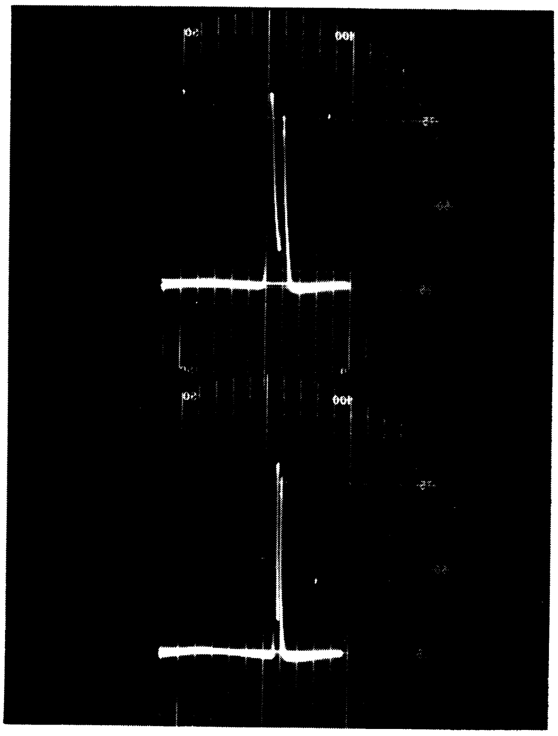

VIDEO AMPLIFIER CIRCUIT
VERTICAL AMPLIFIER CIRCUIT
HORIZONTAL AMPLIFIER CIRCUIT
CRT POWER SUPPLY CIRCUIT
CRT POWER SUPPLY
60 ~ SYNC. CIRCUIT
INPUT FROM I.F. AMP.
V15 6AU6
V1 1/2 12AU7
V2 12AU7
V3 12AX7
V4 12AU7
V5 12AU7
V6 6X4
V7 1V2
V8 5BP1
V13 1/2 6AL5
300 V. REGULATED
300 V XFMR
TO STEP CIRCUIT
TO PIN 2 V11
TO PIN 6 V14
TO PIN 7 V10
TO PIN 1 V10
HORIZ. GAIN
HORIZ. CENT.
ASTIG.
FOCUS
INT.
PILOT LAMP
115 V
WITHOUT EXPAND
WITH EXPAND
SHEET 1 OF 2
ENGINEERING RESEARCH INSTITUTE
UNIVERSITY OF MICHIGAN
ANN ARBOR MICHIGAN
PROJECT 2262
DATE 5-22-57
DISPLAY CIRCUITS
PETR-PAN RECEIVER
DWG. NO. C-S4-47
2262 DWG. NO. C-S4-47 HW 5-22-57

FIG. 18.



VARIABLE POSITION NOTCH SWEEP



SEPARATION OF TWO SIGNALS VERY CLOSE IN FREQUENCY BY USE OF VARIABLE POSITION NOTCH.

FIG. 19.

frequency band is displayed on the remaining 3/4 of the sweep width. The ratio of sweep speed in the notch to sweep speed outside the notch and the ratio of sweep speed in the notch to normal sweep speed (i.e., expanded sweep off) can be obtained from a consideration of Figure 20. The time required to sweep out the frequency range of any one FE unit, as shown in Figure 14 is $T_3 - T_o$ or about 11 ms. If in the time $T_2 - T_1 = 1/16\ (T_3 - T_o)$ the sweep width is $\frac{L}{4}$, where L = total sweep width, then in the time $15/16\ (T_3 - T_o)$ the sweep width is 3/4L; thus, the ratio of sweep speeds

$$\frac{V_2}{V_1} = \frac{\left(\frac{\Delta L}{\Delta T}\right)_2}{\left(\frac{\Delta L}{\Delta T}\right)_1} = \frac{5}{1}$$

The same type of reasoning shows that the ratio of notch sweep speed, V_2, to normal sweep speed, V_3, $= \frac{4}{1}$.

The delay multivibrator (Figure 21) is a single-shot cathode-coupled circuit. It is fired by a positive trigger on pin No. 2, V-12. The positive time of pin No. 6, V-12 is adjustable by means of the frequency dial on the front panel. The positive square wave from pin No. 6, V-12 is differentiated and fed to pin 5 of V-13 which discriminates against the positive spike. The negative spike which occurs at a selected delayed time is fed to pin No. 2, V-14. The notch (or expand multivibrator) is a plate-coupled one-shot multivibrator with one grid held at a positive potential. Negative output is taken from pin No. 6, V-14 and provides a faster RC discharge time in the cathode network of V-3, the linear time base generator. Thus, using FE-A (35-70 MC) approximately 2 MC can be displayed in the notch. Using FE-B (70-130 MC), approximately 4 MC can be displayed, and using FE-C (130-200 MC) approximately 4.5 MC can be displayed in the notch.

2.5.4 Passive Step-Marker Circuit. A passive marker circuit is designed to provide a vertical step in the base line at the frequency indicated on the calibrated slide-rule dial. To determine the frequency of any signal displayed on

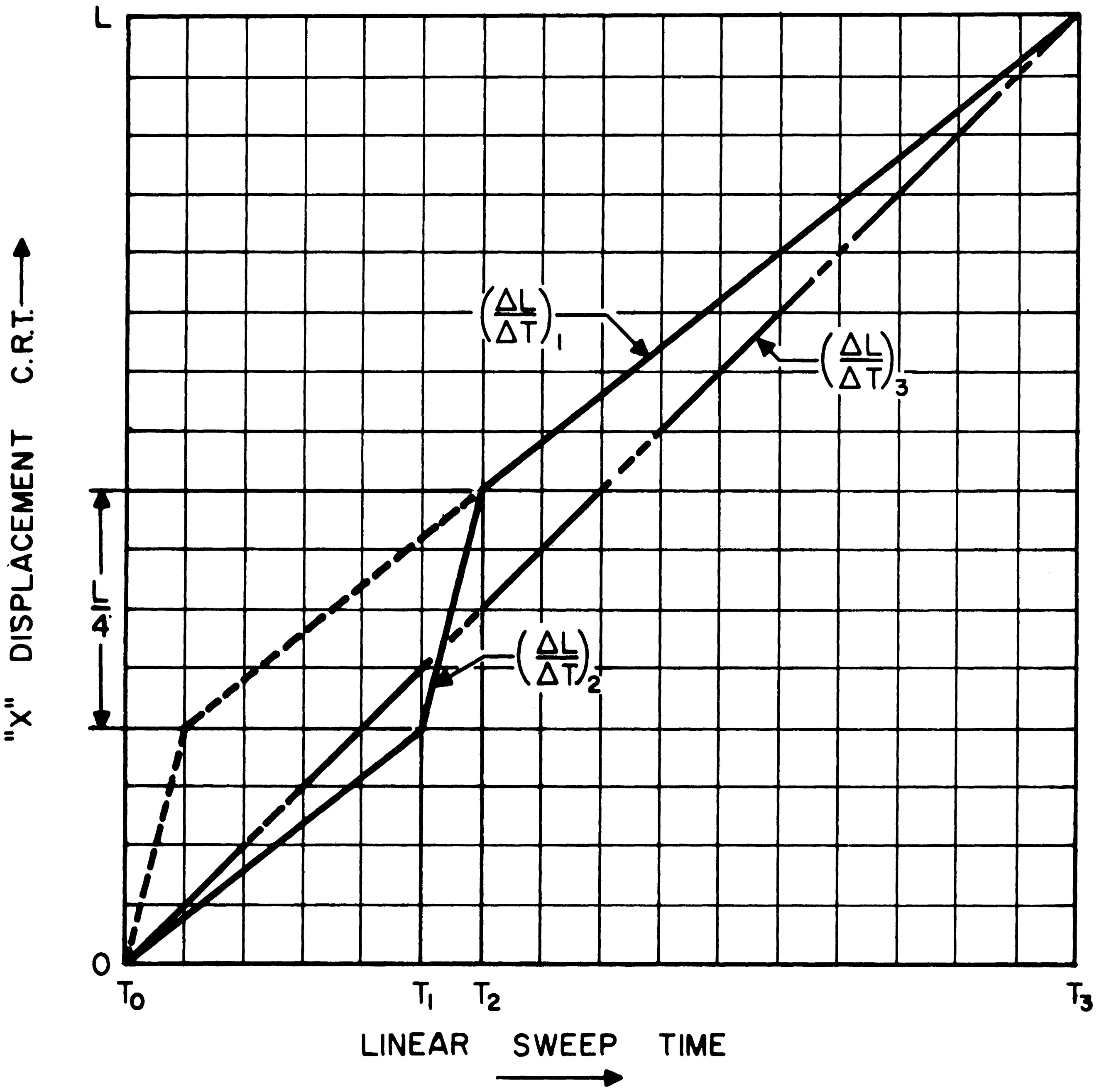


FIG. 20. NOTCH SWEEP DESIGN CONSIDERATION

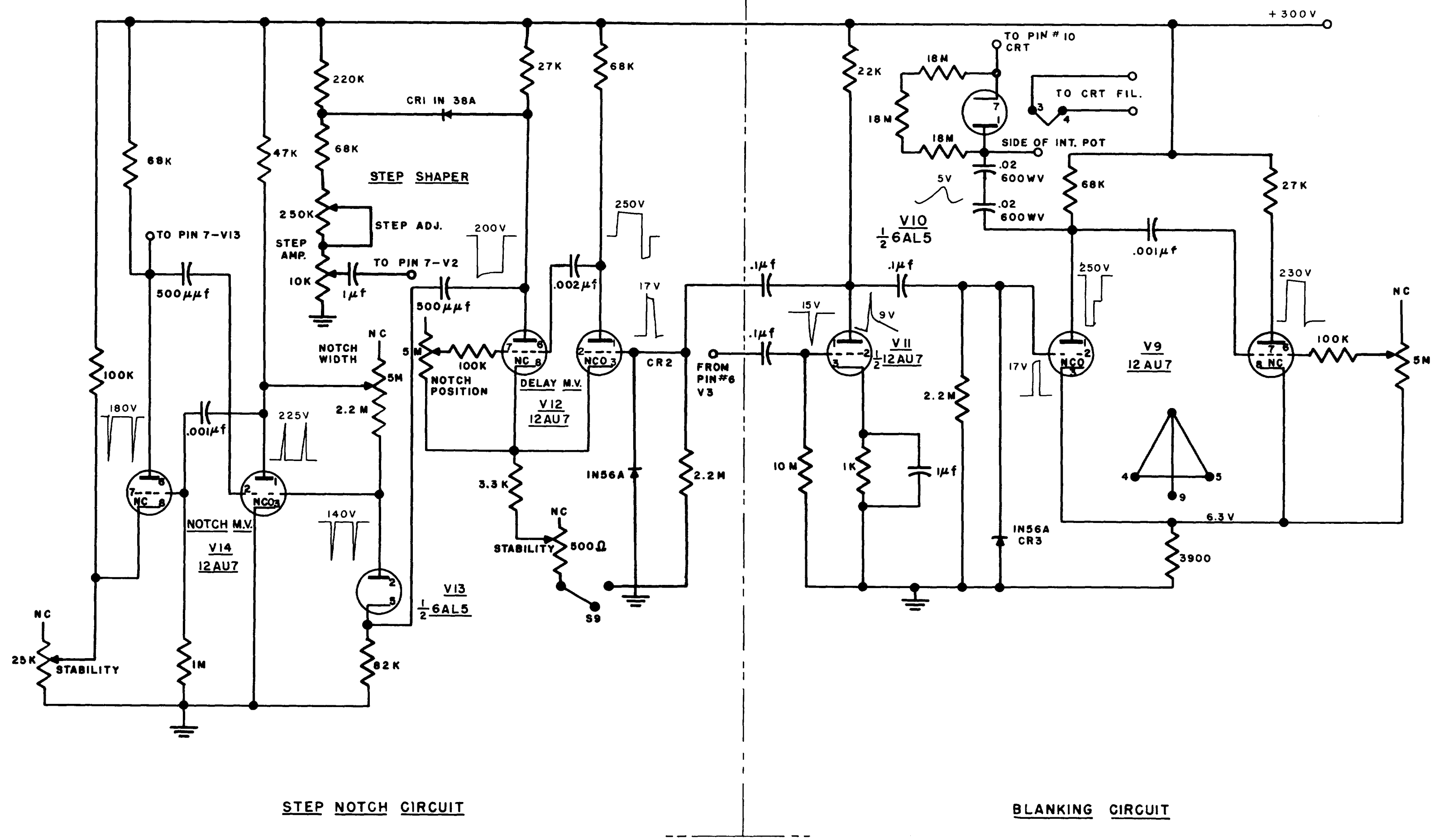


FIG. 21.

SHEET 2 OF 2

ALL DIMENSIONS UNLESS OTHERWISE SPECIFIED MUST BE HELD TO A TOLERANCE - FRACTIONAL ± 1/64" DECIMAL ± .005" ANGULAR ± 1/2°

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PROJECT 2262	TITLE DISPLAY CIRCUITS PETR-PAN RECEIVER
CLASSIFICATION ——	DWG. NO. C-S4-46

2262 DWG. NO. C-S4-46 HNW 5-12-57

the scope it is only necessary to tune the dial knob until the base-line step coincides with the signal spike and read frequency directly from the large slide-rule dial.

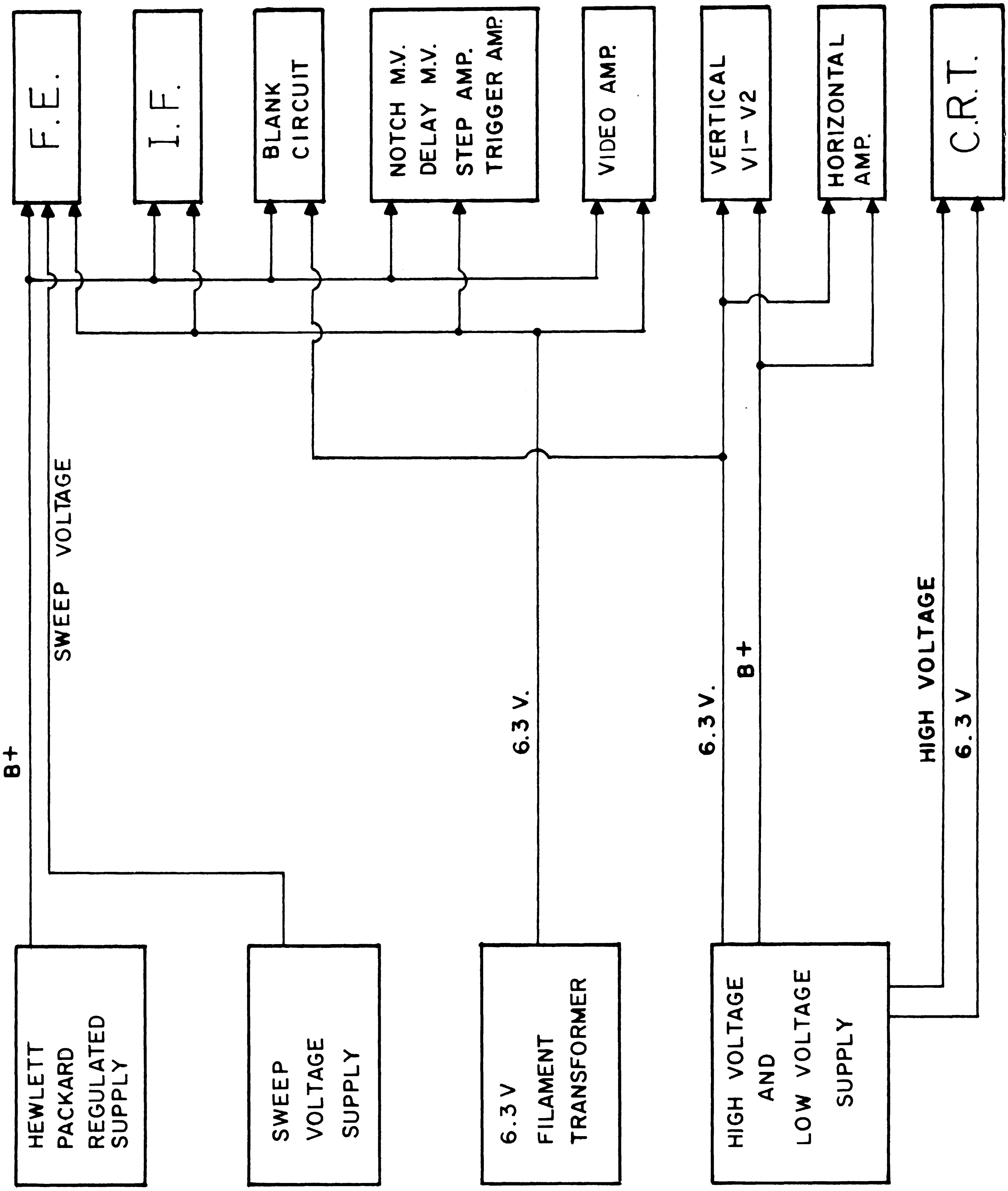
As shown in the schematic of Figure 21 the step shaper is a diode clipping circuit designed to shape the square wave from the delay multivibrator. The step occurs simultaneously with the notch and is applied to the vertical amplifier as indicated. The slide rule dial is coupled mechanically to the 5 megohm notch-position potentiometer in the grid-cathode circuit of the delay MV. The dial is calibrated for each FE unit by applying signals of known frequency to the receiver and marking these locations on the dial.

2.5.5 Blanking Circuit. The blanking circuit of Figure 21 must not only give positive blanking during the sweep multivibrator retrace time but it must also blank the sweep during the flyback time of the shaped tuning voltage (see Figure 14). This eliminates the effect of split-image signals. The split-image effect is caused by the charge and discharge cycle of the shaped tuning voltage impressing equal voltages at slightly different times on the tuning capacitors. Blanking is accomplished by applying a negative output voltage from pin 1 of the single-shot cathode-coupled blanking MV to the clamper diode across the grid-cathode circuit of the CRT. During blank time the coupling capacitors from pin 1 (V_9) to pin 1 (V_{10}), and the high resistance connected across the clamper diode, provide a long time-constant. During the sweep interval the clamper diode conducts through the coupling capacitors to restore the negative DC level to the control grid of the CRT.

2.6 Power Supply

Figure 22 is a block diagram of the power distribution in the receiver. The power supply is made up of three sections.

These are: (1) A low voltage positive supply (Hewlett-Packard 712-A) which provides regulated power for operating the FE units, IF strip, video amplifier,


F.E.
I.F.
BLANK CIRCUIT
NOTCH M.V.
DELAY M.V.
STEP AMP.
TRIGGER AMP.
VIDEO AMP.
VERTICAL V1-V2
HORIZONTAL AMP.
C.R.T.
B+
SWEEP VOLTAGE
6.3 V.
6.3 V.
B+
HIGH VOLTAGE
6.3 V
HEWLETT PACKARD REGULATED SUPPLY
SWEEP VOLTAGE SUPPLY
6.3 V FILAMENT TRANSFORMER
HIGH VOLTAGE AND LOW VOLTAGE SUPPLY

trigger amplifier, notch circuit, marker circuit and the blanking circuit.

(2) A high voltage negative supply provides the potentials necessary for operating the various electrodes of the CRT and furnishes a negative supply to the positioning controls.

(3) A low voltage positive supply provides the power for operating the "X" and "Y" amplifiers, the time base generator, and the positioning circuits of the CRT.

III. CONCLUSIONS

The design of a packaged, electric tuned, panoramic receiver was carried out to acquaint engineers with a method of utilizing the voltage tuning characteristics of ferroelectric capacitors, and to indicate the direction engineers can take in developing and packaging a dielectric-tuned receiver for operational use.

The characteristics of the ceramic capacitors and considerations leading to their use in the tuning elements of the receiver front end assemblies are discussed in Appendix A. Included also is a brief discussion concerning the reliability of tuning elements and an evaluation of the recently developed Michigan Hi-Q material.

The principle points of difference in the design of the three receiver front end assemblies as compared with conventional circuitry has been discussed.

A simple method of shaping the tuning voltage in order to obtain a linear frequency response with time has also been discussed.

Receiver performance tests are described and the results obtained are given in Appendix B.

Receiver alignment and operational procedures are given in Appendix C.

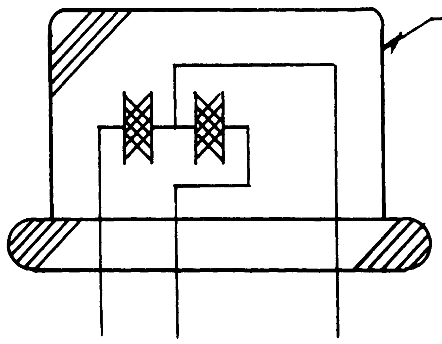
During the course of this investigation several problems were encountered which seemed worthy of investigation.

To facilitate replacement of the tuning elements the sub-miniature capacitors could be placed inside a plastic filled transistor case as shown in Figure 23. In addition, development work should be carried on to compare the suitability of plastic-bead packaging versus vacuum packaging. Since the tuning elements have a greater tendency to break down under conditions of high humidity vacuum packaging would provide a convenient means of moisture proofing.

It was pointed out previously that it is desirable to use titanate materials which have a small temperature coefficient of dielectric constant over a wide temperature range. If the materials do not possess this quality, as in the case of the Michigan Hi-Q material, or if the receiver is to be used over an extended temperature range as it would be under field conditions, then methods of temperature compensation or thermostating techniques will be required to maintain receiver tuning range and sensitivity.

The present receiver incorporates three plug-in front end assemblies and since it takes approximately 45 minutes for each assembly to warm up, frequency determination becomes a very slow procedure. If temperature-sensitive materials are used, future receiver design should include turret construction with all tunable elements thermostated. If materials are available which are not too temperature-sensitive, the plug-in assemblies could be retained and a fence generator which would provide 2.5 or 5 MC marker pips for calibration purposes could be incorporated.

Since the receiver was designed to monitor communication signals, high resolution and probability-of-intercept problems were not considered important, and since a 60 cycle source was available it was used as the basic sweeprate. If rapid sweep rates are to be employed it will be found that the P-E loop will degenerate as the frequency of the driving field is increased. This is due to the finite time of nucleation of the domains and the propagation rate of domain boundaries. As the driving frequency is raised, a point is reached where the extreme value of polar-



SECTION THRU PACKAGED DUAL UNIT
SHOWING ARRANGEMENT OF NON LINEAR
CAPACITORS

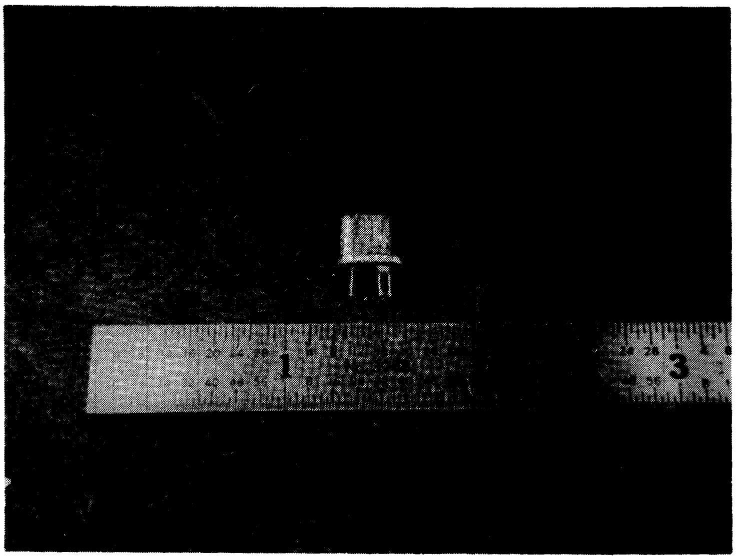


FIG. 23. PHOTO OF PACKAGED DUAL UNIT

ization is not obtained because of insufficient time. Tests indicate that degeneration of the P-E loop becomes evident above sweep rates of 100 KC[1]. Dielectric heating may become a problem at rapid sweep rates but this may be overcome, to a large extent, by keeping the dielectric volume sufficiently small.

It should be noted that the development of the Michigan Hi-Q tuning elements will have a decided effect upon the design of future FE assemblies. Test results indicate that a definite improvement in gain and bandwidth, with the same overall tuning range, may be realized by the use of the Michigan material. Although these improvements will lead to the design of FE assemblies with improved image frequency rejection, greater rejection against undesired signals, greater gain, and improved signal-to-noise ratio, they will also lead to increased tracking problems and greater difficulty in eliminating general FE instability.

It was pointed out previously that the effects of aging are not completely known or understood at the present time. If the tunable element were to undergo a change in Q of 2:1 over a three month period, which is not unreasonable, the receiver characteristics would change drastically during this period also. If, the tunable elements become stable after a certain period of time, however, the aging program could be instituted before the elements were put to use. In any case, more development work has to be done in this area before the effects of aging on the realibility and stability of dielectric tunable elements can be specified.

It is not advisable to continue designing front end assemblies at frequencies much in excess of 250 MC using presently available dielectric materials and lumped constant design techniques since the losses become too great for satis-

1. "Ferromagnetic and Ferroelectric Tuning," L. W. Orr, Electronic Defense Group Technical Report No. 32, The University of Michigan, Engineering Research Institute, Ann Arbor, Michigan.

factory operation of a swept receiver. However, the development of new materials,[1] and the use of transmission line techniques[2] will undoubtedly make possible the design of front end assemblies for use at much higher frequencies.

1

Progress Report No. 17, Task EDG-4, Engineering Research Institute, University of Michigan, page 5, July 1956.

Progress Report No. 18, Task EDG-4, Engineering Research Institute, University of Michigan, page 4, February 1957.

2

Progress Report No. 17, Task EDG-4, Engineering Research Institute, University of Michigan, page 22, July 1956.

Progress Report No. 18, Task EDG-4, Engineering Research Institute, University of Michigan, page 10, February 1957.

APPENDIX A

FERROELECTRIC TUNING ELEMENTS

A.1 Introduction

The ceramic capacitors used in the tuning elements in the receiver front end assemblies are of the barium-strontium titanate class and were developed in the Electrical Engineering Laboratories at the University of Michigan.

To obtain the maximum tuning range with good receiver stability and sensitivity, the capacitor body material should have the following characteristics:

(1) A large change in dielectric constant with applied electric field;

(2) Small variation of dielectric constant with temperature;

(3) Low dielectric loss;

(4) Freedom from breakdown with large DC fields.

Figure 24 shows the small-signal dielectric constant as a function of both temperature and biasing field for a typical commercial ferroelectric ceramic. The data are shown in three dimensions as a convenient means of presenting the temperature field behavior. Note that the material is very temperature-sensitive at low biasing fields and that the peak dielectric constant shifts to higher temperatures as a biasing field is applied. Representation of the capacitor characteristics by means of a surface is very useful in the design of tuning elements as illustrated by the following example.

Consider the design of an RF amplifier circuit which is to tune from 50 to 100 MC. Assuming a value of 135 $\mu\mu f$ for the total tank capacitance at 50 MC gives 32 $\mu\mu f$ for the total capacitance at 100 MC. If the stray capacitance is 10 $\mu\mu f$, the tunable element must vary from 125 to 22 $\mu\mu f$. This element will be a pair of ceramic capacitors connected in series, each having a value of 250 $\mu\mu f$ at zero applied field.

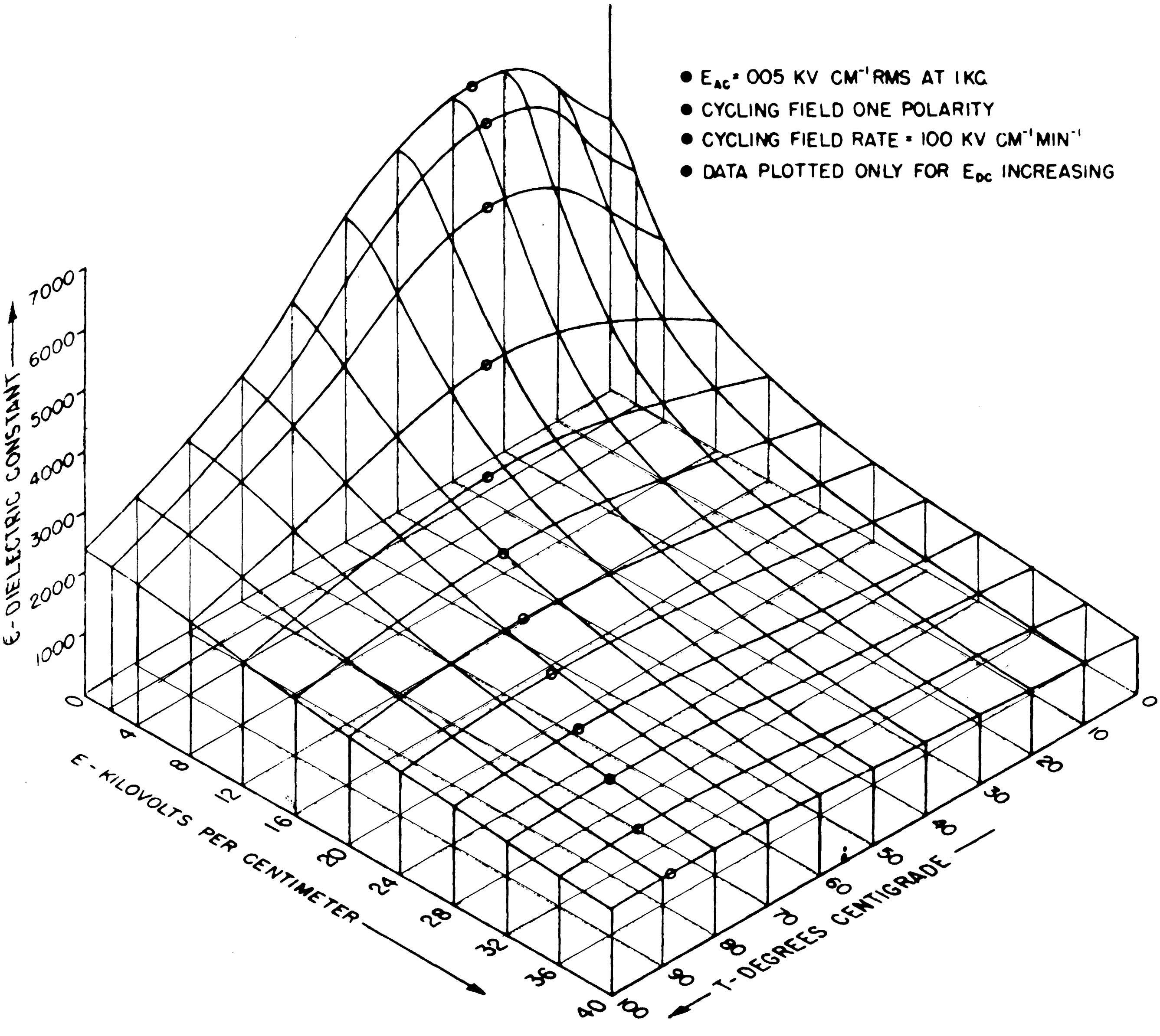


FIG. 24.

ε-T-E SURFACE

FOR AEROVOX "HI-Q" 40

(From "ε-T-E Surfaces of Ferroelectric Ceramics" by L. W. Orr, Electronic Defense Group Technical Report No. 53, The University of Michigan, Engineering Research Institute, Ann Arbor, Michigan.)

The capacitance variation as the electric field is applied may be obtained from the contour curves of Figure 24. Consider the curve for 30°C. It is noted that a 5:1 variation in dielectric constant is obtained when the applied field varies from zero to 30.5 KV/cm. If the material is .02 cm thick the required field is furnished by an applied voltage of 610 volts.

The effect of a temperature change on tuning may be calculated by noting the dielectric constant at 20°C is almost identical with that at 30° with zero applied field. Thus the lower frequency limit will still be 50 mc. At a field of 30.5 KV/cm the dielectric constant is approximately 4% lower than its value at 30°C. The minimum capacitance value at 20°C will therefore be 23 μμf, thus raising the upper frequency level to 101 mc, which is an increase in the tuning range of 2%. Thus, as the receiver warms up from 20°C to 30°C the tuning range will shrink by this amount. For receiver applications where large temperature variations are expected, a material exhibiting a Curie[1] temperature of approximately 60°C should be used as the tuning elements and thermostating techniques should be employed to maintain the tuning elements at this temperature.

Figure 25 shows the Q, which is the inverse of the loss tangent, plotted as a function of applied field and frequency. The frequency range considered here is from 25 to 250 mc, while the biasing field varies from zero to approximately 70 V/mil. In this frequency range, the Q's of the tuning elements increase with an increasing biasing field. As shown in Figure 12b the frequency increases with increasing biasing field. Thus the path traced on the QEF surface in such a case runs diagonally from the low frequency-low field region to the high frequency-high field region. In this manner a reasonable Q(Q=30) is maintained over this range. Low loss is an important factor in electric tuning, and tests show that this is

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[1] The Curie point is defined as the zero field temperature at which the dielectric constant of the material is a maximum.

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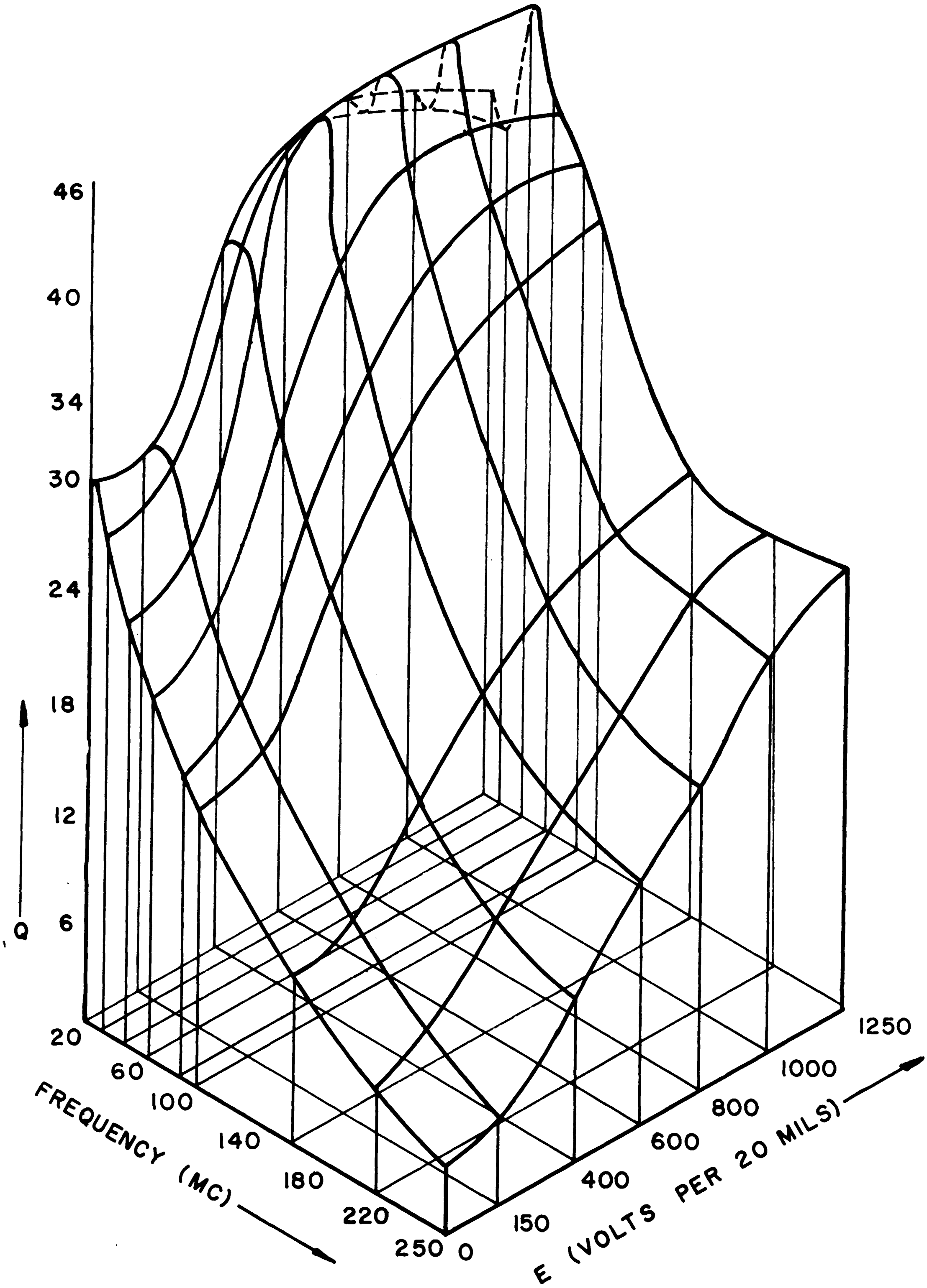

FIG. 25.
Q-E-F SURFACE

not entirely a property of the ceramic material. The type and quality of metal used for electrodes, and its thickness and uniformity, all have important effects on the loss. Capacitors for the tuning units were made with .020" cubes of ceramic having one mil thick electrode plating extending to the edges of the ceramic faces. Figure 26 shows typical capacitors made in this manner.[1] A plastic coating over the dielectric seals out moisture, preventing excessive losses and possible electrical breakdown.

As shown in Figure 27 the resonant frequency of a voltage-tunable tank is controlled by applying a variable DC bias voltage to the junction of the two capacitors with a ground return at one end of the coil. The DC bias voltage defines the operating point, C_o. To electronically sweep the tank in frequency it is only necessary to superimpose an AC signal (e.g., 60 sine wave) which will vary the capacitance about the operating point. It should be noted that the two capacitors are in parallel with the bias and sweep voltages but in series with the RF voltage. This particular arrangement helps to obtain a low RF field across each ferroelectric capacitor and still have the required polarizing field held to reasonable values.

A.2 Reliability of Tuning Elements

The reliable life of tuning capacitors under operating conditions can not be completely evaluated since life-test results are incomplete. However, the present FE assemblies have been operating normally over a period of approximately three months without a failure. The tuning voltage has been approximately 1400 volts, and forced air cooling has been employed.

No effects due to aging in the day-to-day operation of the front end assemblies has been noted. Tests conducted on a long term basis to determine the effects of aging are incomplete at this writing. However, there is some evidence

1. "Miniature Non-Linear Capacitors," H. Diamond, Electronic Defense Group Technical Report No. 54, The University of Michigan, Engineering Research Institute, Ann Arbor, Michigan.

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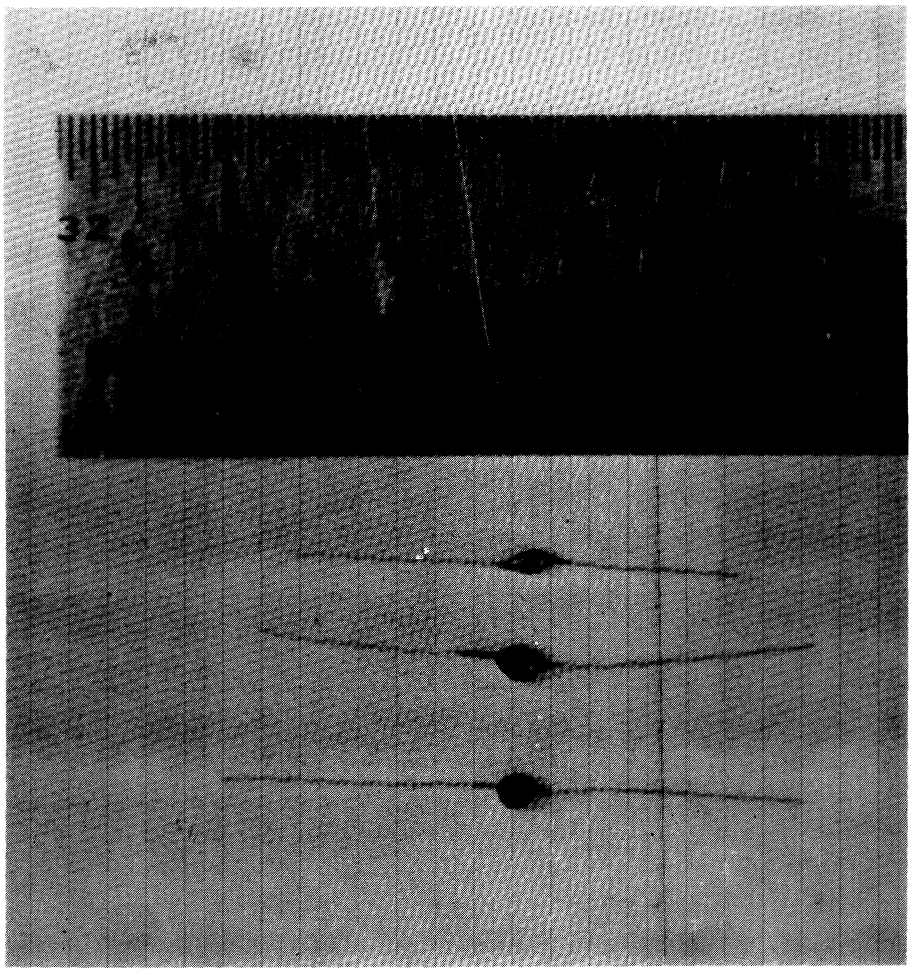


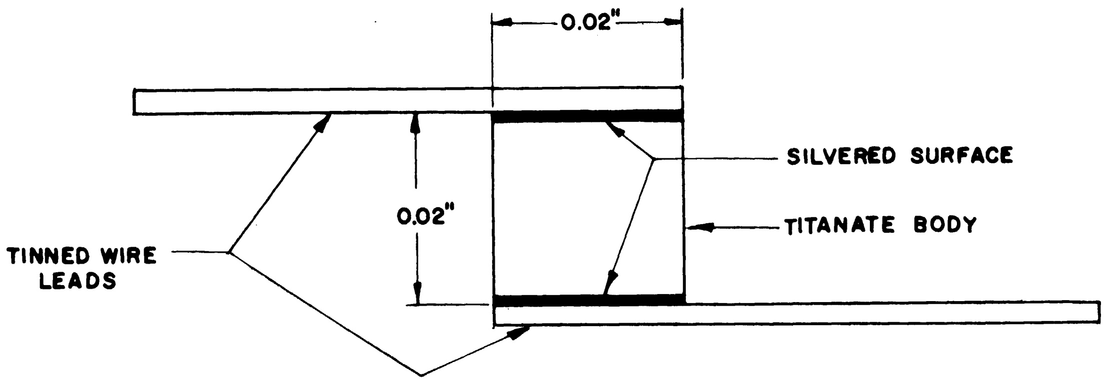
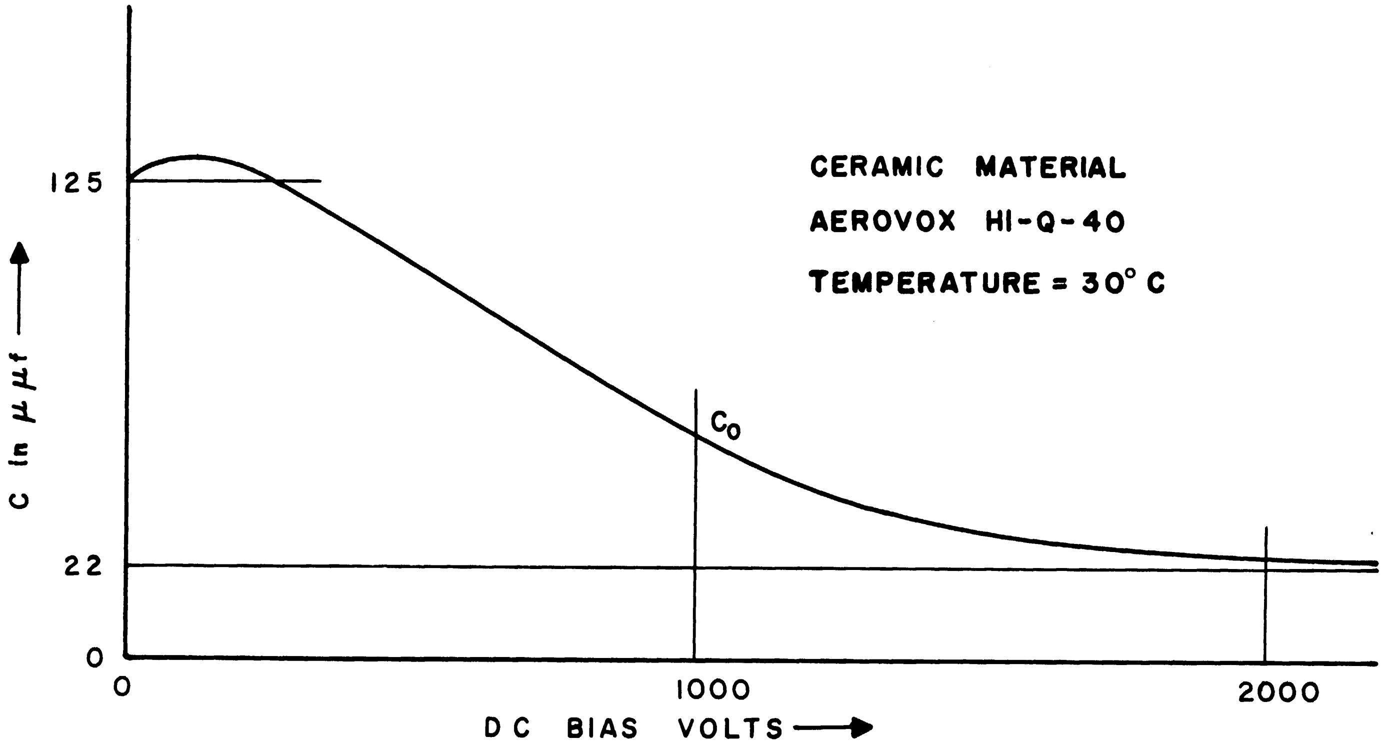


FIG. 26. CONSTRUCTION OF THE LOW VALUE TITANATE CAPACITORS.



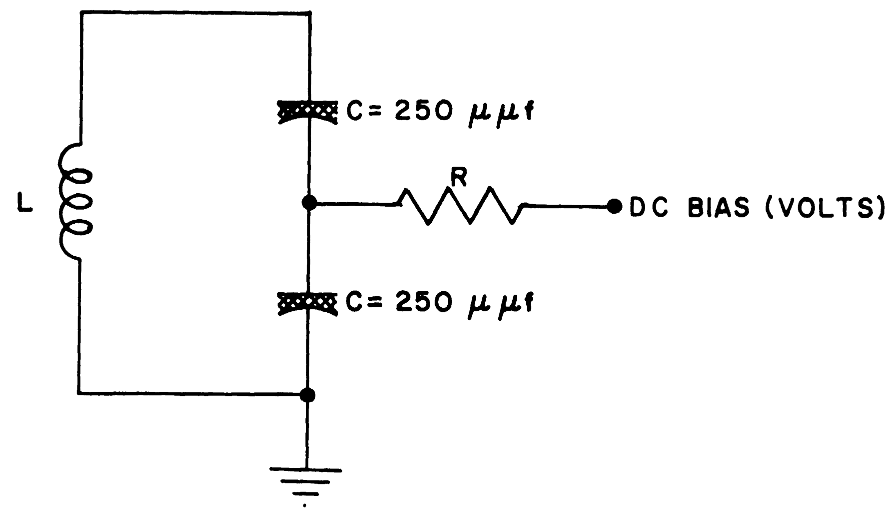


FIG. 27.
TUNED RESONANT CIRCUIT

to support the theory that a gradual change in the basic properties of the material does take place. One effect was the gradual increase in Q of the material over a three month period.

A.3 Evaluation of Michigan Hi-Q Material as Tuning Elements

Materials[1] have been developed in the University of Michigan Solid State Laboratory which have Q's ranging from about five to ten times as high as commercial materials. While this material also has about the same degree of non-linearity, it was decided not to implement their use in the front end assemblies at the present time for the following reasons:

(1) Machining and packaging these materials in a form suitable for use as tuning elements has proved to be difficult.

(2) The material is very temperature-sensitive.

(3) The voltage breakdown strength of the dielectric material, when used as a tuning element, was found to be low.

(4) Definite changes in the material structure due to aging are in evidence.

The methods used for machining commercial materials are not satisfactory for the Michigan materials since they are exceptionally hard and tend to shatter under ordinary machining processes. New machining techniques are presently being investigated which should greatly simplify this problem.

Since the material is quite temperature-sensitive when compared with presently used commercial materials it would be necessary to utilize a fairly precise thermostating arrangement even under normal laboratory use in order to obtain proper operation of the front end assemblies. Recent work in the material

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1 Progress Report No. 17, Task EDG-4, Engineering Research Institute University of Michigan, July 1956.

Progress Report No. 18, Task EDG-4, Engineering Research Institute University of Michigan, February 1957.

development program indicates that it is quite feasible to reduce the temperature-sensitivity while maintaining the desirable properties of the materials. This, in turn, would greatly reduce the thermostating requirements.

Although the voltage breakdown strength of the dielectric material when used as tuning elements was found to be somewhat low, it is felt that this could be due to improper firing techniques and machining procedures. The voltage break-down properties will undoubtedly improve with advances in firing and machining procedures.

Aging, i.e., gradual changes in both the color and Q value of the material, indicates that the state of the material immediately following firing was only metastable. Over a period of approximately three months following firing the material turned from jet black to brown, and zero-field Q values increased from approximately 150 to 250 at 50 MC.

When the problems connected with the development of stable, trouble-free tuning elements are solved it should not prove to be a difficult task to implement their use in receiver work.

APPENDIX B

RECEIVER PERFORMANCE TESTS

B.1 Introduction

The three major parameters that have been generally accepted for describing receiver performance are discussed in this appendix. These are: (1) sensitivity; (2) selectivity; and (3) dynamic range.[1]

The tests involved in determining these factors for the Petr Pan Receiver are described in Sections B.2, B.3 and B.4.

B.2 Receiver Sensitivity

In measuring receiver sensitivity, the following factors must be considered: (1) noise limitations imposed by the RF and IF sections of the receiver; (2) adequacy of over-all receiver gain to amplify weak signals to a useful level at the output; and (3) amount of signal required to produce some arbitrary output level.

When the sensitivity of a receiver, which can be considered a measuring device, is described, both "full-scale" sensitivity and noise threshold must be specified. Full-scale sensitivity depends upon over-all receiver gain, while noise threshold depends upon noise sources and signal handling efficiency of the receiver input circuits.

Although the noise limitation has come to be defined in terms of a noise figure which relates only to the theoretical input noise of an ideal receiver without regard to total gain, the total gain must be taken into consideration. The gain depends upon the service for which the receiver is intended.

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[1] "Standard on Radio Receivers", Proceedings of the IRE, Inc., New York, page 14-15, 1938.

Receiver sensitivity can be tested and described in at least two ways:

(1) Compare the actual noise threshold of a receiver with that of a theoretically noise-free receiver having the same bandwidth.

(2) Determine the input CW signal which produces a "pip" on the scope that is twice the height of the residual noise.[1]

B.2.1 Noise Figure. The figure of merit that compares the actual noise threshold of a receiver with that of a theoretically noise-free receiver having the same bandwidth is termed the noise figure. Since the principles of noise figure measurements can be found in the literature, only the basic relationships and the methods used to obtain these relationships will be presented here.

Consider a receiver designed to operate from a voltage source of internal resistance, R. The thermal noise voltage developed in R (theoretical minimum noise of the receiver) can be thought of as an equivalent constant-voltage generator in series with R as shown in Figure 28. The rms voltage (E) of this generator is given by the relation

$$E^2 = 4\ KTR\ \Delta f \tag{B.1}$$

where: K = Boltzmann's Constant (1.37^{-23} x 10 joules per degree Kelvin);

T = temperature of R (degrees Kelvin);

Δf = receiver noise bandwidth.

Let E' be the actual noise of the receiver, including the theoretical noise referred back to the voltage source at the input. Again this is an equivalent noise voltage in series with the resistance R replacing E in Figure 28. The noise figure may now be defined as:

$$F = \left(\frac{E'}{E}\right)^2 \tag{B.2}$$

This is a power measurement, although the measurement of E' can be made from either voltage or power. The ratio is usually expressed in db as:

[1] J. B. L. Foot, "Wide Band V.H.F. Panoramic Receiver", Wireless World, Sept. 1953.

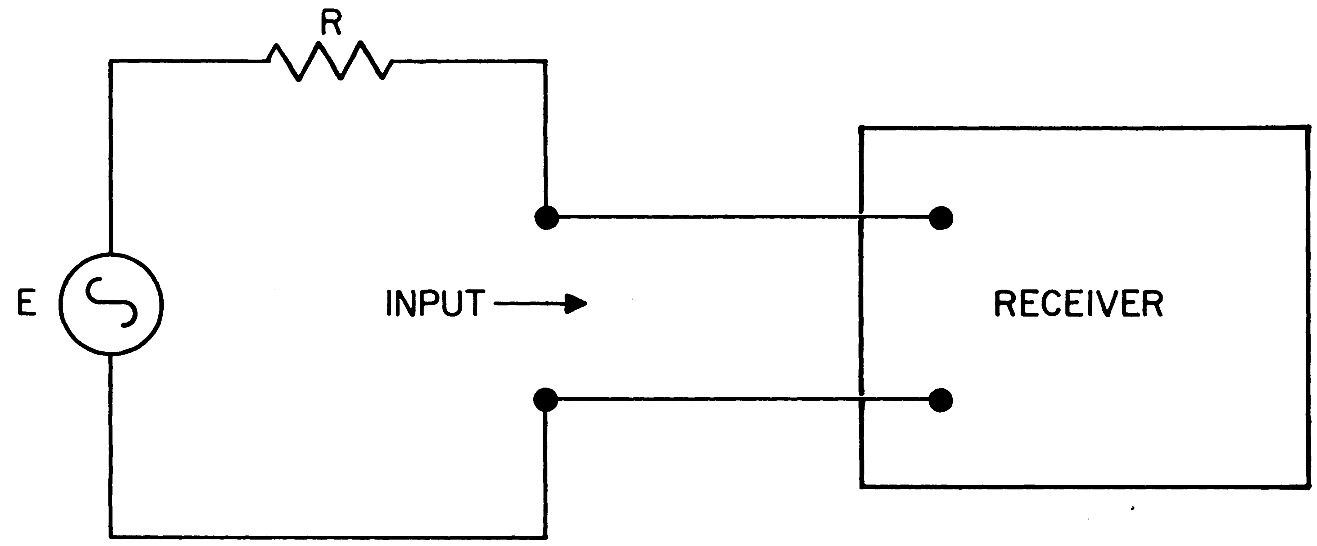

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FIG. 28. EQUIVALENT CIRCUIT FOR NOISE VOLTAGE, E, GENERATED IN A RESISTANCE, R, CONNECTED ACROSS THE INPUT TERMINALS OF A RECEIVER.

$$\text{Noise Figure (in db)} = 20 \log_{10} \frac{E'}{E} \tag{B.3}$$

E' can be measured by two methods. In one method the test generator provides noise of adjustable power and known amount having a uniform frequency spectrum over the receiver passband, and presents the specified source impedance for which the receiver is intended. In the frequency range of interest a temperature-limited diode for which the noise component of the plate current is proportional to the direct current through the diode can be used as a noise generator. This relation has been shown to be:

$$I^2 = 2e\, I_{dc}\, \Delta f \tag{B.4}$$

where: e = charge on the electron (1.59×10^{-19} coulomb);

I_{dc} = direct current in the diode (amperes);

Δf = bandwidth for which I is the rms noise current.

The noise generator (shown in Figure 29) is connected to the receiver through the proper source impedance. Noise from the generator is increased from zero until the output noise of the receiver is doubled. The added noise is then equal to the actual receiver noise, referred to the input, plus thermal noise from the source resistance, R. In making the measurement of output noise the principle requirement is that all noise sources be included. It can be assumed that noise generated in the receiver beyond the first IF stage is ineffective, since the signal is large compared with the noise at this point. Thus, the noise can be measured at any point following these first stages. However, the indicator must correctly add the rms noise of the receiver and that from the noise generator.

It should be noted that receivers of differing band-pass characteristics can be compared more readily by means of the noise generator method. The noise power produced by the generator is directly proportional to Δf, as indicated by Equation B.5.

$$(E')^2 = (IR)^2 = 2e\, I_{dc}\, R^2\, \Delta f \tag{B.5}$$

But, since $F = \left(\frac{E'}{E}\right)^2$ and $E^2 = 4\ KTR\ \Delta f$,

then $F = \dfrac{e\ I_{dc}\ R}{2\ KT}$

Note that Δf has dropped out of the solution. This is quite helpful because Δf is not easy to measure.

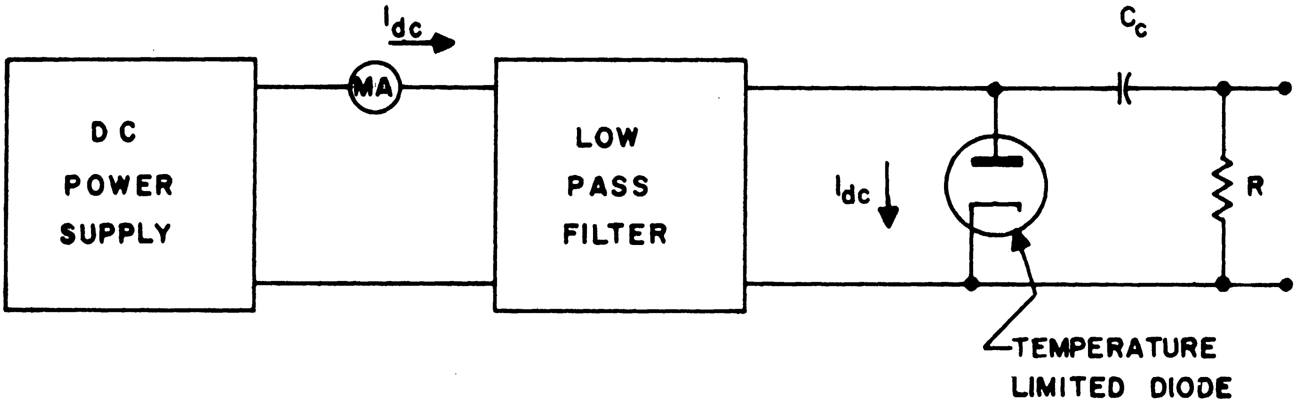

FIG. 29. BASIC NOISE GENERATOR

The second method of measuring the actual receiver noise, referred to the input terminals, is to use a CW signal to supply sufficient energy to double the receiver output indication due to the noise voltage E'. The basic requirement is that the output indicator respond identically to the summation of CW-plus-noise energy as it does to noise alone. The expression for noise figure then becomes

$$F = \frac{E_s^2}{4\ KTR\ \Delta f} \tag{B.6}$$

where E_s is the CW signal applied in series with R to double the output indication due to the actual receiver noise, E', and Δf is obtained from the selectivity characteristics of the receiver.

B.2.1.1 Noise Figure Measurements. The method used in obtaining the noise figures of the FE units plus IF strip is described in this section.

After determining that the 2nd detector is operating in the linear portion of its range during normal receiver operation, the test equipment was set up as shown in Figure 30. The scope was calibrated in volts/inch of deflection on the "Y" axis, and a convenient base line selected. With the dial of the noise generator set at zero (which terminates the line in its characteristic impedance), and the receiver operated under normal operating conditions, the DC voltage level due to the internal noise of the FE under test (plus IF strip) is measured from the base line. The noise figure (i.e., the noise power introduced at the receiver input terminals which causes the noise power at the 2nd detector to increase by a factor of 2 over that value obtained due to the internal noise alone) can be read directly in db from the dial plate of the noise generator. It should be noted that, since the scope is a voltage measuring device, the DC level of noise voltage measured from the base line increases by a factor of approximately $\sqrt{2}$ as the noise power at the 2nd detector is doubled.

The noise figure measurements made under non-sweeping conditions at room temperature and input impedance levels of $\sim 50\ \Omega$ are given in Table B.1.

Table B.1

FE Assembly	Frequency in MC	Noise Figure in db
A	55	~ 7
B	100	~ 11
C	160	~ 13

However, when noise figure measurements were taken in the electronic tuning mode it was found that the overall noise figure increased and seemed to be a function of the tuning voltage. Most of the noise observed can be traced to properties of the

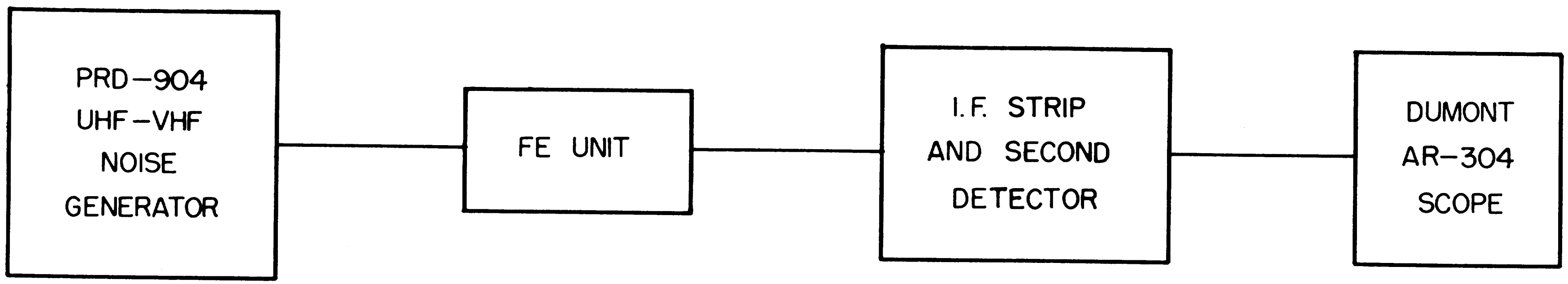

FIG. 30. TEST SET UP FOR NOISE FIGURE MEASUREMENTS.

ferroelectric material used to manufacture the tuning capacitors.

There are several possible mechanisms of noise generation in dielectric materials,[1] such as noise associated with ionic conduction through the volume and over the surface of the specimen, noise resulting from imperfectly applied electrodes, noise as a function of temperature, noise as a function of electrical stress, etc.

Noise as a function of electrical stress plays a most important part in the contribution of noise when the receiver is switched from manual to electronic tuning. When a slowly varying electric field (e.g., 60 ~ sinewave) is applied to a ferroelectric capacitor, discontinuous changes occur in the polarization due to domain wall jumps which give rise to pulses of current. These pulses of current probably constitute the major part of the noise in the ferroelectric capacitors.

The P-E loop (polarization versus electric field) in Figure 31 shows the relationships between polarization and applied electric field for a typical ferroelectric ceramic operated below its Curie temperature. Due to the rapid changes in polarization on the steep portions of the loop a large noise output is generated. Conversely, nearer the tips of the loop, the noise output is much smaller and frequently absent altogether. Thus, it can be seen that the noise in the ferroelectric material is a varying quantity, with its magnitude being dependent upon four main parameters. These parameters are:

(1) Rate of change of the applied electric field;

(2) Instantaneous value of the driving voltage;

(3) The characteristics of the material, and;

(4) The temperature at which the sample is being operated.

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[1] A. C. Kibblewhite, "Noise Generation in Crystal and in Ceramic Forms of Barium Titanate when Subjected to Electric Stress," The Journal of the Institute of Electrical Engineers, Vol. 102, Part B, page 59-68.

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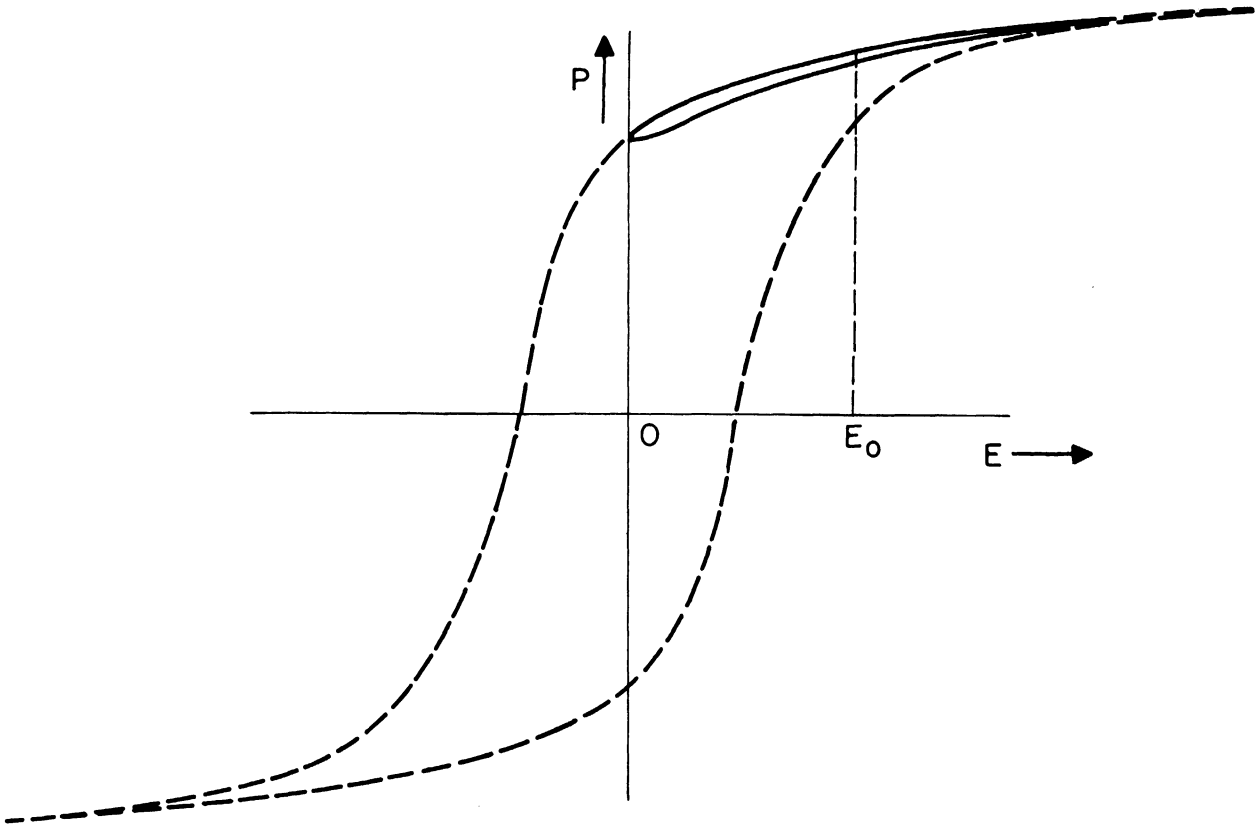


FIG. 31. POLARIZATION AS A FUNCTION OF ELECTRIC FIELD IN A TYPICAL TITANATE CERAMIC.

Figure 32 shows qualitatively the noise (central trace) in a typical ceramic capacitor as a function of the rate of change of the applied electric field. It should be noted that the maximum noise output does not occur at the maximum value of dE/dt but some time later at maximum dP/dE.

Under actual operating conditions the receiver is not cycled around a complete P-E loop, but is cycled over the portion of the P-E loop as shown by the solid lines of the curve of Figure 31. To determine the expected increase in noise output of a dielectric-tuned receiver being swept at a 60~ rate, the following test was performed. The noise figure of a front end assembly was obtained at a DC bias level of 1000 volts (i.e., E_o = 1000 v), and with zero sweep. The sweep voltage (60~ AC) was applied in 100 volt steps and the noise figure of the front end assembly measured with each incremental increase in sweep. The results of this test are shown by the graph of Figure 33.

B.2.2 Measured CW signal Sensitivity. Figures 34, 35, and 36 indicate the response or the three front end assemblies. In each case the photo on the left was obtained by connecting a Hewlett-Packard Model 608-D signal generator by means of a short piece of 50-Ω coaxial cable to the input terminals of the front end under test and measuring directly. The photos on the right were made with a short antenna connected to the input terminals and indicate the local strong spectrum. Each front end assembly was biased to 700 volts DC and swept by means of a 500 volt AC (effective value) sinusoidal signal.

B.3 Receiver Selectivity

In measuring receiver selectivity, the following factors must be considered: (1) image signal rejection; (2) harmonic interference effects; and (3) IF break-through.

The measurement of spurious responses consists in determining the sensitivity of the receiver at one or more arbitrary frequencies within the normal

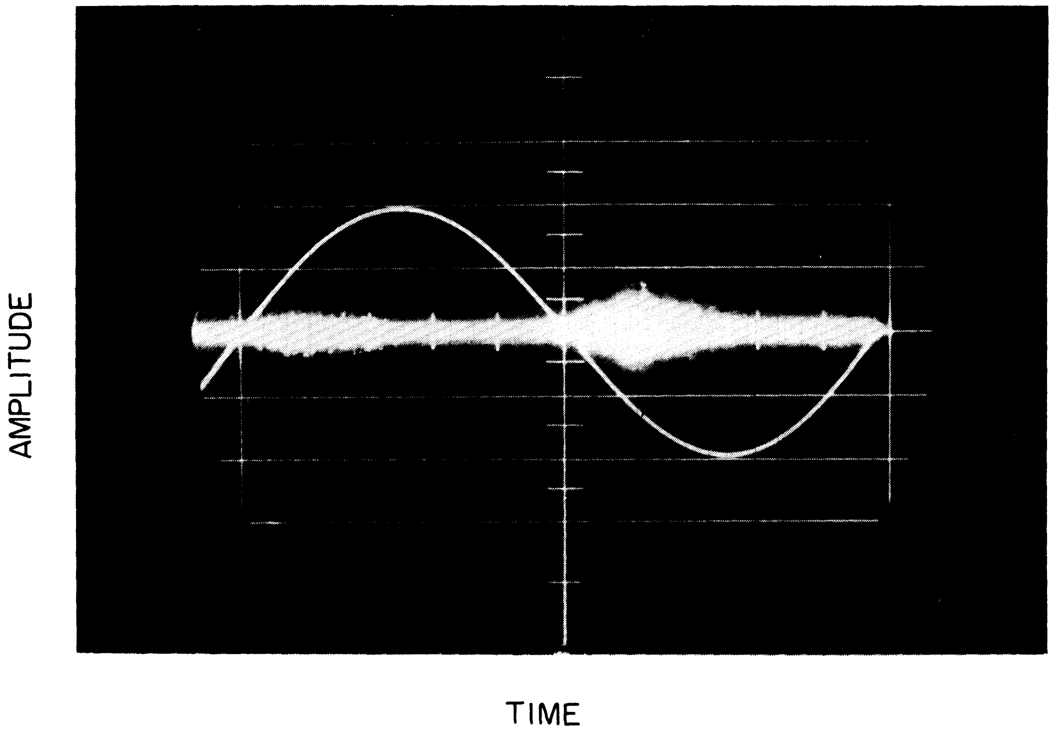


FIG. 32. NOISE AS A FUNCTION OF ELECTRICAL STRESS IN A TYPICAL TITANATE CERAMIC AT 27° C. (From an unpublished study of Barkhausen noise, by M. H. Winsnes)

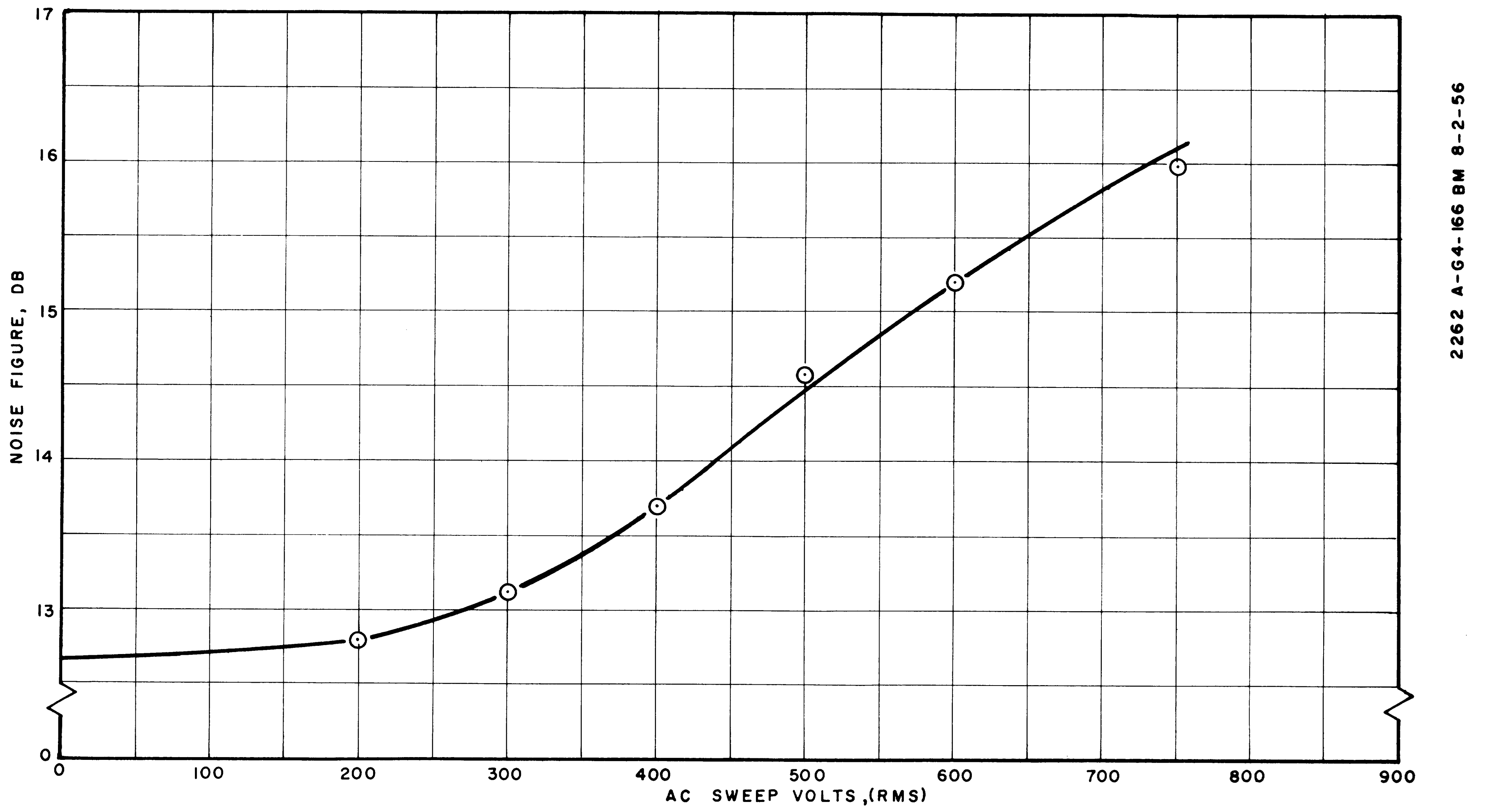
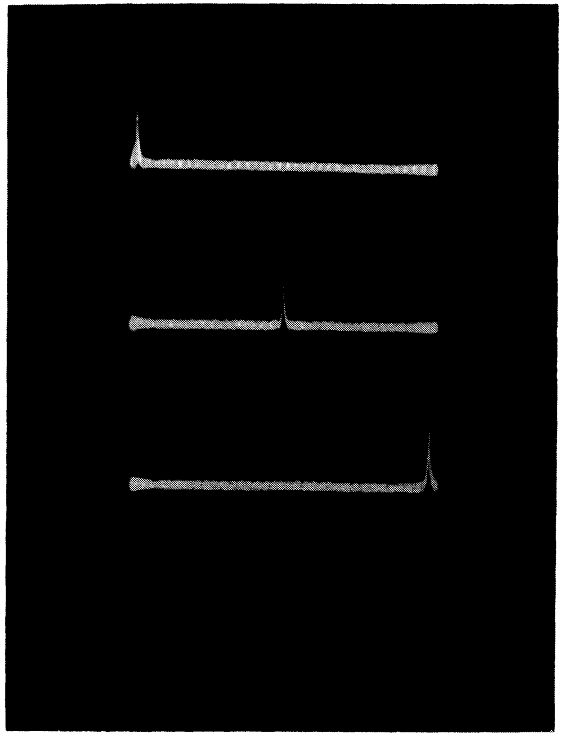
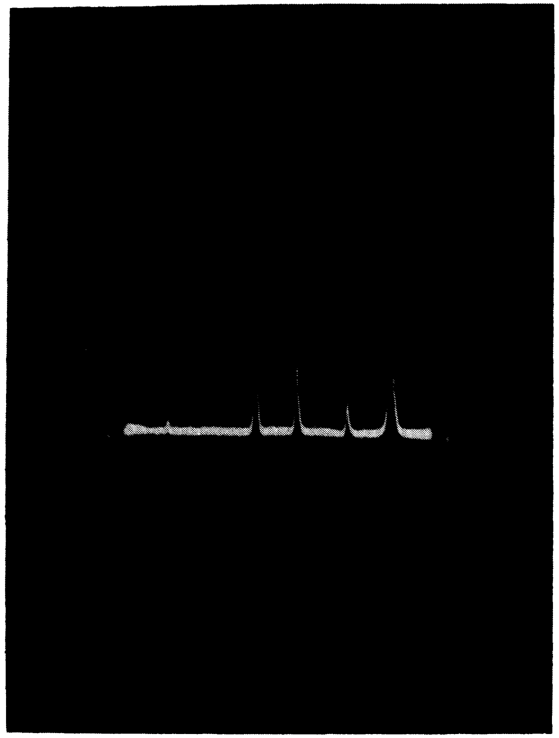


FIG. 33. RECEIVER NOISE FIGURE
FE-B (70 - 130MC)
E_0 = 1000V (DC BIAS LEVEL)



HEAD A
700 DCV 500 ACV
RF 150 V IF 60 V
35 MC 5 μV INPUT
58 MC 5 μV INPUT
69 MC 5 μV INPUT



HEAD A
700 DCV 500 ACV
RF 150 V IF 60 V
TYPICAL SIGNAL DISPLAY
WITH ANTENNA.
2ND SIGNAL FROM RIGHT
IS 5 μV GEN. SIGNAL.

FIG. 34.

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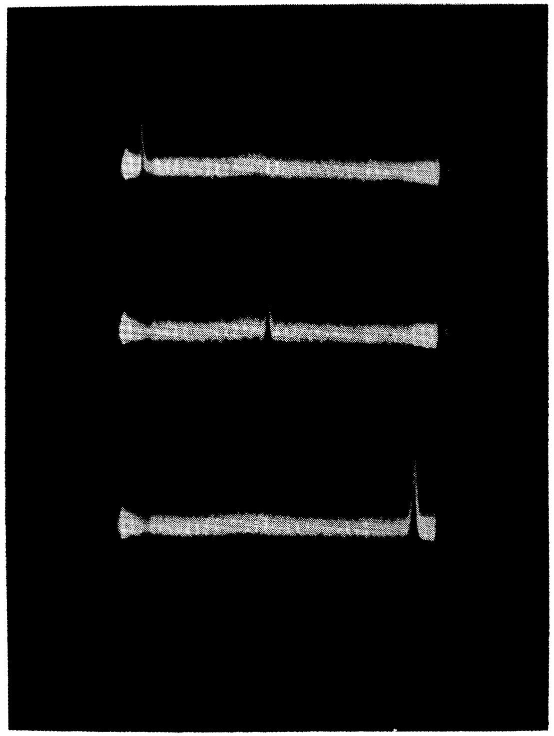
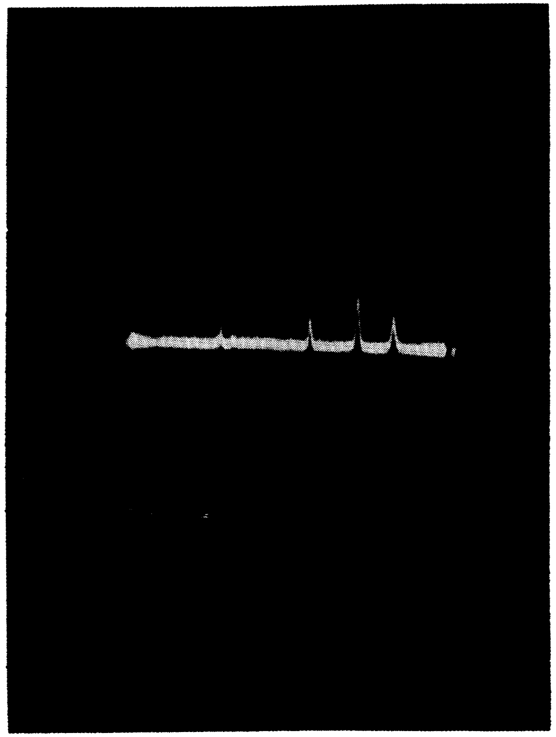
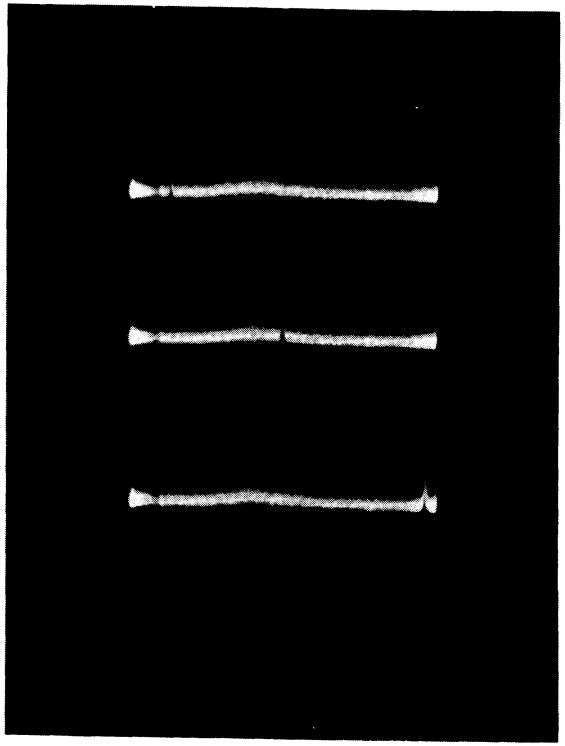


HEAD B
700 VDC 500 VAC
RF 150 V IF 60 V
75 MC 15μV INPUT
110 MC 5μV INPUT
128 MC 5μV INPUT

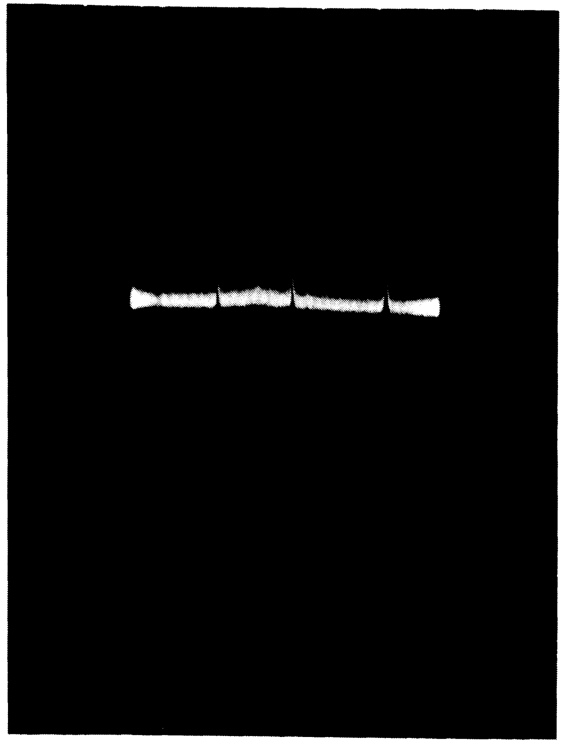


HEAD B
700 VDC 500 VAC
RF 150 V IF 60 V
TYPICAL SIGNAL DISPLAY
WITH ANTENNA.
SIGNAL ON RIGHT IS 5μV
GEN. SIGNAL.

FIG 35



HEAD C
700 VDC 500 VAC
RF 150 V IF 60 V
130 MC 15μV INPUT
175 MC 7μV INPUT
197 MC 7μV INPUT



HEAD C
700 VDC 500 VAC
RF 150 V IF 60 V
TYPICAL SIGNAL DISPLAY

FIG. 36.

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tuning range, and, for each of these, determining the sensitivity for spurious signals by tuning the signal generator to the appropriate spurious frequencies. The ratio of sensitivity for undesired to that of desired signals, expressed in decibels, is defined as the rejection for each particular type of spurious response. The most serious of these spurious signals is usually the image; i.e., the signal whose frequency differs from that of the local oscillator by an amount equal to the intermediate frequency, but whose frequency appears in the spectrum on the opposite side of the local oscillator frequency from the desired signal. The difficulty of obtaining good image rejection makes its determination an important test for receivers intended to indicate a given signal at only one spot on the dial. The image-signal sensitivity was found by adjusting the receiver tuning frequency to a convenient value, f_s, and then varying the signal generator carrier frequency over a range covering a frequency of $f_s + 2f_i$, where f_i is the intermediate frequency. The carrier voltage was adjusted (when the frequency point of maximum video output had been found) until a convenient voltage output level, V_1, was obtained. The signal generator was then set to the receiver tuning frequency, f_s, and adjusted until the same voltage output level, V_1, attained with the image signal was again obtained.

Image rejection data are given below in Table B.2 for the three FE assemblies.

Table B.2

FE Assembly	Signal Frequency (MC)	Image Rejection (db)
A	55	45
B	100	35
C	160	26

The difficulty of building good RF preselection to remove the image makes it necessary to test VHF receivers for spurious responses of other sorts, principally those involving multiples of the local oscillator frequency and of the signal frequency. These are commonly referred to as harmonic responses. Even though the receiver oscillator voltage and the signal voltage may initially be devoid of harmonics both of these RF voltages are applied to the mixer, which is a nonlinear circuit element. In the output of such a nonlinear circuit there appear frequencies comprising differences (and also sums) of multiples of the original frequencies. When the difference frequency of any pair of multiples is equal to the intermediate frequency of the receiver the following formulas tell where the signal generator should be set in order to measure each of these harmonic responses.

For FE units in which the local oscillator is tracked at a frequency higher than the signal frequency:

$$f_s = \frac{m}{n} \cdot f_{dial} + \frac{fi}{n} (m \pm 1),$$

where f_s = signal-generator setting where spurious response will be found;

n = order of the multiple of signal-generator frequency;

m = order of the multiple of local oscillator frequency;

f_{dial} = indicated receiver frequency; and

f_i = intermediate frequency.

The most serious harmonic response is due to 2nd harmonic local oscillator output. To determine the magnitude of oscillator 2nd harmonic interference effects, the same test procedure as that described above can be used. Oscillator 2nd harmonic interference effects are shown in Table B.3.

Table B.3

FE Assembly	Signal Frequency in MC	db Rejection of 2nd Harmonic Oscillator Interference Effect
A	36	>100
B	70	> 65
C	130	>100

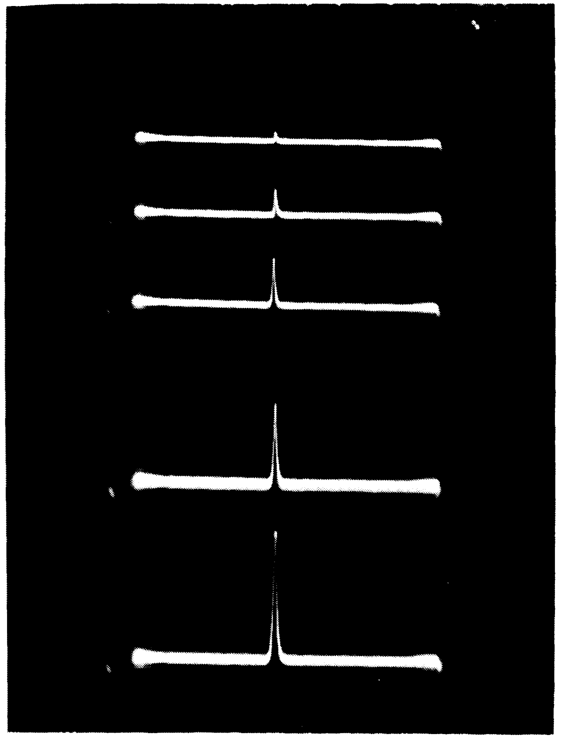
The other spurious response to be tested is IF break-through; i.e., the response due to a strong signal at intermediate frequency reaching the IF amplifier in spite of the selectivity and shielding in the RF circuits. The test consists simply in applying a signal of intermediate frequency to the receiver input and comparing the sensitivity thus measured with that for a signal to which the receiver is tuned. To determine the magnitude of IF break-through, the same test procedure as described above can be used. The magnitudes of the IF break-through are shown in Table B.4.

Table B.4

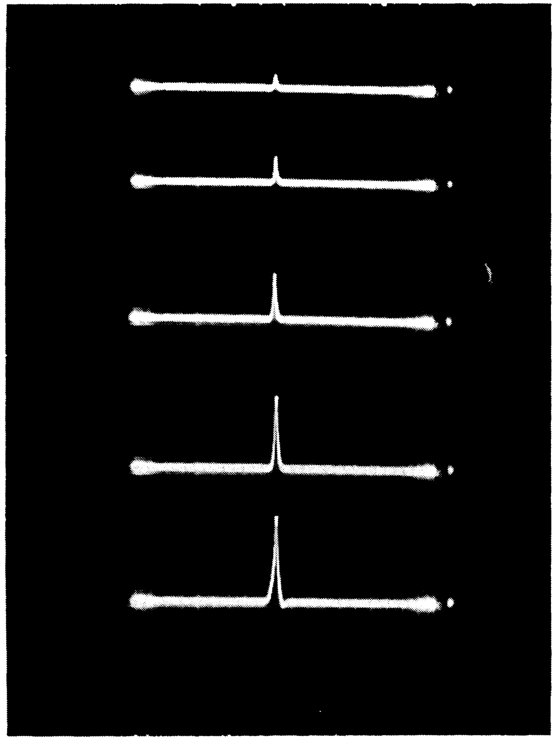
FE Assembly	Frequency in MC	db Rejection of IF Break-through Frequency
A	55	> 50
B	100	> 50
C	160	> 50

B.4 Dynamic Range of the Receiver

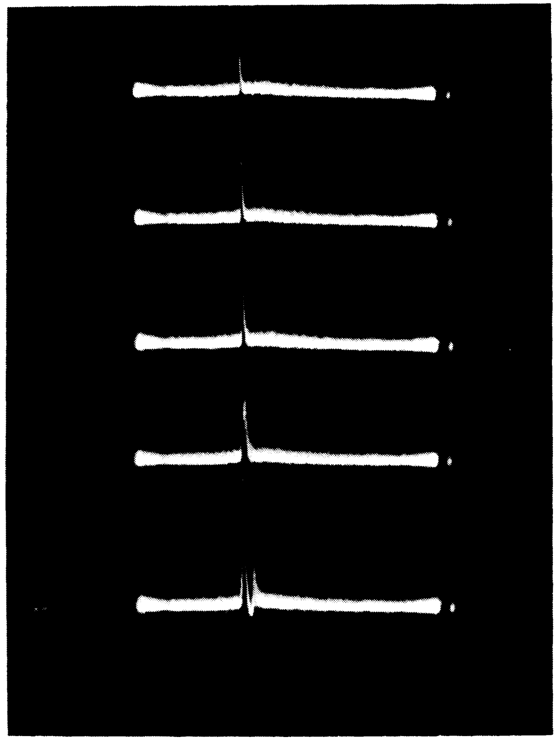
The dynamic range of the receiver can be adequately described by reference to Figure 37. In each case a Hewlett-Packard 608-D signal generator was connected to the input terminals of the front end under test by means of a short length of 50 Ω coaxial cable and the input signal was increased until overload tendencies were indicated. Each front end assembly was biased to 700 volts DC and swept by means of a 500 volt AC (effective value) sinusoidal signal.



HEAD A
700 VDC 500 VAC
RF 150 V IF 60 V
"DYNAMIC RANGE"
AT 55 MC
2 μV INPUT
5 μV INPUT
10 μV INPUT
20 μV INPUT
50 μV INPUT



HEAD B
700 VDC 500 VAC
RF 150 V IF. 60 V
"DYNAMIC RANGE"
AT 110 MC
5 μV INPUT
10 μV INPUT
20 μV INPUT
50 μV INPUT
500 μV INPUT



HEAD C
700 VDC 500 VAC
RF 150 V IF 60 V
"DYNAMIC RANGE"
AT 160 MC
5 μV INPUT
10 μV INPUT
20 μV INPUT
50 μV INPUT
500 μV INPUT

FIG. 37.

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APPENDIX C

ALIGNMENT PROCEDURE FOR PANORAMIC RECEIVER

C.1 Horizontal Sweep Circuit

The sawtooth generator operates as follows. The 60~ line voltage is stepped up and positively and negatively clipped. The semi-square wave is differentiated and the positive going spike is discriminated against. The negative spike is fed to pin No. 2 of V-3, amplified and fed to pin No. 7 of V-3. This positive spike results in a surge of current through this section of V-3 which charges the cathode capacitor. When the spike is gone the cathode capacitor charge is sufficient to hold this section of the tube in cut-off through the long RC discharge in the cathode network. A sawtooth is taken off pin No. 8 of V-3, amplified, and applied to the horizontal plates of the CRT in push-pull fashion.

In aligning the sawtooth generator, the following procedure is observed.

(1) Trigger amplitude to pin No. 2 of V-3 is not critical, but should be of sufficient amplitude for positive action. Adjust synchronizing control (back plate) accordingly.

(2) With a scope in pin No. 8, of V-3 adjust frequency vernier (back plate) so that no clipping of sawtooth occurs. If the RC time in the cathode network is too short, this section of V-3 will come out of cut-off before trigger time. RC time (frequency vernier) should not be made too long since this will affect the horizontal width. The "X" position and "X" gain controls for the horizontal sweep circuit are front panel adjustments.

C.2 Delay Multivibrator

The delay MV is a one-shot cathode-coupled circuit. It is fired at zero time by a positive trigger on pin No. 2, V-12. The positive time of pin No. 6, V-12, is adjusted on the front panel.

In aligning the delay MV the stability control is used to adjust bias on the normally cut-off section of V-12. Adjust for stability while viewing the step on the display unit scope.

C.3 Step Shaper

The step shaper is a diode clipping circuit to shape the square wave from pin No. 6 of the delay MV.

To align the step shaper, with the step on the display unit scope, adjust setup for sharpest possible step. This is not a critical adjustment.

C.4 Expand Multivibrator

The positive square wave starting at zero time from pin No. 6 of the delay MV is differentiated, and V-13 discriminates against the positive spike at zero time. The negative trigger which occurs at a selected delayed time is fed to pin No. 2 of V-14. The expand MV, V-14, is a plate-coupled one-shot multivibrator with one grid at a positive potential. Negative output is taken from pin No. 6, V-14, and provides a sharper RC discharge to the horizontal cathode network for expand MV time.

To align the expand MV, the following procedure is observed.

(1) The stability control should be adjusted to cut-off of normally cut-off section. The expanded step may be observed on display unit for any sign of free running.

(2) Expand width may be varied over a small range by adjustment of the notch width control.

C.5 Sweep Voltage

The high voltage rectifier tied to the top of the high voltage transformer serves as a peak charging path for the sweep voltage capacitor.

The sweep capacitor discharges through an RC time toward a negative potential supplied by a high voltage rectifier connected to the center tap of the high voltage transformer. The total discharge time is 240°.

The sweep voltage alignment is accomplished as follows:

(1) With an RF head connected and using a grid dipmeter or Hewlett-Packard signal generator for a signal input, turn power-stat (back plate) until signal of highest desired frequency appears at right hand side of the sweep while facing the display unit.

(2) Adjust RC time potentiometer and bias control (top of chassis) simultaneously for maximum possible linear change in frequency on the calibrated dial, and to a point just beyond which no noise spikes appear on the scope. The noise spikes are generated by the ferroelectric capacitors if the sweep voltage is allowed to sweep through the zero potential point.

C.6 Front End Assembly

The RF section consists of a cascode input (for low noise) and a mixer. All tanks are ferroelectrically tuned and sweep approximately a 2:1 frequency ratio, tracking 20 MC below the local oscillator.

The local oscillator is a Colpitts ultra-audion type whose tuned circuit consists of a coil parallel with a high voltage center-tapped set of ferroelectric capacitors, parallel with the interelectrode capacity from plate to ground and grid to ground in series, parallel with the interelectrode capacity from plate to grid. A quartz trimmer is placed in parallel with the plate to ground interelectrode capacity for optimum feedback over a wide frequency range.

The RF head alignment procedure is as follows:

(1) Using a Hewlett-Packard signal generator, the RF section should be peaked over the sweep range of the head by slug-tuning all tanks concerned.

(2) The oscillator should be tuned by varying the quartz trimmer and tank slug simultaneously until a compromise between maximum tuning range and maximum sensitivity is obtained.

(3) It may be necessary to tune out oscillations with the 20 MC transformer slug and the mixer tank slug.

C.7 Blanking Multivibrator

The blanking MV is a one-shot cathode-coupled circuit which receives a positive trigger at zero time. The variable-length negative square-wave from pin No. 1 of V-9 is clamped and fed to the control grid of the CRT.

In aligning the blanking MV, the following procedure is observed.

Since the FE sweep voltage rises sharply for 90° before the RC discharge begins, the entire range of frequencies is swept by the RF head before it reaches its highest tuned point in frequency and is ready to display a signal in its spectrum during sweep time. This results in a double image of a signal at the high frequency end of the spectrum. Vary the blanking time (back plate) until a signal from a Hewlett-Packard signal generator or a grid dip meter will just run off the right side of the sweep high frequency end without showing a double image.

C.8 Calibration

If a complete alignment procedure is followed, it may be necessary to recalibrate the scale. Chances are this will not be necessary if the RF head and FE sweep voltage alignments are not made. Check the calibrator occasionally; the difference from an accurate source should be within 1%.

Recalibration is accomplished by using a standard RF source such as a Hewlett-Packard signal generator or grid-dip meter. A signal of approximately 2 micro-volts should be used. Vary the frequency knob until the step bisects the displayed signal; that is, the top corner of the step should be at the topmost part of the signal. Mark the frequency on a piece of scotch writing tape which can be easily applied to the face of the frequency dial. Do this at as many frequencies as desired.

C.9 Operation

The operational sequence described below should always be observed.

(1) Turn on display unit power supply (front panel) and auxiliary power supply (+300V) 30 seconds later. The equipment should be operated at room temperature and should be allowed a 45-minute warm-up period for accurate frequency readings. The warm-up period is an important procedure since the ferroelectric tuning capacitors used in the RF head are sensitive to temperature.

(2) Signals may be displayed on a linear sweep by setting the expand-step control (front panel) to the "off" position.

(3) To display an expanded signal the following steps should be observed:

a) Turn expand-step control to the "on" position, but keep the control fully counter-clockwise.

b) Move the expanded portion of the sweep to the desired signal by varying the frequency dial. If a signal is symetrically expanded, the calibrated dial should <u>not</u> be taken for an accurate frequency reading.

(4) To determine signal frequency the following procedure is recommended.

a) Turn expand-step control CW to desired step amplitude.

b) Vary the frequency dial until the vertical edge of the step bisects the signal; that is, the top corner of the step should be set at the topmost part of the signal. Read frequency directly off calibrated scale. For better accuracy always approach the signal by turning the dial clockwise. This eliminates variation due to mechanical play.

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